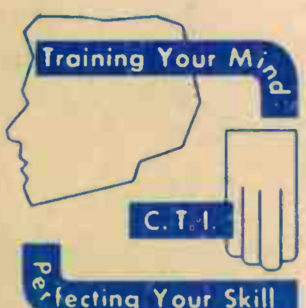
1→5

COMMERCIAL TRADES INSTITUTE

KNOWLEDGE + AMBITION = SUCCESS

LEARN . . .

ACT . . .



ENJOY . . .

"The heights by great men reached and kept
Were not attained in sudden flight,
But they, while their companions slept,
Were climbing upward in the night."

COMMERCIAL TRADES INSTITUTE

CHICAGO, ILLINOIS

TELEVISION

LESSON NO. 1 WHAT MAKES TELEVISION WORK?

Training Your Mind

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

COMMERCIAL TRADES INSTITUTE

Training Your Mind — C.T.I. — Perfecting Your Skill

LESSON NO. 1

WHAT MAKES TELEVISION WORK?

At two and one-fifth seconds after quarter past four in the afternoon, at Wrigley Field in Chicago, the batter connects with a fast one. In that same fraction of a second, although you may be hundreds of miles away, it is possible to see the hit, hear the crack, and watch the ball travel out, up, and over the wall for a homer. That's television!

The television "signal" which lets you see and hear that play may have come through hundreds of miles of coaxial cable, and through amplifier after amplifier on the way. Always the signal has originated in a television camera and has passed through the vast complexities of a transmitter. From the transmitter antenna that unbelievably minute force which is the signal flies through space to the antenna of your television receiver.

At your receiver the signal starts a whole train of events, with power building up higher and higher in paths which divide and subdivide. Finally, all but one of these paths come back together at the picture tube, and there you see the swing of the bat and the flight of the ball. The one remaining path leads to the loud speaker, from which you hear the "smack" and the cheers of the crowd. The whole thing happens in little more time than it takes for light to travel from this page to your eyes.

Nothing which ever came to pass in the world of science is more wonderful than television. A television receiver utilizes more different electrical principles than any other device you can think of. The ordinary television set contains four times as many tubes as an ordinary radio set. But it isn't the number of tubes which tells the real difference, for the television receiver employs all the principles found in sound radio and dozens of others which the radio man never has to think about.

When you look at the top of a television receiver which has been removed from its cabinet you see quite a few tubes and a number of rather large parts, as pictured in Fig. 1-1. It is from underneath the chassis that you see most of the small parts, especially those which become familiar to every service technician. In Fig. 1-2 we are looking at the underside of a 20-tube television set. This is a rather simple job, as receivers go, but in it there are 283 principal parts – not counting such minor things as wires, lugs, clips, insulators, mountings, screws, knobs, and all the other small parts. Eighty per cent of the principal parts are capacitors, inductors, and resistors of one type or another.

The words capacitor, inductor, resistor, amplifier, signal, and all the other technical terms are part of the language of television. Soon this language will be yours. A little later one of your fellow technicians may ask, "What was wrong with that job?" And you will answer, "Well, the discriminator in the horizontal afc was out of phase and put a blanking bar right down through the pattern. Not only that, there was a leaky capacitor on the grid of the vertical sweep oscillator, and the picture wouldn't stop rolling."

This television receiver, a scientific marvel whose working depends on the most delicate balances between forces great and small, brings forth living pictures and the accompanying sound at the flick of a switch and the turn of a dial – for anyone. Even a child can operate a modern television set. Still more remarkable, this

A

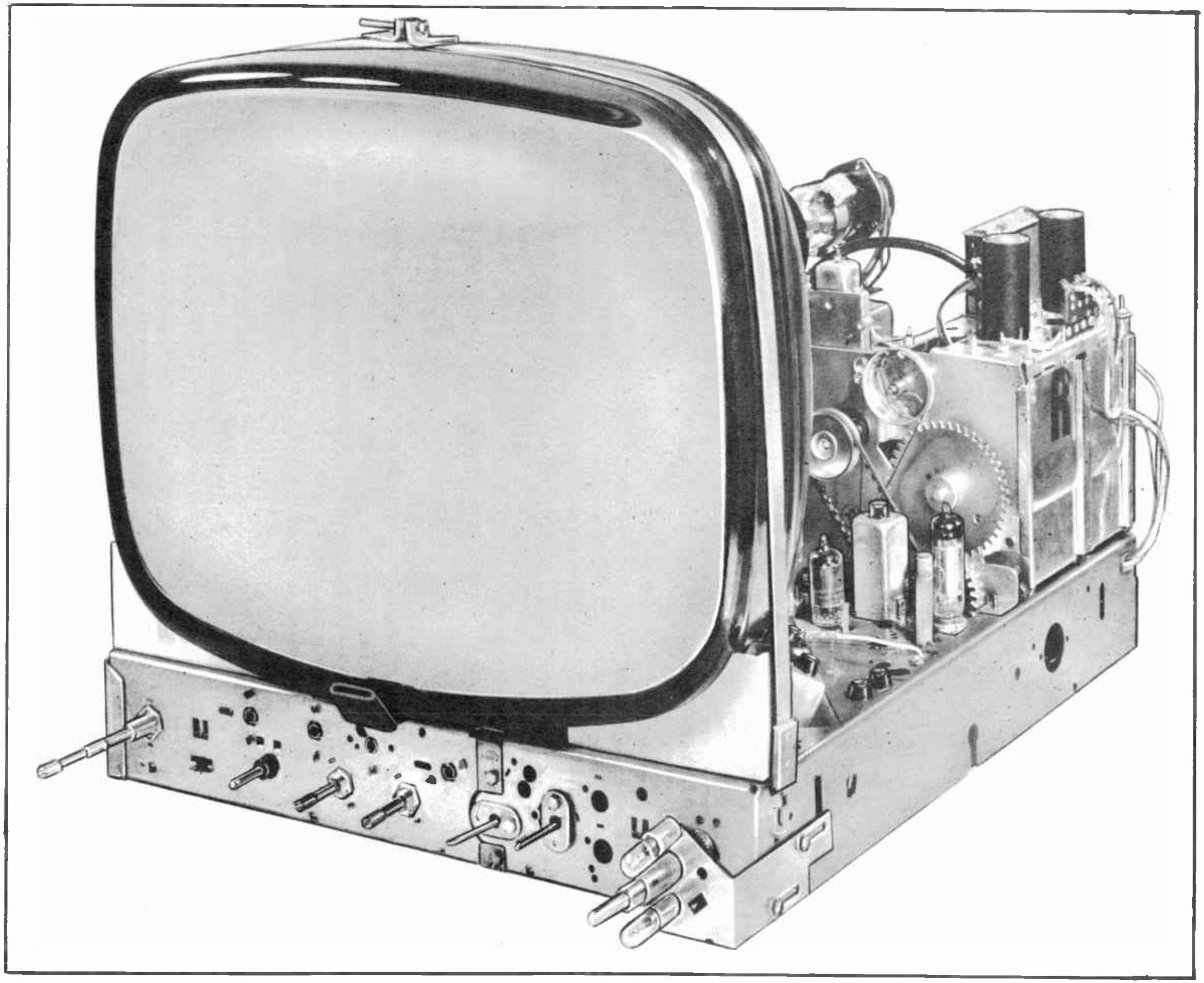


Fig. 1-1. On top of a television receiver chassis we find many tubes and most of the larger parts. (Courtesy Zenith Radio Corp.)

precision instrument is surprisingly tough. There is nothing the child can do with any of the front panel controls to put the receiver out of business. He can only throw it out of tune or out of synchronization.

Sad to relate, however, the forces which carry the ball game to your eyes and ears can be jostled about in space until you see blurs, ghosts, flashes, herringbone patterns, and lots of other things which certainly aren't pictures. Even though the signal gets into the receiver without mishap it still has a long ways to go, electrically speaking. The final picture, if one exists, may be out of focus, off center, full of "snow", all twisted and torn, too dark, or, as the photographers say, nothing but soot and whitewash. This is where you, the service technician, come in.

Now we don't intend to belittle the abilities of the radio service man, for reasons you soon will discover. A good radio service man has to know plenty. Servicing a radio is child's play, however, compared with servicing a television set. From the standpoint of the set owner this isn't so good. For one thing, it will be years before the supply of competent television technicians catches up with the need for them. If television sets continue to sell at the present rate of thousands every day, the need never will be fully satisfied.

From the standpoint of your own self-interest this prospect is decidedly rosy. When a television set goes haywire it may mean merely a burned out tube, which anyone can replace. This remark, about "anyone" replacing a tube, would bring a smile from the man who knows television. Replacement in any one of at least half the positions has to be followed by a job of realignment before performance comes back to normal. A tube which checks OK on any number of tube testers may or may not work in a television receiver. A tube which fails in one position may be entirely satisfactory in another spot.

Just to illustrate how trouble shooting really is carried out, let's assume that you are further along in your studies and are called to fix a set from which the sound is fine, but on the face of the picture tube there appears nothing but white, with some diagonal lines even whiter. Being a trained technician, you know instantly that the sound section and the sweep section are working. Doubtless you would put a test signal on the antenna terminals or on the mixer tube, and check results at the video detector output. Were everything OK there you would check for the test signal at the picture tube input. There you might find no signal at all. The rest would be easy, for the fault is between the detector and the picture tube. Maybe it is a blown coupling capacitor, maybe something else, but easily located with your electronic voltohmmeter.

The big point is that television servicing requires training of a very special kind, and familiarity with highly specialized testing equipment. The result, naturally, is more money per job, or per hour or minute on the job. Maybe you already know about the "service insurance" offered for many high grade receivers. The cost per year, for fixing whatever may happen, runs from forty to seventy-five dollars. It goes up as the receiver grows older. This means that the average yearly service expense on every receiver has been figured by experts as forty to seventy-five dollars. The makers say that during the next twelve months at least two million new television sets will go into the hands of owners. Multiply sets by dollars – and remember, this is the increase for just one year.

How are we going to learn to give television service which makes the owners happy, and makes us pay more income tax? What is the road to be traveled between here and real proficiency? Our "road map" will appear in this and the following lesson.

We have a lot of ground to cover. To travel the route in a reasonable time we must get into the actual practice of television without delay. Therefore, we shall not spend time just now on so-called theory and fundamentals, but rather commence learning how television works. Don't, however, get the wrong ideas about theory and basic principles. Many times you will encounter troubles which do not yield to any ordinary serv-

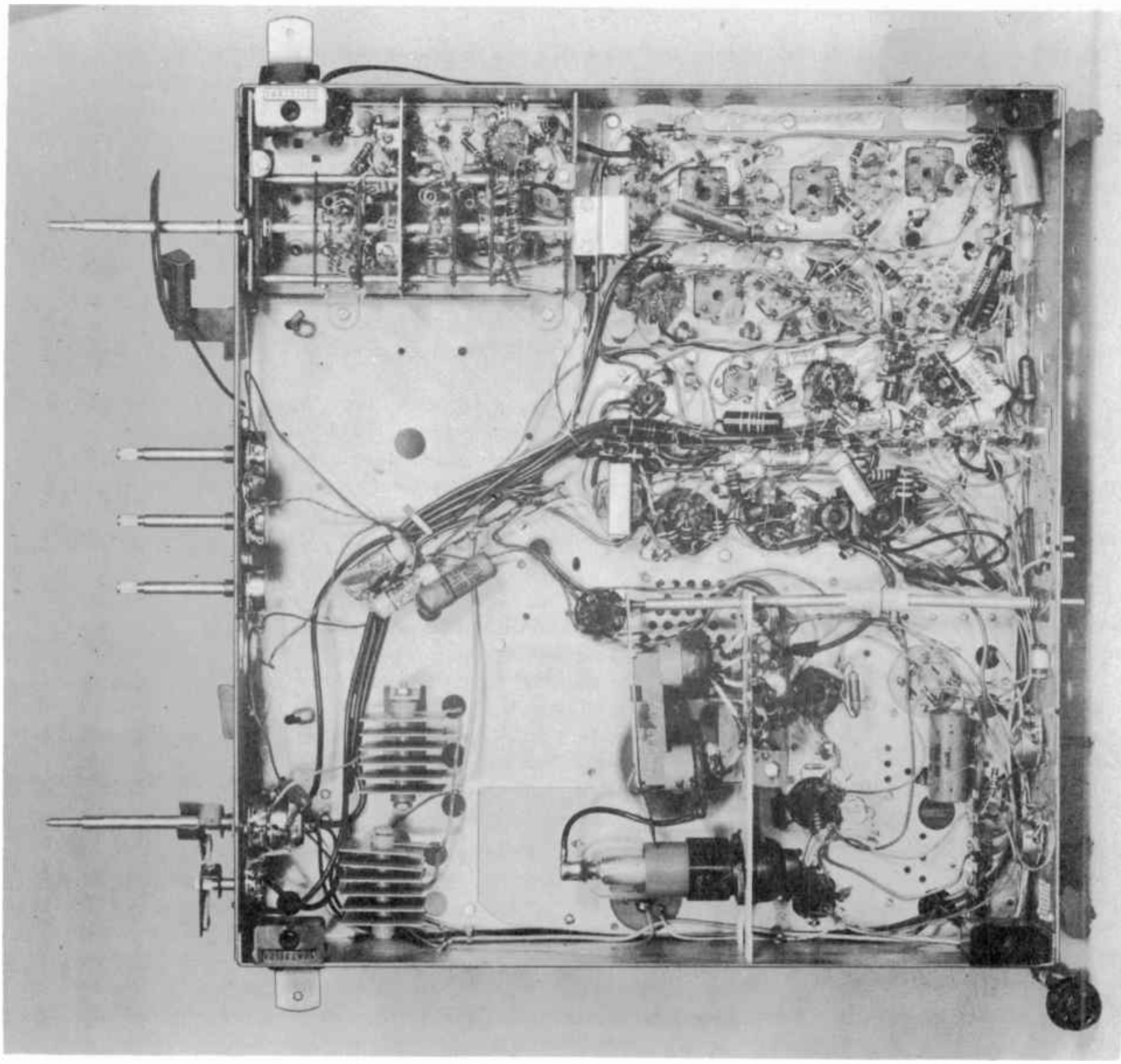


Fig. 1-2. Underneath the chassis are the smaller parts with which the service technician is most often concerned. (Courtesy Motorola, Inc.)

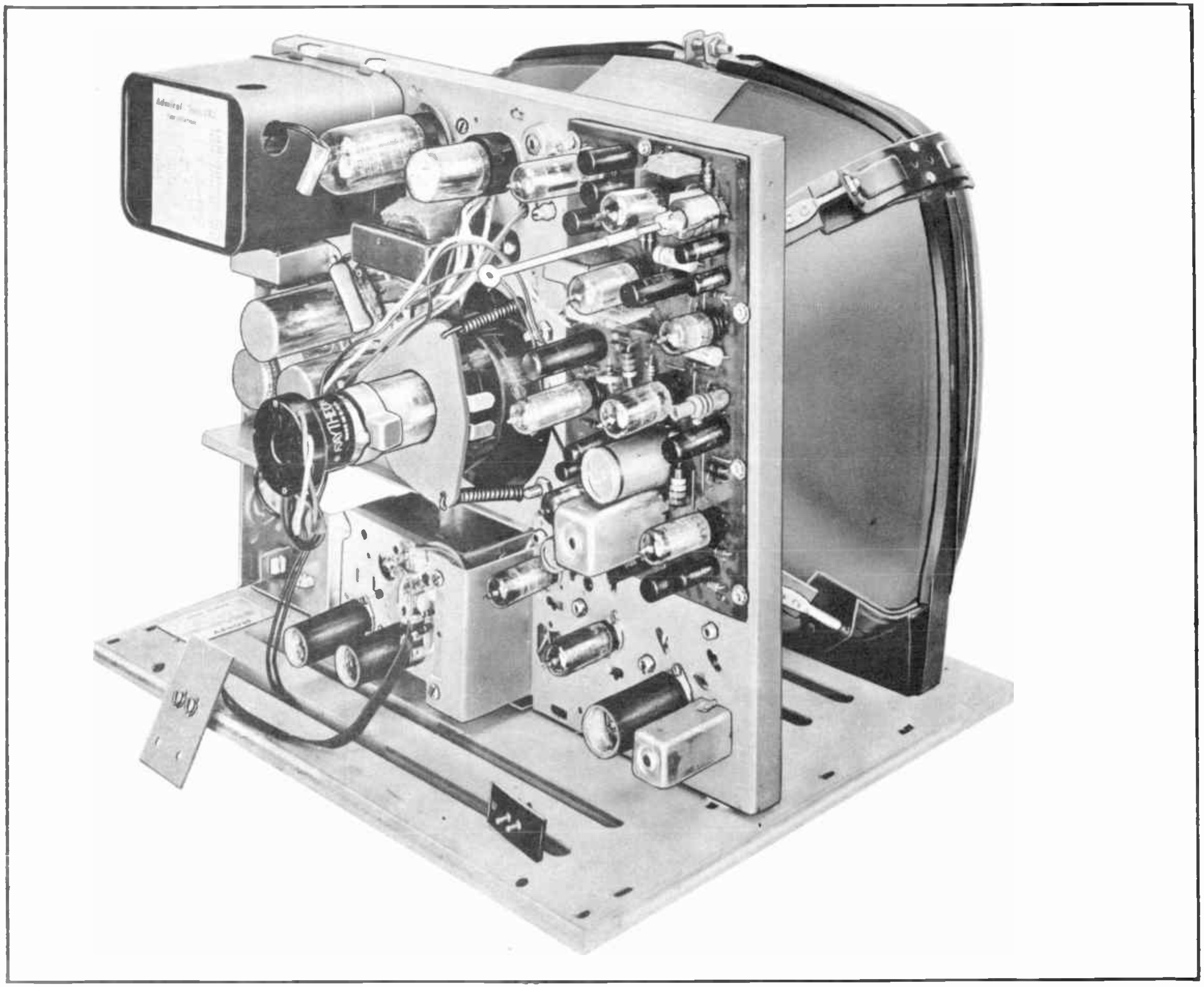


Fig. 1-3. This is a rear view of a large screen television receiver using a vertical type receiver chassis. (Courtesy Admiral Corp.)

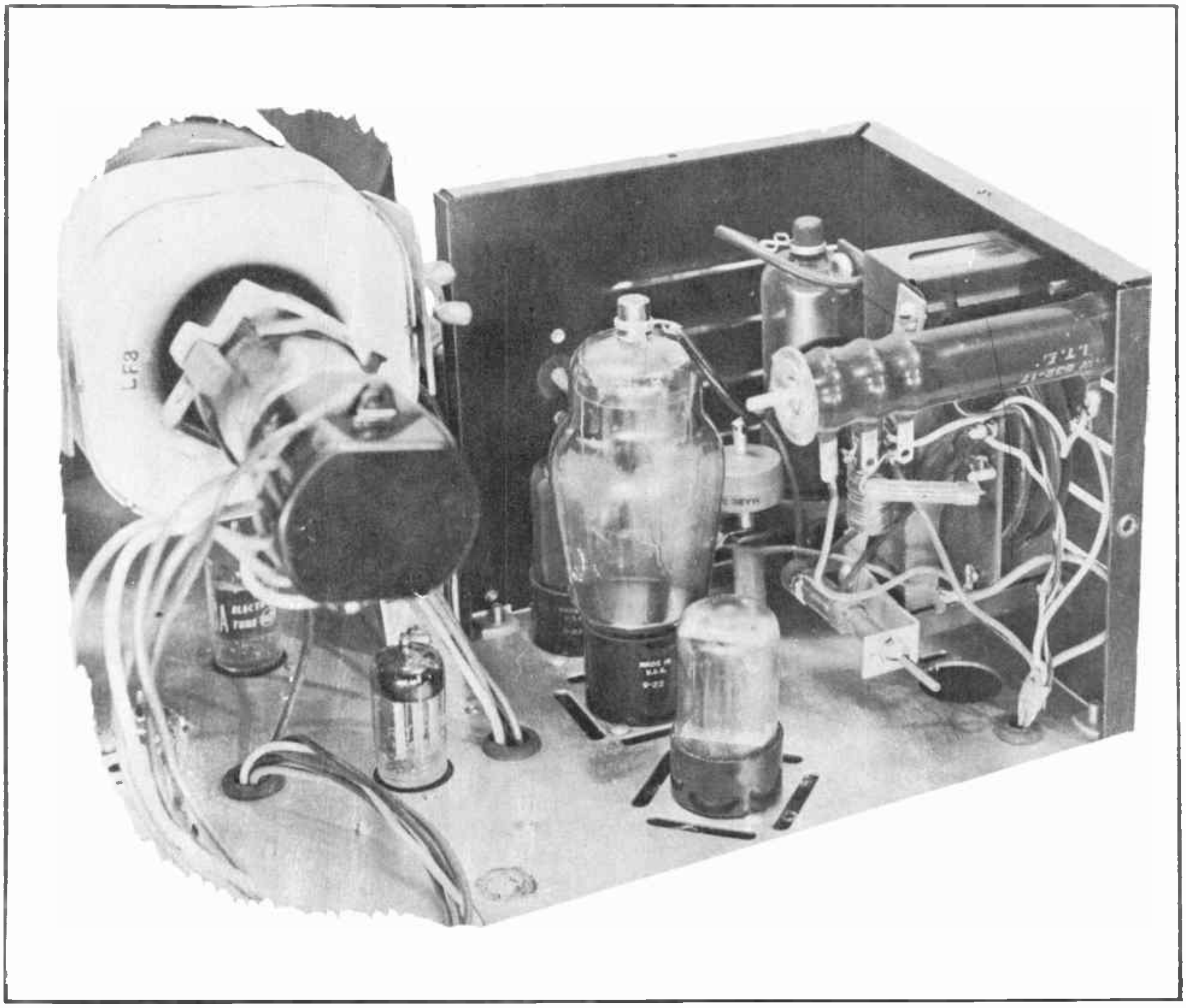


Fig. 1-4. One style of power supply which delivers a potential of 14,000 volts for the picture tube visible at the left.

ice methods. Then an understanding of principles will save the day, for you can figure out what must be wrong, and will know what must be done for correction – even though it is "not in the book".

There is just one preliminary warning. If you slight any part of any lesson you surely will get off the main road and have rough going. This may happen because you decide that some subjects are too easy, or that you already know all about them. No one knows everything about anything in television. You may know a lot, especially if you already are a radio man, but don't pass up the new information on new ways of doing things in television.

We are not going to learn to turn some certain adjustment to the left and another to the right by a turn and a half to accomplish some result. If the result shouldn't appear, as many times it won't, you must understand what really is happening in that particular circuit. We shall proceed on the assumption that if you know how a thing should work you can fix it when it doesn't work.

Now let's get down to business. Fig. 1-5 is a photograph of one of the television receivers with which we shall carry out many investigations. The biggest and most important looking part is the picture tube. Why do all the lights and shadows of an apparently living picture come and go on the face of that tube?

HOW PICTURES ARE FORMED

The answers to why pictures appear on the television tube will explain the basis of television itself. As a preliminary we shall consider the problem which confronted the scientists who conceived the idea of sending moving pictures by radio. They were up against these hard facts. A radio signal can be varied in strength only from one instant to another. Its strength cannot vary in two ways at the same time. During some particular millionth of a second the signal may be weak, in the following millionth of a second it may be strong, and at other instants may be of intermediate values, but the signal cannot be both strong and weak in the same instant. This is all right for transmission of music and speech, because sounds consist of changes of pitch and intensity which occur one after another. At any one instant there is only one kind of sound, and at that instant the radio signal need represent only that sound.

But in a moving picture there are different changes of light and shadow occuring at the same instant in many places on the image. Some spots become lighter while others are getting darker, and still others remain unchanged – all at the same time. How would you go about transmitting all these different but simultaneous changes with a radio signal which cannot possibly vary in more than one way at one time?

The only possible solution for this problem has been known for more than sixty years. The solution is to divide the picture into parts so small that each consists of only a single shade or single degree of brightness. These tiny parts of the picture must be viewed one after another. Signals corresponding to the brightness of each small area must be transmitted one after another in time. Then, at the receiving end, all the parts, each with its appropriate brightness, must be assembled in the same positions they occupied in the original scene. This will reproduce the picture being viewed at the transmission end of the system.

This first part of the solution gives us only one stationary picture. All the people and objects will be in fixed positions. To have movement in television pictures we utilize the same method employed in motion pictures. You know that a motion picture consists of a rapid succession of still pictures. Each still picture shows the positions of people and objects at one instant of time. The next still picture shows the positions during a following instant, and so on. These still pictures, in the movies, are projected onto the screen at the rate of 24 each second. When we look at one of these pictures our eyes retain the impression for about 1/15 second. Consequently, we see one picture until the next comes along. The slight changes of position of the people and other objects which occur between one still picture and the next one all blend together, and there is the illusion of movement.

The ability of our eyes to retain the impression of a picture for a fraction of a second, called persistence of vision, is used in television just as in motion pictures. Complete pictures are formed on the television screen at the rate of 30 per second. That is, all the small areas of light and shade which form one "still" picture are reproduced in their correct positions and degrees of brightness during 1/30 second. We continue to see that picture until another one is formed during the next 1/30 second. Then we see, or seem to see, all the smooth and continual movements that are occurring in the televised scene.

A

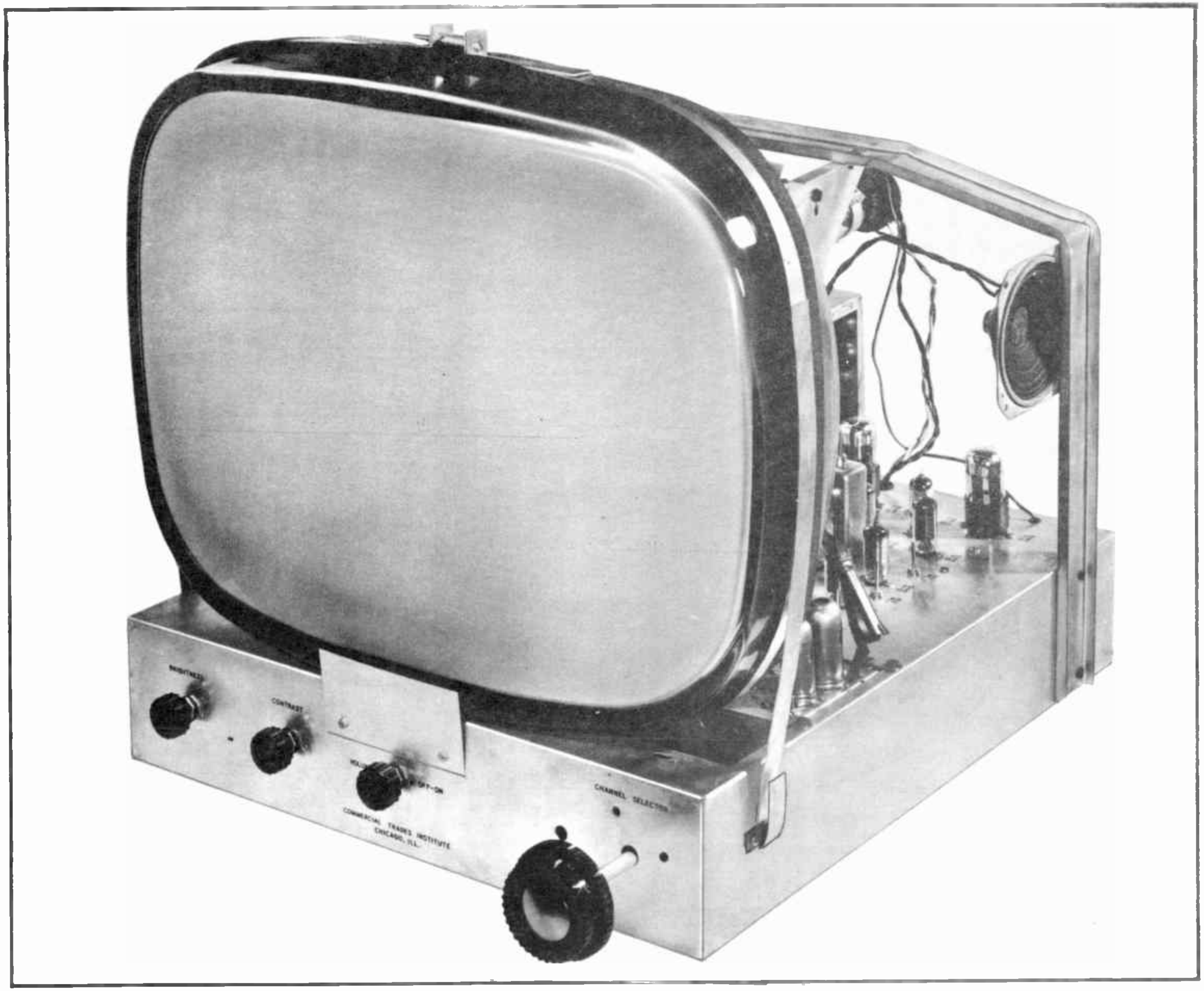


Fig. 1-5. The chassis and picture tube of a large screen television receiver.

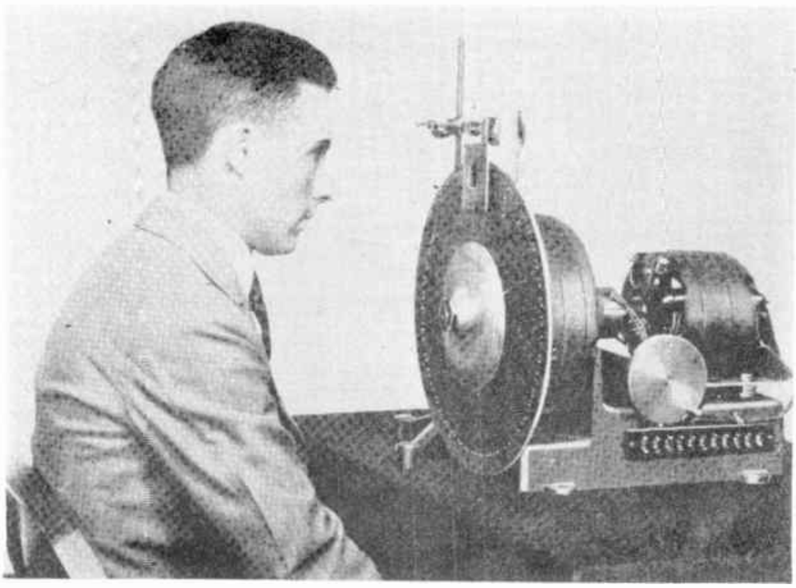


Fig. 1-6. Television – vintage of 1928. The observer is looking through holes in a spinning Nipkow disc which divides the picture into many small areas.

DIVIDING THE PICTURE

Now to apply these principles in practice. To begin with we shall split the scene or its image into many small parts. The first step is to divide the picture into a great many horizontal lines. Fig. 1-7 shows this division. The football scene has been ruled off into 75 horizontal lines. In actual television reproduction the picture would be divided into about 500 horizontal lines, each much narrower than shown here. We commence with wide lines because it will be easier to see what happens.

To examine the makeup of a single horizontal line we shall look at part of one which cuts through the numeral "30" on the jersey of one of the players. This portion of a line is seen more easily in Fig. 1-8. By following along this line from left to right we observe that it consists of a series of shades; some black, some white, and a great many intermediate grays. Any other line or part of a line, is similarly made up of lights and shadows.

Across the bottom of the picture is a horizontal scale whose divisions are each just as long from left to right as the vertical distance between lines in the picture. Were each picture line divided into sections of this horizontal length the areas would be squares. Each square would be predominantly black, white, or of some particular gray tone. At least, this would be the case were the lines of less height and the squares smaller than shown here, or were the lines of the size actually found in television pictures. This division of the picture brings us close to the final solution of our problems.

In the television camera there is a lens and optical system almost exactly like those in high quality cameras used by the ordinary photographer. An image of the televised scene is focused by the television camera

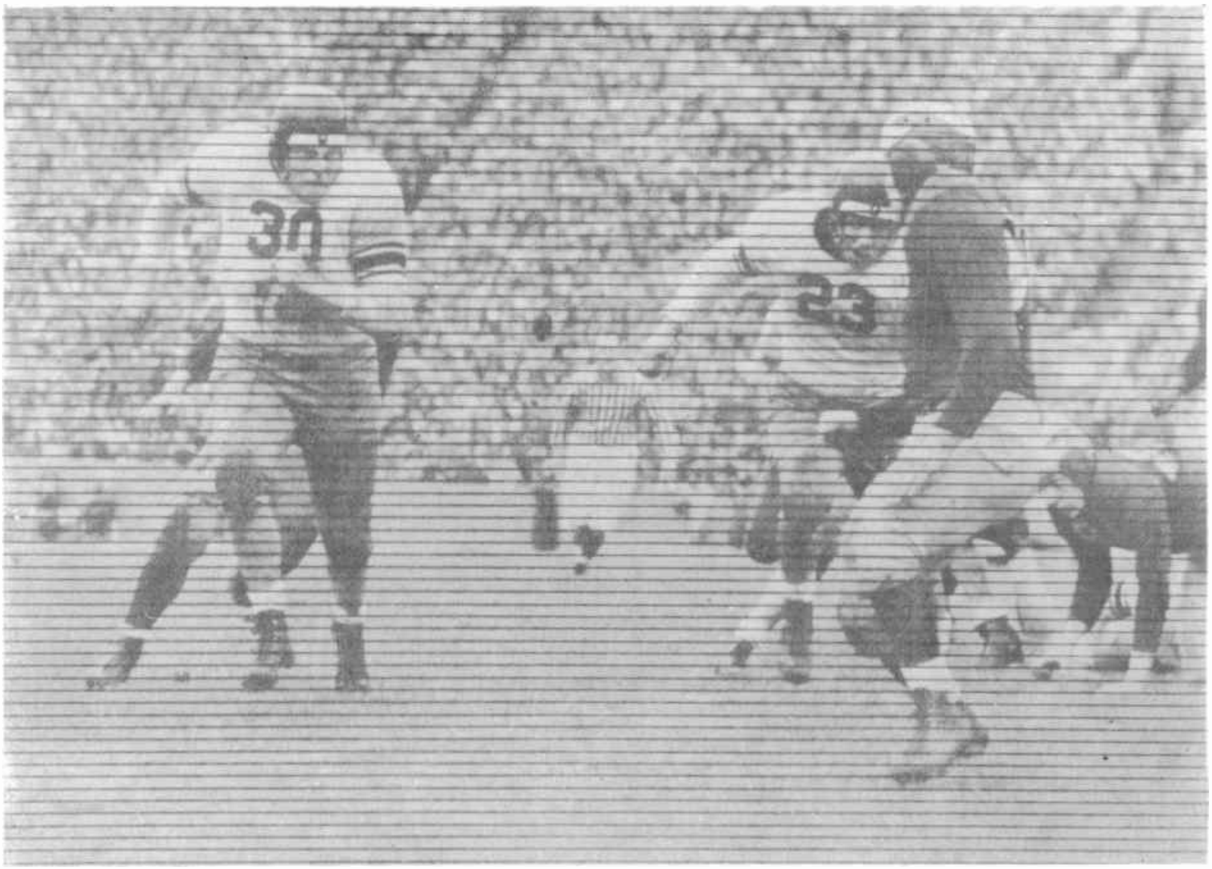


Fig. 1-7. Any picture or image may be divided into many narrow horizontal lines.

onto a light-sensitive surface, just as an image is focused onto the photographic film in any other camera. The light-sensitive surface in the television camera is not a photographic film. Rather it consists of countless tiny photoelectric cells which produce electrical impulses proportional to the brightness of light in the focused image.

The photocells in the camera tube deliver their electrical impulses only when they are affected first by a stream of electric particles (electrons) which is directed against the cells. This stream of electric particles is made to travel across the focused image along horizontal lines like those discussed in connection with Figs. 1-7 and 1-8. In effect, the television camera looks along each of the image lines from left to right, and follows all the lines from top to bottom one after another. Thus the camera sees only rapid changes of brightness. The television picture signal which originates at the sensitive surface in the camera then will vary according to these successive changes of brightness. The camera looks at only one small area of the image at any one instant, but it looks over the entire image during every 1/30 second of time.

Fig. 1-8. Every horizontal line, or part of a line, consists of areas in which there is predominantly a single shade.

The picture signal will arrive at the picture tube in your television receiver as fast as it originates in the camera. Now, undoubtedly, there are two big questions in your mind. First, how does this signal cause all the changes of light and shadow which make up the picture during each 1/30 second of time? Second, how are the lights and shadows distributed over the screen of the picture tube to appear in precisely the same positions they have in the original scene? To arrive at the answers to these questions it will be well to commence with an examination of the picture tube, and how it works.

THE PICTURE TUBE

When removed from its mountings a picture tube of one type appears as in Fig. 1-9. Underneath is one of the smallest tubes from the same receiver. The base of the picture tube is located at the left hand side in this illustration. Protruding from the base are a number of metal pins which allow making electrical connections to parts inside the tube. Attached to the base is a straight cylindrical glass section called the neck of the tube. The neck ends at the beginning of the flare, just above the little tube. The flare ends at the flat-

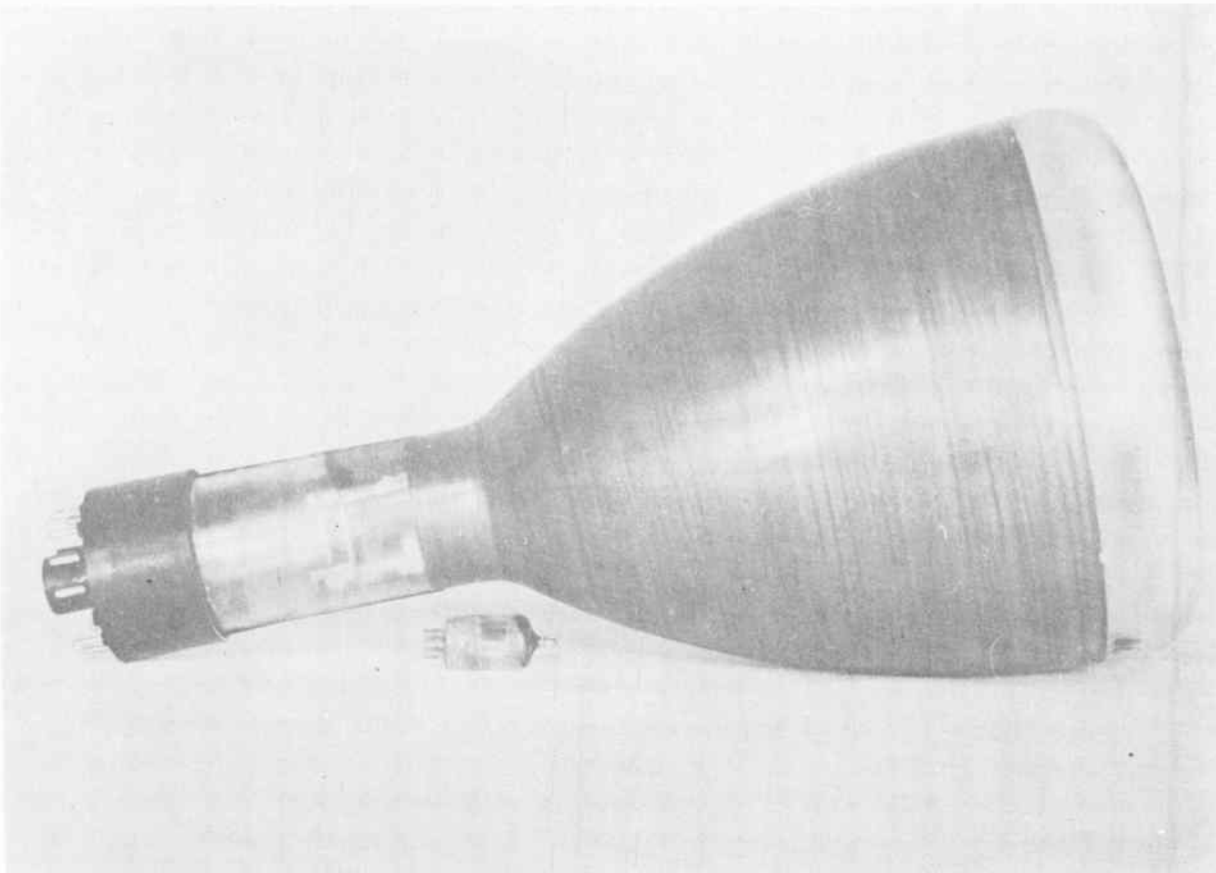


Fig. 1-9. A picture tube and its smallest relative, the video detector tube.

tened front face of the tube. On the inner surface of the face is the screen where pictures appear.

The entire glass portion of the picture tube is called the envelope. The envelope is highly evacuated and tightly sealed. That is, almost every last trace of air and other gasses have been pumped out of the envelope. This leaves practically no pressure at all on the inside of the glass, while on the outside surface there is the full pressure of atmospheric air, 14-7/10 pounds on every square inch of glass. The total excess air pressure on the outside of a 21-inch television tube is about six and one half tons. Needless to say, we handle picture tubes with great respect.

When we look into the neck of our picture tube, as in Fig. 1-10, we see several metal cylinders end to end, and numerous other small parts. Inside the cylinder which is nearest the base of the tube is a part called the cathode. The cathode is heated red-hot while the tube operates, and the heat causes particles of electricity to boil out of the cathode surface. This action is much like that in which particles of water in the form of vapor or steam come from the surface of any wet object which is heated. The particles of electricity are

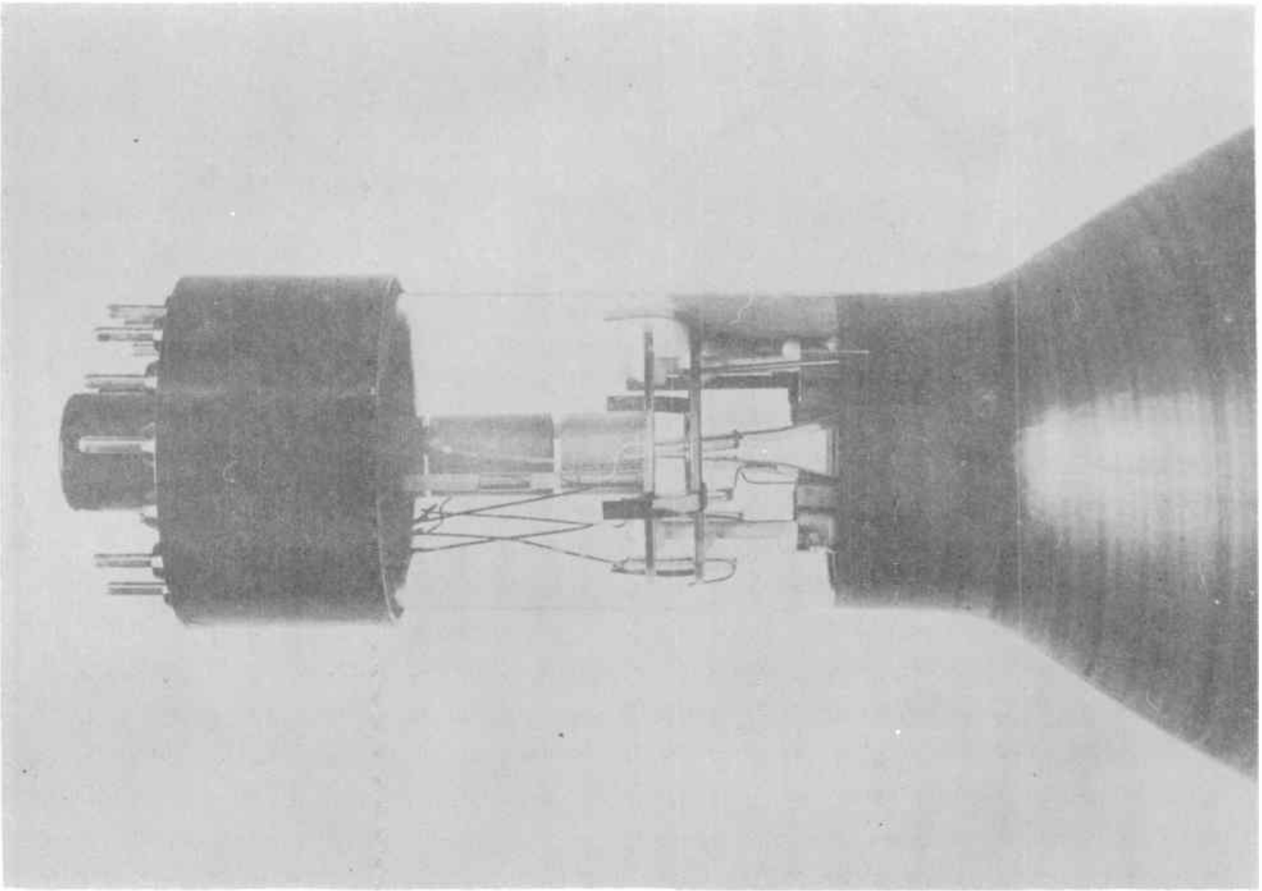


Fig. 1-10. Inside the neck of the picture tube are the parts forming the electron gun.

electrons. We shall have a lot to do with the behavior of electrons, for they are all-important in television and radio. For the present it is enough to know that electrons are electricity itself, and that they are forced out of the cathode by heat.

As soon as an electron gets out of the cathode its natural inclination is to fall right back in again. But many electrons, before they can drop back into the cathode, are caught by a very strong force. This force results from electrical potential of the metal cylinders which are farther from the base of the tube. Potential commonly is called voltage, because its strength is measured in the unit called a volt. Potential in house lighting circuits is usually between 110 and 120 volts. In the picture tube being examined there are potentials as high as 5,000 volts. In other common types there are potentials of 9,000 to 16,000 volts, and in a few picture tubes this electrical force exceeds 25,000 volts.

Many electrons which have boiled out of the cathode are pulled away by the strong potentials in the picture tube. Not only are the electrons drawn away, they are accelerated to speeds which are beyond the imagination. By the time an electron gets through the parts in the neck of a picture tube it is traveling at more than 50 million miles an hour. The part of the tube in which electrons are liberated and accelerated to such tremendous velocities is, appropriately, called the electron gun.

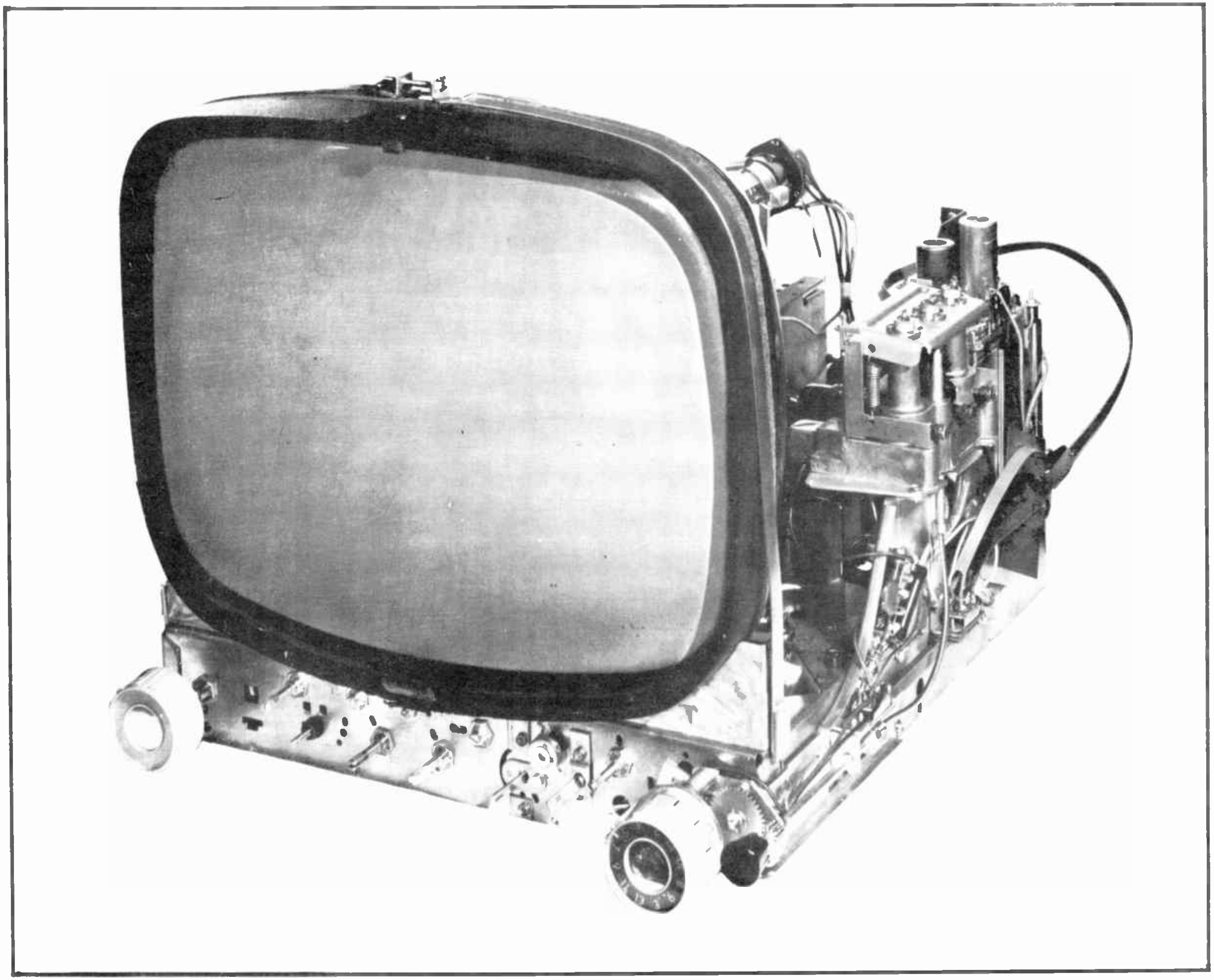


Fig. 1-11. Inside the face of the picture tube is the screen with the phosphor material which emits light when bombarded by flying electrons. (Courtesy Zenith Radio Corp.)

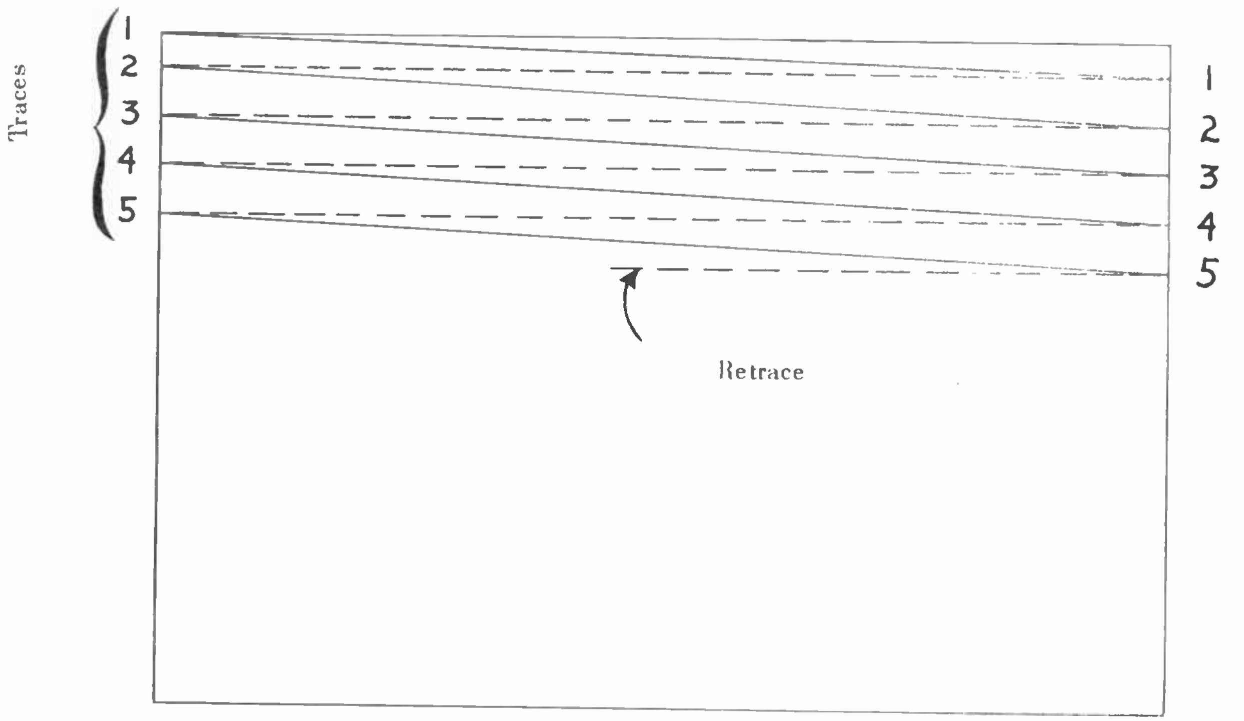


Fig. 1-12. Successive lines of lights and shadows are formed as the electron beam moves gradually downward on the screen.

Billions of electrons will be shot straight through the center of the picture tube to hit the center of the screen. The screen, which is facing us in Fig. 1-11, consists chiefly of a coating called the phosphor on the inside of the glass face of the tube. The phosphor material is similar to that used inside the long glass tubes of flourescent lamps which are commonly used in homes and places of business. It is a substance which glows brightly when struck by electrons.

The stream of high velocity electrons hitting the center of the screen causes a small bright spot to appear there. During their travel through the tube these electrons have been drawn together or focused into a beam which produces a light spot only about 1/100 inch in diameter. The light spot may be made of any desired brightness by varying the quantity of electrons coming to the screen. The greater the number of electrons hitting the screen per second the brighter is the spot, Reducing the rate of arrival makes the spot less bright. Shutting off the electron beam leaves the screen dark. Thus it is possible to control a small spot to produce any degree of brightness, or even darkness.

Now we have learned how brightness may be varied on the screen of the picture tube. To form a horizontal line it is necessary only to vary the brightness while the spot being formed by the electron beam is made to travel from left to right. Any line may be built up with changes of brightness or of light and shadow.

SWEEPING THE BEAM

To form one complete picture the electron beam is first directed toward the upper left-hand corner of the screen. Then the beam is swept across the width of the picture space all the way to the right. During this

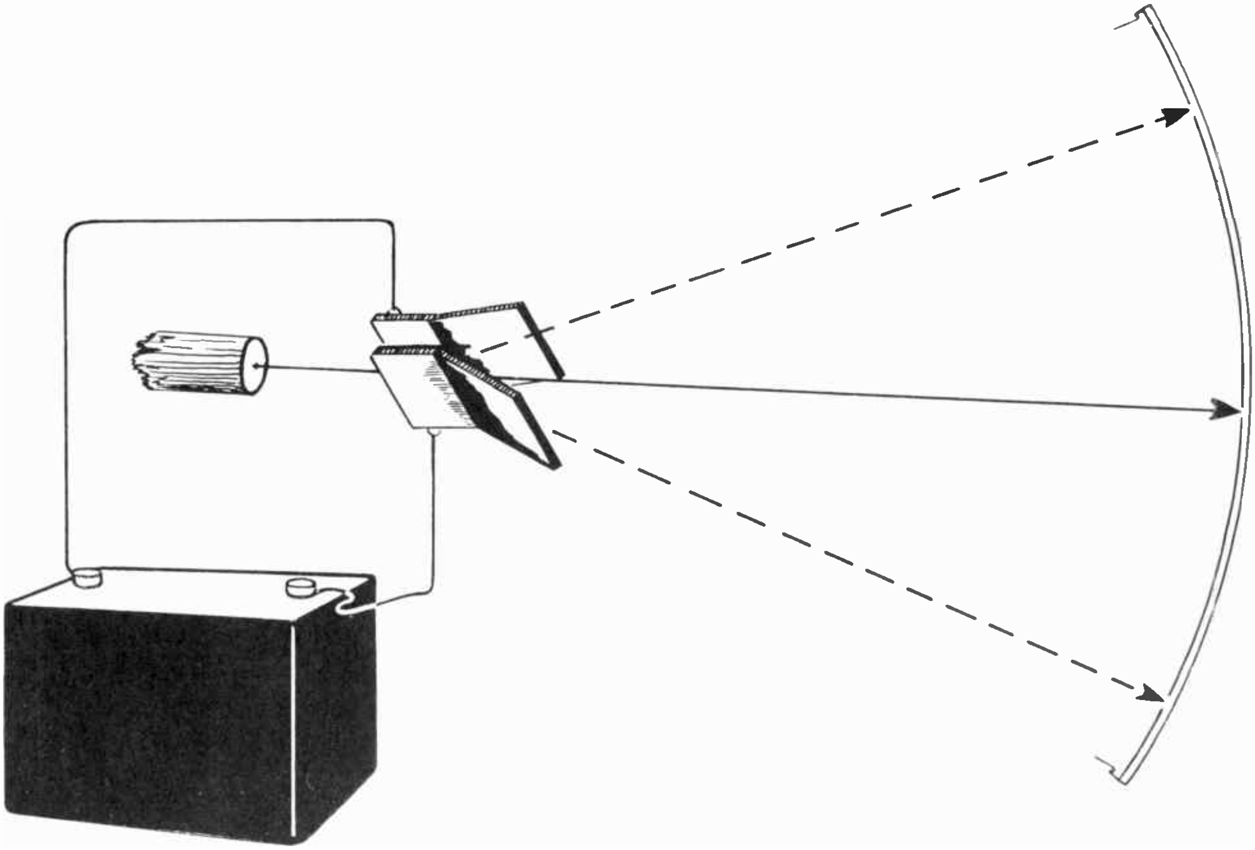


Fig. 1-13. The electron beam is bent or deflected as it passes through a space between two metal plates.

travel the rate of flow of electrons in the beam is varied according to the picture signal coming to the tube. This variation produces such changes in brightness of the spot as to place at each point along this one line the same degree of light or shadow that exists at the corresponding point on the televised image. The horizontal line consisting of lights and shadows may be called a trace.

To form a second line or second trace the beam must again start from the left-hand side of the picture space, as may be seen from Fig. 1-12. But we don't return the beam itself from right to left along one of the broken lines in that figure. Such a return would cause a spot of light to move backward and form a faint white line between each two adjacent traces composed of picture lights and shadows. During the time it would take to bring the beam back to the starting point of another line the electron flow is shut off. We say that the beam is blanked during the horizontal retrace period. Then, when the forces which control the sweep of the beam are all ready for the next line, the beam again is turned on and the light spot is formed.

It is apparent that, between each pair of horizontal lines, the beam must move downward by an amount equal to the distance between luminous traces. When this is done, all the lines from top to bottom of the picture will be reproduced.

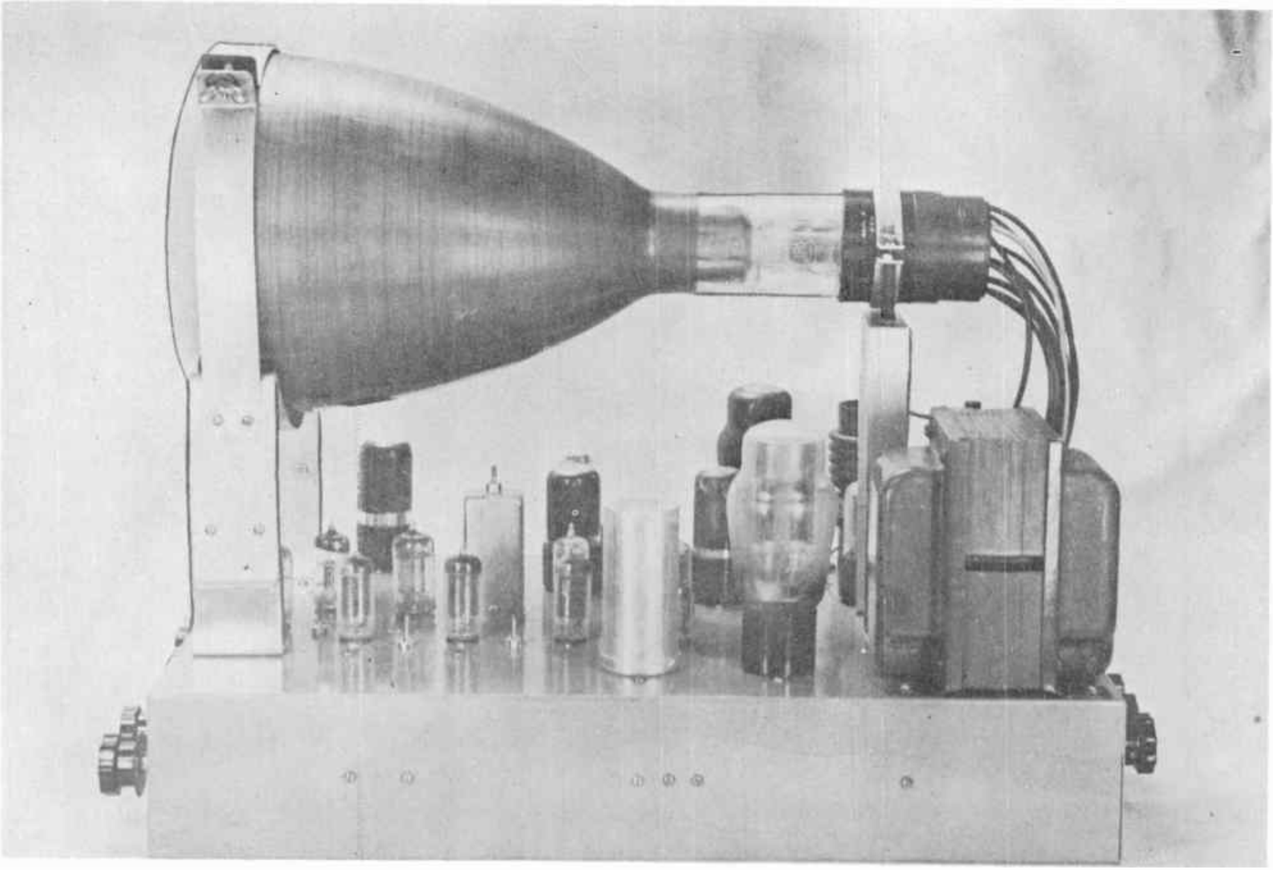


Fig. 1-14. The plates for electrostatic deflection are within the neck of the picture tube.

After one picture has been completed, and the beam is at the lower right-hand corner of the picture space, the beam again is blanked. Then the forces which control vertical travel of the beam get all set to start a new picture in the upper left-hand corner, and there the beam is turned on. The time between shutting off the beam at the end of one picture and turning it on again at the beginning of the next picture is called a vertical retrace or a vertical blanking period. The action of forming one picture after another continues as long as the program lasts or until something happens which requires the help of a service technician.

By the way, don't try too hard to remember all the television words and terms which we are using in these first few pages. As we proceed, you will meet these words so often that you can't possibly forget them nor what they mean.

The light spot travels so fast that we do not see it as a spot, only as a line of light. Unless we are too close to the picture tube we don't even see the separate lines. And no matter how close we get it never becomes possible to see the separate pictures completed in only 1/30 second each. Persistence of vision lets us see only the complete scene with all its motion.

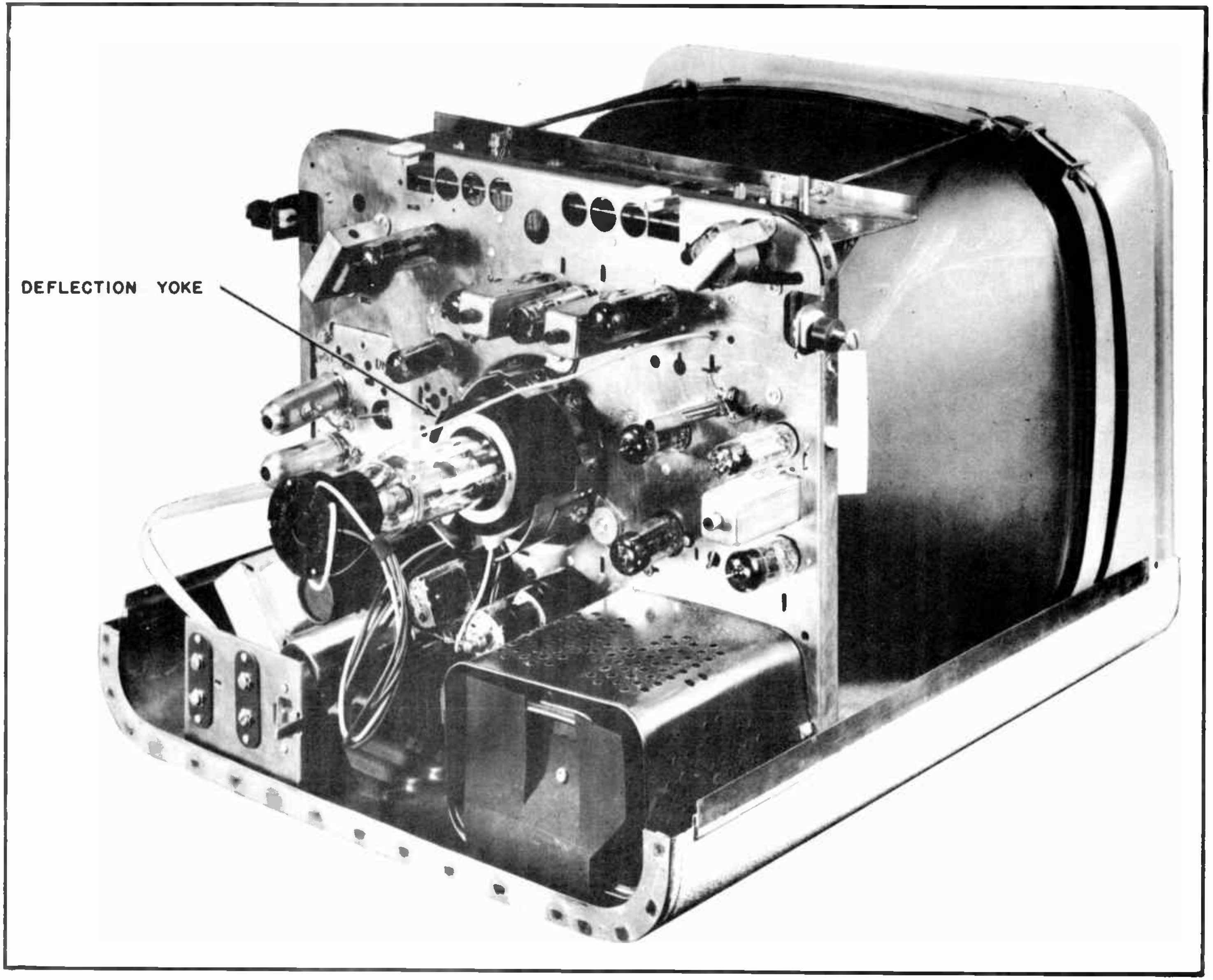


Fig. 1-15. The yoke, containing magnet coils for magnetic deflection, is around the outside of the picture tube neck just back of the flare. (Courtesy Raytheon Mfg. Co.)

The beam and the spot travel all the way from left to right in about 53 millionths of a second. The period for horizontal blanking lasts about 10 millionths of a second. In forming one complete picture the electron beam travels from top to bottom of the picture space in about 15 thousandths of a second, and vertical blanking lasts for about 1½ thousandths of a second.

No doubt you noticed that this last statement, about the beam traveling downward in 15 thousandths of a second, hardly fits in with the formation of a complete picture in 1/30 second, for this time for a complete picture would be about 33 thousandths, not 15 thousandths of a second. The apparent discrepancy comes about in this way. Only every alternate horizontal line is formed during one downward travel of the electron beam. Then, on the next trip downward, the intervening lines are filled in. This is called interlaced scanning, a matter for discussion later on. We cannot cover all the little details in this preliminary talk about television - it isn't quite so simple.

DEFLECTION METHODS

There are two methods of deflecting the electron beam in a picture tube. One method is to pass the streams of electrons between two metallic plates arranged as in Fig. 1-13. These plates are inside the neck of the picture tube, just beyond the electron gun. Were you to connect the two terminals of an electric battery to the two deflecting plates the electron beam would be bent or deflected to one side in passing between the plates. Reversing the connections between battery and deflecting plates would bend the beam to the other side. The stronger the voltage from the battery the greater would be the degree of bending or the greater the deflection.

Naturally, we don't deflect the beam in a picture tube by means of battery connections, partly because we need 15,750 horizontal deflections every second. This job requires various synchronizing tubes, sweep oscillators, sweep amplifiers, and other parts, all of which will be examined in due time. By varying the deflection voltage at a suitable rate it is possible to make the beam travel across the screen at just the right speed, and by suddenly reversing the voltage we accomplish the retrace between lines.

This method of sweeping the electron beam and light spot is called electro-static deflection. The picture tube on the receiver of Fig. 1-14 employs electrostatic deflection. The deflecting plates and their connections are inside the neck of the tube.

The other method of deflecting the beam is with magnets. In the space near a magnet is a force which we recognize as attraction. It is this force that pulls a nail toward a magnet even while there is still an air space between the two. This magnetic force will exert a pull on electrons in the picture tube beam, and will bend or deflect the beam. Magnets used for deflection are electromagnets. They are made with pieces of iron around which are coils of insulated wire in which electricity flows. You can see small electromagnets in every door chime, bell, and buzzer.

The deflecting magnet is placed around the outside of the neck of the picture tube, with the two poles of the magnet on opposite sides of, or above and below, the neck. Then the electron beam inside the tube travels between the magnet poles. By varying the rate of flow of electricity in the magnet coils, and by reversing the direction of flow, the electron beam and light spot are made to travel from left to right across the screen while forming a trace, and to start over again for the following trace.

This method is called magnetic deflection or electromagnetic deflection. On the neck of the picture tube of the receiver in Fig. 1-15 is a set of deflecting coils. These coils are built into a deflecting "yoke" which is supported around the tube neck just back of the flare.

A single pair of electrostatic deflecting plates may be used for sweeping the beam either horizontally or vertically. A second pair is required for the other direction of sweep. Similarly, a single deflecting magnet will sweep the beam either horizontally or vertically, and a second magnet is required for sweep in the other direction.

HOW FAR WE HAVE TRAVELED

Near the beginning of this lesson the language of television was mentioned. We have been talking it ever since. How many television words and terms do you suppose have been used up to this point? Count them in the list which follows. Look through the list and check the words which you could explain, at least in a very elementary way, to someone who knows nothing about this subject. If you check fifteen words you have done exceedingly well.

We have talked about what might be called the beginning and the end of television, about the camera and the picture tube. It has been taken for granted that a television signal picked up by the antenna on your receiver has gotten through to the picture tube in some manner still not explained. It is between the antenna and the picture tube that we find a majority of the parts with which we are most concerned in television servicing. In the following lesson we shall look at many of these parts and discuss their purposes.

Here is the list of technical words and terms which have already been used.

Antenna

Blanking Bar

Capacitors
Cathode
Chassis

Deflecting magnet
Deflecting plates
Deflection
Discriminator

Electromagnets
Electron flow
Electron gun
Electronic
Electrons
Electrostatic deflection

Flare

Ghosts

Horizontal afc
Horizontal blanking

Inductors
Interlaced scanning

Magnetic deflection
Mixer tube

Neck

Phase
Phosphor
Picture tube
Potential

Realignment
Resistors
Retrace

Screen
Snow
Sweep
Sweep oscillators
Sync
Synchronizing tubes

Trace

Vertical blanking
Video detector
Voltage
Voltohmmeter

DO NOT TEAR — CUT ALONG THIS LINE AND SEND IN FOR GRAD

TELEVISION

LESSON NO. 2 FOLLOWING THE TELEVISION SIGNAL

"Training Your Mind"

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

COMMERCIAL TRADES INSTITUTE

LESSON NO. 2

FOLLOWING THE TELEVISION SIGNAL

We have learned how a moving picture is formed on the screen of the picture tube in a television receiver. Now we shall learn what happens between the receiving antenna and the tube in order that a picture may appear. This means following the signal through the receiver. First we must know a little bit about signals in general. Then, to understand the peculiarities of the television signal, we must determine what it is supposed to accomplish by the time it gets to the picture tube.

A television signal is only one of a great variety of radio signals which are being transmitted and received everywhere at all times. Radio signals are employed for airplane control, for guidance of ships, for radar detection of objects beyond our sight, for experimentation and development by the radio amateurs, for communication with moving trains, trucks, and automobiles, and for other purposes almost too numerous to mention. We are most familiar with signals used for entertainment and education by means of sound alone in standard broadcast radio, in frequency-modulation or f-m radio, in international short-wave radio, and now in television where sight is added to sound.

Any and every radio signal passing through space consists of forces called electromagnetic radiation. All we need to know about this radiation, just now, is that it produces exceedingly weak voltages in receiving antennas. These weak signal voltages must be strengthened in the receiver. About half the tubes in a television receiver are amplifiers, whose purpose is to step up the voltage of the signals.

An amplifier works something like a faucet or valve in a water system. You may apply a small force to the handle of a faucet or valve and thereby control or regulate a flow of water capable of exerting much greater force. This likeness is so real that the English don't use the word tube, they call them valves. The faucet doesn't produce the force of the water, it merely controls the force. An amplifier tube doesn't produce voltage, but it does control the flow of electrons from other sources of voltage.

There are other tubes which take power from electric lighting circuits in buildings and deliver this power in the kinds of electron flows and voltages needed for reception. It is these voltages which are controlled by the amplifiers. The voltage controlled by one amplifier may be used by a second amplifier to control a still stronger voltage. This process may be repeated a number of times, until the final controlled voltage is ample to operate the picture tube. It all begins with the very weak voltage induced in the antenna by the received signal.

The strong signal voltages delivered to the picture tube will be the final result of weak signals coming to the antenna. Consequently, to determine what is needed in the received signal we may make a list of things needed for control of the picture tube. You are well enough acquainted with the workings of a picture tube to write this list for yourself. Here are the requirements..

1. The rate of electron flow in the beam must be varied in order to form lights and shadows for pictures.

2. The start of each horizontal sweep of the picture tube beam must occur at the exact instant of the start of a horizontal sweep of the beam in the television camera. Otherwise the light and dark spots on the screen

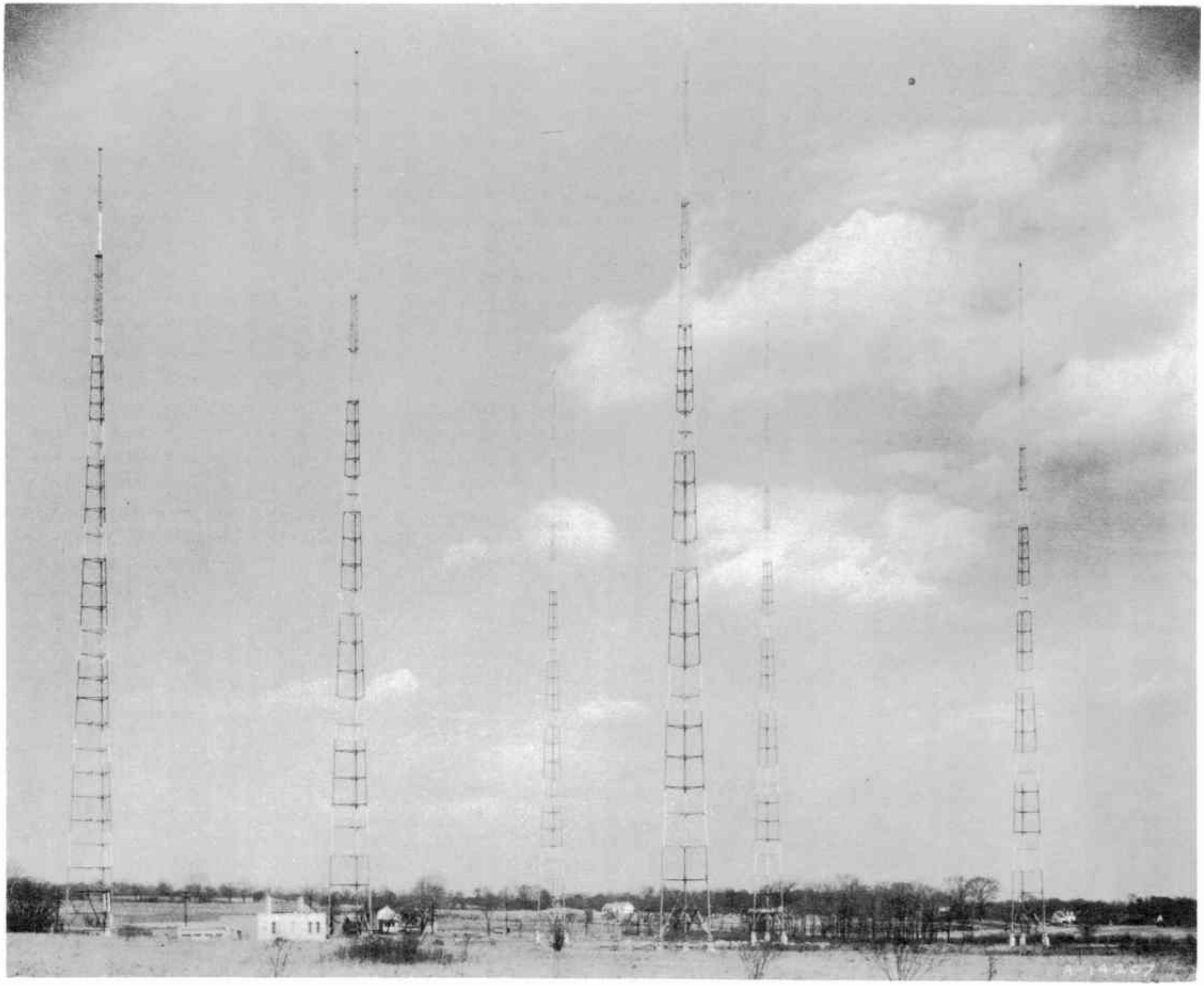


Fig. 2-1. Transmitting antennas of station WFMJ at Youngstown, Ohio. One of these Truscon self-supporting towers is 346 feet high. The other five are each 400 feet high.

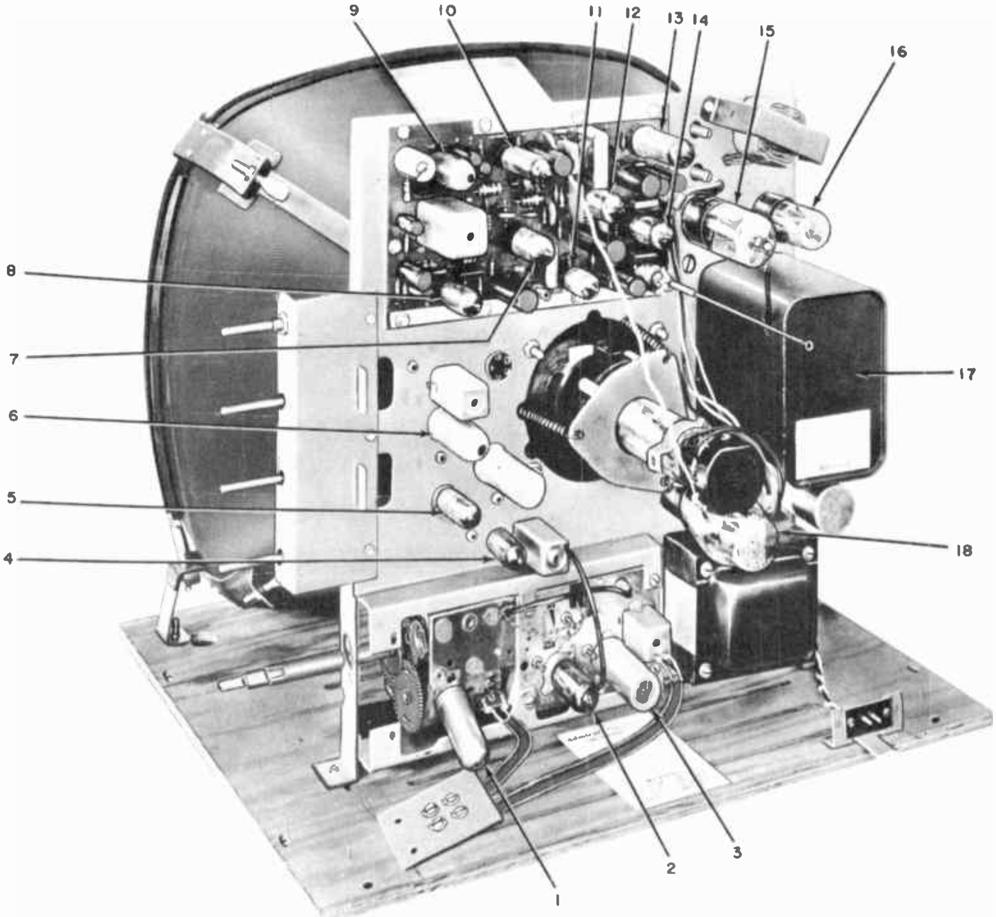


Fig. 2-2. Each tube in the television chassis illustrated performs a function which is essential in the correct operation of a television receiver. Tubes 1, 2, and 3 are in the tuner which selects and amplifies the desired sound and picture signals. Tubes 4, 5 and 6 further amplify the picture and sound signals. Tube 7 amplifies only the picture signal, and tubes 8, 9 and 10 amplify only the sound signals. Tubes 11, 12, 13, 14, 15 and 16 develop and amplify the voltages necessary to synchronize and sweep the beam in the picture tube. Tube 17, which is under the shield can, provides the high voltage necessary to operate the picture tube, while tube 18 provides the necessary low voltages.

of the picture tube would not be where similar light and dark spots exist in the televised scene or on its image in the camera tube. The picture would be thoroughly scrambled.

3. The start of each downward travel of the electron beam in the picture tube must be precisely in time with the start of a downward travel of the electron beam in the camera tube. If this were not done you might see a man's head below his feet, or his body might be sliced in two anywhere in its height.

4. There must be means for blanking the electron beam in the picture tube between horizontal traces, and for blanking the beam during the time between the end of one downward sweep and the beginning of the next one. Were there no blanking of the beam during retraces, the whole picture would be too light in shading, and it would be decorated with sloping white lines which don't belong there.

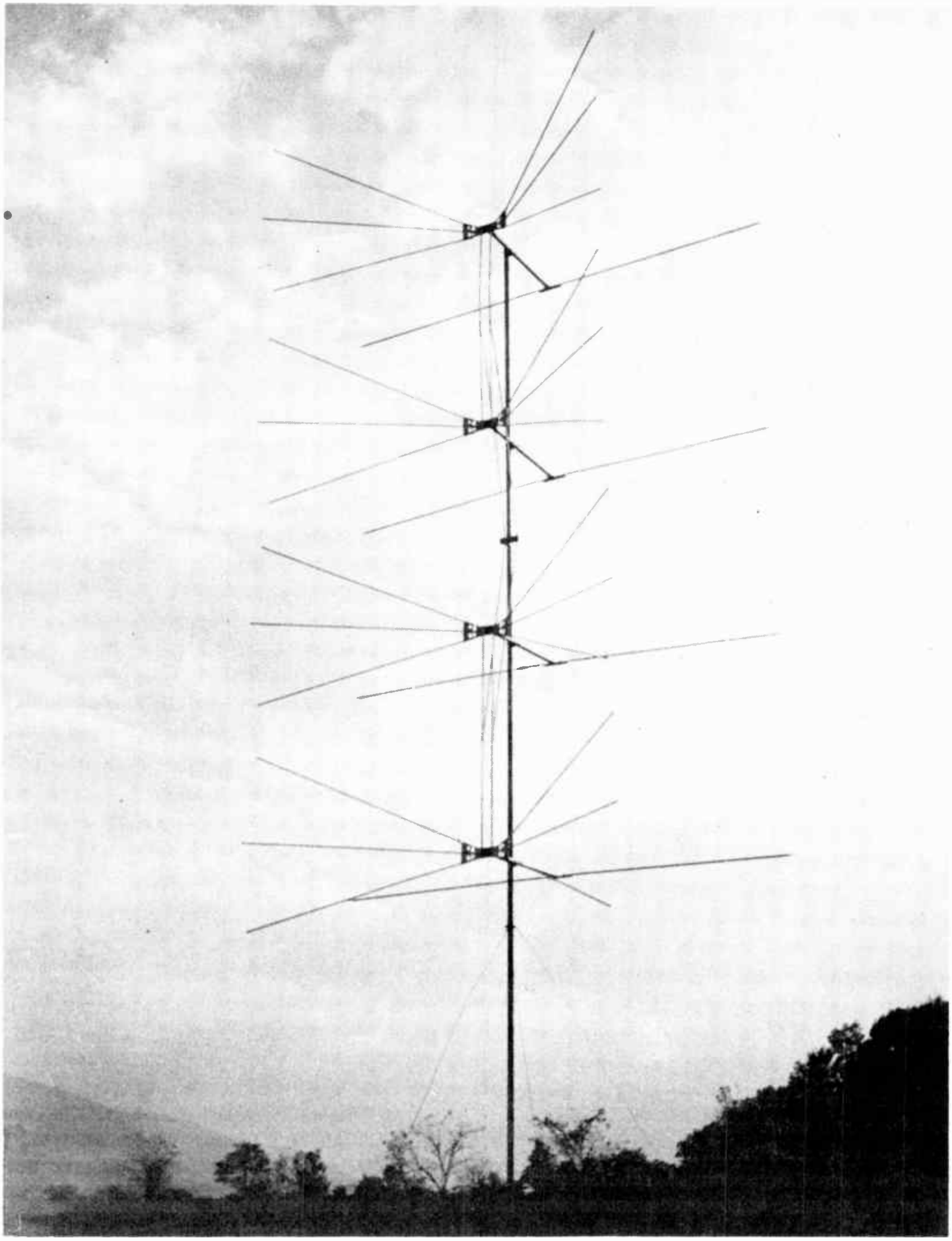


Fig. 2-3. This is a Channel Master four-bay multiple-stacked antenna array for long-distance television reception.

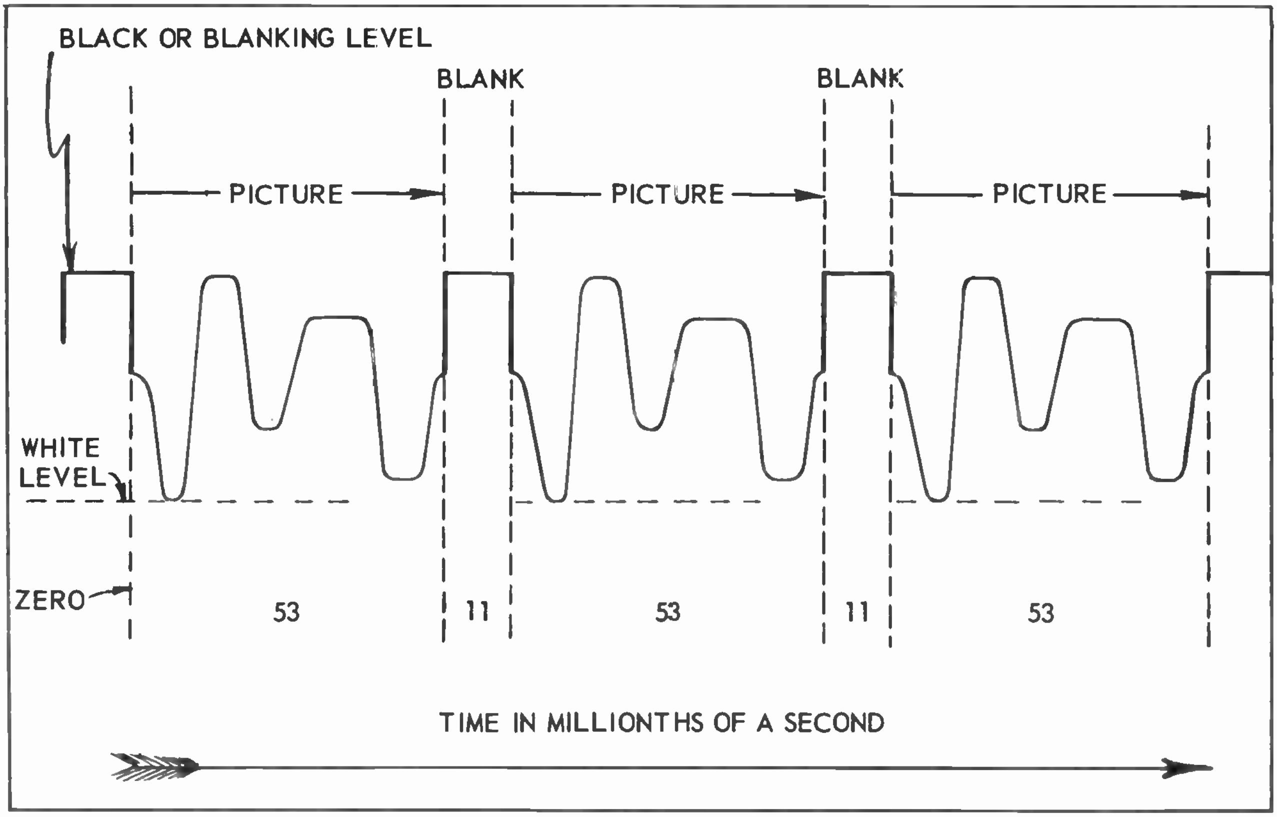


Fig. 2-6. The portions of a composite television signal representing three horizontal traces and four horizontal blanking intervals.

brightness. This picture portion of the signal may have to vary at a rate of nearly four million times a second in order to have good reproduction of small details.

2. *Horizontal sync.* In the television camera tube there is an electron beam which sweeps across the light-sensitive surface to activate the tiny photo-cells. In the picture tube we have an electron beam which sweeps across the screen to activate the phosphor and produce light. These two beams must remain precisely in time or in step with each other, always.

Any two actions which are forced to occur at the same instant and in the same manner are said to be synchronized. This is a long word, so we abbreviate it to "sync", and pronounce the abbreviation just like the name of the sink you have in the kitchen. Sync, in the language of television, means synchronized, synchronizing, synchronous, or any other word commencing with the letters s-y-n-c.

Now we may give a name to the voltages of the composite signal which time the horizontal sweep of the beam. These voltages are the horizontal sync pulses. They arrive right along with the picture signals, in between each two successive lines. Each horizontal sync pulse lasts for about 5 millionths of a second. That is enough to start the beam on its way.

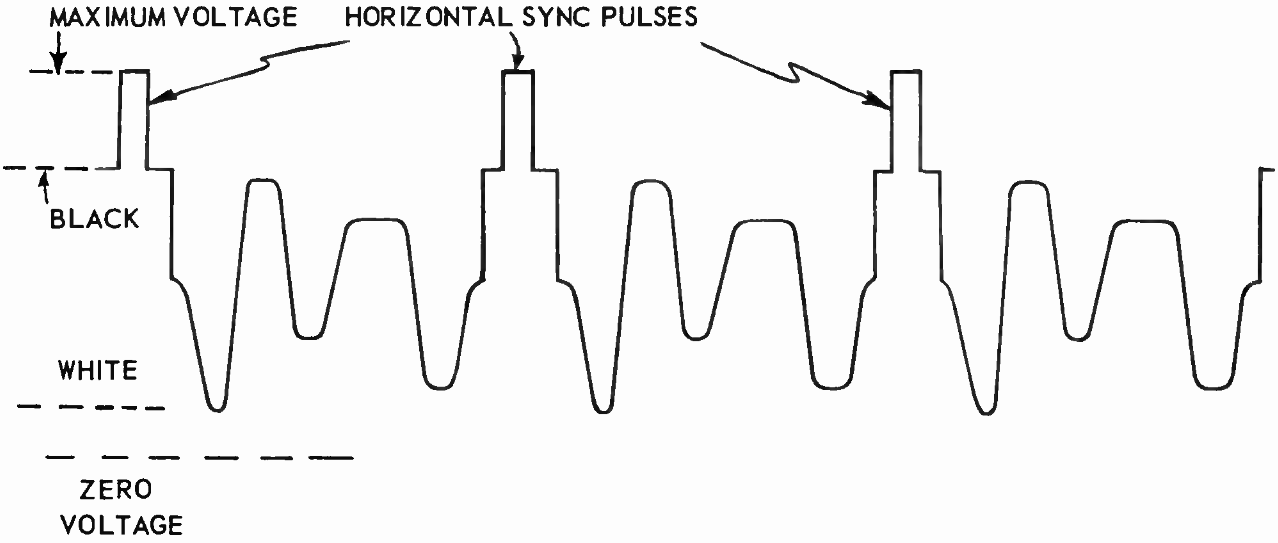


Fig. 2-7. The composite signal with horizontal sync pulses added.

3. *Vertical sync.* The start of each downward travel of the picture tube beam must be precisely in time with the start of the corresponding downward travel of the beam in the camera tube. What would you call the composite signal voltages that control this starting? Naturally, you would call them the vertical sync pulses. The entire pulse of signal voltage which starts the beam on its downward travel lasts for about 185 millionths of a second.

Were you a service technician to whom the sync section of a television receiver is being explained we would use a diagram such as that of Fig. 2-5. This is just a preview of the manner in which tubes, connections, and other parts are shown in the written language of television. It is a schematic circuit diagram including the parts which take the sync pulses from the composite signal and deliver corresponding voltages to other parts which sweep the beam horizontally and vertically. Everything is shown by symbols. The circles represent tubes, with small lines inside to show the working elements. Zig-zag lines represent resistors. Pairs of short lines separated by an open space represent capacitors. A series of small horizontal lines growing shorter and shorter from top to bottom means a connection to the main body of metal in the receiver chassis; these are ground connections. The rest of the lines represent wires between the parts.

The sync separator tube takes both horizontal and vertical sync pulses and some of the picture signal from the last of a string of amplifiers which have strengthened the entire composite signal. Then comes the sync amplifier, which strengthens the signal delivered by the separator. Next, the sync stripper gets rid of any portions of the picture signals which may have gotten through to this point. Finally, the sync limiter trims all the pulses to uniform strength and sends them to the filters. The filters separate the vertical and horizontal voltages which correspond to the original pulses.

4. *Blanking the beam.* A few paragraphs back we learned that signal voltage applied to the control grid

of the picture tube will make the light spot brighter or dimmer. To shut off the beam entirely it is necessary only that the voltage applied to the control grid exceed a certain strength. Then the electrons cannot get through the grid at all, they cannot get away from around the cathode. Thus the beam is blanked. The period of blanking between each pair of horizontal traces lasts a little more than 11 millionths of a second. The blanking period between the end of one downward sweep and the beginning of the next one is about 1/800 second.

In talking about the voltage of the composite television signal we don't use the words voltage or potential. We speak of various "levels". The value of voltage which is just sufficient to blank the beam is the blanking level. It may also be called the black level, because this and any greater voltage will keep the beam shut off and leave the screen black. There is also a white level. When signal voltage reaches the white level we have the brightest possible spot of light on the screen.

5. *Interlacing.* The subject of interlaced scanning is something with which we need not concern ourselves yet. The portions of the composite signal which permit this action are called equalizing pulses. They occur during 185 millionths of a second preceding each vertical sync pulse, and again during an equal time period following each vertical pulse.

In our discussion of the effect of signal voltages on the picture tube beam you may have wondered why a strong signal voltage blanks the beam, and a weak voltage causes the spot to become very bright. Well, this is the action we get with what is called negative transmission. Negative transmission is used in the United States and many other countries. Others employ positive transmission, with which a strong signal voltage makes a bright picture, and a weak voltage makes a dark picture. Our method reduces the chance for outside electrical interference to upset synchronizing of the beam, for reasons which will appear in due time.

THE COMPOSITE SIGNAL

Before starting the composite television signal through all the parts of a receiver it will be helpful to have some simple way of representing on paper the changes of voltage which make up this signal. The easiest way is the one universally used by television technicians and engineers. It shows variations of voltage or signal strength by a line which rises to show increases and falls lower to show decreases.

Fig. 2-6 shows how picture signals are represented. Here we have the signal for three successive horizontal lines. The image is practically the same in all the lines, indicating little or no movement. At the left is marked the voltage level for blanking, which leaves the screen black, and also the level for the whitest possible trace. Even for this whitest trace there is some signal voltage, as is evident from the fact that the level for zero voltage or no signal is quite a ways below the white level.

Time proceeds from left to right. That is, the left-hand end of this diagram represents a time earlier than the right-hand end. We commence, at the extreme left, with the beam blanked – the signal voltage is at the blanking level. Then begins a horizontal trace. During this trace the signal goes first away down to the white level, making the spot very bright. Then the signal rises all the way to the black level. Next comes a drop of signal voltage, which would produce a light gray part of the trace. This is followed by a rise of signal voltage, for a dark gray portion. Finally, the signal goes almost to the white level just before the beam again is blanked at the end of this first complete trace.

Formation of this first horizontal trace has taken up about 53 millionths of a second. The following blanking interval lasts for about 11 millionths of a second. Then comes another trace line, another blanking interval, and so on. Our entire diagram, including three picture lines and four blanking intervals, shows what happens in about 200 millionths of one second. To show on the same scale what happens during only one

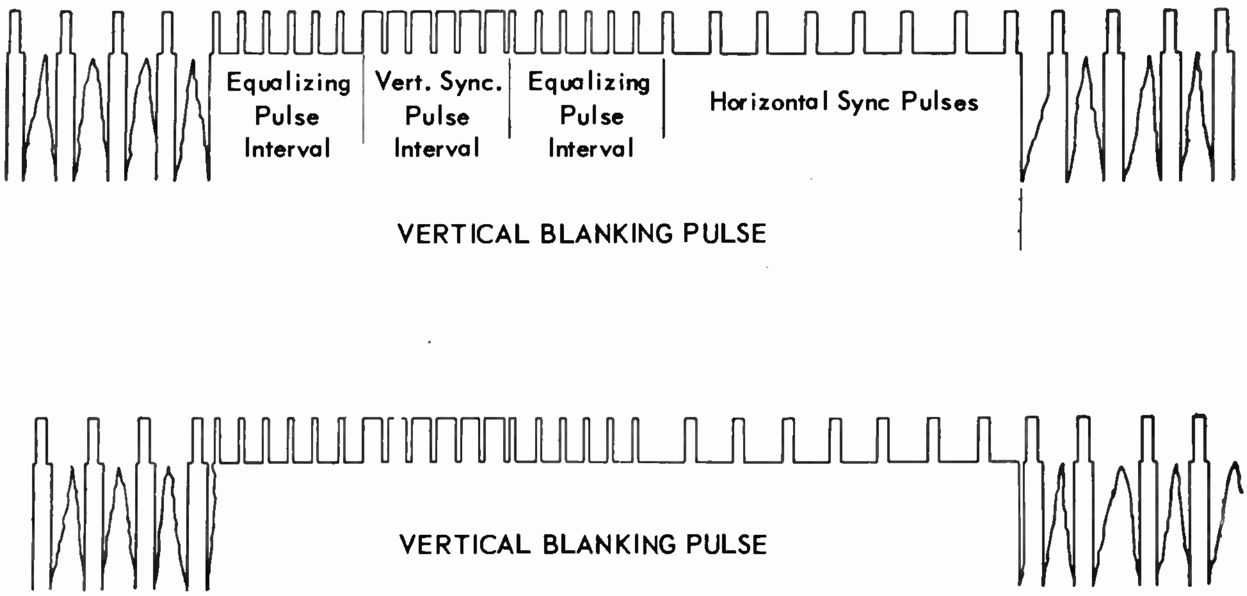


Fig. 2-8. To this composite signal have been added the vertical sync pulses, equalizing pulses, and horizontal pulses which appear in the two vertical blanking intervals occuring during formation of one complete picture.

second of time would require a diagram five thousand times as wide as this one.

When we add horizontal sync pulses to the signal it may be represented as in Fig. 2-7. These pulses are increases of signal voltage above the black level. During each sync pulse the signal voltage rises from the black level to the maximum voltage of that signal, then drops back to the black level. The horizontal sync pulses occur during the intervals of horizontal blanking.

For the present we shall not go on with an examination of vertical sync pulses, vertical blanking, and equalizing pulses. We have learned enough about the general form of the composite signal to talk intelligently about its adventures between the antenna and the picture tube.

FREQUENCIES

There is just one more thing which should be looked into before taking up the parts of the receiver. This thing is the matter of frequency. It is almost impossible to talk about the performance of different parts of a television receiver without understanding frequency, for it is the difference between operating frequencies that is responsible for most of the differences in construction and electrical behavior.

When we speak of the frequency of an electric current or an electron flow we tell how many times per second the electrons move back and forth, or alternate. Probably you are able right now to look at some wire which is carrying current for a lighted electric lamp. Quite likely you know that the lamp is operating on what we call 60-cycle current, or, in some parts of the country, on 50-cycle or even on 25-cycle current. If your electric supply current is of the 60-cycle variety, the electrons are moving lengthwise of the wire and back again 60 times during every second. If you have 50-cycle current the electric particles (electrons) in the wire are moving one direction and back again 50 times per second, and so on for any frequency.

If you watch a television receiver which is tuned, for example, to channel 4, the electricity in the antenna wires and in the first tubes of that receiver is moving back and forth about 70 million times a second. We say that the frequency is about 70 million cycles per second. Usually, when speaking of frequency, we don't say "per second". That part is understood. All frequencies are per second of time.

Should your television receiver be tuned to channel 11, as another example, the frequency in the antenna and connected parts is around 200 million cycles. Television in the ultra-high frequency channels is carried out with frequencies between 500 and 1,000 million cycles.

Possibly you have heard that electricity travels at the speed of light. This is a common misunderstanding. It is not electricity or electrons in wires, but only the electromagnetic radiation of radio signals which travels at nearly the speed of light. Electrons in wires move at speeds you could easily match with a fairly brisk walking pace. If the electrons moved much faster they would generate so much heat that the metal of the wire would disappear in a flash of fire and a puff of vapor.

Many a good electrician would doubt your statement that electricity merely walks through wires, but is easily proven. As yet you aren't well enough acquainted with the behavior of electrons or electricity to fully understand the proof, but you can read the next paragraph to any doubter.

First, the National Electrical Code prohibits a current of more than 15 amperes in rubber covered copper wire of number 14 gage size, the size used in most house lighting circuits. An ampere is a measure of rate of flow of electricity or electrons, it is a rate of one coulomb per second. A coulomb is a measure of quantity of electricity, just as quarts and gallons are measures for quantities of liquid. In one coulomb there are about 6¼ millions of millions of millions of electrons. On the generally accepted assumption that for each atom in a substance there is one electron capable of moving about, in one foot of number 14 wire there are about 3 3/4 coulombs of such electrons. If all these electrons move along at only 4 feet per second, the flow rate is 15 amperes. This is a speed considerably less than 3 miles an hour, and you can easily walk 4 miles an hour.

How about the movement of electrons when the frequency is 70 million cycles? Were you an electron, and couldn't move at a speed of more than 4 feet per second, and had to move back and forth 70 million times per second, how far would you move each time? You would merely stand there and vibrate. That's what the electrons do.

Frequency describes the rate of back and forth movement not only of electrons in wires, but of many other things associated with television. Television signals are distinguished from other radio signals largely by differences in frequency. The frequency of a signal is the rate per second at which its forces change direction.

In a radio signal as illustrated by Fig. 2-10 there are two forces. One is the same kind of force which exists in the space around a magnet, it is magnetic force. The other is the same kind of force that allows a piece of sealing wax to pick up a bit of paper after the wax has been rubbed on a woolen cloth, and is

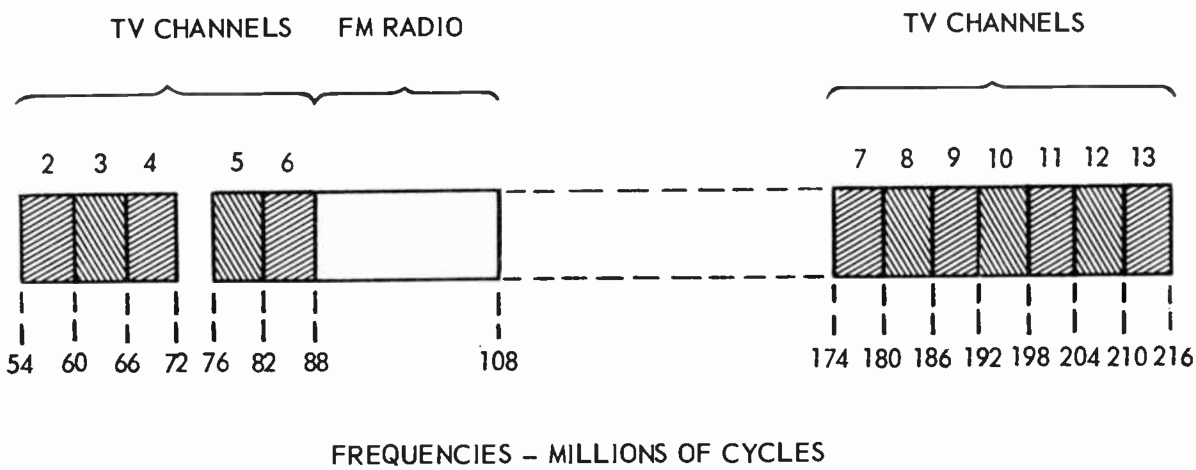


Fig. 2-9. Television channel frequencies in the low-band and high-band ranges, channels numbered 2 through 13.

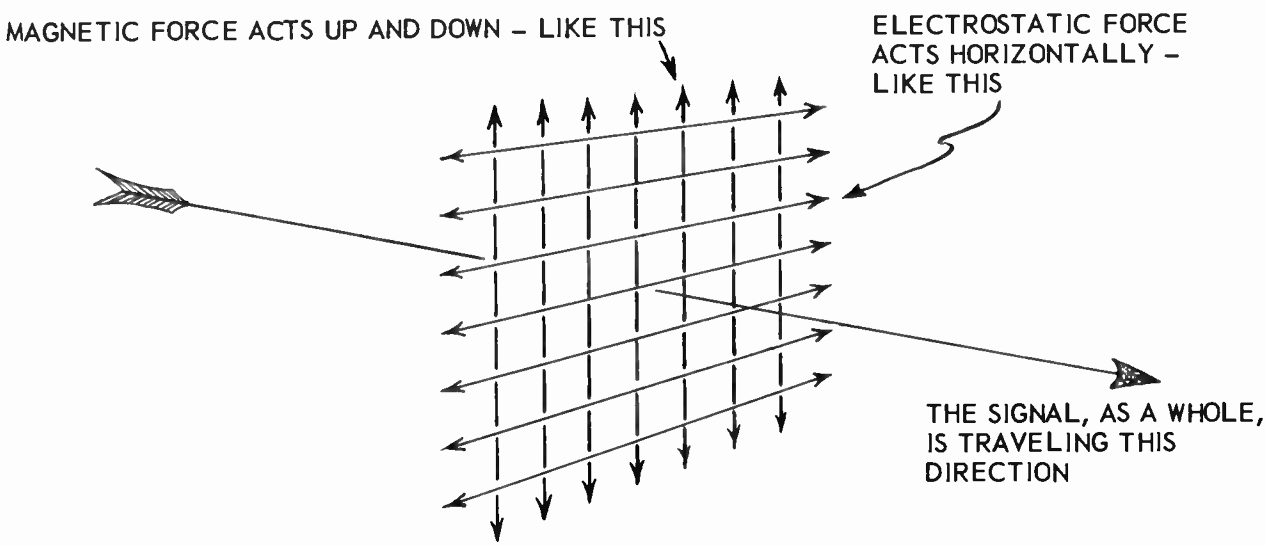


Fig. 2-10. Magnetic and electrostatic forces in a television signal act at right angles to each other and to the direction of signal travel.

Fig. 2-11. This is one type of television generator such as used by service technicians to supply all television frequencies during adjustment of receiver parts.

the kind of force that causes a tiny spark to jump from your fingertip to a radiator or water pipe after you have dragged your feet across a carpet. This is electric force or electrostatic force. Because the radio signal consists of both electric and magnetic forces we call it electromagnetic radiation.

When the kind of radio signal used for television is traveling east, its electric force acts alternately toward the north and the south, while its magnetic force acts alternately up and down. The number of times per second at which these electromagnetic forces go through a complete change of direction, as from north to south and back to north again, is the frequency of the signal.

It is the electric or electrostatic force in a television signal that acts on the receiving antenna. This force induces voltages in the antenna. These voltages reverse, or alternate, in time with the signal force. The frequency of the signal voltage induced in the antenna is exactly the same as the frequency of the signal itself. Signal voltages induced in the antenna cause movements of electrons in the wires and other metallic parts of the antenna. The electrons vibrate back and forth at exactly the frequency of the induced voltage, and at exactly the same rate as the signal force. Then, in the antenna, we have current or electron flow at the signal frequency.

In an earlier paragraph it was mentioned that frequency is measured in cycles per second. A cycle is any series of actions which repeats over and over again. If it is your habit to get up in the morning, eat three meals during the day, go to bed, and get up the next morning, that completes one cycle of eating and sleeping. When the electric force in the television signal acts in one direction, then in the opposite direction, and returns to the first direction, it has completed one signal cycle. When electrons in a wire start from

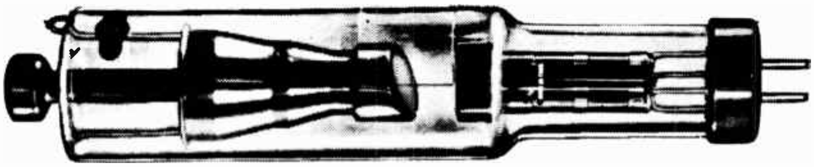


Fig. 2-12. A tube which delivers radiation at frequencies of hundreds of billions of megacycles. It is an X-ray tube for operation at 220,000 volts.

some certain position, move as far as they have time for in one direction, then in the opposite direction, and return to the starting point, they have completed one cycle of electron flow or one cycle of alternating current.

Most electric power and light systems carry voltages and currents which alternate at 60 cycles per second. This is the most common power frequency. When a loud speaker or other vibrating object causes particles of air to vibrate at frequencies between about 50 and 10,000 cycles per second the vibration affects our ears as sound. Some people can hear sound frequencies as high as 15,000 cycles, and a few have ears which respond at frequencies approaching 20,000 cycles. These frequencies between about 50 and 20,000 cycles per second are called audio frequencies, because they are audible as sound.

Frequencies measured as thousands of cycles per second usually are specified in kilocycles. One kilocycle is equal to 1,000 cycles. Kilo-, as a prefix, always means 1,000 times the unit. A frequency of 150,-000 cycles may be called a frequency of 150 kilocycles. Radio signals in the neighborhood of 125 to 400 kilocycles are used for guidance of ships and planes. Radio frequencies between 550 and 1,600 kilocycles are used for standard broadcasting.

Still higher frequencies usually are specified in megacycles. One megacycle is equal to one million cycles per second. Mega- is a prefix meaning one million times the unit. A frequency of 5 megacycles is one of 5,000,000 cycles per second. Short-wave radio communication and broadcasting are carried out at freqnencies between about 2 and 300 megacycles. Frequency-modulation or f-m radio uses signal frequencies from 88 to 108 megacycles. Television channels number 2 to number 6 carry signals at frequencies between 54 and 88 megacycles. (Channel 1 no longer is used for television broadcasting) Channels number 7 through 13 carry television signals at frequencies between 174 and 216 megacycles.

At frequencies around 100 million megacycles we have radiant heat, the kind that comes from the sun to the earth through 93 million miles of completely empty space. When the frequency of electromagnetic radiation reaches the range around 500 million megacycles it affects our eyes as light. When such radiation is at frequencies between 100 billion megacycles and 100,000 billion megacycles we have industrial and medical X-rays. This last frequency means the astronomical number of 100,000,000,000,000,000,000 cycles per second, which is something to think about.

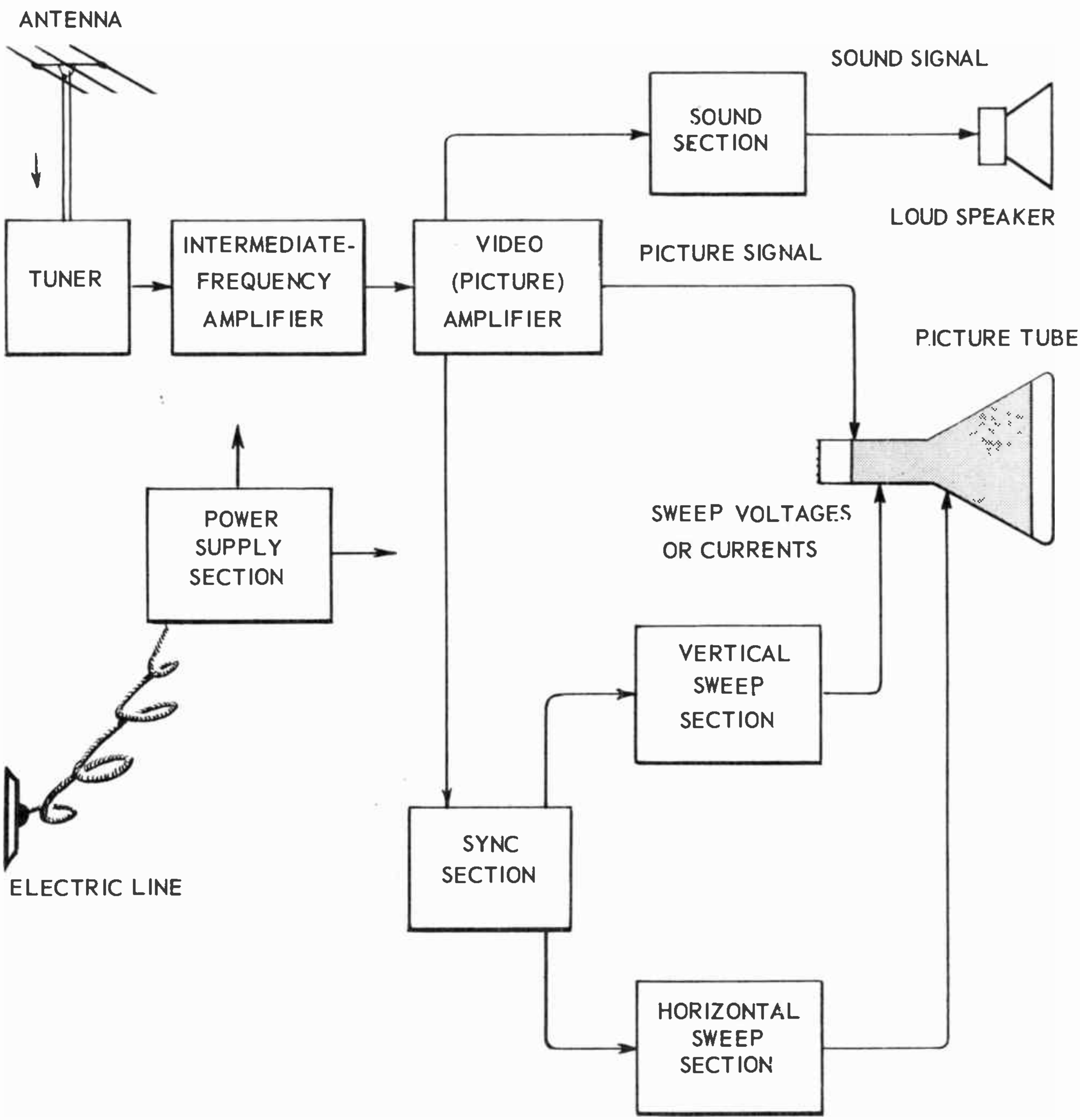


Fig. 2-13. The principal parts of a typical television receiver in the form of a block diagram, with arrows indicating the signal paths.

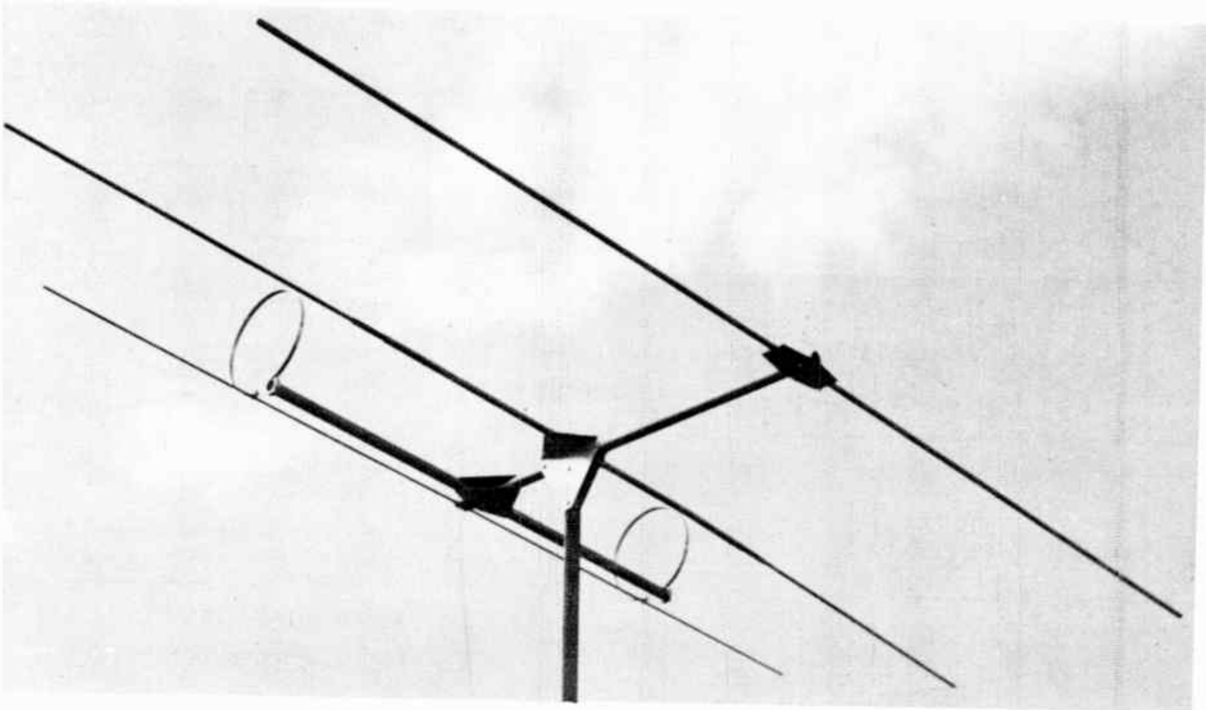


Fig. 2-14. One type of Tricraft television antenna for reception at all the channel frequencies in the low-band and high-band ranges.

SIGNAL PATHS IN A RECEIVER

Fig. 2-13 shows the paths of television signals through the principal parts or sections of a typical television receiver. The signal flying through space is picked up by the antenna and carried through wires to the tuner of the receiver. The television antenna may be of the outdoor type, supported on the roof or a chimney of the building in which is the receiver. One style of outdoor antenna is pictured in Fig. 2-14. Outdoor antennas of many designs are familiar sights wherever television signals are available.

In the earlier days of television it was possible to spot the location of every receiver, for, extending above the building housing the set would be an outdoor antenna. In England, where you pay a tax for the privilege of owning a radio or television set, the tax collector had an easy time. But as television receivers became more highly perfected and more sensitive the signals could be picked up by indoor antennas provided the receiver were only a few miles from the transmitter, and if other conditions were suitable. One type of indoor antenna is illustrated by Fig. 2-15. In some of the more recent television receivers the antenna is built into the cabinet. Then, if conditions are favorable, the only external connection to the receiver is the one from the electric power line.

The type of antenna required depends on these three things: First, the distance between receiver and transmitter. Second, the presence or absence of obstructions between receiver and transmitter. Third, the sensitivity of the receiver or its ability to operate from weak incoming signals.

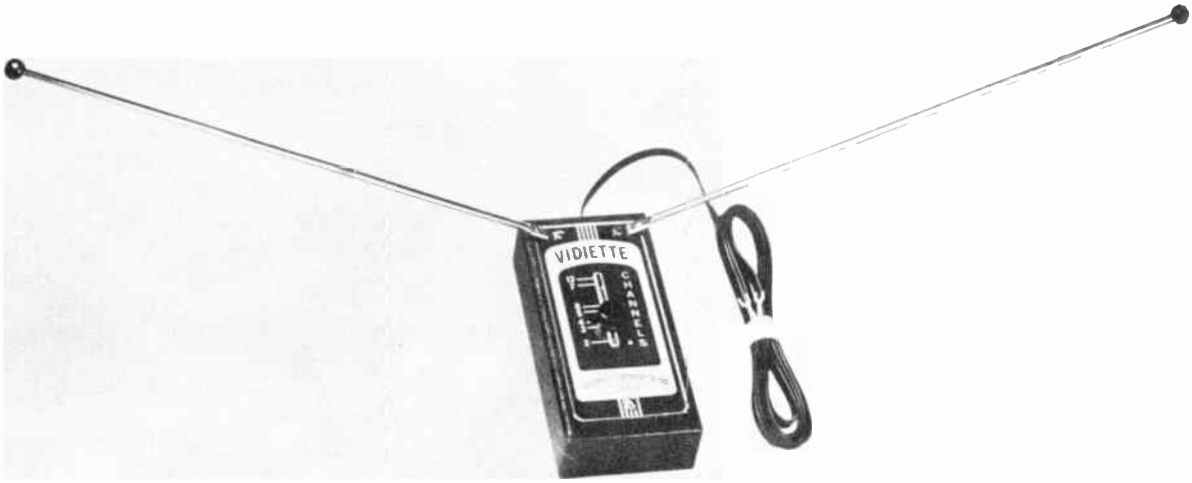


Fig. 2-15. A Vidiette indoor television receiving antenna with a unit allowing adjustment for different channels.

Signals at the high frequencies used for television and f-m radio transmission don't behave like those at the much lower frequencies of standard broadcast and international short-wave radio. When the lower-frequency radio signals leave a transmitting antenna they may be reflected back and forth between the surface of the earth and a layer of electrified gases high above the earth, and they can change direction to some extent when passing from one region to another in the atmosphere and when getting past the edges or corners of obstructions. In this manner these signals may travel thousands of miles to a receiver.

Quite different is the behavior of the very high frequency television and f-m signals. They act like the beam from a powerful searchlight, traveling only in perfectly straight lines, with only weak reflections from solid objects, and with rapid loss of strength as they travel away from the transmitting antenna.

The farther apart are the television transmitter and receiver the higher must be the transmitting antenna, the receiving antenna, or both, in order that the signal may pass from one to the other without meeting insurmountable obstructions along the way. This fact has made Mount Wilson, near Los Angeles in California, the world's greatest center for television transmission. Transmitting antennas on Mount Wilson are almost 6,000 feet above the level of nearby towns and cities. Signals reach receivers 100 miles away with good strength, and when everything is favorable the signals travel even farther.

The usual way of insuring good reception at distances greater than 20 to 25 miles between transmitter and receiver is to put up a rather elaborate antenna, or to raise the antenna higher in the air, or do both.

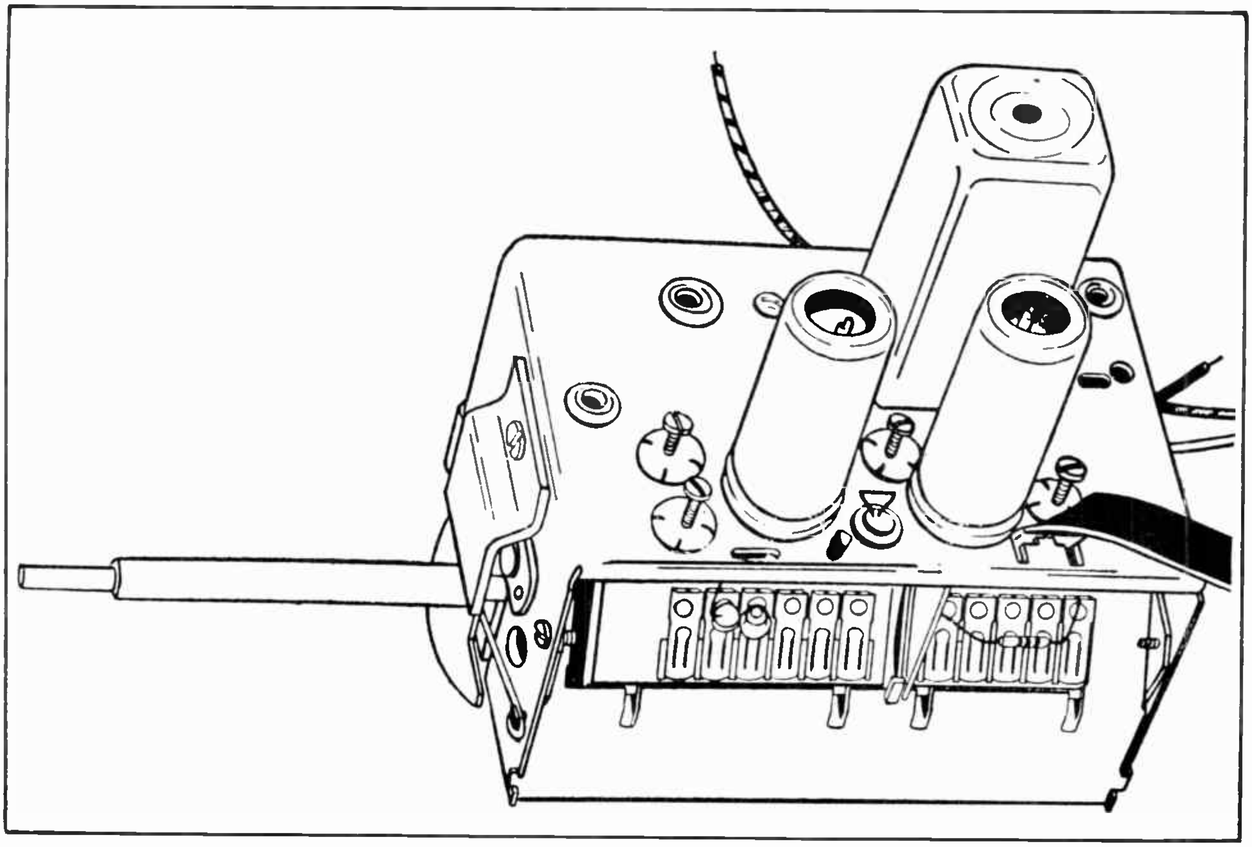


Fig. 2-16. A turret tuner of the kind used in a number of television receivers.

The limit of complexity in receiving antennas for the ordinary installation is on the order of the four-bay stacked array pictured back in Fig. 2-3. If you cannot get satisfactory signal strength from an antenna of this general style, mounted as high as is practicable, you are beyond the range for good reception.

The tuner of the television receiver is the part operated by the dial or pointer which you turn when selecting the channel on which you wish to look and listen. A tuner, in radio and television, is the part of the receiver which selects from all the signals reaching that receiver just the one in which you are interested, and rejects all the others. The tuner of any television receiver, of any f-m radio receiver, and of every superheterodyne radio receiver for any class of service does other things too. But let's stick to the tuning functions here in the beginning.

When talking about tuning for television and f-m radio we should consider the antenna, the tuner itself, and the connection between them as a single electrical assembly. This is because a lot of tuning or a lot of signal selection may be accomplished by suitable design of the antenna.

For every signal frequency there is one particular length of the horizontal rods or metallic tubes of the antenna which allows maximum pickup at that frequency. There are types of antennas which "peak" quite

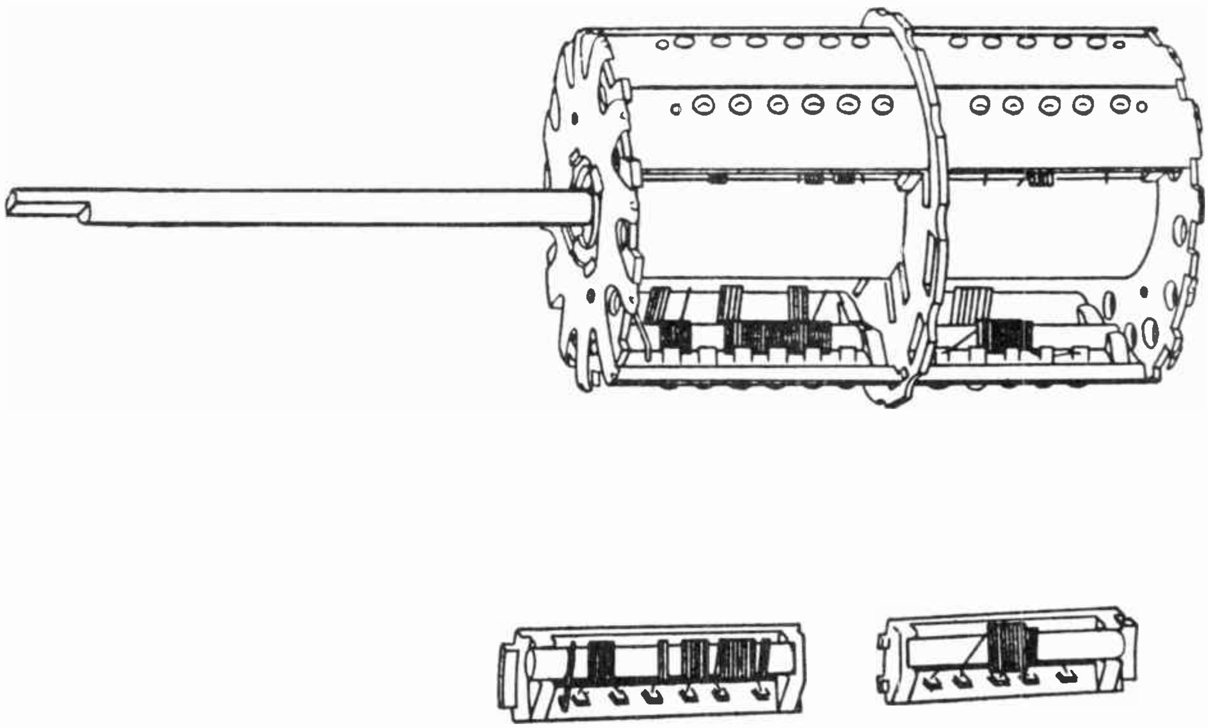


Fig. 2-17. The turret and two of the tuning strips.

sharply for one signal frequency. But we seldom employ these types because it is highly desirable to use only a single antenna, or at most two antennas mounted together, for all the signals to be received. We can, however, make this single antenna or dual antenna of such design as to be responsive throughout the entire television frequency band (of Fig. 2-9) and unresponsive to other transmissions at higher and lower frequencies.

If the antenna delivers to the tuner all the television signals which are reaching your locality at a given time, and does not bring in too many other unwanted signals or electrical impulses, this is all we ask of the antenna. Then it is the work of the tuner to make the final selection of the one signal or one program which you wish to receive, and to get rid of everything else so far as possible.

The tuner is the most intricate single part of the television receiver, at least from the standpoint of mechanical construction. The tuner is not so wonderful in its action as is the picture tube, but easily rates second in interest. Tuners are made in a great variety of mechanical designs. Fig. 2-16 shows what is called a turret construction. Inside the cylindrical metal "cans" on top of this unit are two tubes. Inside the rectangular can is a combination of capacitors, inductors, and resistors which handle part of the work of signal selection.

The part of this tuner which allows you to pick the desired channel is shown by itself in Fig. 2-17. This

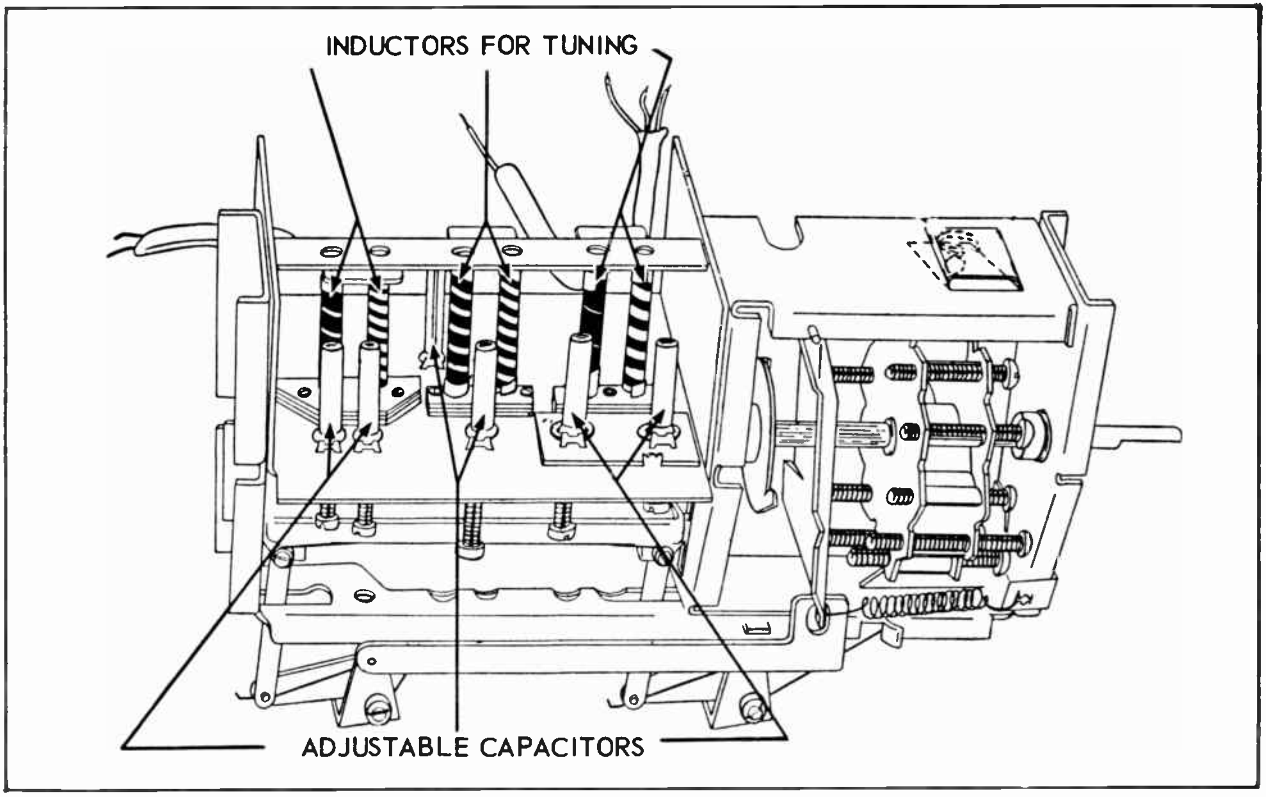


Fig. 2-18. Another type of television tuner. Here the tuning is by means of adjustable inductors.

is the turret. The tuning knob is placed on the outer end of the long shaft so that you may rotate the turret. For each channel the inductors and capacitors on one lengthwise section of the turret are thus electrically connected to the tubes on top. Two of the strips which carry these inductors and capacitors are shown below the turret.

You must have noticed that inductors, capacitors, and resistors have been mentioned many times. Yet their purposes and actions have hardly been touched upon. This is because these subjects are too big and too extensive to get into before you appreciate the real importance of these three kinds of units in every part of every receiver. If you knew all that a service technician needs to know about inductors, capacitors, and resistors, you would have mastered a large percentage of the practice of servicing.

Fig. 2-18 shows a different type of tuner. The mechanical construction is wholly unlike that of the tuner in Fig. 2-16, and the electrical action is decidely different. Yet, were you to feed certain signals from an antenna to one of these tuners you could have from the tuner output exactly the same signals as produced in the output of the other tuner with the same input signals.

We haven't gotten very far in following the television signal through the diagram of Fig. 2-13, but maybe far enough for one session. In the next lesson we shall continue along with the signal, and talk about some of the things which happen to it, about some of the problems encountered, and how they are solved.

TELEVISION

LESSON NO. 3 PRODUCING PICTURES AND SOUND

Training Your Mind...
C. T. I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

COMMERCIAL TRADES INSTITUTE

Training Your Mind — C.T.I. — Perfecting Your Skill

LESSON NO. 3

PRODUCING PICTURES AND SOUND

To the tuner of the television receiver are delivered all the signals which are reaching the antenna. Then the tuner goes to work.

First, the tuner selects the composite and sound signals for the one channel to which you have tuned the receiver at that particular time.

Second, the tuner strengthens or amplifies these selected signals and weakens all others.

Third, the very high frequencies of the selected and amplified signals are changed to lower frequencies at which further amplification may be more easily and efficiently carried out in parts of the receiver which follow the tuner.

To perform these three jobs the tuner may utilize two, three, or even more than three tubes in connection with many other parts. However, the very first function of the tuner, that of signal selection, is not performed by tubes at all. An amplifier tube has no power of selection, it amplifies whatever signal you give it. Signal selection is accomplished by parts which are connected in between the tubes. These parts are couplers.

A coupler is a device for transferring signal voltages from the output side of one tube to the input side of a following tube, or for transferring signal voltages from the antenna connections to the input of the first tube. There are many possible ways of handling this signal transfer, there are accordingly many types of couplers and many different designs for each type. At the left in Fig. 3-2 is one style of tuned transformer coupler. At the center is one kind of tuned impedance coupler. The screws sticking out from the tops of these two couplers are service adjustments which permit you to select the frequencies at which there is most effective signal transfer. These screws move a part of the coupler called its core. A core, with its adjusting screw, is pictured at the right.

When a coupler is adjusted or tuned for signals of some one frequency or a small range of frequencies it allows signals at these frequencies to pass through with the greatest of ease. Signals at all other frequencies meet with great opposition and are decidedly weakened. The signals for which couplers are adjusted or tuned pass from one amplifier tube to another, becoming progressively stronger. The opposition of the couplers to unwanted signals is so great that these other signals become weaker and weaker as they go through one coupler after another, in spite of the amplification applied by the tubes.

With signal selection taken care of by couplers only two things remain to be done by tubes in the tuner. These are, first, to strengthen or amplify the selected signals, and second, to change the signal frequency.

The wires which bring signals from the antenna to the tuner lead first to a coupler. The other side of this antenna coupler is connected to the first tube in the tuner. This tube is a radio-frequency amplifier. Its pur-

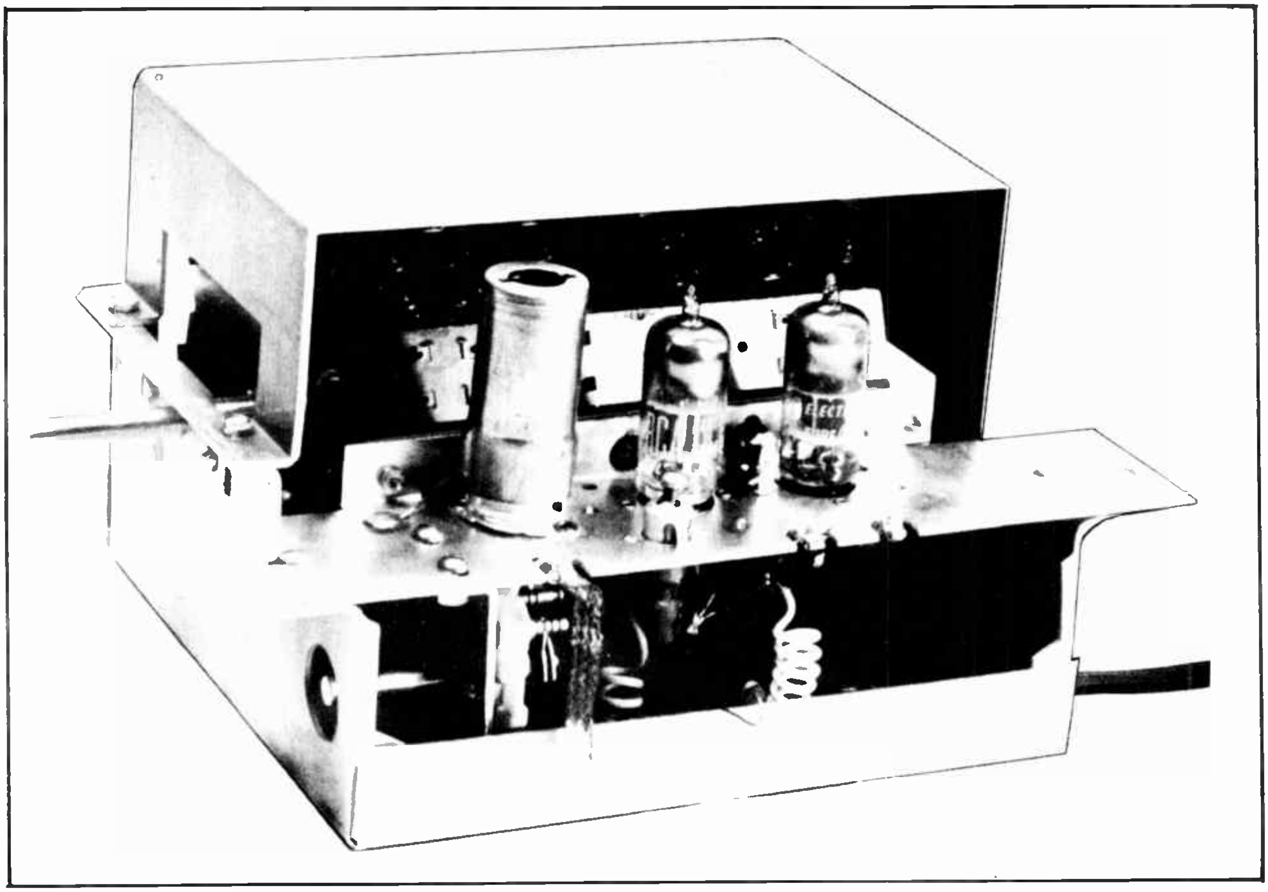


Fig. 3-1. This turret tuner employs three tubes. Two are exposed, one is inside a cylindrical shield at the left.

pose is to strengthen or amplify signal voltages coming to it from the antenna coupler. As a rule there is only one radio-frequency amplifier tube, although some receivers have two.

Signal voltages from the output of the radio-frequency amplifier tube go through a second coupler and to the input of another tube called the mixer. This is the tube in which the very high frequencies of the received signals are changed to lower frequencies for additional amplification farther along in the receiver.

This change of frequency is brought about by feeding into the mixer tube, right along with the signal voltages, another voltage whose frequency is 20 to 40 megacycles higher than frequencies of the received signal. When voltages at the two frequencies combine in the mixer tube the result is other voltages at frequencies equal to the difference. For example, were the frequency of a received signal to be 70 megacycles, and were the frequency of the additional voltage applied to the mixer to be 95 megacycles, the output of the mixer tube would contain a new voltage having a frequency equal to the difference, or to 25 megacycles. This difference frequency is called the intermediate frequency, because its value is in between the frequencies of the received signal and the frequencies finally delivered to the picture tube and the loud speaker. Signal voltages at the intermediate frequency are amplified in the following parts of the receiver.

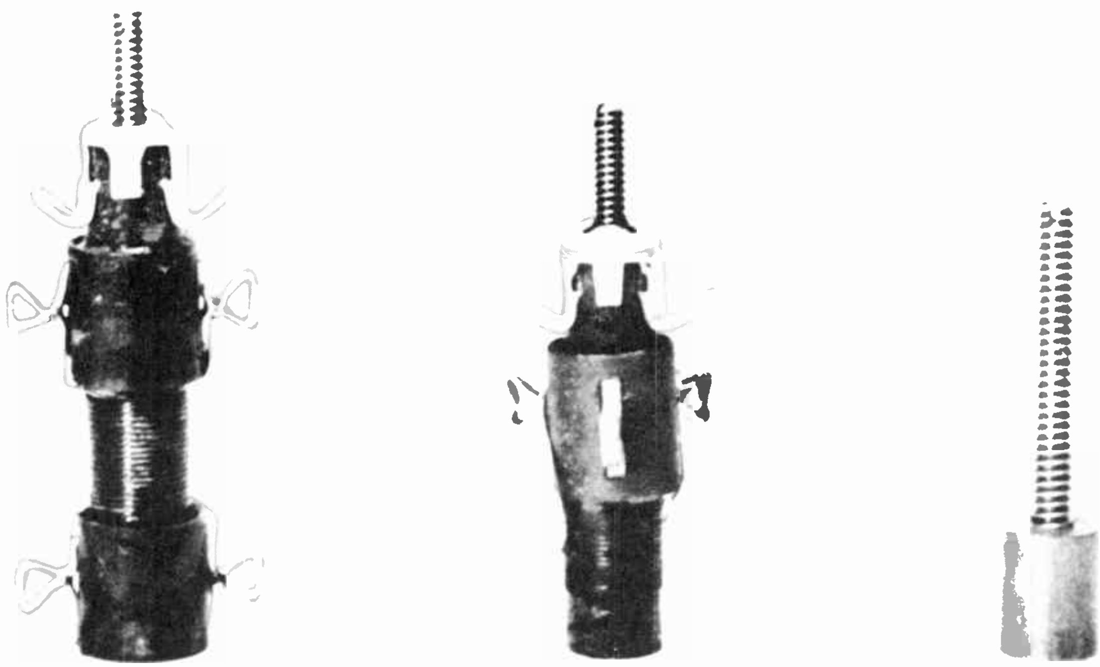


Fig. 3-2. Two types of couplers and an adjustable core used in these types.

Voltages at the intermediate frequency vary in precisely the same manner as voltages in the signal from the antenna. Therefore, these intermediate-frequency voltages contain picture signals, a blanking level, sync pulses, and all other characteristics of the original signal. There are also intermediate-frequency voltages varying in the same manner as the television sound signal at the antenna.

The additional voltage applied to the mixer tube, the voltage whose frequency is 20 to 40 megacycles higher than the original signal voltages, comes from a radio-frequency oscillator or a local oscillator as it is sometimes called. An oscillator, of any kind whatever, is a tube or part of a tube producing voltages which rapidly change their direction. We would say that an oscillator produces alternating voltages and accompanying alternating currents. An alternating current is the kind which exists when electrons vibrate back and forth in wires and other parts.

Although the tubes in the tuner have to perform three jobs, we don't have to have three separate tubes. In order to save space, cut down the number of wiring connections, and provide manufacturing economies, all the internal parts needed for doing two different things may be built into a single tube. In a single tube there may be two radio-frequency amplifiers, with the output of one fed to the input of the other. In a single tube we often have both the mixer and the oscillator for television.

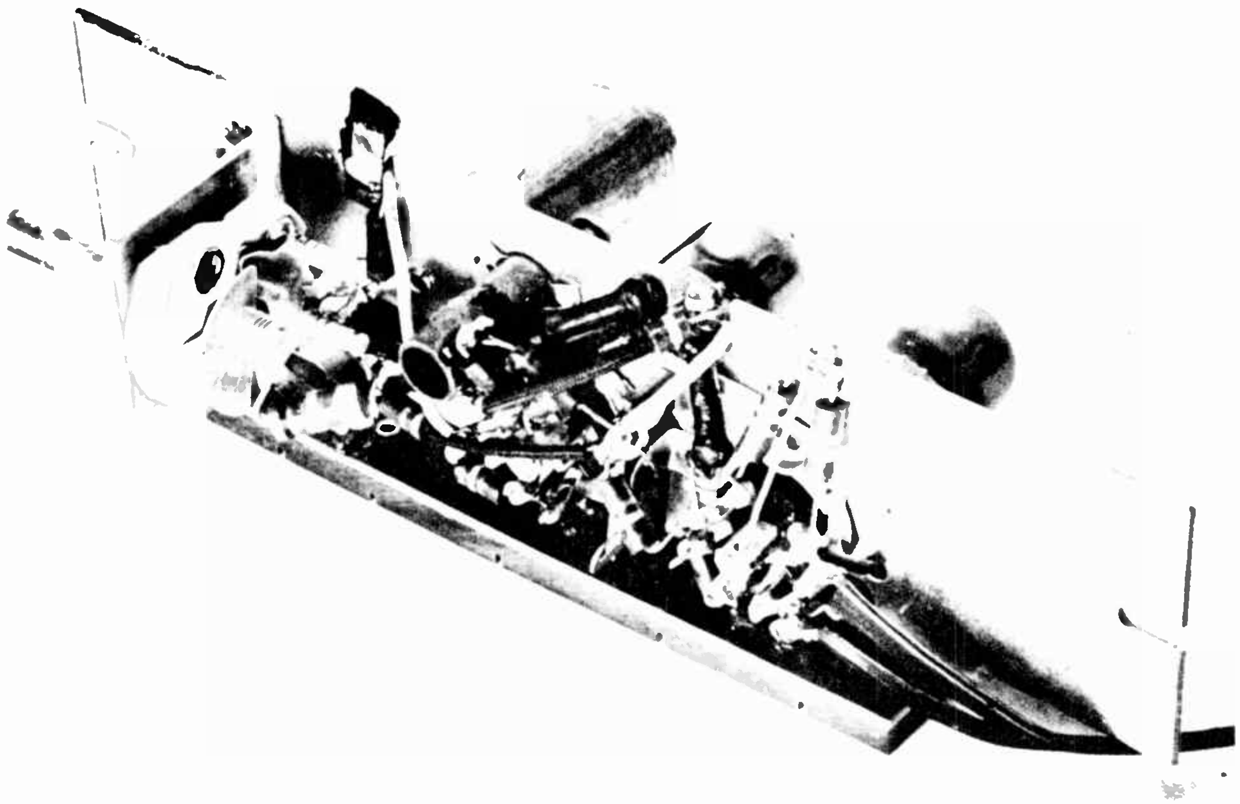


Fig. 3-3. The underside of the tube shelf of a television tuner.

If you are familiar with broadcast receivers for sound radio you will recognize that in the television tuner we have tubes and couplers performing just like tubes and couplers in the "front end" of any superheterodyne receiver. In every superheterodyne receiver the voltage of the received sound signal is amplified by one or more radio-frequency amplifier tubes. The oscillator of the superheterodyne produces voltages at a different frequency. The signal and oscillator voltages are combined in a mixer or converter tube, and the result is signal voltages at an intermediate frequency suitable for further amplification.

The parts of a television receiver which handle the composite signal and the sound signal employ exactly the same principles found in ordinary superheterodyne sound receivers. Although the principles are the same, frequencies are so much higher in the television receiver that we find constructions and performance decidedly different than in sound receivers. After the television sync pulses are separated from the composite signal, and we follow these pulses through the sync section and sweep section of the receiver, even the principles are unlike anything found in sound receivers.

Let's pause here for just a few moments to add more items to our television and radio vocabulary. First, the term radio-frequency is so long that we abbreviate it to the letters r-f. Then, instead of speaking of a radio-frequency amplifier or oscillator we talk about r-f amplifiers and r-f oscillators. Intermediate-frequency,

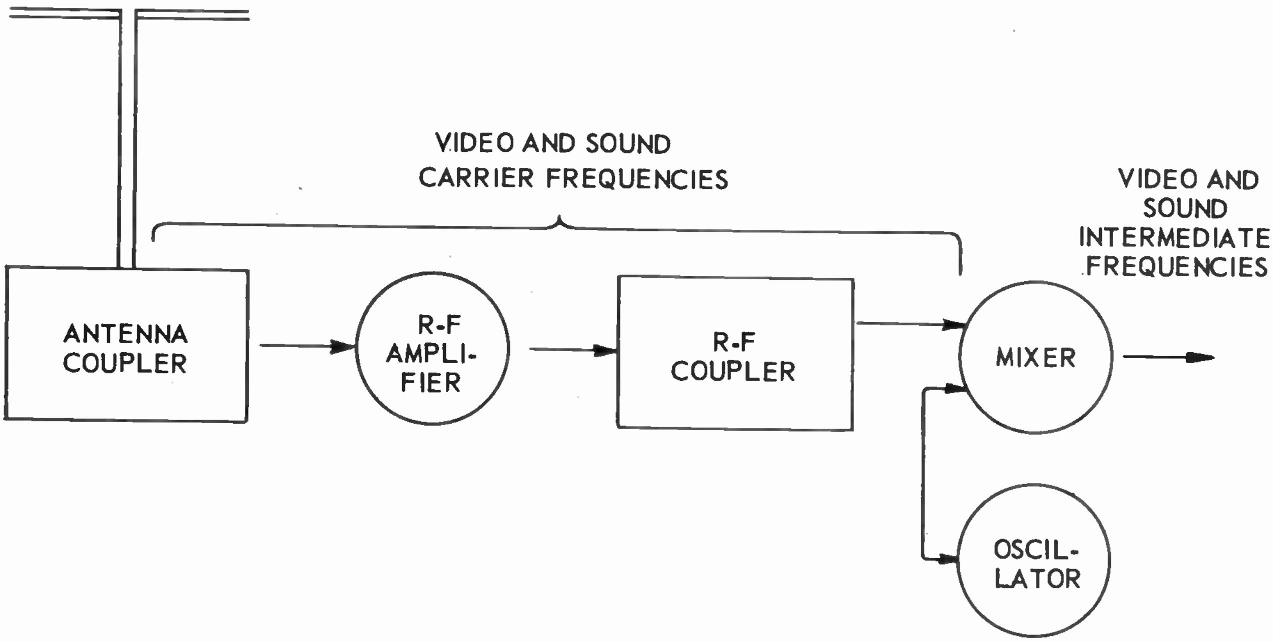


Fig. 3-4. Where various frequencies are found in the tuner.

another long term, is abbreviated to the letters i-f. Later we shall have much to do with such things as if-amplifiers.

Units of frequency greater than cycles nearly always are abbreviated. Kilocycles are abbreviated to kc. Megacycles are abbreviated to mc. Instead of writing 20 megacycles we write 20 mc, and so on.

When the composite television signal and the television sound signal leave the mixer tube of the tuner these signals are at intermediate frequencies. To distinguish between these intermediate frequencies and the original signal frequencies at the antenna it is common practice to speak of frequencies coming to the antenna and existing in the antenna as carrier frequencies. They are the frequencies at which the signals are carried through space.

In talking about the portion of the television signal which contains the picture variations, sync pulses, equalizing pulses, and everything except the television sound, we have been using the term "composite signal". It is more usual to call this the video signal. Then the remainder of the complete television signal is called the sound signal.

By using these new words we may easily identify the frequencies of four signals which are highly important in our work. Fig. 3-4 shows where these frequencies are found. Tubes are here represented by circles, couplers by rectangles. We have the video carrier frequency, which is the frequency around which centers the video (composite) signal formed at the antenna and going as far as the mixer tube in the tuner. There is also the sound carrier frequency, the frequency around which centers the sound signal reaching the antenna, and which goes as far as the mixer tube. Then, following the mixer tube, we have the video intermediate frequency and the sound intermediate frequency, which are the frequencies around which center the video sig-

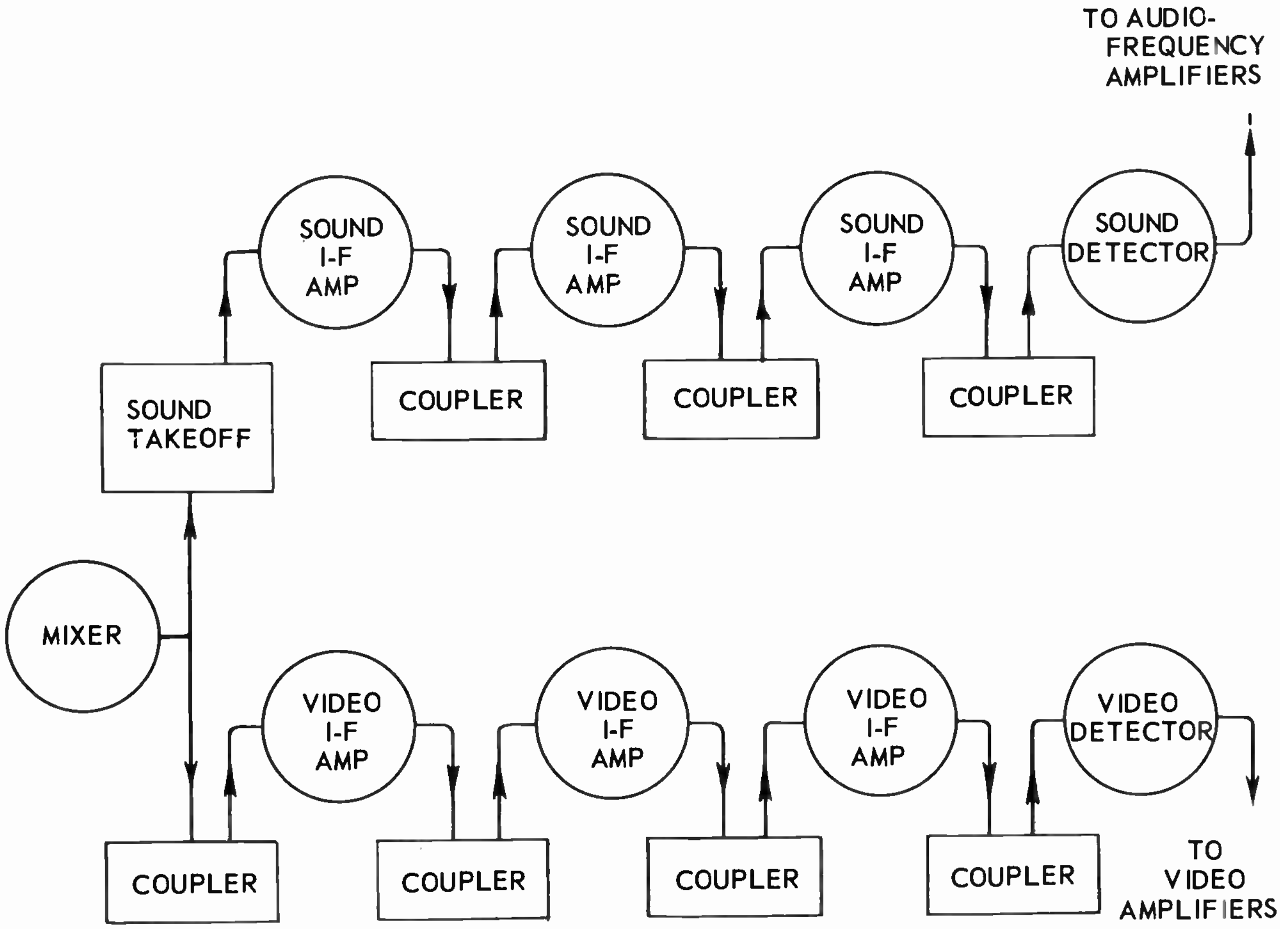


Fig. 3-5. Separation of sound and video signals immediately following the mixer.

nals and the sound signals after they leave the mixer tube.

This talk about some signals being carried by others, and about some frequencies centering around others, may be a bit confusing. To make a complete explanation would take us deep into the subject of modulation, for which we are not quite ready. By means of the process called modulation it becomes possible for a signal at some relatively low frequency to ride along on another signal at higher frequency. The reason for employing modulation is that high frequency signals are easily transmitted through space, while low frequency signals are not.

You really are quite familiar with the effects of modulation. When you tune the dial of a standard broadcast receiver for some certain station you turn the pointer to a number which corresponds to kilocycles of carrier frequency used by that station. These carrier frequencies range all the way from 550 to 1,600 kilocycles. Yet you listen to voices and music at audio frequencies, which are between 50 and 5,000 cycles as they come from the speaker of an ordinary set. These relatively low audio frequencies have been carried through

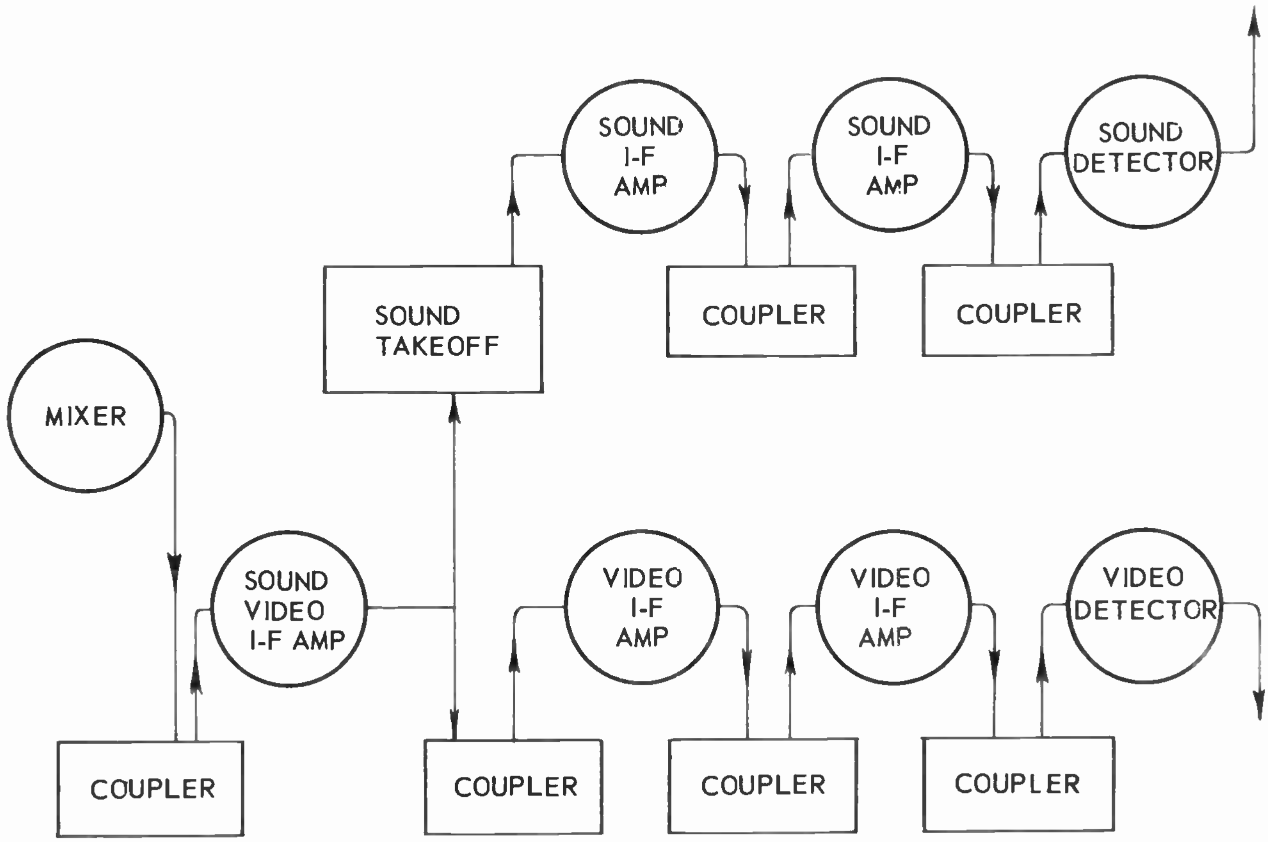


Fig. 3-6. Both the sound and the video signals may go through several i-f amplifiers before the two signals are separated.

space by the much higher carrier frequencies. The carrier frequencies originally were modulated with the audio frequencies back at the radio transmitter. In your receiver the carrier is demodulated, which means to separate the sound frequencies so that you may listen to them.

THE INTERMEDIATE-FREQUENCY AMPLIFIERS. In the output of the mixer tube we have the video intermediate frequency and the sound intermediate frequency. In all early television receivers the signals at these two frequencies were separated from each other as soon as they came from the mixer. This method is represented in Fig. 3-5. Signals at the video intermediate frequency then pass through the video i-f amplifiers, and signals at the sound intermediate frequency pass through an entirely separate set of sound i-f amplifiers. Many of the newest television receivers still employ this method.

You should note that in this diagram, and in others to follow, the output of one amplifier tube is not connected directly to the input of the following tube. Rather the connection from one tube to the next is made through a coupler in every case.

After the sound i-f signal has been amplified by as many sound i-f amplifier tubes as may be used in the receiver this signal goes to the sound detector or sound demodulator. The detector or demodulator separates

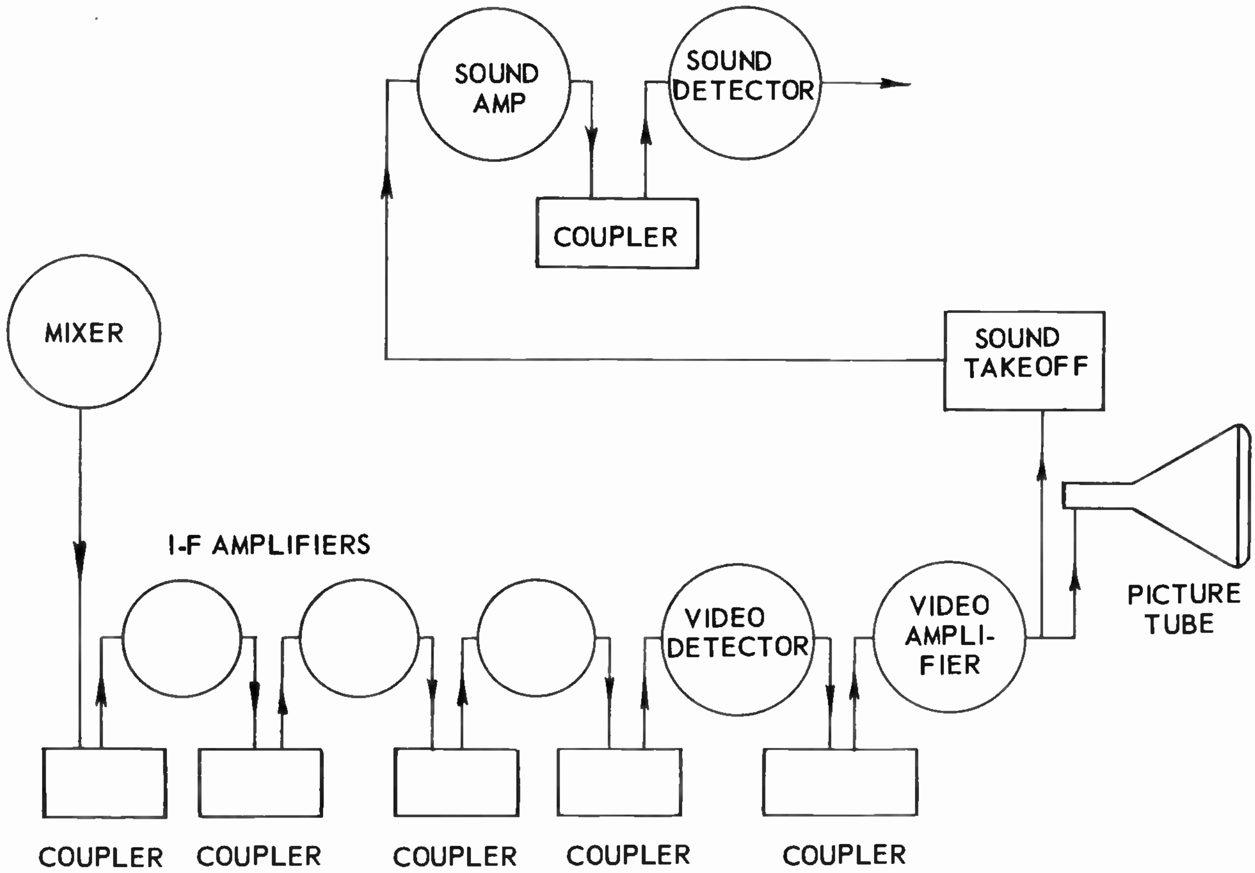


Fig. 3-7. With intercarrier sound systems the sound and video signals pass together as far as the video detector and sometimes beyond.

the actual sound signals, which are at audio frequencies, from the voltages at sound intermediate frequencies which have served to carry the audio-frequency signals through to this point.

These sound i-f signals, whose frequencies are between 20 and 40 megacycles, have been modulated with the audio-frequency sound signals whose frequencies are between 50 and 20,000 cycles. Now the detector or demodulator gets rid of the intermediate-frequency voltages and passes the remaining audio-frequency signal voltages to one or more audio-frequency amplifier tubes from which the strengthened sound signals go to the loud speaker.

The video i-f signals go through a series of couplers and video i-f amplifier tubes to the video detector. These video i-f signals, at frequencies of 20 to 40 megacycles, have been modulated in accordance with the video signal voltages. These video signal voltages correspond to picture variations, the blanking level, sync pulses, and equalizing pulses.

The video detector gets rid of the intermediate-frequency voltages and passes the picture signals, sync pulses, and all the rest of the modulation to one or more following video amplifier tubes. From the last of

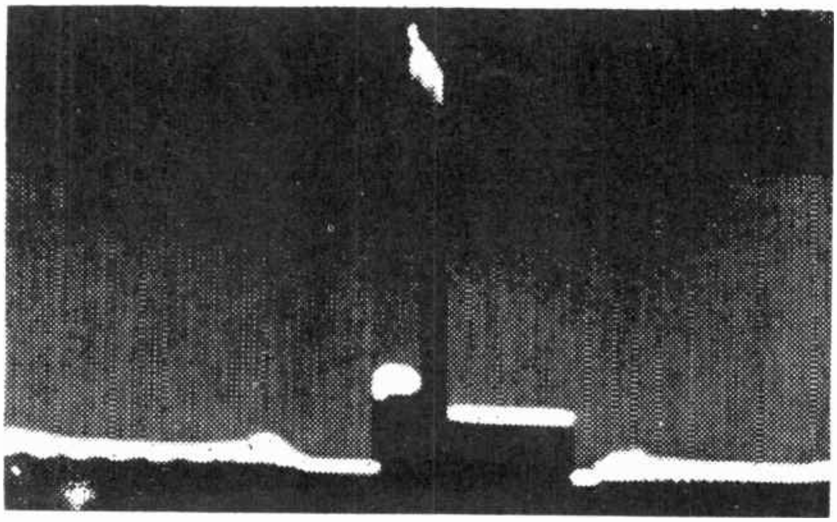


Fig. 3-8. An oscilloscope trace showing a vertical blanking interval and vertical sync pulse.

these amplifiers the video signal goes to the picture tube.

Still later in the development of television there appeared many receivers in which both the video and sound i-f signals pass together through one or more i-f amplifier tubes, but not through all the i-f tubes. This is shown by Fig. 3-6. At some point along the string of i-f amplifiers the sound i-f signal is separated from the video i-f signal. The video i-f signal then goes on by itself through remaining video i-f amplifiers, while the sound i-f signal passes through one or more separate sound i-f amplifier tubes. This method too still is followed by many of the newest receivers.

In a great number of recently designed receivers both the video i-f and sound i-f signals pass together through all the i-f amplifier tubes, and both signals enter the video detector. Now, for the two frequencies of these two signals, the video detector acts as a mixer tube. Just as the mixer tube back in the tuner produces an output frequency equal to the difference between frequencies of the carriers and the r-f oscillator, so the video detector now delivers in its output a signal frequency equal to the difference between video and sound intermediate frequencies. The signal at this difference frequency from the detector still is modulated with the original sound signal, it carries the sound signal. This method is illustrated by Fig. 3-7.

In the output of the video detector of Fig. 3-7 there is also the complete video i-f signal which carries the pictures, sync pulses, and so on. At some point following the video detector the sound signal is separated from the video signal. The sound signal then goes through an additional sound amplifier to a sound detector or demodulator, while the video signal goes to the picture tube. The separation of sound from video may be immediately following the video detector, but more often this separation is made only after both the video and sound signals have passed through one or more additional amplifier tubes.

This method last described is called the intercarrier sound system. The frequency of the sound signal, as finally separated from the video signal, is the same as the difference between sound carrier and video carrier frequencies coming to the antenna. This difference always is 4.5 megacycles, regardless of the channel to which the receiver is tuned.

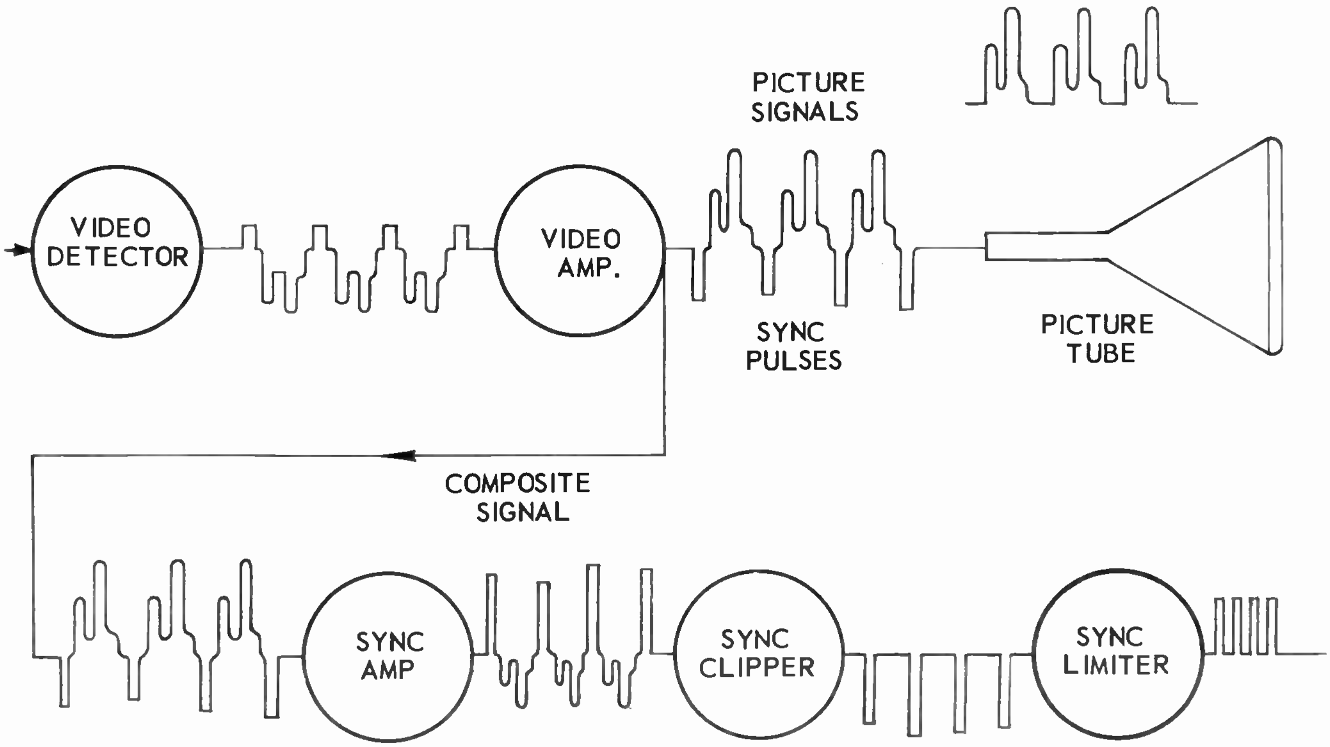


Fig. 3-9. Forms of signals in parts of the receiver following the video detector.

TELEVISION SOUND. As mentioned in an earlier lesson, the television sound signal is frequency modulated. This signal for television sound is of the same form as the frequency-modulated sound signals coming to the antennas of f-m sound receivers which are in common use. This f-m sound signal goes through the television tuner just as through the tuner of an f-m sound receiver, and comes from the mixer as a frequency-modulated i-f signal. Then, in the television receiver, the f-m sound signal proceeds right along with the video i-f signal until reaching the point of sound takeoff. The sound takeoff is merely another coupler, adjusted or tuned to the sound intermediate frequency so that this frequency passes easily through the takeoff while video signals are opposed and are kept out of the sound amplifiers.

The sound i-f amplifiers of Figs. 3-5 and 3-6 are similar to the i-f amplifiers in an f-m sound receiver, as is also the single sound amplifier of Fig. 3-7. The sound detectors or demodulators are exactly like those used in f-m sound receivers, and the audio-frequency amplifier section of the television set is like the audio-frequency system in any sound receiver.

Because of this similarity between television sound systems and f-m sound receivers we shall defer further examination of television sound for the time being. Just now we are interested in following the video signal, for it is in what happens to this signal that we find the great differences between receivers for television and those for sound alone.

LOOKING AT SIGNAL VOLTAGES. While following the signal through remaining parts of the television receiver it will be convenient to show picture signals, sync pulses, and blanking levels by means of diagrams somewhat similar to those used during explanations of signal voltages in an earlier lesson. Because changes of signal voltage involve times of only a few millionths of a second you may have assumed that diagrams showing such changes are merely theoretical. Actually you will watch such voltage changes during much of your trouble shooting work, and will watch the effects on these voltages of whatever adjustments you may make.

The instrument used to observe changes in signal voltages is called the oscilloscope. An oscilloscope contains a cathode-ray tube, which is like a television picture tube employing electrostatic deflection. The electron beam in the cathode-ray tube of the oscilloscope is deflected vertically by whatever signal voltage you wish to observe. At the same time the beam is swept horizontally so that you may see how the signal voltages change from one instant to another.

Were you to connect an oscilloscope to some part of a receiver in which the composite signal is acting, and adjust the oscilloscope for observation of a vertical blanking interval, the "picture" on the screen of the cathode-ray tube would appear somewhat as shown by Fig. 3-8. Here you can plainly see the gap which is the vertical blanking interval, also the vertical sync pulse near the beginning of this interval, and you can see as a sort of haze on each side the effects of picture signals and horizontal sync pulses occuring before and after the vertical blanking.

In a hand-drawn diagram the details of the signal might be shown more distinctly than by the oscilloscope trace, and the diagram might make it easier to understand what the signal is supposed to do. But the oscilloscope trace is the signal itself, as it actually exists at the moment of observation. Because he already is familiar with diagrams of signals, the service technician understands what the oscilloscope trace should look like if everything is working OK, and he can identify many kinds of trouble by the actual appearance of the signal.

Keep this in mind. When we look at signal diagrams and other representations of voltages in the lessons we are becoming acquainted with the meanings of voltages which you will see many times on the screen of the oscilloscope during service work. Now we shall continue following the signal through the receiver.

SEPARATING THE PICTURE SIGNALS. At the output of the video detector we have the entire composite signal as shown at the upper left in Fig. 3-9. This signal goes through the video amplifier tube, which increases the voltages. Then the entire composite signal goes to the picture tube. The picture portion of the signal will be used to produce lights and shadows on the screen, but the sync pulses of the signal must not be allowed to effect the pictures.

To get rid of the sync pulses and retain the picture variations of the signal we make use of the fact that all picture variations of voltage are on one side of the blanking level, with all sync pulses on the other side of the blanking level. It is necessary only to operate the control grid of the picture tube in such a way that the beam is blanked whenever the signal voltages go over onto the sync pulse side of the blanking level. This effectively removes the sync pulses from the signal used in the picture tube.

THE SYNC SECTION. Although we don't want sync pulses in the picture tube, we do want them for other parts of the receiver where these pulses are needed for control of vertical and horizontal deflection of the beam. In these other parts of the receiver we don't want picture signals, because sweeps of the beam must not be affected by picture signal voltages. So our problem becomes one of separating the sync pulses from the composite signal, of saving the sync pulses, and getting rid of the picture signals. This is the function

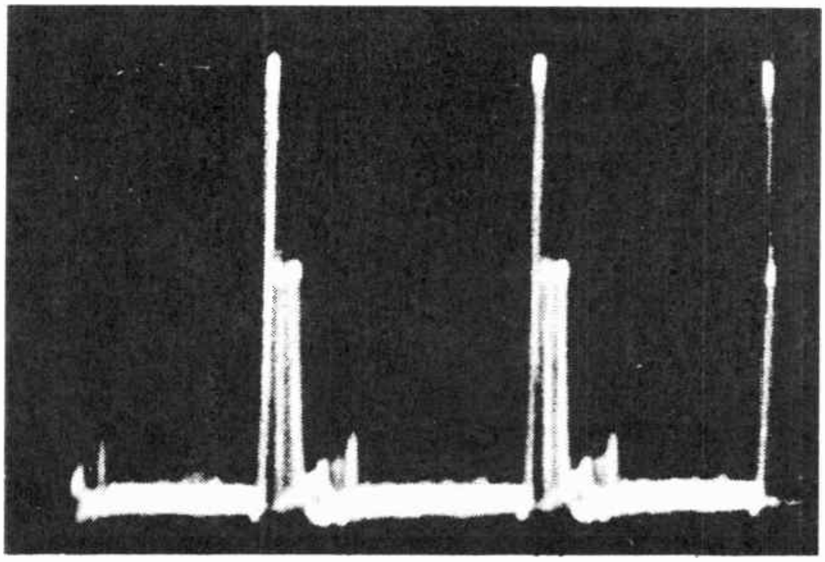


Fig. 3-10. How the oscilloscope shows the output voltage from a sync clipper.

of the sync section of the receiver.

Going back to the output of the video detector in Fig. 3-9 we find there the entire video signal; the picture variations, sync pulses, and everything else. This signal goes through the video amplifier tube, which increases the voltages. Here we should note that the video amplifier does something else of importance. It turns the signal upside down or inverts the signal. All the changes of voltage which, in the output of the video detector, acted upward with reference to the blanking level now act downward in the output of the amplifier. All changes which at first acted downward now act upward.

Inversion of signal voltages occurs in every tube which is acting as an amplifier. When signal voltages are amplified by a tube they always are inverted at the same time. To get the signal voltages right side up again we put them through a second amplifier. If this is followed by a third amplifier the voltages come out of this last amplifier upside down. Signal inversion comes in handy many times, for it allows obtaining voltage changes in whichever direction is more useful. But whether inversion is helpful or not, it always takes place when signal voltages go through any amplifier tube.

There may be one or more video amplifier tubes, depending on how much the signals are to be strengthened. From the last video amplifier the signal goes to the picture tube. As previously explained, the picture tube discards the sync pulses and uses the picture signal to vary the intensity of illumination on the screen.

From the output of one of the video amplifier tubes, or sometimes from the output of the video detector, we take the composite signal or entire video signal and carry it to the sync section of the receiver. In Fig. 3-9 this entire composite signal is applied to a sync amplifier tube, which here is the first tube in the sync section.

This sync amplifier is operated in such a way as to amplify the sync pulses more than the picture variations of the signal, as you may observe from the relative heights of sync pulses and picture signal voltages

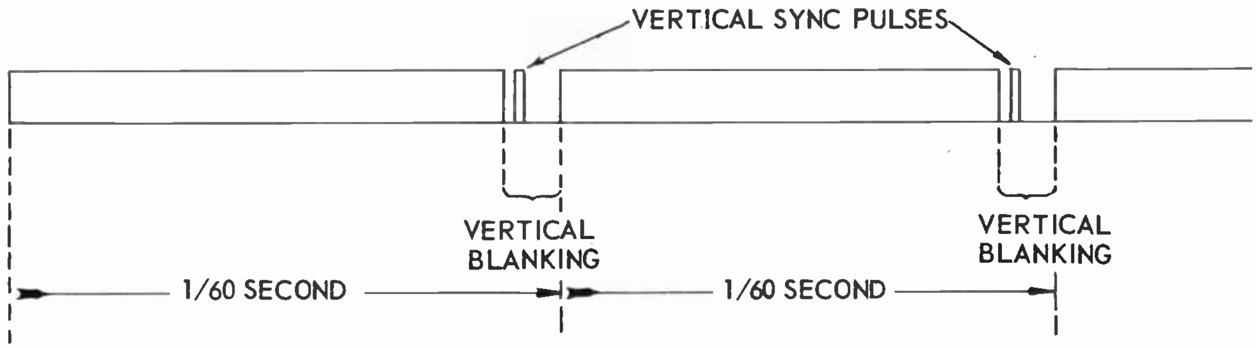


Fig. 3-11. Vertical sync pulses occur 1/60 second apart, and last much longer than a horizontal sync pulse.

before and after this tube. Note also that the signal voltages are inverted in going through this amplifier.

The next tube in the sync section here illustrated is the sync clipper. This tube is operated in such manner as to clip off the remaining picture signal voltages while retaining the sync pulses. When using your oscilloscope to look at the output of the sync clipper in one particular receiver you would see the signal of Fig. 3-10. This is another photograph of an oscilloscope screen.

It often happens that the sync pulses become of unequal strength, some pulses become stronger or weaker than others. So, in Fig. 3-9, we apply these uneven sync pulses to a sync limiter tube. This tube is operated to limit the voltage in its output and thus to bring all the sync pulses to uniform strength or uniform voltage.

The final result of all the operations which have been described is complete separation of sync pulses from picture signals. Sync sections of television receivers may be designed and constructed in many different ways, but the final result always is the same. Sync amplifiers, clippers, and limiters may follow one another in an order different than illustrated. There may be two or more tubes performing the same kind of work. Should the signal from the last of these tubes be wrong side up we may add an inverter tube for the sole purpose of turning the signal back again.

There is no general agreement on the names applied to tubes performing different jobs in the sync section. Various manufacturers use names such as clipper, stripper, limiter, separator, and others to mean different things. The clipper of Fig. 3-9 might be called a separator or a stripper in some receivers. However, from measurements of tube voltages or from observing the input and output signals with the help of an oscilloscope the service technician can tell what a tube really is doing and usually can tell what the tube is supposed to do.

In Fig. 3-9 only the horizontal sync pulses are shown. Vertical pulses and equalizing pulses come right along with horizontal pulses and are treated in the same way throughout the sync section so far as we have examined it. The signal issuing from the output of this sync section would consist of all kinds of pulses.

Horizontal and vertical pulses must be separated from each other. Then horizontal deflections of the picture tube beam may be controlled by the horizontal sync pulses, while vertical deflections are controlled by vertical sync pulses. Fig. 3-11 will help show how the separation becomes possible. Here are represented two time periods of 1/60 second each. During the total of 1/30 second one complete picture is formed on the screen.

Horizontal sync pulses occur at the rate of 15,750 every second. These pulses never cease. They continue all through the pictures, between successive lines, and they continue through all the vertical blanking in-

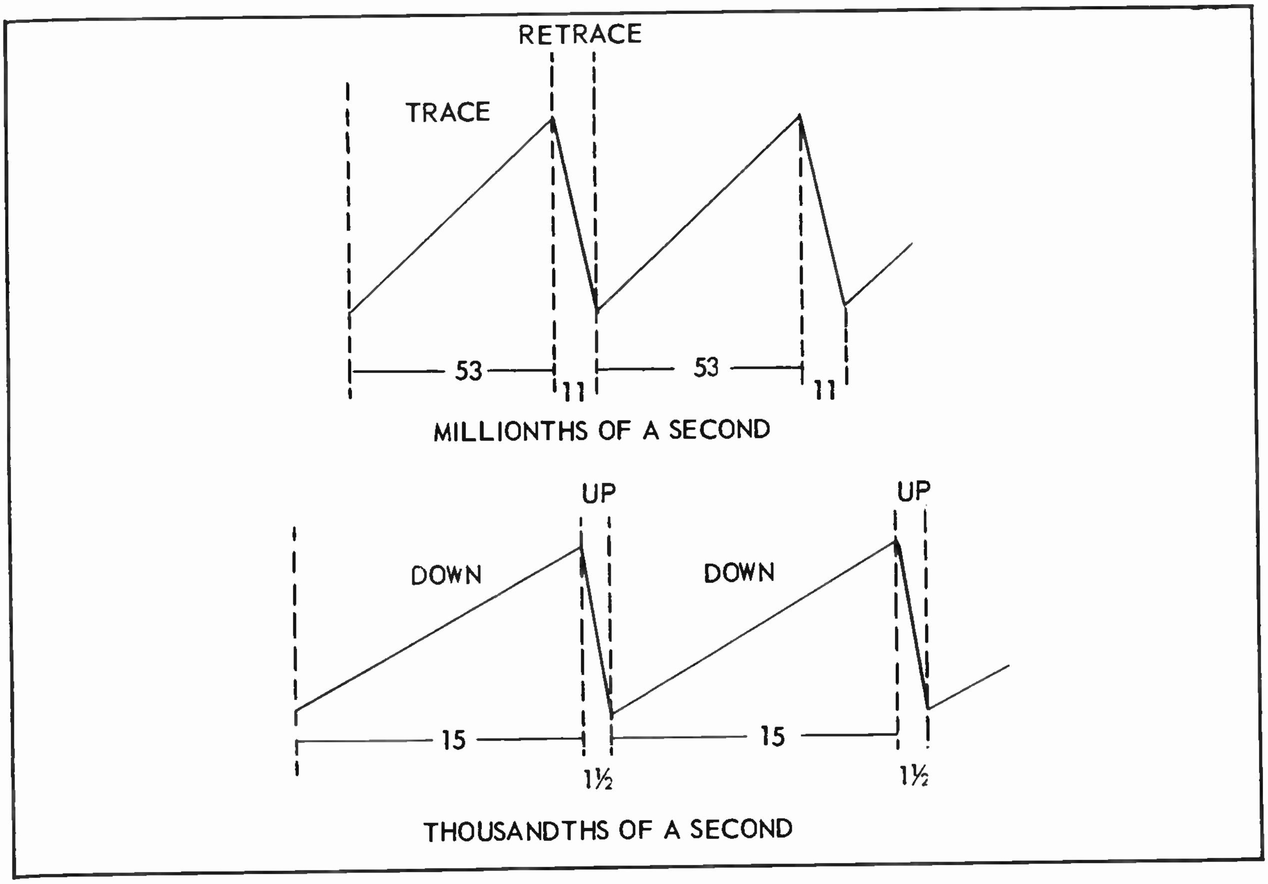


Fig. 3-12. The manner in which deflection voltages or currents rise and fall.

tervals. At the end of each downward travel of the picture tube beam there is a vertical sync pulse. These vertical pulses occur at a rate of only 60 per second. The chief difference between horizontal and vertical sync pulses is in how long each kind lasts. One horizontal sync pulse lasts for about 5 millionths of a second. Each vertical sync pulse consists of six closely spaced sections which last for a total of 185 millionths of a second.

Connected to the output of the last tube in the sync section are two electrical filters. Each filter consists of nothing more complicated than capacitors and resistors. One filter is so designed as to produce from each and every horizontal sync pulse a sharp rise and fall of voltage. These voltage "pips" are used to control the timing of all following actions which cause horizontal deflection of the picture tube beam.

The other filter is so designed that it is not affected by the exceedingly brief horizontal sync pulses. But when a relatively long vertical pulse comes along, the voltage at the output of this vertical filter rises higher and higher. Before the vertical pulse is completed the voltage at the filter output is strong enough to control all following actions which cause vertical deflection of the beam.

In technical literature about television the filter which produces horizontal voltage pips usually is called a differentiating network, and the one providing control for vertical deflection is called an integrating net-

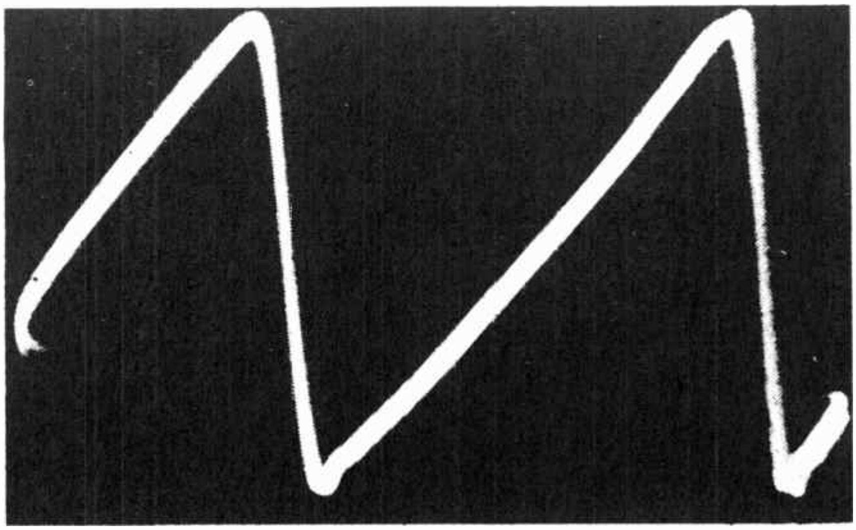


Fig. 3-13. A sawtooth voltage as seen on the screen of the oscilloscope.

work. The filters themselves are so simple that the parts in each would cover little more space crosswide of this page than needed to write their technical names.

THE SWEEP SECTION. Horizontal and vertical synchronizing or timing voltages from the filters which terminate the sync section are not strong enough, nor do they have the kind of voltage changes needed to sweep the electron beam in the picture tube. Consequently, we need in our television receiver two additional sections before it becomes possible to deflect the beam.

The first of these sections will produce voltages having the type of changes needed for deflection. The other section will strengthen those voltages for tubes having electrostatic deflection, or will change the voltages into strong currents or electron flows of the kind needed for tubes having magnetic deflection.

Let's see what kind of voltage changes or current changes are needed for sweeping the electron beam. For horizontal deflection the beam must move from left to right, during one horizontal trace, within a time of about 53 millionths of a second. Then follows a horizontal blanking period of about 11 millionths of a second. From these two time periods we may conclude, as shown at the left in Fig. 3-12, that a voltage for deflecting the beam during a horizontal trace must increase steadily during 53 millionths of a second, then must come back to the starting value in no more than 11 millionths of a second, and must continue such changes.

Now what about vertical deflection? The beam must travel from top to bottom of the picture space in about 15 thousandths of a second, and voltages which control vertical travel must be ready for another downward travel within the next 1½ thousandths of a second. Increase and decrease of vertical deflecting voltage then must be as shown at the right in Fig. 3-12.

Here again we are dealing not with theoretical voltages, but with real ones as they exist in the television receivers which you will be servicing later on. When you connect your oscilloscope to some part carrying

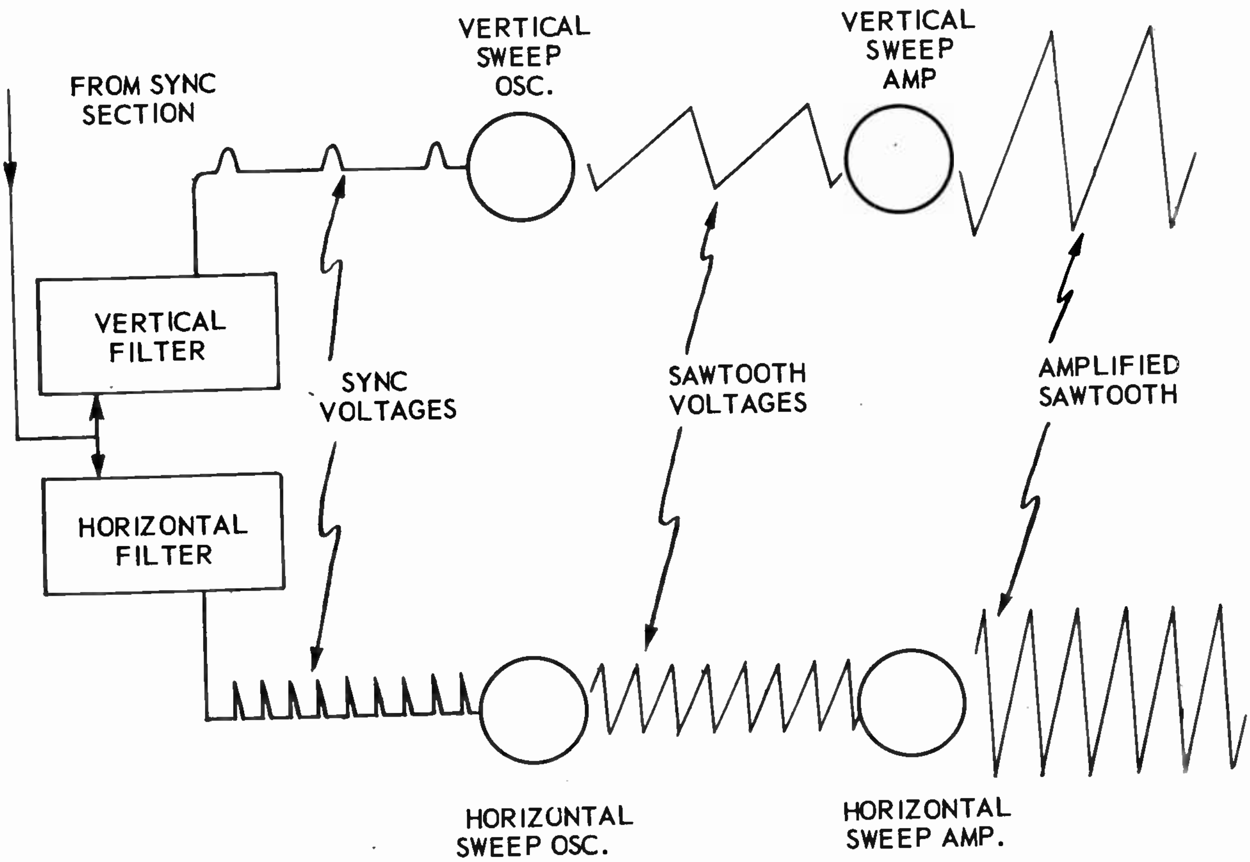


Fig. 3-14. Synchronizing voltages and sawtooth voltages in the sweep section.

one of these deflecting voltages the trace on the screen of the cathode-ray tube will appear about as pictured by Fig. 3-13. This is one more photograph of what you really see.

Are you reminded of anything in particular by the shape of these lines which show changes of deflection voltages? They look like the teeth of a saw. Such deflection voltages are called sawtooth voltages. If our picture tube employs magnetic instead of electrostatic deflection we require magnet currents which go through the same kind of changes, we require sawtooth currents.

No matter which kind of deflection is used, we always must have sawtooth voltages to begin with. For electrostatic deflection the sawtooth voltages are amplified and applied to the deflecting plates in the picture tube. For magnetic deflection the sawtooth voltages are changed to corresponding sawtooth currents, and these currents flow in the deflecting magnets.

Sawtooth voltages are produced by tubes called sweep oscillators. As represented by Fig. 3-14, the synchronizing voltages from the two filters which terminate the sync section are applied to these sweep oscillators, and sawtooth voltages which are in time with the sync pulses appear in the oscillator outputs.

Because of differences between shapes of the synchronizing voltages applied to the sweep oscillators and of the sawtooth voltages obtained from these oscillators it is apparent that the sawtooth voltages are

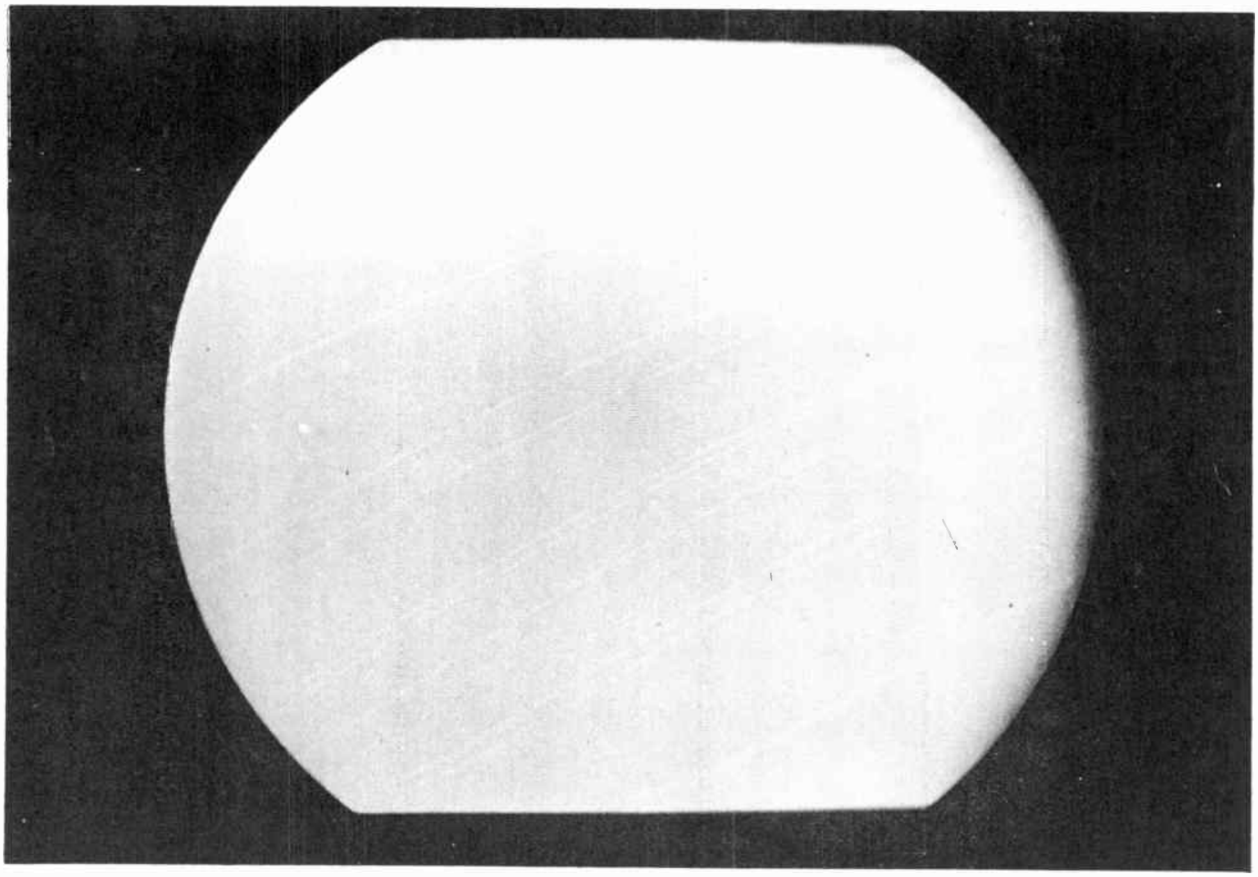


Fig. 3-15. Illumination of the picture tube screen when no pictures are present is called the raster.

not the result of amplifying the synchronizing voltages. Amplifying a voltage does not materially alter its form. The synchronizing voltages do nothing except keep the sawtooth voltages precisely timed with the original sync pulses. Even though you disconnect the sync filters from the sweep oscillators, the sawtooth voltages will continue.

The electrical action of a sweep oscillator is a lot like the mechanical action of the pendulum on a clock. The number of swings per second of an ordinary pendulum depends only on its length, or rather on the distance between the pivot and the center of weight in the bob. The number of times per second that the voltage from a sweep oscillator increases from zero to maximum and returns to zero depends on the electrical size of capacitors, inductors, and resistors connected to the oscillator tube.

You could stand to one side of a clock pendulum and, just before it naturally would start each return swing, you could give it a little push. Thus you could speed up the number of swings per second. You would be doing for the pendulum just what the synchronizing voltages do for the sweep oscillators.

The sweep oscillators are constructed and adjusted so that the number of sawtooth voltages produced

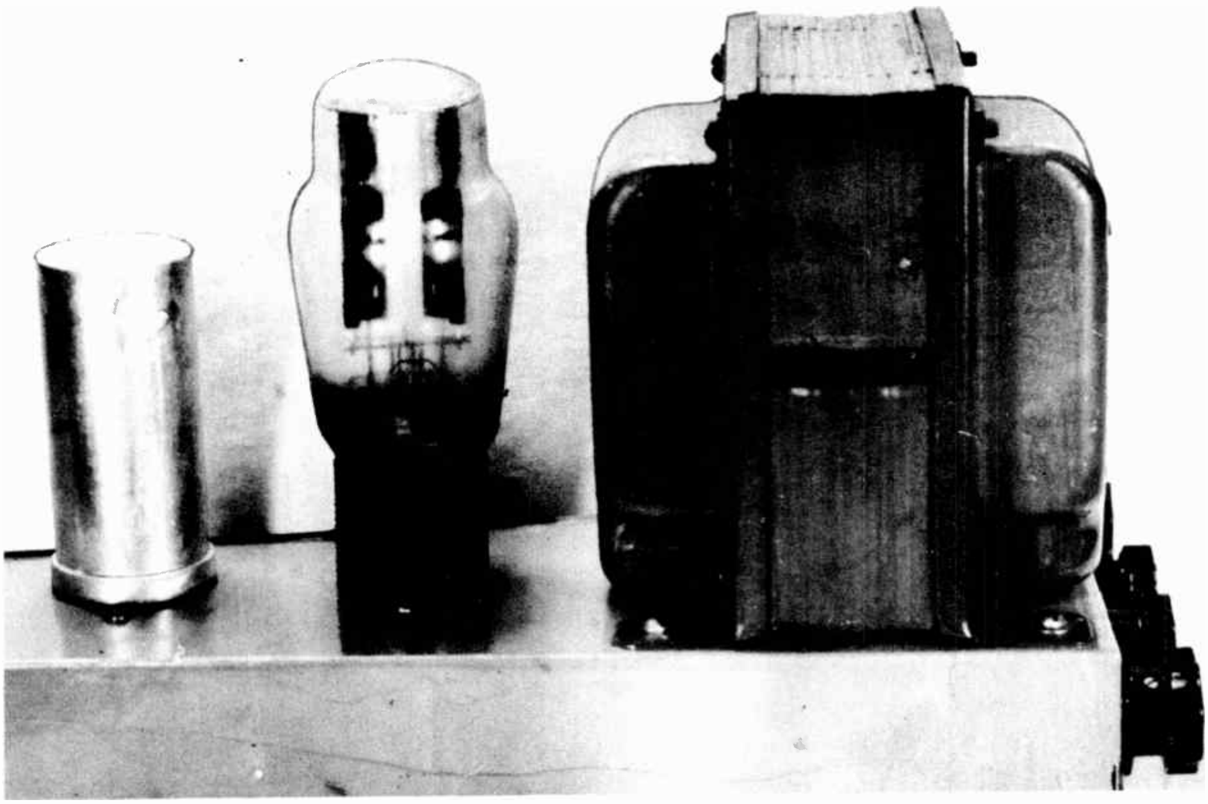


Fig. 3-16. From left to right; the filter capacitors, the rectifier tube, and the transformer for a low-voltage power supply.

naturally during each second is just a little less than the number of sync pulses per second. That is, the oscillators are set to run a little too slow. Then the synchronizing voltages come along and speed up the oscillators so that sawtooth voltages are produced at the correct rate per second, and at precisely the correct instants, to sweep the picture tube beam in time with the electron beam away back in the television camera tube.

The sweep oscillators will deflect the electron beam of the picture tube horizontally and vertically even though no television signal is being received, and there are no sync pulses. The clock pendulum will continue swinging even though you don't push it. If the electron flow in the beam is sufficient to produce a spot of light, the screen will be illuminated over the entire area which would be occupied by pictures were any picture signals present. Such illumination of the picture tube screen is called the raster. A raster is shown by Fig. 3-15.

We have considered only the simplest possible sweep section, because we are dealing with only the most elementary principles of television reception. Later we shall take up all the interesting variations found in the many makes and types of receivers. Merely as a hint of what is to come, there are more than a dozen

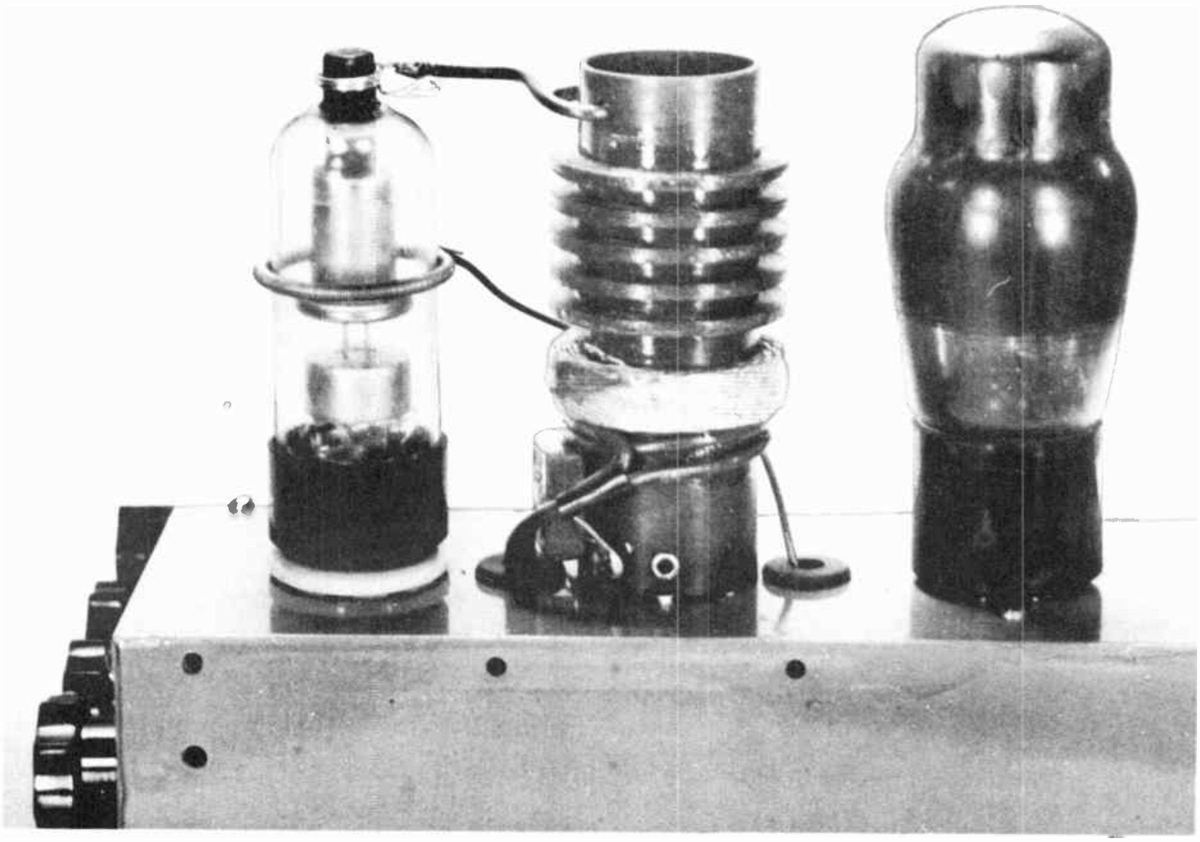


Fig. 3-17. Principal parts of a high-voltage power supply. From left to right; the rectifier tube, the transformer, and the oscillator tube.

ways of insuring that sweep voltages and currents remain precisely in time with the incoming sync pulses. These are called automatic frequency controls for the horizontal and vertical sweep oscillators.

POWER SUPPLIES. Earlier it was mentioned that amplifier tubes control electric power furnished by other tubes and other parts of the receiver. Not only the amplifier tubes, but all other tubes in every section of the receiver require power for their operation. These operating powers must be supplied at potentials all the way from five or six volts for parts which heat tube cathodes, through several hundred volts for tube plates and screens, and to many thousands of volts for acceleration of beam electrons in the picture tube. These voltages are furnished by sections called the power supplies.

Most television receivers have at least two separate power supply sections, a high-voltage power supply for the electron-accelerating elements in the picture tube, and a low-voltage power supply for everything else. The principal parts of a low-voltage power supply for one receiver are pictured by Fig. 3-16. At the right is the power transformer. This transformer takes alternating-current electric power from the building lighting circuits at whatever potential exists in these circuits, usually 110 to 120 volts. The transformer lowers the voltage to values suitable for heating tube cathodes, and raises the voltage to values suitable for tube plates and screens.

In the center of Fig. 3-16 is a power rectifier tube. This tube changes the alternating voltages for tube plates, screens, and some other parts into voltages which act always in the same direction. That is, the rectifier tube changes the alternating voltages into direct voltages which are needed for operation of tubes in the receiver.

At the left in Fig. 3-16 is a metal can containing a number of filter capacitors. These capacitors, in conjunction with other parts of the power supply not shown, smooth out the direct voltages. As these voltages come from the rectifier tube they always act in one direction, but they are not at all steady. Smoothing is done by the filter capacitors in the metal can and by filter inductors which do not show in the picture.

Fig. 3-17 is a picture of the principal parts in one kind of high-voltage power supply, the supply which provides electron-accelerating voltages for the picture tube. At the right is a high-frequency oscillator tube. This tube takes direct voltage and current from the low-voltage supply section and produces alternating voltages at frequencies of 200 kilocycles, and even higher in some designs.

These high-frequency voltages from the oscillator are put through the transformer which you can see at the center of Fig. 3-17. This transformer increases the potential of the high-frequency voltage to about 5,000 volts. Then the high-frequency voltage at its high potential is applied to the high-voltage rectifier tube which you see at the left. This rectifier changes the high-frequency voltage, which is an alternating voltage, into a direct potential at equally high voltage. Finally the very high direct voltage goes through a smoothing filter consisting of capacitors and resistors, not visible in the picture, and the steady high voltage is delivered to the picture tube. This type of high-voltage supply is found in many television receivers, but also in common use are other types which will be examined later on.

Maybe it has struck you as rather strange that we enter the power supply with alternating voltage from the building circuits, then change it to a direct voltage in the low-voltage power supply, next back to alternating voltage in the oscillator of the high-voltage supply, and again back to direct voltage at the output from this latter power supply.

The reason for all the changing back and forth is that the simplest and most economical way of either raising or lowering any voltage is to have that voltage in alternating form and to put it through a transformer. To raise or lower the value of direct voltage is either a complicated process or is very wasteful of power. The best way is to use an oscillator when we wish to change direct voltage to alternating, to use a rectifier when we wish to change alternating voltage to direct, and to use a transformer for either raising or lowering the potential of an alternating voltage.

We have followed the television signal all the way from the receiving antenna to the picture tube and the loudspeaker. We commenced with electromagnetic radiation coming through space to the antenna. We finished with music, voice, and other sounds from the speaker, and with moving images on the screen of the picture tube. Dozens of parts have been shown and described. If you go back through these first lessons and list all the parts of a television receiver you will realize that a service technician has to know a great deal about a great many things in order to shoot trouble and handle servicing in general.

Although we already are better acquainted with the processes of television reception than are most people not actively engaged in this business, we know but little about how all the parts really operate, or why they operate, or why they might fail to operate. Learning all about the how and why of television won't be so very difficult if we go about it in the right way. There is one key which will unlock most of the mysteries about normal and abnormal behavior of television and radio apparatus. This key is an understanding of electrons and of how and why they act as they do. This will be our subject for the next lesson.

TELEVISION

LESSON NO. 4 ELECTRONS, THE FOUNDATION OF RADIO

"Training Your Mind"

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind
C.T.I.
Perfecting Your Skill

COMMERCIAL
TRADES
INSTITUTE

LESSON NO. 4

ELECTRONS, THE FOUNDATION OF RADIO

When men learned to control the flow of electrons in wires we entered the age of electricity, which gave us the telephone and telegraph, electric light, electric power, and everything these things do for us. When men learned how to force electrons out into the open, and still control the flow, we entered the age of electronics. Electronics gave us radio in all its applications, including television and countless uses in commerce and industry.

Everything in radio and all its branches depends directly or indirectly on the controlled flow of electrons through vacuums and through gases. Possibly we should have commenced our study of television by examining electrons and what they accomplish. Had we done so, many actions which could be mentioned in the first lessons only as accomplished facts, with no apparent reasons for their existence, would have appeared as the only things which possibly could happen under the circumstances.

The fact that we commenced studying the practice of television rather than its foundation may make it even more interesting to examine the foundation, which consists of electrons. To get acquainted with electrons we shall first consider a piece of aluminum. By various mechanical treatments the aluminum could be divided into particles so small that a powerful microscope would be needed to see them. By chemical action the aluminum could be divided into particles not visible even through the microscope. But none of these divisions can break the aluminum down into any other substances. The smallest possible particle is an atom of aluminum, still a particle of the original silvery white metal.

Atoms of all substances are of about the same size, and this size is small. Had every man, woman, and child commenced at the beginning of the Christian era to drop one atom per second into a cup, and had all who have lived since then done the same thing, the cup today would contain one-third of a cubic inch of atoms. If each man, woman and child in the United States and all the rest of the countries in North, Central, and South America were to lay down one atom, with all the atoms close together and in a straight line, that line would be somewhat less than one and one-quarter inches long.

Although we cannot conceive of anything so small as an atom, we now must accept the fact that every atom is mostly empty space. An atom is made up in much the same way as our solar system. At the center of the solar system is the sun, and in empty space around the sun are whirling the earth and all the other planets. Were any microscope strong enough to make an atom visible you might see something such as shown by Fig. 4-2. At the center of the atom is a part called the nucleus. Moving in orbits around the nucleus are electrons. Were an atom to swell up to an overall diameter of 1,000 feet, the nucleus at the center would be the size of a basketball, and each electron would be the size of a pea.

In each atom of aluminum there would be 13 electrons. Could you take away one electron, and could the atom adjust itself to the permanent loss of one electron, it no longer would be an atom of aluminum but

Fig. 4-1 The F-m signal generator, the oscilloscope, and the receiver being aligned all depend for their operation on the flow of electrons through vacuums.

would have changed to an atom of the metal magnesium. With still another electron taken away, and another readjustment, we should have an atom of the metal sodium. With electrons removed, and nine remaining, there would be an atom of the gas neon. By the time only two electrons remained there would be an atom of the gas called helium. Finally, with one solitary electron around a nucleus you would have an atom of the gas hydrogen.

Electrons can be added to atoms as well as removed. Our sun shines and gives off heat because its atoms of hydrogen are changing to atoms of helium and gaining one electron in the change. Matter of which the sun is composed is disappearing in this process at the rate of about five million tons per second. But don't let this worry you. Less than one-tenth of the sun's hydrogen has changed to helium in the last billion years.

Electrons move at tremendous speeds in the space around the center of an atom. But the atom holds together. It holds together for the same reason that our solar system holds together, with the earth and all the other planets continuing to revolve around the sun instead of flying away into limitless space. It is the force of attraction between sun and earth that holds the earth on its orbit around the immensely heavier sun. It is the same kind of attractive force between you and the earth which holds you in place, so you don't shoot off into space as the earth's surface spins around at about 1,000 miles an hour. This earthly force is called gravity. The kind of force having similar effect in the atom is called electrostatic attraction between the nucleus and the electrons.

With strong attractive forces at work why doesn't our earth fall into the sun? And why don't the electrons fall into the nucleus of the atom? We stay out of the sun because of another kind of force, centrifugal force. This is the kind of force which you feel as a pull when whirling a weight tied to the end of a string. Cen-

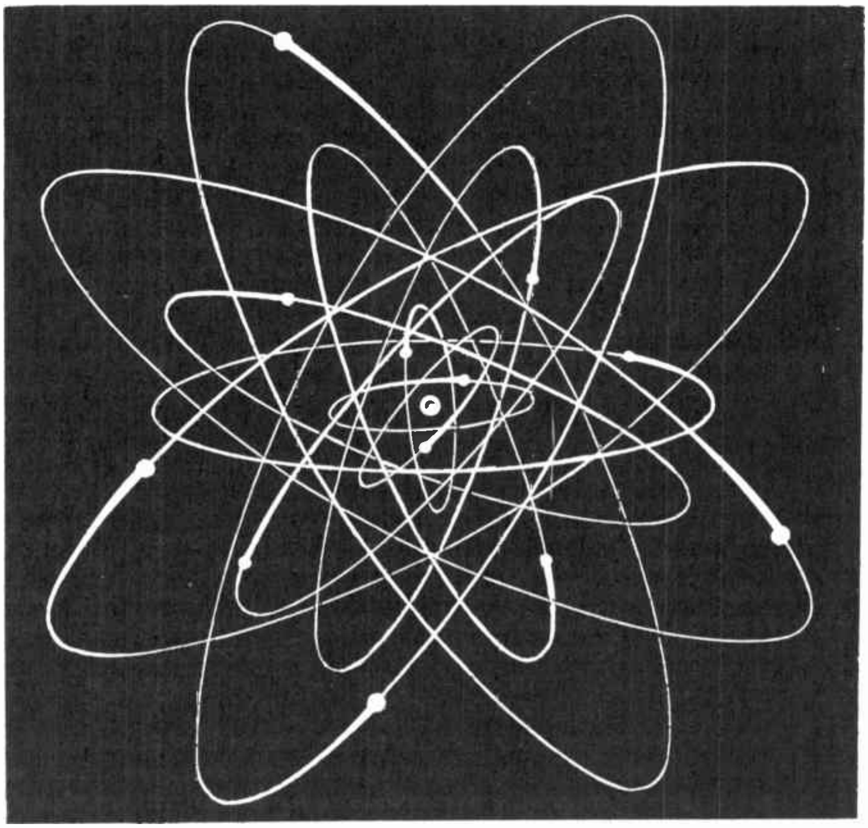


Fig. 4-2 Electrons move in orbits around the central nucleus of an atom.

trifugal force alone would carry the earth away into space. Attractive force alone would pull us into the sun. But the effects of the two forces are exactly balanced, so we neither fall into the sun nor move farther away. Electrons don't fall into the nucleus of the atom because they are moving at speeds which would enable them to fly right out of the atom were they not held by the equal force of attraction.

The attractive force existing between bodies such as the sun and the planets is of the same kind in both bodies. It is the result of the weight or the mass of the materials in these bodies. In the atom, however, the kind of attractive force existing in the nucleus is not the same as the kind existing in the electrons. We know these two forces are not of the same kind because, in some ways, they act differently. The nucleus attracts electrons, and the electrons have attraction for the nucleus. But the nucleus of one atom does not attract the nucleus of another atom, they repel each other. And one electron does not attract another electron, the two electrons repel each other. It is the repulsion of each electron for every other electron that keeps them from colliding with one another inside the atoms.

To distinguish between the two forces which act differently we call the kind possessed by the nucleus a positive force, and call the kind possessed by electrons a negative force.

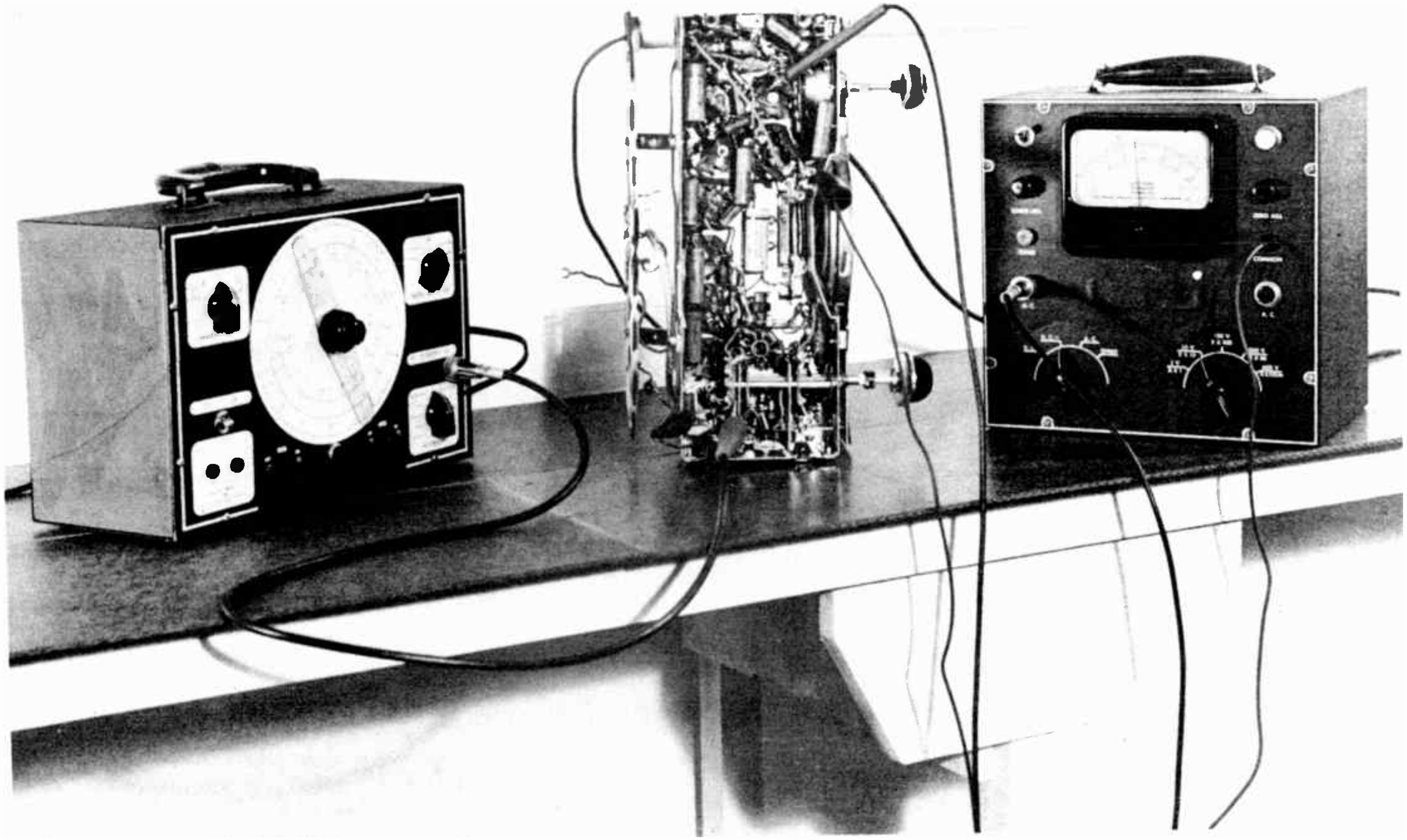


Fig. 4-3 With a signal generator and electronic voltmeter used for receiver alignment we still depend on flow of electrons through vacuums for operation of all three instruments.

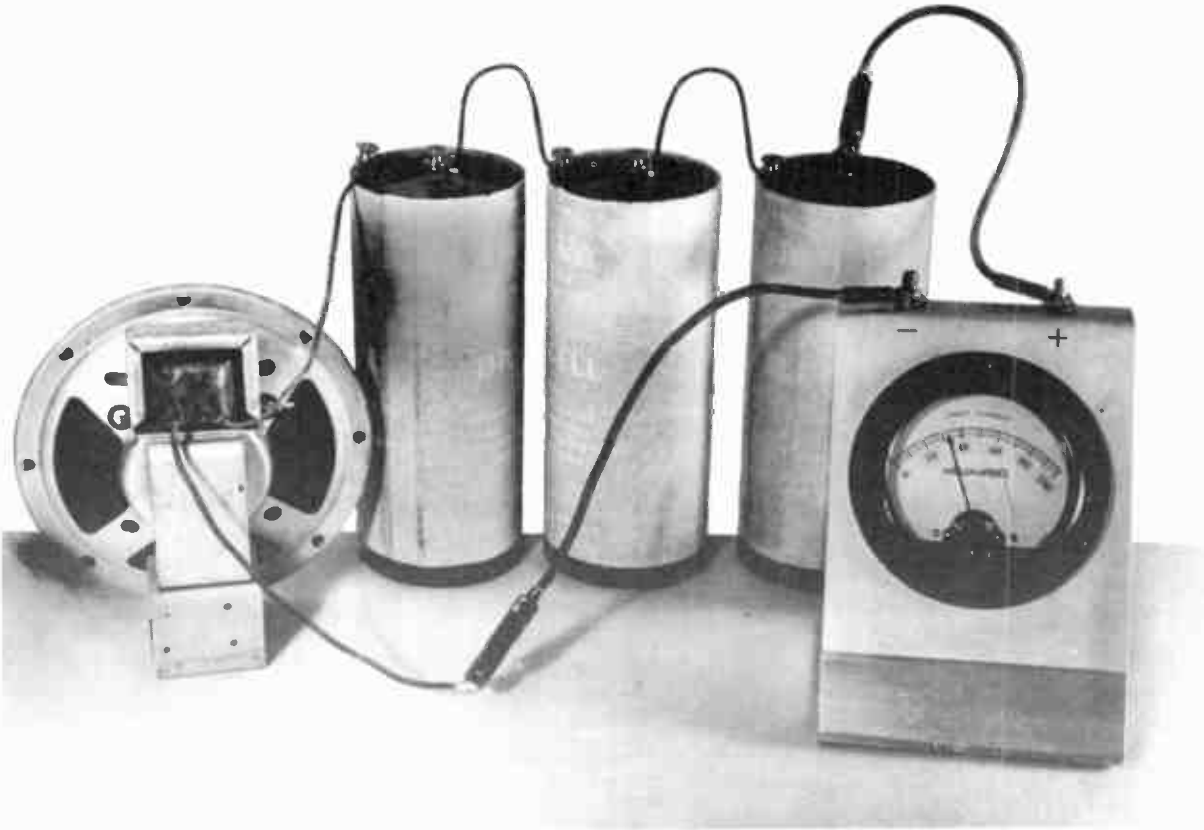


Fig. 4-4 The dry cells force electrons to flow in themselves, the speaker transformer, and the meter – while the meter measures and indicates the rate of flow.

By using the words positive and negative we may describe the forces in the atom as follows. When one particle is negative and another is positive there is attraction, each for the other. When both particles are negative they repel each other. When both particles are positive they repel each other.

While an atom has its normal number of electrons, as 13 electrons in the case of the aluminum atom, the total negative force of all the electrons is exactly equalled by the total positive force in the nucleus. This means simply that all the electrons are pulling just as hard on the nucleus as the nucleus is pulling on all the electrons. The two kinds of forces are balanced in the normal atom. There is no excess of either kind of force which might act on some other particle outside the atom. So far as the atom as a whole is concerned it can exert neither positive nor negative force on anything outside its own boundaries. This, remember, is the condition of an atom which possesses its normal quota of electrons.

Although the electrons within an atom always are in violent motion, the nucleus of the atom does not move unless the substance of which the atom is a part undergoes movement. If you move our piece of aluminum you move all the atoms which compose the piece, but the atoms remain in fixed positions within

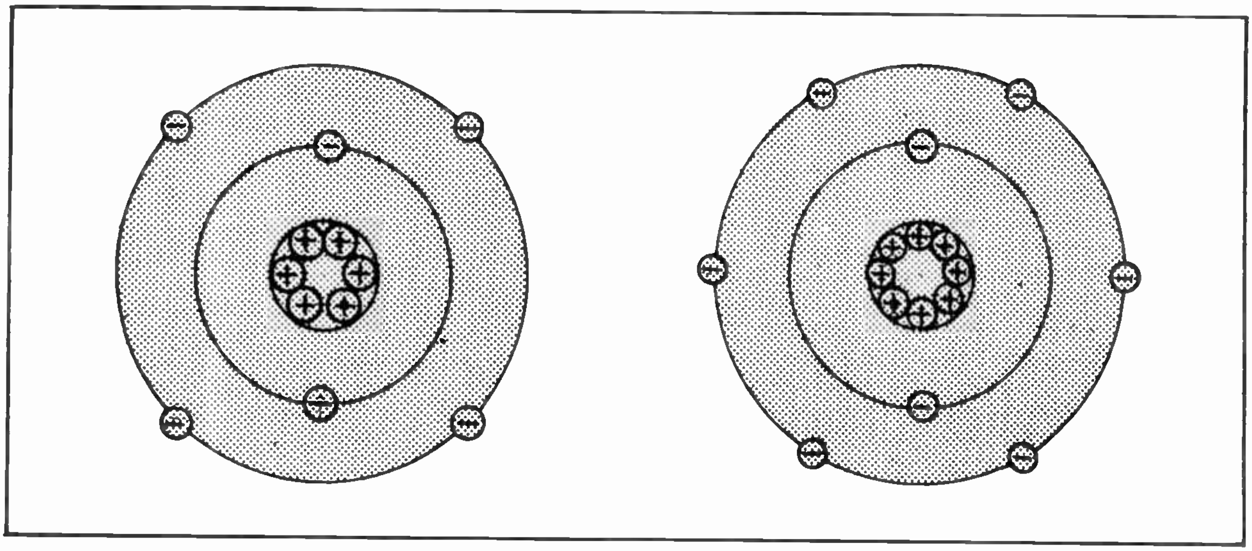


Fig. 4-5 How electrons are arranged in two orbits within atoms of carbon and of oxygen.

the piece of aluminum. While the nucleus of an atom is either stationary or moving, the attractive forces within each atom hold the electrons with the nucleus, because even the lightest nucleus is nearly 2,000 times as heavy as one electron.

In diagrams it is convenient to identify positive forces with a plus sign (+) and to identify negative forces with a minus sign (–). Using these signs we might represent atoms as in Fig. 4-5. Each electron is shown by a small circle enclosing a minus or negative sign. The central nucleus is shown to have a total positive force equal to the total negative force of all the electrons by placing within a central circle a number of smaller circles enclosing positive or plus signs, with the number of these units of positive force equal to the number of negative electrons. Incidentally, the atom with six electrons would be an atom of carbon, and the one with eight electrons would be an atom of the gas oxygen.

FREE ELECTRONS. Now we shall leave the atoms for a few minutes and talk again about the solar system which on a grand scale, is so much like the atom on a submicroscopic scale. Consider what might happen to a planet away out at the farthest limits of the solar system. The farthest planet we know of today is called Pluto. The distance from the sun to Pluto is more than 3,500 million miles. Our earth is only about 93 million miles from the sun. Naturally, the force of attraction acting through the great distance between the sun and Pluto cannot exert nearly so strong a pull as between the sun and the earth.

Supposing now that some other heavenly body of size comparable to that of the sun should come close to our solar system. Pluto would be quite likely to feel the attraction of that other body so strongly as to leave our solar system and join another one.

This is just what happens continually to one of the outermost electrons in an atom. Between the outermost electron and the nucleus of the atom the attractive forces act much more weakly than between the nucleus and electrons closer to it. All around are other atoms whose positive nucleuses might pull hard on the electron which is flying around the outside of its own atom. Under certain conditions this outermost

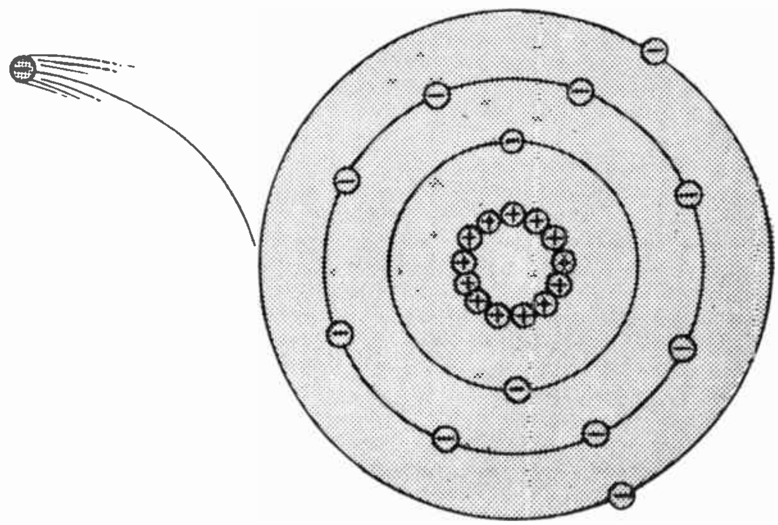


Fig. 4-6 One of the outermost electrons escapes from the aluminum atom and becomes a free electron.

electron may jump from the original atom to another nearby atom.

Let's see what these certain conditions must be. A few paragraphs back it was stated that the positive and negative forces in a normal atom are balanced. In a balanced atom the total positive force of the nucleus is just equal to the total negative force of all the electrons, and there is no force remaining to attract any other electron which might have freed itself from some other atom.

Supposing, however, that this normal or balanced atom were to lose one of its own electrons. Then this atom no longer would have balanced forces, it would possess a surplus of positive force. Thus surplus of positive force would pull the wandering electron into the atom, and the balance would be re-established.

Now we'll go back to Pluto. Supposing something were to happen which imparted much greater speed to this distant planet in its path around the sun. The increase of centrifugal force might completely overcome the weak attraction between Pluto and the sun, with the result that Pluto would start traveling outward away from the sun. However, Pluto wouldn't get very far, as interplanetary distances go, because the universe is full of other big suns. One of these other suns soon would make a captive of the wandering planet, and Pluto would join some other solar system.

We do not know of anything which might give Pluto additional speed, but we do know of several ways to increase the speed of electrons which are near the outer limits of their atoms. One way is to heat the substance containing the atoms. When a substance is made very hot, all the electrons in its atoms gain energy. Energy is the ability to do work. The electrons expend their extra working ability in increasing their speed. The added speed lets many of the electrons move farther from the central nucleus, and it lets one of the outermost electrons jump clear out of the atom. But, like Pluto, this electron which has become a free

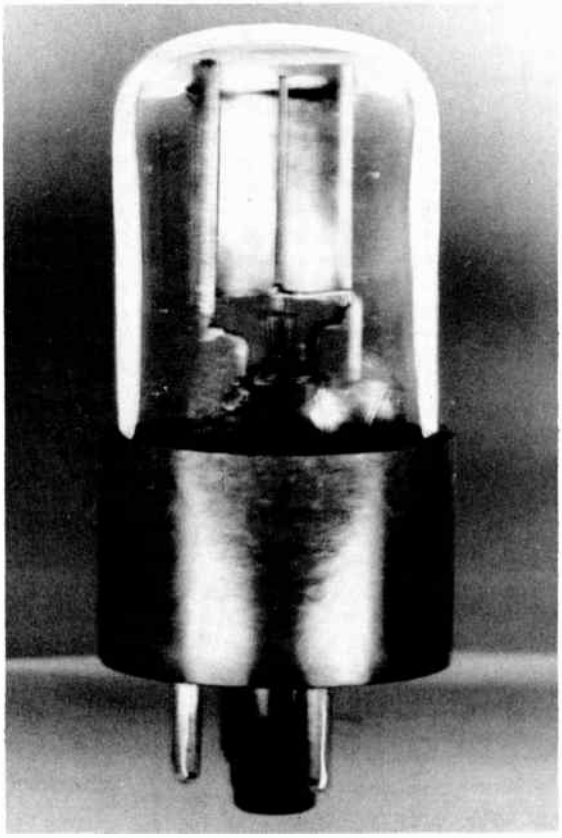


Fig. 4-7 Light entering this phototube gives cathode electrons extra energy, and the electrons emerge into the vacuum.

electron won't get very far. It immediately is pulled into some other nearby atom which has lost one of its own electrons and is looking for a replacement. Fig. 4-6 represents one of the outermost electrons escaping from an atom of aluminum.

Here we have discovered the real reason why heating the cathode of a picture tube forces electrons out of the cathode material. It is the reason also why the cathodes of practically all other tubes in radio and television receivers are heated, as you may see by looking at the red glow inside every tube which is working. Heat gives some electrons enough additional energy to enable them not only to leave their individual atoms, but to jump entirely clear of all atoms and to emerge from the surface of a cathode material. These electrons will exist for a moment in the surrounding space. If nothing else happens, these electrons almost immediately are pulled back into the cathode material to join atoms which lack an electron.

Heat isn't the only force which gives electrons the ability to jump free and clear of atoms. Light will do it. Light, like heat, is a form of energy. When the energy which is light hits certain substances this energy changes into the kind of energy which gives more speed to some electrons. This is the energy of motion, the kind of working ability which enables a baseball to break a window when aimed in the wrong direction. It is light energy which gives us phototubes and photocells, and makes possible the television camera tube. Electrons which have been speeded up by light will jump right out of the surface of some kinds of substances.

There is still another force which gives electrons jumping ability. This is the force we call electric voltage or electric potential. If you put two bodies close together in a vacuum, and make the potential of one body very much greater than that of the other body, electrons will be pulled out of one and will pass through the vacuum to the other body. This is the action employed in what are called cold-cathode rectifier tubes and in tubes employing what is called ionic heating of their cathodes.

It doesn't require much imagination to realize that, if we get electrons out of a cathode by means of heat, a part which pulls the electrons to itself and clear away from the cathode needn't be nearly so close to the cathode as when voltage of this other part is the sole means for freeing the electrons from atoms. Here we have the basis for all thermionic tubes, a classification which takes in just about every kind of tube used in radio and television. By heating the cathode to get electrons into the clear, and by having a difference of voltage or potential between the cathode and another part inside the tube, free electrons are pulled all the way through the vacuum which is inside the tube. We wouldn't have to follow this line of investigation very far to be right in the middle of the whole subject of tubes for radio and television, but there are few more of the elementary things to be talked about before we come to tubes.

Let's finish first with the subject of free electrons. Each atom in any substance normally possesses just the number of electrons which allows balance of positive and negative forces. For instance, in any piece of aluminum there are normally 13 electrons per atom. Were there something like a billion, billion atoms there would normally be 13 times a billion, billion electrons. Lots of these electrons are continually popping out of some atoms and right back into other atoms.

With all this movement of electrons into and out of atoms there is, in every substance at every instant, about one free electron per atom. Unless we are applying some external force these free electrons don't really get anywhere, they simply jump about from one atom to another in all kinds of random directions. This random motion of free electrons is somewhat as shown by Fig. 4-8.

If we apply the right kind of force in the right way we can make all the free electrons move in the same general direction as they progress from atom to atom. In a length of wire the free electrons may be made to move either always to the right or always to the left. Then we have a direct electron flow or a direct current. The free electrons may be made to move first to the right and then to the left, and made to keep this up. Then we have an alternating electron flow or an alternating current. If the free electrons are made to move right and left at very short intervals we have a high-frequency alternating current.

Even with a steady direct current or a one-way current in a length of wire no one electron need go very far at a time. Rarely, if ever, would a particular electron travel all the way from one end to the other of the wire. But all the free electrons would move the same direction between the instant of escape from one atom and the instant of entering another atom.

EXPERIMENTING WITH ELECTRONS. You may experiment with free electrons and move them from one body *to* another at any time you choose. All that is needed in the way of equipment is a stick of sealing wax from the stationery store, a piece of woolen cloth from anywhere, and a few bits of paper.

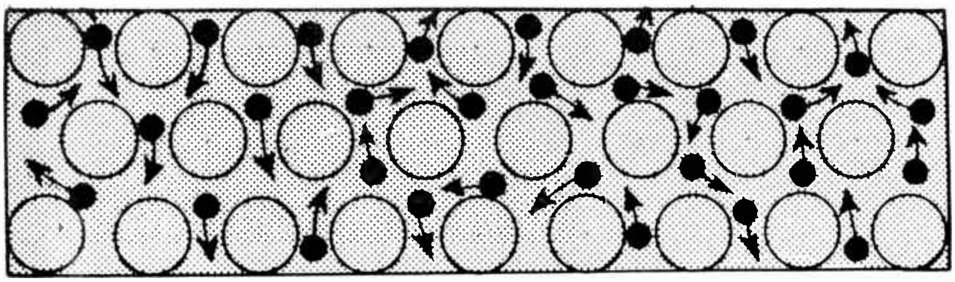


Fig. 4-8 When not affected by some external force the free electrons move in random directions from atom to atom.

First, rub the end of the wax with the woolen cloth. Then, as in Fig. 4-9, bring the end of the wax near a few small pieces of paper. Even before the wax touches the paper, the paper will jump up and cling to the wax. This is what happened. By rubbing the wax and wool together you actually rubbed many billions of free electrons off the wool and onto the wax. Since all these electrons are negative the wax accumulated a great excess of negative force. When you brought the wax near the bits of paper the strong negative force of the wax repelled free negative electrons which were in the side of the paper toward the wax, and drove these electrons to the other side of the piece of paper. Moving these negative electrons away from one part of the paper left that part with an excess of positive force. Then the paper was drawn to the wax by attraction between the negative force of the wax and positive force on one side of the paper.

You may restore an equilibrium of forces in the wax by wiping it with your moist hand. The excess of negative electrons on the wax then passes to your hand, and the wax no longer will pick up pieces of paper.

In the atoms of substances from which paper is made the atoms hold onto their electrons very tenaciously. The negative force transferred to the wax by rubbing it with the wool is a strong force, and it is capable of forcing electrons from one side of the paper to the other in spite of all difficulties. Once the free negative electrons are driven to the far side of the paper they stay there, because they are held tightly by the atoms and would require a considerable force in the opposite direction to drive them back again.

In metals the atoms do not hold their outermost electrons at all securely. It is easy for electrons to escape from atoms of metals and wander all around the place. If you use some force which drives free electrons from one side to the other of a piece of metal, then take away that force, the electrons will come right back again and distribute themselves uniformly throughout the metal.

We may check this latter statement by performing another experiment. This time a piece of aluminum foil or any thin metal foil is hung from a thread, as in Fig. 4-10. Rubbing the sealing wax with the woolen cloth and bringing the end of the wax near the foil allows just the same action as described for the wax and the paper, and the foil comes over and touches the wax.

Because it is so easy for electrons to pass all through any kind of metal, much of the excess of free negative electrons now leaves the wax and passes into the foil. Then the foil has more than its normal complement of negative electrons. If the foil is aluminum it now has more than 13 electrons per atom. This means that the whole piece of foil possesses a surplus of negative force. The wax does not lose all its extra negative electrons to the foil, and so now we have two bodies with each a surplus of negative force. The

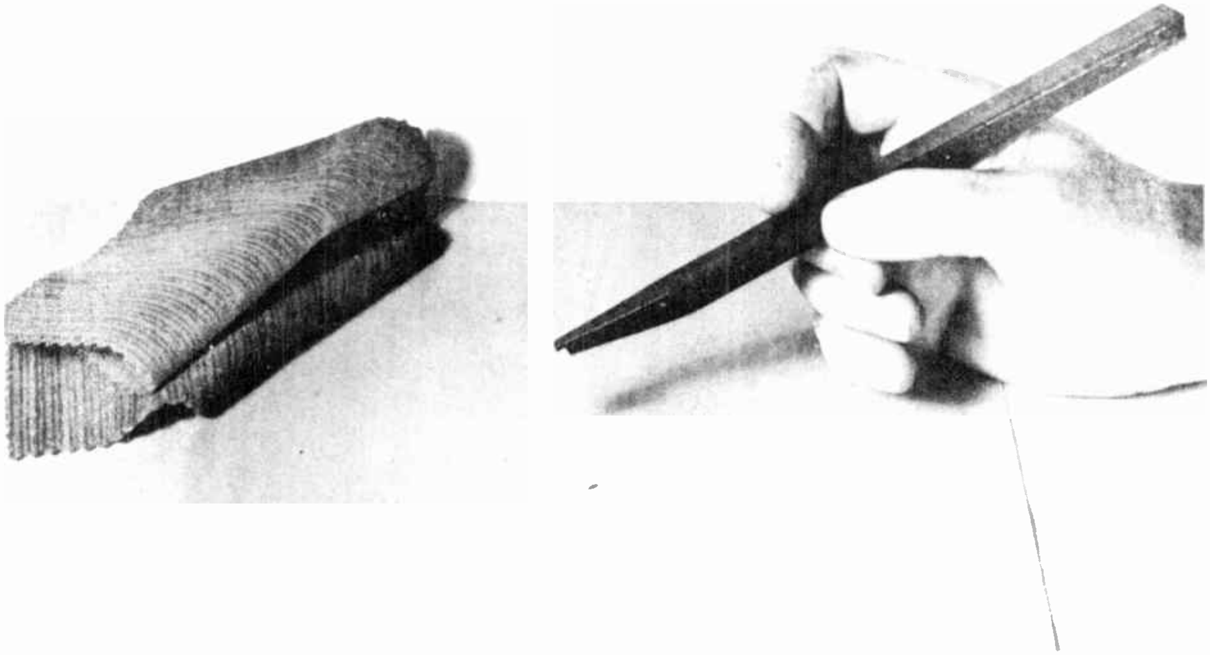


Fig. 4-9 When extra electrons are rubbed off the cloth onto the wax their attractive force allows picking up the paper.

result is shown by Fig. 4-11, the wax and the foil repel each other.

Experiments such as these are easy to carry out, and if you wish to see electrons at work the experiments are well worth performing. If you wipe the wax on your hand and touch the foil with your finger, to restore the balance of forces, you may start over again. First the foil will come over and touch the wax, and almost instantly will jump away from the wax. The foil must be supported by a thread because thread is composed of substances generally similar to those in paper, and electrons cannot easily move from one part to another in such substances. If the foil were supported by a piece of metal wire, in which electrons pass from place to place very easily, all the excess negative electrons passing from the wax to the foil would run up through the wire and the supports. Then the concentration of negative electrons would not be sufficient to cause attraction and repulsion.

CONDUCTORS AND INSULATORS. Any substance in which free electrons get from place to place very easily is called an electrical conductor. A conductor is composed of any substance whose atoms do not hold very tenaciously to their electrons, and in which a great movement of free electrons may be caused by a fairly weak external force of any kind which attracts or repels electrons. When voltages or differences of electrical potential are applied to a conductor there are movements of immense quantities of free electrons in the conductor.

The metal foil of our last experiment is a conductor. All reasonably pure metals are excellent conductors. Silver is the best of all conductors, because a greater rate of electron flow or a greater current is caused in silver than in any other metal by a given voltage applied to this metal. Next in conductive excellence

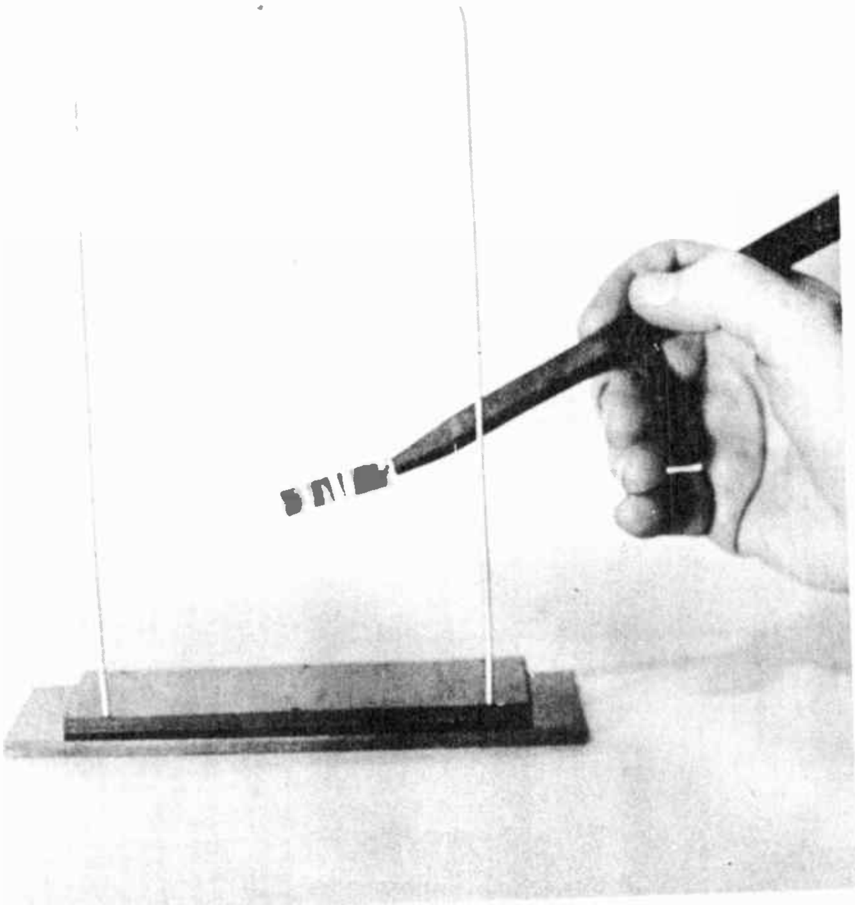


Fig. 4-10 The attractive force of electrons on the wax pulls a piece of metal foil against the wax.

comes copper, which is used for nearly all the wires between parts of radio and television apparatus. Aluminum, brass, bronze, iron, steel, zinc, lead and cadmium are good conductors. All are used in radio and television as supports and fastenings through which electron flows are to take place.

Any substance whose atoms hang onto their electrons very tightly is an electrical insulator. In an insulating material strong forces are needed to cause any appreciable rate of flow of free electrons. The paper of our first experiment and the supporting thread of the second experiment are insulators. It was possible to force free electrons from one part to another in the paper because we really had a very strong force in the rubbed wax. It is easy, by means of rubbing, to build up a force which would be comparable in strength to that in power wires for electric lighting when measured in volts. But you couldn't feel this force because there isn't enough of it. All the electrons you could rub onto a piece of wax wouldn't furnish even enough total energy to make a flash lamp blink, yet the potential or voltage of the relatively few electrons might

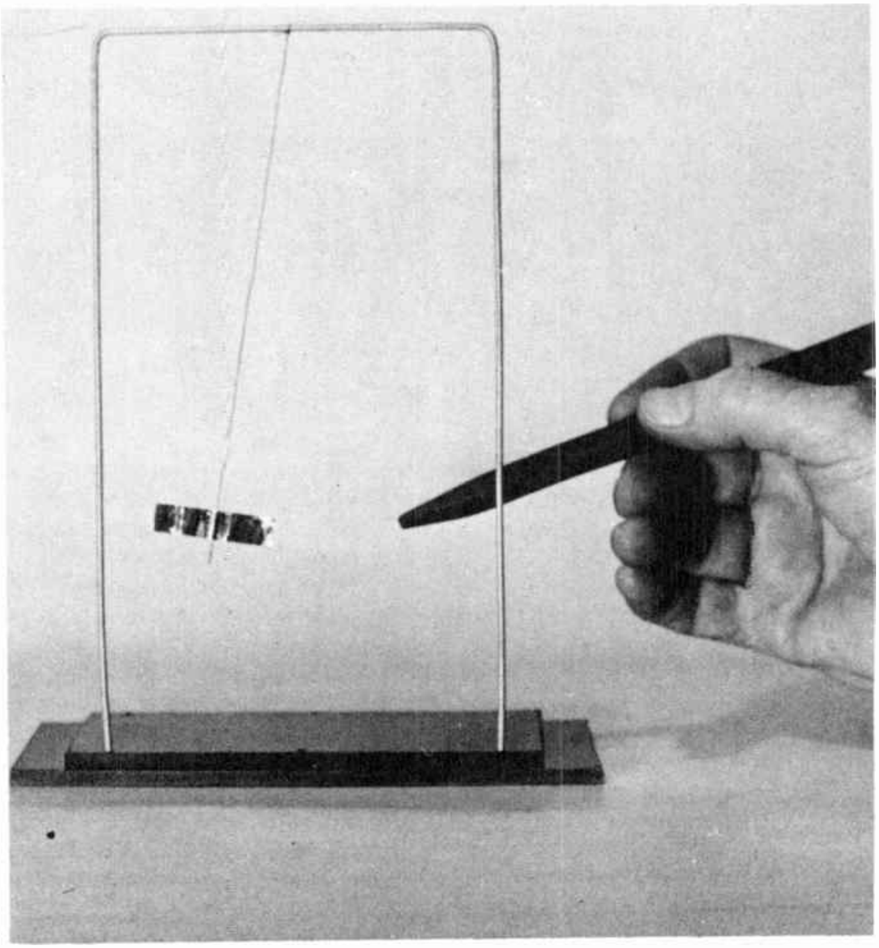


Fig. 4-11 When electrons flow from the wax into the metal foil both bodies have excess quantities of electrons, and there is repulsion.

be very high.

Among the insulating materials most generally used in radio and television are the many plastic compounds, of which Bakelite was the first and still probably is the best known example. Well known among newer plastic insulating materials are polystyrene and vinylite. In the old standbys for radio insulation we find rubber, paper, linen, cotton, many ceramic materials, mica, fibre, and various waxes and varnishes. The prime purpose of all insulation is to confine free electrons within the conductors where the electrons are supposed to move. Of equal importance in many places is the ability of good insulation to protect us from electric shocks which might be received from conductors in which electrons are being moved by high potentials or voltages.

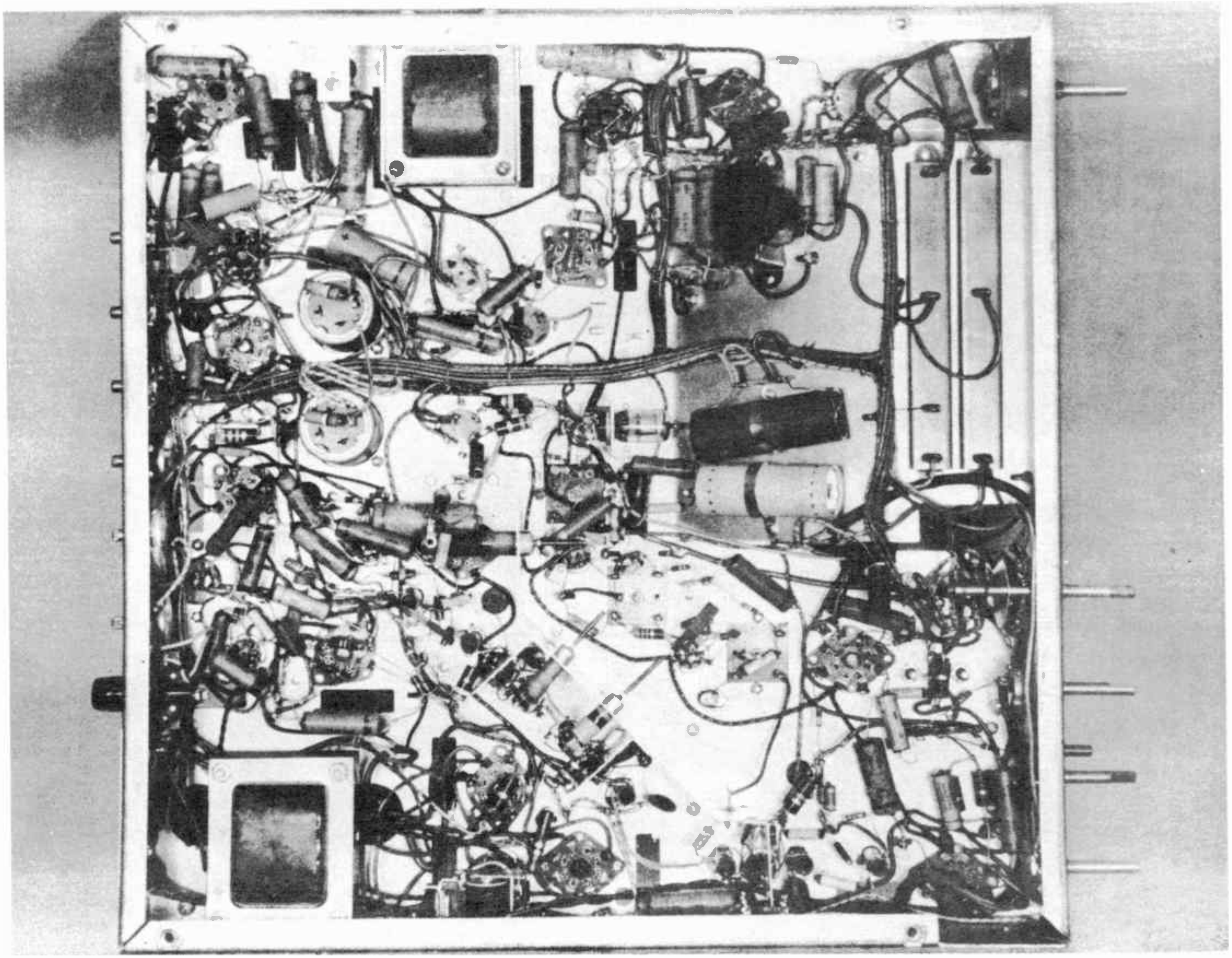


Fig. 4-12 A service technician knows how to trace electron flows in all these parts, and knows how to measure the results of such flows.

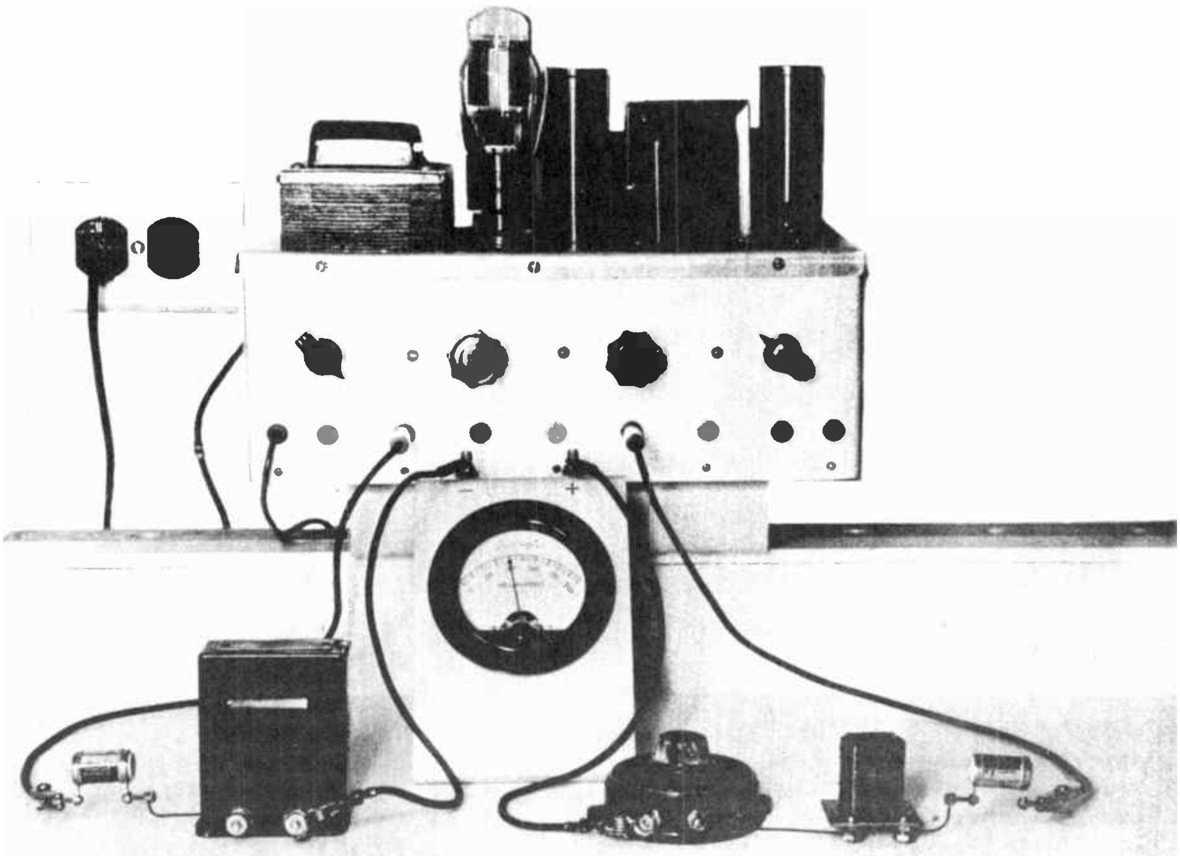


Fig. 4-13 The power supply unit up above is forcing free electrons to flow in one direction through all the other parts while the meter measures the rate of flow in milliamperes.

ELECTRIC CHARGES. When any piece of material of any kind has more than a normal number of free electrons that piece of material possesses an extra amount of negative force. The body which thus has an excess of negative electrons is said to have a negative charge, or to be negatively charged. The sealing wax in our first experiment became negatively charged.

Any body which has less than a normal number of electrons possesses an extra amount of positive force That body is said to have a positive charge, or to be positively charged. The woolen cloth which lost free electrons to the wax was left with a positive charge.

When a body is neither negatively nor positively charged, but has its normal number of free electrons, that body is said to be neutral or may be said to be uncharged. If a body which originally is neutral

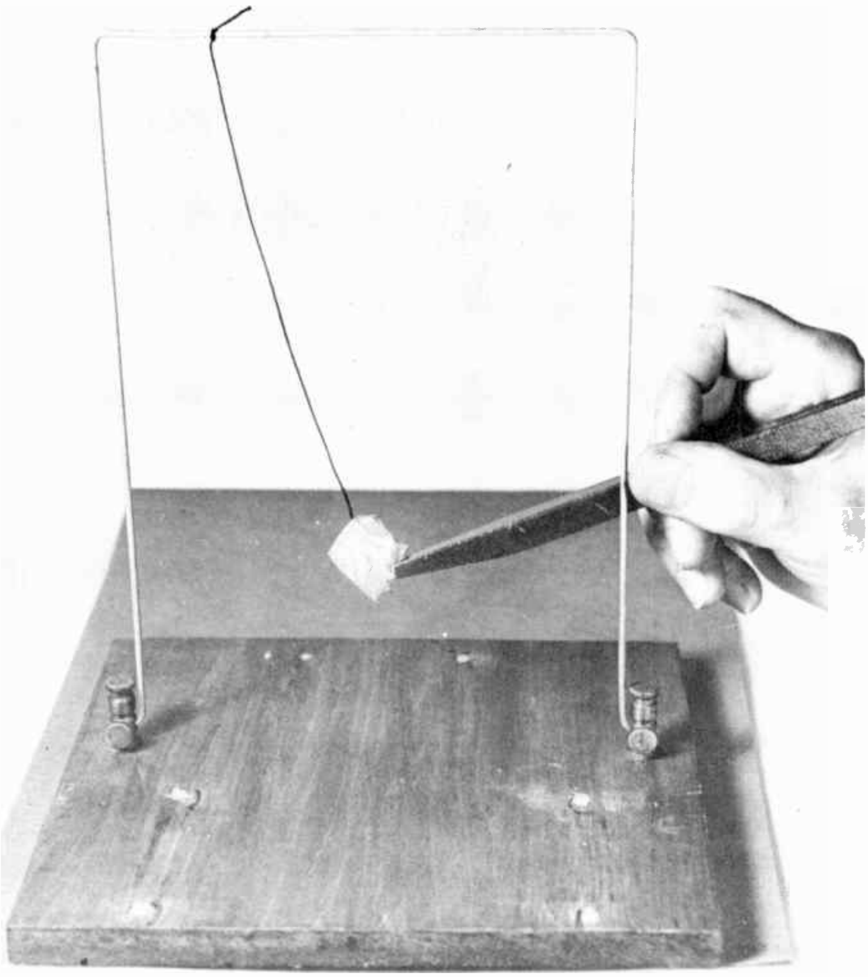


Fig. 4-14 Once more the negatively charged wax attracts the metal foil.

receives extra electrons that body is given a negative charge. If the neutral body loses some of its negative electrons it then has a positive charge.

You have just read the definitions of a negative charge, a positive charge, and a neutral body. The definitions are so exceedingly simple in the reading that you cannot possibly appreciate the importance of these three conditions in all our future work. There is nothing we can do now to bring home a realization of the importance of electric charges in radio and television, we shall just have to wait until their importance becomes apparent in all practical work.

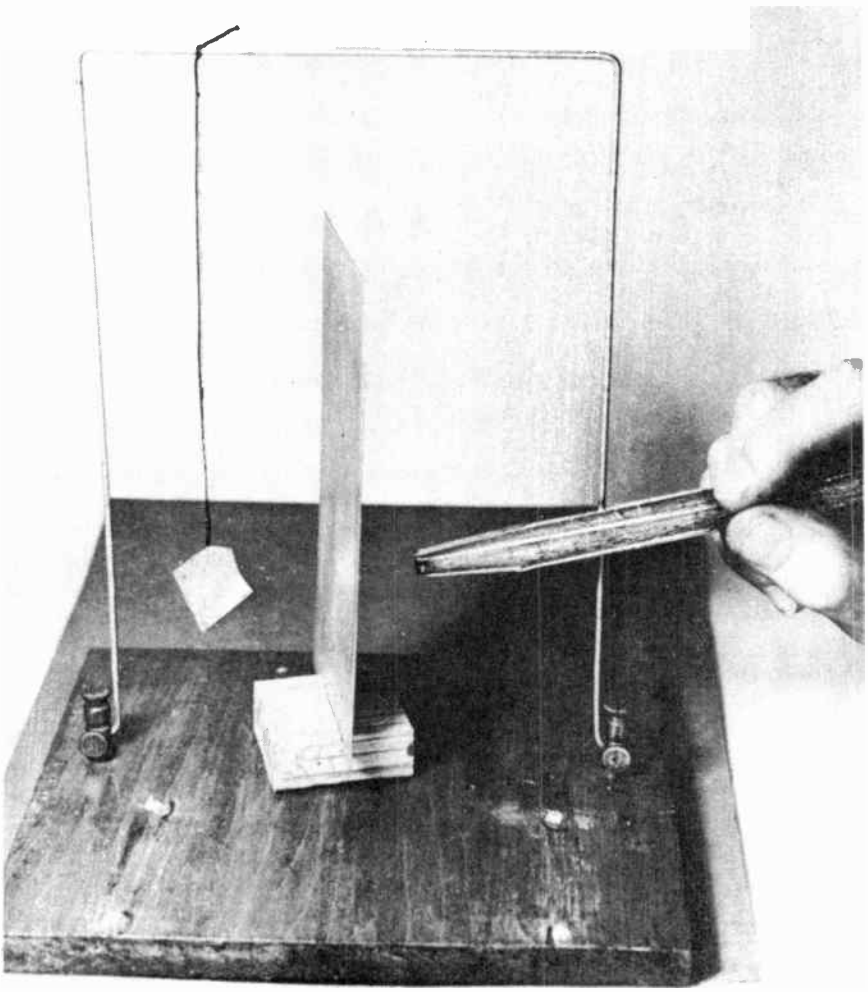


Fig. 4-15 With a metal shield between the wax and the foil the force from the negatively charged wax cannot have any effect on the foil.

As you might expect, there is a unit in which we measure the quantity of electric charges. This unit is called the coulomb. A charge of one coulomb means a quantity of 6,280,000,000,000,000,000 free electrons. If some certain body has 6,280,000,000,000,000,000 fewer free electrons than it would have when neutral, that body has a positive charge of one coulomb. If the body has this number of free electrons in excess of what it would have when neutral, the body has a negative charge of one coulomb.

Just as there is attraction between a positive nucleus and a negative electron in a single atom, so there is attraction between any body which is positively charged and any other which is negatively charged. And just as there is repulsion between one negative electron and another negative electron, so there is repulsion between any two bodies when both are negatively charged. Naturally, there is repulsion also between any two bodies when both are positively charged.

Don't spend any time on trying to memorize the definition of a coulomb, it may never be mentioned again. The coulomb is important chiefly because it affords a sort of stepping stone to an understanding of another unit in which we measure the rate of electron flow in conductors. When free electrons move past a point at the rate of one coulomb per second this rate of flow is called one ampere.

The first thing to be noted about a rate of flow is that it is not a measure of electron speed. If enough free electrons are moving in one direction, one coulomb of electrons can pass a certain point during one second even though the speed is very slow. If you have a river wide enough and deep enough to carry great quantities of water, the rate of flow past a given point might be millions of gallons per second with the speed of the water only a few miles per hour. If the stream of water were a narrow, shallow brook, the speed of the water would have to be very great to get millions of gallons past a given point during one second. Similarly, if there are but few free electrons available, a rate of flow of one ampere (one coulomb per second) might require an impossibly high speed.

The second thing to be noted about a rate of flow in amperes (coulombs per second) is that it is not a measure of voltage or potential or force. A rate of flow of water, in gallons per minute or cubic feet per second, does not tell about the force which is causing the water to flow. The force pulling a certain rate of water flow down Niagara Falls is very great, but the force causing an equal rate of flow along some quiet stretch of the river below the falls is relatively very small. Similarly with an electron flow rate measured in amperes. In a poor conductor, where electrons are held by their atoms rather firmly, it might take a considerable force or considerable voltage to cause some certain flow in amperes. In silver or copper the same rate of flow could result from a relatively small voltage.

Rates of electron flow may be measured in amperes whether the flow is always in one direction (direct current) or reverses in direction (alternating current). Even though the frequency of an alternating current were so high that electrons moved only a small fraction of an inch in each direction, their rate of flow past the middle of that fraction of an inch could be measured in amperes.

We have talked so much about amperes as a measure of the rate of flow of electrons or electricity that we may as well get acquainted with two other units for measuring rates of flow. There are very few places in radio and television receivers where the rate of electron flow is so great as one ampere. Most currents are measured in milliamperes. One milliampere is equal to 1/1000 of an ampere. Currents in many parts of the picture tube, and in some other places as well, are so small that we measure them in microamperes. One microampere is equal to 1/1,000,000 (one millionth) of an ampere.

In this lesson we have covered more of the road to a complete understanding of radio and television than you will realize until we commence putting all this information to use. Yet we hardly have begun to investigate all that electrons do. As an example, when parts of a receiver are being rapidly charged, first negatively and then positively, forces radiated from these parts may completely upset the performance of other parts around them. These forces may be confined to where they originate by using electric or electrostatic shielding.

Shielding consists of metal partitions or enclosures. Its effect might be illustrated with our stick of sealing wax to represent some part producing a force to be confined, and a piece of metal foil representing another part to be protected. In Fig. 4-14 the foil has been attracted, as usual, by the force which extends all around the charged wax. In Fig. 4-15 there has been interposed between wax and foil a thin sheet of metal, a shield. No matter how hard you rub the wax with the woolen cloth, the force from the charged wax is stopped dead by the shield and there is not the slightest tendency to attract the foil. It will be some time before we examine the practical applications of shielding in receivers and testing instruments, but here we have seen it in action.

One of our subjects for the next lesson will be these forces which extend from all charged bodies, and which account for electric potential or voltage and all that it does in tubes and other parts of receivers.

TELEVISION

LESSON NO. 5 HOW ELECTRONS DO THEIR WORK

"Training Your Mind"

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

COMMERCIAL TRADES INSTITUTE

Training Your Mind

C.T.I.

Perfecting Your Skill

LESSON NO. 5

HOW ELECTRONS DO THEIR WORK

Men who once get into radio, and now television, seldom leave these fields. Whether or not they own up to it, the real reason for sticking is the absorbing interest in watching seemingly impossible things happen right before your eyes. Good technicians make good money, but they might make good money in other fields. The difference is this: In television and radio we have fun while making money. Every case of trouble shooting is a puzzle to be solved. It is a challenge to your ability. And there is nothing more pleasant than the feeling of having solved a difficult puzzle.

In earlier lessons we dealt with things so interesting that you didn't have to study very hard in order to learn. Also, we have been taking our time. Now some time must be saved. There is so much ground to be covered on actual service operations in lessons to come that we cannot afford to take our time getting there. Consequently, in this lesson, we are going to cover a lot of important ground – and do it fast.

First, read this lesson all the way through without trying to learn a thing. Some facts will stick anyway. Then read it again, and try to comprehend the meaning of each statement while you read it. Don't try to memorize; we are not learning rhymes to be recited in front of a class. If you understand a fact while reading about it, you never can wholly forget it. When you need that fact later on it will come popping out of the place in your mind where it has been stored all the while.

Finally, if you feel that you understand the general meanings of all the facts presented, you have done a good job. Everything mentioned here will come up so many times in actual service work that, eventually, you will learn it anyway. The general understandings which you will gain from this lesson simply make the rest of the road easier and quicker to travel.

Now let's see whether you can take it, whether you can take a few straight facts without sugar coating.

HOW ELECTRONS DO THEIR WORK

In Fig. 5-1 we are looking at a small fraction of the parts and their connections in a television transmitter. This is a modulator power supply. Fig. 5-2 is a picture of all the wiring and a good share of the parts in a simple radio receiver. Both the big transmitter and the small receiver operate because free electrons move in all the conductors and all the circuit parts.

When a technician looks at parts and wiring he thinks in terms of electron flow. He sees, in his mind's eye, the electrons sometimes moving steadily in one direction, again moving back and forth at perhaps millions of times a second. He knows why the electrons should move in certain places and in certain ways, and he knows what probably has gone wrong when tests and measurements show that electrons are failing to flow in the right directions, in the right places, and at correct rates.

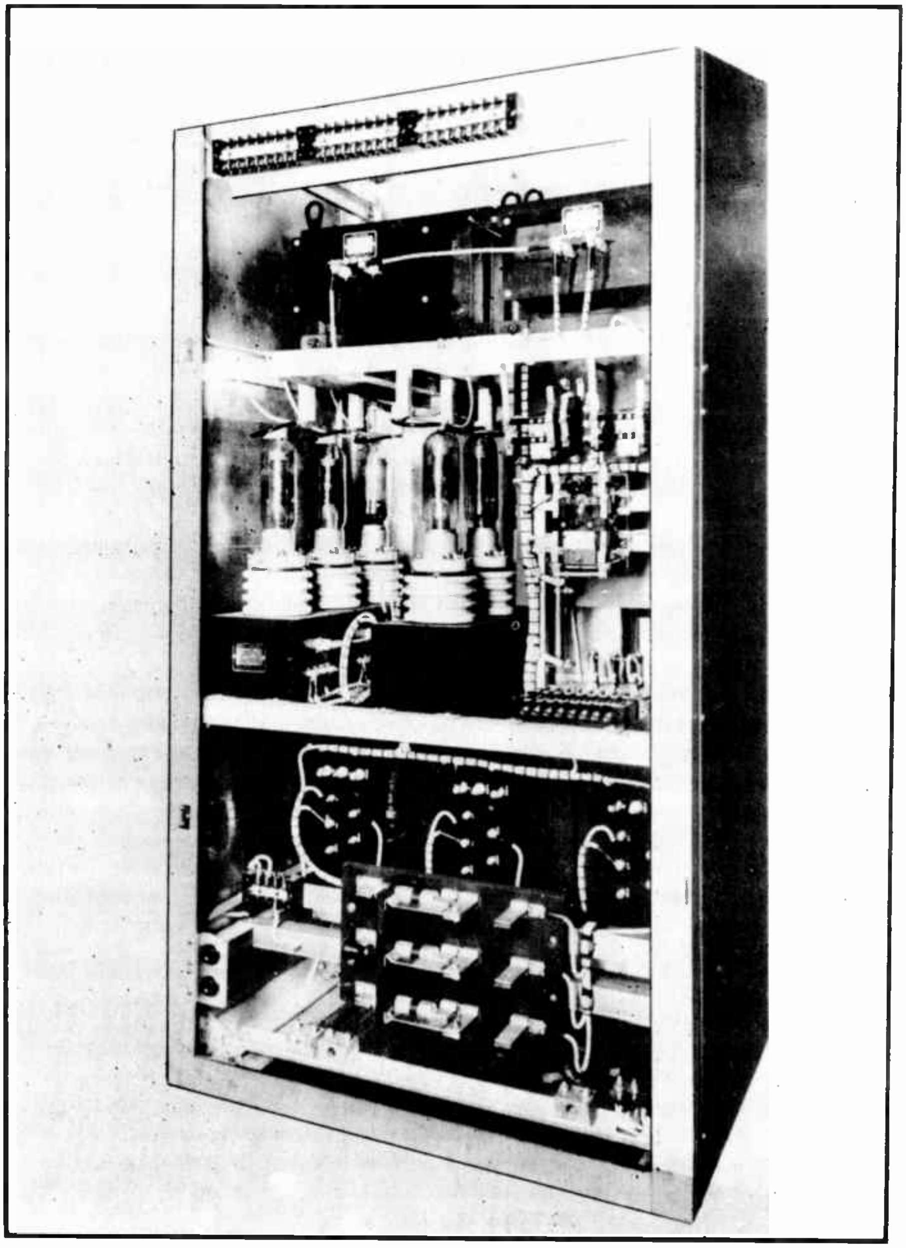


Fig. 5-1. Heavy-duty rectifiers in the modulator power supply of a television transmitter.

Our next big subject is to be that of radio tubes; how they work and why, and how to use them to best advantage for certain jobs. We cannot talk intelligently about the behavior of tubes without knowing quite a

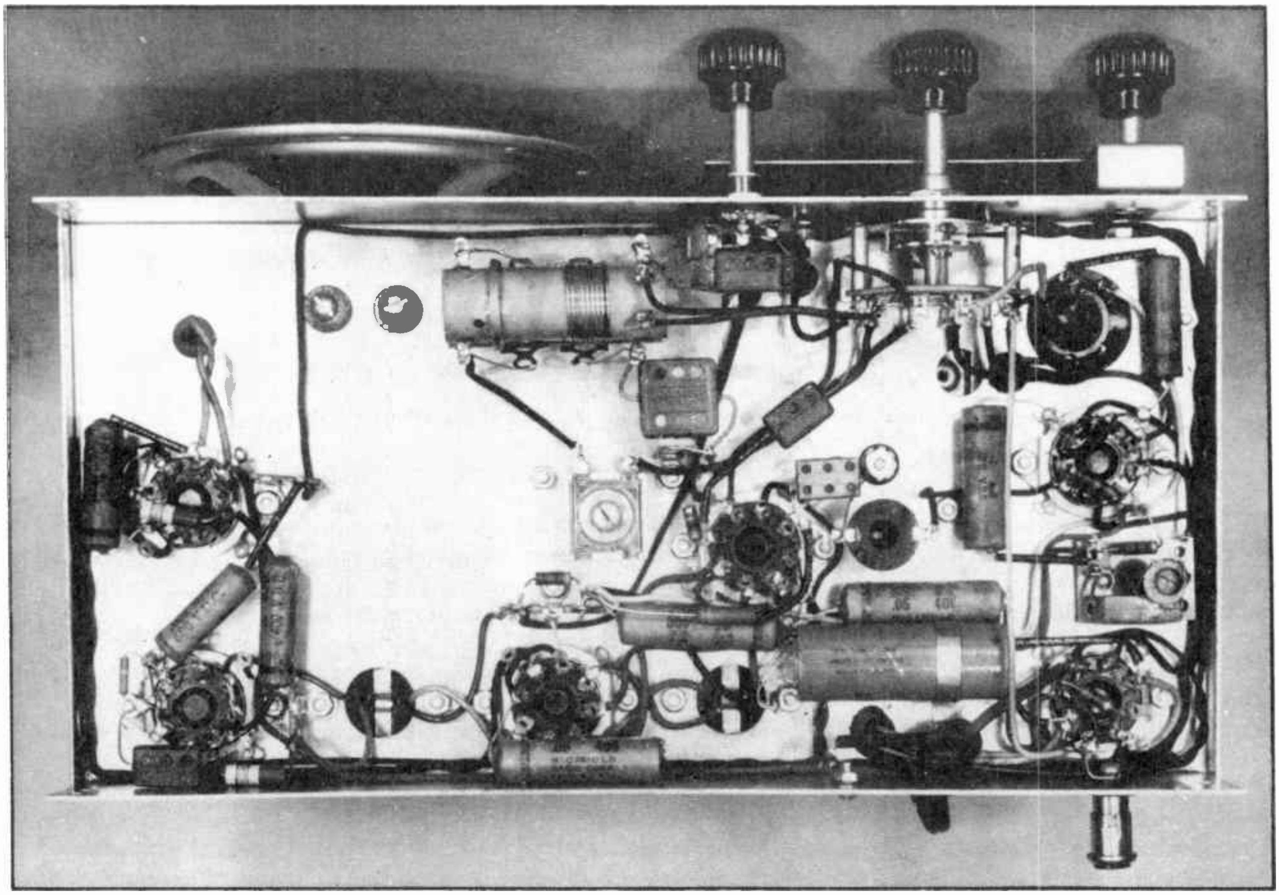


Fig. 5-2. Parts and wiring underneath the chassis of a six-tube radio receiver.

bit about amperes, volts, and ohms. We do know the real meaning of an ampere of electron flow. We have talked about volts, but never have said what a volt really means. And as for ohms, they haven't even been touched on.

To really understand volts and ohms we must commence with some very simple and elementary facts relating to work and energy. These facts will be understood easily enough by imagining that we do some work on a single free electron, knowing that what happens to this one electron is happening at the same time to billions of its fellows.

To begin with, what do we mean by work? Maybe you think you work when you hold a 50-pound weight above your head for an hour, but that is not work. If, however, the weight originally were on the ground, and you lifted it off the ground, you would have done work during the time the lifting continued. You then worked because you employed your muscular force to overcome the force of gravity which tends to hold the weight down, and you moved the weight to a higher position against this force of gravity. Work, in the technical sense, is done only when one force acts against another force and causes motion or a change in the rate of motion.

Fig. 5-3. Work is done when electrons are driven contrary to the forces of attraction and repulsion.

Now about the free electron. We intend to follow it through a power supply and a conductor connected between the negative and positive ends or terminals of the power supply. The simplest of all power supplies is a battery, so we shall talk about a battery for easy understanding. What happens inside and outside the battery will happen in a generally similar fashion inside and outside any other power supply.

A battery provides a direct voltage, a voltage which tends to drive free electrons always in the same direction. It will furnish a direct current, a current of electrons flowing always in the same direction. This is true also of all television and radio power supplies used in connection with all the parts of tubes except the parts which heat the cathodes.

At one end of the battery or any other direct-current power supply is a terminal or connection which is negative. At the other end is a terminal or connection which is positive. The first terminal is negative because here there is an excess of free electrons, or there is a negative charge. The second terminal is positive because here there is a deficiency of free electrons.

A battery, or any other direct-current power supply, is simply a device inside of which the free electrons are pulled away from one terminal, which thus becomes positive or positively charged, and are driven over to the other terminal, which thus becomes negative, or negatively charged.

The free electron whose travels we shall follow is first at the positive terminal of the power supply, as at the left in Fig. 5-3. This electron wants to stay where it is, because the electron is negative, the terminal or its charge is positive, and there is attraction. Furthermore, the electron is being repelled by the negative charge over at the negative terminal.

Supposing that, by some means, we force the electron to move away from the positive terminal and toward the negative terminal, as in the right-hand diagram. The electron is being moved against the forces of attraction and repulsion, and it takes work to overcome those forces. Moving the electron against the forces of attraction and repulsion is like moving the weight off the ground against the force of gravity.

It is easy to measure the amount of work done on any job. Work is measured in a unit called the foot-pound. One foot-pound is the amount of work done in lifting one pound through one foot against the force of gravity. The total number of foot pounds of work is found from multiplying the number of feet of distance by the number of pounds moved through this distance. If you lift the 50-pound weight upward through a distance of 6

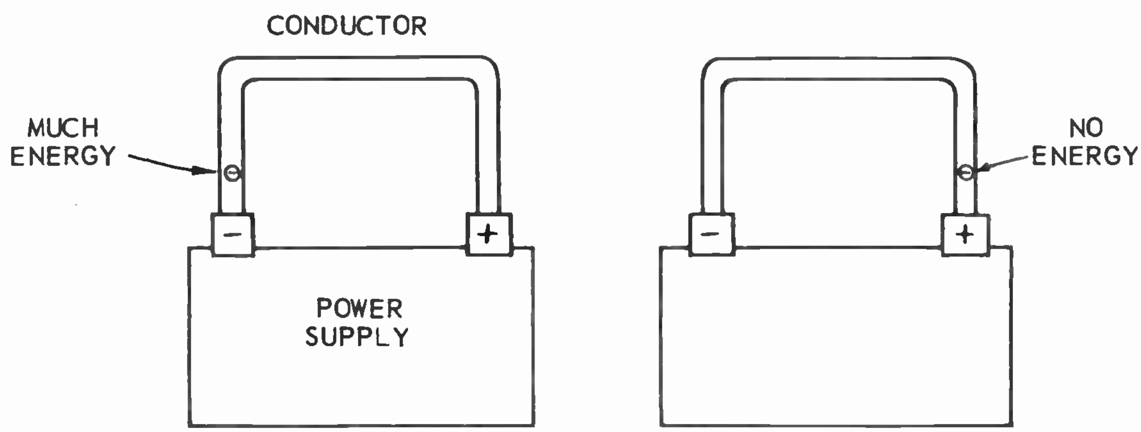


Fig. 5-4. Electrons lose their energy while traveling from negative to positive outside a power supply.

feet you do 6 times 50 or 300 foot-pounds of work. In a few moments we shall come to an equally simple unit of electrical work.

ENERGY. Here is a question. What does the weight possess when elevated that it does not possess when on the ground? Answer: It possesses more energy. Energy means the ability to do work. With the weight elevated it is capable of doing more work than before it was lifted. The elevated weight will do work when it falls. Maybe the work will be no more than knocking some object out of the way, but that is work because there would be motion resulting from a force.

If the 50-pound weight falls back 6 feet to the ground it will do 300 foot-pounds of work (6 times 50), either while it falls or when it lands. When again resting on the ground the weight will have lost its 300 foot-pounds of energy or working ability.

Now back to the electron. Assume that the electron has been forced all the way over to the negative terminal, as in Fig. 5-3. The electron now has been moved as far as possible from its original position inside the power supply. In the original position, at the positive terminal, the electron is like the weight lying on the ground; neither has any energy. But the electron at the negative terminal is like the weight when fully elevated. Then both the electron and the weight have been given energy or working ability, because work had to be done on the electron to get it from positive to negative, and on the weight to get it off the ground and to its maximum elevation.

The electron is forced to travel from positive to negative inside the battery by energy which was stored in the chemical compounds inside the battery. This chemical energy is released as the compounds break down into new substances, just as heat energy is released when fuel gas breaks down into other gases and vapors when burning. Energy released from the chemicals does work on the electron and moves it against the forces of attraction and repulsion. Work done on the electron puts energy into the electron, just as work done by your muscles in lifting the weight puts energy into the weight when it is elevated. The electron finds itself at the negative terminal in possession of the energy which resulted from the work of driving it from positive to negative inside the battery.

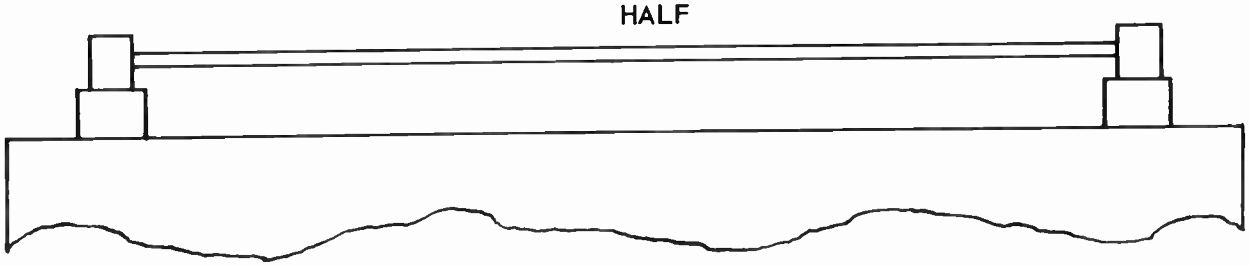


Fig. 5-5. Potential decreases from negative to positive charges or from negative to positive terminals outside a power supply.

Now, as at the left in Fig. 5-4, we shall connect a wire conductor between negative and positive terminals on the outside of the battery or power supply. When the electron, full of energy or working ability, is at the negative terminal it immediately starts through the conductor toward the positive terminal. This happens because the negative electron is repelled by the negative charge at the negative terminal while being attracted by the positive charge at the positive terminal. The electron cannot travel back through the battery, for there it would encounter the chemically released force which acts oppositely. And so the electron travels through the external wire.

That wire conductor is not easy to get through. For one thing it is full of stationary atoms with which the electron collides no matter in what direction it moves. Lots of the atoms have temporarily lost one of their own electrons, and they try to pull our struggling free electron into themselves. The task of the electron in getting through the conductor is like your trying to rush through a hallway thickly studded with stationary pillars, many of which have long, sharp hooks to catch you. You would have to do a lot of dodging, a lot of pushing and pulling, a lot of work in getting through the hallway, and so does the electron have to do a lot of work in getting through the conductor.

The store of energy (working ability) possessed by the electron when leaving the negative terminal enables the electron to do the work of getting through the conductor. But, as the electron progresses, more and more of its energy is used up in doing this work. By the time the electron gets to the positive terminal, at the right in Fig. 5-4, all the acquired energy is gone.

It is natural to ask what happens to the energy lost by the electron in getting through the conductor. Well, every time the electron hits an atom, work is done on that atom at the expense of energy lost by the electron. The moving electron, and billions of others traveling with it, strike the atoms as you might repeatedly strike a piece of steel with a hammer. Try it, and the steel will become hot from the work done on it. Electric energy lost by the electron is changed into heat energy in the atoms. Heat is another of the many forms in which energy may exist.

If you think that tiny electrons cannot strike tiny atoms hard enough to produce heat, remember that there are countless billions of collisions in every small fraction of an inch of conductor. For further proof, feel a wire that is carrying electricity, look at the filament in a lighted electric lamp, or at the heater inside an electronic tube in a radio or television receiver.

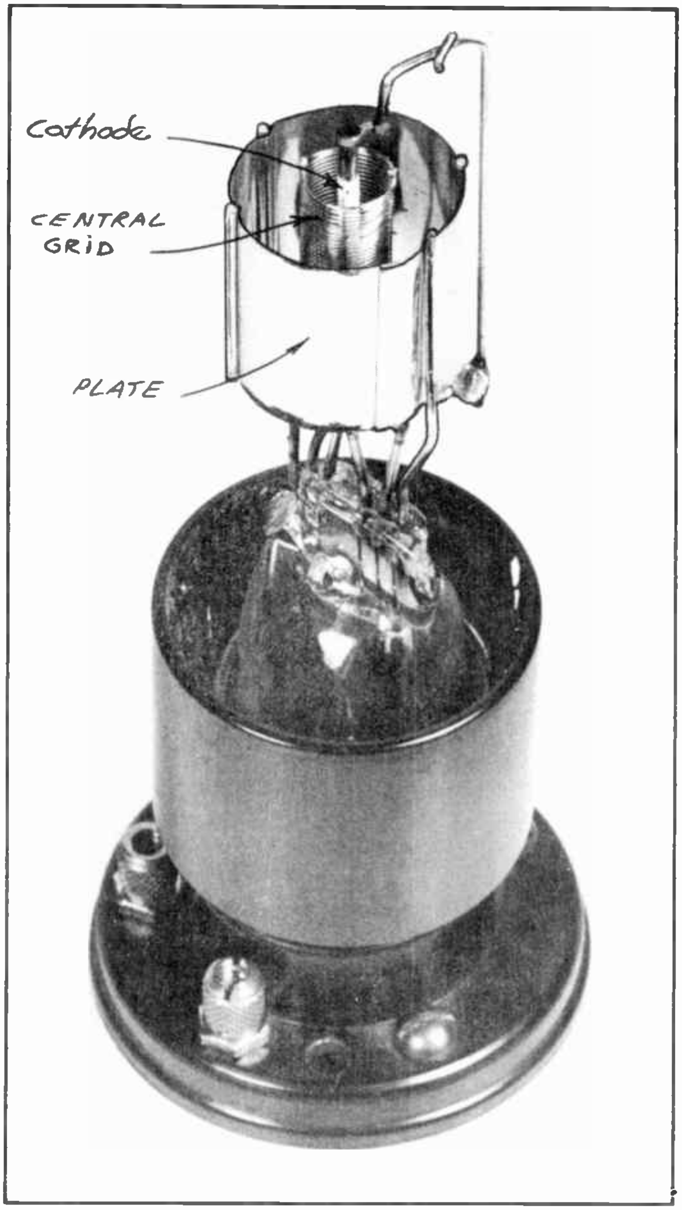


Fig. 5-6. The cathode, the control grid, and the plate of an electron tube.

POTENTIAL. Once more back to the elevated weight. When the weight is off the ground, at any height, it possesses more energy than when lower down. Energy in a body which results from the position of that body is called potential energy. The word potential, coming from "potent", means having ability to do things, such as work.

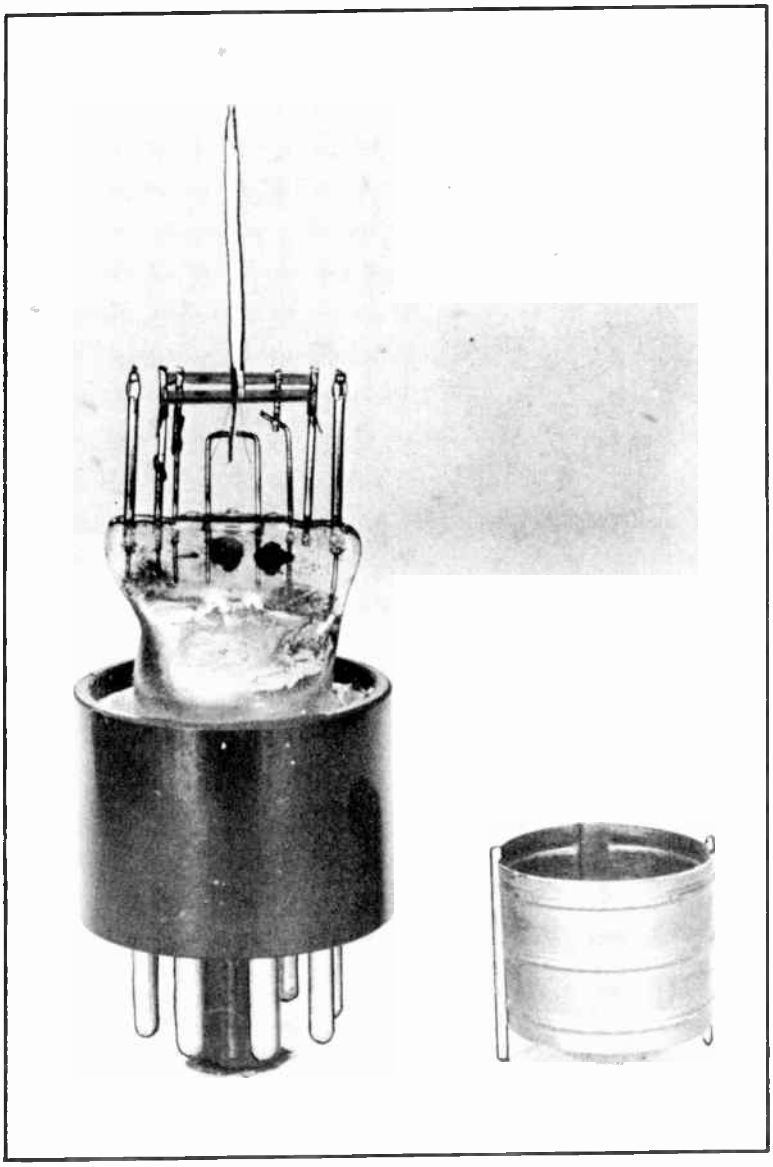


Fig. 5-7. The cathode of an electron tube, with the plate and grid removed.

The electron at the negative terminal of the battery possesses electric potential energy. The elevated weight possesses mechanical potential energy. When the electron is at the negative terminal, Fig. 5-5, the electron possesses maximum potential energy, just as the weight possesses its maximum potential energy when fully elevated.

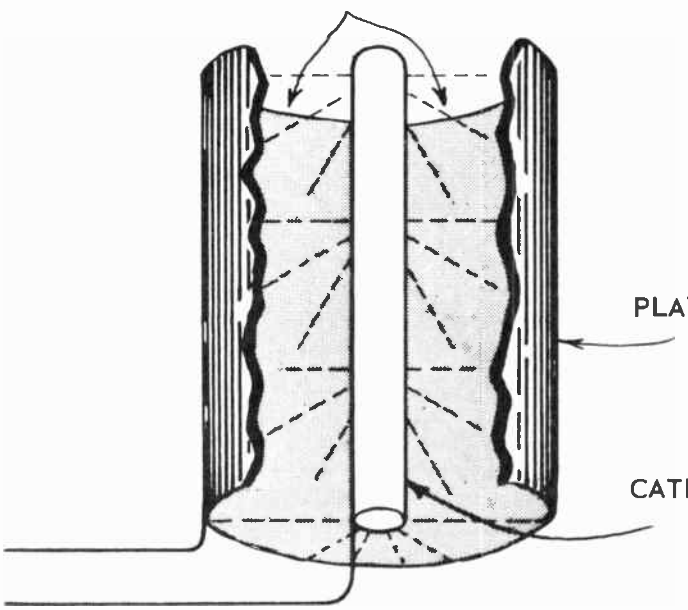


Fig. 5-8. There is an electric field between cathode and plate inside the tube.

If it is just as hard for the electron to work its way through any particular inch of conductor as through every other inch, the electron will have lost half its potential energy when it gets half way from negative to positive. The weight will have lost half its potential energy when it gets half way through its fall back to the ground, for at that point the weight will be capable of doing only half the original number of foot-pounds of work. When the electron arrives at the positive terminal its potential energy will be zero, as is also the energy of the weight when it reaches the ground.

The quantity of potential energy remaining in the electron is strictly proportional to how far the electron has traveled from negative to positive. The energy is proportional to the distance from the negative terminal to where the electron might be, or to the distance remaining between that point and the positive terminal. Now, even were no electrons traveling through the conductor, every point along that conductor still could be described as having some certain potential. That potential would be the value which would have to be in an electron at that point, were any electron present.

It really is this characteristic of points or positions between electric charges that is called potential. The electric potential at any point will tell us how much energy must be possessed by whatever electrons happen to be at the point where the potential is measured.

This thing called potential is not a property of the wire conductor between terminals of the power supply, it exists in just the same manner with the wire removed. Potential relates to points or positions between negative and positive charges. Were an electron to travel through empty space from the negative terminal to the positive terminal, the change of energy in the electron and the decrease of potential along the path would be exactly the same as though the electron were traveling through a connecting wire.

Now let's come up to catch our breath. Quite likely you were beginning to feel that we dove off the deep end into a sea of theory. Not at all. We simply were talking about what happens inside all the tubes used in television and radio. If you remove the glass or metal outer envelope from one of the simpler types of tubes the internal parts will appear as in Fig. 5-6. At the center is the cathode which is heated to make it emit free electrons. Around the outside is a hollow cylinder called the plate. In the space between cathode and plate is a coil of fine wire with its turns widely spaced. This is the control grid for this type of tube.

Fig. 5-7 shows the tube base and internal supports with only the cathode left in place. The plate has been set down alongside the tube. The control grid is not in this picture. Inside the tube there are no conductors of any kind between the cathode, grid, and plate. Each of these "elements" are separated from the others by space which is practically empty, because inside the tube there is a nearly complete vacuum. Electrons emitted from the cathode travel right past the grid wires and to the plate, through nothing but space.

Free electrons go from cathode to plate inside the tube because the cathode is negatively charged and the plate is positively charged. The elements are kept charged by connecting them externally to a power supply, through conductors which go down through the tube base and into the base pins. It is fairly easy for the electrons to flow through the space inside the tube, for in the highly evacuated interior of the envelope there are relatively few atoms of gases with which the electrons may collide.

ELECTRIC FIELDS. Even though there is a nearly complete vacuum inside the tube and between the elements, there still is something of great importance in this space. This thing is an electric field whose position is shown by broken lines in Fig. 5-8. The electric field is only a force, it is nothing which we can see. To understand the meaning of an electric field, think away back to your experiments with the sealing wax and the metal foil. The wax would pull the foil while the two were as much as three or four inches apart. Were you to hang the foil from a very flexible thread, and give the wax a brisk rubbing, you would find that the force of attraction extends for a long distance.

The space in which are acting the forces of attraction and repulsion is an electric field or electrostatic field. The words electric and electrostatic, as used in television and radio, may mean the same thing.

Even though you have an electric charge all by itself in space, nowhere near an opposite charge, there still is an electric field all around the first charge. There is no particular point at which the field ends; it just spreads out through more and more space and becomes weaker and weaker as we move away from the charge that is maintaining the field. Although such spreading fields have important effects, we are more interested for the present in electric fields which exist between negative and positive charges not so very far apart.

Of all the electric fields with which we shall be concerned in television and radio none is more interesting in its effects, and in what we may do with it, than the field between charges on cathodes and plates of tubes. At every point in this field there is a certain potential. If an electron is at some certain point in the field, that electron possesses energy proportional to the potential at that point in the field.

THE VOLT. Potentials at points in an electric field, potentials at the charges which maintain the field, and the potential energy of electrons at various points in the field or in the charges, all are measured by a unit whose name is most familiar. This unit is the volt.

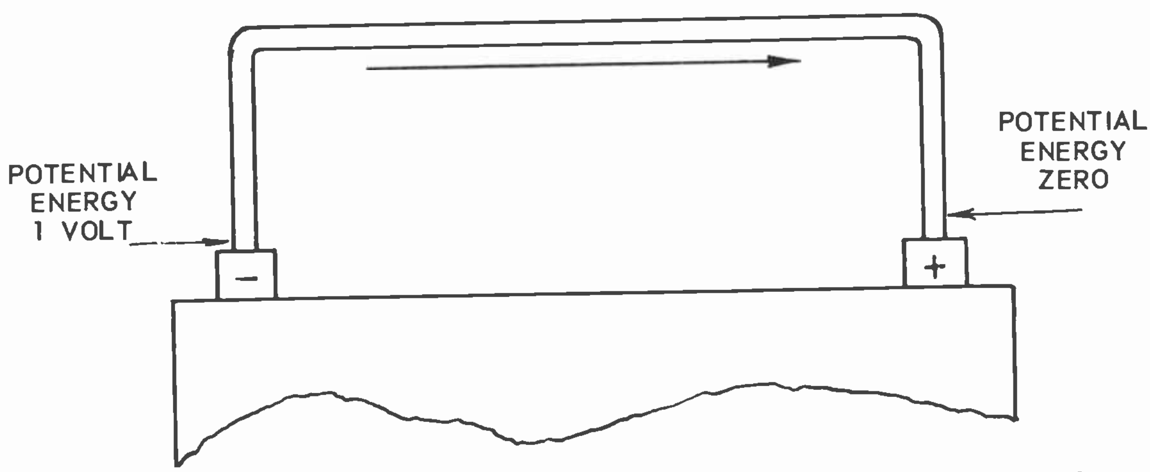


Fig. 5-9. One coulomb of electrons does 0.7376 foot-pound of work while losing one volt of potential energy.

The meaning of a potential energy of one volt is shown by Fig. 5-9. If one coulomb of electrons would have to do 0.7376 foot-pound of work in traveling from a negative charge to a positive charge, or from a negative to a positive terminal, the potential energy in the coulomb of electrons when leaving the negative terminal would have to be one volt. That is, in order that one coulomb of electrons may have the ability to do work equivalent to 0.7376 foot-pound of work, this quantity of electrons must have energy proportional to one volt of potential.

Another way to define a volt is to say that it is the difference of potential through which one coulomb of electrons will move when doing 0.7376 foot-pound of work. Still another way is to say that one volt is the difference of potential through which one coulomb of electrons will be moved when 0.7376 foot-pound of work is done on the electrons. This latter definition would apply to the inside of a battery or other power supply, in which work is done on electrons to move them from positive to negative.

In Fig. 5-10 is represented a potential difference of 6 volts between charges or terminals, also the decrease of potential from one charge to the other, and the fraction of the original maximum energy remaining in electrons which have reached various potential points in between the charges or terminals.

It is quite probable that, after finishing this lesson, you may never have to think about these precise definitions of a volt. You will become so familiar with what may be accomplished by potentials and potential differences of certain numbers of volts that you won't need to think about definitions. But here in the beginning it is essential to realize that there really are definite relations between potentials and electron energy or working ability. If you keep it in mind that there actually is such a relation, everything will be fine.

ELECTROMOTIVE FORCE. The force which gives energy to electrons as they go through the inside of a battery or any other kind of power supply is called electromotive force. This is the force which makes the

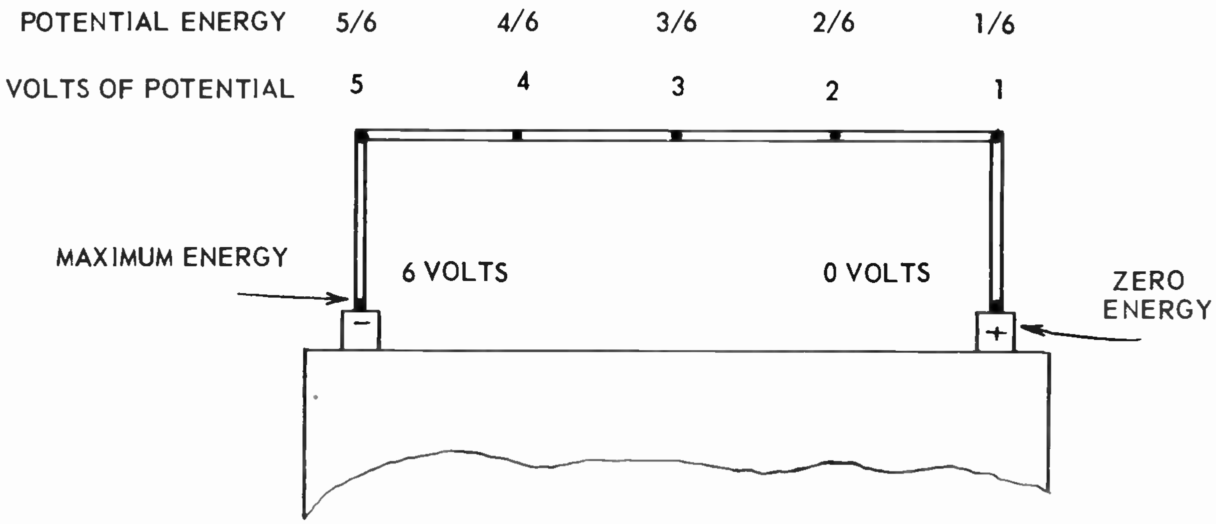


Fig. 5-10. Potential energy in electrons decreases as the volts of potential decrease.

electrons travel from positive to negative inside the battery or power supply. Electromotive force may come from chemical energy in batteries, from light energy in photocells, from heat energy in thermocouples, and from energy of motion in electric generators and in certain kinds of crystals. All these are devices for converting their original form of energy into the potential energy imparted to electrons passing through them. The name electromotive force is abbreviated to emf. When coming to this abbreviation in reading we hardly ever say "electromotive force", we just pronounce the three letters, and say "ee-em-eff".

The unit of measurement for emf is the volt. If a power supply is capable of doing 0.7376 foot-pound of work on each coulomb of electrons passing through it, that source is furnishing an emf of one volt. Thus we learn that emf, in volts, is a measure of the energy imparted to electrons as they pass through a power supply. A single dry cell furnishes an emf of slightly more than 1½ volts, and does so regardless of the size of the cell. How much emf is furnished depends on the kinds of chemicals in a cell, not on how many ounces or pounds may be used.

ELECTRICAL RESISTANCE. Probably you have noticed when using a flashlamp that the light finally becomes quite dim as the cells or battery run down or become discharged. Yet when the battery is so far gone as to make the flashlamp filament hardly red the battery is furnishing just as much emf as when brand new and fresh. Then why doesn't the lamp light brightly? The answer would be easy if we knew more about opposition to electron flow, and before long we must know a lot more about this opposition or there will be many service questions which cannot be answered.

We have talked about electrons and their flow through conductors and through empty spaces. We have learned about quantities of electrons measured in coulombs, and about rate of flow measured in amperes. We have learned about work and energy, about potential and potential difference, about emf's, and about measuring these things in volts. But knowing all about amperes and volts and what they measure isn't going to help solve very many problems in servicing until we know about the opposition to electron flow which exists in all things.

Opposition to electron flow is called electrical resistance. Since everything we talk about is electrical, we usually call this opposition just resistance.

When it comes to measurement of resistance the television-radio technician really has things easy, because there is a unit of resistance which is as simple as the ampere for measurement of electron flow rate, and the volt for measurement of potentials and emf's. In no branch of science other than electricity is there any simple unit for opposition to flow.

Were you learning to be a heating and air conditioning engineer you could specify opposition to air flow only by saying that a certain duct requires so many pounds per square inch of air pressure to maintain a flow of so many cubic feet per minute. Were you learning to be a hydraulic engineer or a master plumber you would deal with pounds per square inch per cubic foot per minute per foot of pipe. But the radio-television technician can say that some certain wire, resistor, or tube has resistance of so many ohms — just the word ohms tells the whole story. Knowing the ohms of resistance the technician can tell in a jiffy how many amperes or milliamperes will flow with any given potential difference in volts, and what potential difference must exist to cause any given flow rate.

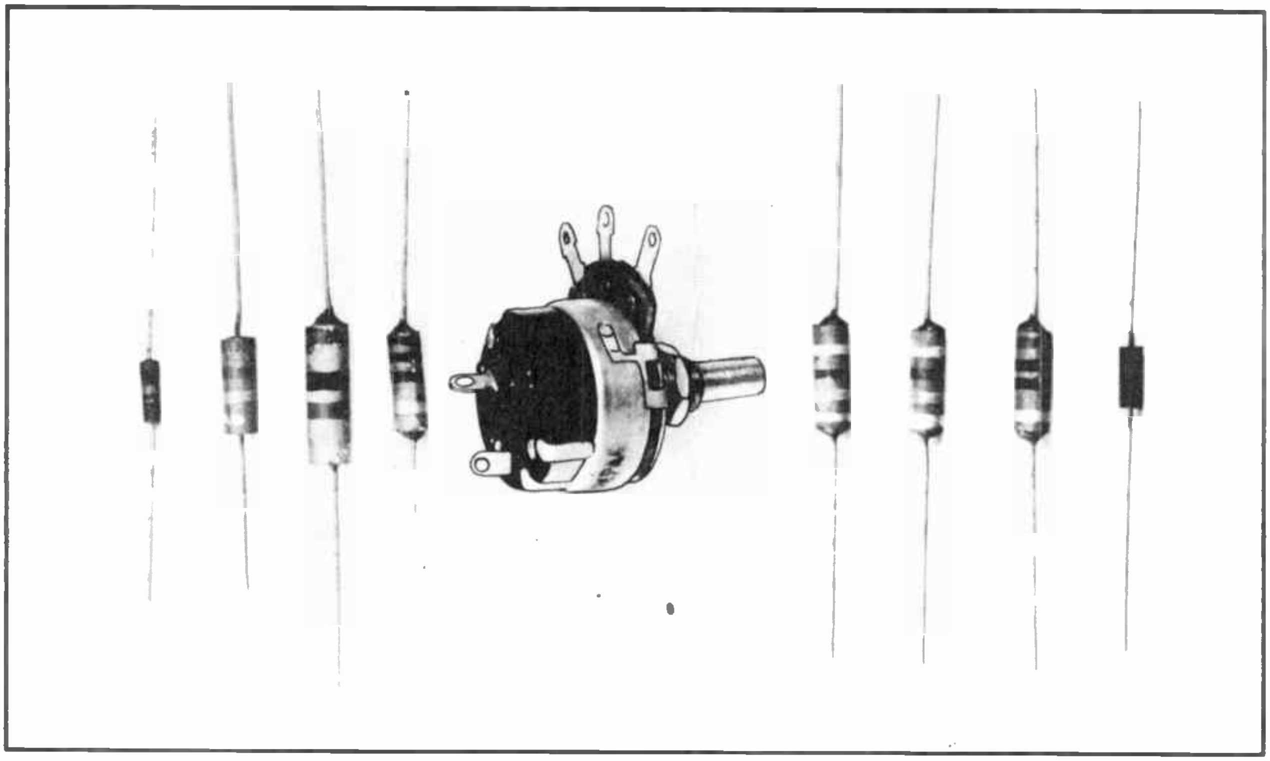


Fig. 5-11. Resistors such as used in television and radio receivers.

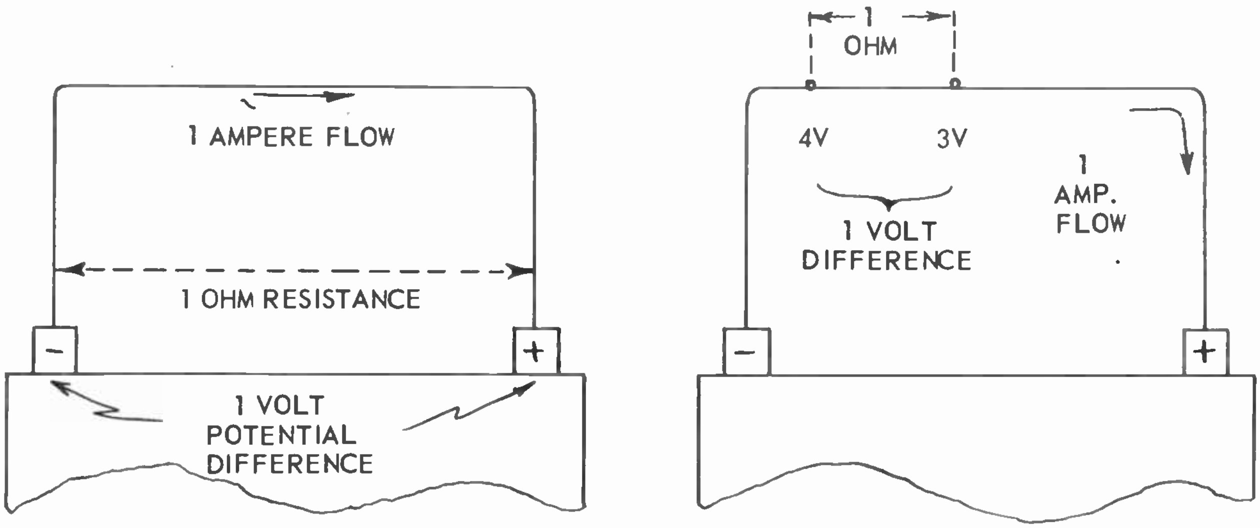


Fig. 5-12. When electron flow between two points is at the rate of one ampere, and potential difference between the same two points is one volt, the resistance between these two points is one ohm.

Resistors, whose chief purpose is to provide resistance where we need it, are one of the "big four" in all television and radio circuits. The other three are capacitors, inductors, and tubes. We have looked at many pictures showing the wiring and small parts underneath the chassis of receivers. Dozens of resistors were visible in those pictures. Fig. 5-11 shows a few resistors by themselves. At the center is a type whose resistance may be adjusted to various values while a receiver is operating. All the others are fixed resistors, whose resistance in ohms cannot be altered after they are manufactured.

The unit of electrical resistance is called the ohm. At the left in Fig. 5-12 are represented two points between which there is a potential difference of 1 volt. The rate of electron flow between these points is 1 ampere. The resistance to flow between these two points is 1 ohm. One ohm of resistance is the amount of resistance which allows a flow rate of one ampere when the potential difference is one volt; this is the definition for one ohm. At the right in Fig. 5-12 there is a potential of 4 volts at one place and of 3 volts at another place. The potential difference is 1 volt. The flow rate is 1 ampere. Resistance between these two places along the conductor must be 1 ohm.

Resistance is something which exists in a conductor whether or not there is any electron flow or any potential difference in the conductor. Resistance is a property of the conductor itself. There are several features of a conductor which affect its resistance. First of all, the resistance or opposition to electron flow depends on the kind of material in the conductor, on the extent to which free electrons get tangled up with

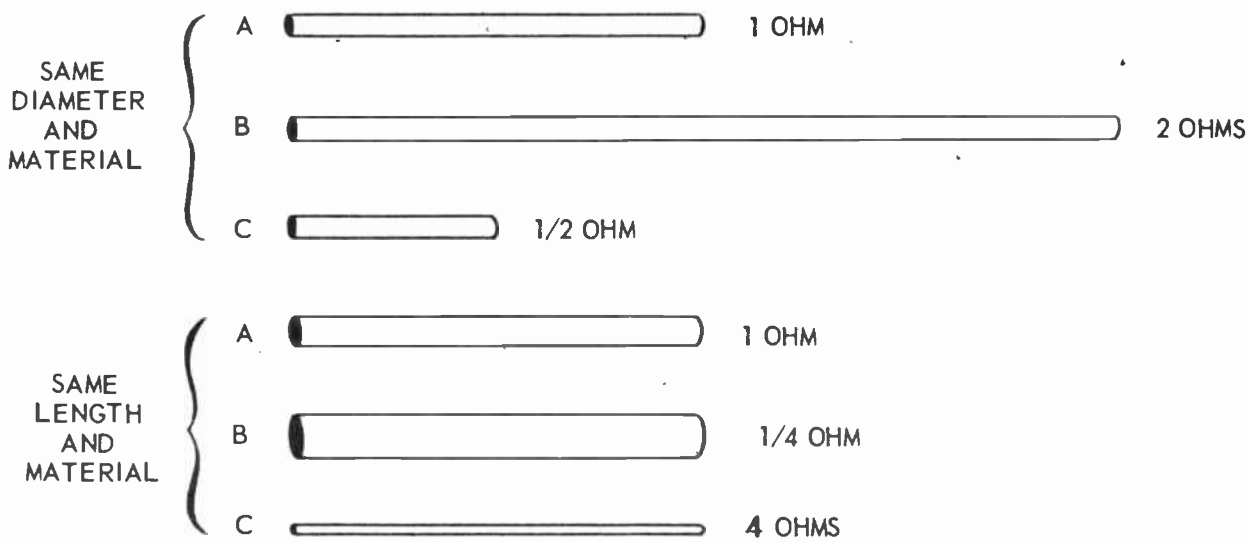


Fig. 5-13. Resistance varies directly as the length of a conductor, but varies inversely as the square of the diameter.

the stationary atoms and on how tightly the atoms hang onto their outermost electrons.

The atoms in mild steel hang to their electrons more than six times as hard as do the atoms in copper used for hook-up wiring. You might cut off a piece of copper wire having a resistance of exactly 1 ohm, then cut a piece of galvanized steel wire of the same length and same diameter. The resistance of the steel wire would be about 6 1/3 ohms. The same length and same diameter of the kind of wire used in many resistors would have resistance of about 60 ohms.

Two other factors which affect resistance are illustrated by Fig. 5-13. At the top are three conductors, all of the same material and of the same diameter. Doubling the length doubles the resistance. Halving the length halves the resistance. Resistance is directly proportional to length of a conductor when nothing else is changed.

At the bottom are three more conductors, all of the same material and of the same length. The diameters differ. The diameter of conductor B is twice that of conductor A, and the diameter of C is half that of A. Cross sectional area varies as the square of diameter. This is the area, in square inches, exposed on the end of a conductor which has been cut straight across. When we multiply diameter by 2 the cross sectional area is multiplied by the square of 2, which is 2 times 2, or 4. For 1/2 the original diameter the cross sectional area is proportional to the square of 1/2, which is 1/2 times 1/2, or 1/4.

Resistance varies inversely as the cross sectional area of a conductor, all other things being unchanged. That is, with twice the area there is half the resistance, with half the area there is twice the resistance, and so on. Conductor B has 2 times the diameter of A, so has 4 times the cross sectional area. With 4 times the original area we have 1/4 the resistance. Conductor C has 1/2 the original diameter, so has 1/4 the original cross section. With 1/4 the cross section we have 4 times the original resistance.

It is easy to see why resistance must vary directly with length. It is twice as hard for electrons to get through 2 feet of a given conductor as through 1 foot of the same conductor, just as you would have to do twice as much work in getting through twice the length of a hallway full of pillars to oppose your progress.

It is not difficult to see why resistance varies as it does with diameter or size of a conductor. With twice the diameter, and four times the cross sectional area, there will be four times as much material and four times as many atoms and four times as many free electrons. With four times as many electrons acted upon by attraction and repulsion there will be four times the flow rate. Four times the flow rate, with the same potential difference, must mean only one-fourth the original resistance.

We have used some numbers and fractions, and some arithmetic, to show in a definite way why certain things are true. Don't worry about remembering the arithmetic, but remember two facts. Resistance varies directly with length of a conductor. Resistance increases as diameter gets smaller, and does so proportionately to the squares of the diameters.

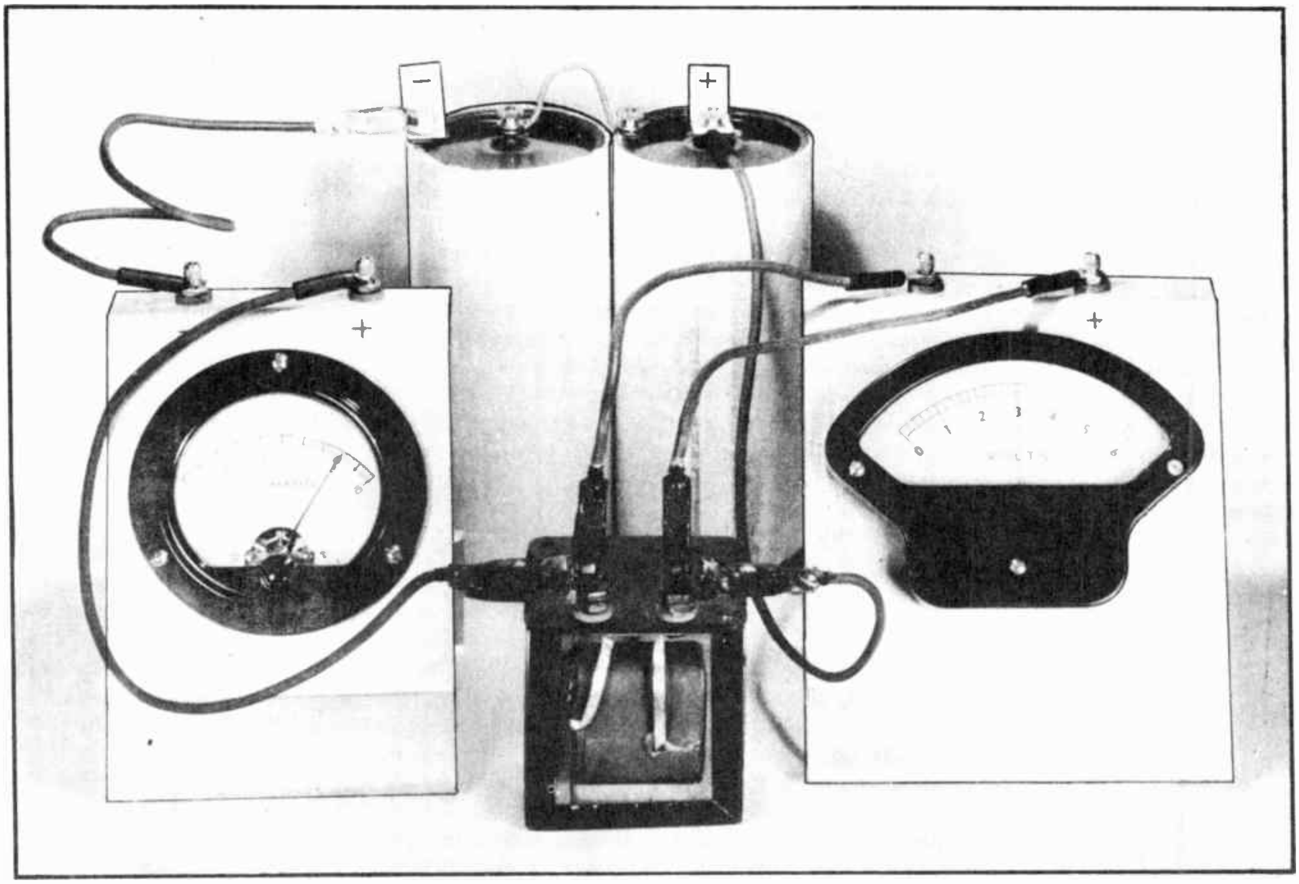


Fig. 5-14. Measuring the resistance of a filter choke by means of a voltmeter and a current meter.

There is one more thing which alters the resistance of conductors. It is temperature. In all pure metals, and in most alloys or mixtures of metals, resistance increases as temperature rises – and decreases as temperature drops. We shall look into the relations between temperature and resistance a little later on.

It is possible to measure the actual resistance of any conductor by applying to the ends of the conductor some certain difference of potential, which can be measured in volts, then measuring the current or rate of electron flow caused by this potential difference. The number of volts potential difference divided by the number of amperes of electron flow rate, equals the number of ohms of resistance.

This method of resistance measurement is being used in Fig. 5-14. The resistance of a filter choke (a kind of inductor) is being measured by using two dry cells as the power supply or the source of potential difference, by using a voltmeter to measure the actual potential difference, and by using a current meter to measure the rate of electron flow in the unit being measured.

There are easier ways of measuring resistance. Service technicians make such measurements with an instrument called an ohmmeter. When you connect an ohmmeter across any conductor, the number of ohms resistance is directly indicated on the scale of the meter. When you know a little more about relations between volts, amperes, and ohms, you will know why an ohmmeter gives direct readings of resistance.

A BAD MISTAKE

This is the story of one of the most unfortunate errors ever made in science. Long before there was radio or television, and before men knew about free electrons, it was arbitrarily assumed by scientists of that day that electricity flows from positive to negative in conductors outside a power supply or source.

During all the years in which production and distribution of electric power had their beginnings and much of their industrial development, all text books, instructions, and wiring diagrams were based on the wrong assumption that electricity flows from positive to negative in conductors outside a source of potential, whereas we now know that the only actual flow of anything at all is that of electrons from negative to positive.

By the time the truth became known it was too late to undo the damage; it was impossible to have everyone agree that, at some certain day and hour, they would reverse their idea about the direction of electric flow. Consequently, even to this day, a big part of the electrical world talks about flow from positive to negative outside the sources.

Some people try to get around the contradiction by saying that electricity flows from positive to negative, while electrons flow from negative to positive — but they cannot explain of what their kind of electricity consists. Others say that "conventional" flow is from positive to negative, while actual flow is from negative to positive.

Only in radio, television, and some other electronic fields does everyone recognize that the only flow is of electrons, and the only direction of their flow while outside a source is from negative to positive. In our work with alternating potentials and currents the direction of flow need not be considered, for it is first one way and then the other. But even then, the instantaneous flow always is from negative to positive, and when the flow reverses it is because the negative and positive polarities of charges and potentials have reversed.

Listed below are various reference books dealing with the direction of electron flow and current. You may be able to secure one at your library if you wish to do some additional research on this interesting subject. The exact page where our particular theme can be found is also noted.

Elements of Electricity, W. H. Timbie, 2nd Edition, page 539

Manual of Radio Telegraphy and Telephony, U. S. Naval Institute, 8th Edition, page 9

Radio Instruments and Measurements, U. S. Bureau of Standards, 2nd Edition, page 8

Van Nostrand's Scientific Encyclopedia, 2nd Edition, under Electronics.

Standard Handbook for Electrical Engineers, Knowlton, 7th Edition, page 33.

6 → 10

COMMERCIAL TRADES INSTITUTE

KNOWLEDGE + AMBITION = SUCCESS

LEARN . . .

ACT . . .

Training Your Mind

C. T. I.

Perfecting Your Skill

ENJOY . . .

"The heights by great men reached and kept
Were not attained in sudden flight,
But they, while their companions slept,
Were climbing upward in the night."

COMMERCIAL TRADES INSTITUTE

CHICAGO, ILLINOIS

TELEVISION

LESSON NO. 6 CIRCUITS FOR RADIO AND TELEVISION

Training Your Mind

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

COMMERCIAL TRADES INSTITUTE

Training Your Mind... C.T.I. Perfecting Your Skill

LESSON NO. 6

CIRCUITS FOR RADIO AND TELEVISION

PREFACE

As you work on television receivers, and radio receivers, with the help of service manuals issued by the manufacturers, you will find detailed information on how to trace trouble to some one part of the set. Then, time after time, you will read a statement something like this: "The exact point of trouble now may be found with routine tests of continuity, resistance, and voltage." You are supposed to know how to make these all-important tests, which really are ordinary routine procedure for the experienced technician.

A continuity test shows only whether conductors extend all the way between certain points, and whether there may be electron flow in parts and connections between these points.

Resistance tests mean measurements of actual resistances between various points, in comparison with values of resistance which should exist when the receiver is in good order.

Voltage tests mean measurements of actual potential differences, usually between chassis metal and certain specified points, in comparison with values of potential difference which should exist when the receiver is in good condition.

As part of their service data, manufacturers usually furnish either tables, charts, or diagrams showing resistances or voltages or both.

In this lesson we shall commence making tests of resistances and of potential differences in a simple circuit of an actual receiver. Then we may proceed to interpreting the results of these tests as they indicate possible troubles. Continuity tests are so simple that we shall practice them later on. Our job won't be completed in this one lesson, but there will be a good start.

CIRCUITS FOR RADIO AND TELEVISION

If you ask a radio service technician what one instrument he uses more often than any other he is almost certain to say it is his volt-ohm-milliammeter. Maybe he will call this instrument a set analyzer, but it means the same thing. Fig. 6-1 is a picture of a volt-ohm-milliammeter. There are many other sizes, and many special features in various makes and models. But all of them do what their name implies, they measure volts of potential difference, ohms of resistance, and milliamperes of electron flow or current.

In the preceding lesson we learned enough about potential difference, resistance, and electron flow to proceed with some practical measurements in a receiver. Of course, we are not yet very expert in these matters, consequently it will be well to pick something simple to begin with. From the electrical standpoint nothing in a receiver is simpler than the connections for the cathode heaters of the tubes, so that is where we shall do some checking.

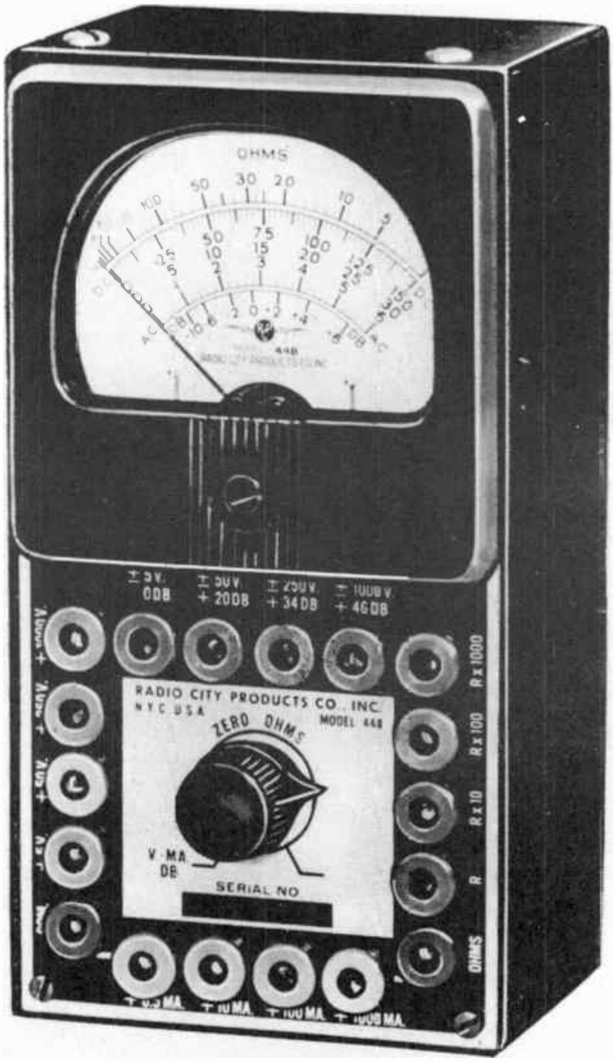


Fig. 6-1. A volt-ohm-milliammeter used for service tests.

We already have looked at a tube with its envelope removed and with only the cathode left in place. Were you to remove the cathode carefully the heater would be exposed. The heater is a length of wire having considerable resistance. Spread out, this wire would appear as in Fig. 6-2. When we apply potential difference to the ends of the heater wire, by way of pins on the tube base, the resulting electron flow will raise the temperature of the heater wire high enough to make it bright red. Heat passes into the surrounding cathode, which becomes dull red. Then the cathode emits electrons.

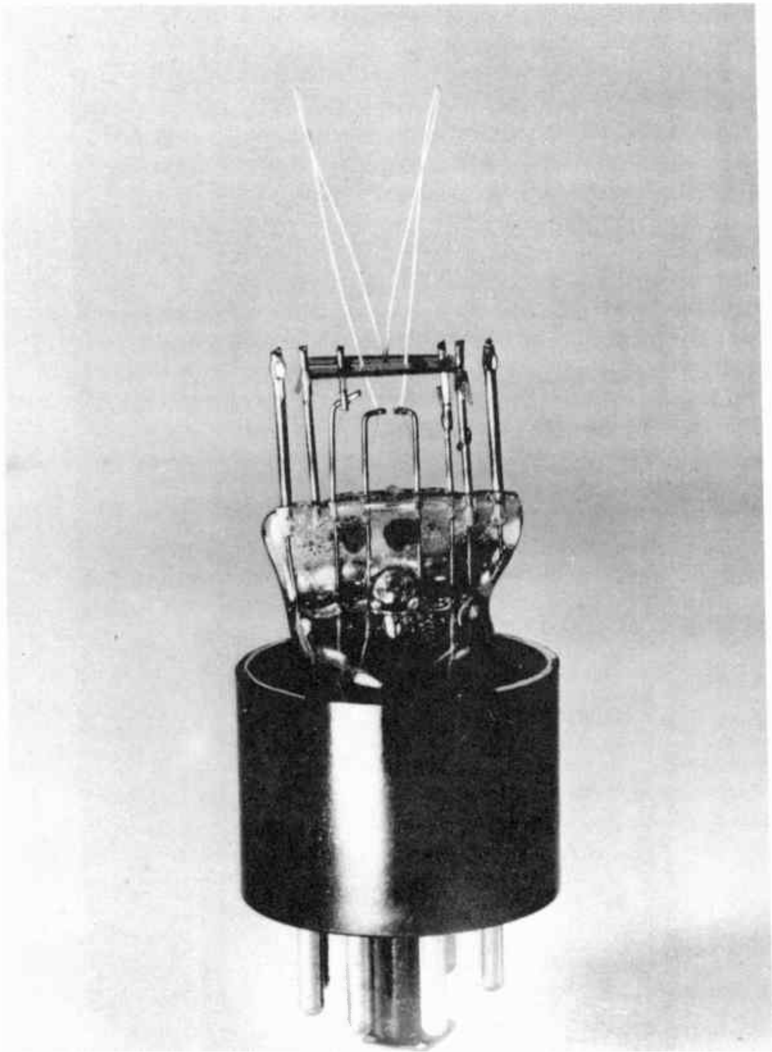


Fig. 6-2. The heater as it appears when the cathode is removed.

Our first measurements will be made on a radio set having a total of six tubes. The two ends of the heater in each tube are connected to two of the pins on the tube base. When the tube is pushed into a socket the base pins make contact with extended lugs on the bottom of the socket. To these lugs are connected the wires which bring potential difference to the heater.

In Fig. 6-3 you can see how the six sockets for the six tubes appear from underneath the chassis. The only wires which have been installed and connected are those for the tube heaters. In spite of the fact that we have here the simplest of wiring it obviously is difficult to show the paths for electron flow and the paths

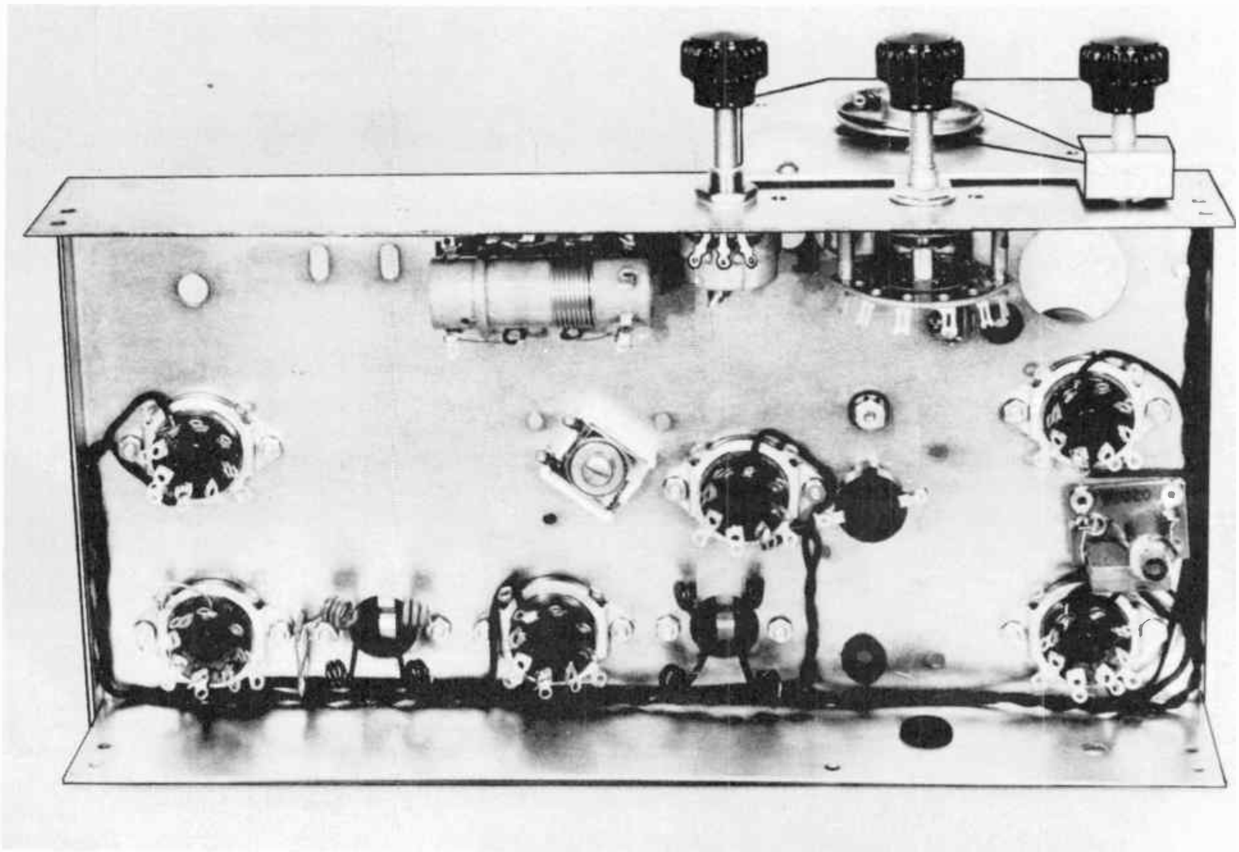


Fig. 6-3. Wiring for the tube heaters, as seen from underneath the chassis.

in which there is change of potential energy by using a photograph.

Service technicians have little use for photographs. They much prefer wiring diagrams. A technician would rather have one good wiring diagram than a basket full of photographs. Fig. 6-4 is a wiring diagram for our heater connections. In addition to the wires shown in the photo, this diagram includes connections to the a-c (alternating-current) power line and to the on-off switch of the receiver.

SOCKETS AND TUBE BASES. Each of the sockets in this receiver has eight lugs, they are called octal sockets. As shown by Fig. 6-5, in the center of each socket is a hole into which fits a round extension on the tube base. On this base extension is a lengthwise ridge or key, and in the socket hole there is a lengthwise groove or keyway to take the key when the tube is pushed home. The key usually is called the locating

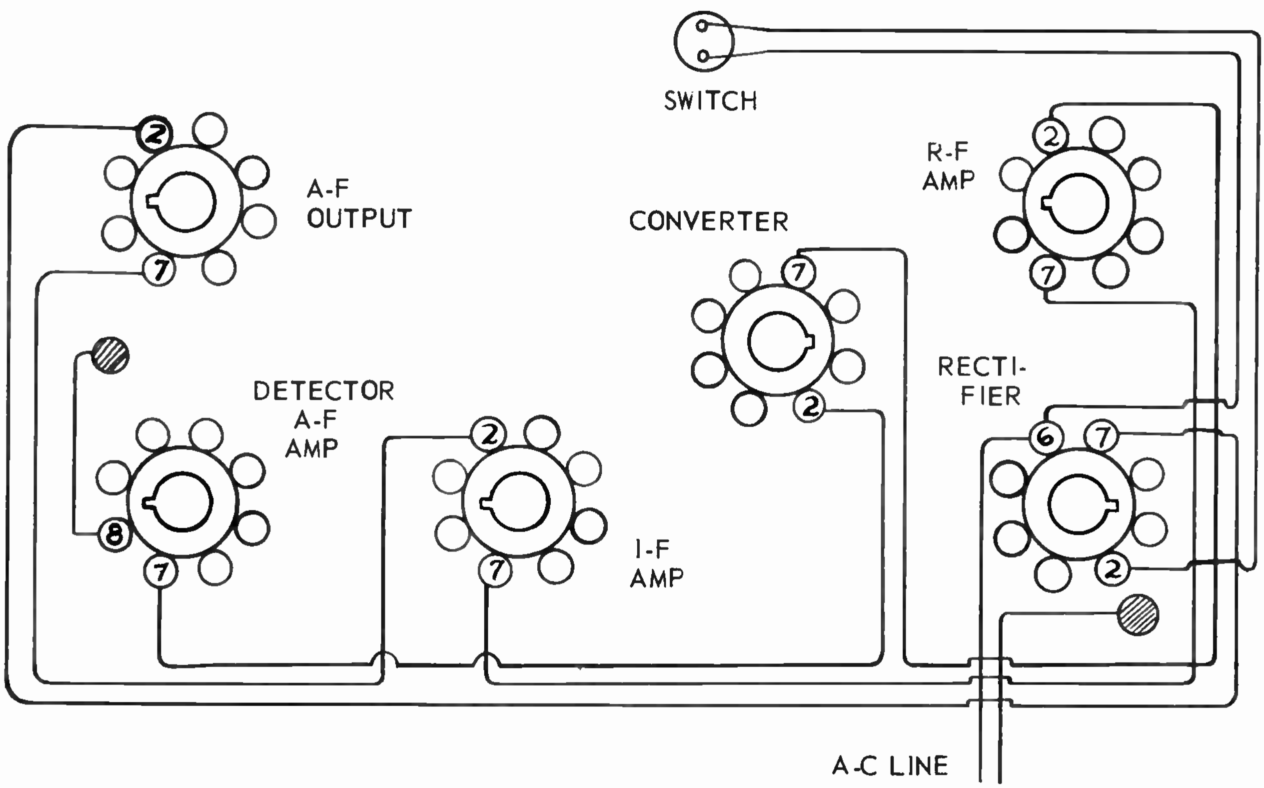


Fig. 6-4. A wiring diagram of the heater circuit.

key, for it locates or determines the position of the tube in the socket.

When looking at the bottom of a socket, and the locating key is pointing towards you, the lugs are numbered from 1 to 8, clockwise, starting at the left of the locating key. When looking at the top of the socket, where the tube pushes in, the pin holes are given corresponding numbers. The numbers go around counter-clockwise when looking at the top. Pins on the tube base are assigned numbers to correspond with numbers of the socket lugs.

Some sockets have these numbers formed right on them, but whether or not the numbers appear you are supposed to know their positions with reference to the locating key. There seldom are numbers on the tube base, but again you are supposed to know the number of each pin according to the diagram of Fig. 6-5.

A HEATER CIRCUIT. Now we may go back to Fig. 6-4 and trace the heater circuit. A circuit is a path

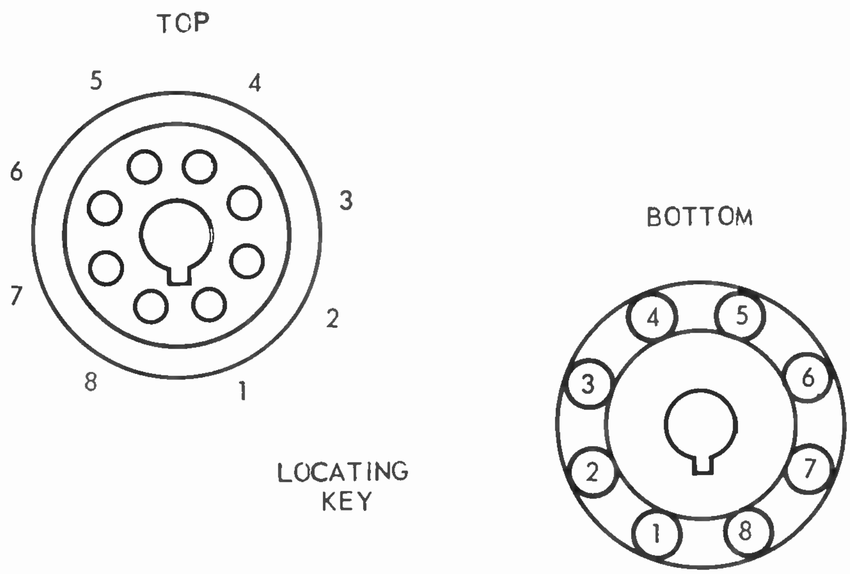


Fig. 6-5. Standard numbering for lugs or terminals, and pin holes, on an octal socket.

in which electrons may be made to flow, either in one direction or back and forth, when a potential difference is applied to the ends of the path. The greater portion of a circuit or a flow path ordinarily consists of wires and other metallic conductors. However, electrons may be made to flow in the evacuated space inside the tubes, and they may be made to flow in liquids contained in some kinds of batteries and some photocells – so a circuit need not consist wholly of metallic conductors.

All useful circuits contain one or more loads. A load is anything wherein the electrons may expend some of their energy and do work that is useful or desired. The entire receiver is a load on the power line. Tubes, resistors, inductors, capacitors, and other parts are loads on the receiver power supplies. Of course, all the wires and other metallic current-carrying parts of the receiver have some resistance to electron flow, and some energy is lost by electrons traveling in them. But in well designed apparatus the resistance of these conductors is so very small that loss of energy in them is negligible in comparison with that lost in the loads. As a consequence, when talking about resistances in a circuit we seldom need to pay any attention to the slight resistance of wiring and other conductors which are not parts of any load.

The chassis of receivers are made of steel, aluminum, or other metals. All metals are good conductors, so the chassis metal is used as a circuit conductor in nearly all cases. The total cross sectional area of this metal is so great that, even with a metal such as steel, the resistance is exceedingly small.

As the word circuit is used in practice it may apply to only a few parts and connections. We may have a heater circuit including only the tube heaters and their connections. We may have a plate circuit, or a grid circuit, or a speaker circuit, and so on. Using the word in this manner, a receiver is made up of a great many circuits.

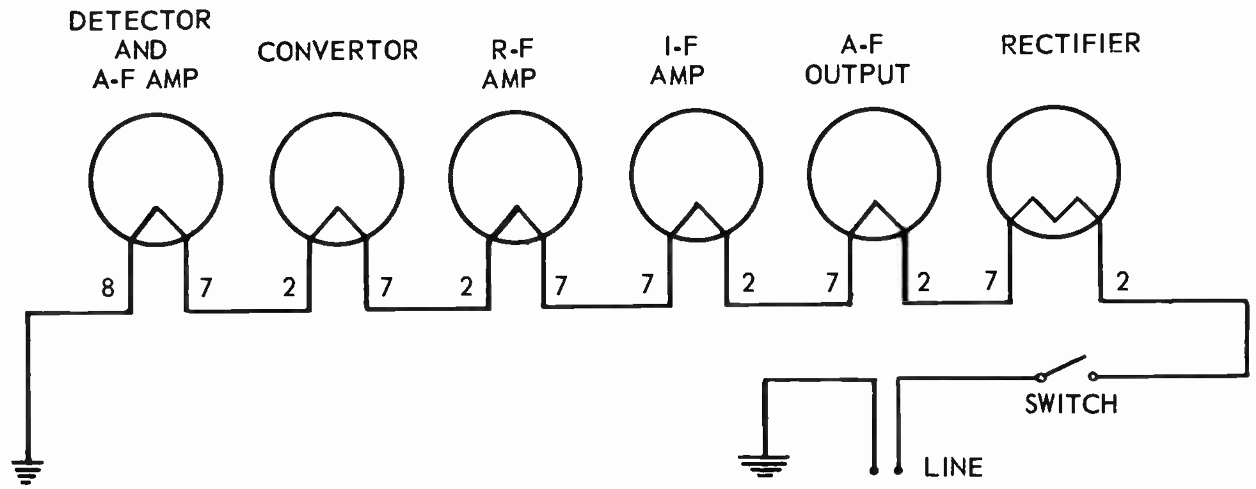


Fig. 6-6. A schematic diagram for the heater circuit.

Before tracing through our heater circuit we shall look at a kind of diagram which most experienced technicians prefer even over wiring diagrams. This other kind is called a schematic diagram. A wiring diagram, as you may see by comparing Figs. 6-3 and 6-4, shows the parts and the wiring connections in the positions they actually occupy. A schematic diagram pays no attention to actual positions but shows parts and connections in whatever way makes the electrical performance most clear. A schematic shows the "scheme of things".

Fig. 6-6 is a schematic diagram of the heater circuit. First we note the simplicity of the diagram and the ease with which connections may be followed. Then we note that this schematic diagram shows the electron flow paths not only as far as the socket lugs, but right through the heaters in each tube. The tubes are represented by circles and the heaters by inverted V's. These are standard symbols for tube envelopes and for heaters. The switch is represented by a simple symbol which makes plain the action of opening to break the conductive path, and of closing to complete the path.

Terminal 8 of the combined detector and a-f (audio-frequency) amplifier tube, also one side of the a-c line, are shown connected to ground. The symbol for a ground connection is a number of unequal horizontal lines. Ground, in this particular case, is the metal of the receiver chassis.

GROUNDS. A ground originally referred to, and still may refer to, a connection into the great mass of the earth. When the soil is moist, as usually it is after penetrating only a little way below the surface. the earth is a good electrical conductor. It contains the greatest imaginable quantity of free electrons. It is assumed, correctly, that the quantity of free electrons in the earth is so great that no quantities which might be added or withdrawn could have any measurable effect on the potential of the earth. For this reason, the potential of the earth or ground is considered as zero when no other reference value is specifically mentioned.

If we speak of a potential of some certain number of volts, without stating or implying that the voltage is with reference to some one point in a circuit, this potential is with reference to the earth or ground as zero.

When we wish to make a ground connection from a receiver to the earth it is usual practice to run a heavy wire from chassis metal to a cold water pipe or to some other good conductor running deep into permanently moist soil. In this ground wire there is so little electrical resistance and so little possible change of potential that the potential of the receiver chassis becomes practically the same as that of the earth, which is zero potential.

In most modern receivers we do not use a wire connection from the chassis metal to a cold water pipe or other earth ground, yet it is general practice to speak of the chassis metal as ground. More correctly it is called the chassis ground. All conductors connected to the chassis ground are effectively connected together through this path of low resistance. All are at the same potential so far as direct-current circuits and low-frequency alternating current circuits are concerned. Other points in any connected circuit may be either positive or negative with reference to chassis ground, they may have either deficiencies or excesses of free electrons with reference to the normal or neutral quantity in the chassis or ground.

ALTERNATING POTENTIALS. As mentioned before, the potential difference applied to our heater circuit comes from the a-c (alternating-current) power line. In any a-c power line the potential in each of the two wires becomes alternately positive and negative. If, for example, the line frequency is 60 cycles per second the potentials in both wires go through one complete change during every time period of 1/60 second.

The potential of each line wire varies as shown at the top of Fig. 6-7. Potential goes first from zero to the maximum positive value, then back to zero and to the maximum negative value, and back to zero again during each 1/60 second. In the lower diagram are represented the simultaneous changes of potential in the two wires of the power line, with the alternating potential of one wire marked A and of the other wire marked B. During each half-cycle in which wire A is positive, wire B is negative. During each half-cycle in which wire A is negative, wire B is positive. The two potentials always are opposite except when there is no potential at all, as both pass through their zero values.

The potential difference between the two wires of the power line is continually varying between zero and maximum. Were this varying potential difference applied to a resistance the resulting electron flows would produce heat in the resistance material. Were we to disconnect the a-c supply wires from the resistance and apply instead a steady potential difference, it would be possible to adjust this steady potential difference to a value producing just the same heating effect as produced by the alternating potential.

We might measure the steady potential difference and find it to be 100 volts. Then the "effective voltage" of the alternating potential from the a-c line would be called 100 volts. To have an effective (heating) value of 100 volts, the alternating potential difference would have to change between zero and 141.4 volts. All ordinary service voltmeters for measurement of alternating potentials are designed to indicate effective values of potential on their dials. This includes the volt-ohm-milliammeters.

Voltages of a-c power lines always are specified in effective volts. If your home supply is at 110 to 120 volts, this is the range of effective voltage. All television and radio receivers designed for operation on

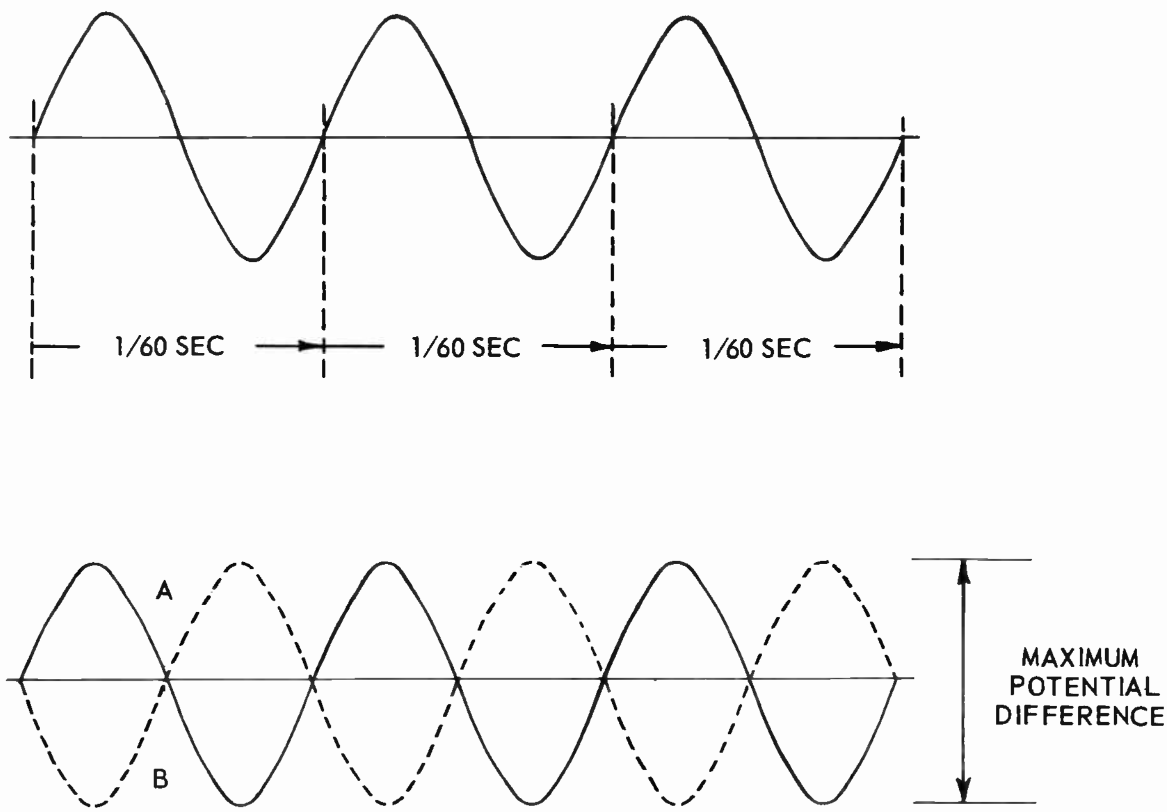


Fig. 6-7. How potential varies at one wire (above) and how potentials vary at the two wires (below) of an alternating-potential power supply.

110-120 volt a-c power lines are adjusted for normal operation at an effective alternating potential difference of 117 volts.

POTENTIAL DROP. Now, with the help of the a-c volts scale of our volt-ohm-milliammeter or with any suitable a-c voltmeter, we shall check the potential differences in the heater circuit. You have learned that there is variation of potential from a negative to a positive charge, or from a positive to a negative charge, or between positive and negative terminals of a power supply. The a-c line is our present power supply, Neither end is constantly positive or negative, they alternate continually. But our voltmeter will show what happens, just as though one end were always negative and the other always positive. When we apply across this heater circuit an alternating potential difference of 117 effective volts, meter indications will be the same as though we were applying a steady one-way (direct) potential of 117 volts.

Our first measurements will be of the potential differences across the heaters of each of the tubes while the effective potential difference across the line is 117 volts. Voltages across the individual heaters are

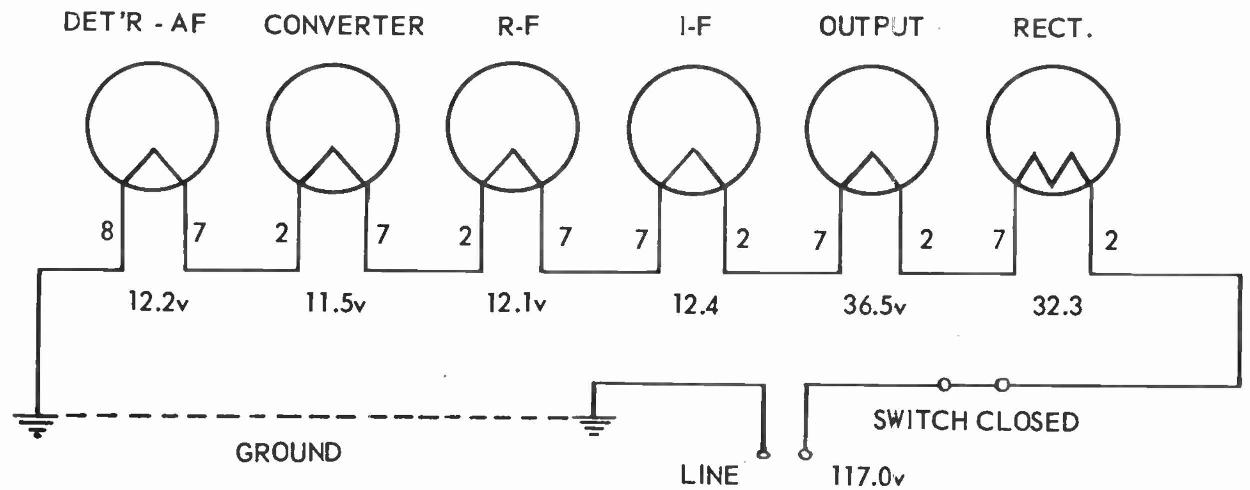


Fig. 6-8. Potential differences as measured across the heaters of each of the six tubes.

shown by Fig. 6-8. Add all these separate voltages together. Their sum is 117.0 volts. The total potential difference from the line provides all the separate potential differences for the heater circuit. Incidentally, these heater potential differences were measured on a real receiver with the applied a-c potential adjusted to 117.0 volts. Service shops often use voltage adjusting transformers to bring testing voltages to a desired value when actual power line voltage is too low or too high.

Our next voltage check is illustrated by Fig. 6-9. One terminal of the testing meter is connected to one side of the power line in the receiver and left there. The other terminal of the meter will be connected successively to points marked a, b, c, and so on through to h. With the meter connected to point a the voltage reading is zero.

This zero reading is the result of having both sides of the testing meter connected to the same conductor (chassis metal), which is of negligible resistance or of practically zero resistance. One side of the meter connects to this chassis metal at grounding point a, the other side connects to chassis metal through the power line connection to ground. If a conductor has no resistance the free electrons lose no energy as they move in the conductor. Electron energy is measured by potential. If there is no change of electron energy there can be no change of potential, no potential difference — and so the meter reads zero.

When the testing meter connection is taken away from point a and applied to point b of Fig. 6-9 the meter is effectively connected across the resistance of the heater in the left-hand tube. Can you see why this is true? If not, look at the small sketch at the bottom left of Fig. 6-9. When the right-hand terminal of the meter is connected to ground anywhere in the receiver the results are the same as with this terminal connected to ground anywhere else, because all the ground metal is at the same potential. A connection to the power line

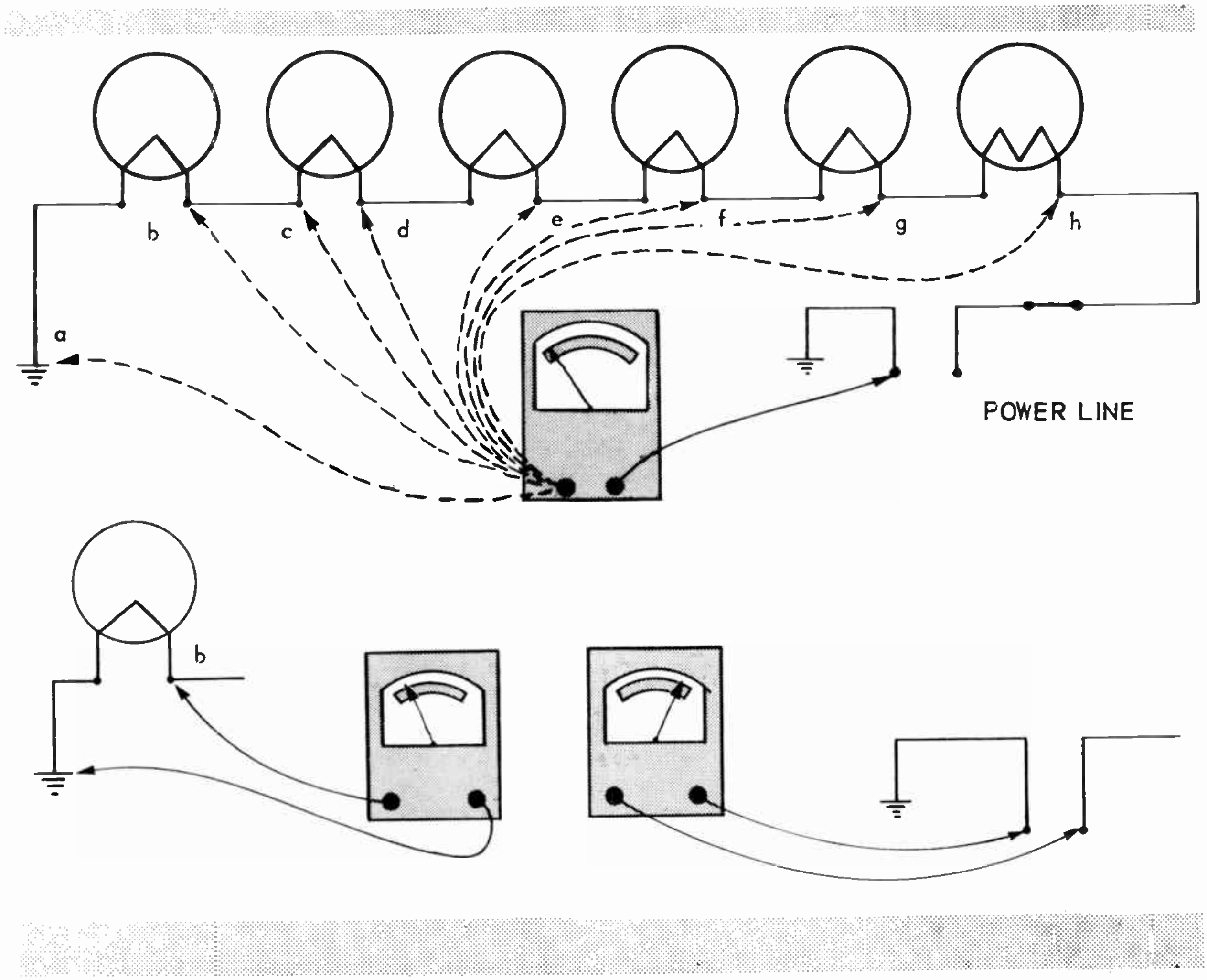


Fig. 6-9. The measured potential difference increases as the meter is connected across more and more heaters.

ground, in the main diagram, is equivalent to a connection at the ground on one side of the heater in the left-hand tube. With the meter connected to point b the reading is 12.2 volts, just as with the connection across the left-hand tube in Fig. 6-8.

The next test connection is to point c. Points b and c are connected together by a wire of negligible resistance. There is no difference of potential in a conductor of negligible resistance. Therefore, the connection at c gives the same voltage reading as the connection at b.

Next we make a connection to point d. Now the testing meter is effectively connected across the heater resistances of two tubes, the combined detector and a-f amplifier, and the converter. The reading is 23.7 volts. This is the sum of the potential differences across these two tubes as shown by Fig. 6-8.

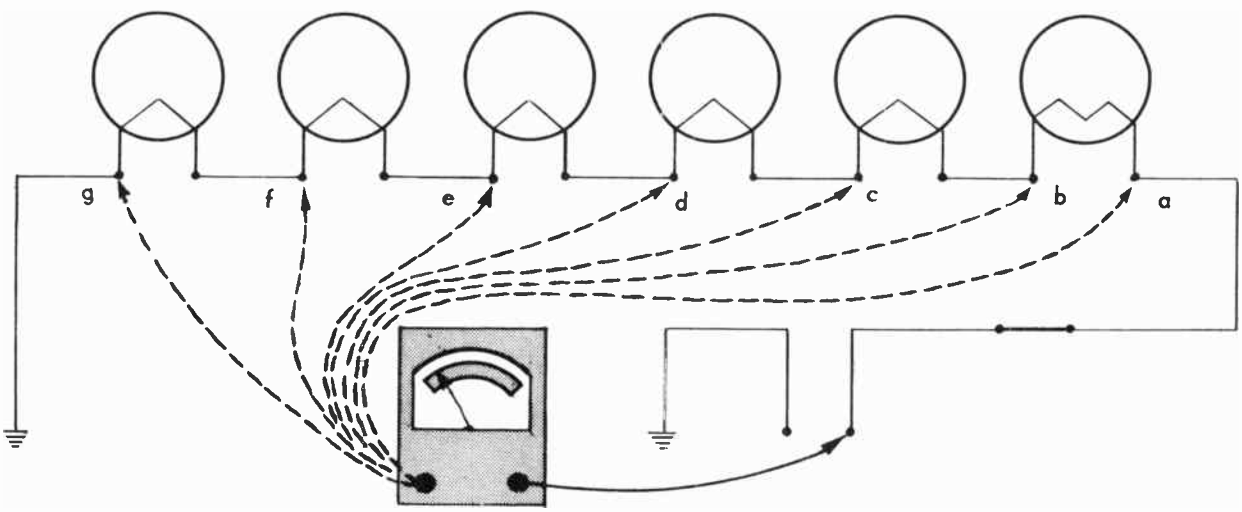


Fig. 6-10. Successive potentials are measured in the reverse order.

Here is a list of voltage readings with the meter connected to each successive point along the heater circuit. The reading at c is omitted because it is the same as at b.

a = 0.0 volts.

b = 12.2 volts.

d = 23.7 volts.

e = 35.8 volts.

f = 48.2 volts.

g = 84.7 volts.

h = 117.0 volts.

Each time the meter connection is changed we bring in the heater resistance of one more tube. Then the voltage reading increases by the potential difference across the added heater resistance. When the meter finally is connected to point h it is effectively connected to the right-hand side of the power line, because in the wires and on-off switch between h and the line there is negligible resistance and negligible difference of potential. We are then measuring power line voltage, just as though the meter were connected as in the small sketch at the lower right.

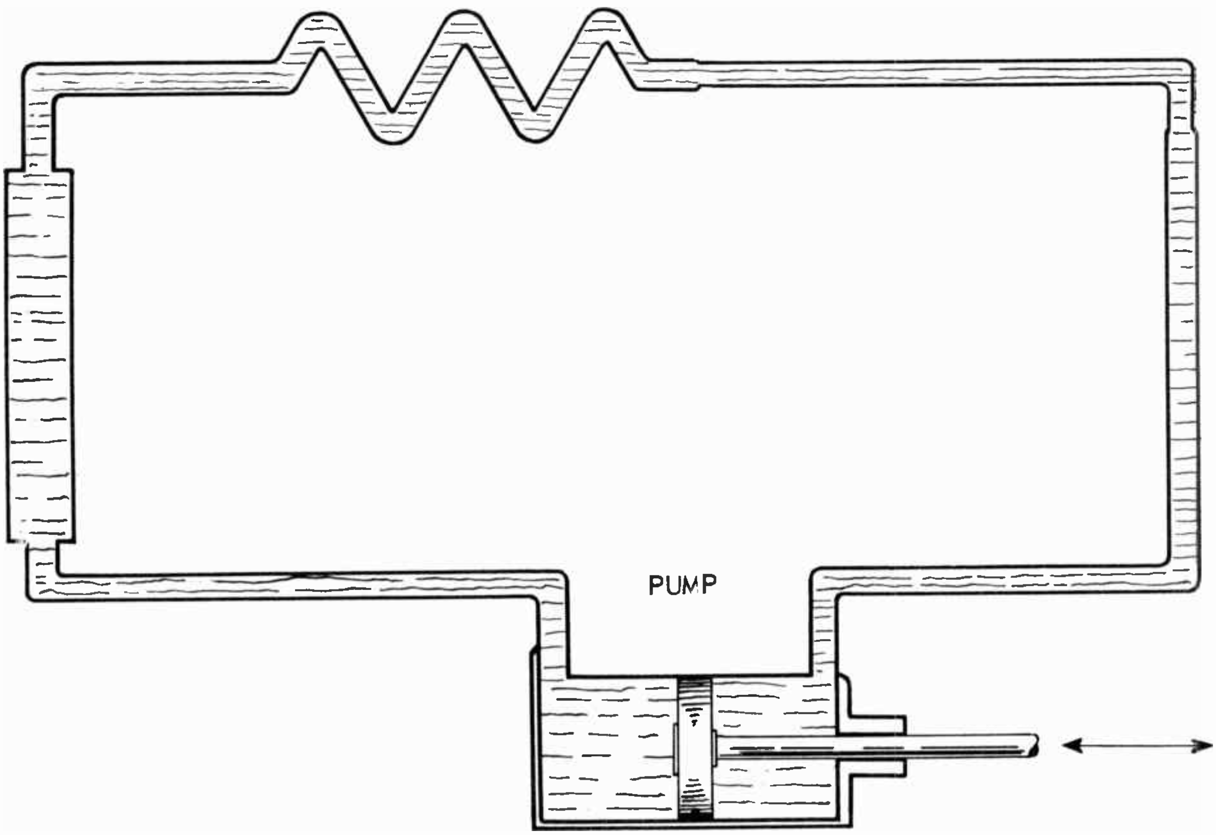


Fig. 6-11. The behavior of water particles in a closed water circuit is much like that of free electrons in a closed electric circuit.

In Fig. 6-10 the voltage tests are made in the reverse order. One terminal of the testing meter now is connected to the "high" side of the power line instead of to the grounded side, and is left there. With the other meter terminal connected to point a the voltage reading is zero, because between a and the high side of the line there is negligible resistance in the wire connections and the on-off switch. Here are the successive voltage readings from point a to point g.

a = 0.0 volts.

b = 32.3 volts.

c = 68.8 volts.

d = 81.2 volts.

e = 93.3 volts.

f = 104.8 volts.

g = 117.0 volts.

With the connection at point b the meter is effectively across the heater resistance of the right-hand tube,

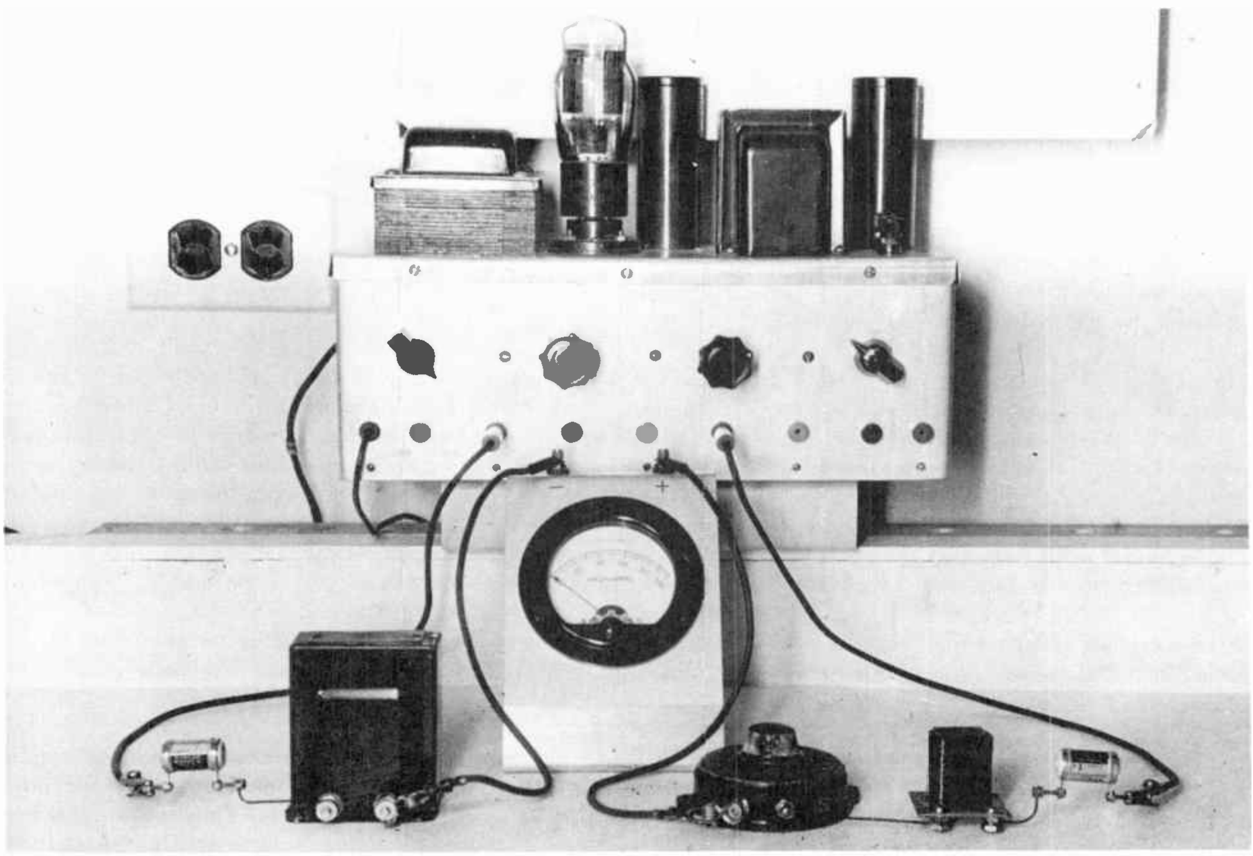


Fig. 6-12. A current meter inserted anywhere in a series circuit will indicate the rate of electron flow which exists everywhere in the circuit.

and we read the potential difference across this heater. This potential difference is the same as shown for the same tube in Fig. 6-8. Then, with each following connection we bring in the heater resistance of one additional tube, and have voltage readings corresponding to potential differences across the sum of all resistances across which the meter is connected. At point g the meter connection is through the wire conductor to chassis ground, through ground to the grounded side of the power line, and through the grounding wire to the line. Now the meter is effectively connected across the power line, and reads line voltage.

HEATER RATINGS. In manufacturer's specifications of tubes are listed the potential differences in volts and the currents or electron flows in amperes at which the heaters are supposed to be operated in order to produce correct cathode temperatures. Here are the heater ratings for the six tubes with which we have been working.

Detector and a-f amplifier	12.6 volts	0.15 ampere
Converter	12.6 volts	0.15 ampere
R-f amplifier	12.6 volts	0.15 ampere

COMMERCIAL TRADES INSTITUTE

I-f amplifier	12.6 volts	0.15 ampere
A-f output tube	35.0 volts	0.15 ampere
Rectifier	35.0 volts	0.15 ampere

The rated operating voltages do not exactly match the measured voltages shown by Fig. 6-8. Voltage is somewhat too high on the heater of the a-f output tube, and somewhat too low on the heaters of all the other tubes. In spite of this, the receiver operates satisfactorily. Actual operating voltages need not be exactly the same as voltage ratings, but the variation preferable is no more than 5 per cent and never should exceed 10 per cent in practice. On our rectifier the voltage is 7.7% low, on the converter it is 8-7% low, and on the a-f output tube it is 4.3% high. Substituting other tubes of the same type numbers would doubtless alter the actual voltages or potential differences.

CURRENT IN A SERIES CIRCUIT. Current ratings are identical for all the tubes, all are rated for operation with heater current or electron flow of 0.15 ampere. Our heaters are connected in what is called a series circuit. In a series circuit all the parts are connected end to end, one after another. Were we to apply a steady one-way potential difference to a series circuit, electrons starting their travel at one end of that circuit would have to proceed through every other part until reaching the far end of the circuit. This assumes, of course, that all parts of the circuit are well insulated or are surrounded by materials through which it is practically impossible for electrons to flow. Then electrons traveling through the circuit conductors and working parts cannot escape, nor can other electrons enter except at the ends of the circuit.

The rate of electron flow must always be exactly the same in every part of a series circuit. That is, the amperes, milliamperes, or microamperes of flow rate in any one part must be identically the same as in every other part of the same series circuit.

The reason for the uniform rate of electron flow in a series circuit may be understood with the help of Fig. 6-11. Here we have a water circuit including pipes of various sizes and shapes, and a reciprocating water pump. All the parts are connected end to end, so this is a series water circuit. This water system is assumed to be completely filled with water. When the piston or plunger of the pump is moved back and forth, water will be forced to move back and forth in every part of the water circuit.

Water must flow in every part of the water circuit because water is as nearly incompressible as anything we know of. With water confined in a closed system, no pressure which can be applied by any ordinary means will reduce the water volume in the least. The pressure might burst the conductors, but it would not compress the water. Consequently, the water flow rate, as measured in cubic feet per second, must be the same in every part of the water system. Speed of the moving water may vary, it will be slow in the bigger pipes and relatively fast in the smaller pipes. But the rate of flow (cubic feet per second) will be the same everywhere. When you send water into one end of this closed water circuit at a rate of so many cubic feet per second, water must come out at the other end at the same number of cubic feet per second. There is nowhere else for the water to go unless it bursts the piping.

Free electrons in a series electric circuit composed wholly of conductors behave like water in the closed water circuit. Every conductor in the electric circuit is as full of free electrons as it can be. The electrons may be moving only in random directions between atoms, but they are there.

Free electrons in a conductor are as incompressible as are water particles in a pipe. To understand why, we must remember that a conductor is a material in which free electrons move very easily from place to place. Were you to push more electrons into one end of a conductor in a series circuit the resulting extra negative charge at that end would instantly repel negative electrons farther along, they would repel electrons still farther along, and so on until, from the far end would emerge the same quantity of free electrons you pushed into the first end. Were you to pull electrons out of one end of a conductor the resulting positive charge at that end would instantly pull other electrons from farther along in the conductor. Into the farthest end would be pulled the same quantity of free electrons you took out of the first end.

The insulation which confines free electrons in the conductors of the electric circuit is like the pipe which confines water in the water circuit. The alternating potential difference which makes free electrons surge back and forth in the conductors of the electric circuit is like the reciprocating pump which makes water surge back and forth in the water circuit.

When the alternating potential difference at one end of the power line becomes something like 10 volts negative it simultaneously becomes 10 volts positive at the other end. The 10-volt negative charge will push electrons into one end of the electric circuit at some certain rate of flow, measured in coulombs per second or amperes. The 10-volt positive charge will pull electrons out of the other end at the same rate. Since free electrons act as though completely incompressible, the flow rate everywhere in the series circuit will be exactly the same as the flow rate at the two ends. Now we have the reason why all the heaters in our series circuit must be rated to carry the same rate of electron flow, in amperes.

HEATER RESISTANCE. In an earlier lesson it was stated that resistance in ohms may be computed when we know the potential difference across the resistance and the rate of electron flow in amperes through the resistance. The rule is exceedingly simple. All you need do is divide the number of volts potential difference by the number of amperes electron flow, and the answer is the number of ohms of resistance. The rule may be shown this way:

$$\text{Resistance in ohms} = \frac{\text{Potential difference in volts}}{\text{electron flow in amperes}}$$

The same rule may be shown in more compact form by using letter symbols for the electrical values. The letter R is the universally recognized symbol for resistance in ohms. The letter E is the symbol for potential difference in volts. The letter I is the symbol for current or electron flow rate in amperes. By using the symbols we may write the rule for resistance like this:

$$R = \frac{E}{I}$$

Let's use this rule to compute the resistance of the heater in the tube at the left in Fig. 6-8, where the potential difference is 12.2 volts and the current may be taken as 0.15 ampere. The easy way to work with decimal numbers is to add enough ciphers to any of them so that all have the same number of numerals at the right of the decimal point, then take out the decimal point, and proceed as in the simplest arithmetic. Here is our example worked out.

$$R = \frac{E}{I} \qquad \text{Putting in the values gives: } R\text{ (ohms)} = \frac{12.2}{0.15}$$

$$\text{Simplifying: } R = \frac{12.20}{0.15} \quad \text{or} \quad \frac{1220}{15} = 81\ 5/15 \quad \text{or } 81\ 1/3 \text{ ohms.}$$

To check the accuracy of our computation we shall take the tube out of its socket, let it cool off, and make a direct measurement of resistance with an ohmmeter. You may be astonished to find that the resistance measures only 12 to 13 ohms. Which is wrong, the formula or the ohmmeter? Neither is wrong, both are correct. The discrepancy comes about because you measured "cold resistance" with the ohmmeter, while

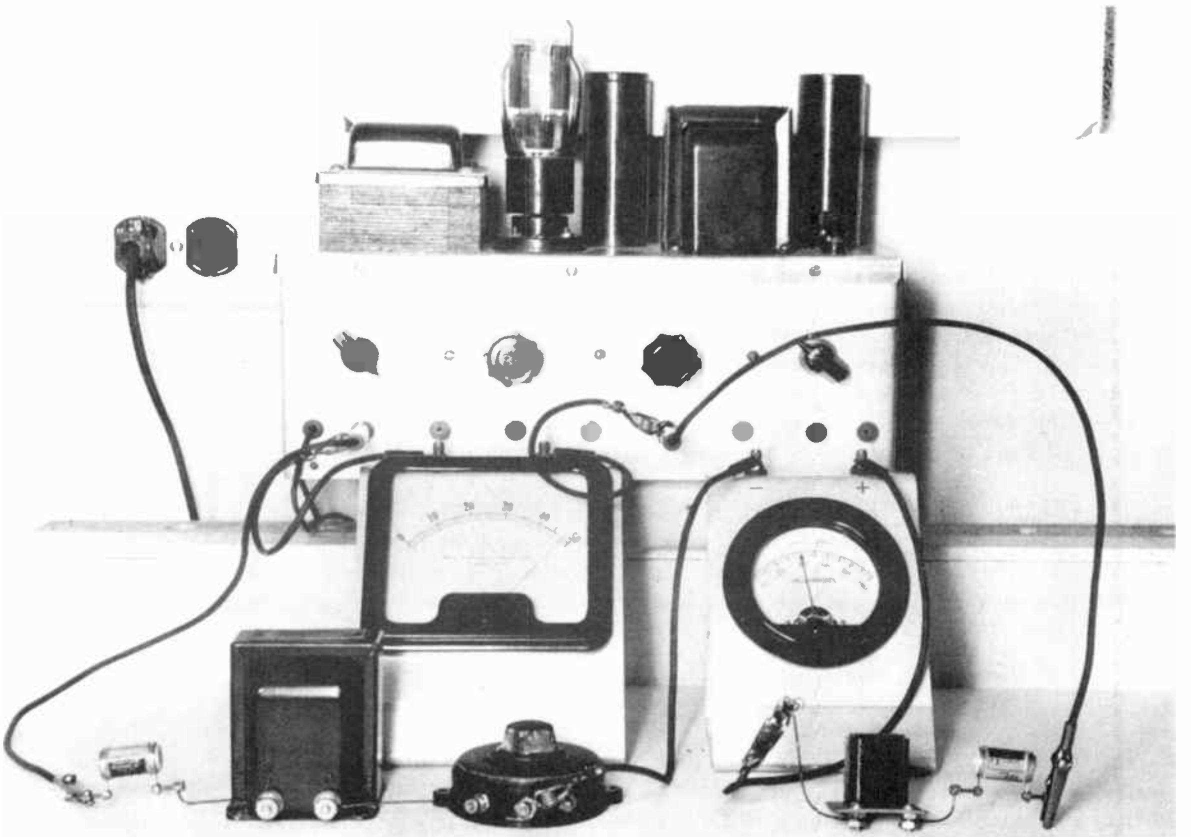


Fig. 6-13. A voltmeter and a current meter allow measurements for computing resistance at actual operating conditions.

the potential difference and rate of electron flow used in the formula are measured while the tube is working and while the heater is very hot.

An ohmmeter furnishes its own potential difference used for measurement of resistance. Inside the ohmmeter is a battery or a power supply. If an ohmmeter is connected across a resistance to which is being applied some other potential difference at the same time, that other potential difference is being applied also to the ohmmeter. The result will be a wholly incorrect resistance measurement. More important, in nearly every case the ohmmeter will be seriously damaged. An ohmmeter should be used for resistance measurements only on parts which are disconnected from their circuits.

If we wish to measure resistance of parts while they are connected into live circuits, and operating normally, it is best done by measuring potential difference and electron flow rate, then using the resistance rule or formula. This method is just as useful with parts operating on direct potentials and with direct currents as with those, like the heaters, which are operating with alternating potentials and currents.

At the top of Fig. 6-12 is a direct-current power supply such as often used for testing in service shops and laboratories. Down below, in a series circuit, are a number of parts. From left to right these parts include a fixed resistor, a filter choke, a current meter, an adjustable resistor, the winding from a transformer, and another fixed resistor. When the power supply is turned on, the meter will indicate the current which flows in all the parts – because current is the same everywhere in a series circuit.

In Fig. 6-13 the voltmeter is connected across the end of the series circuit, which means this meter is connected also across the terminals of the power supply. The power supply has been turned on. The voltmeter indicates 48 volts. The current meter indicates 40 milliamperes, which is 40/1000 of one ampere or 0.040 ampere. To use our rule for resistance we add onto the right-hand end of 48 (volts) as many ciphers as there are numerals at the right of the decimal point in 0.040 (ampere). This gives 48.000. Then we discard the decimal points and divide 48000 by 40, to find that the combined resistance of all the parts in this series circuit is 1200 ohms.

TEMPERATURE AND RESISTANCE. The increase of resistance in the tube heaters, from 12 or 13 ohms when cold to more than 80 ohms when hot, is due entirely to change of temperature in the heater wire. The high temperature of the heater while operating results from work done and energy expended by electrons as they move against the opposition which is called resistance. Then the higher temperature increases the resistance.

The reason why resistance increases with temperature is as follows: What we feel as temperature is due to vibration of the atoms of the substance. In a cold substance the atoms are relatively quiet. When energy in the form of heat is imparted to the atoms they commence to jump about. The more heat and the higher the temperature the more violent becomes the atomic motion. The atoms actually push themselves apart. This is why heated bodies expand. If you heat a piece of iron red hot it gets bigger. Raise the temperature high enough and the atoms get so far apart that the iron becomes molten. At still higher temperature the iron will vaporize.

About resistance. It becomes harder and harder for free electrons to get through a conductor in which the atoms are vibrating more and more violently, just as it would be harder for you to push through a crowd of people jumping in all directions than through a crowd standing still. Resistance increases with temperature because higher temperature means more atomic vibration.

Tables which list resistance of various kinds and sizes of wire are based on a temperature of 68° Fahrenheit (20° centigrade) unless stated otherwise. Resistances at such temperatures usually are called "cold resistances".

Radio and television apparatus always is designed on a basis of "hot resistance", which is the resistance of conductors, and of parts containing conductors, after they have reached normal operating temperatures. In circuits composed wholly of metallic wires and chassis metal the increase of resistance with temperature reduces the electron flow rate or current as the parts warm up.

For resistor wires, and for some other wiring, we may use mixtures or alloys of metals whose resistance decreases slightly when their temperature rises. There are other alloy metals whose resistance remains almost constant with change of temperature either up or down through ordinary operating ranges. With all pure metals, and with most alloys not specifically intended for certain resistance characteristics, resistance increases when temperature rises.

Materials whose resistance decreases with rise of temperature are said to have a "negative temperature coefficient of resistance". Resistors of such material may be used in a circuit consisting otherwise of copper wires or other conductors whose resistance increases with temperature rise. Then the drop of resistance in the "compensating resistors" balances or nearly balances the rise of resistance in other parts, and the overall resistance of the entire circuit remains nearly constant with change of temperature.

The fact that some materials have negative temperature coefficients hardly fits with our explanation of why resistance increases when higher temperature makes atoms vibrate more strongly and impede electron flow. It is thought that in these negative temperature materials many electrons are so loosely held by the atoms that increase of electron energy due to heat causes immense quantities of these electrons to become free electrons. The added quantities of free electrons more than make up for the greater opposition to flow through the vibrating atoms, with the result that electron flow increases. More flow means less resistance.

Our measurements of voltages and resistances in the heater circuit have brought out many facts which will be useful in all work to come. In the following lesson we shall move up to the plate circuit of the audio output tube. This circuit includes not only the output tube but also the rectifier tube, part of the d-c power supply, and part of the loud speaker. Needless to say, we shall run into many highly interesting and worth while methods of service testing.

TELEVISION

LESSON NO. 7 SERVICE TESTS IN A PLATE CIRCUIT

"Training Your Mind"

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind

C.T.I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

LESSON NO. 7

SERVICE TESTS IN A PLATE CIRCUIT

Now we are ready to follow electron flow in a path quite a bit more intricate than that for the tube heaters. This new path is the plate circuit for an audio output tube, the tube which furnishes signal power to operate the loud speaker. When working with the heater circuit we dealt with only alternating potential differences and alternating current or electron flow. In the plate circuit we shall be dealing with direct or one-way potential differences and with direct current or electron flow.

This does not mean that we have only direct potentials and currents in a plate circuit, for signal voltages appear in plate circuits, and they alternate at various frequencies. It does mean that we require direct potentials to keep free electrons traveling from cathode to plate inside the tube. Then signal voltages vary the rate of this electron flow in ways which we shall talk about when getting into the subject of signal amplification in tubes. In this lesson we are going to make measurements of only the direct potentials. This is one of the first things a competent technician does when commencing to look for elusive troubles in a receiver.

The receiver on which our measurements will be made is pictured in Fig. 7-1. It will serve our present purposes as well as the biggest television receiver, for the plate circuit of the output tube in this simple radio set is just about as tricky as anything you will come across. The audio output tube is the one just behind the speaker, the tube with a glass envelope. In the plate circuit of this tube is included part of the speaker coupling transformer or output transformer, which you can see on top of the speaker. Also in the plate circuit is the rectifier tube located away over at the left-hand corner, and the filter choke which is immediately at the right of the rectifier tube. The power supply for this plate circuit is the a-c power line which enters the chassis just below the filter choke. Finally, on the front of the receiver there is the on-off switch that will get itself into our plate circuit.

Fig. 7-2 shows all the larger parts of the plate circuit taken out of the receiver and connected together in essentially the same way as when working, except that here the connections between parts are shorter. Were one electron to travel all the way through our plate circuit, this electron would go from one side of the power line into chassis ground, and through ground to the free end of the resistor which is at the extreme left in the picture. When mounted in the receiver this free end of the resistor is connected to chassis ground.

Then the electron would go into the output tube, from cathode to plate inside this tube, and from plate to the speaker coupling transformer. After going through the transformer the electron would pass through the filter choke to the rectifier tube, would go from cathode to plate inside the rectifier, and from there to the ungrounded side of the power line. That's all there is to the plate circuit, so far as direct potentials and currents are concerned. But it is astonishing how complicated a plate circuit can appear when it is interconnected with a lot of other circuits in a receiver.

Now to answer your question, for without a doubt you have noted that we connect the plate circuit across

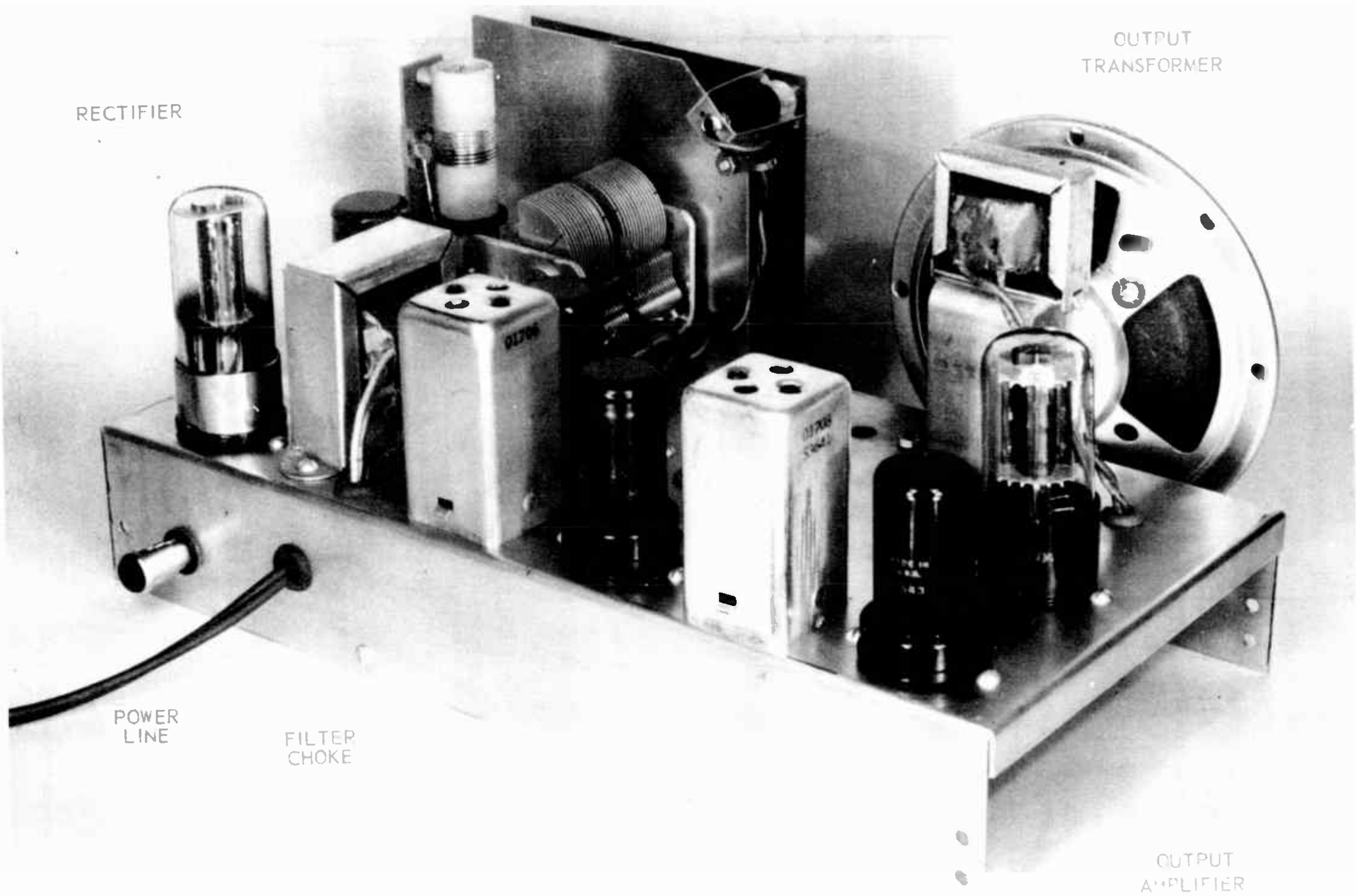


Fig. 7-1. Many parts which carry plate current for the output tube are on top of the chassis, but the wiring connections are underneath.

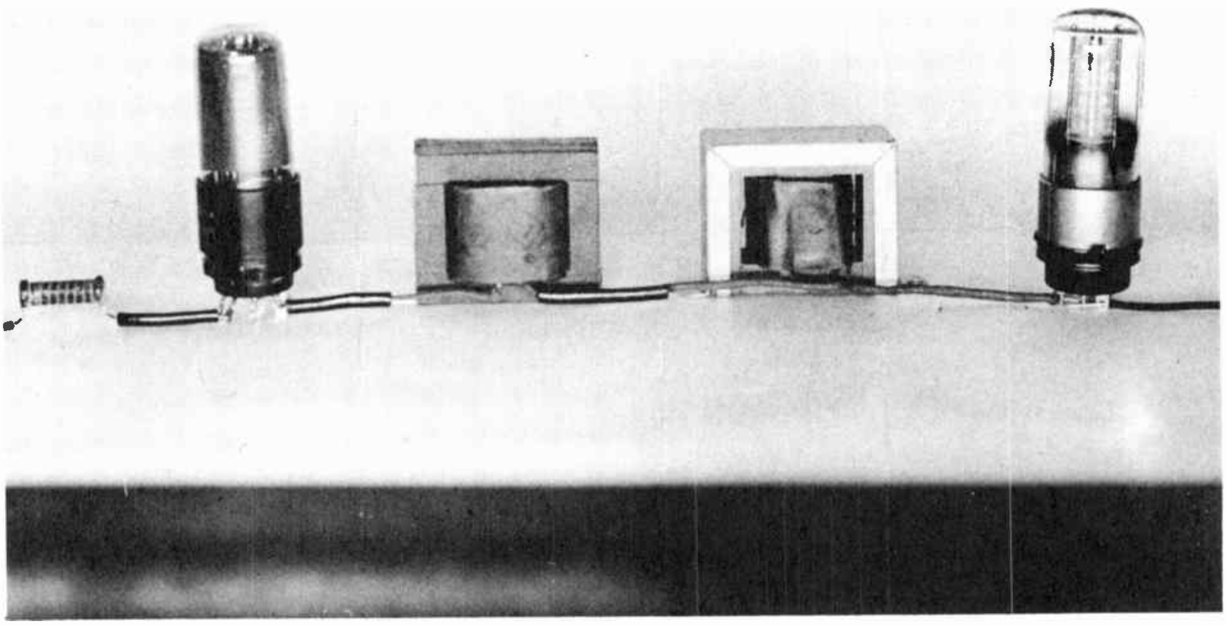


Fig. 7-2. The principal parts of the plate circuit as they would appear when taken out of the receiver.

a power line furnishing only alternating potential, yet talk about having direct potential and direct current in the plate circuit.

THE RECTIFIER TUBE. It is the rectifier tube that delivers direct current or direct one-way electron flow when alternating potential is applied to this tube. Later we shall have more to do with rectifiers and their operation, but the elementary principle is so simple that we may as well examine it right here.

At the top of Fig. 7-3 a rectifier tube is shown by its symbol. Inside the rectifier are only a plate, a cathode, and a heater to raise the cathode temperature so electrons may be emitted from the cathode. This is the sole purpose of the heater, so we shall forget it from here on. The same explanation applies to the heater in the audio output tube, so we won't even show the heater in diagrams.

The rectifier plate is directly connected to one side of the power line. The other side of the power line connects through the load to the rectifier cathode. Remember, a load is any conductive path in which electrons may flow and do work.

Electrons emitted from the rectifier cathode are negative. If the plate is made more positive than the cathode, these emitted electrons will be attracted to the plate and will enter the plate. If the plate is made more negative than the cathode, the negative electrons will be repelled by the negative plate and will neither go to nor enter the plate. Electrons can flow through the space within a tube only while the plate is more

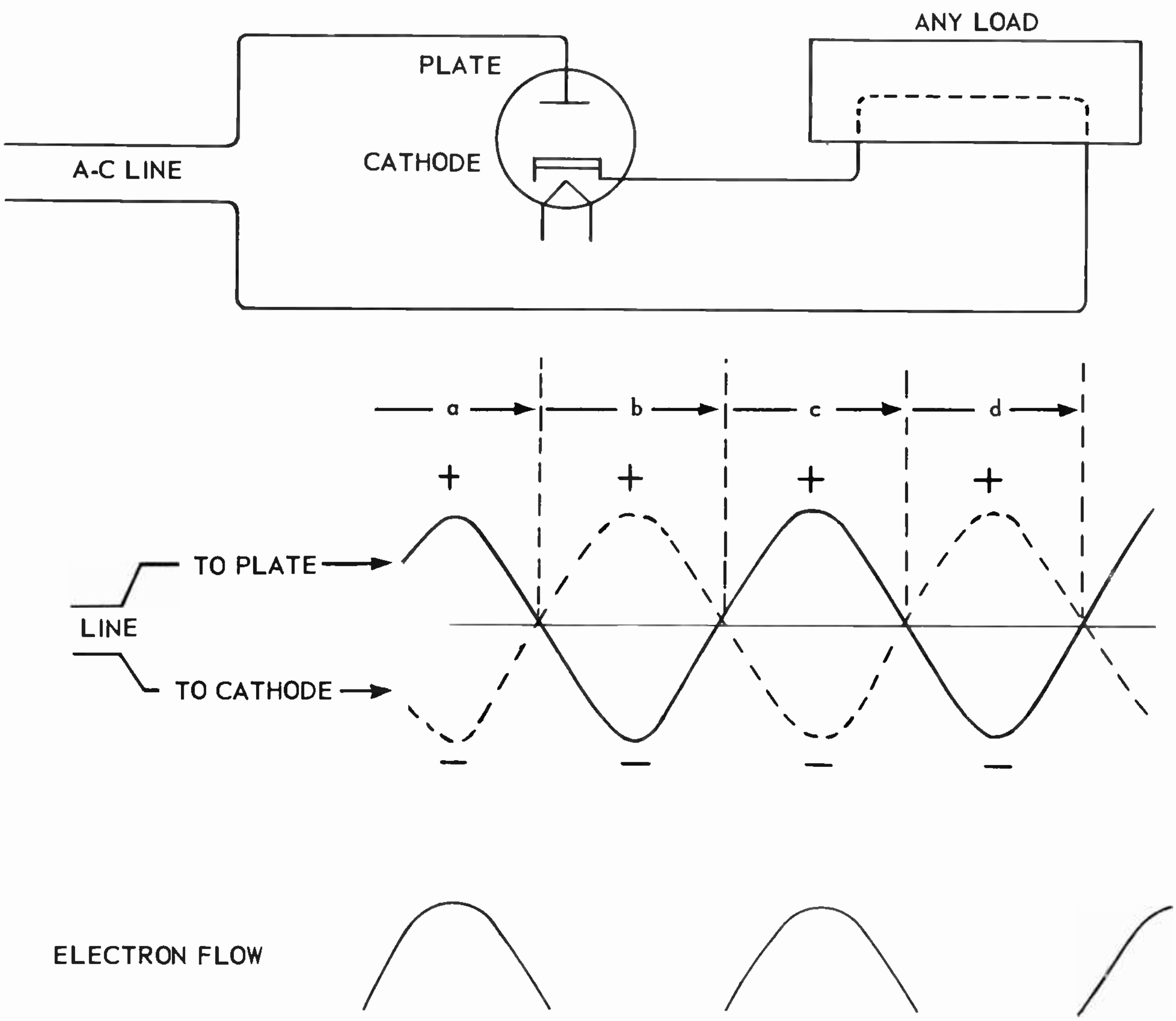


Fig. 7-3. How a rectifier delivers direct current when alternating current is applied.

positive than the cathode. Be sure you get this straight, it is one of the most important facts in the whole science of radio and television. With some of the rectifiers in television sets you could make the plate 30,000 volts more negative than the cathode with perfect safety (to the tube) and without getting a trace of electron flow in the reverse direction.

Now look at the middle diagram of Fig. 7-3. Here are shown the variations of alternating potentials existing in the two sides of the power line. One side of the line is connected to the plate and the other side to the cathode of the rectifier, as in the upper diagram. During the half-cycle marked "a" the rectifier plate is made positive and the cathode negative. Will there be electron flow in the rectifier? Yes, of course there

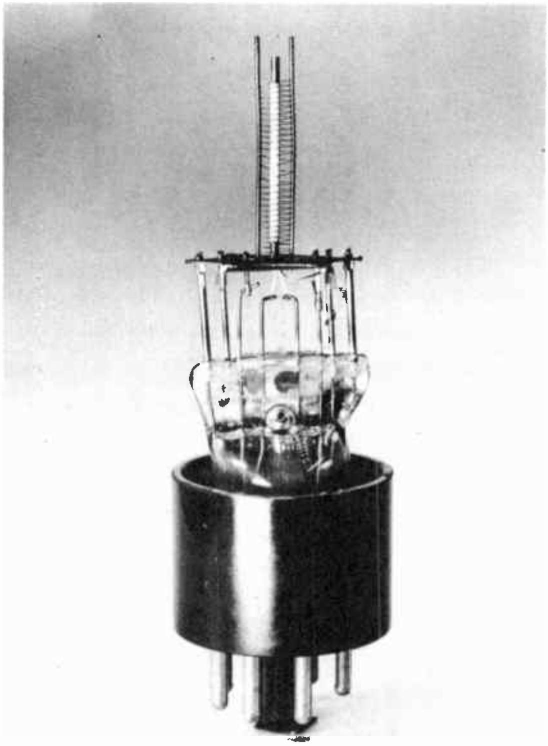


Fig. 7-4. The control grid is an open spiral of wire around the cathode.

will be. During the following half-cycle, marked "b", the plate is negative and the cathode positive. Will there be electron flow? Not a trace. Then there will be electron flow during half-cycle "c," no flow during half-cycle "d," and so on, and on, and on.

During every half-cycle in which the plate is positive and the cathode negative there will be electron flow from one side of the power line through the load to the rectifier cathode, from cathode to plate inside the tube, and from plate back to the line. During every intervening half-cycle there will be no flow at all. Electron flow will occur in intermittent pulses, as shown at the bottom of Fig. 7-3. All the pulses will be of flow in the same direction, because flow can occur in only one direction through the rectifier. Thus we have direct electron flow or direct current from an alternating potential.

THE FILTER CHOKE. It is better to have a steady, smooth flow of direct current rather than the pulsating flow. The filter choke does a lot to smooth out the pulsations. A choke consists of nothing more than some iron, called the core, around which are many turns of insulated copper wire. Free electrons may flow in this copper wire when there is a potential difference across the choke.

From the manner in which a choke affects electron flow we may think of the choke as a sort of electrical

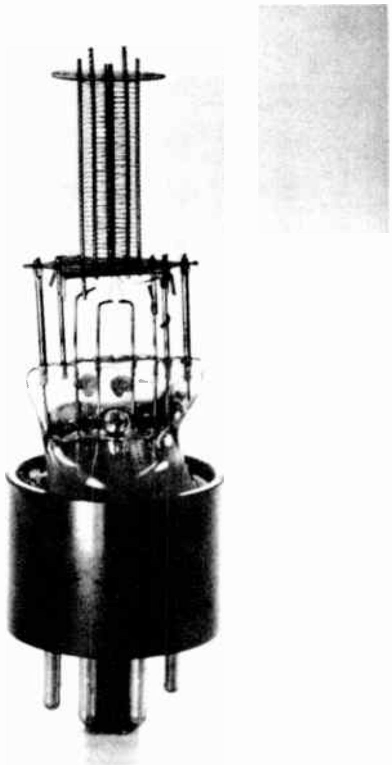


Fig. 7-5. Around the outside of the control grid is another spiral of wire, the screen.

weight. Imagine yourself holding a weight of maybe a couple of hundred pounds. You couldn't start running very suddenly, but once under way it would be hard to stop you. That's what the choke does to the electron flow. The rate of electron flow cannot become so great during a half-cycle, but at the end of that half-cycle the electrons cannot stop very quickly and the flow carries on into the time of the following half-cycle during which otherwise there would be no flow at all. Now we have talked about our first inductor, for the filter choke is one of the many inductors used in all television and radio receivers.

There are other parts in the filter system. They are capacitors not shown in our pictures nor diagrams. These capacitors help keep the electron flow going during half-cycles in which the potentials are reversed. They don't keep flow going through the rectifier, they store up electrons while there is a strong flow, then send these electrons on through the circuit while the rectifier furnishes no flow. The capacitors are like storage tanks for electrons. However interesting it might be to continue talking about filter systems, just now we must get back to measuring potentials and currents in the plate circuit.

ELECTRON FLOW IN A TUBE. We have talked about a rectifier tube in which the only active elements are a cathode and a plate. The heater is not counted as an active element, because it takes no part in carrying or controlling the electron flow that passes through the evacuated space within the tube envelope.

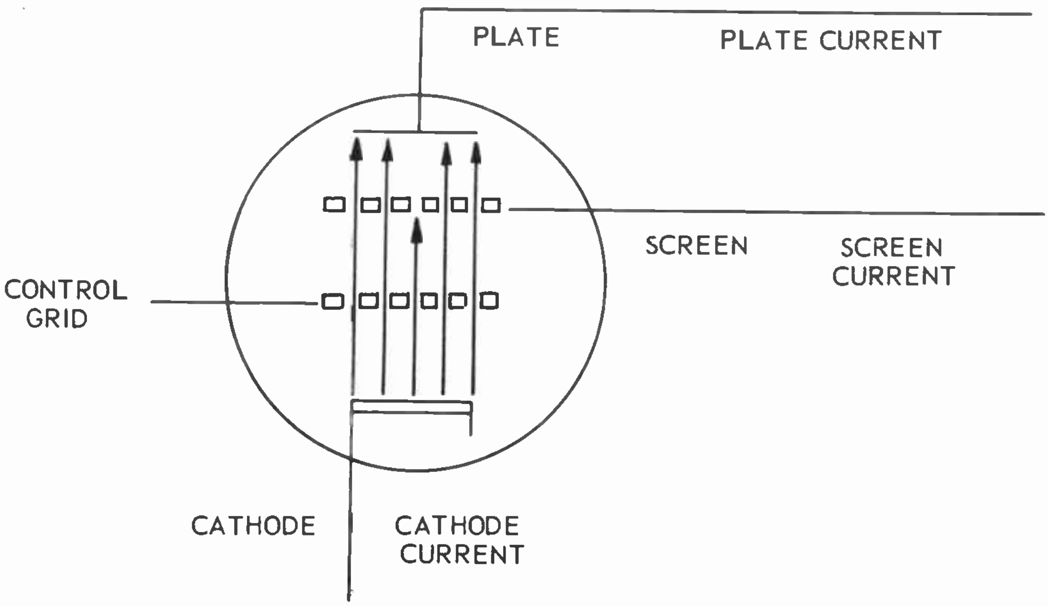


Fig. 7-6. Cathode current divides within the tube into plate current and screen current.

We have talked also about amplifier tubes in which the active elements include a cathode to emit free electrons, a control grid to regulate the rate of flow of these free electrons through the tube, and a plate to receive the electrons and pass them on to the external circuit. But in the output amplifier tube on which we are to make measurements, and in a great many other amplifiers, there is a fourth active element. It is called the screen grid or, more commonly, just the screen. One purpose of the screen is to help pull free electrons away from the cathode and speed them on their way to the plate. The only reason for mentioning the screen just now is because we shall need to measure screen voltage, the potential at the screen lug on the tube socket. How the screen does its work will appear a little later.

Were we building up a tube which is to have a screen we should begin, as in Fig. 7-4, by placing around the cathode the spiral of wire which is the control grid. Next, as in Fig. 7-5, we should place around the outside of the control grid another wire spiral of greater diameter, but otherwise quite similar. This is the screen. Around the outside of the screen would later be placed the plate, which is an open-ended cylinder of thin metal having no openings through its sides.

In Fig. 7-6 is a symbol for a tube containing a screen. The screen is between the control grid and the plate inside the tube. The screen, like the plate, is maintained positive with reference to the cathode in order that the screen may do its work of pulling negative electrons away from the cathode. Electron flow inside this tube is shown by vertical arrows. All electrons drawn away from the cathode go through the spaces between turns of the grid wire. Most of these electrons are then traveling so fast that they go right on through the spaces between turns of the screen and reach the positive plate. There are no openings in the plate sides, so all the electrons stop there. A relatively small fraction of the flying electrons get caught by the positive screen instead of going on through to the plate.

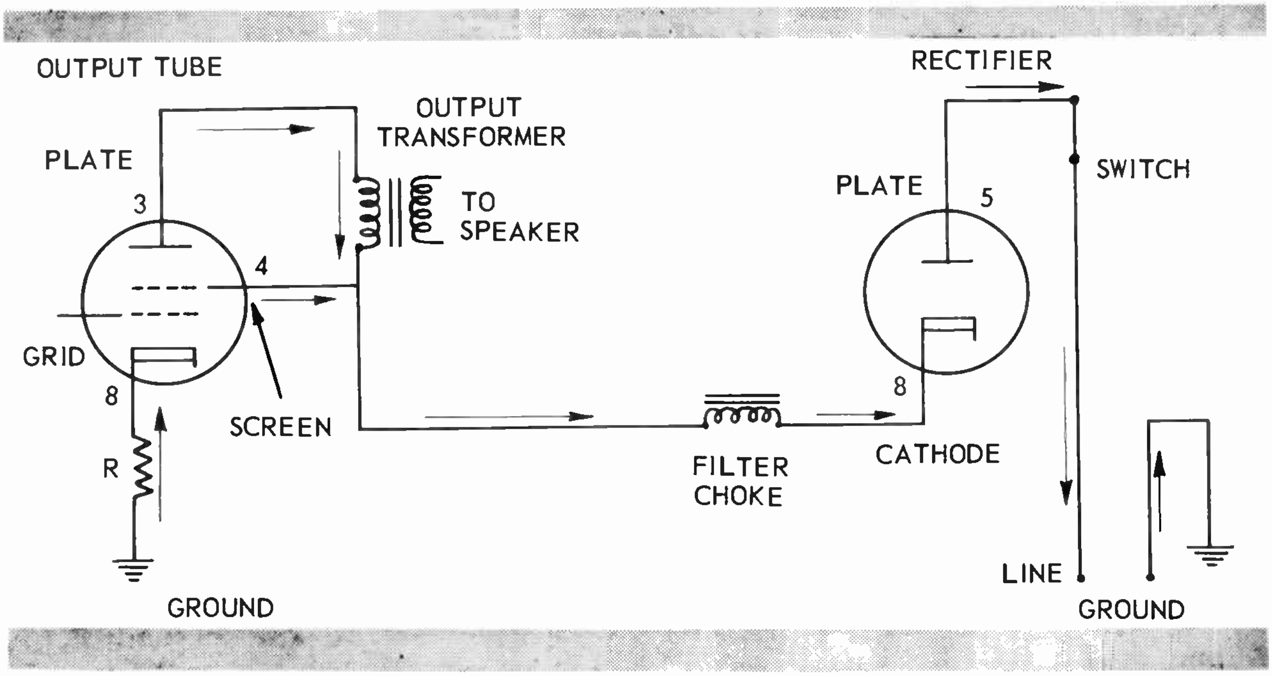


Fig. 7-7. Paths of electron flow in the plate circuit.

In the external parts of the circuit leading away from the plate are flowing all the electrons which reached the plate and entered it. This flow is the plate current. In the external circuit leading away from the screen are flowing all the electrons caught by the screen. This flow is the screen current.

Now for something of importance. All free electrons which leave the cathode and go both to the plate and the screen must be replaced by free electrons flowing into the cathode from the external circuits. Were the emitted electrons not continually replaced, the loss of electrons quickly would leave the cathode so highly positive that no more free negative electrons could be pulled away from it. All the electrons which form both the plate current and the screen current must come from the cathode. Therefore, enough electrons must continually flow into the cathode to equal the sum of the plate current and the screen current. This total flow is the cathode current. Cathode current must equal the sum of plate current plus screen current, as measured in amperes, milliamperes, or microamperes.

ELECTRON FLOW IN THE PLATE CIRCUIT. Let's get acquainted with how electrons flow in the plate circuit before commencing actual measurements of voltages. Fig. 7-7 is a schematic diagram of the plate circuit. Directions of electron flow are shown by arrows. Starting from the side of the power line connected to ground, the flow is through chassis metal to the ground connection below the output tube. Then electron flow is upward through resistor R to pin 8 and the tube cathode.

Most of the electron flow now goes through the tube to the plate and to pin 3, then downward through one winding of the output transformer. When there is a signal this transformer will transfer the signal variations of voltage from the plate circuit to the loud speaker. Now, however, we are interested in only the unvarying electron flow which exists without a signal. Part of the electron flow from the cathode inside the output tube goes into the screen and leaves the tube by way of pin 4, to join the flow that has come from the plate through the transformer winding.

The total electron flow goes next to the left-hand side of the filter choke, through the winding of insulated wire in the choke, and to pin 8 on the rectifier tube. This pin is connected to the rectifier cathode. Electron flow is from cathode to plate inside the rectifier, from plate to pin 5, thence through the closed on-off switch and to the ungrounded side of the power line.

MEASURING VOLTAGES. Fig. 7-8 is a wiring diagram of our plate circuit. It is similar, as far as it goes, to service wiring diagrams available for many receivers. Connections are shown as they appear from underneath the chassis. We are looking at the bottoms of tube sockets. Parts which actually are mounted on top of the chassis are here shown with broken lines. These parts include the speaker, the output transformer, and the filter choke. Wires shown going to these parts pass up through holes in the chassis.

Voltages are marked at each of the socket lugs at which readings are taken. Voltages are shown at the cathode, plate, and screen of the output tube, and at cathode and plate of the rectifier tube. Values are given in volts and tenths of volts. These are actual voltages measured on the real receiver. Service diagrams do not always give voltages to tenths, for there may be some variation one way or the other in some circuits without materially affecting performance. Voltages on service diagrams are averages taken from many receivers.

The path of electron flow for connections in the wiring diagram is exactly the same as on the earlier schematic diagram. The schematic shows connections in the simplest manner, so circuits may be most easily traced. The wiring diagram shows connections as actually made in the receiver. Many points of connection which appear on the wiring diagram and not on the schematic are used solely for ease of original wiring, and to make possible a replacement of any one part with least disturbance to connections of other parts.

Two of these convenience wiring connections appear in Fig. 7-8. One is the terminal strip to which a wire is run from lug 4 on the socket for the output tube. Such a strip consists of a piece of insulation on which are one or more soldering lugs or terminal screws, and some means for attaching the insulation to chassis metal. The lugs or terminals are insulated from chassis metal unless they are wanted for ground connections, then one of the lugs may make metallic contact with the mounting and with chassis metal. The particular need for the terminal strip in this layout is to provide a point from which wires (not shown) may be run to plate circuits and screen circuits of other tubes in the receiver.

Another convenience connection is at lug 4 on the rectifier socket. There is no connection from this lug to any internal parts of the tube. If you look at the schematic diagram of Fig. 7-7 you will see that neither the plate nor the cathode is connected to pin 4. Earlier we observed that the heater for this tube is connected between pins 2 and 7. Any socket lugs from which there are no connections into the tube may be used just like an insulated lug on a terminal strip.

The reason for using socket lug 4 in this layout is that the lead coming out of the filter choke is not long enough to reach all the way to the terminal strip. But this lead easily reaches to socket lug 4. Then run another wire from this lug to the terminal strip. The other lead from the filter choke is soldered to lug 8 on the rectifier socket. By taking the choke leads off lugs 4 and 8 it would be easy to remove and replace this unit without disturbing any other part.

All voltages shown on Fig. 7-8 are potential differences between chassis metal and the points indicated. It is usual practice on service diagrams to give all voltages with chassis ground as the reference point. Then all you need do is connect one lead from your testing meter to chassis metal and leave it there while

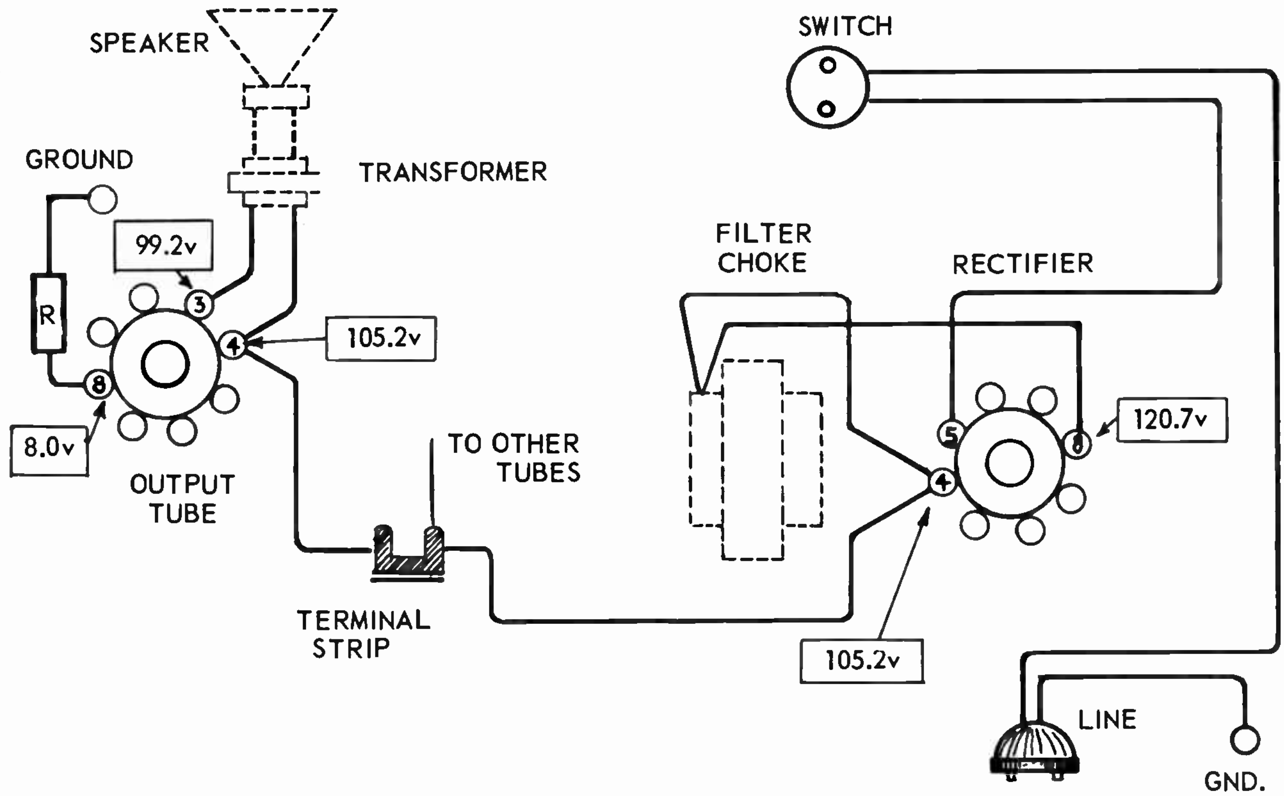


Fig. 7-8. Voltages at test points in the plate circuit.

applying the other meter lead to each point where voltage is to be measured.

The same voltages are shown on a diagram easier to follow in Fig. 7-9. Ground must be the most negative point in this circuit because, as we have seen from Fig. 7-7, all electron flow is away from ground, and electrons can flow only from negative to positive. Then all other points in the circuit must be positive with reference to ground. As we move away from ground, from left to right in Fig. 7-9, the positive potentials become greater and greater. Positive voltage becomes maximum at the rectifier cathode, where it is 120.7 volts.

We make no measurements of direct potentials or potential differences beyond the rectifier cathode, even though the plate circuit continues through the rectifier, the switch, and to one side of the power line. Were we to make a test at the rectifier plate with a voltmeter designed for measuring direct potentials, the meter would be connected across the alternating potential of the power line. This comes about because the meter lead which would be connected to the rectifier plate, at socket lug 5, would be connected through the switch to one side of the line. The other lead from the meter would still be connected to ground, and through ground to the other side of the power line.

Meters designed for measuring differences of direct potential will show a zero reading when connected to

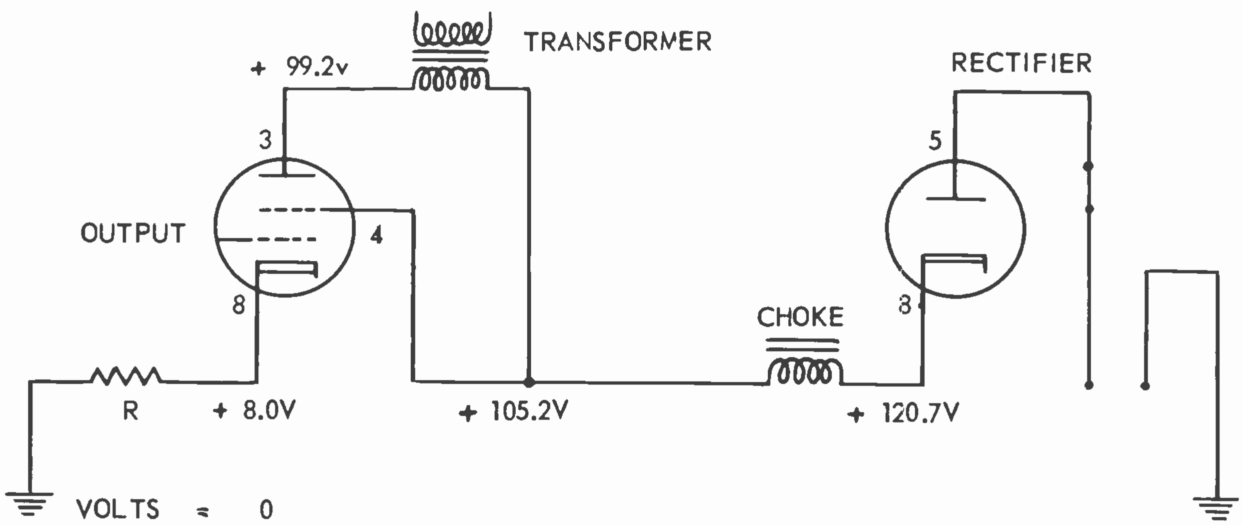


Fig. 7-9. Potential differences across various parts may be determined from voltages read at the test points.

alternating potential, or the pointer of the meter may vibrate. If the maximum alternating potential difference is greater than the maximum direct potential which the meter will withstand, the instrument will be burned out. Many voltmeters designed for measurement of alternating potential differences will give a reading when connected to direct potential, but the reading will be incorrect. Other alternating potential meters or "a-c voltmeters" will show one jump of the pointer and thereafter will read zero.

Probably you have noticed that our maximum positive direct voltage is greater than the effective alternating potential, 117 volts, from the power line. In this connection we must remember that the maximum potential difference of the alternating potential is greater than the effective potential. With 117 effective volts the maximum will be 165.5 volts. Capacitors connected to the filter choke help bring the direct potential above the effective alternating potential, closer to the maximum value of alternating potential. That is why the maximum direct potential is higher than the effective alternating potential from our source.

Now to explain something which might be a little confusing. Note that the potentials shown by Fig. 7-9 become greater and greater as we progress from negative (ground) toward positive at the rectifier cathode. Yet we have learned that electron energy is maximum at the negative end of a circuit outside the power supply, and is zero at the positive end. The confusion is the result of a hangover from the days when everyone thought that electricity flows from positive to negative. That's when voltmeters were first designed and used. Most voltmeters still are made to show potentials increasing in the wrong direction.

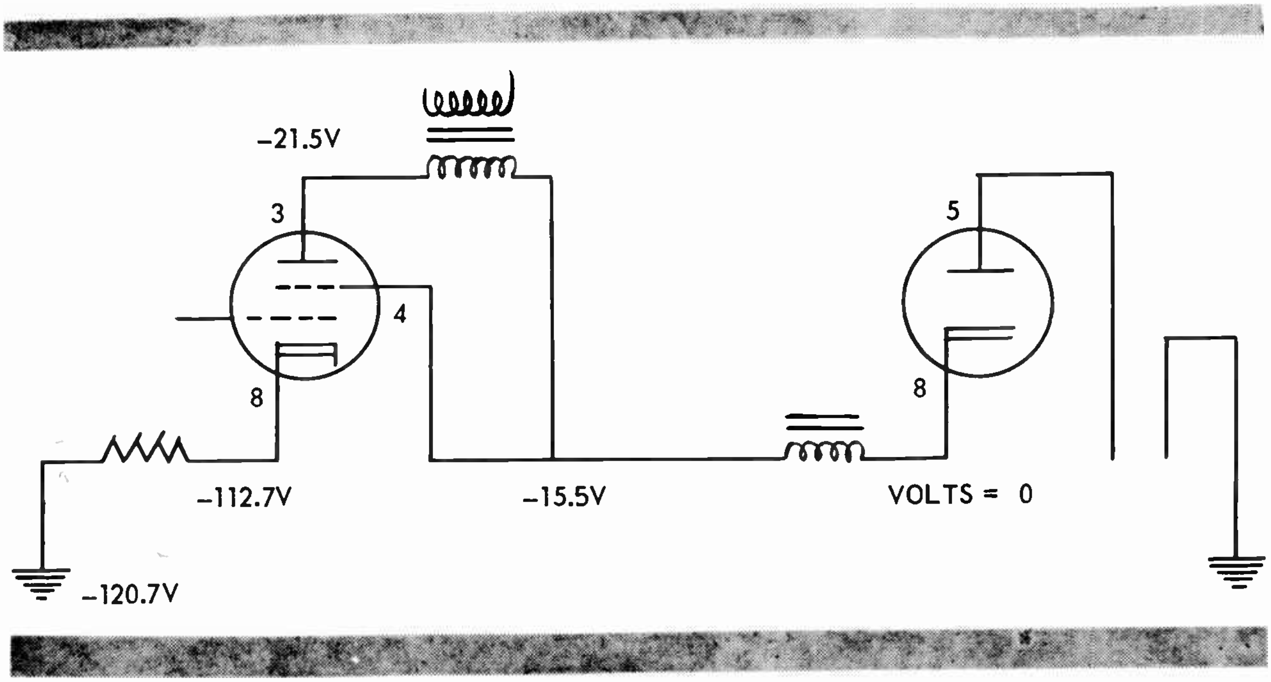


Fig. 7-10. Potentials measured with reference to the rectifier cathode correspond to relative values of electron energy.

It is just as easy to build a voltmeter which gives readings corresponding to electron energy as to build it in reverse. There are some types of voltmeter which will read either way. Some of these meters have the zero mark at the center of their dial scale, with voltages increasing both ways from zero. These are zero-center meters. Many of our electronic voltmeters used for service work have a switch that allows reading voltages either one way or the other. Meters which read increasingly positive from zero may be called positive reading voltmeters. Those which read increasingly negative from zero may be called negative reading meters.

Were we to make measurements with a negative reading voltmeter, its positive terminal would be connected to the rectifier cathode and left there. Then the lead from the negative terminal of the meter would be connected successively to points extending through the circuit until reaching ground. The voltages would be as shown by Fig. 7-10. Differences of potential across each individual part of the circuit are just the same as differences when measuring the other way, but they add up from positive to negative instead of from negative to positive.

No matter how correct may be the method of Fig. 7-10 from the electronic standpoint, there is no use of learning to make routine measurements this way. It just isn't done. All service literature shows voltages increasing positive as we move away from chassis ground into the plate circuits and screen circuits.

At first glance it might appear from Fig. 7-10 that we are trying to operate the output amplifier with its plate negative. The plate is marked –21.5 volts. But the cathode is 112.7 volts negative. The cathode is far more negative than the plate. This amounts to the same thing as having the plate far more positive than the cathode, which it really is. It's all a matter of what reference point we select. In Fig. 7-10 the reference point called zero is the rectifier cathode, and everything else is negative with reference to this point. In Fig. 7-9 the zero reference point is chassis ground, and everything else is positive (or less negative) than this point.

Don't make the mistake of passing too lightly over this matter of negative plates and negative screens. In television receivers you will find plenty of them. But always the cathode of the same tube will be still more negative, which makes the plates and screens effectively positive with reference to the cathodes.

COMPUTING PLATE CURRENT AND SCREEN CURRENT. Fig. 7-11 is still another diagram of our plate circuit. We have gone back to the usual method of marking voltages. To this diagram have been added the values of resistance in ohms of resistor R, of the plate winding in the output transformer, and of the filter choke winding. Resistances of the wire conductors between various parts are so small as to have no effect on measurements, so we shall neglect these resistances.

The potential difference across any of the circuit units is easily determined by subtracting the lower voltage shown for one end from the higher voltage shown for the other end of the same unit. Here are the potential differences, also the resistances across which each potential difference is acting.

Resistor R	8.0	–	0.0 =	8.0 volts.	180 ohms.
Transformer	105.2	–	99.2 =	6.0 volts.	147 ohms.
Choke	120.7	–	105.2 =	15.5 volts.	210 ohms.

Knowing the difference of potential between any two points in a circuit, and the resistance between the same two points, it is a matter of only the simplest arithmetic to compute the current which must be flowing between the same two points. It is easily possible, with only the values in the preceding list, to compute the plate current of the output tube, the screen current of this tube, the cathode current of this tube, the current taken by all the other amplifier tubes connected to the terminal strip (Fig. 7-8), and the direct current through the rectifier tube.

To make the computations we need a rule or formula which gives current when we know potential difference and resistance – we must know the effect of potential difference on current, and the effect of resistance on current. Consider first the effect of potential difference.

At A in Fig. 7-12 is a resistance connected between charges or terminals whose potential difference is 1 volt. Electrons at the negative terminal possess energy proportional to 1 volt of potential. Let's say that this much energy lets the electrons get through the resistance to the positive terminal at such speed that 1 coulomb of them come through during every second. Then the flow rate is 1 ampere, which is a rate of 1 coulomb per second.

At B in the figure we have increased the potential difference to 2 volts. Electrons now leave the negative terminal with twice as much energy as before. With twice the energy the electrons can do work and overcome opposition twice as fast, and it takes them only half as long to get through the resistance. Then twice as many electrons will come through the resistance during each second. The flow rate will be 2 coulombs per second, or 2 amperes.

At C the potential difference has been reduced to 1/2 volt. With only half the earlier amount of energy it takes the electrons twice as long to get through the resistance, so only half as many will come through

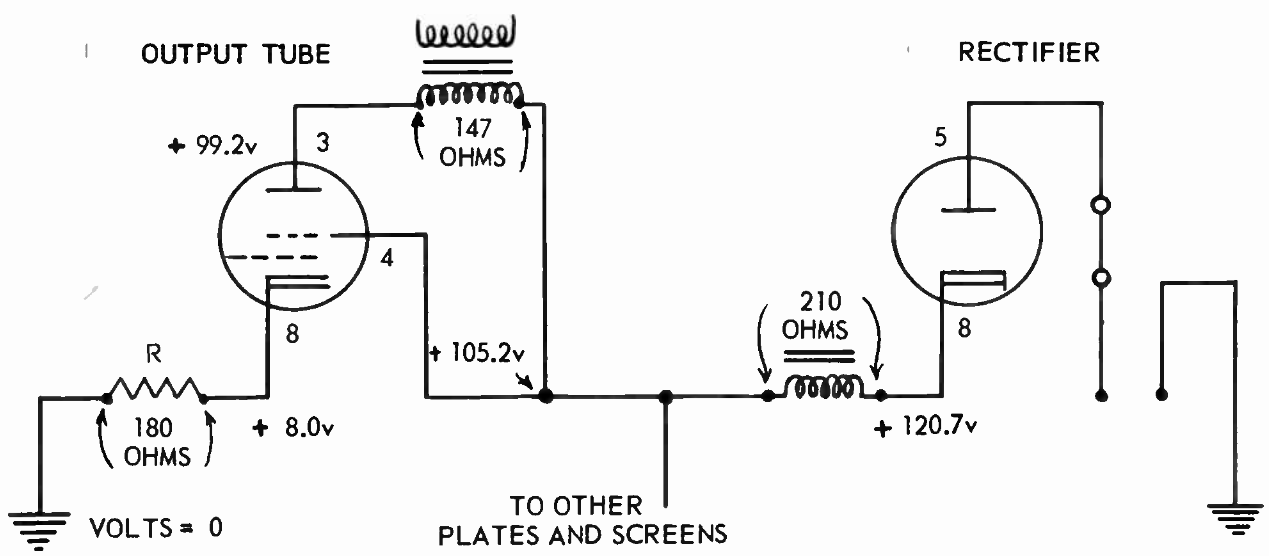


Fig. 7-11. Potential differences and resistances allow computing currents in all parts of the circuit.

during one second. The flow rate now will be 1/2 coulomb per second, or 1/2 ampere.

We have seen that electron flow rate is directly proportional to potential difference. One volt causes flow of 1 ampere, 2 volts cause flow of 2 amperes, 1/2 volt causes flow of 1/2 ampere. This direct relation between volts and amperes exists because resistance remained unchanged in all the examples. This resistance, shown on diagram D of Fig. 7-12, must have been 1 ohm, because potential difference is 1 volt, current is 1 ampere, and $R = E/I$.

In diagram E we still have the same potential difference as at D, but have changed the resistance from 1 ohm to 2 ohms. Electrons starting from the negative terminal still have only the energy corresponding to 1 volt of potential, but they have twice the opposition or resistance to overcome in going from negative to positive. The electron speed will be only half as fast as before, only half as many will come through during a second, the flow rate will be 1/2 coulomb per second, which is a rate of 1/2 ampere.

At F the resistance has been halved from its original value and the potential difference has been kept at 1 volt. With only half the original opposition or resistance to be overcome, the electrons will come through twice as fast. The flow rate will be doubled, it will become 2 coulombs per second, or 2 amperes.

Now we have seen that electron flow rate is inversely proportional to resistance. As resistance goes up, flow rate comes down. As resistance comes down, flow rate goes up.

In diagrams at the left in Fig. 7-12 we changed the potential difference, and thereby altered the electron flow rate. In diagrams at the right we changed the resistance, and again altered the flow rate. The effects

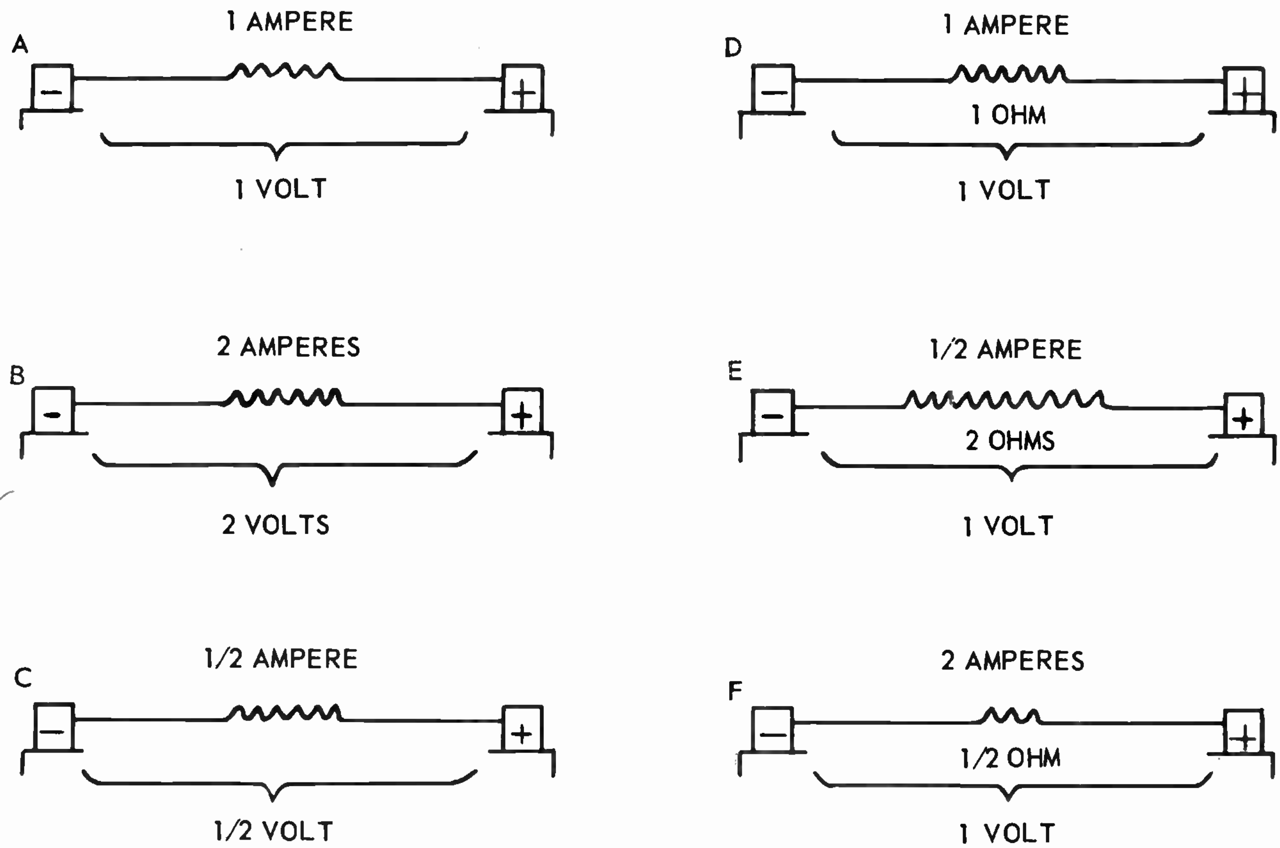


Fig. 7-12. Current is affected by changes of both potential difference and resistance.

on electron flow rate of both potential difference and resistance may be combined. The resulting flow rate in amperes always is equal to the quotient of the number of volts potential difference divided by the number or ohms of resistance, like this:

$$\text{Current, amperes} = \frac{\text{potential difference, volts}}{\text{resistance, ohms}}$$

This rule or formula looks simpler if we use the standard letter symbols; I for flow rate in amperes, E for potential difference in volts, and R for resistance in ohms. Then we have,

$$I = E/R$$

Now look back at the list of potential differences and resistances in various parts of the plate circuit, as shown by Fig. 7-11. By using these values in our new formula it becomes possible to determine the current flowing in each part. This current value would be in amperes. Currents in plate circuits and screen circuits of radio and television receivers never are so great as one ampere. Were these currents measured in amperes all the values would be fractions — hard to work with. Consequently, plate currents and screen

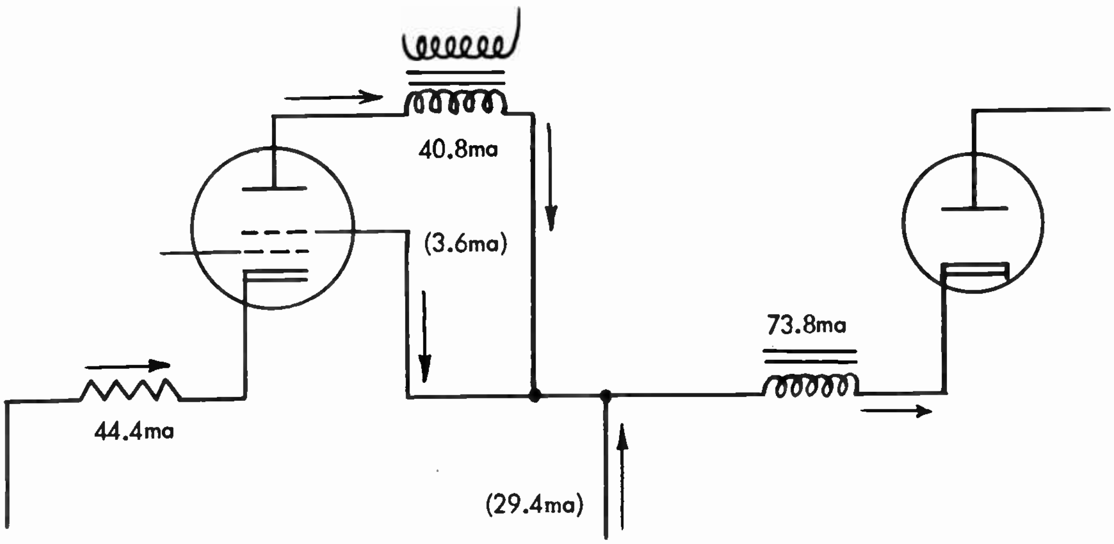


Fig. 7-13. Currents as computed from values of potential difference and resistance.

currents are measured in milliamperes. One milliampere is 1/1000 of one ampere. It is easy to change the formula so that it gives currents in milliamperes instead of in amperes.. Here is the milliampere formula:

$$\text{Milliamperes} = \frac{1000 \times E}{R}$$

To determine current in milliamperes we first multiply the number of volts potential difference by 1,000, then divide that result by the number of ohms resistance. Let's try it on resistor R of Fig. 7-11, across which there is a potential difference of 8.0 volts, and whose resistance is 180 ohms.

$$\text{Milliamperes} = \frac{1000 \times 8}{180} = \frac{8000}{180} = 44.4$$

If you divide 8,000 by 180 and carry out the answer to several decimal places it will be 44.4444... Then why do we say the current is 44.4 milliamperes? Because your measurements of voltage won't be accurate to more than tenths of volts, and it is highly improbable that resistance values will be any more accurate – and there is no object of having more decimal places in the answer than the decimal corresponding to tenths.

To determine current in the transformer winding we use its potential difference and resistance this way:

$$\text{Milliamperes} = \frac{1000 \times 6}{147} = \frac{6000}{147} = 40.8$$

Then we determine current in the filter choke.

$$\text{Milliamperes} = \frac{1000 \times 15.5}{210} = \frac{15500}{210} = 73.8$$

In Fig. 7-13 the values of computed currents have been written on a circuit diagram. Also, in parentheses, are values of current for the screen of the output tube and for all the other plate and screen circuits connected to our power supply system at a point to the left of the filter choke. This point is the terminal strip of Fig. 7-8. Currents are shown in milliamperes. The abbreviation for milliampere or milliamperes is ma. Electron flow directions are shown by arrows.

We know that the screen current of the output tube must be 3.6 ma because the total cathode current in this tube is 44.4 ma, the plate current is 40.8 ma, and the difference must be screen current. This becomes clear if you look back at Fig. 7-6.

Current in the filter choke is 73.8 ma. By observing the arrows which show directions of electron flows it is plain that 44.4 ma must be coming from the plate and screen of the output tube. The remainder of the flow must be coming from the other tubes. Subtracting 44.4 ma from the choke current of 73.8 ma leaves 29.4 ma as the plate and screen currents for all the other tubes in the set.

PLATE RESISTANCE. Potential at the plate of the output tube is 99.2 volts and at the cathode is 8.0 volts, as shown by Fig. 7-11 and other diagrams. Plate current in this tube has been found to be 40.8 ma. Can we use our earlier formula for resistance, $R=E/I$, to compute the plate resistance of this tube? The answer is no. The characteristics of a tube which is called plate resistance is a measure of opposition to signal voltages or currents. Plate resistance is related to the change of signal voltage required to cause a certain change of signal current. This we shall discuss when dealing with the performance of tubes.

The formula, $R = E/I$, would give an apparent plate resistance of about 2,235 ohms. The actual plate resistance with the voltages we are using on plate and screen is about 13,000 ohms. There is no direct relation between the two values.

METER CONNECTIONS FOR TESTING. As a general rule it is easier to compute the approximate values of current in receiver circuits than to measure them with a meter; such as an ammeter for measuring amperes, a milliammeter for milliamperes, or a microammeter for microamperes. To measure current with a meter you have to open the circuit by disconnecting at least one wire, then connect the meter in series with the circuit. There is no other way of making certain that all the circuit current flows in the meter.

To measure plate current of our output tube a milliammeter might be connected as in Fig. 7-14, or anywhere else between plate lug 3 of the tube socket and the point to which a transformer lead connects to lug 4 and thence to the terminal strip. Only in this limited path is there no current other than that coming from the plate of the tube.

The terminals of the meter are shown marked with a plus sign (+) for positive and a minus (–) sign for negative. Not all meters for direct-current measurement have a marking for their negative terminal, but all have one for the positive terminal – and the other terminal then must be negative. The meter must be connected into the circuit so that electron flow through the meter is from its negative terminal to its positive terminal. With this connection the meter pointer will move away from zero and toward higher numbers on its scale when there is electron flow. If connections to the meter are reversed, its pointer will move "off scale" below zero. Reversed current, or electron flow through a current-measuring meter, should cause no damage provided the current is no greater than safely might be measured "up scale".

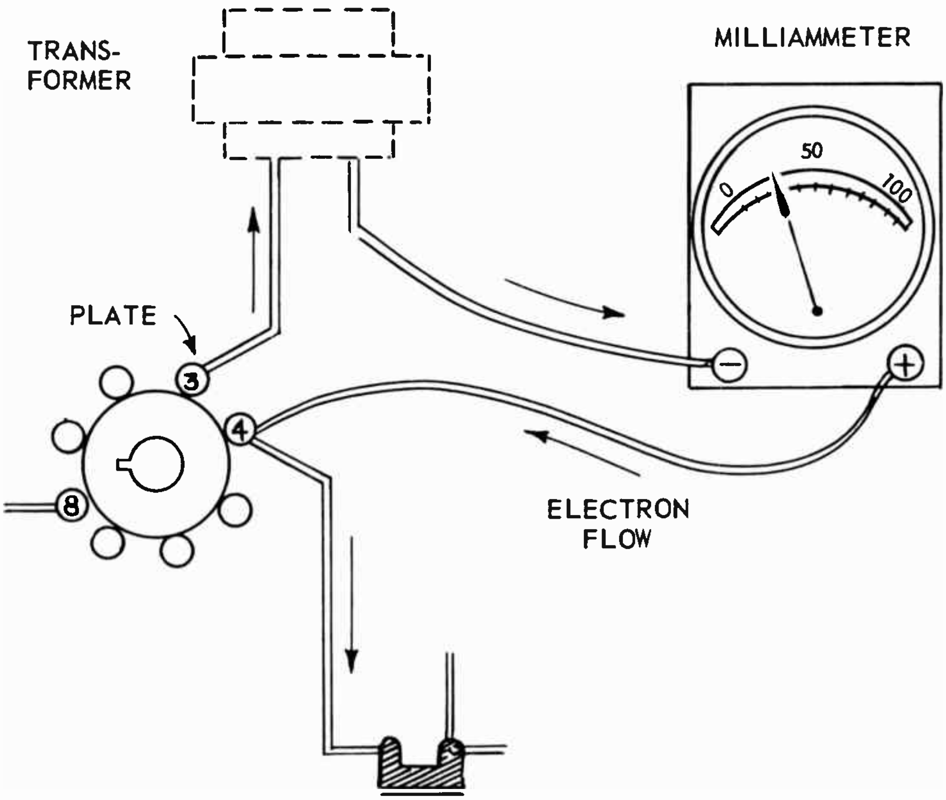


Fig. 7-14. The milliammeter connected for measuring plate current.

The internal resistance of current-measuring meters is very small. This is necessary in order that the meter shall not add enough resistance in the circuit to appreciably effect the current flow. A meter having a maximum range of 100 milliamperes usually has internal resistance of less than 2 ohms. Supposing you should make the mistake of connecting such a meter across a difference of potential rather than in series with the circuit. For example, a meter having 2 ohms internal resistance was accidentally connected between lugs 3 and 4 of the output tube socket, where the potential difference is 6 volts. Current through the meter then would be as computed from our formula for current.

$$\text{Milliamperes} = \frac{1000 \text{ x } 6 \text{ (volts)}}{2 \text{ (ohms)}} = \frac{6000}{2} = 3000$$

With 3,000 milliamperes in a meter designed for no more than 100 milliamperes what do you think will happen to the meter? It will burn out in a fraction of a second after you make the wrong connection. Current-measuring meters of all kinds must be considered as delicate measuring instruments, and handled accordingly.

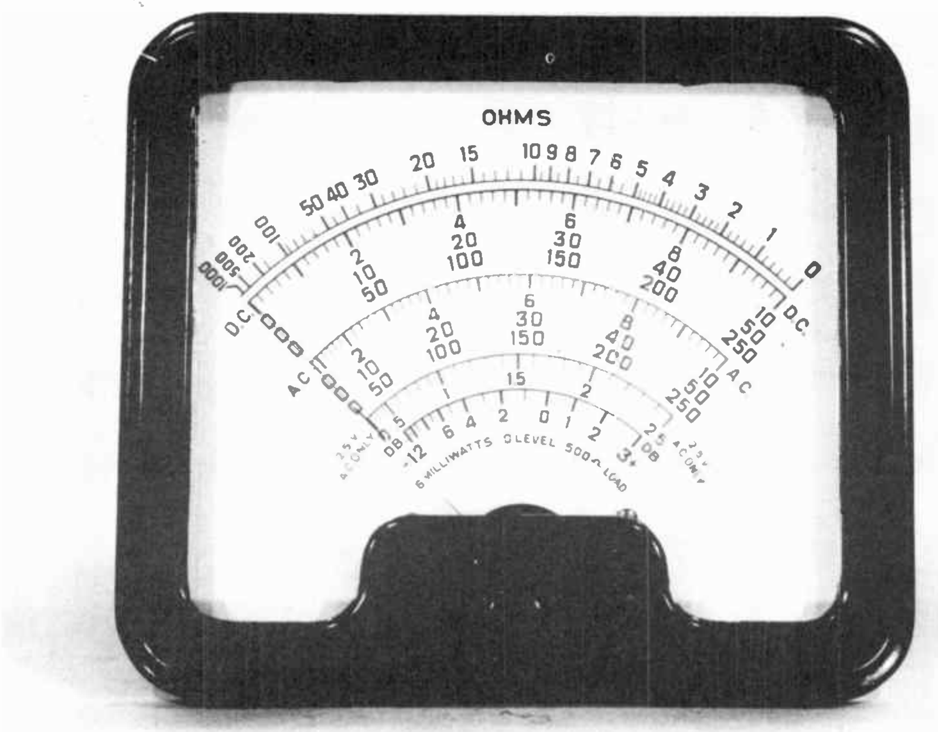


Fig. 7-15. Dial scales of a combination voltmeter-ohmmeter.

Voltmeters are relatively tough, unless you make very bad mistakes. The internal resistance of a direct-potential or d-c voltmeter used for service testing seldom is less than 1,000 ohms per volt. This means that the total resistance of the meter is equal to 1,000 ohms multiplied by the highest voltage reading on the dial scale. For example, such a "thousand-ohm-per-volt" meter, when used with a scale extending to a maximum of 100 volts, will have internal resistance of 100,000 ohms.

Service voltmeters may have resistances such as 5,000, 10,000, or even 20,000 ohms per volt. Fig. 7-15 is a picture of the scales on a combined voltmeter-ohmmeter having d-c voltage scales with maximums of 10, 50, and 250 volts. This is a 20,000 ohms per volt instrument. When using terminals for the 10-volt scale the internal resistance of the meter is 200,000 ohms, for the 50-volt scale the resistance is 1,000,000 ohms, and for the 250-volt scale the resistance is 5,000,000 ohms or 5 megohms.

When using a multi-range voltmeter to measure an unknown potential difference always commence with the highest range. Most meters will withstand a momentary over-voltage up to one and one-half times their full-scale reading. Consequently, when beginning your test with a high range you are unlikely to harm the voltmeter unless you get it across some of the circuits for a television picture tube. If the meter pointer goes off scale at the high end, the potential difference is too high for the meter to measure. If the reading

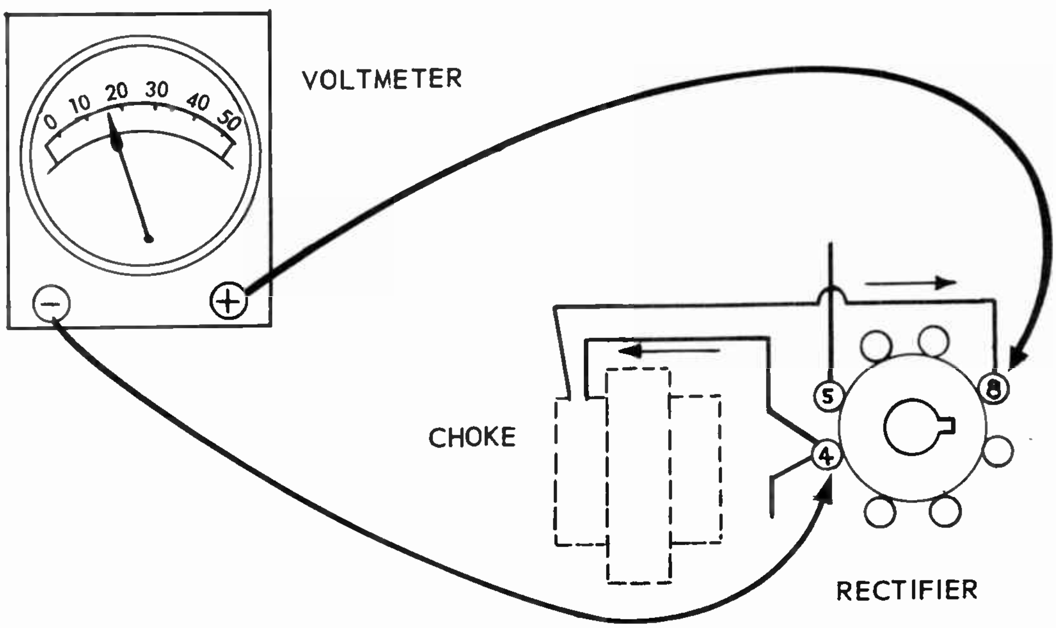


Fig. 7-16. The voltmeter connected for reading potential difference across the filter choke.

is too low for convenient reading on the high range scale, change to a lower range where you can make an accurate reading.

A voltmeter always should be connected across the points between which potential difference is to be measured, never in series. A series connection would insert the high resistance of the meter into the circuit. No circuit wires are removed from the points to which the voltmeter is connected for test. That's why it is easy to use a voltmeter. For measuring potential difference across the filter choke in our plate circuit the meter connections would be made as in Fig. 7-16. The choke is connected between lugs 4 and 8 on the rectifier socket, so leads from the voltmeter are touched to these two lugs.

Terminals of voltmeters for direct potentials are marked positive and negative, or at least one is marked positive, just as are the terminals of current meters. To have the voltmeter read up scale from zero, the negative terminal of the meter must be connected to the point at which electron flow enters the part whose potential difference is being measured. The positive terminal of the meter is connected to the point at which electron flow leaves that part. If you think of electron flow as going through the voltmeter, the connections are made in the same way as with a current meter. If a voltmeter is connected the wrong way around, the pointer will move off scale below zero. No harm will result unless the meter is connected across a potential difference greater than it could safely measure in a forward direction.

TELEVISION

LESSON NO. 8 SERVICE TESTS FOR ALL CIRCUITS

Training Your Mind...

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind

COMMERCIAL TRADES INSTITUTE

C.T.I.

Perfecting Your Skill

LESSON NO. 8

SERVICE TESTS FOR ALL CIRCUITS

HISTORY ALWAYS REPEATS

One of the oldest and best known manufacturers of television apparatus has this to say. "The nature, location, and repair of troubles must be analyzed by the serviceman. This necessitates a good working knowledge of circuits of the sets, as well as an understanding of television principles. It behooves the serviceman to study circuits."

The man who wrote this piece of advice knew that, while nothing changes faster than the details of television design, the basic circuits don't change, they never change. These elementary circuits are combined in new ways to do new things and to do old things better. But if you know your circuits nothing new will stump you.

When Marconi transmitted the first radio signals in 1894 his apparatus was built with conductors, insulators, resistors, inductors, and capacitors. Builders of the first broadcast receivers in 1920 worked out their designs by using the rules which show relations between potential difference, current flow, and resistance. They used detectors and amplifiers. We still have to use all these basic things today, in television as well as sound radio. Circuit elements are put together in different combinations, but when these new combinations are broken down into their essential parts you are right back to conductors, resistors, inductors, capacitors, and all the elementary principles which govern their performance.

We are learning principles while studying actual circuits. These particular circuit combinations may go out of date. But whatever takes their place will employ the same old basic principles. Remember, "It behooves the serviceman to study circuits." Then he won't care whether the set he works on was built in 1940, is built today, or will be built in 1975.

The old timer in radio gets many a smile when he enters this newest field of television. He finds crystal diodes used for video detectors, restorers, and f-m discriminators. Back around 1918 all his detectors were crystals, which then became obsolete. Our new crystals are better mechanically, they contain better materials, but their working principles haven't changed a bit.

Early set builders improved on their crude crystal detectors by connecting a grid leak and capacitor to a tube. Now the same principle, called grid rectification, is used for many other purposes at a half dozen places in television sets.

Around 1937 we all studied automatic frequency control, which then helped people tune their sets correctly and more easily. Later this control was almost forgotten. Now we use precisely the same principles, even the same circuit arrangements, for automatic control of sweep oscillators in nearly all the new television receivers.

Fig. 8-1. A test setup for checking effects of different capacitors in an amplifying stage.

We could keep on telling how things long since obsolete, so we thought, have become the newest discoveries for those who are new in the profession of radio. It would all point up to this. To the man who knows his circuit principles, the new things are just new ways of combining those principles.

SERVICE TESTS FOR ALL CIRCUITS

This must be our last lesson, for the present, on fundamental principles of servicing. After this lesson it will be time to learn about the behavior of the many kinds of tubes used in television and radio receivers. Possibly we should call this a reference lesson, for you, or no one else, going over this material for the first time could remember it all. Even though you memorized every word, you still wouldn't realize all that they are going to mean. Many times in the future you will come back to the information given here, as you find need for it in practice.

POTENTIALS, CURRENTS, AND RESISTANCES. In the preceding lesson we measured some potential differences and computed the currents which flow with these potential differences acting across certain resistances. Here are the values.

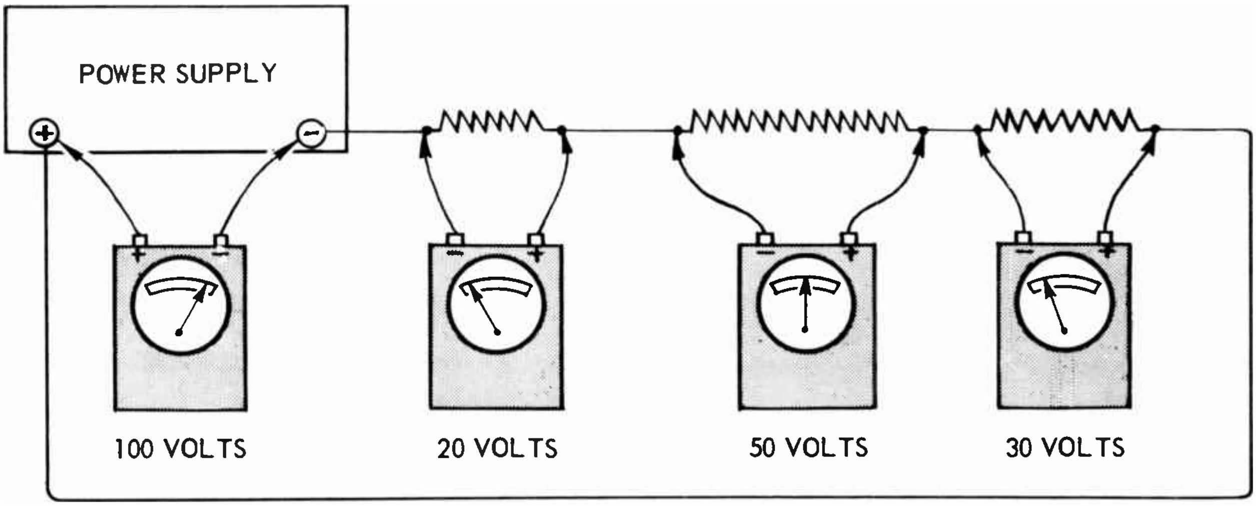


Fig. 8-2. The sum of the potential differences in a series circuit equals the potential difference from the power supply.

	Volts	Milliamperes	Ohms
Resistor on cathode of output tube.	8.0	44.4	180
Plate winding of output transformer.	6.0	40.8	147
Filter choke of power supply system.	15.5	73.8	210

Supposing you knew only the milliamperes and the ohms for each part, could you figure out the volts of potential difference? There is a very simple relation. If you don't quite see it, look at this next list of values, which also were used in the preceding lesson.

Volts	Amperes	Ohms
1	1	1
2	2	1
1/2	1/2	1
1	1/2	2
1	2	1/2

Here the relation becomes plainer. Volts are equal to amperes times ohms, provided all three values are in the same resistance or in the same part of a circuit.

When this new rule or formula for potential difference is written so it uses milliamperes instead of amperes it appears like this.

$$\text{Potential difference, volts} = \frac{\text{current, milliamperes x resistance, ohms}}{1000}$$

Try this formula on the combinations of values given first. It works every time. Much earlier we used a

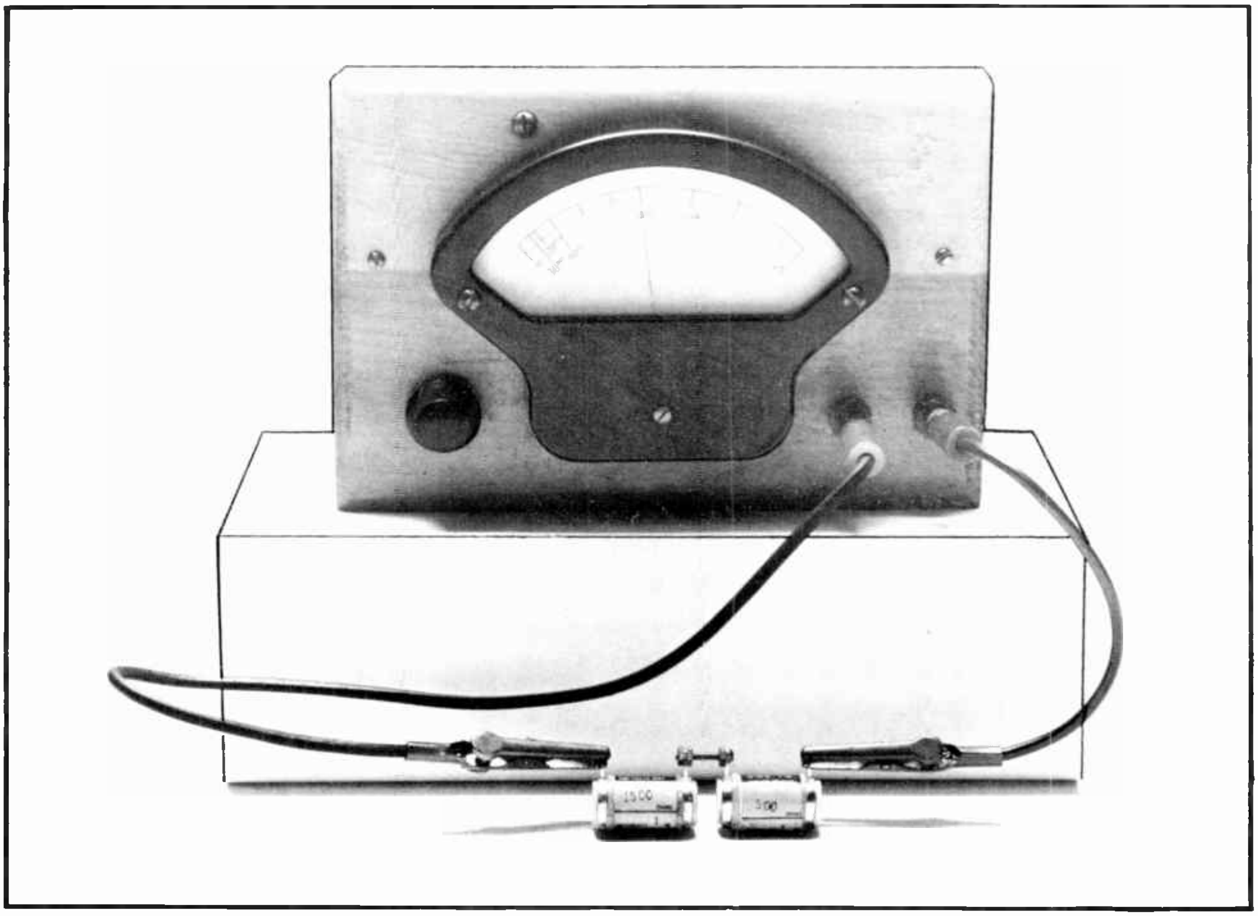


Fig. 8-3. The ohmmeter reads the sum of resistances in series.

formula for resistance, the one that says,

$$R = E/I$$

When the resistance formula is written this way it applies for currents measured in amperes. Most of our currents must be measured in milliamperes. The same resistance formula, written to include milliamperes, is this.

$$\text{Resistance, ohms} = \frac{1000 \times \text{potential difference, volts}}{\text{current, milliamperes}}$$

Or, using letter symbols, we may write,

$$R = \frac{1000 \times E}{\text{milliamperes}}$$

Now we are ready to do a little practicing with the three tools which are indispensable for servicing any

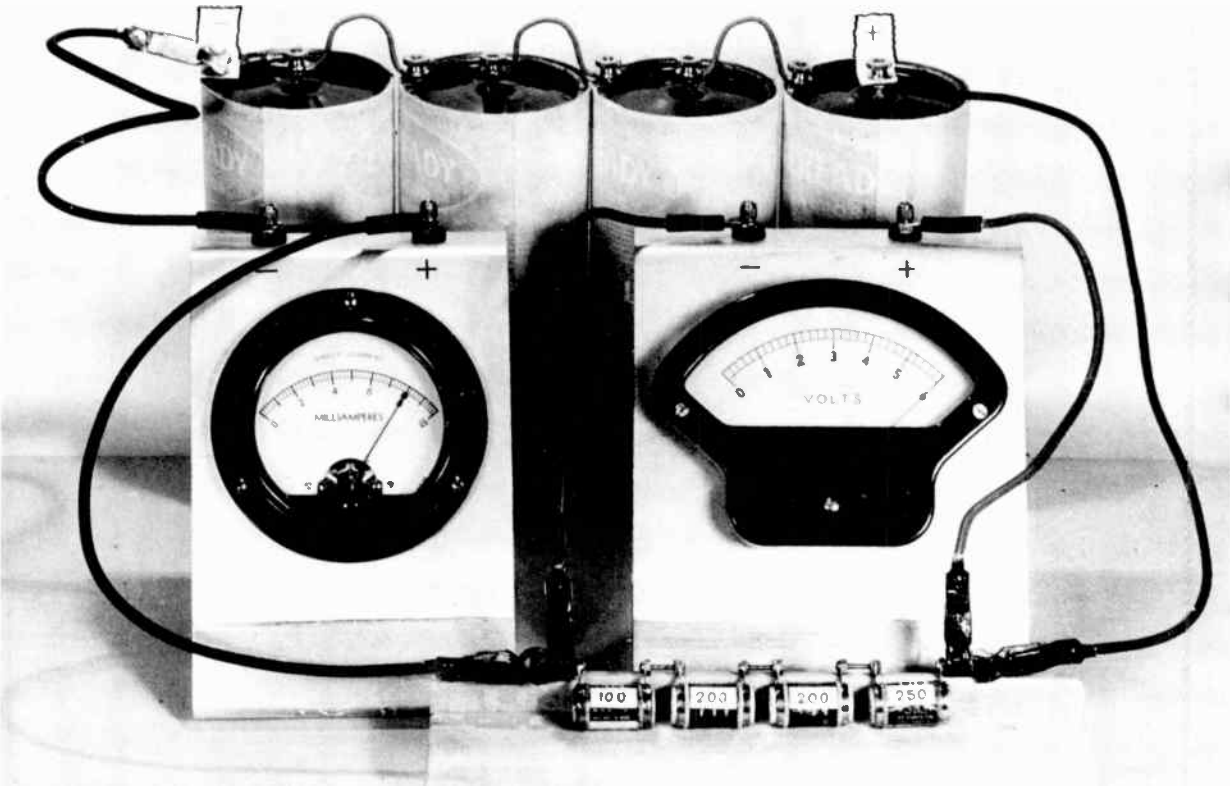


Fig. 8-4. In this experimental setup the dry cells represent any power supply, the resistors represent any loads.

kind of receiver. Even when you are working with equipment like that of Fig. 8-1, you wouldn't get far without these three tools. Luckily, these indispensable tools cost nothing, they never wear out, they take up no space, they never get lost, and they save time and headaches on at least half the problems you run into. The three tools are:

$$R = \frac{1000 \times E}{ma} \qquad ma = \frac{1000 \times E}{R} \qquad E = \frac{ma \times R}{1000}$$

The letter R stands for ohms of resistance. The letter E stands for volts of potential difference. The symbol ma stands for milliamperes of current. Here we don't use the letter I for current, because I should be used to represent amperes, nothing else.

Simple as they are, these three formulas are not too easy to memorize. For the present don't try to memorize them. Copy the three formulas onto a small card, then keep that card handy while we work.

POTENTIAL DIFFERENCES ADD TOGETHER. In Fig. 8-2 a voltmeter connected across the terminals

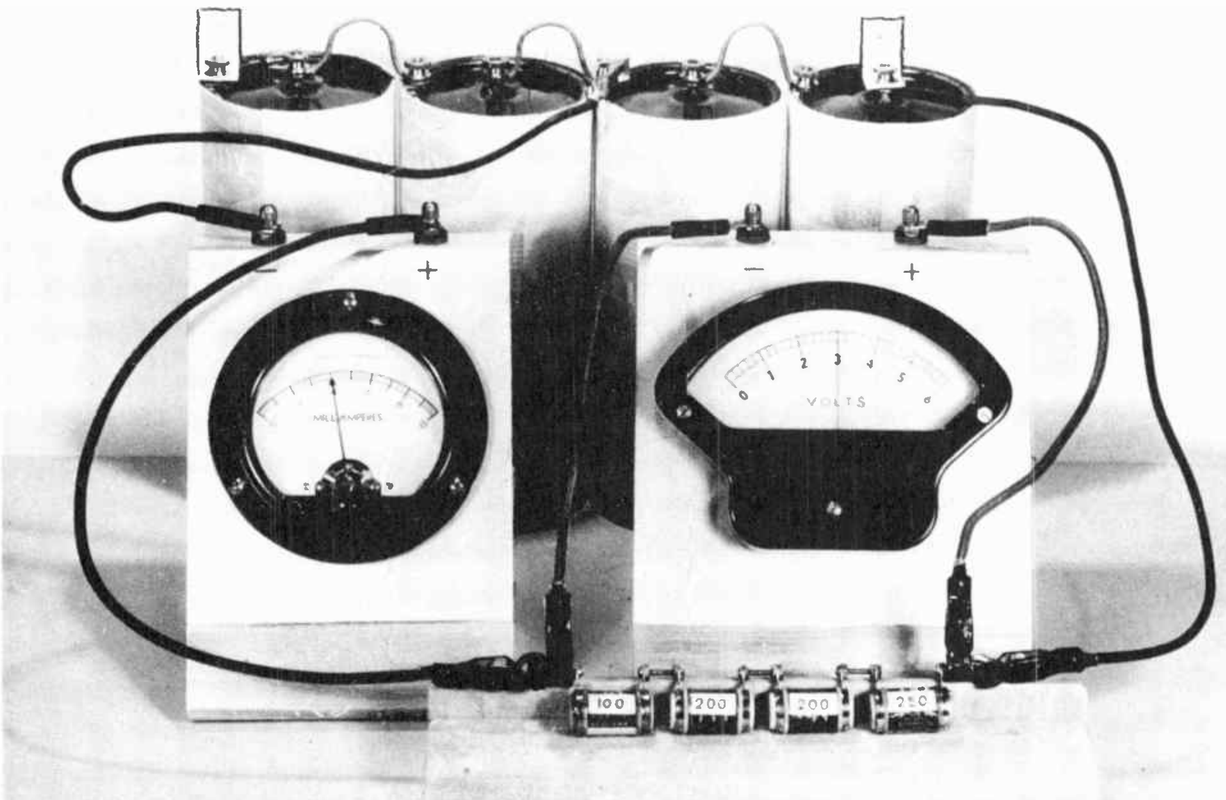


Fig. 8-5. Reducing the voltage reduces current through a given resistance.

of the power supply reads 100 volts. In the external circuit are three resistances, which might represent loads in any series circuit. Other voltmeters connected across the three resistances read 20 volts, 50 volts, and 30 volts. The sum of $20 + 50 + 30 = 100$.

9/ The sum of all the separate potential differences in a series circuit is equal to the potential difference across the entire circuit. We may say that voltage of the power supply must equal the sum of the voltages across all the parts in the connected series circuit.

If you subtract from the voltage of the power supply the voltage across any one part in the series circuit, the remainder is the total of all the potential differences across all the other parts in that circuit.

These simple facts are easy to understand. Electrons leave the power supply with total energy proportional to the voltage or potential difference of the power supply. The electrons lose all this energy in the external circuit. Since the potential difference across each part in the circuit is proportional to electron energy used up in this part, it is plain that the sum of all the energy losses and potential differences must equal the original electron energy and the original potential difference of the power supply.

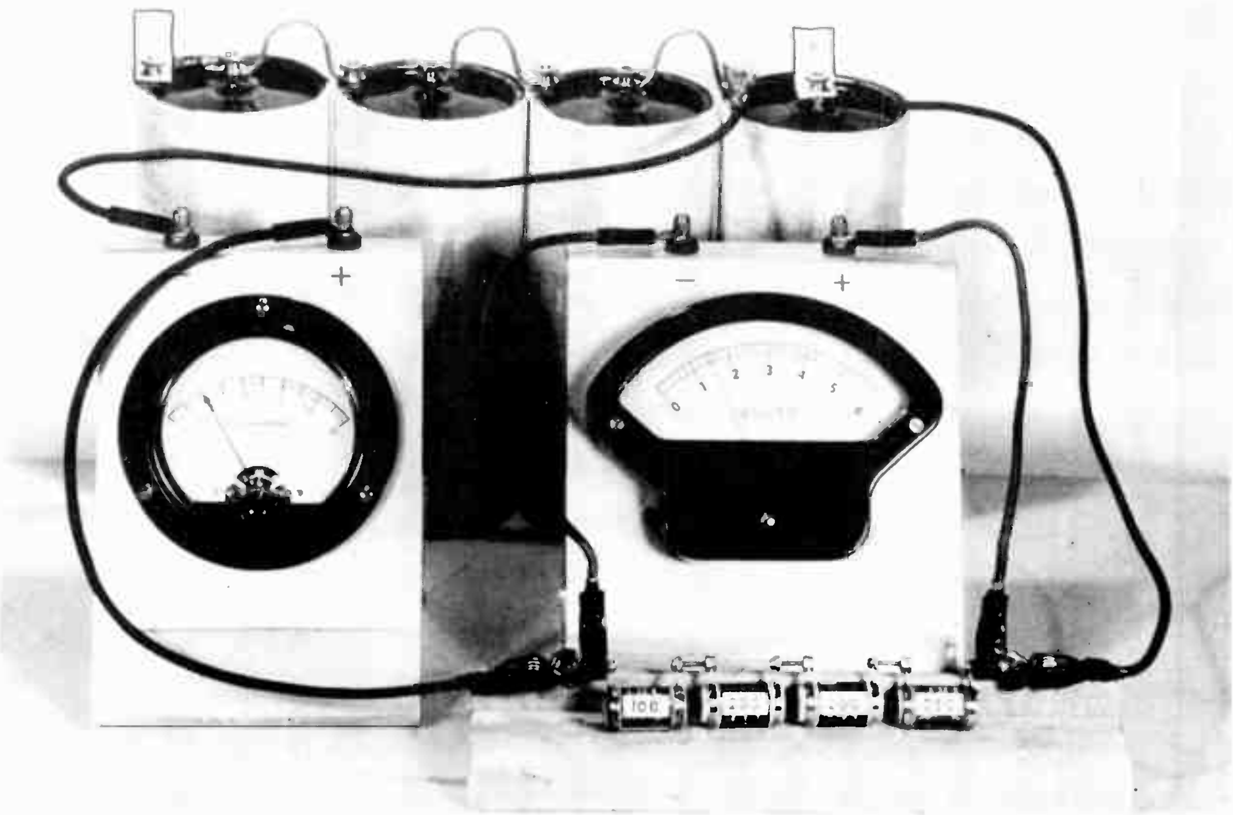


Fig. 8-6. Current remains proportional to applied voltage when resistance is not changed.

RESISTANCES ADD TOGETHER. In Fig. 8-3 an ohmmeter is connected across two resistors which are in series with each other. The resistance of one unit is 1,500 ohms, and of the other it is 500 ohms. The ohmmeter reads a total resistance of 2,000 ohms.

The combined resistance of any number of separate resistances in series is equal to the sum of the separate resistances. The resistance of an entire series circuit is equal to the sum of the resistances of all the parts in that circuit.

This is entirely reasonable. Resistance is a measure of how hard it is for electrons to get through. If there is a certain opposition in 100 ohms, there must be twice as much opposition to the same rate of flow in 200 ohms, three times as much in 300 ohms, and so on. If electrons are forced to flow at a certain rate through a resistance of 10 ohms, then at the same rate through a second series resistance of 20 ohms, and finally at the same rate through a third series resistance of 40 ohms, there is exactly the same total opposition as in one resistance of $10 + 20 + 40$, or 70 ohms.

CURRENTS DO NOT ADD TOGETHER. We have learned that the rate of electron flow or current must

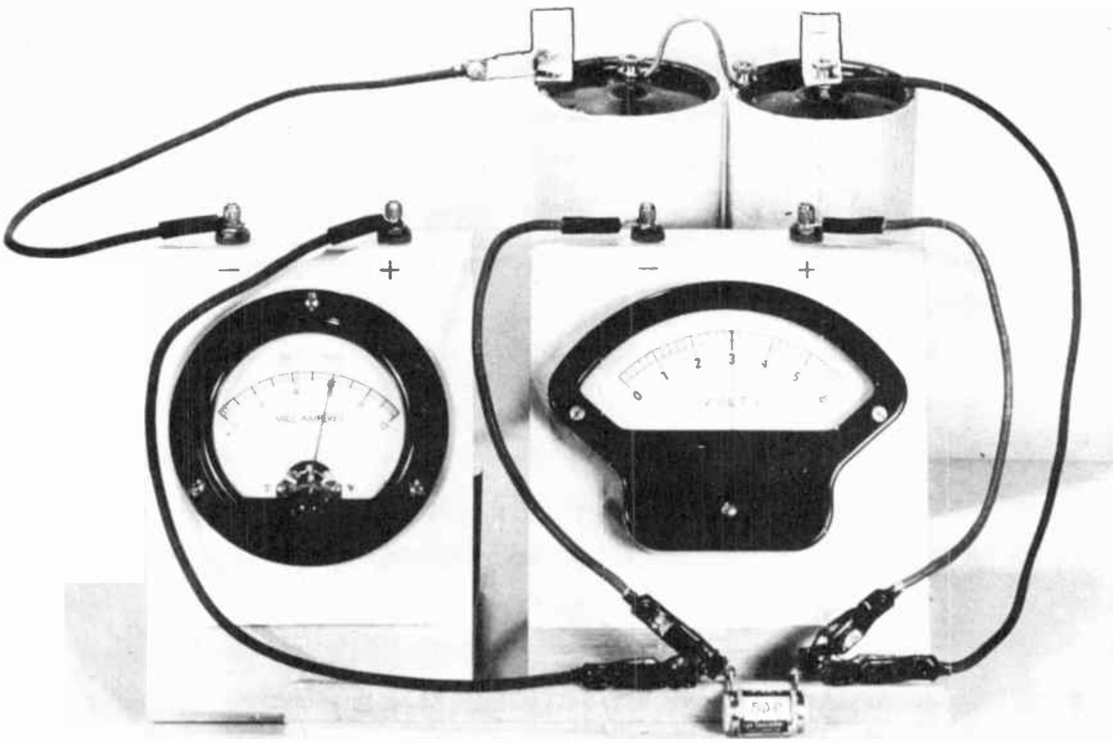


Fig. 8-7. Power supply voltage has been reduced.

be just the same in any one part of a series circuit as in every other part of the same series circuit. If the flowrate is 40 milliamperes from the plate of a tube, and we have a transformer winding in series with this plate, the current in the transformer winding must be 40 milliamperes.

We must not forget that potential differences add together, resistances add together, but current is the same everywhere in a series circuit. Current in one part does not add to current in another part of a series circuit to make a total current equal to the sum. The total current is just the same as the current in any one part.

VOLTAGE OF POWER SUPPLY AND CURRENT IN CIRCUIT. To illustrate principles in the simplest way we shall use some dry cells for our power supply. To represent any kinds of loads in the circuits we shall use some precision resistors whose actual resistances are accurate to within one per cent of their marked values. Then we shall measure potential differences with a voltmeter, and currents with a milliammeter. The setup is pictured by Fig. 8-4.

Here we have four resistances (loads) in series. From left to right the values are 100 ohms, 200 ohms, 200 ohms, and 250 ohms. Total circuit resistance is 750 ohms, is it not? In series with this resistance and the power supply is the milliammeter, at the left. Negative of the power supply goes to negative of the milliammeter. Positive of the milliammeter goes to the left-hand resistor. From the right hand resistor a connection goes to positive of the power supply. The meter reads a current of 8 milliamperes. Across the entire string of resistors is connected the voltmeter, the instrument at the right. This meter reads 6 volts.

Assuming the resistance values to be actually as marked, are the meters telling the truth? Use your three formulas to find out. Really do this: write the figures on a piece of paper. It is invaluable practice. Keep in mind that you might be working with any kind of d-c power supply and with any kinds of loads.

In Fig. 8-4 the voltmeter is effectively connected across the power supply as well as across the loads. From the positive, right-hand, terminal of the voltmeter there is a wire connection down to the right-hand resistor and another wire connection up to the positive terminal of the power supply. From negative of the voltmeter there is a wire connection to the left-hand resistor, another wire from here to the milliammeter, and from the other side of this meter there is a connection to negative of the power supply. The resistance of the milliammeter is approximately 3.5 ohms, entirely negligible in comparison with the total load resis-

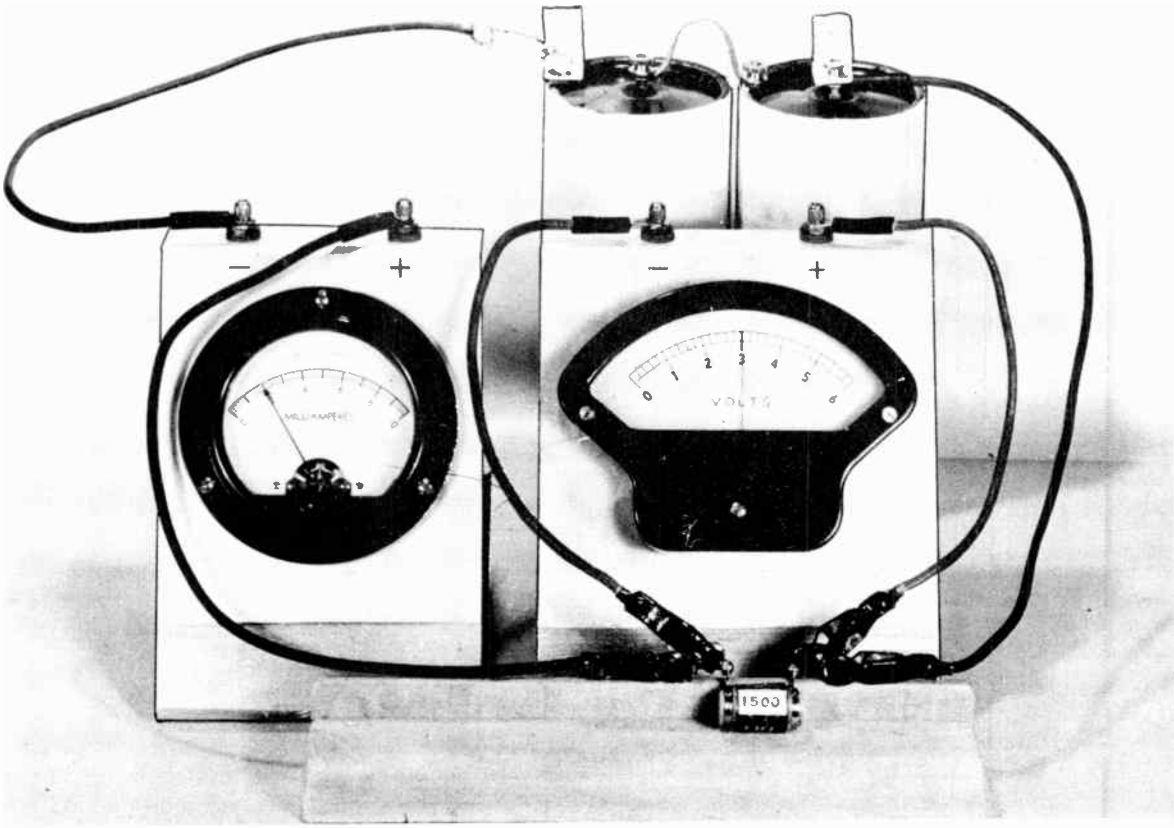


Fig. 8-8. With three times as much resistance there is one-third as much current.

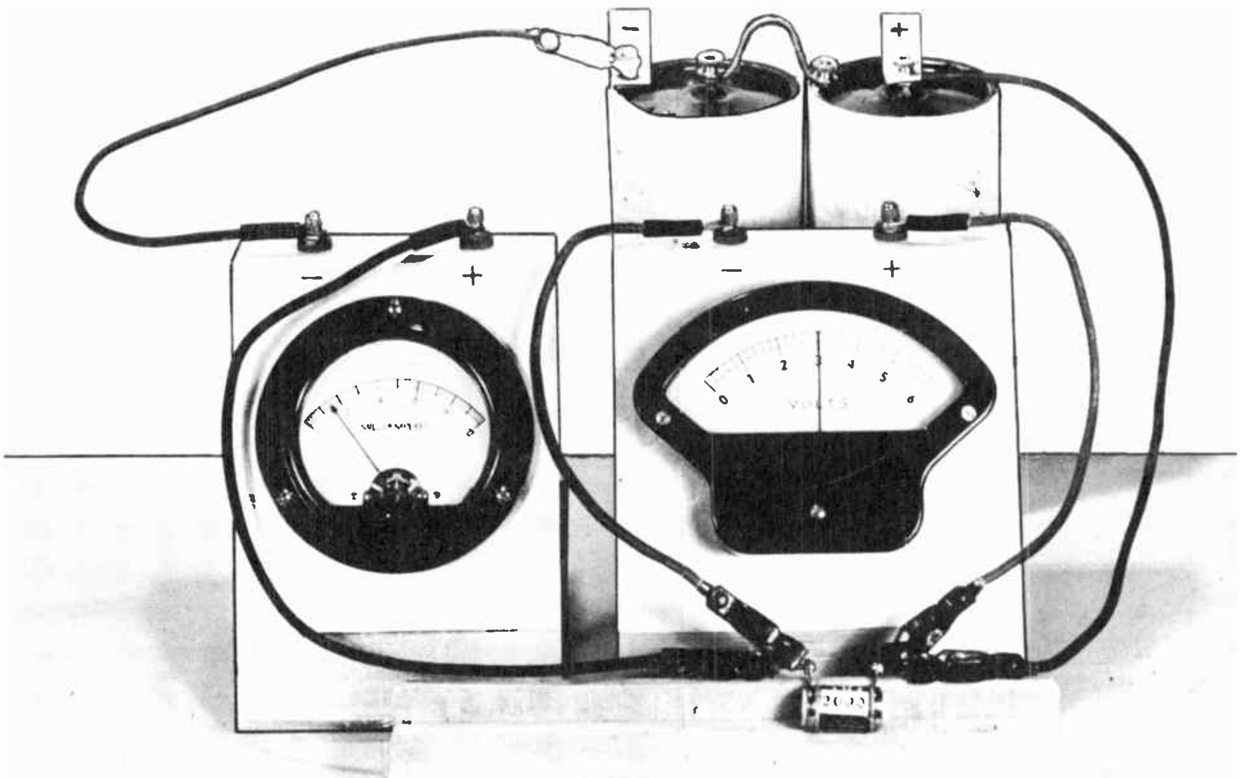


Fig. 8-9. Current is inversely proportional to resistance when potential difference remains unchanged.

tance. Since the voltmeter is connected through negligible resistances to both ends of the power supply, this meter is indicating power supply voltage. The four dry cells are furnishing a potential difference of 6 volts.

In Fig. 8-5 the wire connection from the negative side of the milliammeter has been moved over to a point between the middle dry cells. Now the voltmeter reads 3 volts, although this meter still is connected across the entire circuit resistance. The milliammeter now reads 4 milliamperes. It will be good practice to check this new combination of voltage, current, and resistance with your formulas. That, however, isn't the most important thing.

The important thing to be noted here is the relation between power supply voltage and circuit current. First (Fig. 8-4) we had 6 volts and 8 milliamperes. Now we have 3 volts and 4 milliamperes. Half the voltage has resulted in half the current.

Let's check this voltage-current relation a little further. In Fig. 8-6 the connection from the milliammeter has been moved over to the right-hand dry cell. Now the voltmeter reads only 1.5 volts, although still across the entire load circuit. The milliammeter now reads 2 milliamperes. We have one-fourth the original voltage (of Fig. 8-4) and the current now is one-fourth the original value.

Changes of current in any circuit are directly proportional to changes of voltage across that circuit, when resistance of the circuit remains unchanged. Doubling the supply voltage doubles the current. Halving the supply voltage halves the current. When circuit current increases, there must be a proportional increase of applied voltage provided the circuit resistance has not been changed. If circuit current decreases there must be a proportional decrease of supply voltage.

CIRCUIT RESISTANCE AND CURRENT IN CIRCUIT. Now let's see what happens when we do not change the voltage from the power supply but do change the load resistance or circuit resistance What will happen to current in the circuit. In Fig. 8-7 our power supply consists of only two of the dry cells. With two cells we have 3 volts applied to the circuit load. The voltmeter reads 3 volts. The load, or the circuit resistance, consists of one 500-ohm resistor. The milliammeter, in series with the load resistance, reads 6 milliamperes. Check this reading by means of your formula for current in milliamperes.

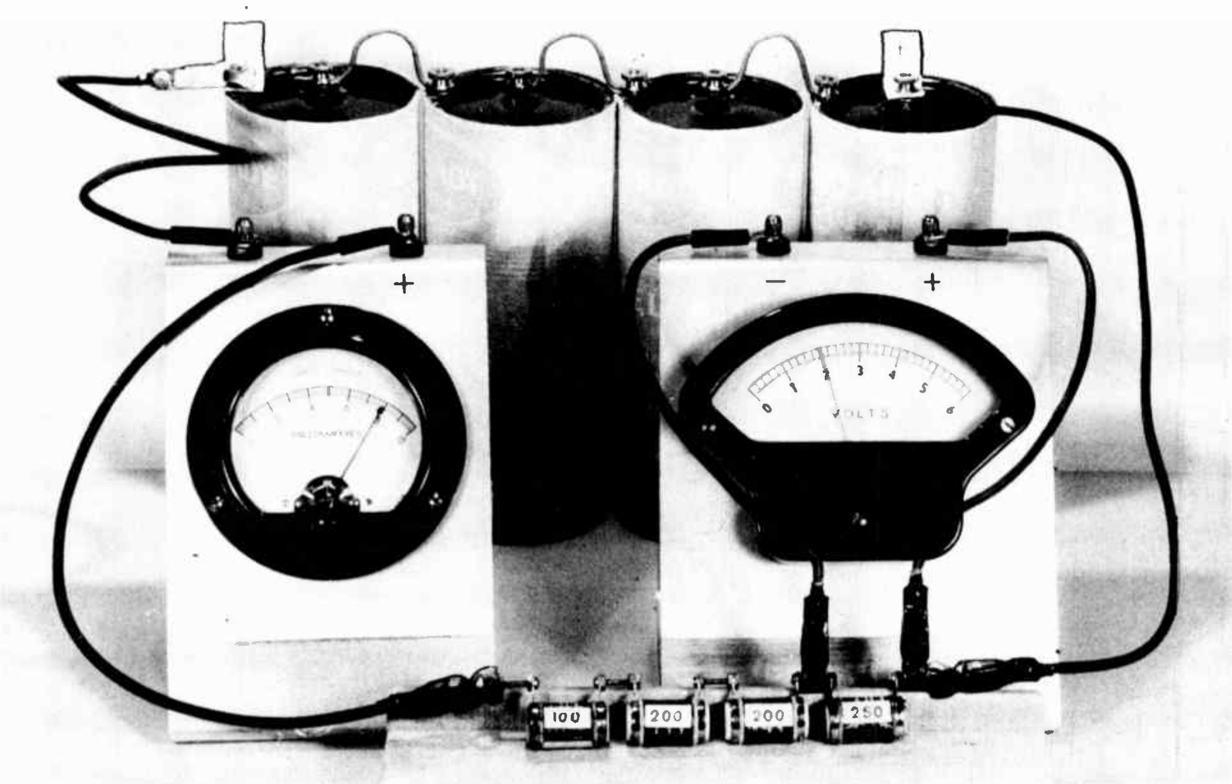


Fig. 8-10. The voltmeter is measuring potential difference across only part of the circuit load.

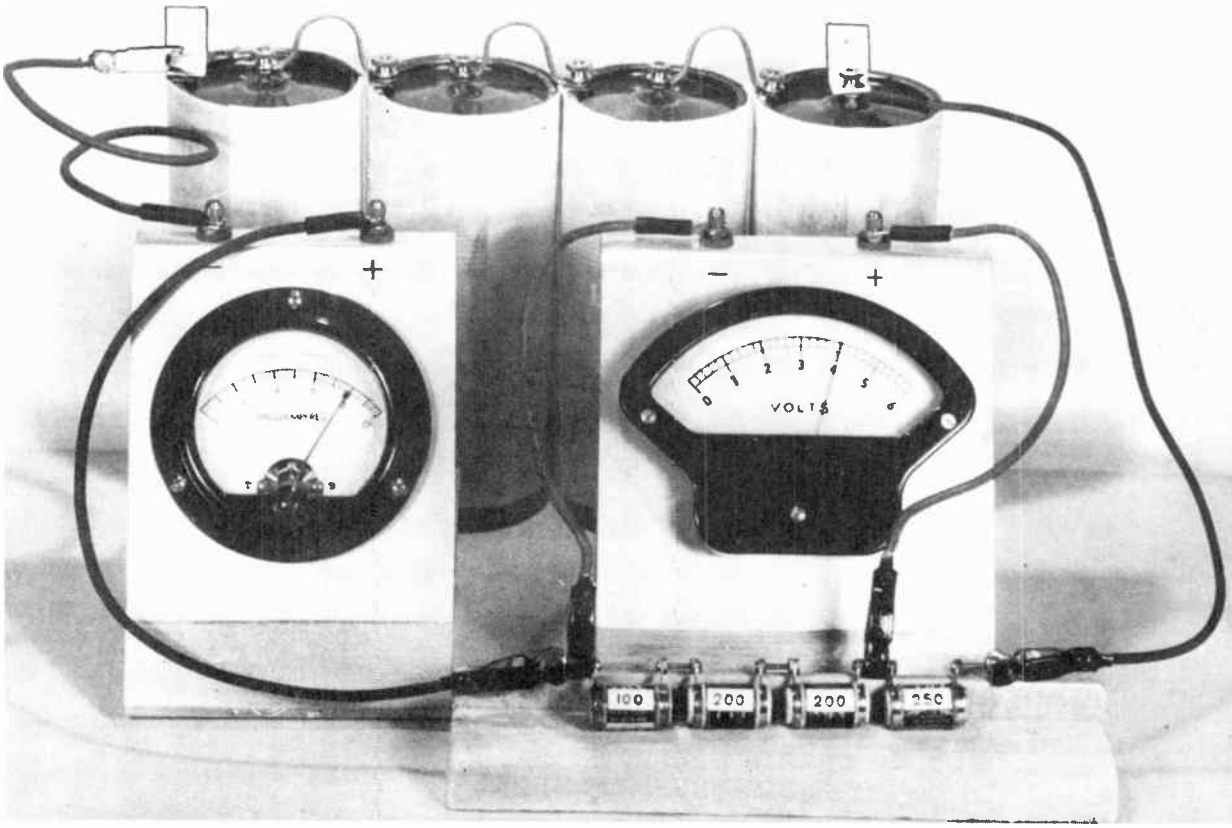


Fig. 8-11. Here the voltmeter is measuring potential difference across the remainder of the circuit load.

In Fig. 8-8 the circuit resistance or load resistance has been changed to 1,500 ohms. We still have 3 volts potential difference from the power supply. Now the milliammeter reads only 2 milliamperes. Using three times as much circuit resistance, with the same applied potential difference, has cut the current to one-third. Should someone look first at this Fig. 8-8, then at preceding Fig. 8-7, they would say that using one-third as much load resistance allows three times as much current. One statement is just as true as the other.

Next we shall change the load resistance to 2,000 ohms, as in Fig. 8-9, while retaining the original applied potential difference of 3 volts. The voltmeter reads 3 volts. The current as read from the milliammeter has dropped to 1.5 milliamperes.

When we started this series of experiments, with load resistance of 500 ohms, the current was 6 milliamperes. Now, with 4 times as much resistance, we have 1/4 as much current.

Comparing Fig. 8-8 with Fig. 8-7 we observe that, with 3 times the resistance (1,500 ohms as against

500 ohms), we have 1/3 the current (2 milliamperes as against 6 milliamperes).

More resistance drops the current. Less resistance allows an increase of current, always with the same applied potential difference. There is an exact relation between resistance and current when there is no change of voltage. Current varies inversely as resistance. With 4 times the resistance there is 1/4 the current, with 3 times the resistance there is 1/3 the current. With 2 times the resistance there would be 1/2 the current. With 1,000 times the resistance there would be 1/1000 of the current.

Resistors in an electric circuit reduce the rate of flow somewhat as valves in a water circuit reduce the rate of water flow. Resistors are not so efficient as water valves, for resistors reduce the electron flow by wasting the electron energy, by changing that energy into heat which raises the temperature of the resistor.

RESISTANCE AND LOSS OF POTENTIAL. Resistance which is connected into a circuit reduces the rate of electron flow in that circuit. The resistance causes a loss of electron energy. In Fig. 8-10 we have gone back to the four-cell power supply and to the string of four resistors in series. In series with the resistors and the power supply is the milliammeter. This meter shows the rate of electron flow to be 8 milliamperes, the same flow as back in Fig. 8-4. The fact that we have the same current and the same total

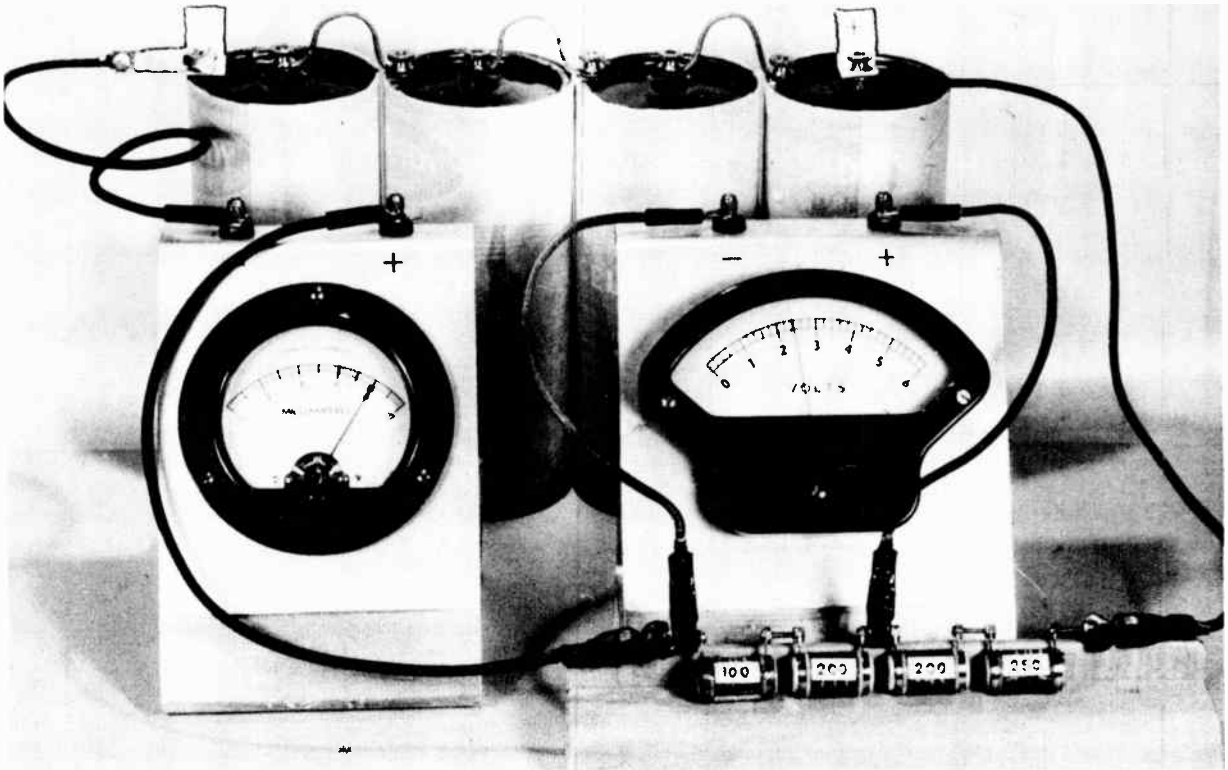


Fig. 8-12. When current does not change, potential difference and resistance always are directly proportional.

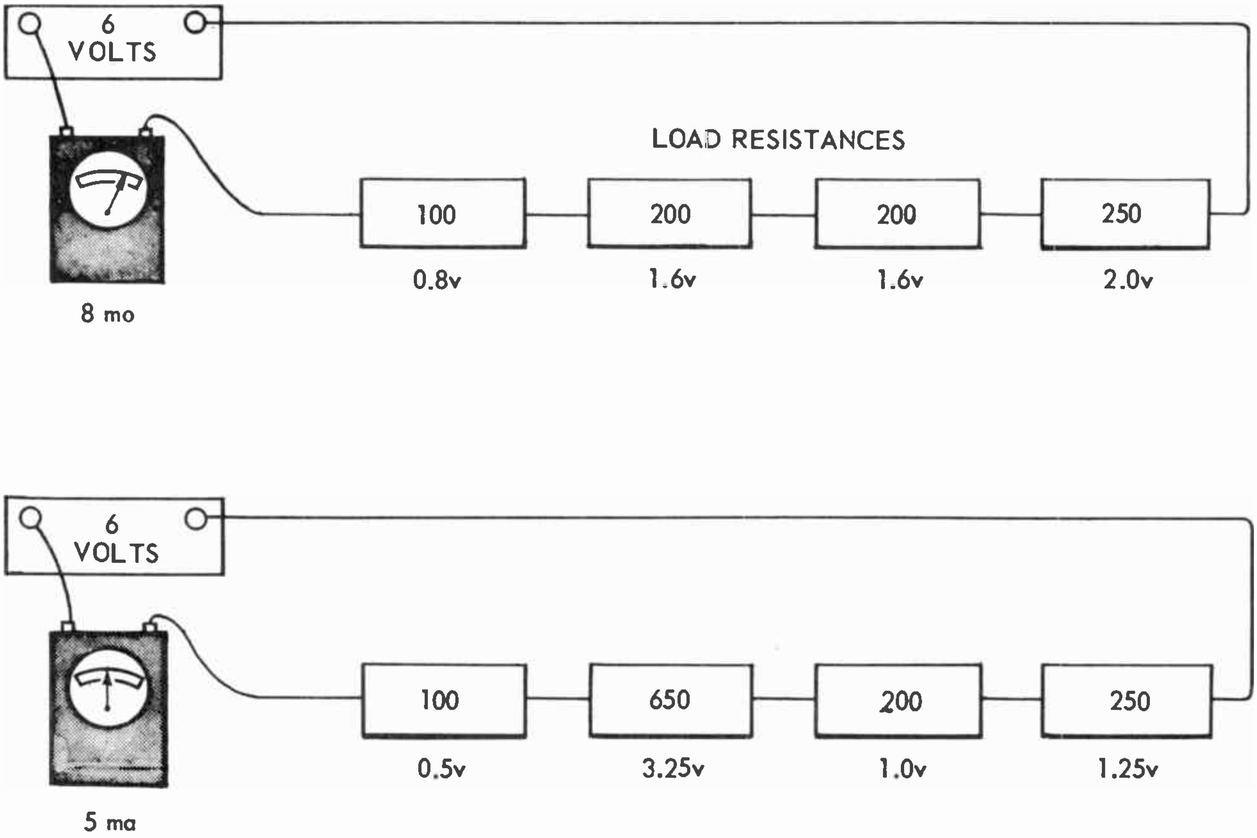


Fig. 8-13. A change of resistance anywhere in a series circuit alters current and potential differences everywhere in that circuit.

resistance as in Fig. 8-4 proves that there must be the same potential difference across the resistors. That potential difference is 6 volts.

In Fig. 8-10 the voltmeter is connected across only the right-hand 250-ohm resistor. The voltmeter shows that the potential difference across this one resistor is 2 volts. In Fig. 8-11 the voltmeter connections have been moved so that they are across the three resistors toward the left. These units have separate resistances of 100 ohms, 200 ohms, and 200 ohms, for a total of 500 ohms. The potential difference across these three resistors is shown by the voltmeter to be 4 volts.

In these last two pictures the voltmeter is connected first across one part of the load, then across all the remainder of the load. The first potential difference is 2 volts, the second is 4 volts. The sum of these two potential differences is 6 volts, which we know to be the potential difference across the entire load. Here we have proof that potential differences in a series circuit add together, and that their sum is equal to the total potential difference across the whole circuit.

Now here is something we should note with care. Potential difference across resistances is exactly pro-

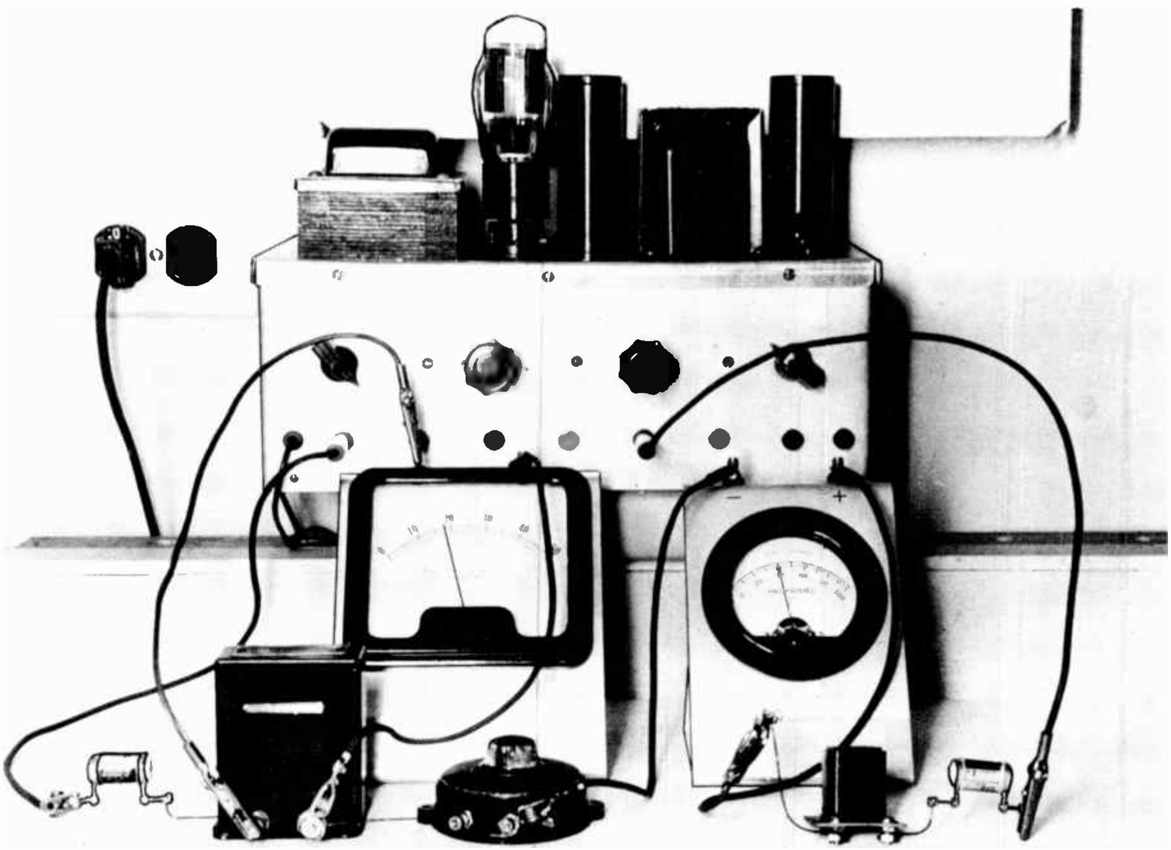


Fig. 8-14. Current measured at one point in a series circuit may be used for computations relating to other parts of the same circuit.

portional to the values of resistance when the current is the same in all units. In Fig. 8-10 the voltmeter reads 2 volts when across 250 ohms. In Fig. 8-11 the voltmeter reads 4 volts when across 500 ohms. The current remains unchanged, as shown by the milliammeter. Across twice the resistance we get twice as much potential difference. Or across half the resistance we get half as much potential difference.

Whether or not this direct proportion or direct ratio between resistance and potential difference always holds true may be checked by one more experiment, in Fig. 8-12. Here the voltmeter is connected across the two resistors toward the left. Their resistances are 100 ohms and 200 ohms, a total of 300 ohms. The current still is 8 milliamperes, as shown by the milliammeter. The resistance of 300 ohms is 300/750 of the total circuit resistance, or 2/5 of the total. The potential difference across the two resistors is shown by the voltmeter to be 2.4 volts. This is 2.4/6.0 or 24/60 or 2/5 of the total resistance in the whole circuit. So we find that the fraction of total potential difference in the circuit is the same as the fraction of the total resistance.

EFFECTS OF INCORRECT RESISTANCE. At the top of Fig. 8-13 is a diagram of the circuit which was pictured by Fig. 8-4. We have the 6-volt power supply, the 8 ma current, and the total load resistance of 750 ohms. By using our formula for potential difference it is possible to compute the potential difference across each resistance in the load. These potential differences are written on the upper diagram.

In the lower diagram the resistance of the load unit second from the left has been increased from 200 ohms to 650 ohms. There might be many reasons why resistance in some part of a radio or television circuit should become abnormally high. The sum of the circuit resistances now is 1,200 ohms. With 6 volts from the power supply acting on 1,200 ohms, the current (from your current formula) will be 5 milliamperes instead of the former 8 milliamperes.

Now, using the formula for potential difference, we may compute the voltage across each of the resistances. These voltages are written below the units in the lower diagram. Compared with "normal" voltages up above, the voltage across the excessive resistance has increased from 1.6 to 3.25. But the voltages across every other unit in the circuit has dropped at the same time.

Excessive resistance in some one part raises the voltage while dropping the current for that part, and at the same time drops both voltage and current for all other parts in the same series circuit. On the other hand, too little resistance in some one part will drop the voltage while increasing the current for that part, while increasing both voltage and current for all other parts.

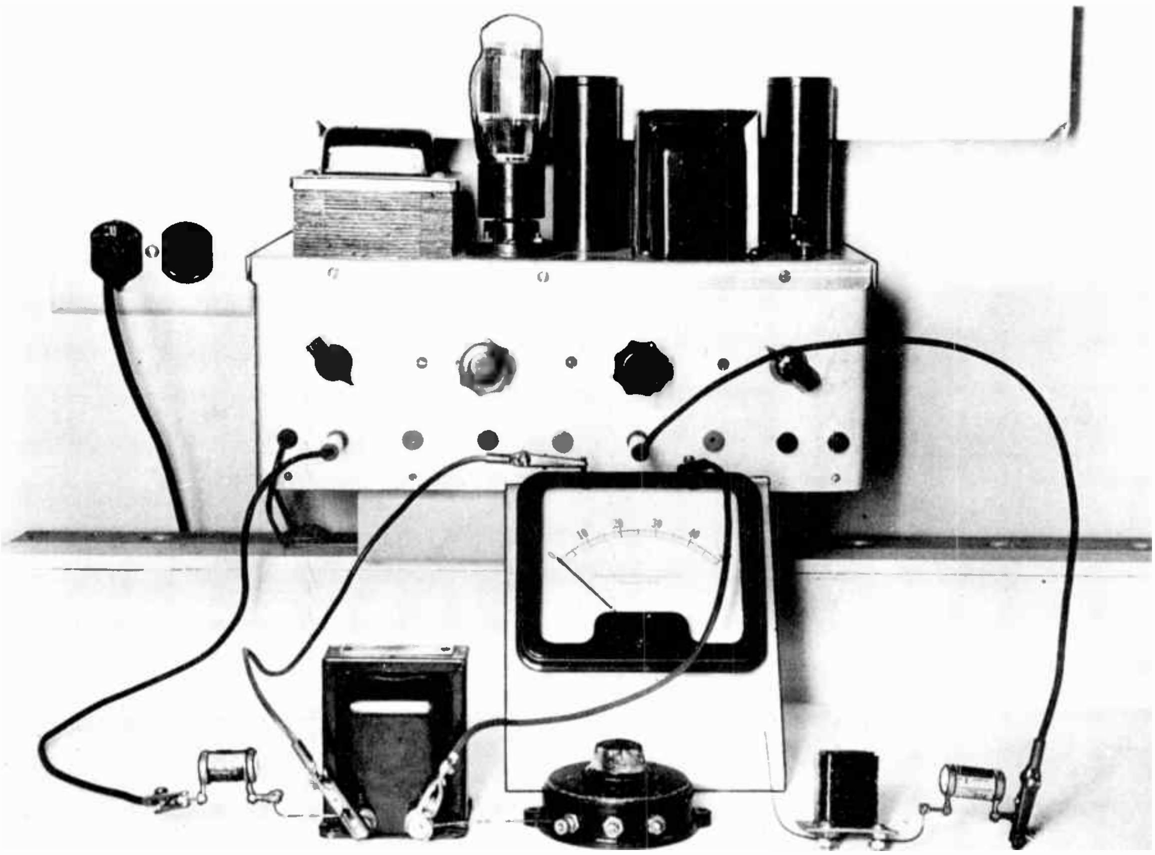


Fig. 8-15. There is no potential difference anywhere in a series circuit which is open.

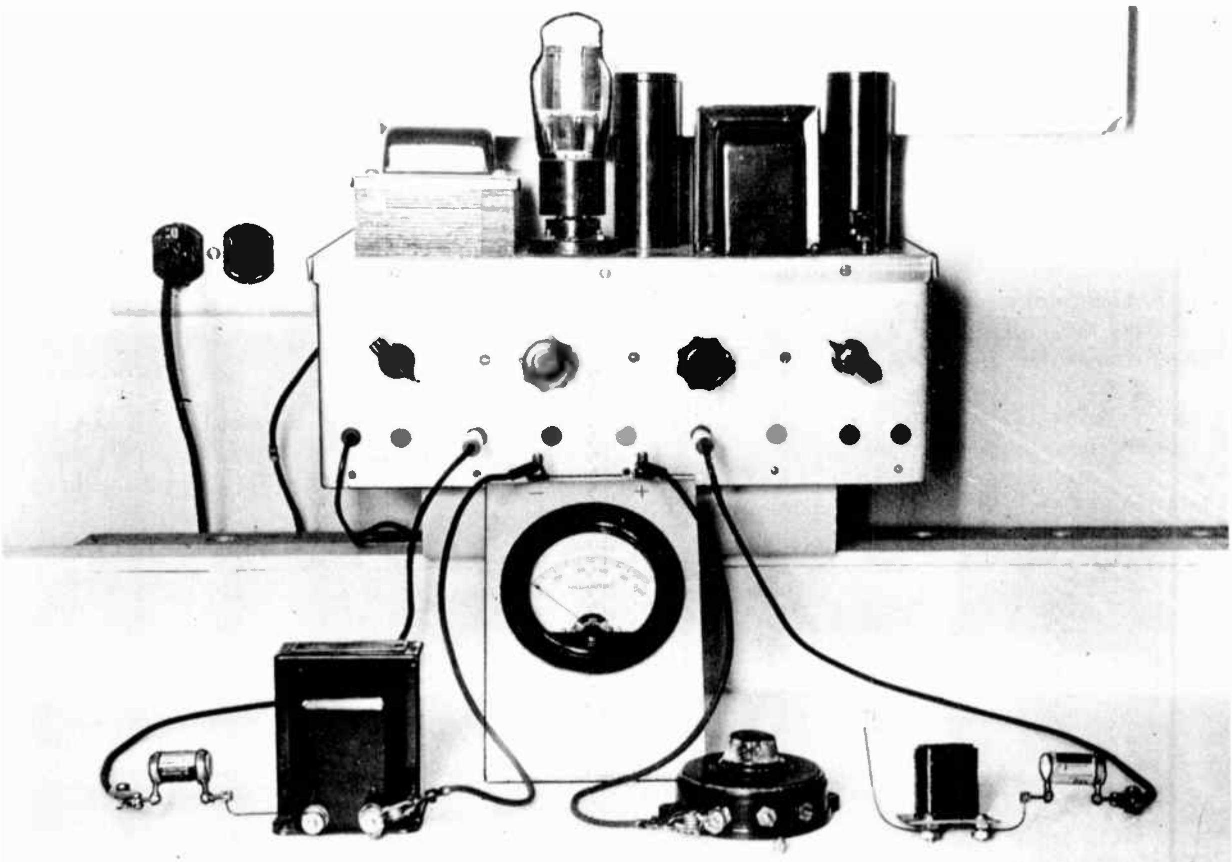


Fig. 8-16. There is no current anywhere in a series circuit which is open.

To prove this last statement, change the resistance of the right-hand unit to 100 ohms. Then your circuit resistances will look like this:

100 ohms	200 ohms	200 ohms	100 ohms

Compute the total resistance.

Use your formula for milliamperes to compute current in the circuit. The power supply still applies 6 volts.

Use your formula for potential difference to compute the voltage across each unit, using in this formula the circuit current and the resistance of each unit.
The voltages should come out like this:

1 volt	2 volts	2 volts	1 volt

WHERE TO MAKE MEASUREMENTS. Whenever you use our trouble shooting formulas the resistance and the potential difference must be in the same part. If you use resistance of one part and potential difference of another part in the same formula the answer will mean nothing at all. Current likewise must be in the same part, but in a series circuit the current is the same in all parts, so we may measure current anywhere and know it is the same throughout the series circuit.

In Fig. 8-14 the series circuit includes many parts. The voltmeter (the left-hand instrument) is connected across the filter choke to measure potential difference on this one part. The milliammeter (at the right) is connected in series between the adjustable resistor and the transformer winding. The current measured here is flowing also in the choke whose potential difference is being measured. We may use the measured potential difference and the measured current to compute resistance of the choke.

Not all circuits are series circuits. In the plate circuit which we examined in an earlier lesson the same current does not flow in the plate and in the screen of the output tube. There we have one example of divided circuits. Later we shall learn some trouble shooting formulas which apply especially to divided circuits. Those formulas will be based on the principle of considering each branch of a divided circuit as though it were a separate small series circuit.

EFFECTS OF AN OPEN CIRCUIT. In a series circuit it must be possible for electron flow to exist all the way from one side of the power supply through every part of the circuit and back to the other side of the power supply. If there is any point where electrons cannot flow, that point is called an "open" or an

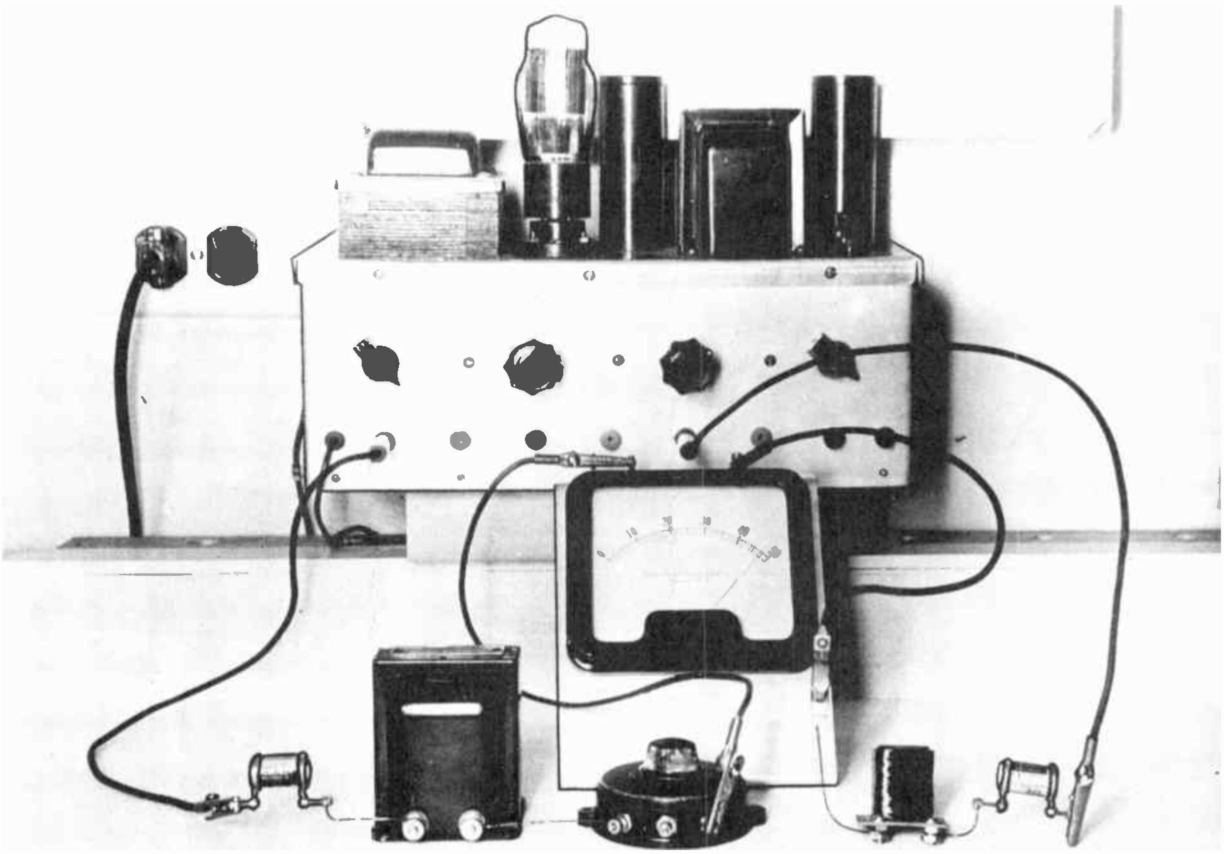


Fig. 8-17. A voltmeter connected across an open point reads nearly the full voltage furnished by the power supply.

open circuit.

In Fig. 8-15 we have disconnected the milliammeter from between the adjustable resistor and the transformer winding, where we had this meter in the preceding setup. Now there is an open between the adjustable resistor and the transformer winding. The voltmeter still is connected across the filter choke, but now the voltmeter reads zero.

The voltmeter reads zero because there is no current in the filter choke. No current means no electron flow, and with no electron flow there can be no loss of electron energy. Loss of electron energy must be accompanied by change of potential. Then with no electron flow there can be no difference of potential.

There is no current in the filter choke because current must be the same everywhere in a series circuit. There can be no current across the open point. Therefore there can be no current anywhere else in this circuit. In Fig. 8-16 the milliammeter has been connected between the filter choke and the adjustable resistor. The open point still exists between the adjustable resistor and the transformer winding. The milliammeter reads zero current. No matter where an open may be, it stops current everywhere in the series circuit.

In Figs. 8-15 and 8-16 the power supply has remained turned on, and has been applying a potential difference across the ends of the series circuit. This potential difference causes no electron flow or current because of the open point. In Fig. 8-17 the voltmeter has been connected across the open point in the series circuit. The voltmeter now reads practically the full potential difference of the power supply, whereas in Fig. 8-15 the voltmeter read zero. Can you think of a good way to locate an open point in a series circuit by using your voltmeter as the testing instrument? With the meter bridging across parts which are not open the readings will be zero. When you get the meter across the part which is open the meter will indicate a voltage, a potential difference.

A voltmeter permits flow through it of some current in spite of the high resistance of the instrument. A voltmeter really measures potential difference across its own internal resistance. When you get the voltmeter across an open point in a series circuit you have placed the meter resistance across the open and have closed the circuit through this resistance. Then there will flow, in the entire circuit, the very small current permitted by the high resistance of the meter. With this very small current there will be hardly any loss of potential in other parts of the circuit, where resistance is very small in comparison with that of the voltmeter. The negligible loss of potential in other parts of the circuit allows practically the whole potential difference of the power supply to act on the voltmeter. That's why a voltmeter reads practically the power supply voltage when the meter bridges an open point.

Every principle which has been illustrated in this lesson is used by competent technicians for trouble shooting in every part of every receiver. We have only hinted at practical applications of these principles, as in the method of locating an open with the help of a voltmeter, and the method of locating resistance which is either abnormally low or abnormally high with principles explained in connection with Fig. 8-13.

In the main we have examined only the principles of trouble shooting, not their practical applications. We must know our tools before trying to use them. Before using these tools to discover what is wrong when parts don't work, we must know much more about how the parts are supposed to work. As we learn about each part it will become apparent how our trouble shooting tools may be used on that part. We won't be delayed by having to talk about the elementary principles of trouble shooting — we already know our tools.

TELEVISION

LESSON NO. 9 TUBES FOR TELEVISION AND RADIO

Training Your Mind...

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

LESSON NO. 9

TUBES FOR TELEVISION AND RADIO

A fairly complete listing of television and radio tubes, according to type numbers assigned by their manufacturers, would include about 1,300 kinds. Fortunately, we won't have to study 1,300 different types of tubes, for the same tube may be given different type numbers by different makers. However, if we eliminate all such duplication, also take out all tubes used exclusively for transmitting, and all such special kinds as picture tubes, to leave only those ordinarily classed as radio and television receiving tubes, more than 400 kinds remain.

In this lesson we are not going to examine television picture tubes because they are entirely unlike any other tubes in our receivers. For the present it will be enough to study amplifiers, detectors, rectifiers, oscillators, and all the other tubes which are constructed quite similarly to one another, leaving picture tubes for later.

Were each of the 400 receiving tubes to differ in principle from all the others we should be undertaking a formidable job in studying them. But this number is considerably reduced because, while quite a few types are identical so far as performance goes, they are given different numbers because of differences in mechanical construction, in voltages required by heaters, and in other relatively minor details.

The most obvious difference between tubes is in the material of the envelope, which may be glass or metal. The largest metal envelope for a receiving tube is pictured at the left in Fig. 9-1, and alongside it is the smallest. Next toward the right is a metal tube with an insulated top cap from which an internal connection goes to the control grid. All other elements are connected to base pins. At the right is a metal type having all internal elements, including the control grid, connected to pins on the base. This latter may be called a single-ended tube.

In Fig. 9-2 are illustrated three sizes of glass envelopes. The tube at the left is a heavy-duty rectifier. The center tube is of a size commonly used almost anywhere in sound radio receivers, and quite often in the sync and sweep sections of television receivers. The tube at the right has a top cap to which is connected its control grid.

As an illustration of how different type numbers may apply to tubes of identical electrical characteristics we may consider first the type 6J5. This is a triode, meaning that it has three active elements consisting of cathode, control grid, and plate. The envelope is metal, as at the right in Fig. 9-1. The base is of the octal type. The heater requires a potential difference of 6.3 volts.

If the same tube is built with a glass envelope such as pictured at the center of Fig. 9-2 it becomes type 6J5-GT. In the two added letters "G" stands for glass and "T" stands for tubular. All "GT" tubes have glass envelopes of the shape and size shown at the center of Fig. 9-2.

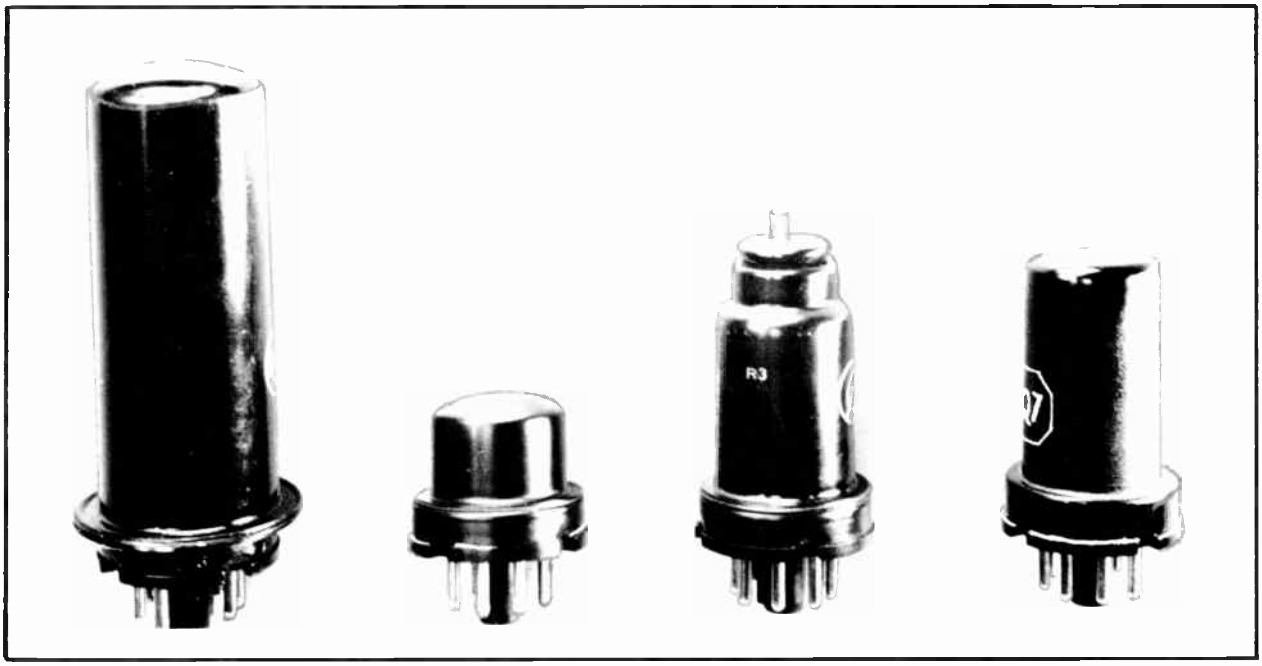


Fig. 9-1. Tubes with metal envelopes.

Now for a third type. If the same cathode, grid, and plate structure of the 6J5-GT are used with a heater which requires 12.6 volts the tube becomes type 12J5-GT. In general, when the first digit of a type number is 6 is means that the heater requires 6.3 volts for normal operation. If the first two digits are 12 it indicates a heater which requires 12.6 volts for normal operation.

So far all our variations of the original tube have had octal bases. Such bases, as you will recall, have positions for eight pins although fewer than eight may be used. On the bottom of an octal base is an extended central pin with a locating key on one side. If we change the type of base while retaining the glass envelope of GT size and shape we have a type 7A4. The type 7A4 has a lock-in base, illustrated in Fig. 9-3. The contact pins are sealed into the glass bottom. Around the bottom is a metal shell and a central pin. Down below the locating key on the pin is a groove which fits into a catch on the socket. Lock-in tubes are pressed down into their sockets like any other type, but for removal you should give a slight off-side pressure to release the socket lock. When the first digit of a type number is 7 it means that the heater will operate with potential difference of either 7.0 or 6.3 volts.

Now for another change. If the lock-in 7A4 tube were built with a heater which will operate with either 14.0 or 12.6 volts potential difference the tube becomes lock-in type 14A4. Remember, we still have exactly the same operating characteristics with which we commenced in the type 6J5. A change of rated heater voltage does not affect operating characteristics.

Next comes a change which adapts our tube for the many applications where a pair of 6J5's would be required. Instead of using two separate 6J5 tubes we may employ only one type 6SN7-GT. This type is precisely like two 6J5's built into a single GT envelope. The 6SN7-GT has an octal base. The heater requires 6.3 volts. If the heater is changed to require 12.6 volts we then have a type 12SN7-GT. These two types of

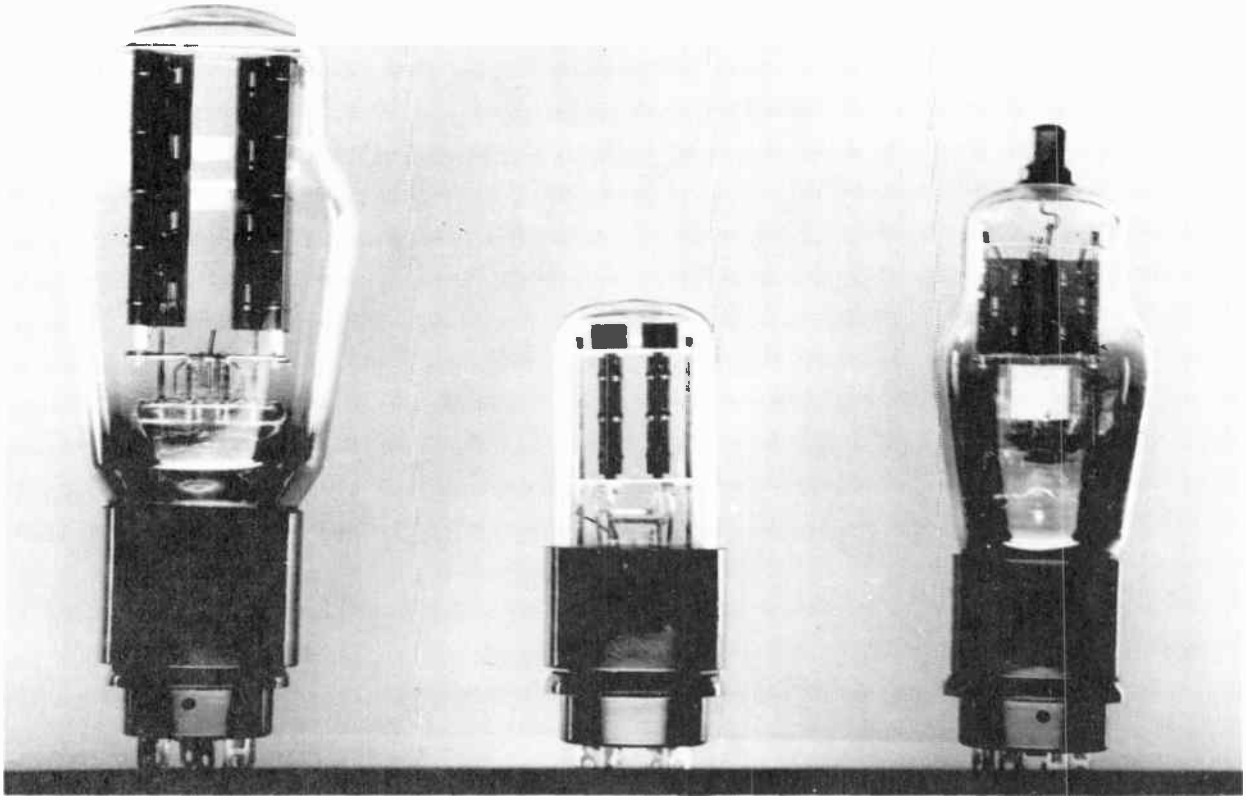


Fig. 9-2. Tubes with glass envelopes.

twin-triode tubes are widely used in television receivers of all kinds. We are not yet through. A tube built like the 6SN7-GT except for having a lock-in base becomes a type 7N7. And one built like the 12SN7-GT except for having a lock-in base becomes a type 14N7. Still, in either triode section by itself, we have the same operating characteristics as in a single 6J5.

We have talked about tubes of nine different type numbers, all having the same operating characteristics. The different numbers are required in order to identify the envelope material, the size and shape of the envelope, the type of base, the heater voltage, and the number of active elements. Different type numbers are used also when into a single envelope are built two or more sets of elements which may operate more or less independently. An example of this is the two 6J5's in a single envelope to form one 6SN7-GT.

In any multi-section tube or multi-purpose tube there will be two or more electron paths inside the envelope. At the left in Fig. 9-4 is a symbol for the 6SN7-GT tube. There are two cathodes, two grids, and two plates, with entirely separate electron streams in the two sections. One section includes a cathode connected to pin 3, a grid connected to pin 1, and a plate connected to pin 2. The other section has cathode, grid and plate connected respectively to pins 6, 4, and 5. Pins 7 and 8, which are not shown on Figure 9-4, are used for the heater. Although this 6SN7-GT tube has six active elements it really has only three kinds

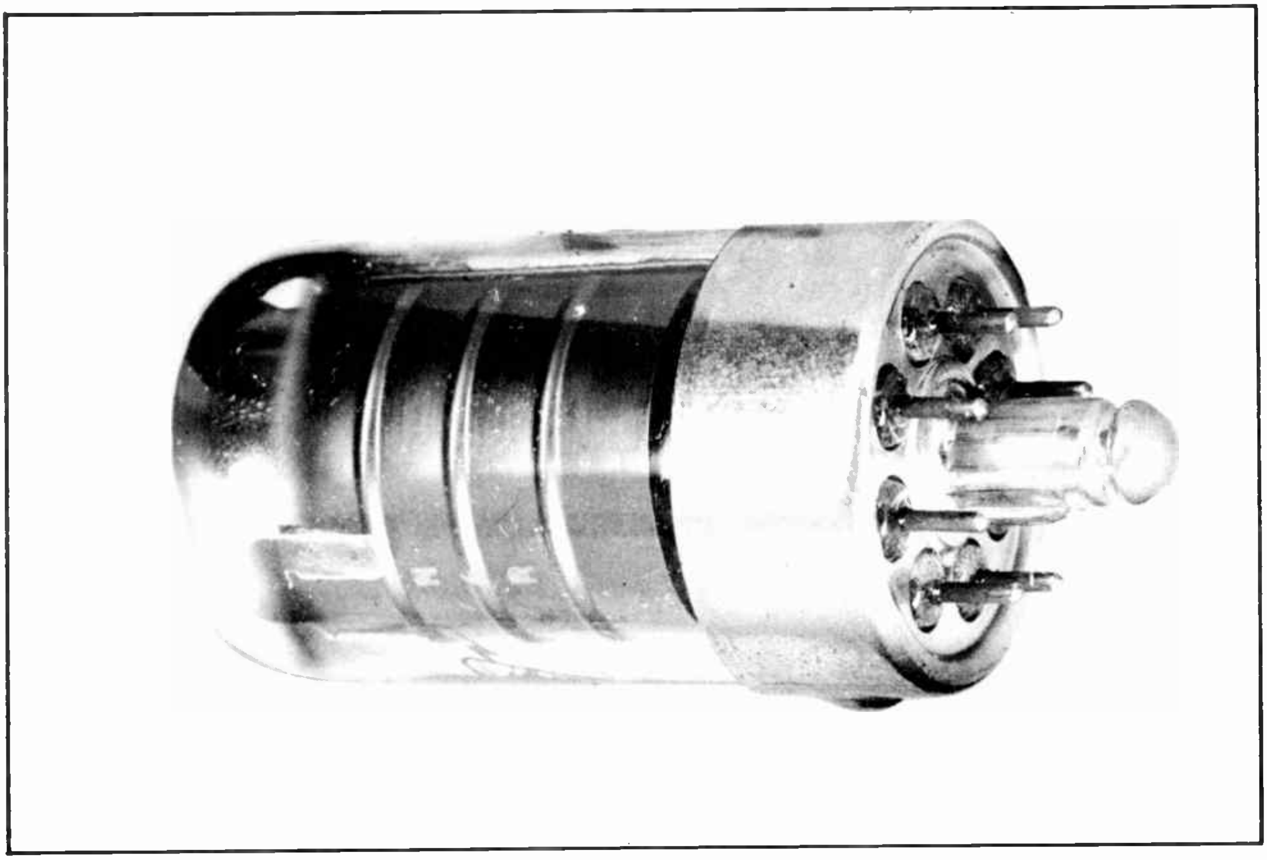


Fig. 9-3. This is a lock-in base.

of elements, with two sets of each.

Now we shall look at a tube in which there are more than two electron streams. At the right, in Fig. 9-4, is a symbol for a 6T8 tube, a type used in the sound sections of many television receivers. This tube has nine base pins. There are two cathodes, connected to pins 3 and 7. Electron flow from the cathode on pin 3 goes only to the plate on pin 2. The cathode connected to pin 7 carries electron flow which separates into three streams; one for the plate on pin 1, a second for the plate on pin 6, and a third which goes through the grid connected to pin 8 and to the plate connected to pin 9. Pins 4 and 5, which are not shown on the diagram, are used for the heater. This 6T8 tube has seven active elements, but consists really of three cathode-plate pairs and of one three-element combination which includes the common cathode with a separate grid and separate plate.

Nearly all receivers employ some multi-section tubes. There are dozens of combinations, but all are made up from a relatively few simple types. If we forget all the multi-section combinations and consider only the basic groups of elements which might be used in a single tube, we may make a convenient classification based on the elements which may act on a single electron stream. The accompanying table, "Tube Types and Their Active Elements", shows this classification.

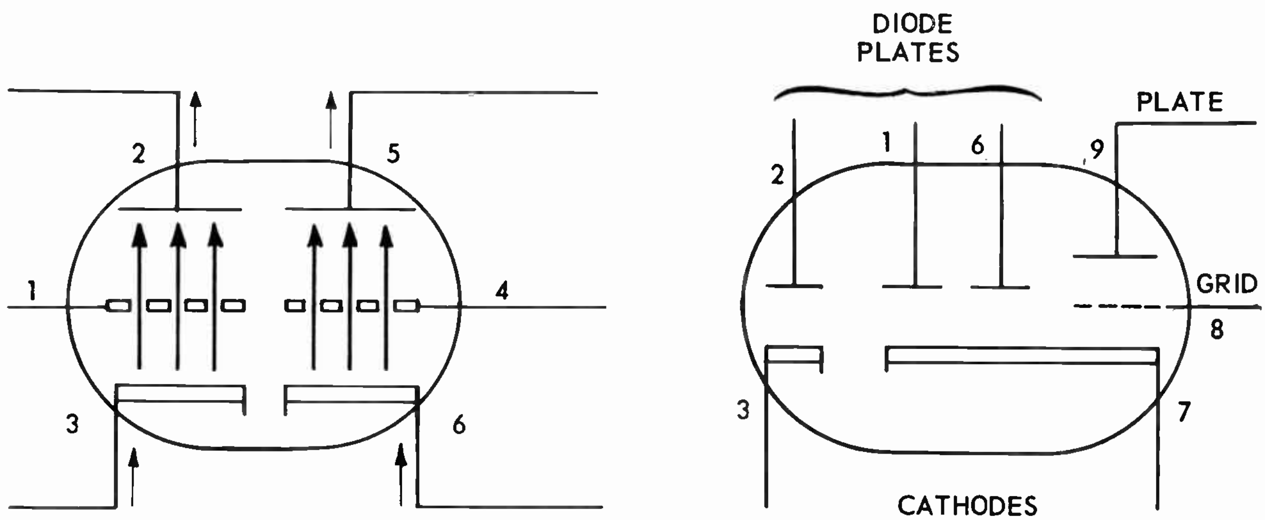


Fig. 9-4. Symbols for multi-section tubes in which there are two or more separated electron streams.

TUBE TYPES AND THEIR ACTIVE ELEMENTS

TYPE NAME	Total Elements	How Many of Each Kind of Element Cathode	Grid	Screen	Suppressor	Plate
Diode	2	1				1
Triode	3	1	1			1
Tetrode	4	1	1	1		1
Pentode	5	1	1	1	1	1
Beam power	5	1	1	1		2
Pentagrid and Heptode Converters	6	1	2	1	1	1
Octode Converter	7	1	3	1	1	1

The type name changes each time an active element is added. We commence on the top line with the simplest possible tube, the diode. A diode has only a cathode and a plate. When we add a grid between cathode and plate the tube becomes a triode. Adding a screen between the grid and the plate produces a tetrode. Adding a suppressor gives us a pentode.

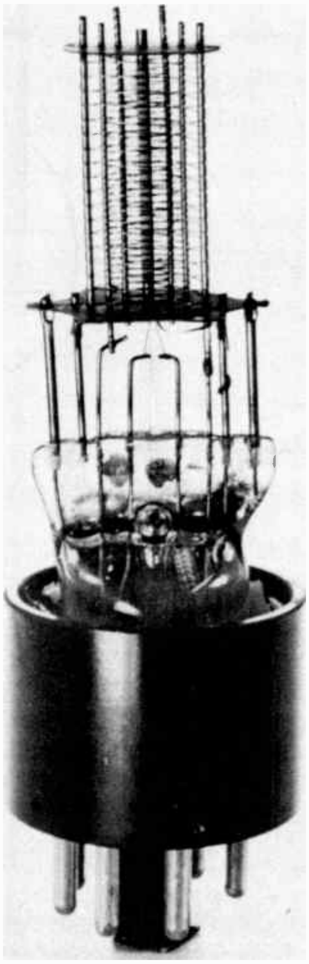


Fig. 9-5. The outermost spiral of wire is the suppressor for this tube which would be a pentode if complete.

This is the first time we have met the name suppressor. This element is a spaced spiral of wire placed between the screen and the plate. The outermost spiral which you see in Fig. 9-5 is a suppressor. The suppressor is around the outside of the screen. The screen is around the outside of the grid. The grid is around the cathode. The active elements for this tube would be completed by placing the plate around the outside of the suppressor.

The purpose of the suppressor is to prevent electrons going backward from the plate to the screen. You will recall that a screen, like a plate, is operated at a positive potential with reference to the cathode. Were it not for the suppressor, a very strong signal might make it possible for electrons to go from the plate to the screen as well as from the cathode to both plate and screen. This would lessen the plate current and reduce signal output.

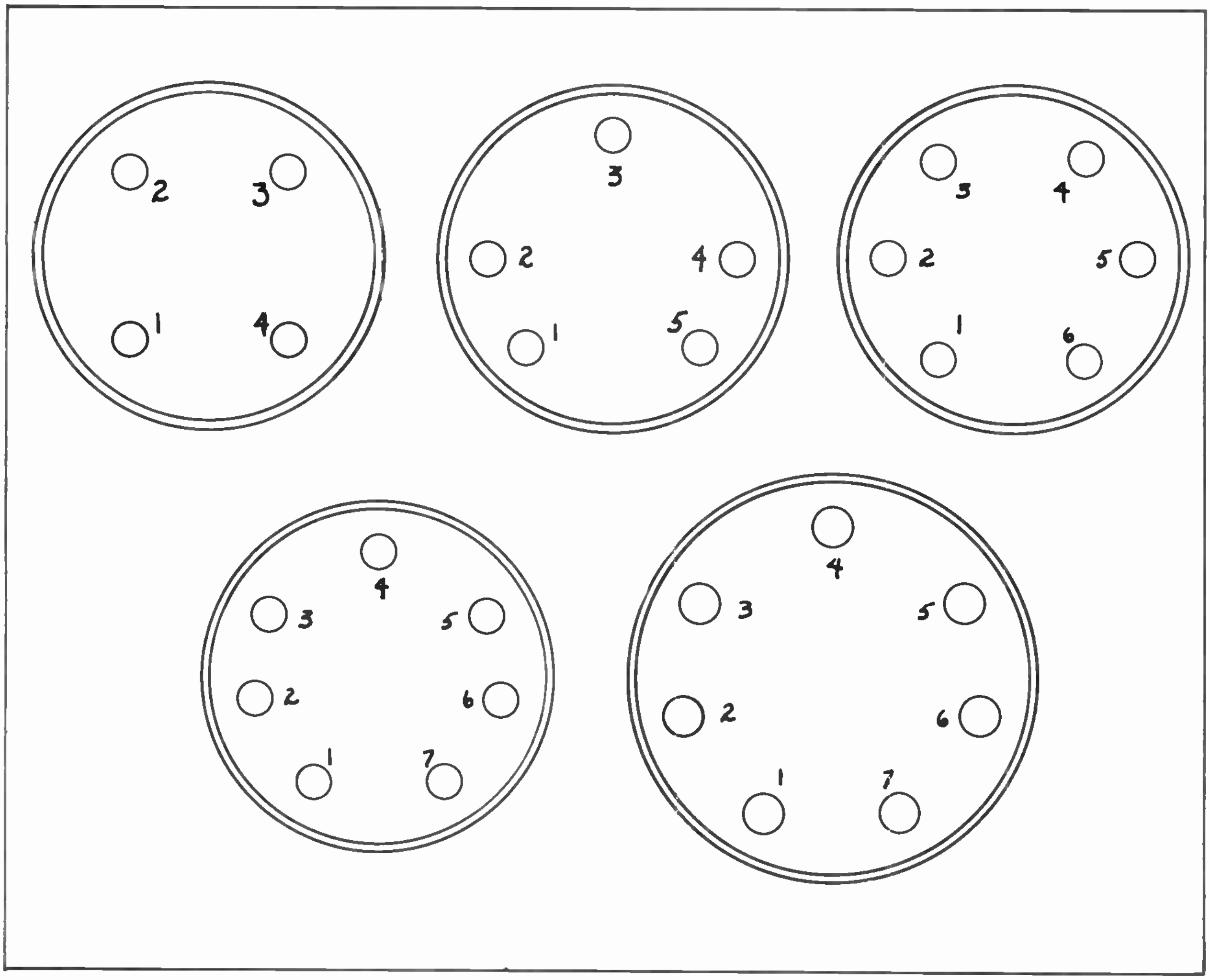


Fig. 9-6. Pin arrangements used for 4-, 5-, 6- and 7-pin tube bases.

Next below the pentode in the table is another 5-element type called a beam power tube. This type is a modification of the pentode which provides smooth and uniform amplification while delivering sufficient power to operate a large loud speaker or to do any other job calling for considerable power in the output signal. The electron stream within this tube is concentrated b y extra elements called beam formingplates.

On the two bottom lines of the table are listed tubes called converters, which have six or seven active elements. These tubes are used as combined mixers and oscillators in superheterodyne receivers for sound. One of the two grids in the pentagrid converter serves the mixer function, the other serves the oscillator function. In the octode converter two of the grids are used as mixer and oscillator grids, with the third grid acting like a plate for the oscillator function. All these features will be investigated more fully in due time.

Tubes may be classified in accordance with what they do as well as with respect to the number and kind of elements. For instance, we may speak of amplifier tubes, oscillator tubes, detector tubes, and so on. The next table, "Tube Functions, and Types Employed", lists the principal purposes for which tubes are used, and shows the types most often employed for each purpose. This table is based on common practices of today. With any listing so general as this there will be exceptions. As one example, there is nothing to prevent using frequency control tubes in sound radio receivers as well as in television receivers. Such tubes have been used in sound receivers, but it is not now a general practice.

You will note that the tetrode, a 4-element tube, is not included in this table of functions. The tetrode, once very popular, is no longer used to any extent. This type turned out to be only a stepping stone between triodes and pentodes. The tetrode overcame shortcomings of the triode in certain applications, but had troubles of its own which were overcome by the pentode.

TUBE FUNCTIONS, AND TYPES EMPLOYED

NAMES ACCORDING TO FUNCTION	WHERE USED		Types Employed			
	Sound Radio	Tele-vision	Diode	Triode	Pentode or Beam	Converter
R-f Oscillators and Mixers		X		X	X	
Converters	X					X
Amplifiers	X	X		X	X	
Demodulators, Detectors and Discriminators	X	X	X			
Restorers		X	X	X		
Limiters, Clippers, Separators		X		X	X	
Inverters	X	X		X		
Sweep Oscillators		X		X	X	
Frequency Control		X	X	X	X	
Dampers		X	X	X		
Rectifiers	X	X	X			

TUBE BASES. Before taking up the manner in which the various elements enter into the operation of tubes it may be well to look at the layouts of some tube bases, and get acquainted with arrangements of the pins through which connections are made to the internal elements.

In an earlier lesson we talked about the octal base for tubes, on which are spaces for eight pins. Some-

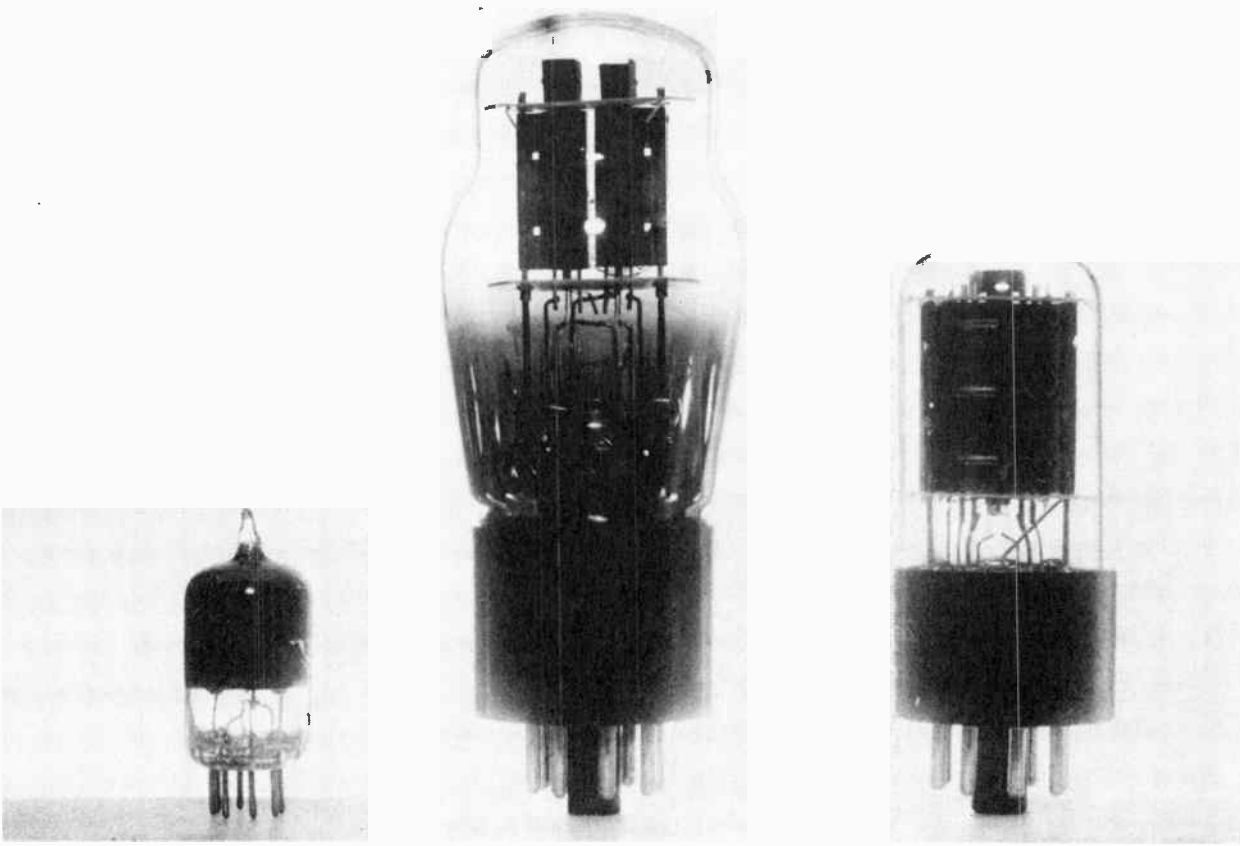


Fig. 9-7. A miniature tube alongside two of its bigger brothers.

times all eight pins appear on the base of a tube. Again only as many pins are used as needed for the active elements and heater, or an extra pin or two may be added to give the tube more support in its socket. Octal bases are used on all tubes having metal envelopes, and on a large percentage of those having glass envelopes.

Early types of radio receiving tubes were fitted with bases of which bottom views are shown in Fig. 9-6. On the 4-pin base the two pins numbered 1 and 4 are 5/32 inch diameter, while pins numbered 2 and 3 are 1/8 inch diameter. Difference between pin diameters insures getting the tube correctly located in its socket. On the 5-pin base all pins are 1/8 inch in diameter. Their unequal spacings insure correct positioning of the tube in its socket. On the 6-pin base the two large-diameter pins are numbered 1 and 6, and on both sizes of 7-pin bases the two large-diameter pins are numbered 1 and 7. When two pins are of larger diameter than others the large ones connect to the heater or to the filament-cathode of the tube.

None of the bases shown by Fig. 9-6 are used on recent types of tubes. However, the 4-, 5-, and 6-pin bases quite often are used on such items as resistors, crystals for frequency control, and other parts which are convenient to work with when provided with a plug-in mounting such as a tube base and socket.

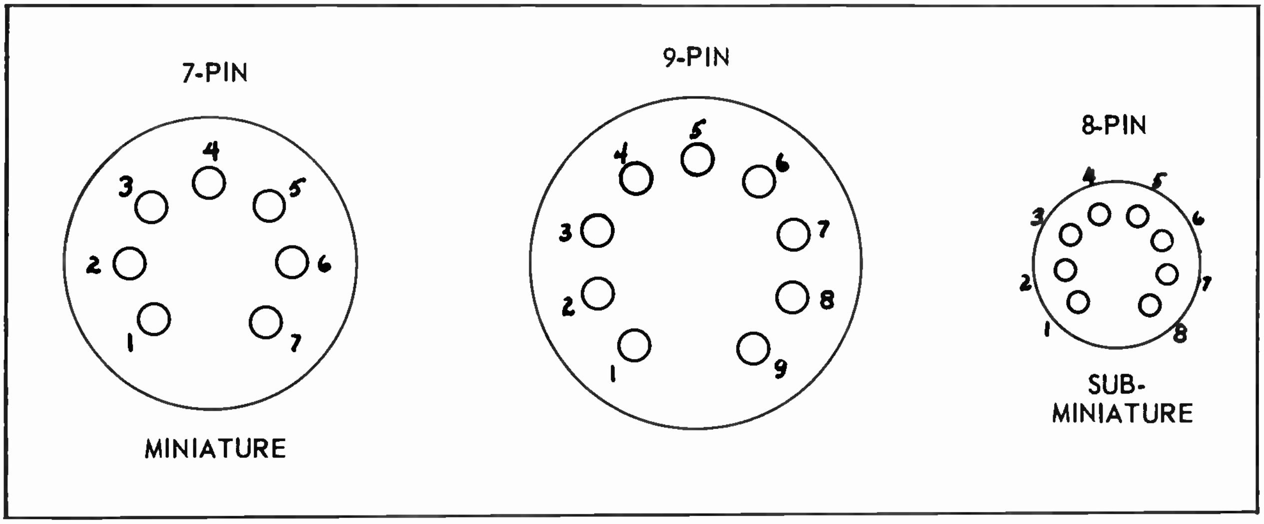


Fig. 9-8. Pin positions and numbering as used for miniature and sub-miniature tubes.

In most recently designed television receivers a great many of the tubes are miniature types. A miniature tube is at the left in Fig. 9-7. In the center is a size often found on rectifiers and on some heavy-duty amplifiers. Over at the right is a GT type. Some miniature tubes are 3/4 inch in outside diameter, and have 7-pin bases. Other miniatures are 7/8 inch in diameter, and have 9-pin bases, A still smaller tube, called a sub-miniature, is only 2/5 inch in diameter. It is fitted with an 8-pin base.

Pin positions on bases for miniature and sub-miniature tubes are shown in Fig. 9-8, as the bases appear from the bottom. Base pins on miniature tubes are 1/25 inch in diameter, and on sub-miniatures are about 1/60 inch in diameter. On all these bases there is a gap at one place in the circle of pins – at a point where another pin might be, but isn't. When looking at the bottom of the tube, or at the bottom of its socket, numbering of the pins always starts with number 1 at the left of this gap, and proceeds clockwise, as shown on all the diagrams.

It is important to remember this numbering system, for there are no visible numbers on the tubes. Even when the small sockets have numbers they may be difficult to see when surrounded and partially covered by wiring.

Now we have a general idea of where and how the principal kinds of tubes may be used, and we know something about the structural features. It is next in order to learn how tubes actually operate. To begin with we shall see what happens when there is only a cathode and a plate. Then other elements will be added in extending the usefulness of the tube.

CATHODES. We have learned that heating gives free electrons in the cathode enough extra energy to let them get through the surface and into the space near the cathode. There are other ways of imparting the energy necessary for electron emission, as with light in the case of phototubes and photocells, but since all tubes in modern radio and television receivers for homes and business places are operated with heated cathodes, we shall confine our studies to this general type.

In most tubes whose construction has been examined the cathode is a small cylinder which surrounds the

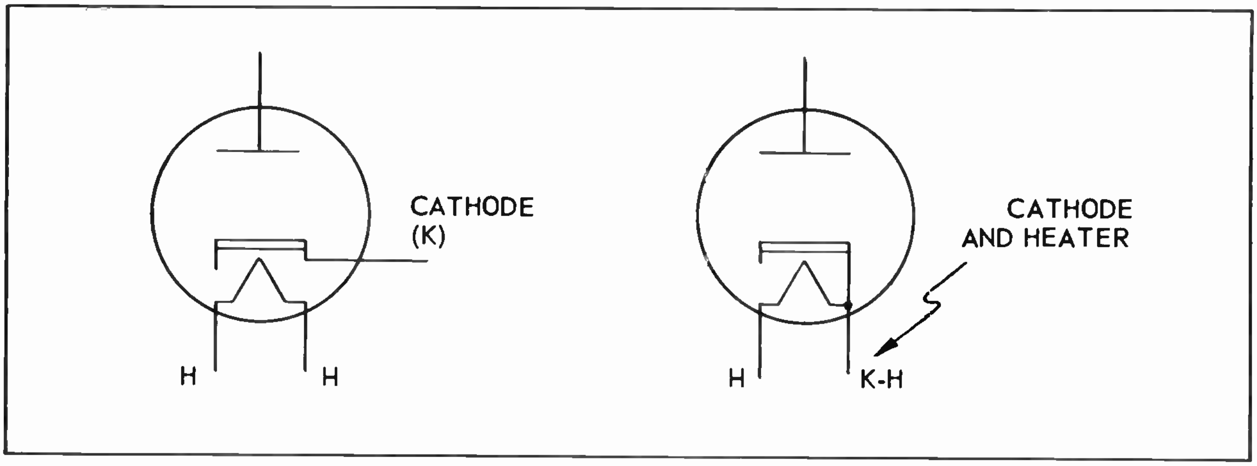


Fig. 9-9. Symbols for diode tubes having heater-cathodes.

heater. The cathode and heater may be completely insulated from eachother inside the tube, with one base pin for the cathode and two separate pins for the heater. This construction is represented at the left in Fig. 9-9. In other tubes the cathode is internally connected to one side of the heater, as shown at the right. With either of these arrangements the cathode is said to be indirectly heated. It may be called a heater-cathode.

In Fig. 9-10 we have an entirely different kind of cathode, the inverted V-shaped conductor which you can see on the right–hand section of the tube where one of the two plates has been removed. The other plate remains in place. There is no separate heater. Cathode temperature is raised by a heating current which flows in the cathode itself. This is called a directly heated cathode, or a filament-cathode, or just a filament.

Circuit connections for a tube with filament-cathode are shown by Fig. 9-11. At the left is a power source which furnishes heating current for the filament. At the right is a separate power supply furnishing the potential difference which causes electron flow in the plate circuit of the tube. The negative side of the plate power supply may be connected to either side of the filament-cathode. The positive side of the plate power supply is connected through the load to the plate of the tube. Thus the entire length of filament-cathode inside the tube, and its heating circuit on the outside, are made negative with reference to the plate. Then electron flow occurs between cathode and plate just as in a tube having a heater-cathode.

There may be any number of additional active elements in a tube having either a heater-cathode or a filament cathode. The kind of cathode has no bearing on the remainder of the design.

Filament-cathodes are used in many rectifier tubes operating with large currents at moderate voltages. Fig. 9-10 is an example of such a rectifier. This type of cathode is used also in rectifiers operating with small currents at very high voltages, as in power supplies for picture tubes.

In all the great variety of tubes designed for operation with battery power in portable radios the cathodes are of the filament variety. Nearly all recent types of battery operated tubes are designed for a normal potential difference of 1.4 volts across their filaments. These filament-cathodes may be operated with any voltage not exceeding 1.6 volts, and will perform quite satisfactorily with filament potentials as low as

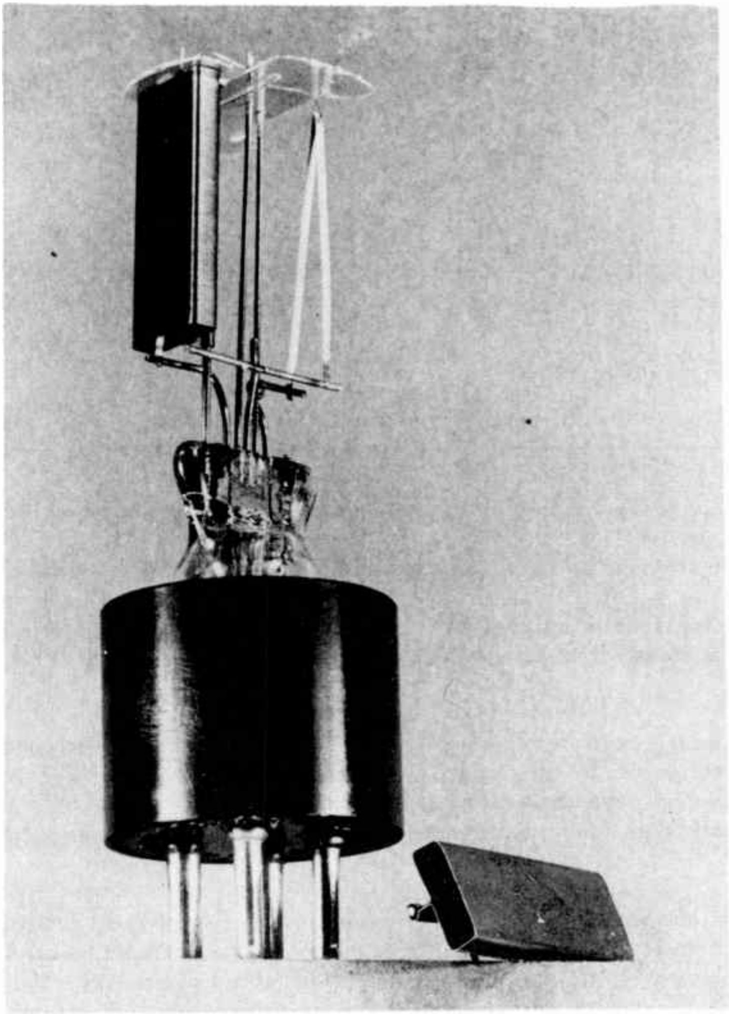


Fig. 9-10. One plate has been removed from this rectifier tube to expose the filament-cathode.

1.25 volts. This range of useful voltages allows connecting the filaments directly across one dry cell, which furnishes a nominal potential difference of 1 5 volts, or across a battery of such cells so connected as to furnish 1.5 volts.

Type numbers of most 1.4 volt battery operated tubes commence with the numeral 1, followed by a letter. Type 1LB4 is an example. Some battery types have a two-part filament which may be connected for operation either on 1.4 volts or else on 2.8 volts as would be furnished by two dry cells connected in series. Type numbers of these tubes commence with the numeral 3, as type 3V4 for example.

There are other battery type tubes designed for operation with a nominal 32 volts on their filaments, as furnished by batteries in some farm-light plants. Still others are designed for 24-volt filament supply from batteries in aircraft.

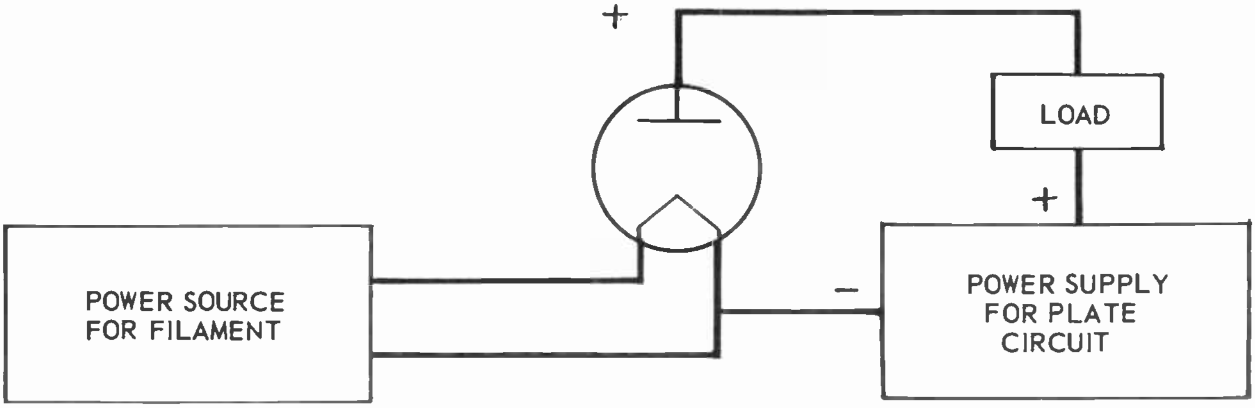


Fig. 9-11. Current for cathode heating and current for the plate circuit both flow in a filament-cathode.

The supporting portion of either a heater-cathode or a filament-cathode is of some metal which withstands high temperature without deterioration. For receiving tubes this support is coated with the active cathode material which contains oxides of barium, strontium, and maybe calcium. These substances have the ability to emit great quantities of free electrons when heated to only a dull red temperature, on the order of 1,500° to 1,700° Fahrenheit. Cathodes constructed in this manner are called coated cathodes.

In earlier types of receiving tubes the entire cathode was made of tungsten containing small quantities of thorium and thorium oxide. This material has to be heated bright yellow, to about 3,000°F., to provide satisfactory emission. The thoriated cathodes are now found in only some transmitting and special purpose tubes. Cathodes of pure tungsten are used in high-voltage transmitting tubes, and were used in the very earliest tubes of all kinds. Pure tungsten must be heated dazzling white, to around 4,000°F., to provide satisfactory emission.

ALL TUBES ARE ONE-WAY ELECTRON VALVES. Examination of Fig. 9-11 shows that the tube would act as a rectifier were the power supply for the plate to furnish alternating potential. We would have one-way electron flow or rectifying action within the tube because only the cathode is made hot enough to emit electrons. Electron flow will occur from cathode to plate only while the plate is positive with reference to the hot cathode, because electrons can flow only from negative to positive.

So far as the matter of element polarity is concerned, electrons could flow from a negative plate to a positive cathode. But no matter what the relations of positive and negative, electrons cannot flow from a cold plate to a hot cathode for the simple reason that no electrons will be emitted from the cold plate.

Direct potential is applied to amplifiers, and to all tubes other than rectifiers, in such polarity as to make the plates positive with reference to the cathodes. However, even were the plates negative with reference to the cathodes there still could be no reverse electron flow, because the plates still would be cold and would not emit electrons. Every tube with a hot cathode acts as a rectifier or a one-way electron valve to the extent of allowing electron flow away from the cathode, and preventing electron flow toward the cathode.

VACUUMS. Almost without exception the tubes in receivers are high vacuum types. Practically all water

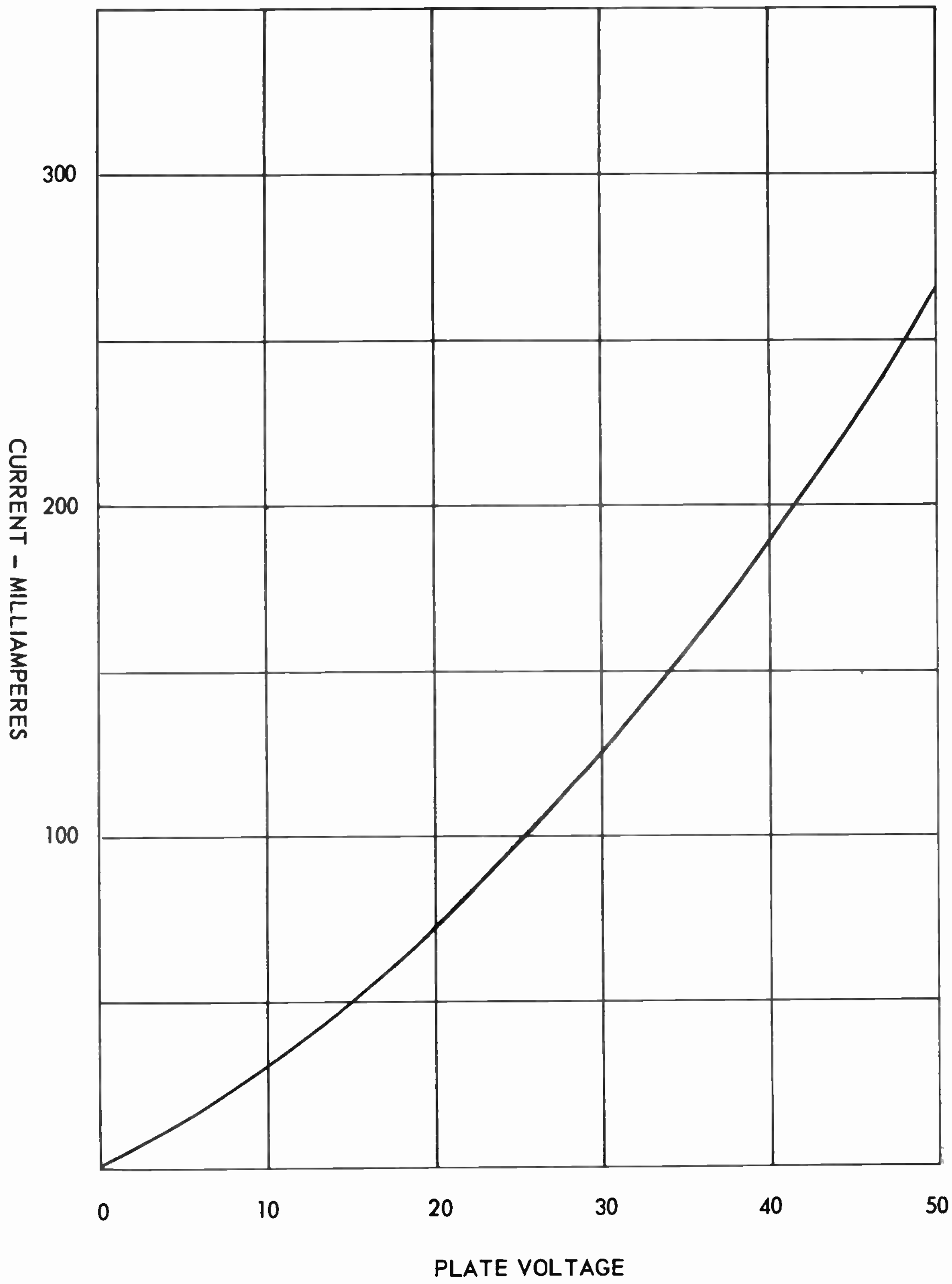


Fig. 9-12. How plate current increases with increase of plate voltage in one type of rectifier tube.

vapor, air, and other gases have been pumped out of the envelope, which then is sealed to retain the nearly perfect vacuum. Were any appreciable quantities of gas to remain, there might be corrosion and oxidation of internal metal parts. Even more important, in a highly evacuated tube there is little chance that atoms of gases will get in the way of electrons speeding from cathode to plate. The chance of an electron colliding with a gas atom isn't one in tens of thousands.

Gases which form air are so dense or so concentrated outdoors at sea level that they exert pressure of 14.7 pounds per square inch on everything exposed to them. The density of gases remaining inside a high-vacuum tube is so little that average internal pressure is only 2½ millionths of one ounce per square inch.

In some tubes there are intentionally small quantities of certain gases, admitted after the tube is first thoroughly evacuated. The purpose of these gases is to permit ionization, an action during which additional electrons are knocked off the atoms of contained gases when the atoms are struck by emitted electrons. About the only purpose for which such gas-filled tubes are employed in home receivers is regulation of voltage in power supplies.

ELECTRON FLOW AND PLATE VOLTAGE. Electron flow from cathode to plate varies inversely with the square of the distance between these elements. This means simply that twice the separation will reduce the flow rate to one-fourth, that three times the separation will reduce the flow to one-ninth, and so on.

For any given distance between cathode and plate the flow rate increases with increase of potential difference between these elements. Potential difference between plate and cathode is called plate voltage. Electron flow or current increases with increase of plate voltage. Current increases somewhat faster than the plate voltage increases throughout the normal range of working voltages for the tube. The relation between plate current and plate voltage is typically as shown in Fig. 9-12, which applies to a heavy-duty rectifier tube.

When electrons are emitted from the cathode surface they move rather slowly at first, but are accelerated to velocities of thousands of miles per second before reaching the plate. The energy of motion (kinetic energy) in the electrons as they strike the plate is enough to raise the temperature of the plate. In large transmitting tubes operating at high plate voltages with large currents this heating effect is so great that the plates have to be cooled with motor driven fans or with circulating water.

ELECTRON FLOW AND CATHODE TEMPERATURE. Supposing we were to operate a cathode at a temperature much below normal. The rate of electron emission from this cathode would be much less than the normal rate. If applied plate voltage then were to be started from zero and gradually increased, the plate current would change about as shown by the full-line curve of Fig. 9-13. At first the increasing plate voltage would draw more and more electrons from the cathode to and into the plate. Soon, however, we should reach a point, A, where all electrons emitted at the low cathode temperature are being drawn away from the cathode. Further increase of plate voltage could draw no more electrons that are being emitted, plate current would increase only slightly, and the current curve would level off as on the graph.

W-e cathode temperature raised to some extent there would be a greater rate of electron emission. Were plate voltage again started from zero and gradually increased, the additional emission would allow plate current to increase up to point B. But again, at B, would come a condition wherein all emitted electrons are drawn from the cathode as fast as emitted. Plate current could increase but little more with further increase of plate voltage, and again the current curve would flatten out for all higher plate voltages.

A third rise of cathode temperature would cause a still greater emission rate, and plate current might follow the curve up to point C before ceasing to increase with increase of plate voltage. When all emitted

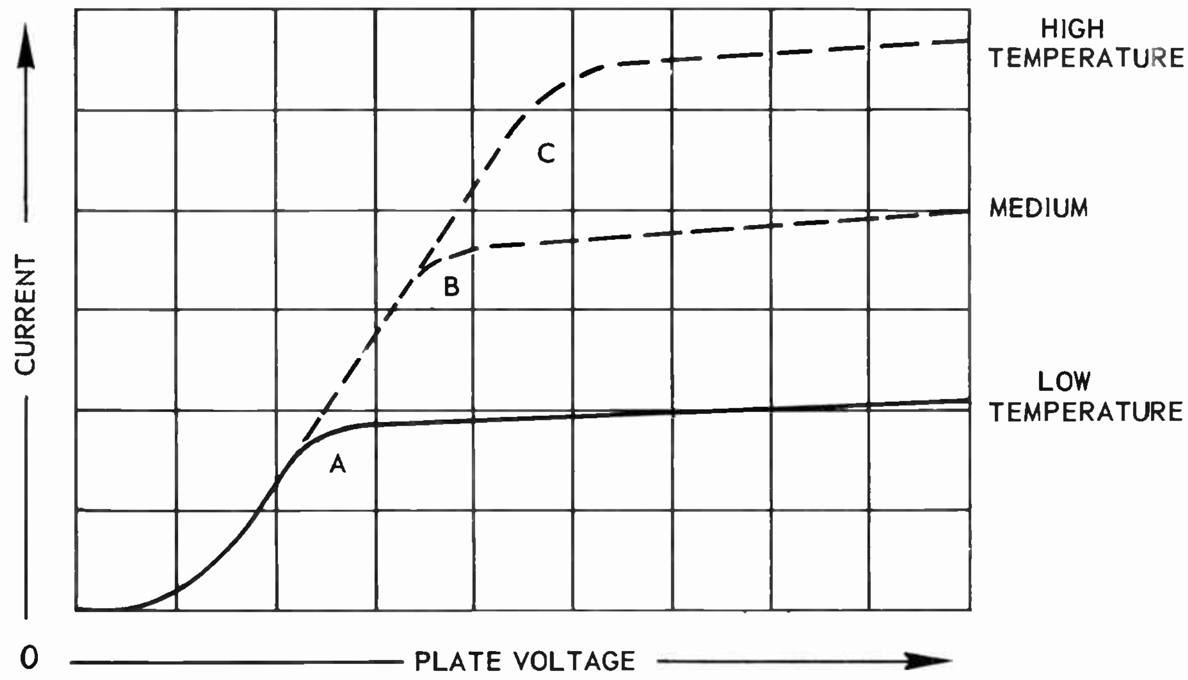


Fig. 9-13. Maximum possible plate current varies with change of cathode temperature.

electrons are drawn to the plate as fast as emitted the resulting value of plate current is called saturation current.

In examples illustrated by Fig. 9-13 the maximum plate current is limited by cathode temperature. This is not a good way in which to operate a tube. There never is any excess of negative electrons at the cathode, and plate potential acts so violently at the cathode surface in an attempt to drag out more electrons that the surface is almost sure to be damaged.

THE SPACE CHARGE. Now let's see what happens when plate voltage is held constant at some moderate value while cathode temperature and emission rate are gradually increased. The constant plate voltage results in a constant positive charge on the plate. Strength of this positive charge, or its ability to attract electrons, depends directly on plate voltage and will remain constant if plate voltage remains constant.

The positive charge of the plate exerts attraction at the cathode. Each atom at the plate surface which lacks an electron is reaching across the tube space and pulling on any negative electron which may be emitted from the cathode. With no emission, a cold cathode, we have the condition represented at A in Fig. 9-14. All the attractive force of the plate is working on the cathode, but it finds no electrons.

At B the cathode has been heated to some low temperature, and there is some emission of negative electrons from the cathode. This small emission furnishes enough electrons to satisfy part of the attractive force (positive charge) of the plate, but the remainder of the attractive force finds no electrons.

At C the cathode temperature has been raised to a point where emission is just enough to satisfy all the

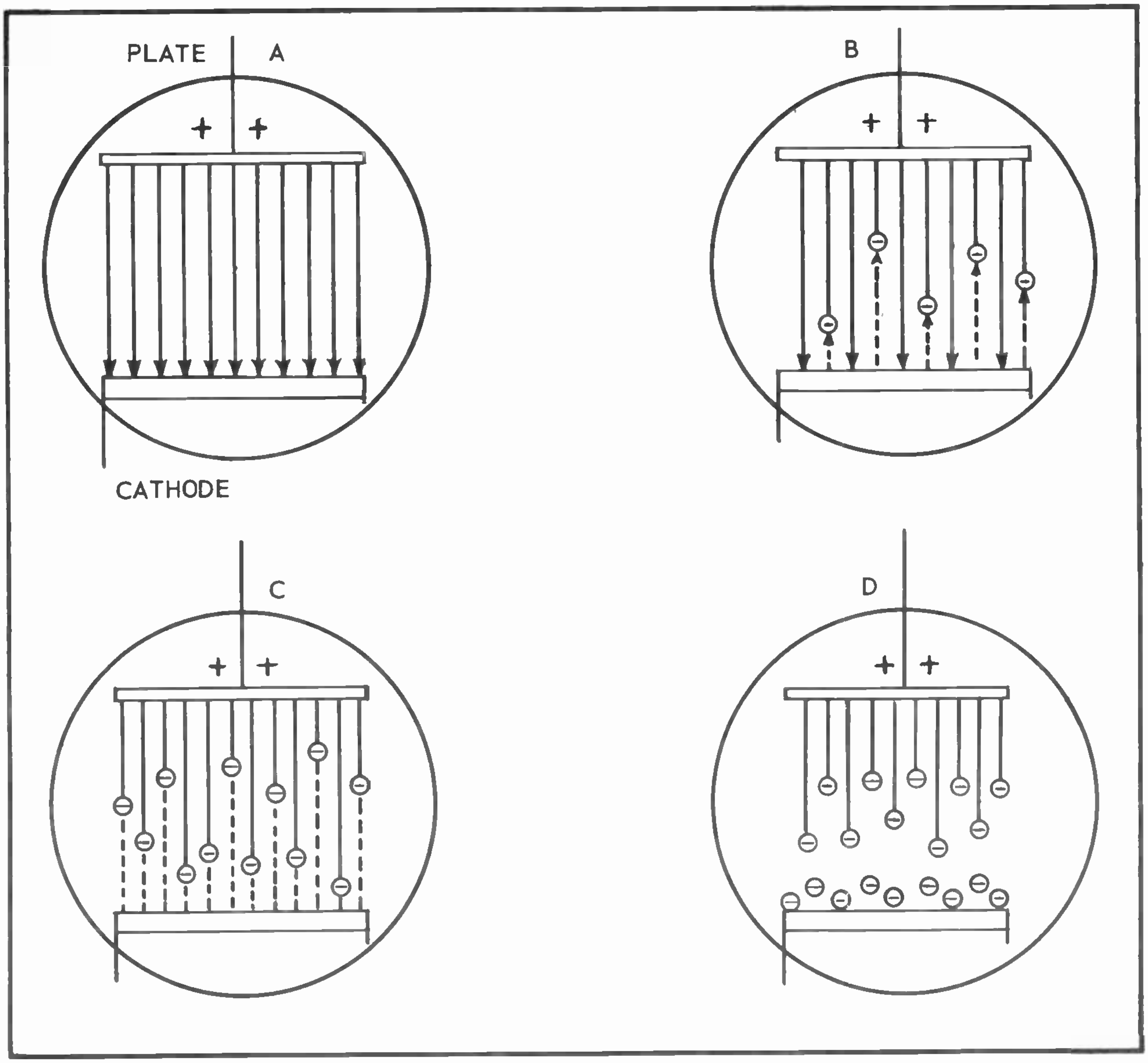


Fig. 9-14. Increasing the emission rate while plate voltage remains constant results in formation of a space charge.

attractive force from the plate. We might think of the situation this way: Every positive atom on the plate exerts one unit of attractive force. There are just enough electrons being emitted to furnish one negative electron for each unit of positive force from the plate. There is no positive force left unsatisfied, but neither are any extra free electrons left at the cathode.

At D we have still greater emission, more than enough to satisfy the entire attractive force corresponding to the applied plate voltage. But eveu with extra electrons available no more of them can be pulled to the plate, for there is no unsatisfied positive charge. Plate cnrrent will be the same as with condition C. The excess of emitted electrons remains in the space near the cathode. This excess is called the space charge.

The space charge is negative, being composed of negative electrons. This negative charge opposes or repels any additional negative electrons which are trying to get out of the cathode because of heat energy. There develops a balance between the emitting force of the heat and the repelling force of the negative space charge. Emission proceeds only fast enough to maintain this balance of forces.

When electrons are drawn from the space charge to the plate there is a reduction of the negative space charge. This means less repulsion for emitted electrons, and more electrons will come out of the cathode. Quite plainly, emission can be at only the same rate as electron flow to the plate – emission and plate current are equal. Plate current is being limited by the space charge, and we say that the tube is being operated "space charge limited". This is the normal and usual method of operating radio and television tubes. The space charge protects the cathode surface. It also forms a sort of "bank" of reserve electrons which may be drawn upon to meet momentary demands for extra plate current.

AMPLIFICATION. A diode does only one thing which is useful in television and radio receivers. This is to provide one-way conductivity. Although a table in this lesson lists many nses for diodes, the action in every case depends solely on the one-way conductivity of these tubes. Of course, all these uses for diodes are important, for otherwise they wouldn't be included in receivers. But of greater importance is something a diode cannot do, it cannot amplify a signal. We have learned that in all receivers there are more amplifiers than any other one kind of tube. The biggest single job always is that of amplification, of increasing the strength of the exceedingly weak signals coming to the antenna.

In order that a tube may amplify a signal, the tube must have a control grid. No matter what other kinds of elements might be placed between the cathode and plate, the lack of a control grid would prohibit amplification. In the following lesson we shall see how a control grid makes amplification possible.

TELEVISION

LESSON NO. 10

THE BEGINNING OF ELECTRONIC TUBES

"Training Your Mind"

C.T.I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON No. 10

AMPLIFYING A VOLTAGE

The three meters of Fig. 10-1 will show us how a triode tube amplifies a signal voltage. You are looking directly at the top of the tube, which is mounted at the center of the small light-colored block below two of the meters. Around the tube are terminal screws connected to the socket and arranged in the same order as the base pins and socket connections.

Up above are the three meters. The one at the left measures grid voltage, which is the potential difference between grid and cathode. The meter at the center measures plate current in milliamperes. At the right is a separate testing instrument which, in this picture, is connected by its flexible leads to measure plate voltage. Plate voltage is the potential difference between plate and cathode of the tube.

Our testing circuit is shown in detail by the connection diagram at the top of Fig. 10-2 and by the simplified schematic diagram at the bottom. The many connections and screw terminals as shown by the diagram and photograph seem more intricate than necessary for this simple hookup, but they will be needed before we are through.

Concealed behind the panel is a power supply unit of the type used in television and radio receivers. The positive side of the power supply section that furnishes plate voltage and current is connected to the panel screw marked B+, and the negative side of this section is connected to the screw marked B-. The section of the power supply that furnishes heater current is connected to the two heater terminals on the tube block.

Broken lines on Fig. 10-2 show that the grid voltage meter is connected behind the panel to two terminal screws marked METER. On the front of the panel the left-hand screw terminal is connected through a wire to the tube grid. The right-hand METER screw is connected on the front of the panel to the cathode of the tube. When later we produce potential differences between grid and cathode of the tube, they will be indicated by this meter.

As shown also by broken lines in Fig. 10-2, the positive side of the plate current meter is connected behind the panel to one of another pair of terminal screws marked METER and, on the front of the panel, to the screw marked B+ and thereby to the power supply. The negative side of this plate current meter is connected behind the panel to the second METER terminal screw and, through front connections, to the tube plate. Therefore, all current flowing in the plate circuit of the tube must pass through and be measured by this meter.

The separate testing instrument is a service type volt-ohmmeter, which will measure either volts of potential difference or ohms of resistance. Throughout this lesson the instrument will be used as a voltmeter. In Figs. 10-1 and 10-2 its positive lead is clipped to a terminal screw from which a wire goes directly to

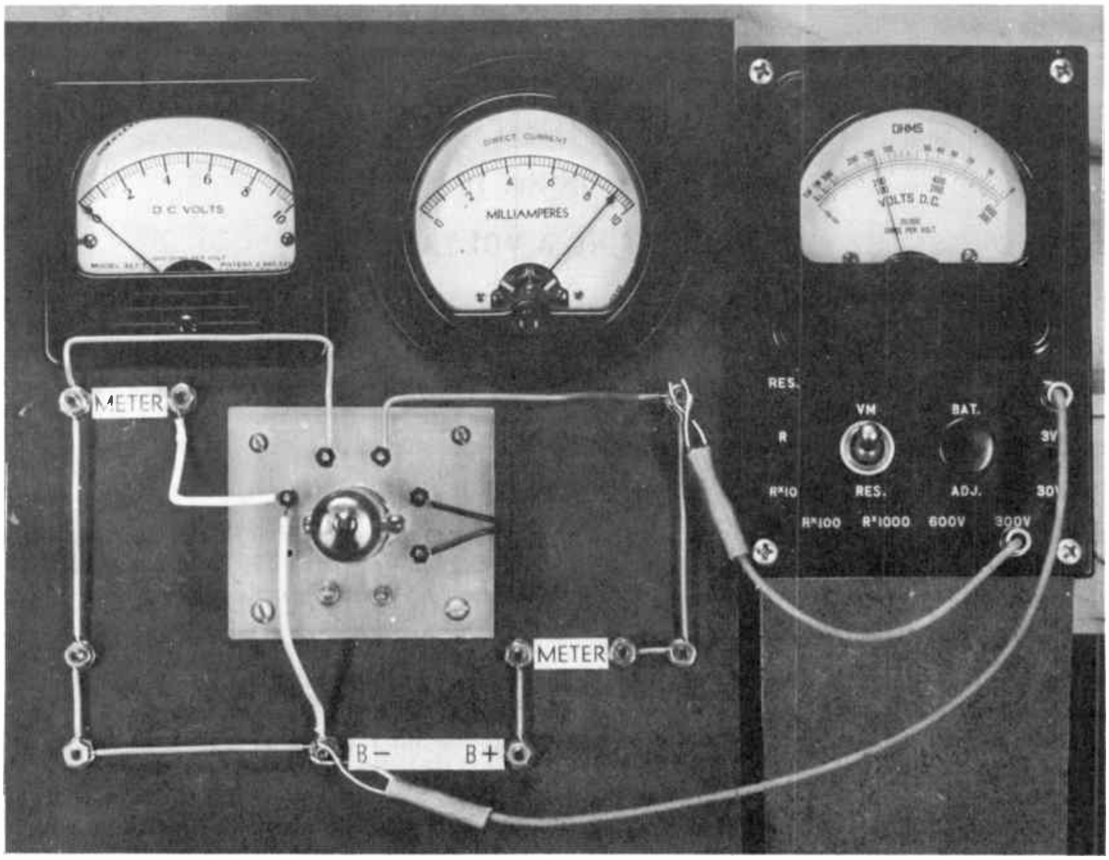


Fig. 10-1. The test setup which will show how a tube amplifies a signal voltage.

the tube plate. The negative meter lead is clipped to the B- terminal screw, and thence through a wire connection to the tube cathode. Therefore, this meter will indicate potential difference between plate and cathode, which is plate voltage.

The dial figures and graduations on the meters for grid voltage and for plate current are large enough to be read easily on a photograph. But the volt-ohmmeter scales have smaller graduations because there are several scales on the one dial. Fig. 10-3 is an enlarged view of the volt-ohmmeter scales. For our present purposes we shall measure nothing greater than 300 volts, with the connecting leads in appropriate jack terminals of the instrument. Then all our readings will be on the bottom row of numbers, which are 0-100-200-300 volts. In Fig. 10-1 you can see that plate voltage as measured on this scale of numbers is about 100.

All of our diagrams show that the grid of the tube is connected through wires on the front of the panel to the cathode of the tube. With this connection, of practically zero resistance, the potential of the grid must be the same as of the cathode. There can be no potential difference between them.

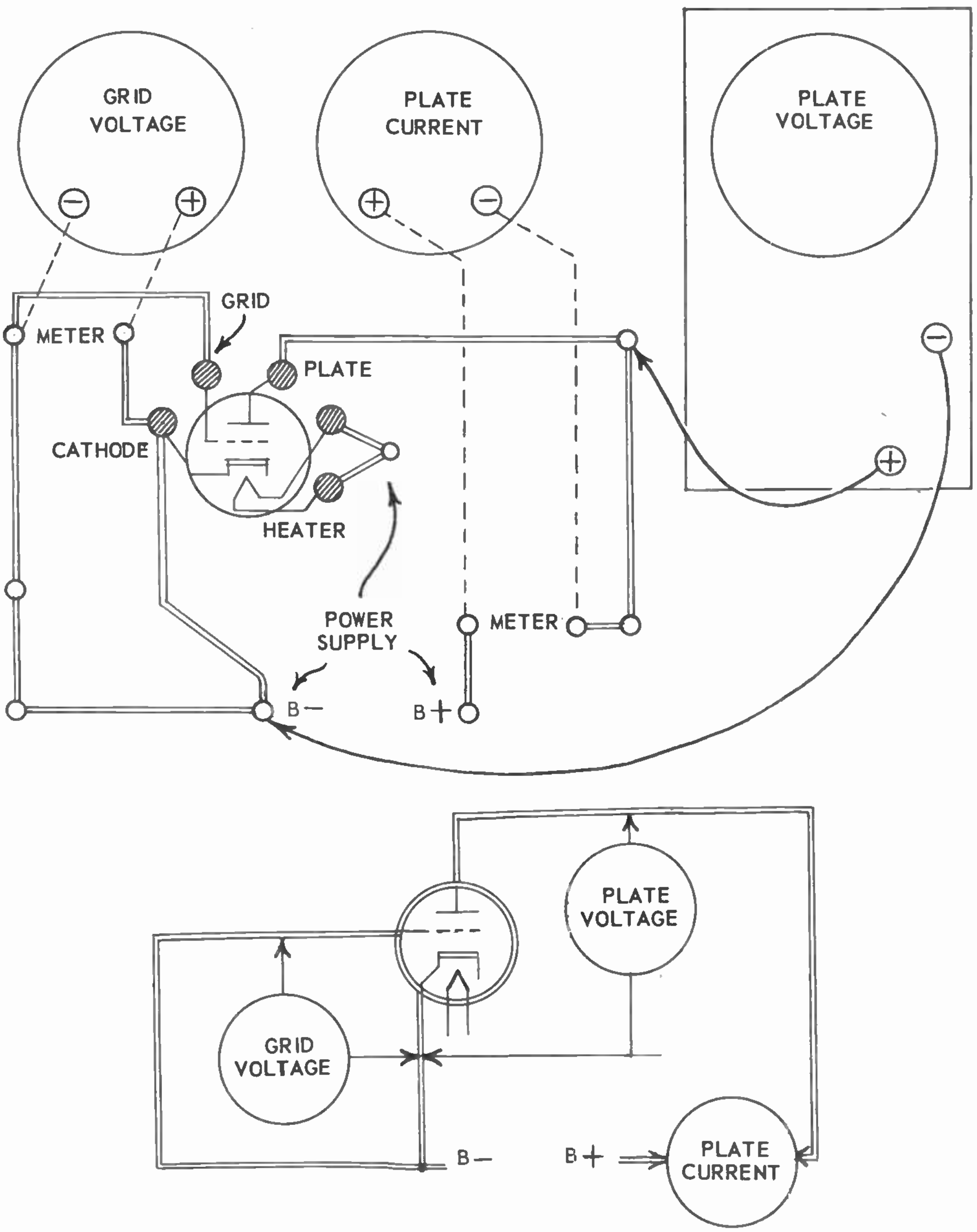


Fig. 10-2. Meter connections and other details of the testing circuit.

Potential of the cathode always is considered to be the reference potential for voltages of all other tube elements; it is considered to be zero. Since we now have no potential difference between grid and cathode, the grid is at zero voltage. Usually we would speak of this condition as "zero grid".

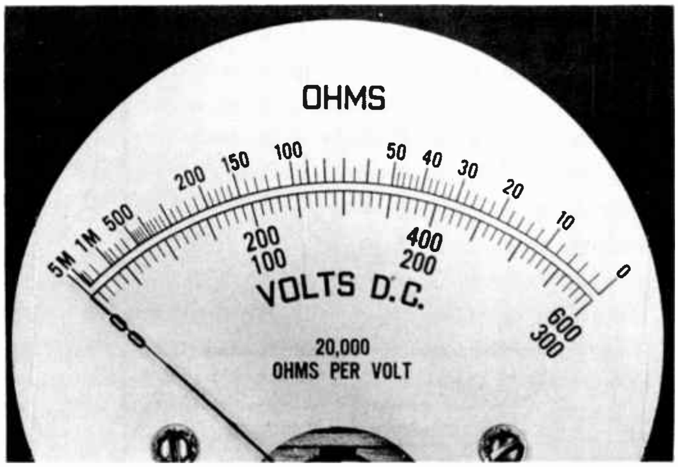


Fig. 10-3. The dial scale markings of the volt-ohmmeter.

The plate is approximately 100 volts positive with reference to the cathode. It is this positive plate potential that pulls electrons from the space charge around the cathode, through the grid, and into the plate. The electrons come from B- of the power supply to the cathode. From the plate they return to B+ of the power supply through the plate current meter. This meter, in Fig. 10-1, shows that the electron flow rate or plate current is almost exactly 9.0 milliamperes.

From the simplified diagram at the bottom of Fig. 10-2 it is apparent that some electrons from the plate could get back to the cathode through the plate voltmeter. Only a negligible quantity of electrons actually take this path, because resistance of voltmeters is much greater than of current meters. The resistance of our plate voltmeter is 6,000,000 ohms (6 megohms) and resistance of the plate current meter is only about 7 ohms.

PLATE CHARACTERISTICS. For our first tests we shall keep the grid at zero with reference to the cathode and determine the effect on plate current of varying the plate voltage. In Fig. 10-1 the plate voltage is 100 and plate current is 9.0 ma. In Fig. 10-4 the plate voltage has been reduced to 70, and plate current has dropped to 5.0 ma. In Fig. 10-5 the plate voltage has been further reduced to 40, and plate current has gone down to 2.0 ma.

Similar measurements may be made with many different plate voltages. Results of a series of actual measurements were as follows.

Volts	Ma	Volts	Ma	Volts	Ma
100	9.00	60	4.00	20	0.80
90	7.60	50	3.00	10	0.30
80	6.35	40	2.15	0	0
70	5.10	30	1.40		

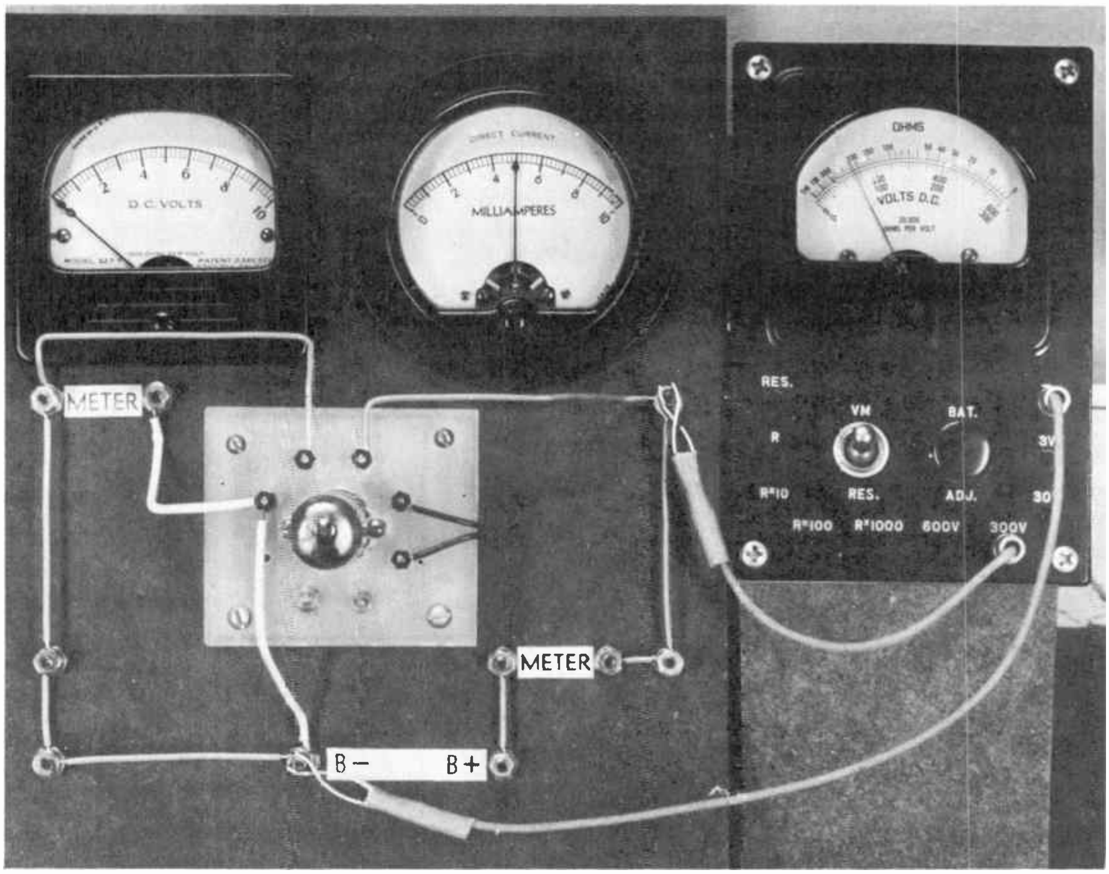


Fig. 10-4. Reducing the plate voltage reduces the plate current.

On a sheet of ruled "graph" paper we may draw a curve showing these relations between plate current and plate voltage, as in Fig. 10-6. The vertical scale at the left is for milliamperes of plate current. The bottom horizontal scale is for plate volts. All the corresponding pairs of values for volts and milliamperes are marked on the graph, and through the marks is drawn a smoothly sloping curve. The curve is drawn with a smooth slope, and if this causes the curve to miss some of the marks it indicates small errors in measurements.

A curve of this kind is drawn with a full line in Fig. 10-6. Engineers would call it a *plate characteristic*. This one characteristic curve applies only when the grid is zero, or at the same potential as the cathode.

EFFECTS OF GRID VOLTAGE. While it is necessary that we understand how a tube behaves with its grid zero, tubes seldom are so operated. The reason will appear when we apply a signal voltage to the tube being tested. In almost every application the grid remains negative with reference to the cathode.

In Fig. 10-7 the grid has been made about 1.5 volts negative by inserting a dry cell between cathode and grid, where formerly there was only a length of wire. The negative terminal of any dry cell in its outer metallic "can" and the positive terminal is a small insulated disc or button in the center of one end of the cell. The dry cell is here mounted so that its negative terminal is connected through wires to the grid of the tube, and its positive terminal to the cathode.

A dry cell of any size furnishes approximately 1.5 volts of potential difference; actually about 1.56 volts when furnishing little or no current. This value of negative grid voltage is indicated by the left-hand grid voltage meter.

The plate voltage meter in Fig. 10-7 shows that the plate of the tube is 100 volts positive with reference to the cathode. But plate current now is only 5.1 ma with the grid 1.5 volts negative, whereas with zero grid and 100 volts on the plate we had plate current of about 9.0 ma.

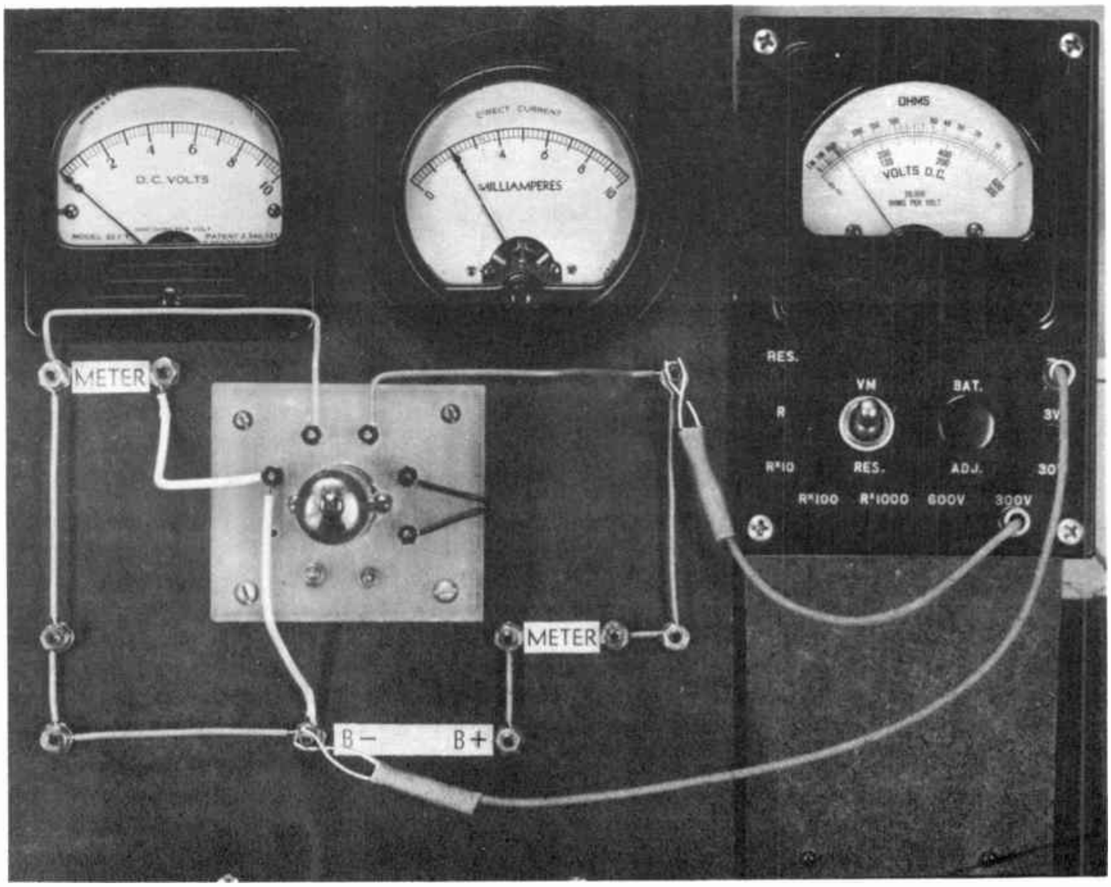


Fig. 10-5. The less we make the plate voltage, the smaller becomes the plate current.

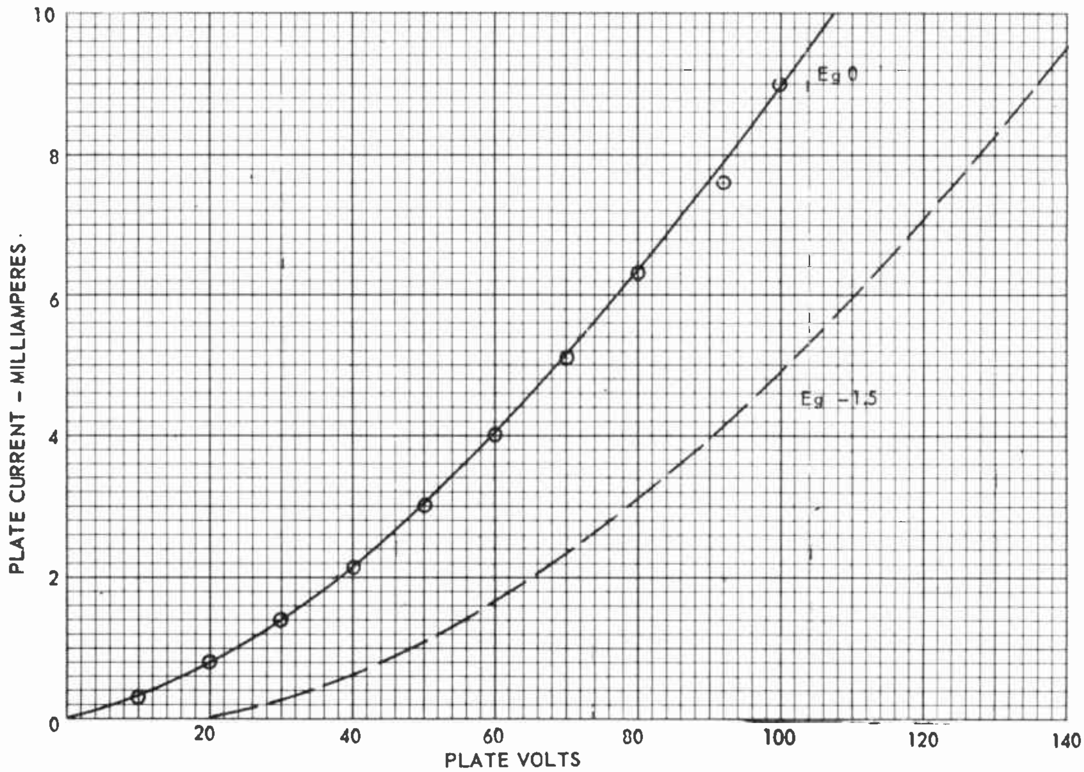


Fig. 10-6. Plate characteristics for the grid at zero and at 1.5 volts negative.

Plate current has been reduced because the negative potential or charge on the grid is counteracting part of the positive potential or charge on the plate. It is the positive charge on the plate that enables the plate to pull to it the negative electrons from the space charge and cathode. The grid is so much closer than the plate to the cathode and space charge that a relatively small negative voltage on the grid will counteract in large measure a much greater positive voltage on the plate.

Now, while keeping the grid at 1.5 volts negative, we again might measure the plate current with various values of plate voltage, and make a second graph curve showing what happens with this grid voltage. Such a curve is drawn with a broken line on Fig. 10-6.

With the grid 1.5 volts negative we can obtain a plate current of 9.0 ma only with an increased plate voltage. This has been done in Fig. 10-8, where the plate voltage meter reads 135 volts instead of the 100 volts which caused this same current with a zero grid.

Fig. 10-6 shows that for any given plate current we require higher plate voltage when the grid is negative than when it is zero. Also, with any particular plate voltage, the plate current is less when the grid is negative than when it is zero.

Down at the lower end of the broken-line curve of Fig. 10-6 we run into a rather peculiar condition, which is shown also by the meters in Fig. 10-9. Plate current has dropped to zero. But still there is 20 volts positive on the plate. This condition is called *plate current cutoff.* For any positive voltage which may be applied to a plate of a tube there will be some negative grid voltage which will reduce the plate current to zero or to cutoff. Our ability to thus cut off the plate current by means of a grid voltage sufficiently negative will prove to be of great usefulness in many television circuits.

Were we to make a complete checkup on the tube being tested it would be in order to change the negative grid voltage in steps, and for each grid voltage draw a plate characteristic curve. Then we should have what is called a *family* of plate characteristics.

Fig. 10-10 shows a family of plate characteristics for a 6C4 triode tube. The left-hand curve shows relations between plate milliamperes and volts when the grid is zero. Following curves are drawn with the grid more and more negative, in steps of 2 volts, until reaching 30 volts negative. At the bottom of each curve,

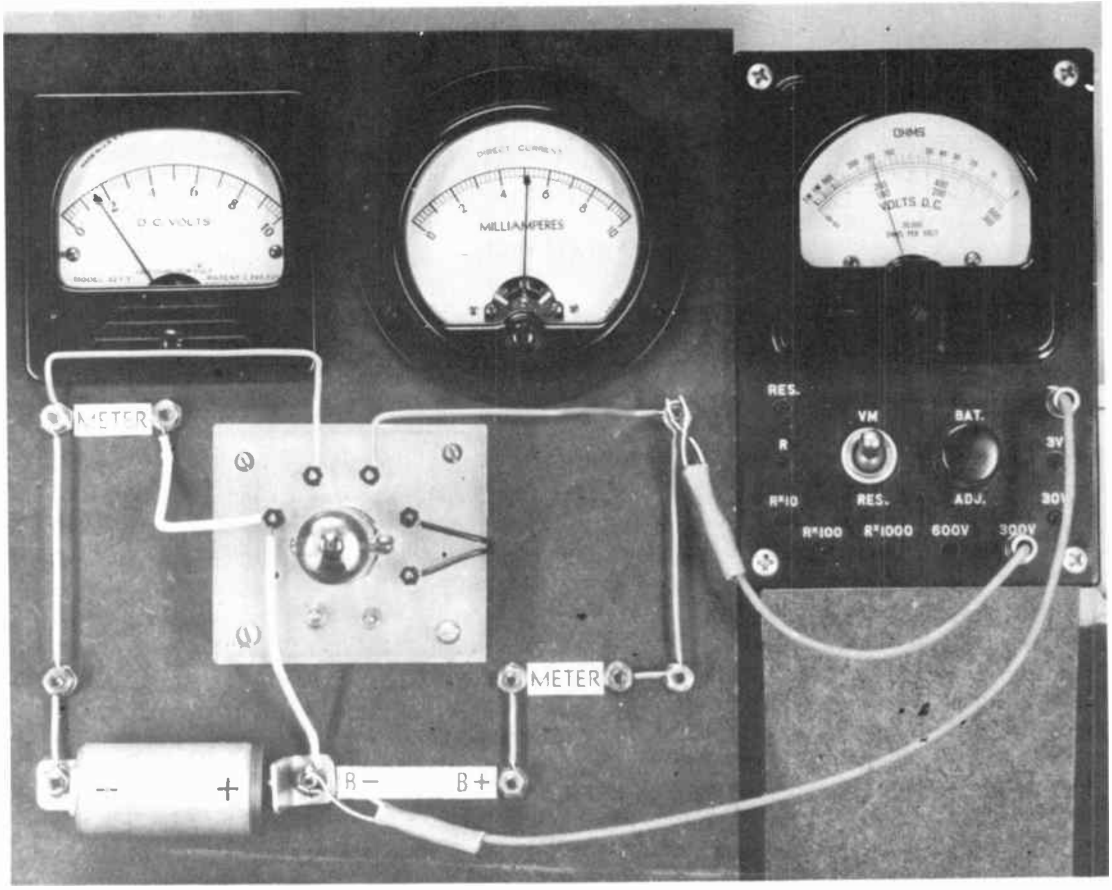


Fig. 10-7. When the grid is made negative there is a decrease of plate current.

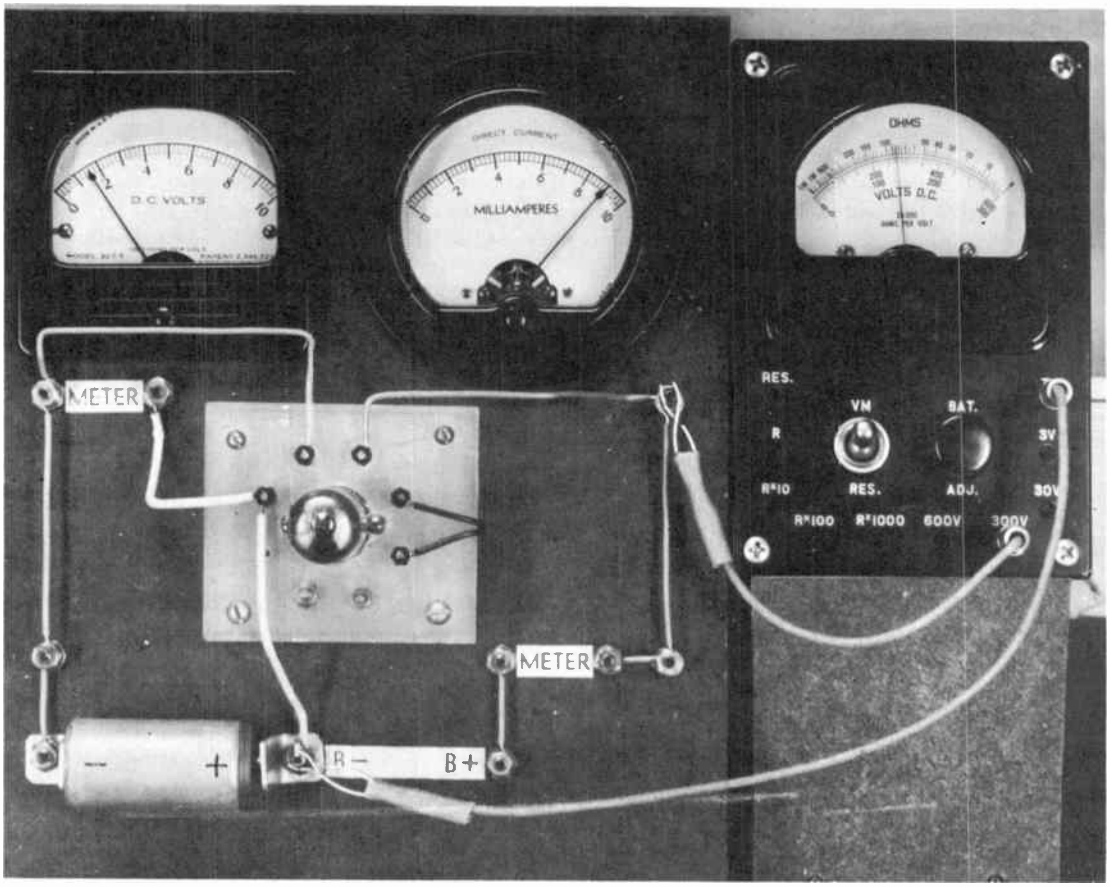


Fig. 10-8. To increase the plate current while the grid is negative we must raise the plate voltage.

where it meets the line for zero plate current, you may read the plate voltages at which there is cutoff for each of the grid voltages. For instance, on the curve for a 10-volt negative grid there is cutoff when the plate still is 140 volts positive.

GRID BIAS. The grid voltage which is furnished by the dry cell added to our test setup in Fig. 10-7 is called a grid bias voltage or simply a *grid bias*. Any voltage applied between grid and cathode in such manner as tends to keep the grid at some certain potential with reference to the cathode is a bias voltage.

The bias voltage may be of any value that suits our purposes in various applications of tubes. The bias voltage need not be furnished by a dry cell or a battery of dry cells. Bias may be obtained from the power supply or in other ways which will be described in later lessons.

GRID VOLTAGE AND PLATE CURRENT. In all signals which we may wish to amplify, the voltages are alternating. Signal voltages act first in one direction or polarity, then in the opposite direction, and continue to alternate. These alternating signal voltages come originally from the antenna, but farther along in the receiver they may come to an amplifier tube from any preceding part or circuit. So far as a given tube is concerned, any part from which the signal comes is the source of signal voltage. This voltage is applied between grid and cathode of the tube, as at *1* in Fig. 10-11.

Now let's see how an alternating signal voltage looks to the grid of the amplifier tube. To begin with we shall assume that there is no signal, or that signal voltage is zero. This may be represented by the straight horizontal line at *2* in Fig. 10-11. Were there no bias voltage, and were this zero signal voltage to be applied between grid and cathode, there would be no potential difference between grid and cathode. We would have the condition of zero grid at this instant.

During a following instant the signal voltage would swing in such direction or polarity as to make the grid positive with reference to the cathode when there is no biasing voltage. This we represent by an upward curve as at *3*. Signal voltage increases from zero to its maximum positive value, then drops back to

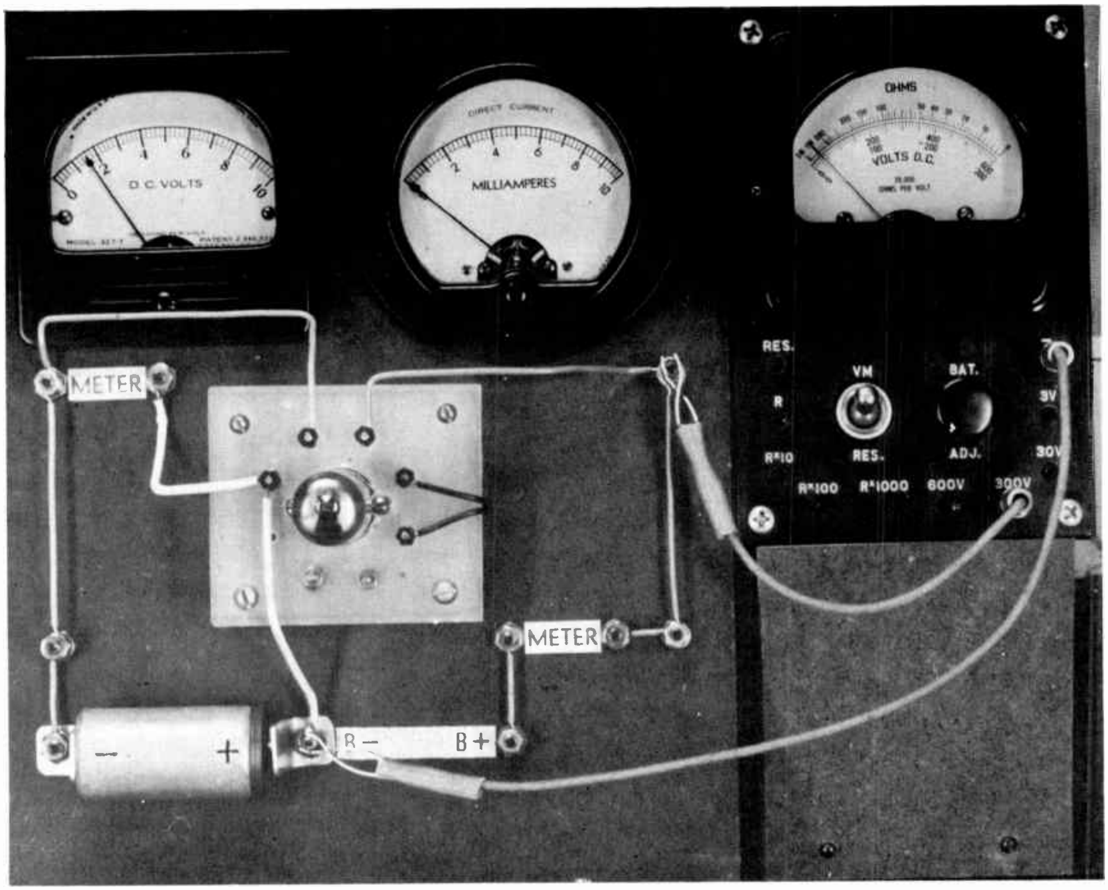


Fig. 10-9. The negative grid causes plate current cutoff, even though a positive voltage is applied to the plate.

zero. Then the signal voltage reverses its polarity, as at *4*, and makes the grid negative with reference to the cathode. Here the signal goes from zero to its maximum negative value and back to zero.

The continually recurring reversals or alternations of signal voltage actually would appear as at 5. There are no pauses at zero. The voltage goes continually from positive to negative and back again, passing through its zero value in between the two maximum values.

A real signal reverses its polarity at the rate of hundreds, thousands, or millions of times every second. We could not see such rapid changes or the resulting currents on our meters. Consequently, we must reduce this rapid rate of alternation to extremely slow motion, or to a standstill at maximum positive and negative values

This has been done in Fig. 10-12. To represent the signal voltage we use a second dry cell mounted vertically on the left-hand side of the panel. The first dry cell, which furnishes 1.5 volts of negative grid bias, has not been disturbed. Our signal-voltage dry cell is mounted with its positive terminal toward the top,

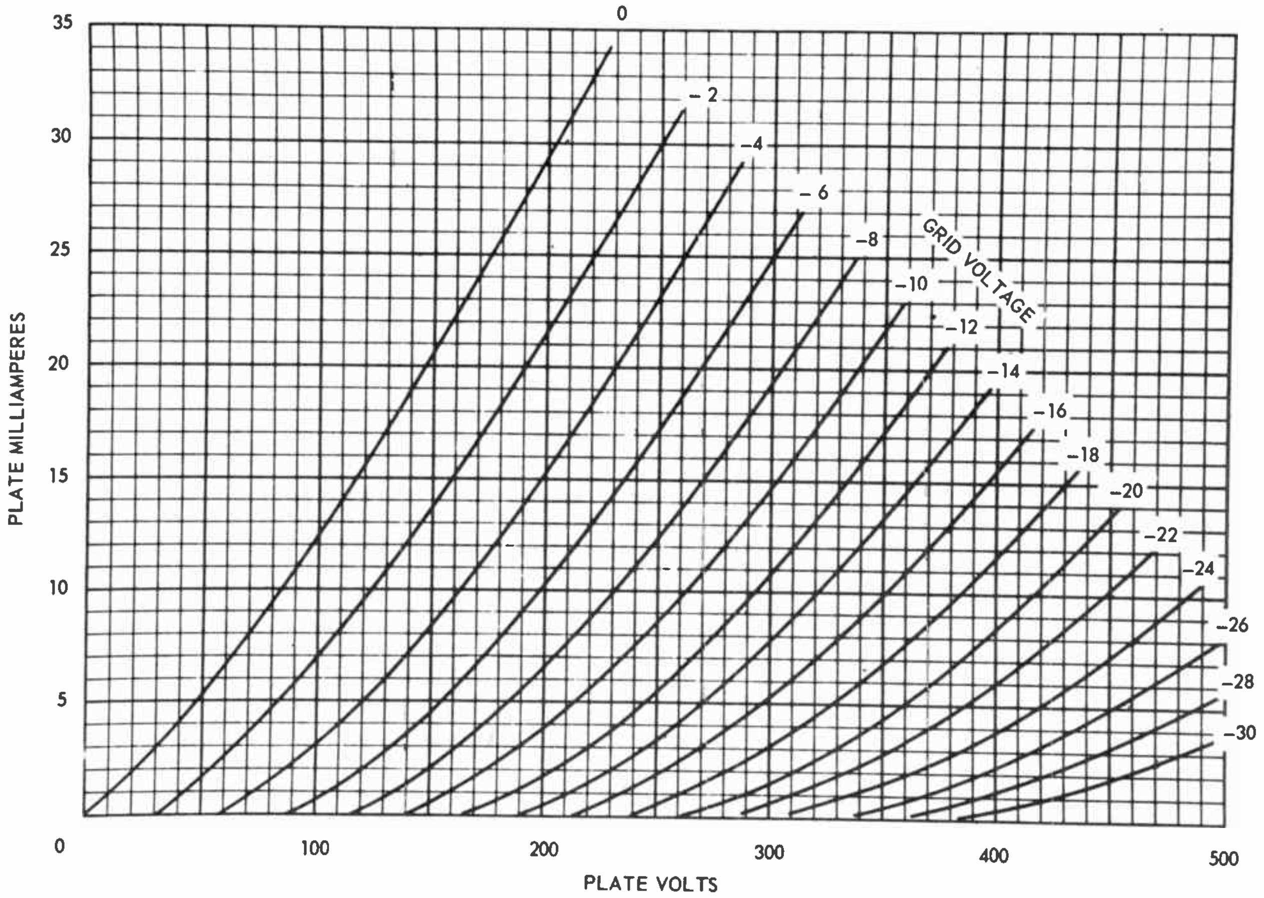


Fig. 10-10. A family of plate characteristics for a 6C4 triode tube.

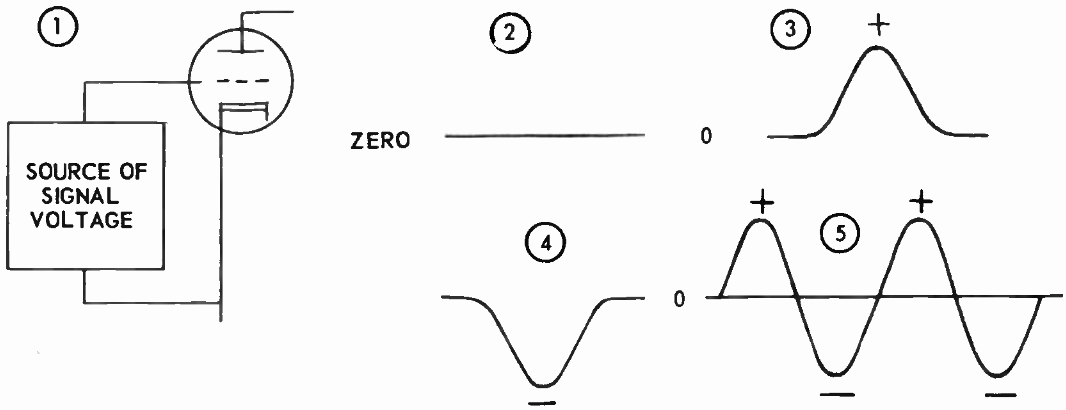


Fig. 10-11. The alternating signal voltage applied between grid and cathode.

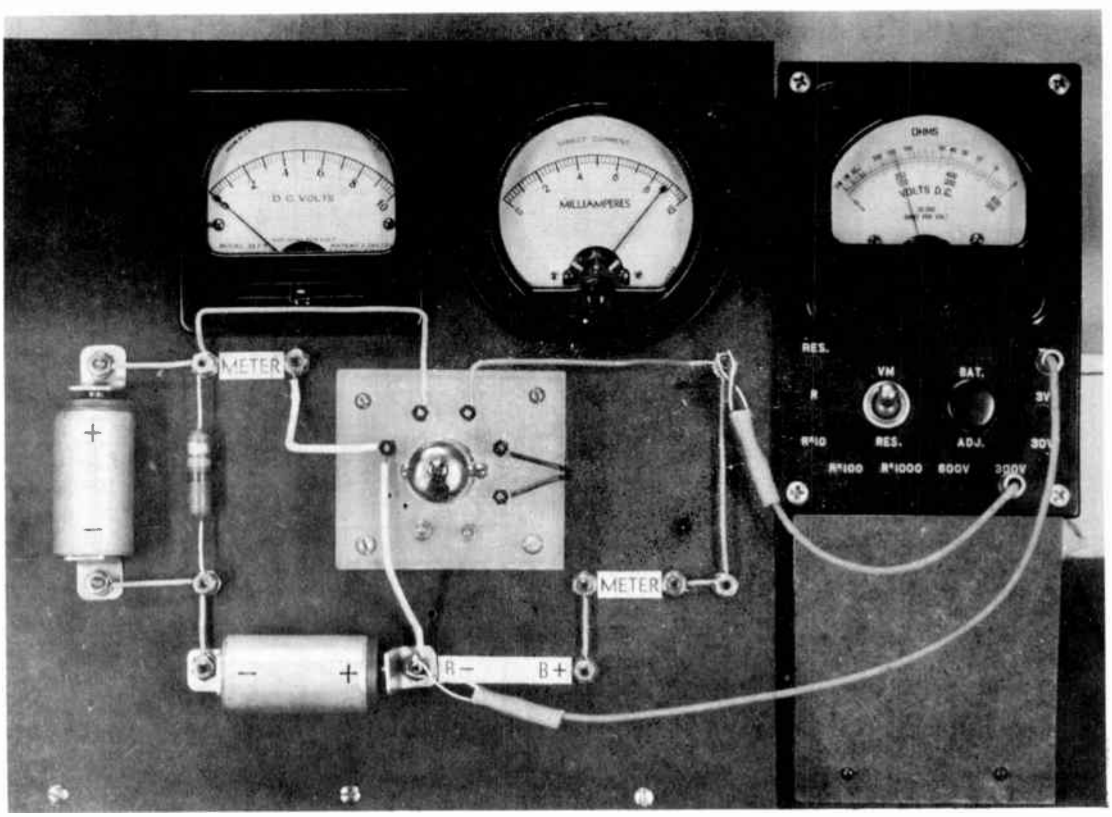


Fig. 10-12. The signal and biasing voltages combine to make the grid zero.

from where it connects to the grid of the tube. The negative side of this cell connects through the biasing dry cell to the cathode of the tube. This gives the effect of diagram *3* in Fig. 11, where the signal tends to make the grid positive with reference to the cathode. The signal does not succeed in making the grid positive, because we have a negative grid bias.

The straight wire which formerly ran from the negative side of the biasing dry cell up to one of the meter terminals and thence to the grid of the tube has been replaced with a resistor. The signal-voltage dry cell is connected across this resistor, and thus is connected at the top to the grid of the tube, and at the bottom is connected through the biasing cell to the cathode.

Had the wire been left in place it would have short circuited the signal voltage. That is, there could have been no potential differences across the negligible resistance of the wire, and no signal voltages. Incidentally, the dry cell which is acting as signal voltage would have been almost instantly discharged and rendered useless by the short circuit. Signal voltage is applied across the inserted resistor. At the same time this resistor maintains a path through which bias voltage would reach the grid whether or not there were a signal circuit.

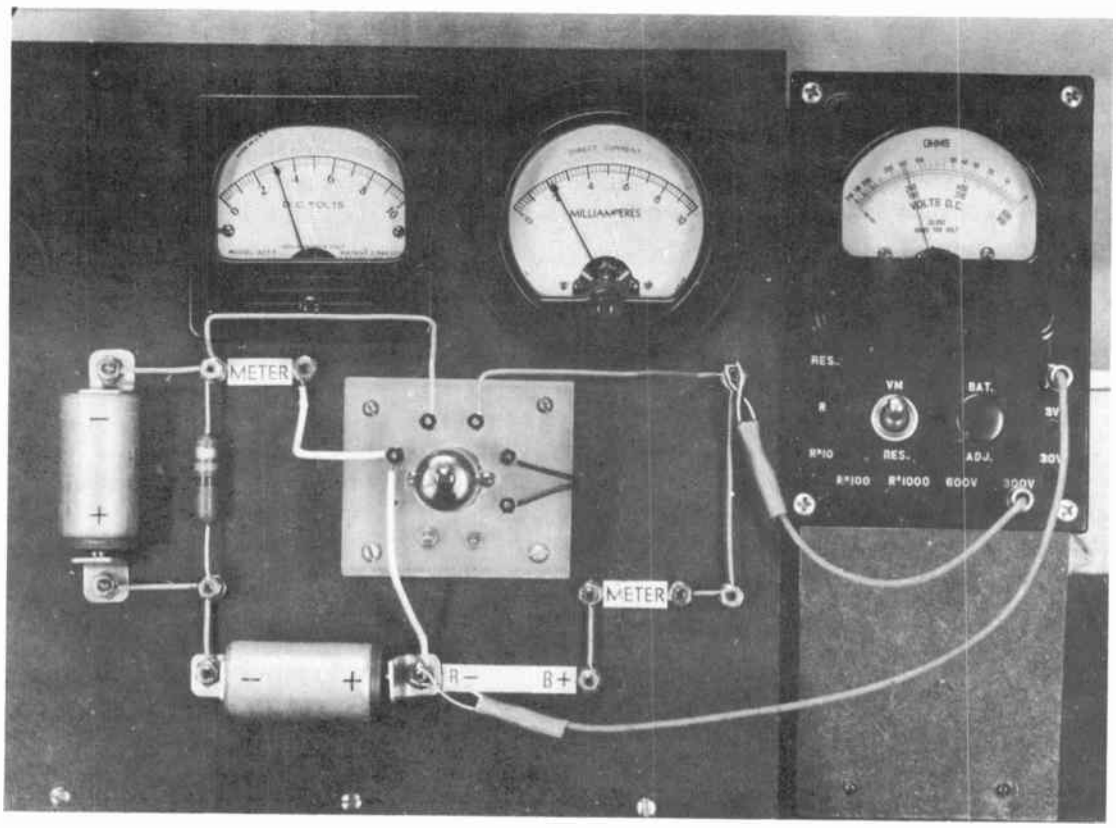


Fig. 10-13. Here the signal and bias combine to make the grid negative.

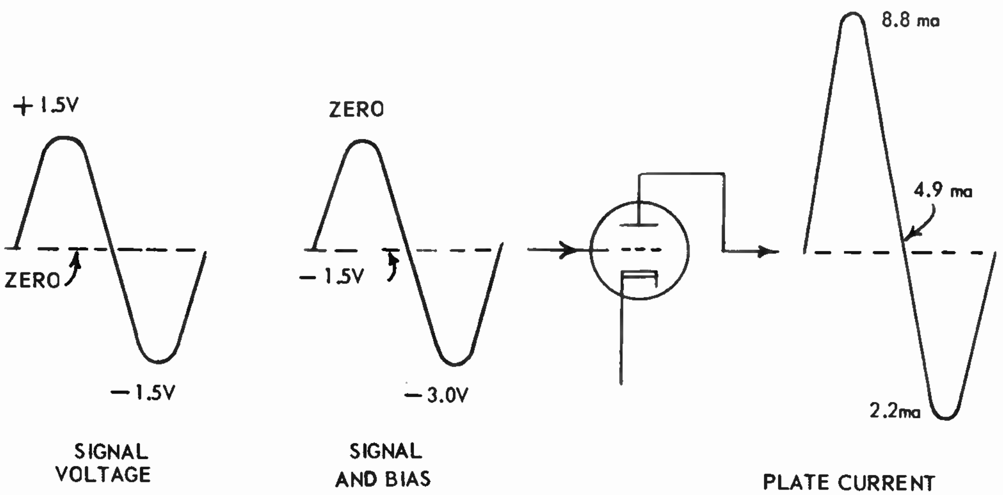


Fig. 10-14. The alternating signal voltage, the combined signal and bias voltages on the grid, and the resulting plate current.

Now let's see what the meters tell us in Fig. 10-12. We note first that the grid voltage is zero. This is because there are, between grid and cathode, two voltages which are equal and which are opposing each other. The signal voltage is trying to make the grid 1.5 volts positive, and the bias voltage is trying to make the grid 1.5 volts negative. Each counteracts the other, and the net difference of potential between grid and cathode is zero.

Plate current is almost 9.0 ma and plate voltage is 100. Back in Fig. 10-1 we had the same plate current with 100 volts on the plate and a zero grid. No matter how any certain grid voltage is obtained, it will allow equal plate currents when there is some given plate voltage.

NOTE: Although plate current reads 8.8 ma in Fig. 10-12 and 9.0 ma in Fig. 10-1 we are justified in considering these currents as alike. They should have been alike, because in both cases we have the same plate voltage and the same grid voltage. The greater plate current in Fig. 10-1 may have resulted from longer operation of the tube before taking the photograph, for there is maximum electron emission only after a cathode reaches its final high temperature. Voltage from the building power line might have changed between times of the two photographs, resulting in a change of tube heater voltage and current. This would not have affected the plate voltage, which was adjusted for each test, while no compensations were made for changes of heater voltage.

You may notice small discrepancies between currents or voltages in other pictures. Similar variations will occur when making receiver tests during service operations. Because we are studying general principles, not precision measurements, these slight differences should be ignored.

For our next test we shall reproduce the effect of a signal voltage such as at 4 in Fig. 10-11, where the polarity of the signal would tend to make the grid negative. It is necessary only to turn the signal-voltage dry cell upside down, as in Fig. 10-13, with the negative side of the signal dry cell toward the grid. Now the signal voltage, by itself, would make the grid 1.5 volts negative. The biasing voltage also, by itself, would make the grid 1.5 volts negative. The two voltages are acting together, in the same polarity. Their effects add together, and make the grid 3.0 volts negative as shown by the grid voltage meter.

Plate voltage still is 100. But plate current has dropped to about 2.2 ma. Between Figs. 10-12 and 10-13 the plate current has dropped from 8.8 to 2.2 ma, which is a change of 6.6 ma. This has resulted from varying the grid voltage from zero to 3.0 volts negative. Plate voltage has remained constant.

What has happened is illustrated by Fig. 10-14. At the left is the signal voltage, swinging to 1.5 volts positive, then through zero to 1.5 volts negative and back to zero. At the same time there has been a biasing voltage which constantly makes the grid 1.5 volts *more negative* than any other voltage simultaneously applied between grid and cathode. The signal and bias voltages combine, as at the center, to swing the grid to zero, then through 1.5 volts negative to 3.0 volts negative and back to zero.

The result of all this, shown at the right, has been to increase the plate current to 8.8 ma, then decrease this current through 4.9 ma to 2.2 ma, and let it come back to 4.9 ma. We know that plate current reaches

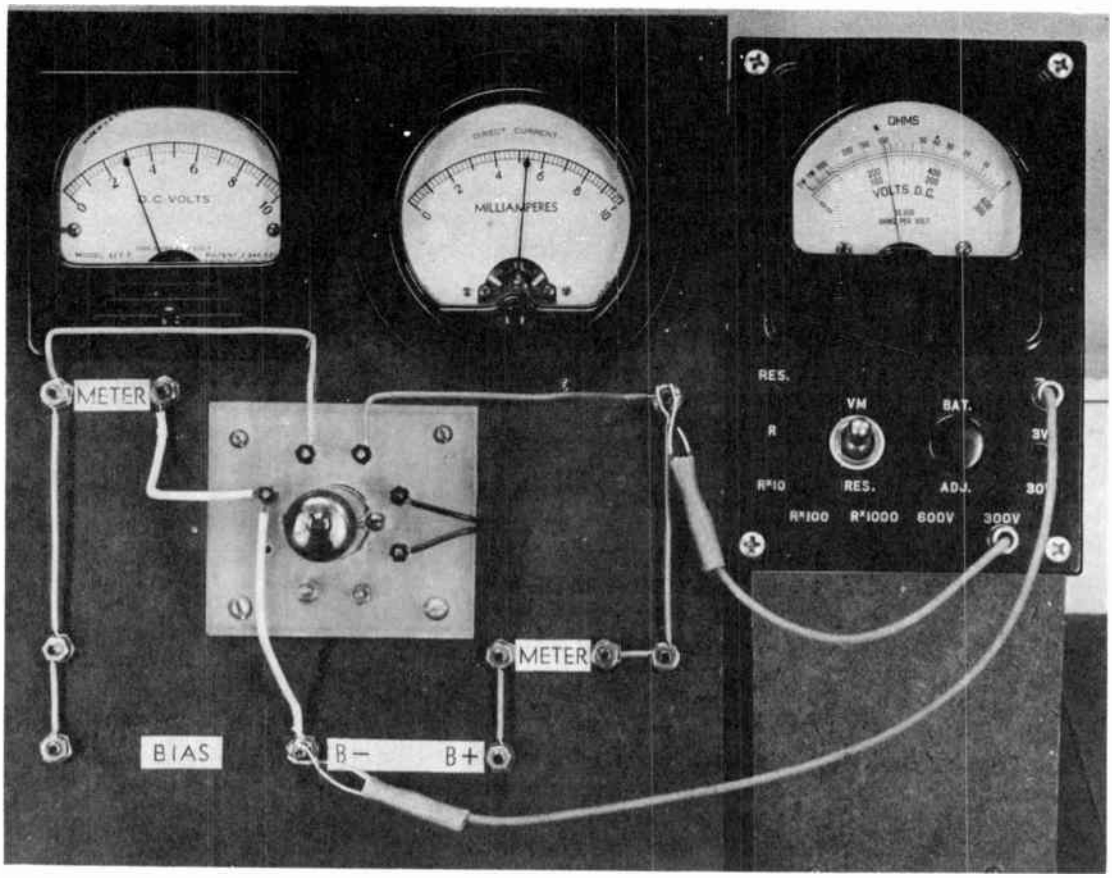


Fig. 10-15. One step during measurements for a mutual characteristic with 125 plate volts.

8.8 ma when the grid is zero by looking at Fig. 10-12, and that current drops to 2.2 ma when the grid is 3.0 volts negative by looking at Fig. 10-13. We know that plate current should be 4.9 ma with the grid 1.5 volts negative and 100 volts on the plate by looking at the broken line curve of Fig. 10-6.

TRANSCONDUCTANCE. We have measured the effect on plate current of changes of grid voltage when the plate voltage remains constant. Fig. 10-14 shows that a *change* of 1.5 volts on the grid, from zero to 1.5 volts negative, causes the plate current to *change* by 3.9 ma, from 8.8 ma down to 4.9 ma. What is the change of plate current *per volt* change of grid voltage? We divide 3.9 (plate ma) by 1.5 (grid volts) to find that there is a change of about 2.6 ma per volt. To avoid using this decimal fraction we may measure the plate current change in microamperes. One milliampere equals 1000 microamperes, so 2.6 ma is the same as 2600 microamperes.

The change of plate current per one-volt change of grid voltage is called *grid-plate transconductance*, or more often is called simply *transconductance*. When the tube is a triode we may call this quantity *mutual conductance*. In the case of a triode the terms mutual conductance and transconductance mean the same thing. When talking about pentodes and beam power tubes we always use the term transconductance.

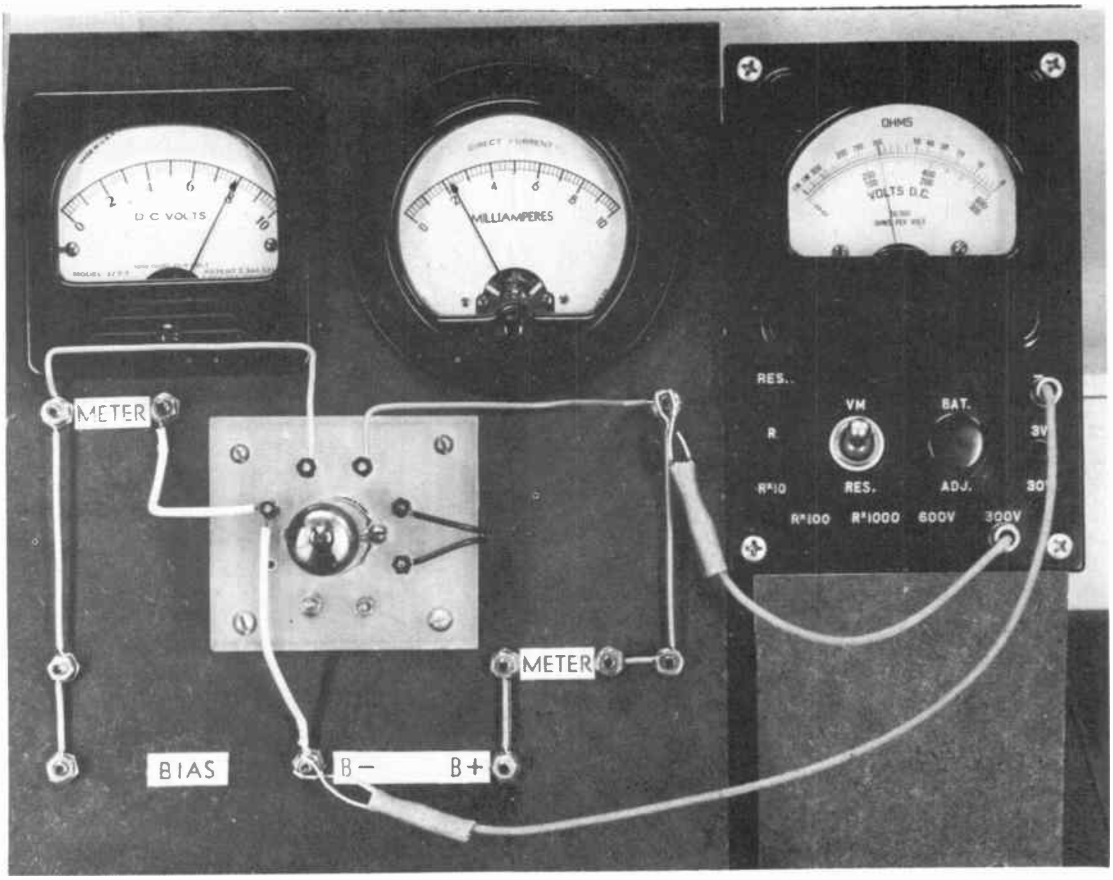


Fig. 10-16. Another step during determination of the mutual characteristic.

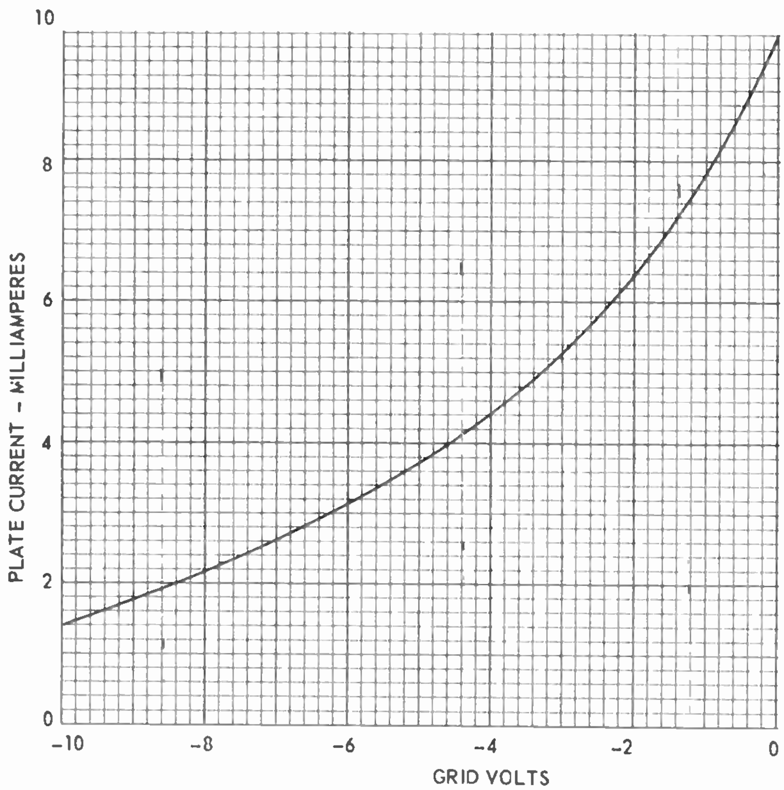


Fig. 10-17. The mutual characteristic for a plate voltage of 125.

Now let's figure out the transconductance when the grid voltage changes from 1.5 volts to 3.0 volts negative in Fig. 10-14. This is another *change* of 1.5 grid volts. The plate current drops from 4.9 ma to 2.2 ma, which is a *change* of 2.7 ma. Dividing the plate current change (2.7 ma) by the grid volts change (1.5) shows that current is varied by 1.8 ma or by 1800 microamperes per volt. Then the transconductance in this case is 1800. There is a big difference between transconductances with the grid working up toward zero and with it working down toward 3.0 volts negative.

Transconductance is not ordinarily computed from changes of grid voltage so great as 1.5 volts, but from a change of a small fraction of one volt near the average grid voltage. This average is the biasing voltage. In lists or tables of tube performance published by manufacturers you will find values of transconductance or mutual conductance as obtained with certain grid voltages and plate voltages. The values will be given as so many *micromhos*, a name which means the same as microamperes per volt.

MUTUAL CHARACTERISTICS. Now we shall measure the effect on plate current of varying the grid voltage while plate voltage remains at a constant or fixed value. One of the tests is pictured by Fig. 10-15. The biasing dry cell has been removed. The two screw terminals on either side of the word BIAS are connected behind the panel to a section of the power supply capable of furnishing any negative biasing voltage from zero to 10 volts.

Plate voltage is 125, and will be kept at this value. Measurements would commence with the grid zero. Then the grid would be made more and more negative in small steps, while keeping the plate voltage constant. Fig. 10-15 shows what happens when the grid has been made about 3.0 volts negative. Plate current is 5.4 ma.

We would make notes of plate currents obtained at each step of grid voltage variation, always with the plate at 125 volts or at whatever plate voltage has been selected for the series of measurements. Fig. 10-16 shows what happens when the grid is made about 8.0 volts negative. Plate current is down to 2.2 ma, while plate voltage still is 125.

After completing a series of measurements with some one plate voltage we should "plot" the results in the form of a graph curve such as the one of Fig. 10-17. This curve shows relations between plate current and grid voltage with the grid varied from zero to 10 volts negative when plate voltage is 125. It would be in order to make curves for several of the plate voltages in the range which normally would be used for the particular tube being tested. Each such curve is called a *mutual* characteristic. Tube manufacturers publish average mutual characteristics for many of the more commonly used tubes.

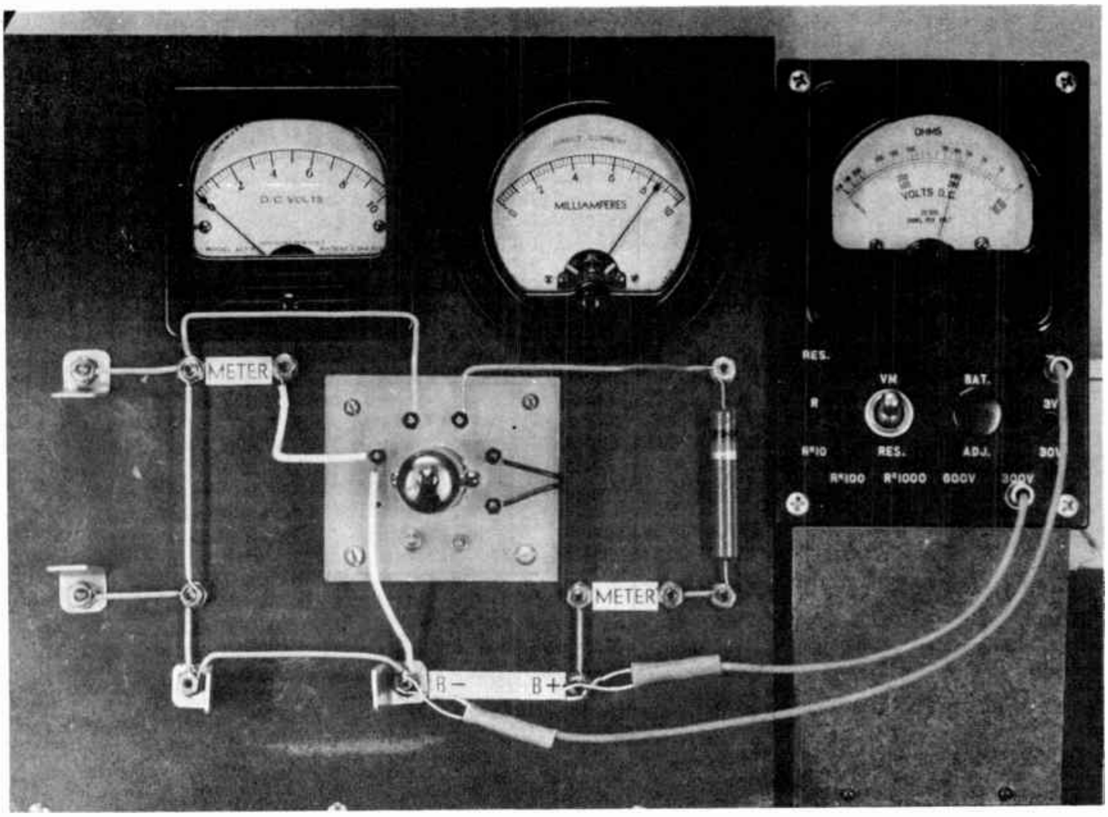


Fig. 10-18. A resistor has been inserted in the plate circuit of the tube.

VOLTAGE AMPLIFICATION. Up to this point we have learned how a varying signal voltage on the grid will cause variations or alternations of plate current. The plate current alternates just as does the grid voltage, and at the same instants. This plate current could be called a signal current. But we are not trying to produce a signal current, what we are after is a signal voltage which is like the grid voltage but much stronger.

The practical way to obtain a signal voltage in the plate circuit is to insert a resistor in this circuit, and let the varying plate current flow in this resistor. Then, whenever there is a change of plate current there will be a corresponding change of voltage between one end and the other of the resistor. From the ends of such a resistor we can take a varying voltage which is like the varying voltage on the grid.

Fig. 10-18 shows how a resistor is put into the plate circuit. This resistor is at the right-hand side of the main panel, connected between two terminal screws where formerly we had a piece of wire. Electron flow in the plate circuit is from the B- side of the power supply to the tube cathode, through the tube from cathode to plate, from the plate to the top of the new resistor, through this resistor and the plate current meter back to the B+ side of the power supply.

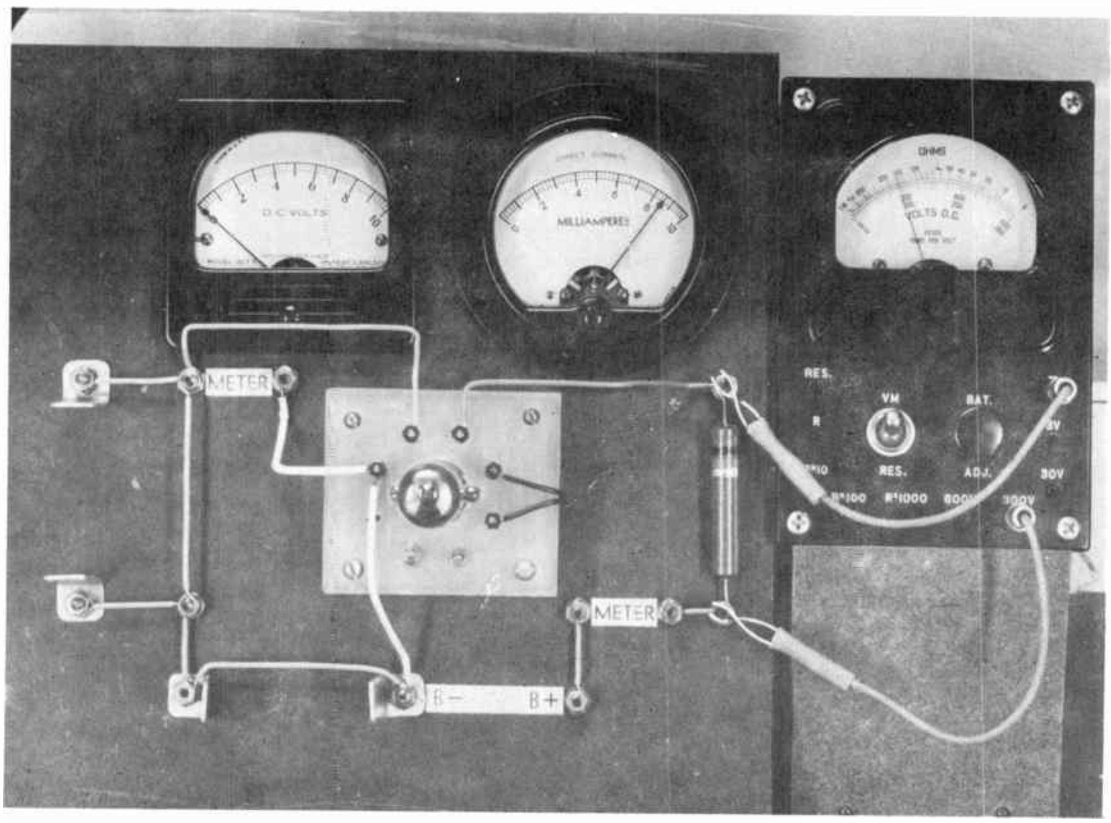


Fig. 10-19. Part of the power supply voltage is used up in the plate resistor.

The biasing dry cell and also the dry cell furnishing signal voltage have been replaced with wires. Thus the grid is connected directly to the cathode, and we have a zero grid. The volt-ohmmeter, used as a voltmeter, is connected between the B- and B+ terminals of the power supply. This meter shows that the power supply is furnishing a 200-volt potential difference for the plate circuit.

Plate current is about 8.6 ma. This is less current than in other tests with zero grid voltage, in spite of the fact that we are using in the plate circuit a voltage higher than ever before. We shall see why this happens.

In Fig. 10-19 the volt-ohmmeter, as a voltmeter, is connected across the plate resistor, and shows that potential difference across this resistor is about 100 volts. This means that 100 volts out of the total 200 volts from the power supply is being used up in forcing electron flow or current through this resistor. Note that plate current still is 8.6 ma.

With 100 volts being used up in the plate resistor, the remaining 100 volts of the total 200 volts from the power supply should be found across the tube. In Fig. 10-20 the volt-ohmmeter is connected between plate

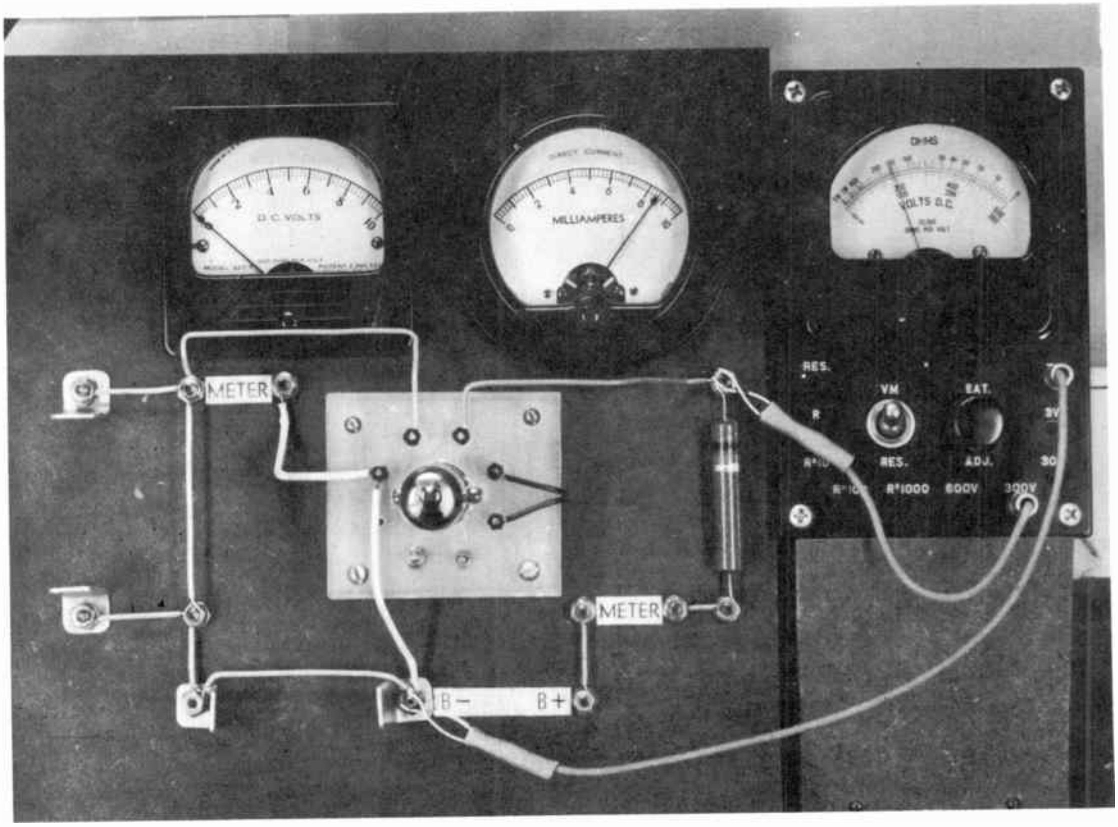


Fig. 10-20. The remainder of the power supply voltage is being used in the tube.

COMMERCIAL TRADES INSTITUTE

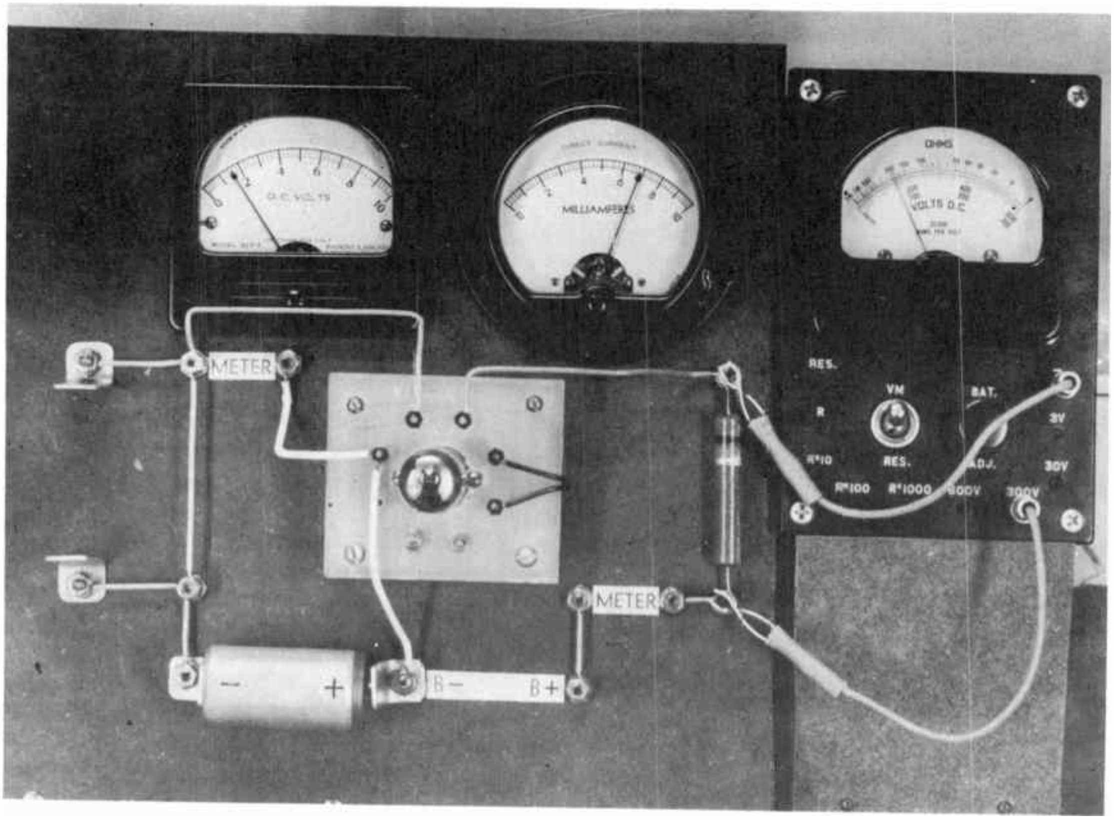


Fig. 10-21. We have no signal or zero signal, with only the biasing voltage acting on the grid.

and cathode of the tube. The positive side of the meter is connected to the plate through the wire running from the top of the plate resistor to the tube plate terminal. The negative side of the meter is connected to the B- screw terminal, from which a wire runs up to the tube cathode terminal. The meter reads about 100 volts, as it should.

It is important to know and remember, when working with any tube, that part of the plate supply voltage will be used in forcing plate current to flow in any resistance in the plate circuit outside the tube. Only the remainder of the total power supply voltage will be plate voltage for the tube. Plate voltage is the potential difference between plate and cathode of the tube. Plate voltage will be less than supply voltage, or, to say this the other way around, supply voltage must be greater than plate voltage.

SIGNAL AMPLIFICATION. A signal voltage is amplified by connecting the signal source into the grid circuit, and taking the strengthened voltage from across a resistor in the plate circuit. We shall commence with signal voltage at zero or with no signal, and with the grid biased 1.5 volts negative as in Fig. 10-21. You can see the biasing dry cell and read the meters. Plate current, which is flowing in both the tube and the plate resistor, is 6.9 ma. Across the plate resistor we measure a potential difference of about 77 volts.

In Fig. 10-22 the signal voltage has been applied, and swung to its positive peak by connecting the signal dry cell with its positive terminal toward the grid. The signal and bias voltages balance each other and there is zero voltage on the grid, as shown by the left-hand meter. Grid voltage is less negative than in Fig. 10-21, with the result that plate current rises to about 8.9 ma. The increased current in the plate resistor causes a greater potential difference across this resistor, which is shown by the meter to be 100 volts. The supply voltage remains at 200.

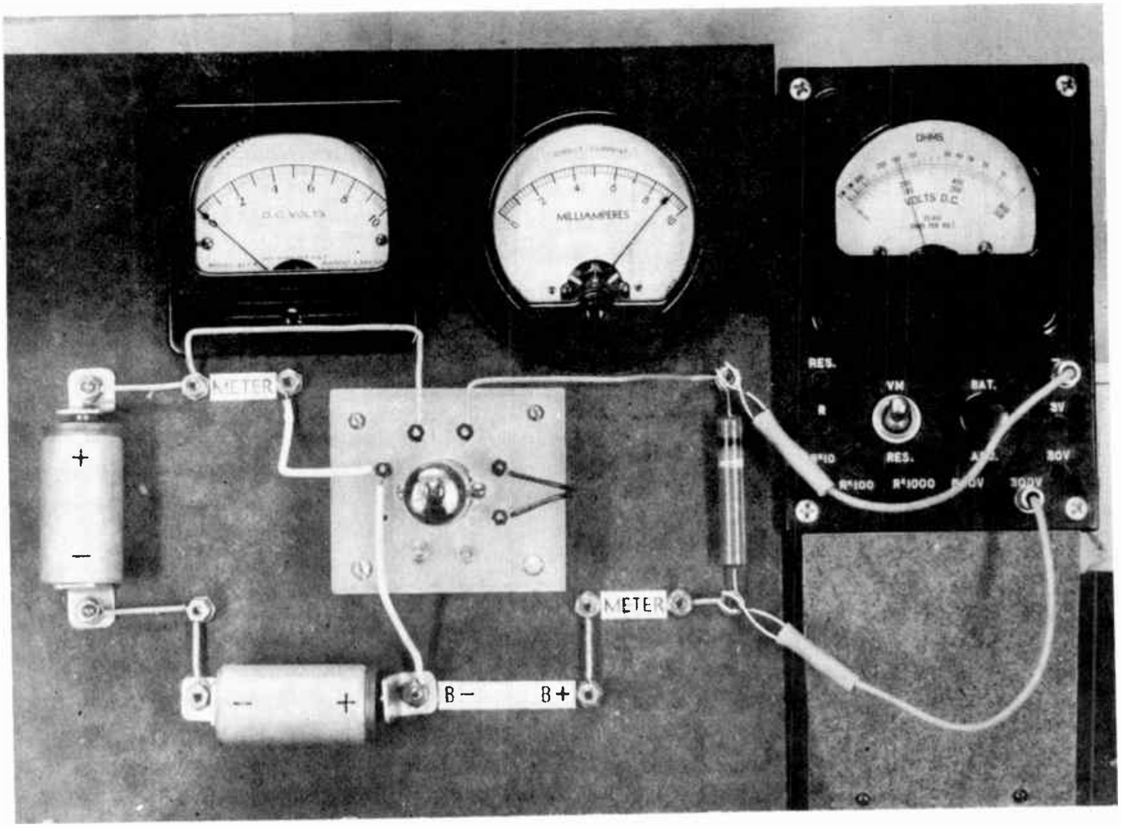


Fig. 10-22. A positive signal voltage increases the plate current and also the voltage across the plate resistor.

In Fig. 10-23 the signal voltage has been swung to its negative peak, by inverting the signal dry cell so that its negative side is toward the grid. Now the signal and biasing voltages are in the same direction or same polarity. They add together to make the grid 3.0 volts negative, as shown by the meter. The grid is more negative than in either Figs. 10-21 or 10-22, and the effect is to drop the plate current to about 5.6 ma. Across the plate resistor we now measure about 62 volts.

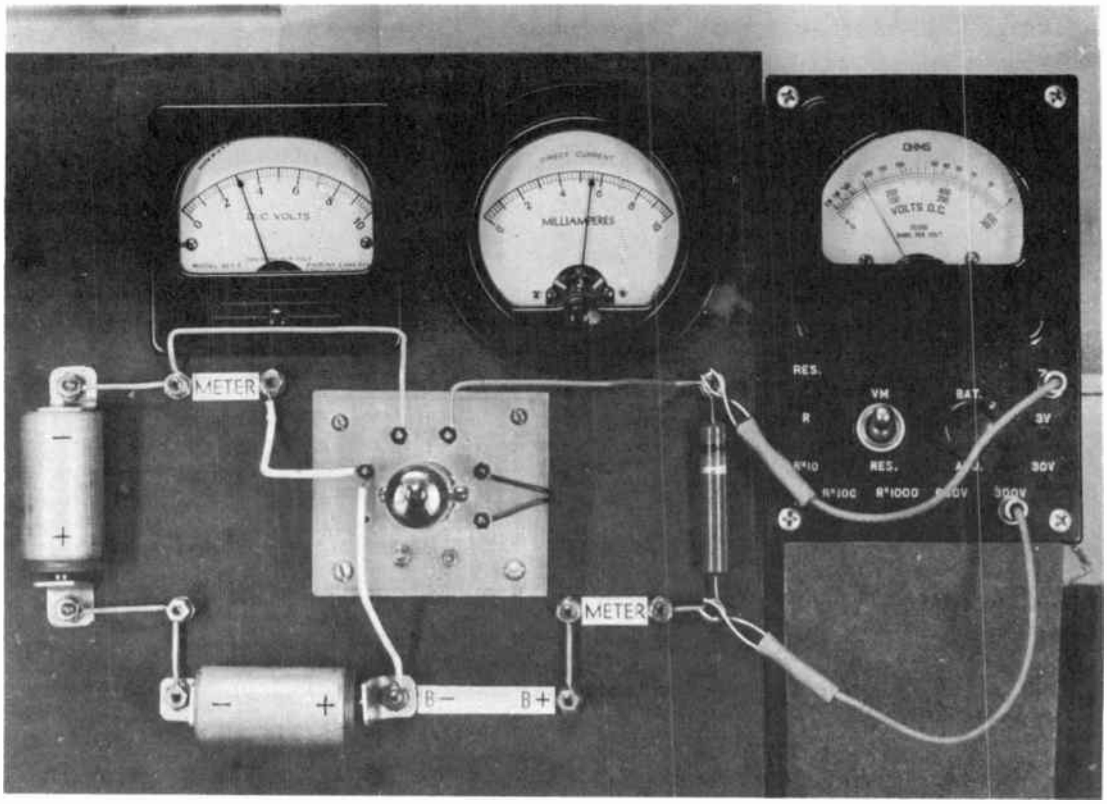


Fig. 10-23. A negative signal decreases the plate current and also the voltage across the plate resistor.

Here is a listing of what has just happened to "input signal voltage", to grid voltage, and to "output signal voltage" as measured across the plate resistor.

INPUT SIGNAL	GRID VOLTAGE	OUTPUT SIGNAL	
+1.5 volts	0 volts	100 volts	(See Fig. 10-22)
0 volts	-1.5 volts	77 volts	(See Fig. 10-21)
-1.5 volts	+3.0 volts	62 volts	(See Fig. 10-23)

The total *change* of input signal, from +1.5 volts to -1.5 volts, has been 3.0 volts. The total *change* of grid voltage, from zero to -3.0 volts, has likewise been 3.0 volts. The total *change* of output signal, from 100 down to 62 volts, has been 38 volts.

A 3-volt change of input signal on the grid has caused a 38-volt change of output signal from the plate. The output signal is about 12.7 times as strong as the input signal. The tube has amplified the signal voltage by 12.7 times.

AMPLIFICATION FACTOR. To obtain amplification of 12.7 times we have used 77 plate volts (while there is no signal or zero signal). We have used a negative grid bias of 1.5 volts. We have used a plate resistor which has measured resistance of about 11,000 ohms. Under these particular conditions the *amplification factor* of the tube being tested is 12.7. Amplification factor means the number of times an input signal voltage is multiplied with certain specified operating conditions.

A symbol for amplification factor is the Greek letter called *mu*. The form of this Greek letter is μ . Often we speak of the *mu* of a tube when referring to its amplification factor.

PLATE RESISTANCE. As mentioned before, tube manufacturers publish lists of typical operating conditions for certain types of tubes. Were such a listing to show conditions which we have been using, the values might be these.

Plate voltage (with no signal)	125 volts
Grid voltage (grid bias)	-1.5 volts
Plate current (with no signal)	6.9 ma
Transconductance (or mutual conductance)	1120 micromhos
Amplification factor	12.7
Plate resistance	11350 ohms

All of the values listed together must exist together. If any one of them is altered, the result will be to alter all the others. Manufacturers' listings give average values for all tubes of a given type. Individual tubes may vary somewhat in their performance, but new tubes of good quality will not vary enough to affect receiver operation except in a very few critical circuits, which will be discussed in due time.

The only one of the listed values not yet explained is *plate resistance.* This is the opposition to flow of alternating plate current between cathode and plate. It would be natural to think that plate resistance of the tube could be computed by putting the values of plate volts and plate current into our regular formula for resistance. For the listed values we then would have,

$$\text{Ohms} = \frac{1000 \times 125 \text{ (volts)}}{6.9 \text{ (milliamperes)}} = \frac{125\ 000}{6.9} = 18\ 120 \text{ ohms, approximate.}$$

It doesn't work out quite this way. To compute what is called plate resistance of a tube we take some small *change* of plate voltage and the accompanying small *change* of plate current, with grid voltage constant, and use these *changes* in our resistance formula.

Here is an example. If we change the plate voltage on our tube by 2.0 volts, as from 122 to 124 volts, the plate current will change from 6.825 to 7.000 ma, which is a change of 0.175 ma. Then we use these changes in the formula, thus.

$$\text{Ohms} = \frac{1000 \times 2 \text{ (volts change)}}{0.175 \text{ (milliampere change)}} = \frac{2\ 000}{0.175} = 11\ 430 \text{ ohms, approximate.}$$

This plate resistance is practically the same as listed in the table of operating conditions.

POSITIVE GRID. In all of our work up to this point the grid of the tube has been either zero or negative, never positive. Even when the signal voltage became 1.5 volts positive it was opposed by a bias voltage 1.5 volts negative, and the grid voltage was zero. As the alternating signal voltage changed from positive through zero to negative, it combined with the negative bias voltage to make the grid more and more negative.

Except in a few types of amplifiers rarely used in home receivers the grid never is allowed to become positive with reference to the cathode. The negative bias voltage is made great enough always to equal or exceed the most positive peak of an applied signal voltage. Then the signal never can completely overbalance the bias, and the grid never goes positive.

Fig. 10-24 shows what happens should the grid become positive. The grid circuit has been opened just below the signal dry cell, and between the opened ends is connected a milliammeter which will indicate any current flowing in the grid circuit. This meter is of the "zero center" type. Its pointer stands at zero in the center of the dial scale when there is no current. Current in one direction swings the pointer to the left, and with current in the opposite direction the pointer swings to the right. The maximum range of this meter is only one milliampere. From the center toward either end the scale is marked 0, then .5, and then 1 at the end.

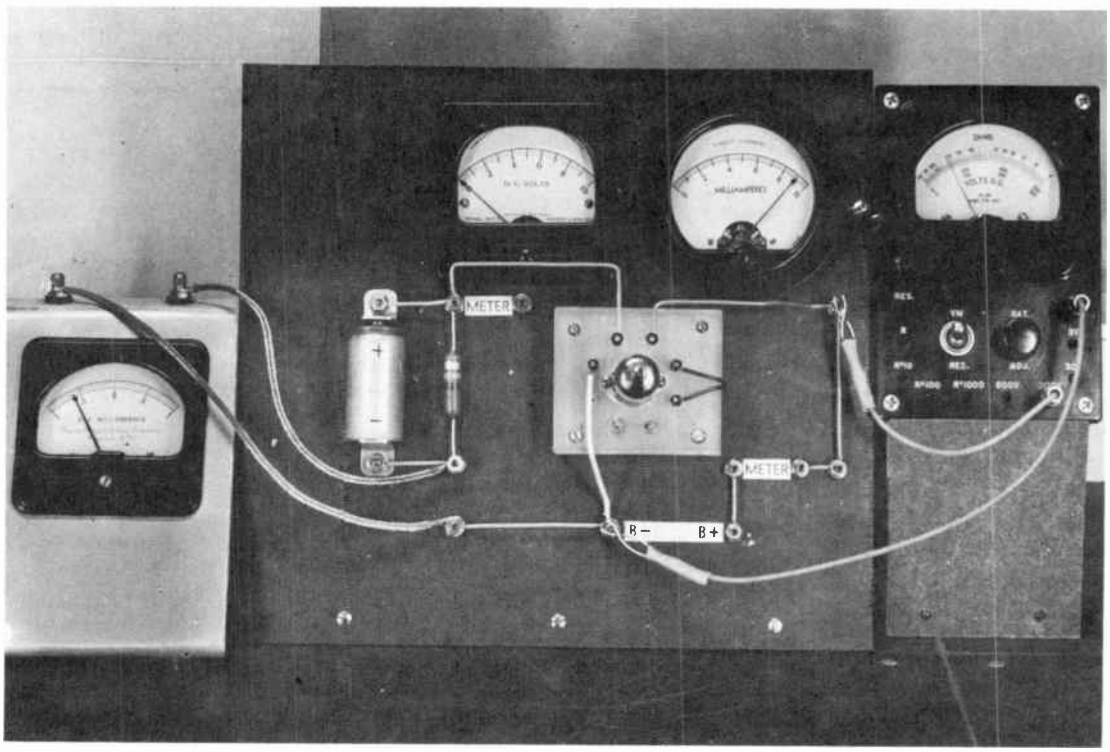


Fig. 10-24. When the grid is positive there is current in the grid and its circuit.

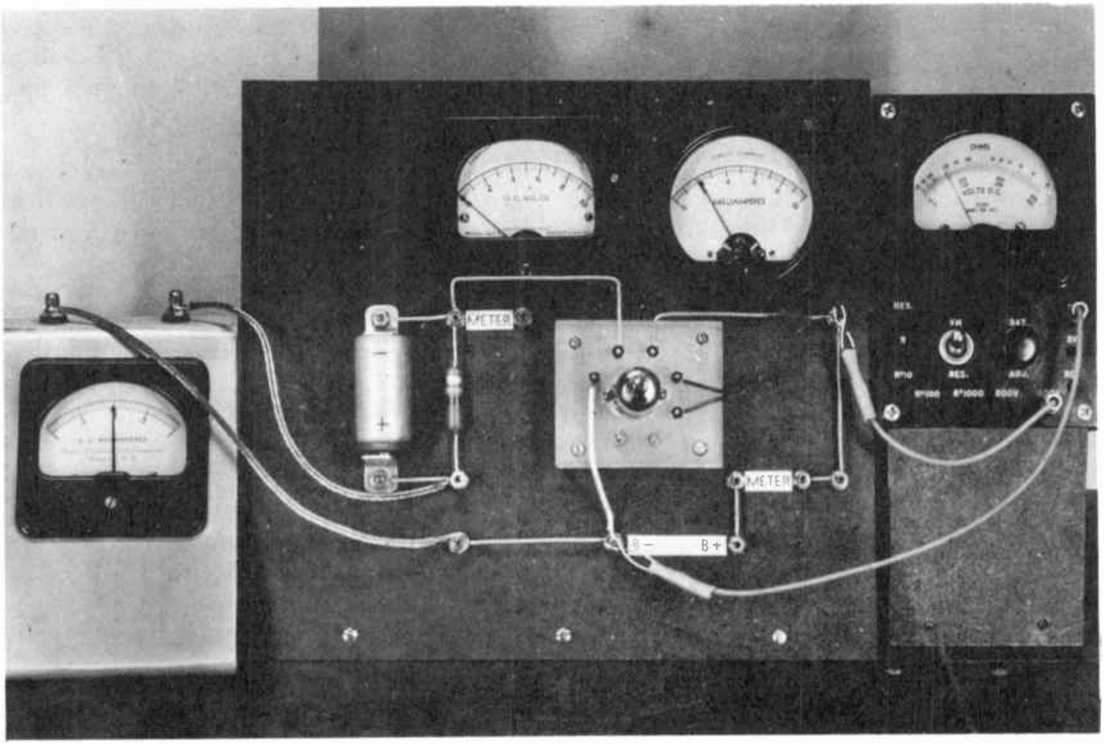


Fig. 10-25. There is no grid current when the grid is negative.

We are using no biasing voltage; the biasing dry cell has been removed and in its place is a wire connection. The signal dry cell is mounted with its positive terminal toward the grid and its negative side toward the cathode. Because there is no opposing bias voltage this makes the grid 1.5 volts positive. The regular grid voltage meter has been disconnected, because it is arranged to read only negative grid voltages. The zero center meter in the grid circuit shows that we have about 0.6 ma of grid current.

With the grid positive it attracts electrons to itself from the space charge and cathode just as does the positive plate. It is only because the positive grid voltage is here so much smaller than the positive plate voltage that the grid current is so small. Were the grid as positive as the plate there would be more grid current than plate current, because the grid is closer to the cathode and gets first chance at the space charge electrons.

When the grid is positive it not only takes electron flow or current for itself, but it gives space charge electrons such a good start toward the plate that great additional quantities go right through the grid to the plate, and we have greater plate current for any given plate voltage. In Fig. 10-24 the plate current is 9.0 ma and plate voltage is about 70. Looking back at Fig. 10-6 you will see that it takes 100 plate volts to get 9.0 ma of plate current with a zero grid. With the grid 1.5 volts negative it takes 135 plate volts to get the 9.0 ma plate current which we now have with only 70 plate volts while the grid is positive.

In Fig. 10-25 we have not replaced the biasing dry cell, but have swung the signal voltage negative by inverting the signal dry cell. The pointer of the zero center meter has not swung to the opposite side of its dial, as would have happened were there grid current in the opposite direction. This meter shows that there is no grid current, even though we have the same 70 volts on the plate. Because of the negative grid, plate current is down to about 2.2 ma.

A negative grid takes no grid current. A zero grid draws grid current so small as to be measured in microamperes, and for practical purposes the zero grid takes no grid current. A positive grid always takes grid current. As we proceed with our work you will learn that grid current can cause plenty of trouble, principally because the output signal voltage becomes decidedly different from the input signal.

Since there can be no grid current in the grid circuit with the grid zero or negative, the only current taken from the signal source is that which flows in the resistor which is across the signal source and between grid and cathode. By using high resistance at this point the current taken from a signal source may be made very small.

It is true also that any source of biasing voltage connected in the grid circuit as in Figs. 10-16, 10-22 and others is called upon to furnish no current when the grid is zero or negative. The bias source furnishes only a voltage, no current.

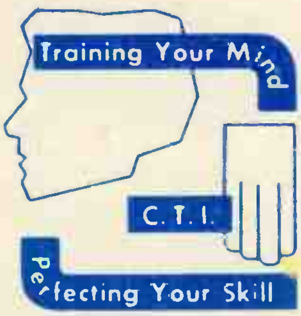
11→15

COMMERCIAL TRADES INSTITUTE

KNOWLEDGE + AMBITION = SUCCESS

LEARN . . .

ACT . . .



"The heights by great men reached and kept
Were not attained in sudden flight,
But they, while their companions slept,
Were climbing upward in the night."

ENJOY . . .

COMMERCIAL TRADES INSTITUTE

CHICAGO, ILLINOIS

TELEVISION

LESSON NO. 11 PENTODES AND BEAM POWER TUBES

"Training Your Mind"

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

COMMERCIAL TRADES INSTITUTE

Training Your Mind

C.T.I.

Perfecting Your Skill

Chicago, Ill.

LESSON No. 11

PENTODES AND BEAM POWER TUBES

In the early days of radio, long before television, only triodes and diodes were available. Since diodes won't amplify a signal, all the amplifier tubes were triodes. The triodes worked very well as audio or sound frequency amplifiers, and performed acceptably at the low-frequency end of the standard broadcast band. But at the high end of this band the triode amplifiers caused plenty of trouble.

The trouble came about because the plate and grid of a triode act like a small capacitor. These two elements are the capacitor plates, and the vacuum between them is the dielectric. The relatively strong signal voltages in the plate circuit can feed back through this *grid-plate capacitance* to the grid circuit, and there they act in such polarities as to reinforce the received signal voltages.

This feedback would be welcome insofar as it strengthens the grid signals, but it doesn't stop there. The feedback voltage becomes progressively stronger as the receiver circuits are tuned to higher frequencies. Feedback increases because opposition of any capacitance to alternating voltages and currents becomes rapidly less as frequency rises. When the early radios were tuned to higher signal frequencies, the alternating signal voltages in the grid circuit lost their control.

The tube and its grid and plate circuits commenced to "oscillate", and produce their own voltages at a frequency determined by the values of feedback capacitance in the tube and of inductances in the grid and plate circuits. The frequency of such oscillating voltages and currents is so far above the audible range that, by themselves, the oscillations produce no sound from the speaker. But the oscillating voltages mix with voltages at other frequencies, and in those old sets the combination caused the speaker to emit uncontrollable whistles and shrieks. Even worse, the receiver became a small transmitter and radiated signals which the neighbor's sets reproduced as screeches and growls.

After trying dozens of trick circuits in their efforts to prevent oscillation, the engineers attacked the real cause for the trouble, the grid-plate capacitance within the tube. Let's see how they did it.

SCREEN GRID TUBES. To build a triode we would commence by placing a grid around the cathode, as in Fig. 11-1. This grid is a spiral of very fine wire closely spaced, for this is the way to give grid voltage the greatest control over plate current. In the triode the plate would be placed around the outside of the grid. Instead, we shall use another spiral of fine, closely spaced wire, as in Fig. 2. This element may be called the screen grid, but usually is called the *screen.* Were a plate to be placed around the screen we would have a tube with four active elements; cathode, grid, screen, and plate. A common name for a four-element tube is *tetrode.*

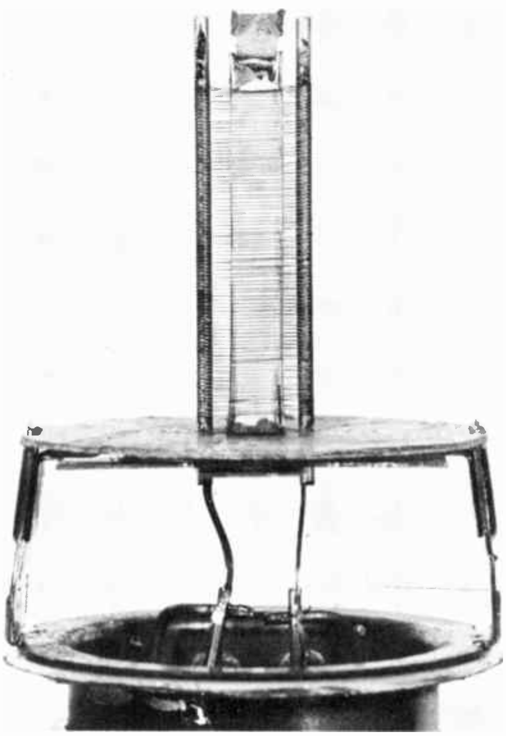


Fig. 11-1. The cathode is surrounded by the grid.

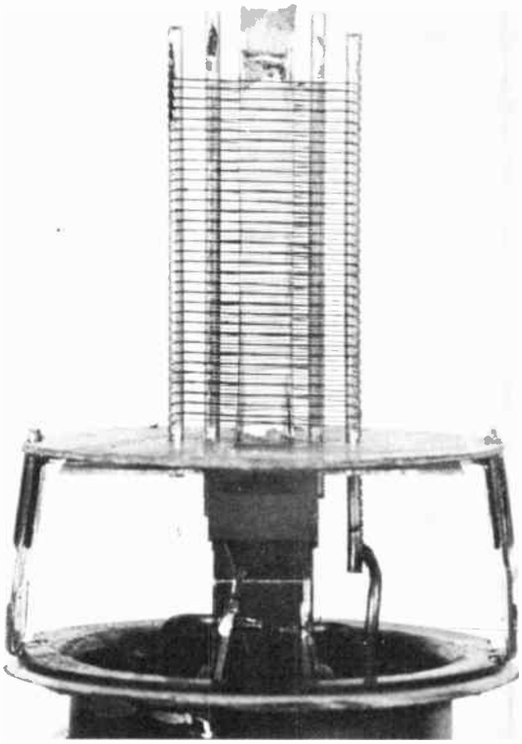


Fig. 11-2. The screen, between grid and plate, reduces grid-plate capacitance.

Fig. 11-3 shows how the screen is connected to the same d-c power supply as the plate, but to a point of less positive voltage. With the screen positive with reference to the cathode, emitted electrons are pulled toward the screen. By the time the electrons reach the screen they are traveling so fast that most of them go right through the openings between wire turns and go to the still more positive plate.

The real purpose of the screen is not to pull electrons from the cathode toward the plate; this could be done as well or better without a screen, letting the plate exert its attraction directly on electrons emitted from the cathode. The real purpose of the screen is to lessen the grid-plate capacitance. The screen is made positive only in order that electrons shall pass from cathode to plate. How the screen reduces grid-plate capacitance depends on the following facts.

Cathode potential remains constant, it does not become alternately positive and negative at the frequency of received signals. Between connections of the d-c power supply to cathode and screen there is very little opposition to alternating signal currents and voltages. Because of this small opposition, the potential of the screen cannot vary with reference to cathode potential, but must remain at whatever positive potential is applied from the power supply.

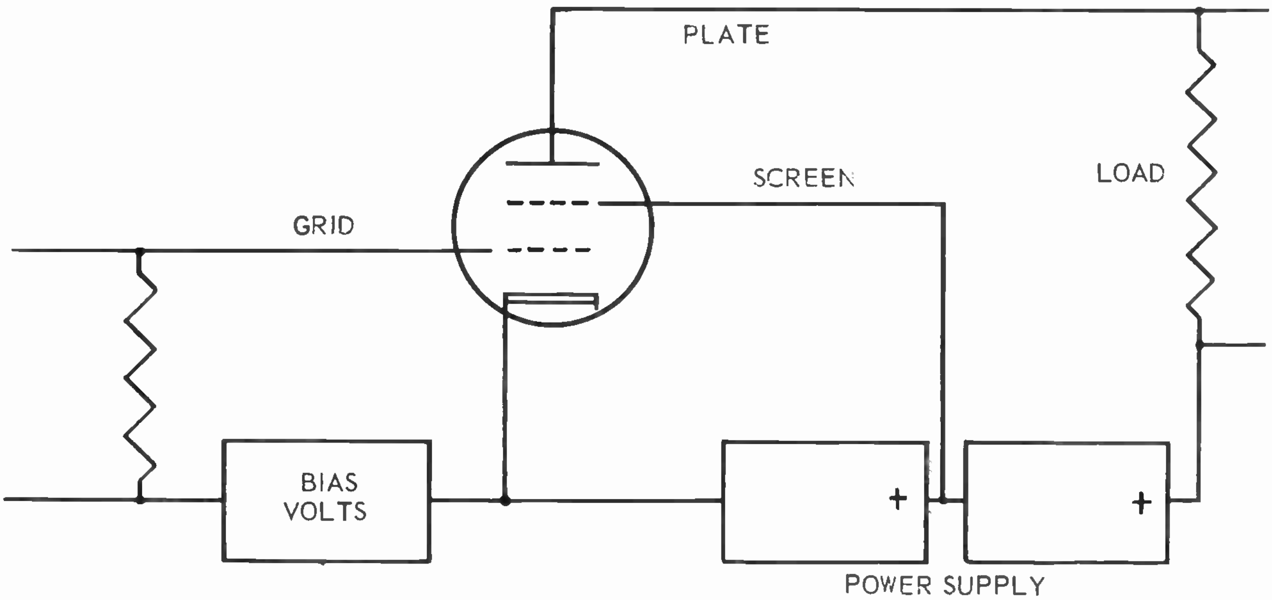


Fig. 11-3. Connections of the plate and screen to the power supply.

The screen, held at constant potential, is located between grid and plate. In order that voltage feedback might occur there would have to be variations of electric field strength in the space between these elements, at the signal frequency. There cannot be appreciable variations so long as screen potential does not vary, and so the effect of grid-plate capacitance is reduced.

In early types of triode amplifiers the grid-plate capacitance was 3.5 to 8.0 mmf. Screen grid tubes or tetrodes have grid-plate capacitance of about 0.007 mmf. This very small capacitance just about does away with plate-to-grid feedback. With feedback so reduced, high amplification becomes practicable. Amplification factors of early triodes were about 9; tetrodes have computed factors of 600 to 800.

SECONDARY EMISSION. The tetrodes worked fine on weak incoming signals, and provided large amplification for such signals. But when a fairly strong signal was applied to the control grid, the amplification would fall off and the form of the output signal would not follow the form of the input signal. That is to say, the tetrodes would distort strong signals.

The drop in amplification and the distortion are the results of secondary emission. Secondary emission means emission of electrons from the plate. We have been proceeding on the assumption that there can be no emission from a cold plate, and it is true that there is no useful emission from the plate. However, there can be electron emission from any body provided some of its free electrons are given enough extra energy to let them get through the surface of that body.

The extra energy which causes emission from the plate comes from electrons emitted from the cathode in the usual way and which are hitting the plate at tremendous velocity. With a plate potential of 150 volts the electrons hit the plate at a speed of more than 16 million miles per hour, and at 300 plate volts the speed is more than 23 million miles per hour. Such bombardment literally knocks electrons out of the cold plate.

There is secondary emission in a triode as well as in a tetrode, but because the plate of the triode is the only positive element, all the negative electrons freed by secondary emission are pulled back into the plate. In the tetrode, however, there is also the positive screen which is willing and able to attract the electrons. If electrons knocked out of the plate enter the screen and leave the tube by way of the screen there will be a reduction of useful plate current.

Secondary electrons will enter the screen only when the screen becomes just as positive as the plate or more positive than the plate. Now we may ask, if secondary emission is bothersome only when plate voltage drops nearly as low or lower than screen voltage, why not keep plate voltage high enough and screen voltage low enough to prevent trouble? There are two reasons. First, if plate resistance of the tube is to be reasonably low, the screen voltage must be quite high – because it takes screen voltage to pull electrons toward the plate. Second, if we are to make use of the high amplification factor the plate voltage must go through wide swings. With wide swings of plate voltage and a fairly high screen voltage, plate potential will come close to or even below screen potential during signal peaks.

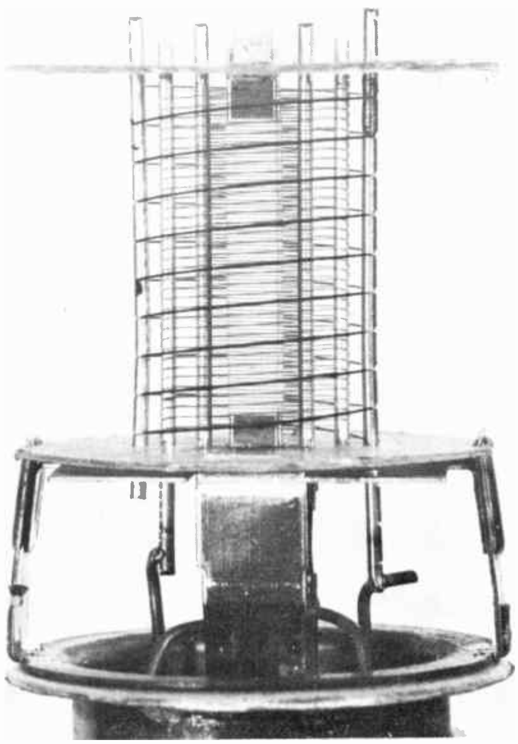


Fig. 11-4. The suppressor is wound around the outside of the screen.

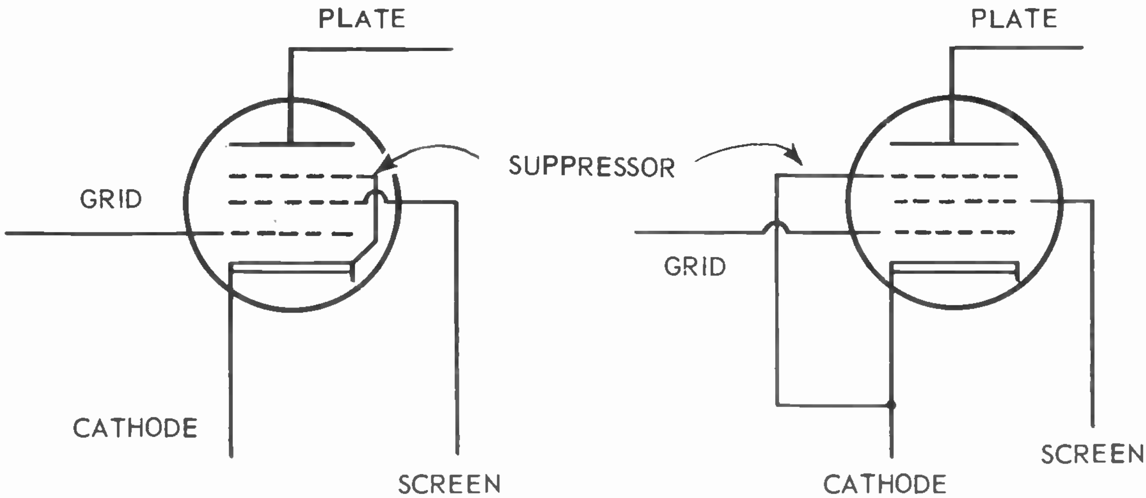


Fig. 11-5. The suppressor is placed between the screen and the plate.

THE PENTODE. To prevent ill effects of secondary emission we shall add still another element to our tube, as pictured by Fig. 4. This element consists of a spiral of wire between whose turns the spacing is greater than for either the grid or the screen. This new element is called a suppressor grid, or simply a *suppressor.*

When a plate is placed around the outside of the suppressor the completed tube will have five active elements; cathode, grid, screen, suppressor, and plate. A five-element tube is called a *pentode.* The position of the suppressor in relation to other elements is shown by Fig. 11-5. In some pentodes the suppressor is internally connected to the cathode. In others there is an extra base pin for the suppressor, which then may be connected at the socket lugs to the cathode, to chassis ground, or to any other point required by circuit design.

Whatever may be the connection of the suppressor, this element is maintained highly negative with reference to the plate. Secondary electrons emitted from the plate encounter the electric field around the negative suppressor before they can reach the screen, and these electrons are repelled back to the plate. Some electrons emitted from the cathode will enter the screen to form screen current. But most of them will be traveling so fast as they reach the screen that they go right on through both screen and suppressor to reach the plate. None of these electrons can get from the plate back to the screen because, on the side of the screen which is toward the plate, the positive attractive force of the screen is nullified by the negative charge of the suppressor. Even though the plate becomes less positive than the screen, the negative suppressor still prevents secondary electrons from passing from plate to screen.

The tube to which we have added a screen and a suppressor is completed by putting on the plate, as in Fig. 6. The plate of this particular tube consists of two sections, one of which is shown by the left-hand picture. At the right both plates are in place and all elements are held in rigid alignment by side supports.

PENTODE PERFORMANCE. The tube whose structure has been shown by Figs. 1, 2, 4, and 6 is a 6AC7 voltage amplifier pentode designed especially for television reception. This tube formerly was used

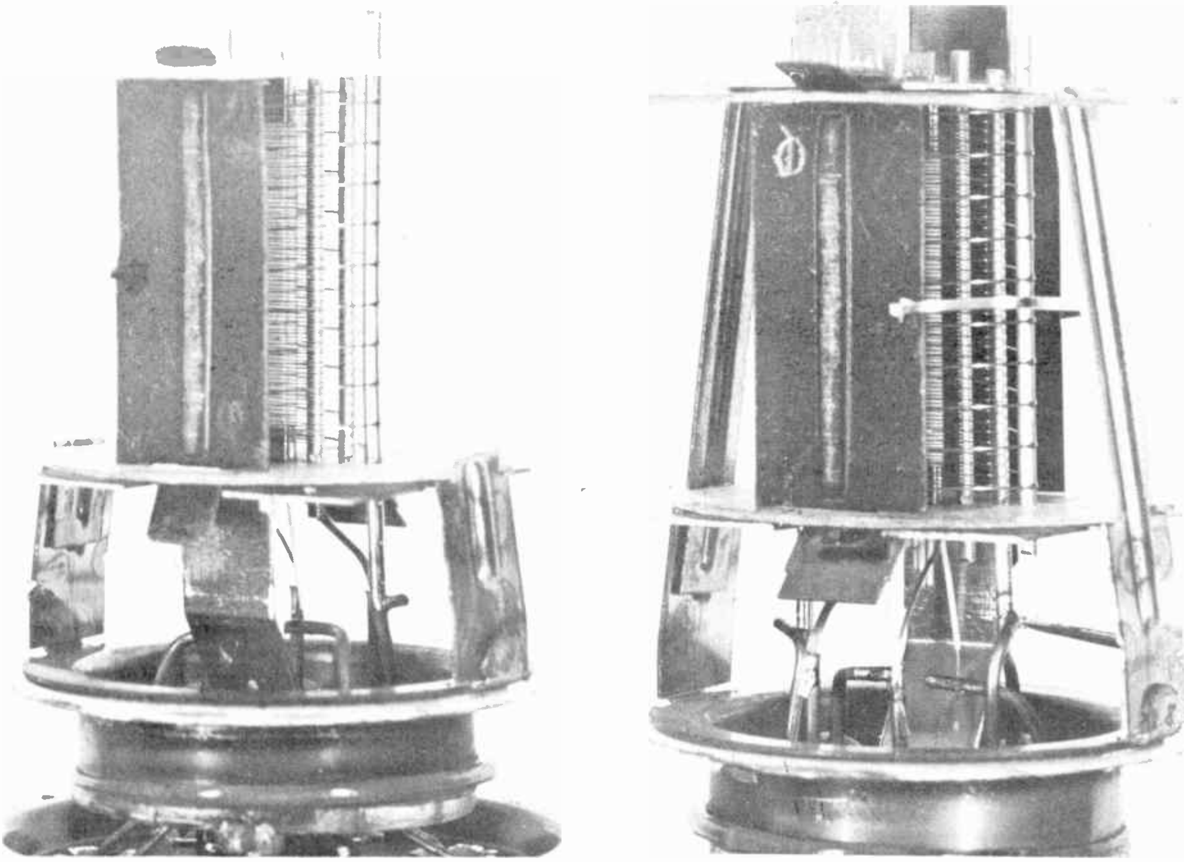


Fig. 11-6. The pentode is completed by adding the plate and supports for the elements.

in tuners, i-f amplifiers, and video amplifiers, but now is found most often in video amplifiers. Operating characteristics of the miniature pentode type 6AH6 are quite similar to those of the 6AC7.

To watch the performance of our pentode we shall measure voltages and currents for all elements simultaneously, using eight meters connected as shown by Fig. 7. Regular circuit connections to tube elements are in heavy lines, while added meter connections are in light lines, indicating that they would not be present for normal operation.

To allow making pictures of the performance it is necessary to use direct voltages. D-C input voltage will be varied from positive to negative through zero. An alternating signal voltage would vary in the same way, but would change too fast for the meters to indicate or for the camera to catch.

The 6AC7 tube is in one of three tube sockets at the top of the test panel shown by Fig. 8. Two other sockets accomodate 7-pin and 9-pin miniature tubes. Back of the panel are terminals which allow connecting the socket lugs to circuits for plate, screen, grid, and cathode, also to the power supply.

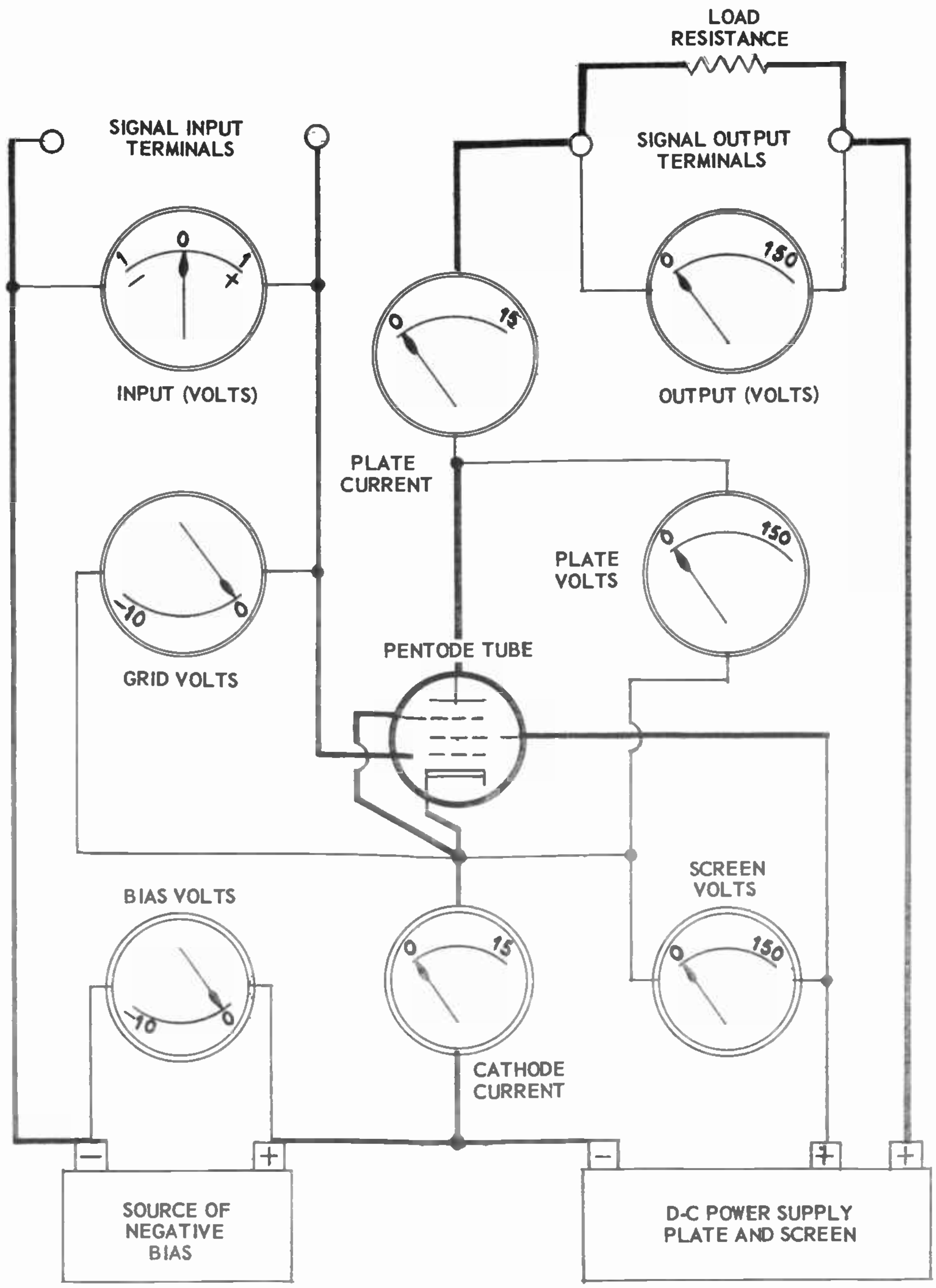


Fig. 11-7. How the meters are connected to the tube circuits for measuring performance of the pentode, shown by its symbol at the center.

One of the signal input terminals, at the upper left, connects directly to the grid of the tube, while the other connects through the bias source and a meter to the cathode. Input voltage is measured by the zero-center *Input* meter, reading from 1 volt negative through zero to 1 volt positive.

Bias voltage is measured by the *Bias Volts* meter, connected across the source of negative bias. Since bias always is negative, this meter reads from zero at the right-hand end of its scale to 10 volts negative at the left. The positive or negative signal input voltage combines with the negative bias voltage to make the actual grid voltage. This grid voltage is measured by the *Grid Volts* meter connected between grid and cathode of the tube. This meter reads from zero at the right to 10 volts negative at the left-hand end of its scale.

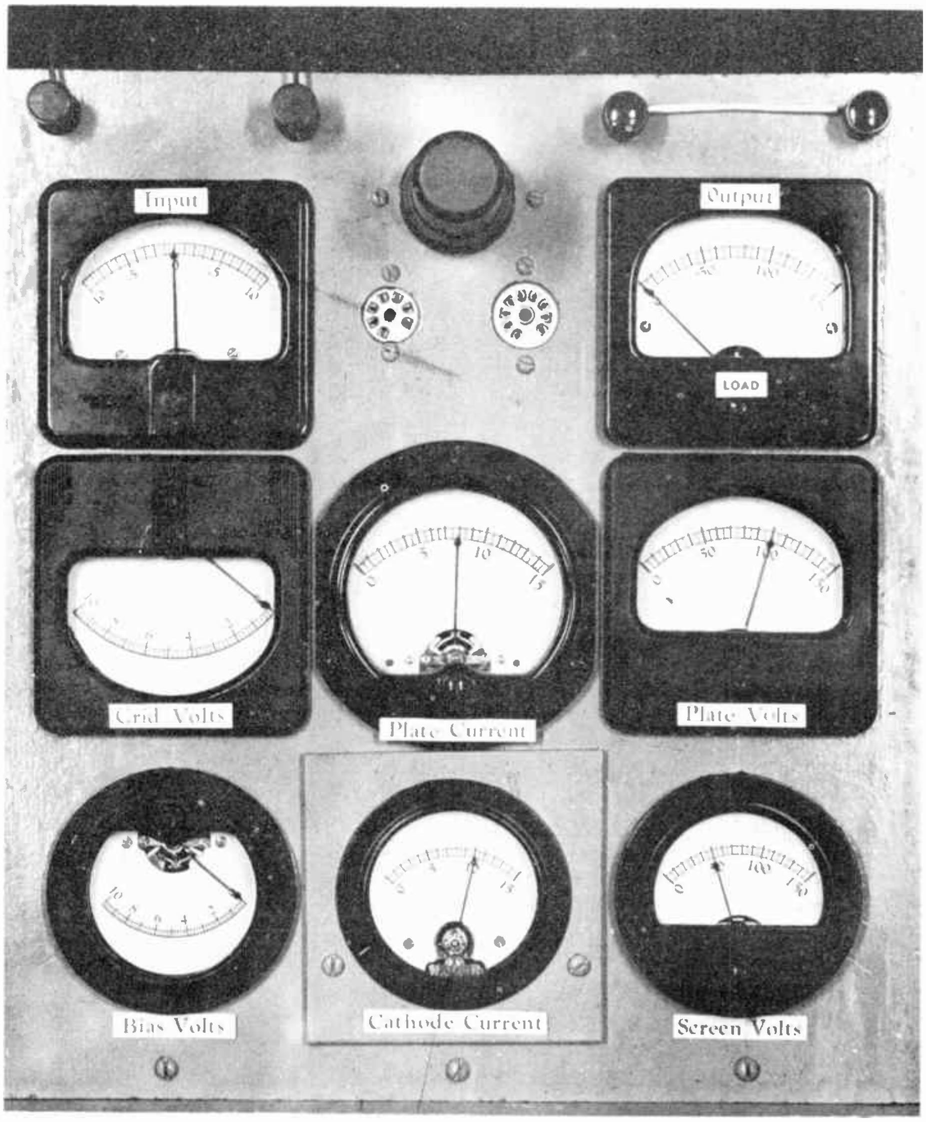


Fig. 11-8. The first step in measuring the effect of grid voltage on tube currents.

Plate current is measured by the 0-15 milliampere *Plate Current* meter in series with the tube plate. In series between the plate and d-c voltage supply is provision for a load resistance between the two terminals immediately above the *Output* meter. This meter, reading 0-150 volts, indicates whatever voltage is across the load.

Plate voltage is measured by a 0-150 volt *Plate Volts* meter connected between plate and cathode of the tube. Screen voltage is measured by the 0-150 volt *Screen Volts* meter connected between screen and cathode of the tube.

Total cathode current is measured by the 0-15 milliampere *Cathode Current* meter in series with the tube cathode. Cathode current is the sum of plate and screen currents, so to determine screen current we need only subtract plate current readings from those of total cathode current.

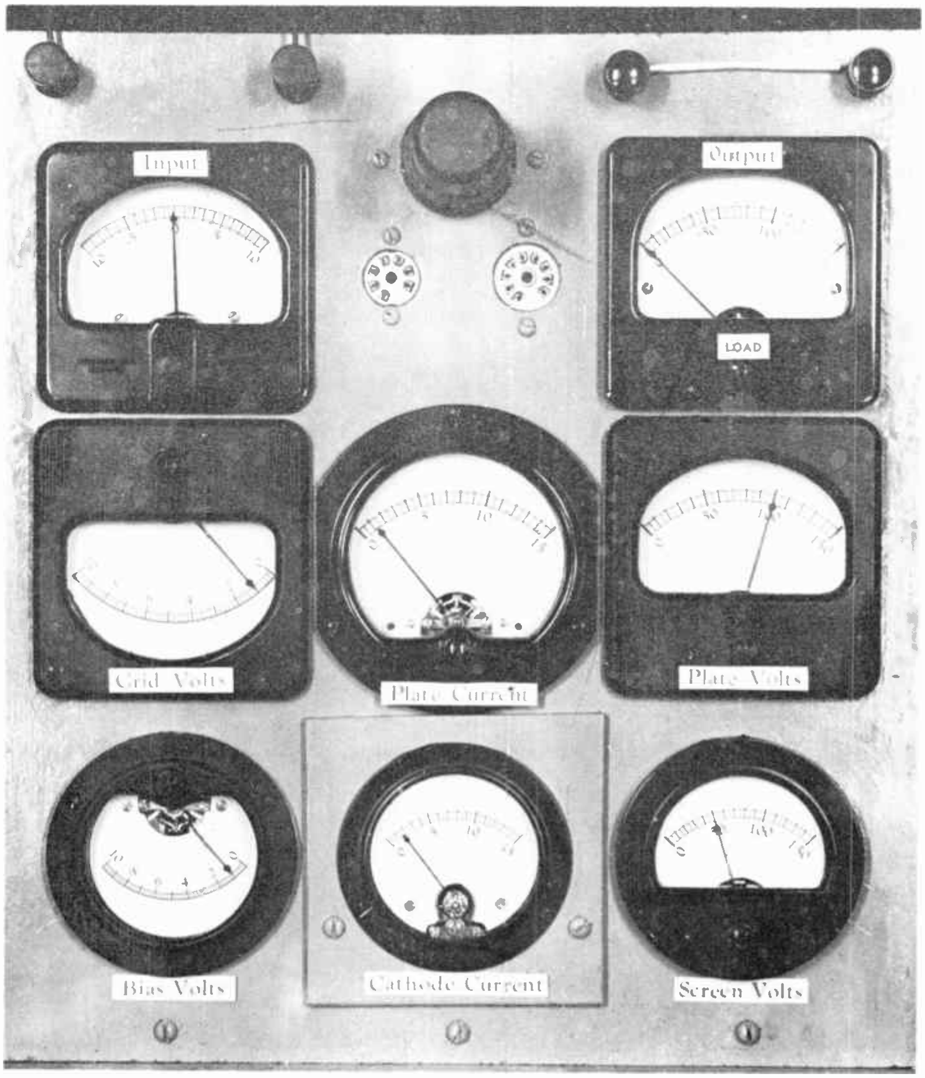


Fig. 11-9. Making the grid negative decreases electron emission and all tube currents.

GRID VOLTAGE EFFECTS. For the first test, Fig. 8, there is zero input or zero signal voltage and there is zero grid bias, as indicated by the respective meters. Consequently the grid is at zero voltage with respect to the cathode, as indicated by the *Grid Volts* meter. There is no load resistance in the plate circuit. The signal output terminals, where such a resistance would be connected, are joined together by a piece of wire. The *Output* meter is short circuited by this wire, and reads zero volts.

Now, still with no signal voltage and without altering the voltages on plate and screen, we make the bias 1 volt negative, then 2 volts negative, as in Figs. 9 and 10. With no signal, the grid voltage is the same as the bias voltage. Observe how the tube currents are changed by altering the grid voltage with everything else unchanged. In the following summary and in others for later tests the voltages and currents are designated by these letter symbols.

Signal input volts	*Esig*	Output volts	*Eo*	Plate current	*Ip*
Grid voltage	*Eg*	Plate voltage	*Ep*	Cathode current	*Ik*
Bias voltage	*Ec*	Screen voltage	*Escr*	Screen current	*Iscr*

Grid Voltage Effect On Plate And Screen Currents

Figure	Esig	Ec	Eg	Ep	Escr	Ik	Ip	Iscr	Load
8	0	0	0	100	50	10.1	7.9	2.2	0
9	0	-1.0	-1.0	100	50	1.6	1.1	0.5	0
10	0	-2.0	-2.0	100	50	0	0	0	0

Making the grid more negative decreases the plate current, and making the grid less negative increases the plate current, just as in the case of a triode. Grid voltage controls total electron emission. Consequently, with the grid more negative there is less cathode current and less screen current, as well as less plate current.

If the grid is made sufficiently negative it stops all cathode current, as in Fig. 10. Then there is zero plate current, or there is the condition of plate current cutoff. With 100 volts on the plate and 50 volts on the screen of the particular tube being tested, plate current cutoff occurs with the grid 1.8 volts negative and, of course, with any grid voltages still more negative.

PLATE VOLTAGE EFFECTS. To observe the effects of plate voltage on currents in a pentode we shall go back to Fig. 8, where there is 100 volts on the plate and 50 volts on the screen, then alter the plate voltage with no change of screen voltage.

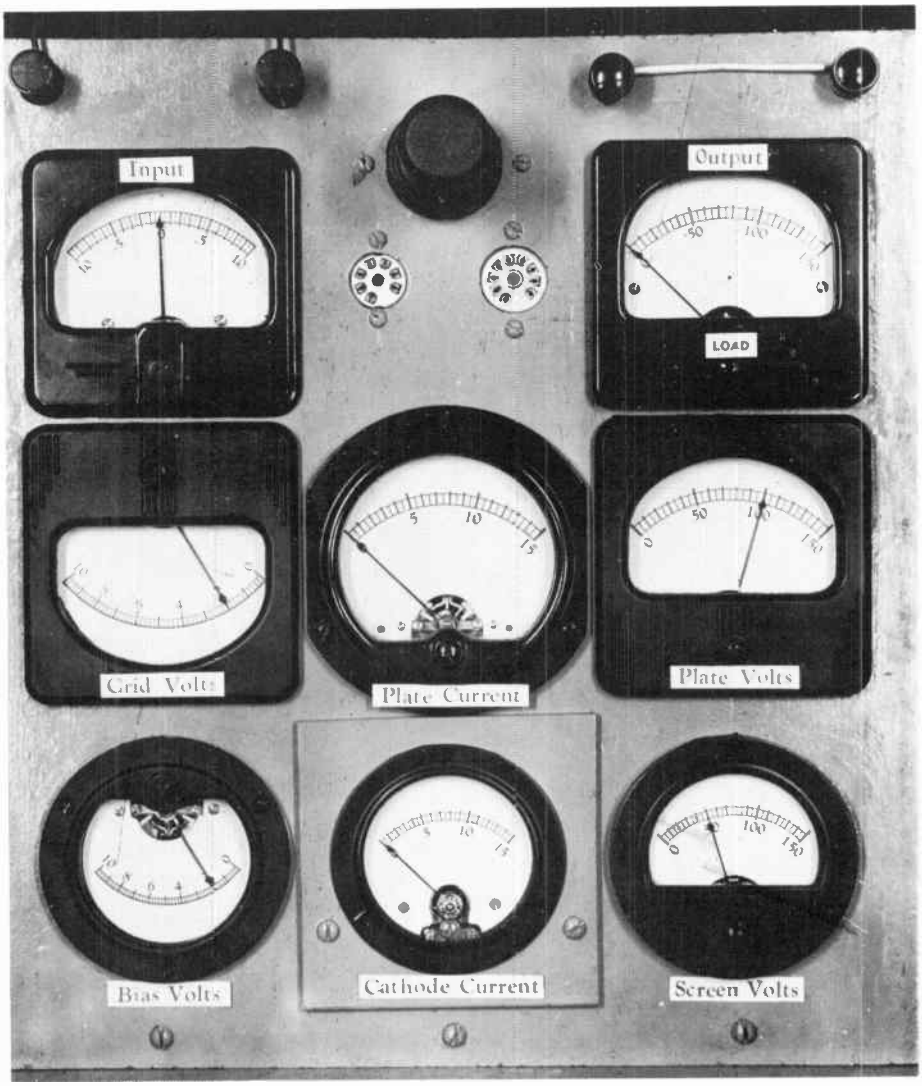


Fig. 11-10. With the grid still more negative there is further reduction of all currents.

Plate Voltage Effect On Plate And Screen Currents

Figure	Esig	Ec	Eg	Ep	Escr	Ik	Ip	Iscr	Load
8	0	0	0	100	50	10.1	7.9	2.2	0
11	0	0	0	50	50	10.2	7.1	3.1	0
12	0	0	0	150	50	10.2	8.0	2.2	0

The tests illustrated by Figs. 8, 11, and 12 bring out this rather peculiar property of pentodes; changes of plate voltage in excess of some certain value have almost negligible effect on plate current. When we reduce the plate voltage to half its earlier value, to 50 volts in Fig. 11, there is some reduction of plate current, but not nearly so much as in a triode when its plate voltage is similarly reduced. When we increase the plate voltage on the pentode by 50 per cent, or to 150 volts in Fig. 12, the plate current increases by only 0.1 ma.

When plate voltage is dropped to the same value as screen voltage, Fig. 11, there is a decided increase of screen current, from 2.2 to 3.1 ma, while there is a drop of plate current. Increasing the plate voltage, Fig. 12, has no noticeable effect on screen current.

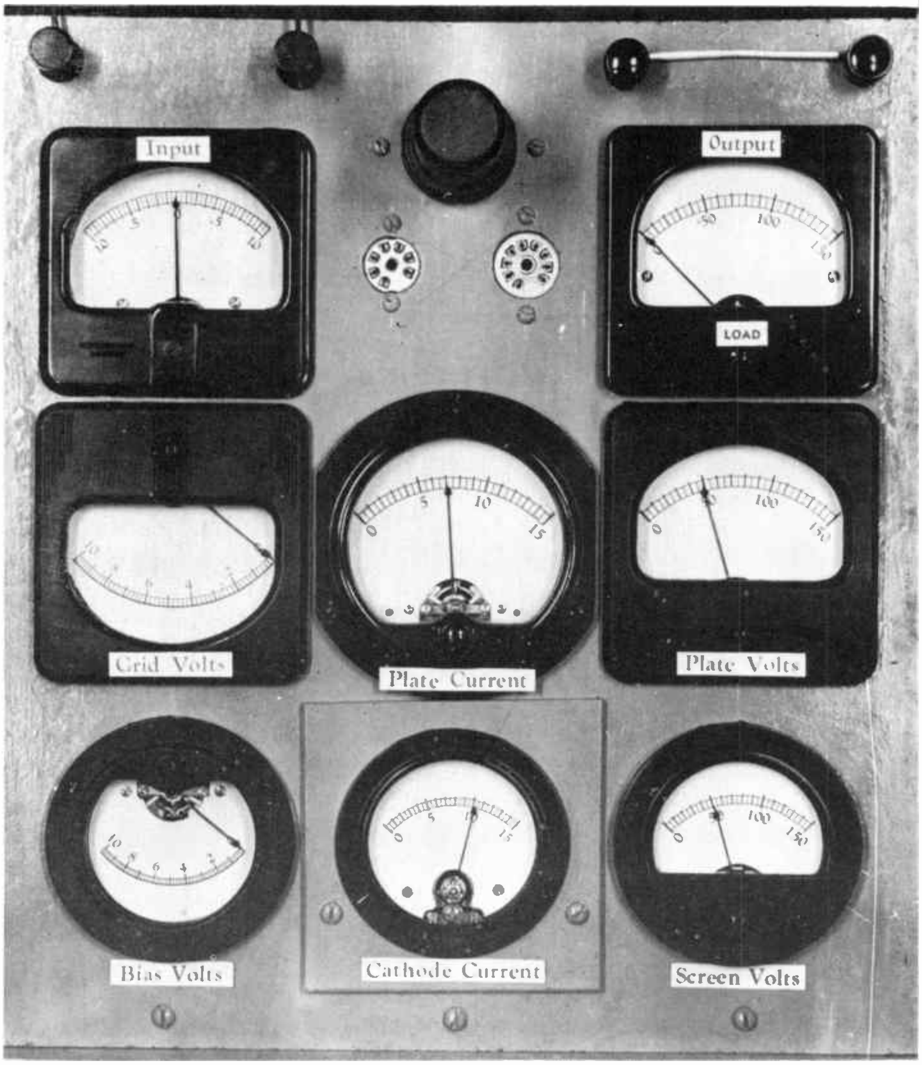


Fig. 11-11. Making the plate less positive causes only a slight reduction of plate current.

SCREEN VOLTAGE EFFECTS. Once again we shall go back to the conditions of Fig. 8 as a starting point, then change the screen voltage while holding everything else constant. The results are shown by Figs. 13 and 14, as follows.

Screen Voltage Effect On Plate And Screen Currents

Figure	Esig	Ec	Eg	Ep	Escr	Ik	Ip	Iscr	Load
8	0	0	0	100	50	10.1	7.9	2.2	0
13	0	0	0	100	30	6.0	4.6	1.4	0
14	0	0	0	100	70	14.8	11.3	3.5	0

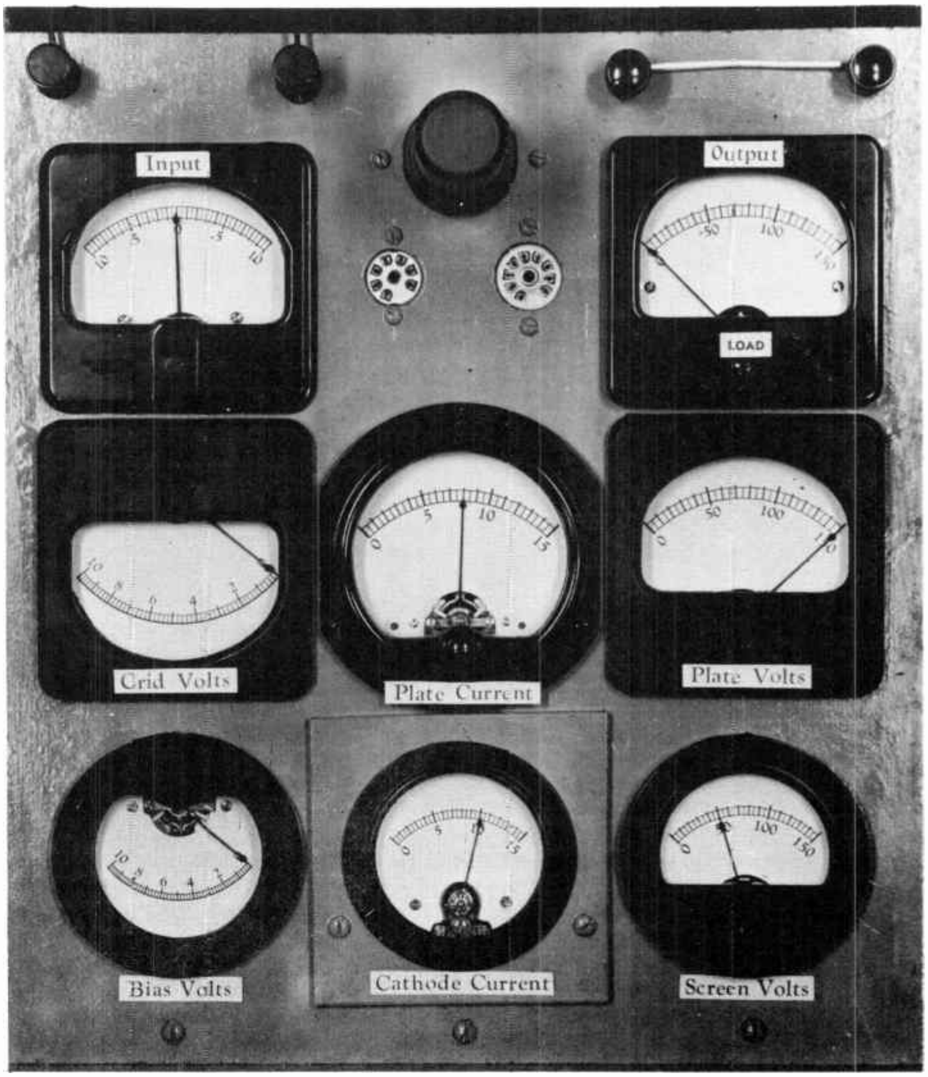


Fig. 11-12. A more positive plate causes hardly any increase of plate current.

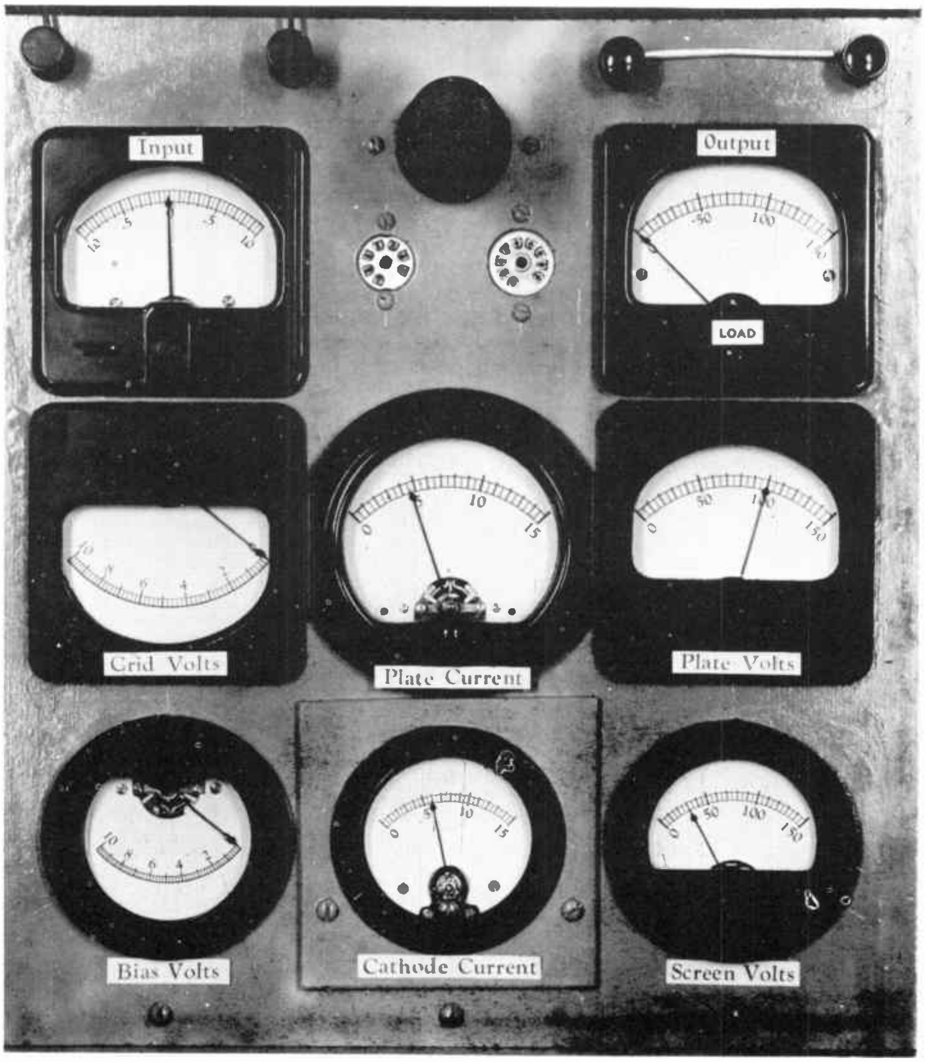


Fig. 11-13. Lowering the screen voltage drops the plate current.

In these tests we discover the factor which has real control over plate current in a pentode, it is screen voltage. Dropping the screen voltage by only 20 volts, from 50 volts in Fig. 8 to 30 volts in Fig. 13, brings plate current down to little more than half its former value. When we increase the screen voltage from 50 in Fig. 8 to 70 in Fig. 14, plate current goes up by almost 50 per cent. This is something you should not forget when working on service jobs; pentode plate current is determined almost entirely by screen voltage, and is little affected by plate voltage.

As we should expect, varying the screen voltage has a decided effect on screen current. For low values of screen voltage the screen current is just about proportional to screen voltage, but for higher screen voltages the screen current increases at a greater rate than screen voltage.

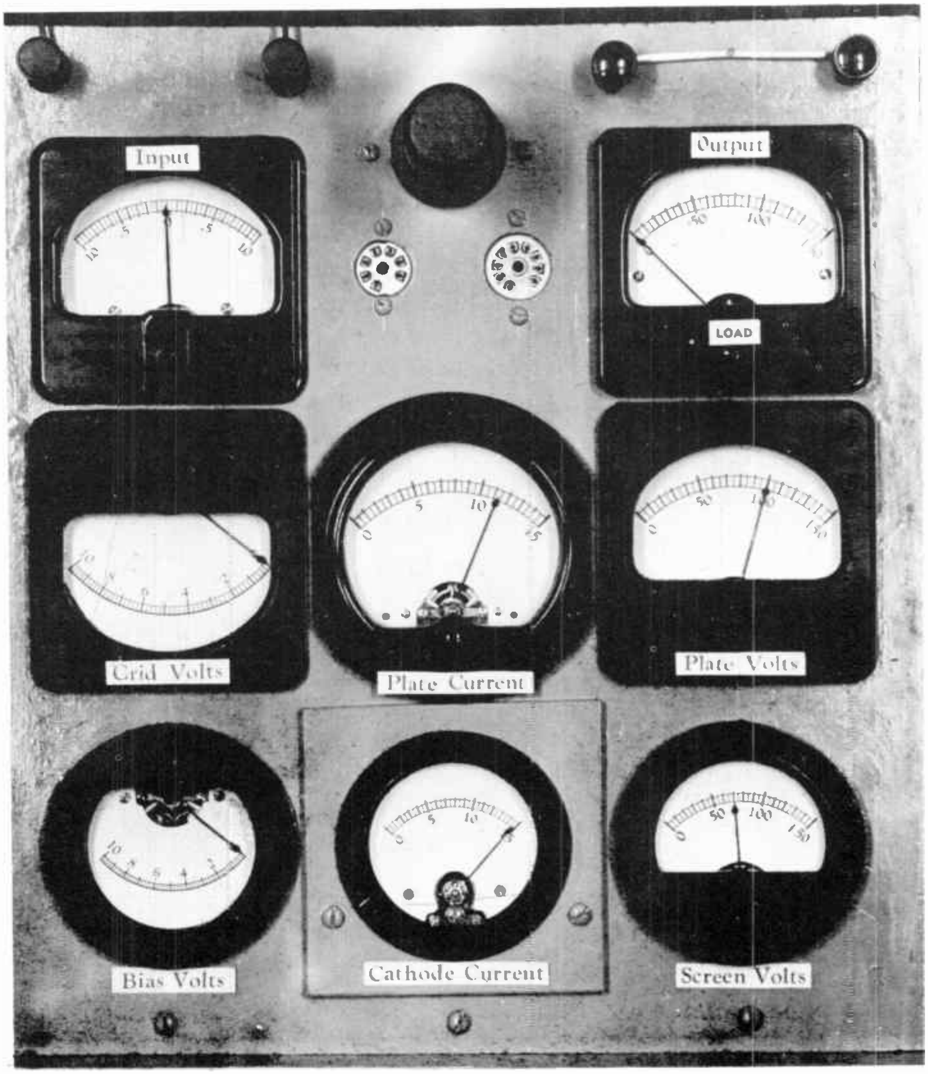


Fig. 11-14 Plate current is controlled almost wholly by screen voltage.

PENTODE PLATE CHARACTERISTICS. Everything which we have seen happen in Figs. 8 through 14 may be shown more completely and for wider ranges of operating conditions by a family of plate characteristics for whatever tube is under consideration. Pentode plate characteristics appear generally similar in form for all types of tubes. We have been working with a voltage amplifier pentode. Characteristics for a power amplifier pentode are shown by Fig. 15. Except that the power tube will safely carry greater currents, and possibly may be worked at higher voltages, the characteristics for our 6AC7 would look much the same.

Fig. 15 has curves for various grid voltages, from zero to 6 volts negative. You can see how plate current decreases from one curve to another as the grid is made more negative. It is also apparent, that above some certain plate voltage on each curve, plate voltage changes have very little effect on plate current. Any family of plate characteristics, such as the one illustrated, applies for only one certain screen voltage, usually for a screen voltage commonly employed in practice. Screen voltage has such great effect on pentode performance that a different family of plate characteristics is needed for each screen voltage. On some graphs of plate characteristics there are additional curves showing screen currents.

MUTUAL CHARACTERISTICS. Another way of showing relations between grid voltage and plate current for some particular voltages on plate and screen is illustrated by Fig. 16. Here we have three curves called mutual characteristics. The upper curve applies with no load resistance in the plate circuit. The middle curve shows performance with a 7500-ohm load, and the lower one with a 10,000-ohm plate load. The mutual characteristics of Fig. 16 are not drawn for the 6AC7 pentode with which we have been working, but they illustrate in a general way the form of such curves for any pentode working at certain plate and screen voltages.

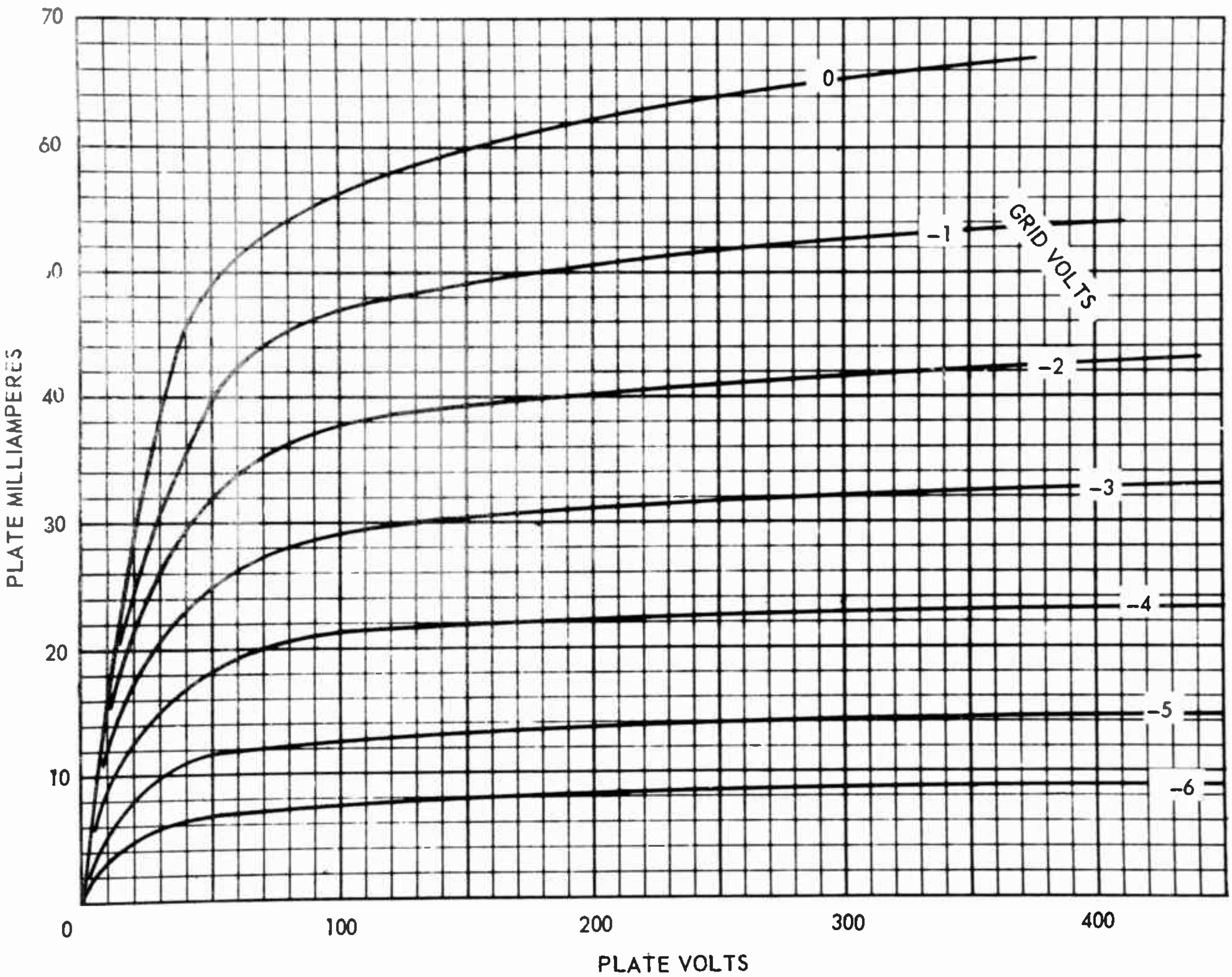


Fig. 11-15. Plate characteristics for a pentode power amplifier.

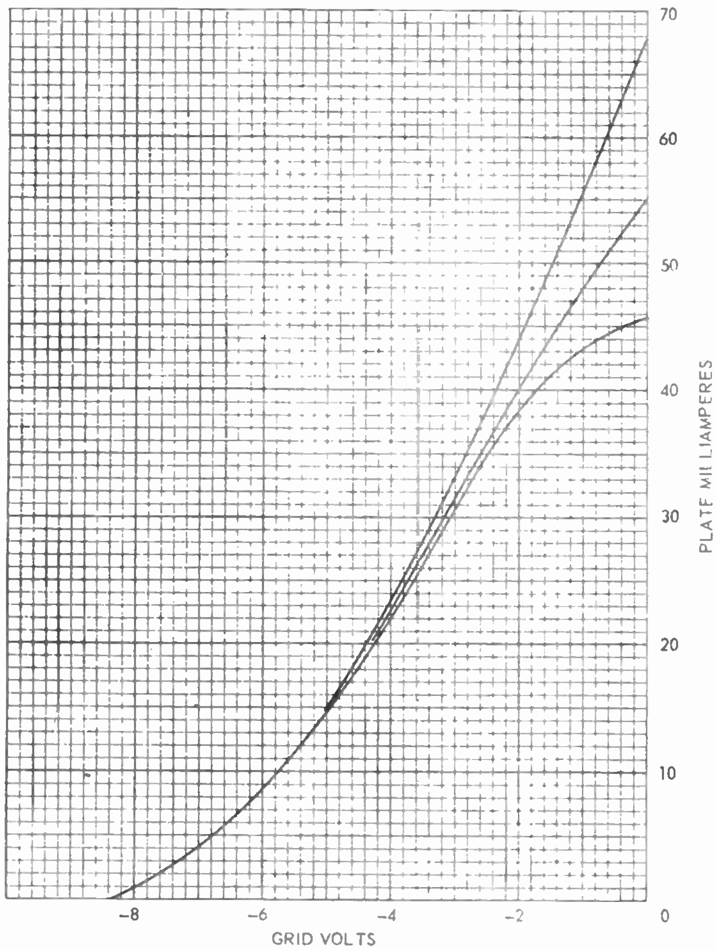


Fig. 11-16. Mutual characteristics of a pentode operated with different loads.

A mutual characteristic sometimes is used to illustrate relations between signal voltage on the grid and current in the plate circuit. Fig. 17 is an example. Alternating signal voltage applied to the grid is shown below the graph. Zero signal voltage is on the line corresponding to the negative grid bias, here 3 volts. The signal swings positive and negative from the bias value, making the grid less negative and more negative during each cycle.

Plate current for each value of grid voltage is spotted along the characteristic curve and lines are carried out toward the right for drawing a representation of the alternating plate current which results from the alternating grid voltage. Although positive and negative swings of grid voltage are equal, the plate current swings to a greater positive value than negative value. This is because negative grid swings bring operation down onto the more sharply curved portion of the characteristic. This could be avoided by using a less negative bias to bring the entire operation onto a nearly straight portion of the curve.

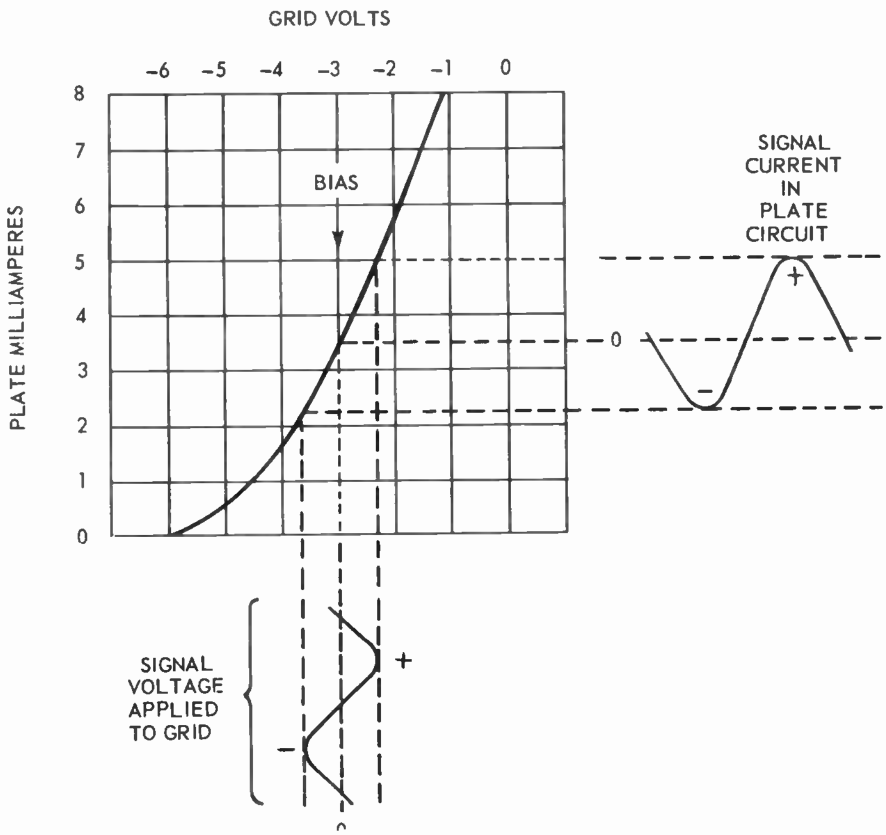


Fig. 11-17. Using a mutual characteristic to show the transfer of a signal from the grid to t... te circuit.

PENTODE AMPLIFICATION. Now we shall go back to the 6AC7 pentode and watch it amplify a signa... voltage. In Fig. 18 the *Input* meter shows a signal voltage 0.2 volt positive. Bias is about 1.0 volt negative, and grid voltage is the difference between bias and signal, about 0.8 volt negative. For amplification we must have a plate load resistance, so a resistor of 40,000 ohms has been connected between the terminals above the *Output* meter. In Fig. 19 the input signal has swung to 0.2 volt negative, bias is 1.0 volt negative, and grid voltage is the sum of negative bias and signal voltages, or is 1.2 volts negative. Here are the comparative readings.

Amplification With 40,000-ohm Load

Figure	Esig	Ec	Eg	Ep	Eo	Escr	Ip
18	+0.2	-1.0	-0.8	90	150	75	4.0
19	-0.2	-1.0	-1.2	135	105	75	2.4
Change	0.4		0.4	45	45		

Voltage gain. 45 (Eo) ÷ 0.4 (Esig) = 112.5

Total change of signal voltage is 0.4 volt, from 0.2 volt positive to 0.2 volt negative. Total change of output voltage (Eo) across the load resistor is shown by the *Output* meter to be 45 volts, from 150 volts down to 105 volts. Dividing the change of output voltage by the change of input voltage shows the voltage gain to be 112.5 times for the particular operating conditions here used.

Voltage from the d-c power supply must equal the sum of voltage drop in the tube, which is plate voltage, and voltage drop in the load resistor, which is output voltage. Power supply voltage was held at 240 volts, as you will find by adding the plate and output voltages during each test.

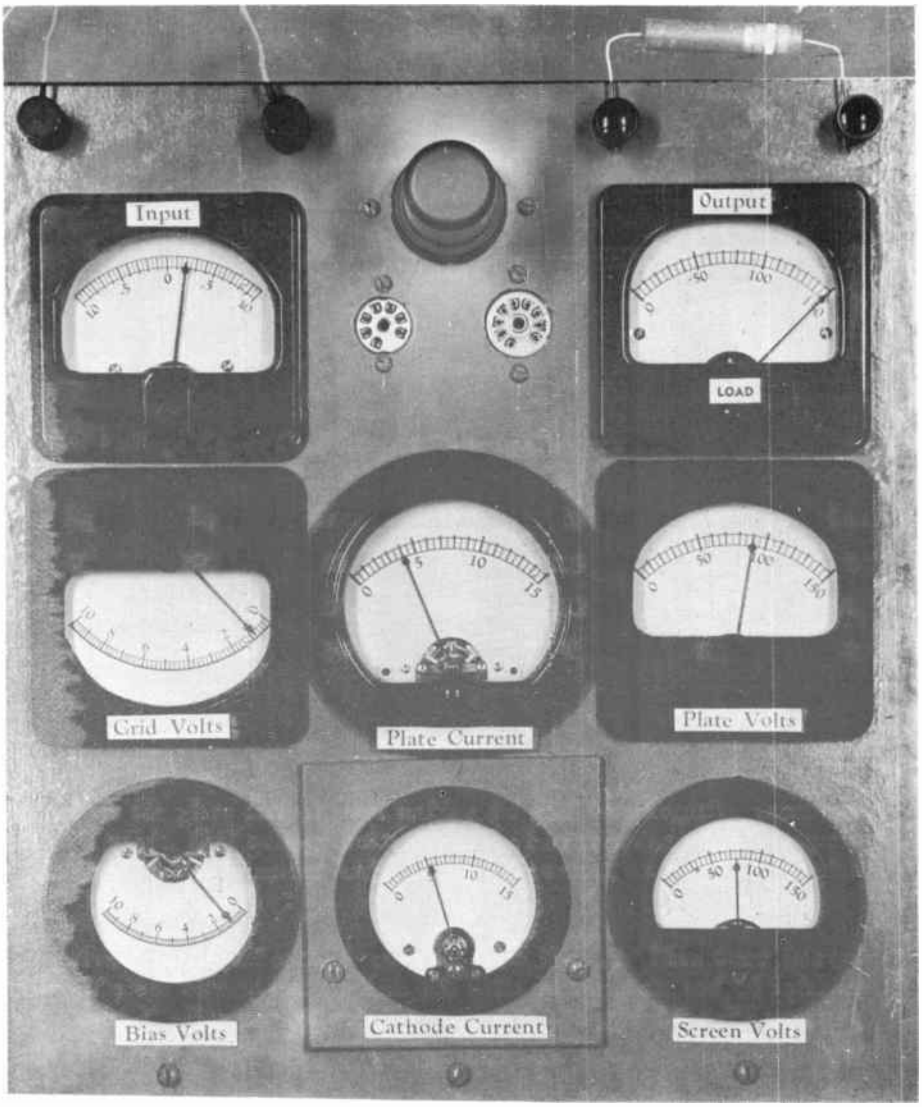


Fig. 11-18. The first step in amplifying a signal of 0.2 volt at the grid.

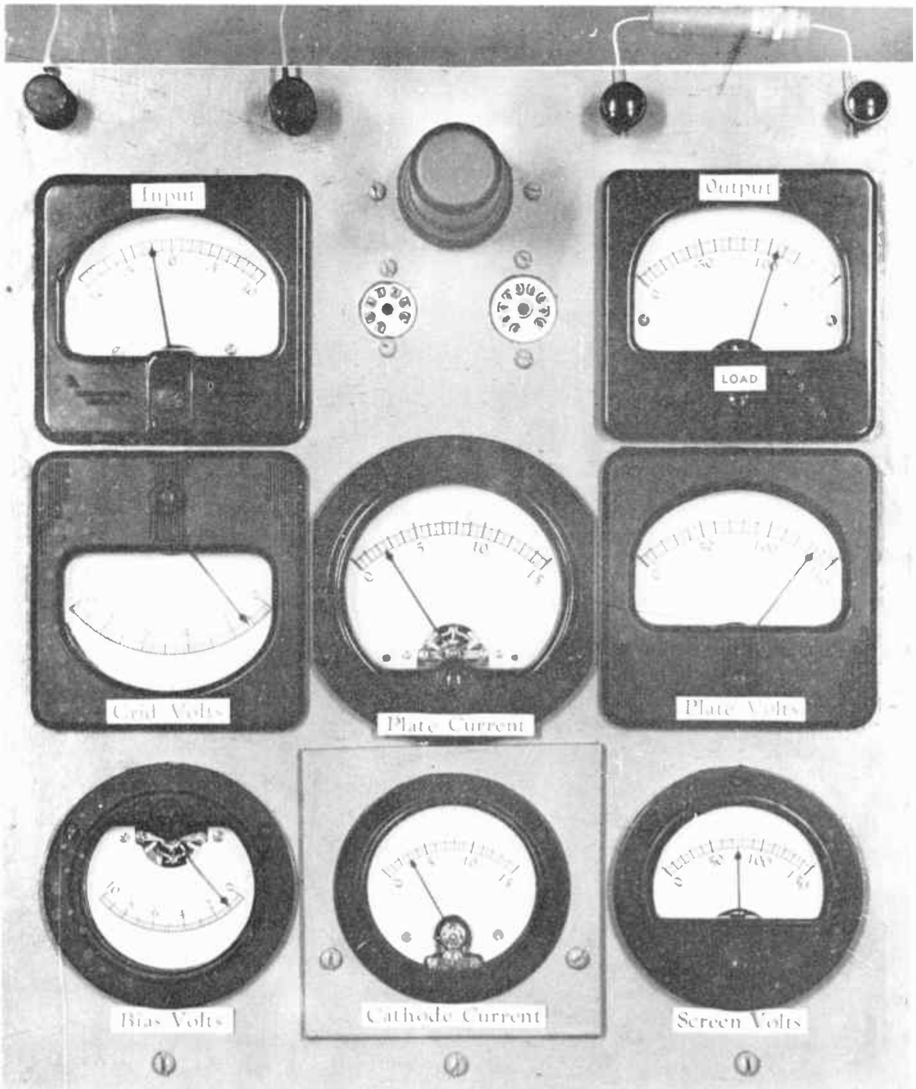


Fig. 11-19. Here the signal voltage has swung to 0.2 volt negative.

When the grid is made more negative the plate voltage increases or becomes more positive, changing from 90 to 135 volts. This illustrates the inversion of voltage changes between grid and plate of any tube; the plate goes positive when the grid goes negative, and vice versa.

With constant voltage from the d-c power supply, load voltage must increase when plate voltage drops, and load voltage must drop when plate voltage rises.

Doubtless you have noticed that the listed values of plate current flowing in a 40,000-ohm load resistor would not cause the voltage drops indicated by the *Output* meter. This is because the meters in the plate

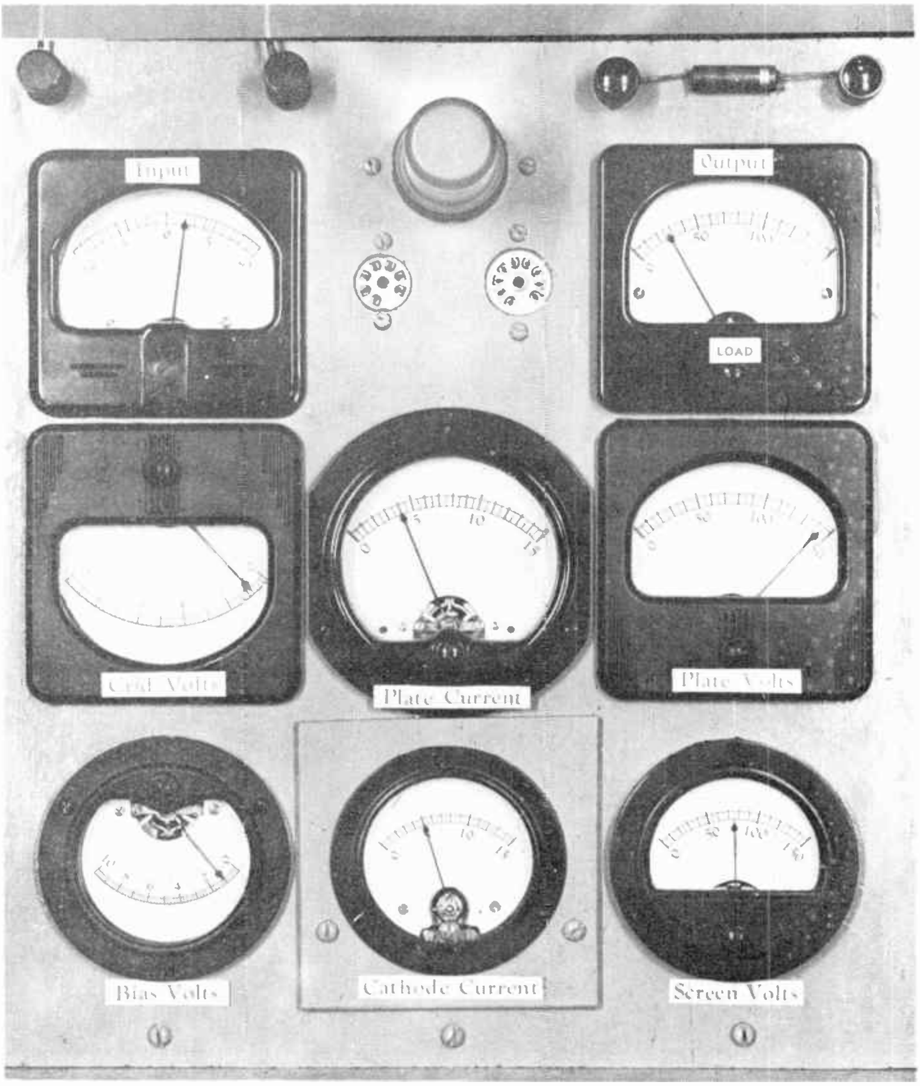


Fig. 11-20. With less load resistance there is less voltage across the load.

circuit are taking small currents to move their pointers, in addition to the true tube cuirent. This effect would not exist were the meters removed, as for normal operation, and, anyway, it is too slight to affect the truth of conclusions being drawn from the various tests.

Now let's check the effect on amplification of changing the load resistance. In Figs. 20 and 21 the load resistor has been changed to one of about 7700 ohms. The input, bias, and grid voltages are the same as in Figs. 18 and 19, and screen voltage is still 75 volts. Check these comparative readings against the previous ones.

Amplification With 7,700-ohm Load

Figure	Esig	Ec	Eg	Ep	Eo	Escr	Ip
20	+0.2	-1.0	-0.8	144	27	75	4.0
21	-0.2	-1.0	-1.2	151	20	75	2.3
Change	0.4		0.4	7	7		

Voltage gain. 7 (Ep) ÷ 0.4 (Esig) = 17.5

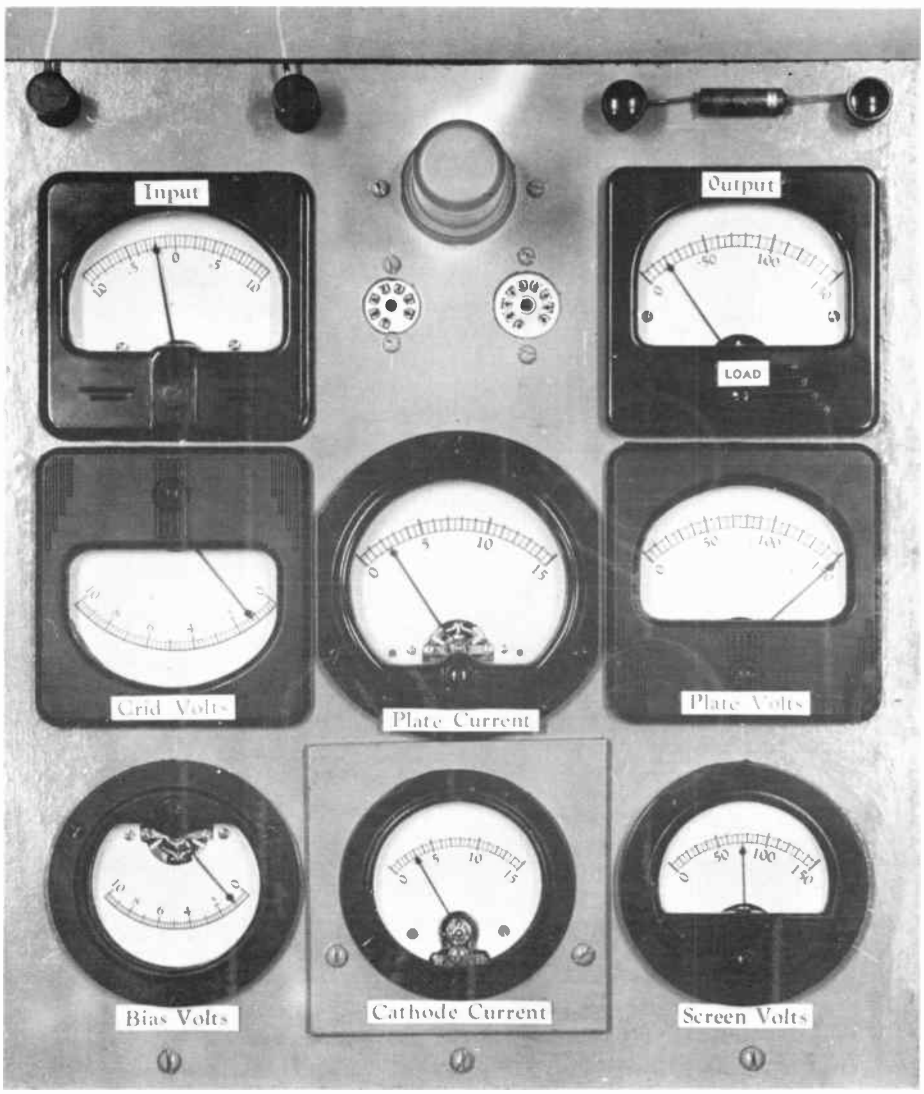


Fig. 11-21. The change of output voltage is small with the small load resistance.

This brings out the all-important fact that voltage gain depends to a large extent on load resistance in the plate circuit. Reducing the load to about 20 per cent of its former value has dropped the gain to about 15 per cent of its former value. Plate current has changed by approximately the same amount with both load resistances. This shows that transconductance of the tube has undergone little change, since transconductance is measured as microamperes of plate current change per volt of grid voltage change.

With any given transconductance, the greater gain is always realized with a greater plate load resistance. An optimum value of resistance is reached when the plate voltage for a given tube is correct, and the plate supply has not been increased above normal for the particular receiver.

It would not be possible economically to obtain the calculated or the theoretical gain of a pentode tube, because of the excessive voltage drop across the plate load resistance. The internal tube resistance of a 6AC7, is 1 megohm when its plate voltage is 300 volts. To obtain the theoretical gain, the plate load resistance must be greater than the internal tube resistance. With a plate load resistance larger than 1 megohm, the plate supply voltage required would be comparable to that of a small transmitter.

The decreased working gain is not a serious loss in obtaining the required overall gain of a receiver. Two stages of amplification using type 6AC7 tubes, and with constants as shown in Figs. 20 and 21 would give an overall gain of 17.5 multiplied by 17.5 or a total gain of approximately 306.

If two tubes having different transconductances are operated with equal plate load resistances, there will be greater gain from the tube of greater transconductance. Neither high transconductance nor high load

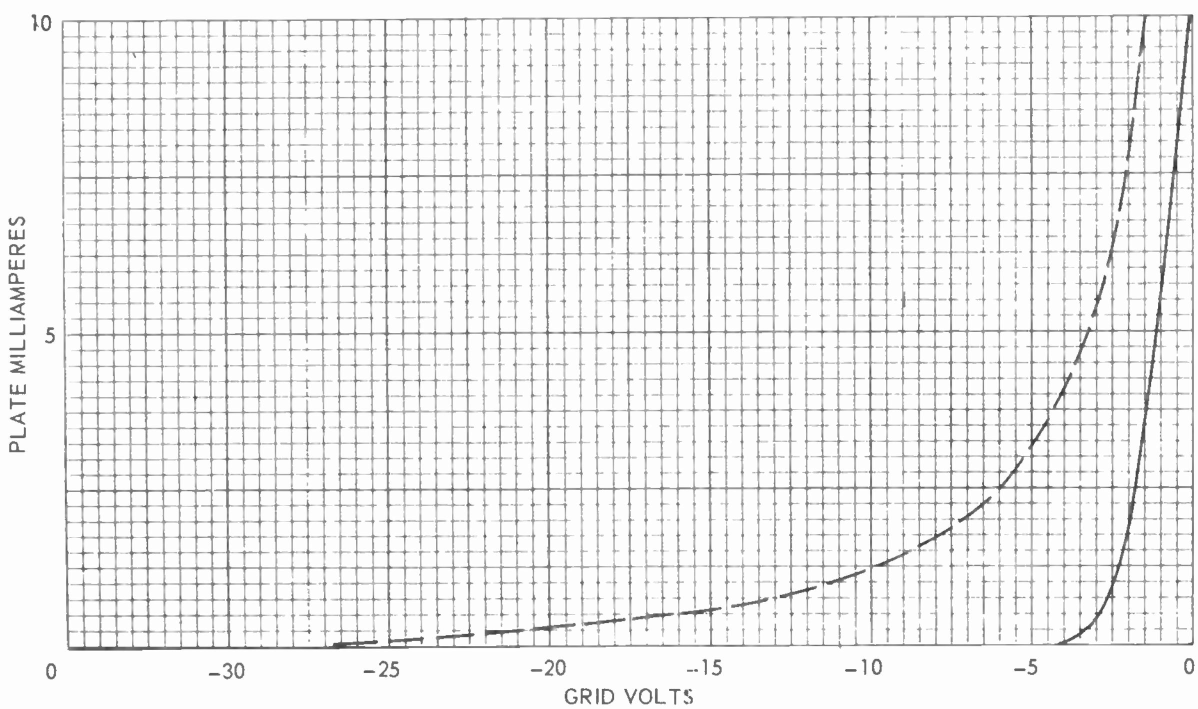


Fig. 11-22. Mutual characteristics for a remote cutoff pentode (broken line) and for a sharp cutoff pentode (full line).

resistance will, by themselves, give high voltage gain. There must be both high transconductance and high load resistance; gain is approximately proportional to their product.

SHARP AND REMOTE CUTOFFS. Pentodes designed for voltage amplification are of two general types, sharp cutoff and remote cutoff. The difference is illustrated by the two mutual characteristic curves of Fig. 11-22. The solid line curve is a mutual characteristic for a sharp cutoff pentode, and the broken line curve is a mutual characteristic for a remote cutoff pentode. Both curves were taken with 250 plate volts and 100 screen volts.

As grid voltage is made more negative the plate current of the sharp cutoff pentode drops rapidly toward zero and, for the tube illustrated, there is plate current cutoff when the grid becomes about 4 volts negative. Plate current of the remote cutoff tube drops rapidly as the grid is made negative through the first few volts, but then drops more and more slowly. Plate current cutoff does not occur until the grid is about 27 volts negative.

The remote cutoff feature is obtained, as shown in Fig. 11-23, by spacing the wires of the grid farther apart at the middle than toward top and bottom. When the grid is only a few volts negative the negative

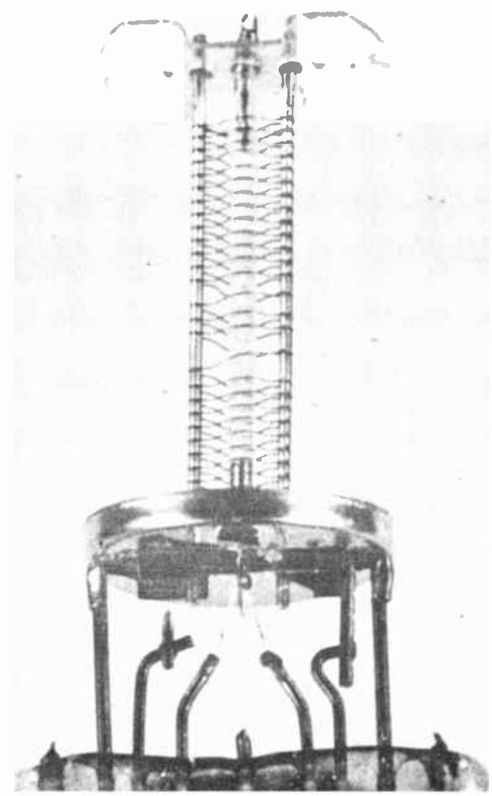


Fig. 11-23. Turns of the remote cutoff grid are widely spaced near the center.

field around the grid wires can shut off all electron flow only where these wires are closetogether, at top and bottom. The negative field cannot extend all the way between the widely spaced center turns until the grid is made highly negative, and some electron flow will continue through the center of the grid until this condition is reached. Electron flow or plate current drops rapidly as the top and bottom parts of the grid cut off the flow through them, but only slowly as increasing negative field strength gradually cuts off flow through the widely spaced center turns.

BEAM POWER TUBES. In a modification of the pentode which is called a beam power tube there is no suppressor, yet secondary electrons are forced back to the plate and kept from reaching the screen. Before explaining the action of such a tube, let's look at its construction. The elements are shown by Fig. 11-24. Inside there is the cathode, and around the cathode are first the grid and then the screen, both formed of the usual spiral wires. Around the outside is the plate, but between screen and plate, where a pentode suppressor would be located, are two "beam forming electrodes". These electrodes are not spirals of wire, they are of continuous metal sheets.

At the left in Fig. 11-25 is a beam power tube from which the plate has been removed. You can see the

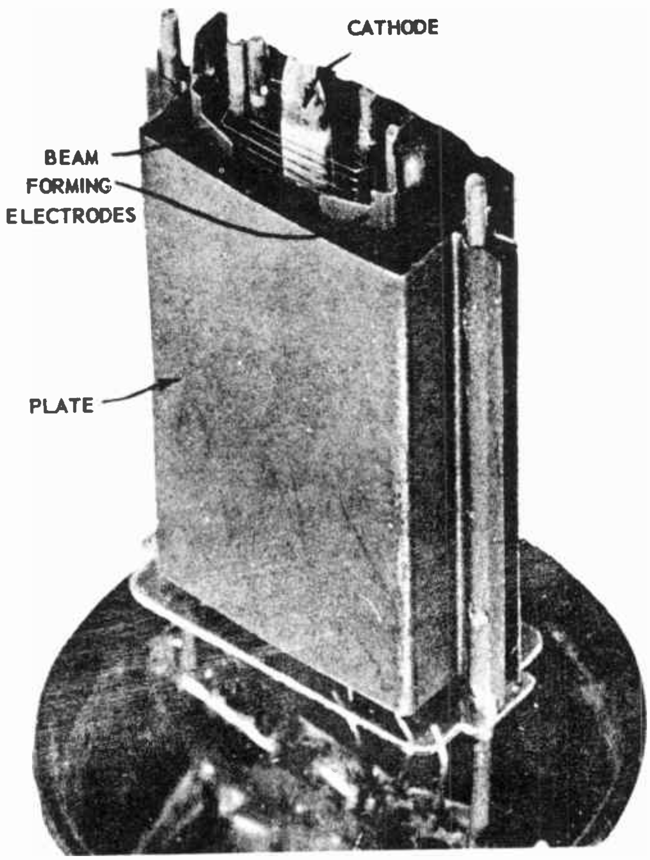
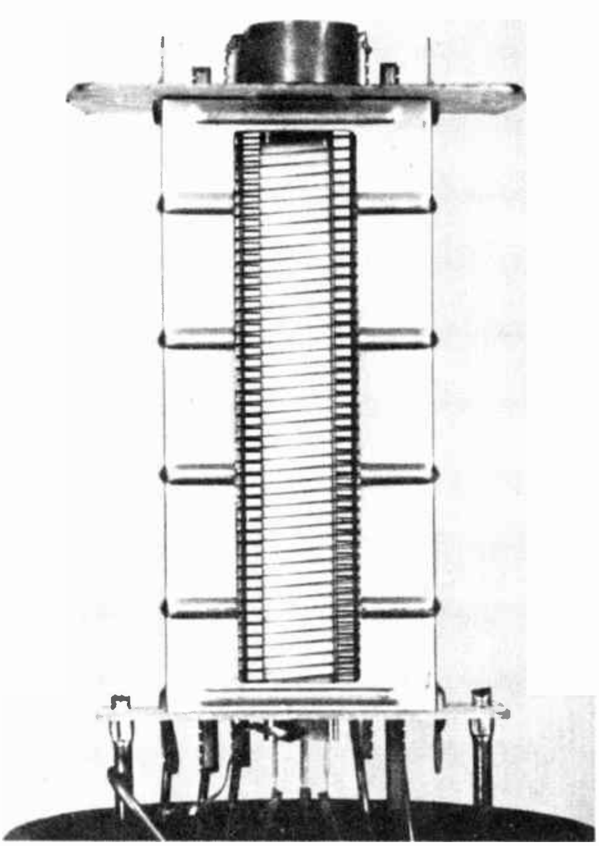


Fig. 11-24. The elements of a beam power tube.



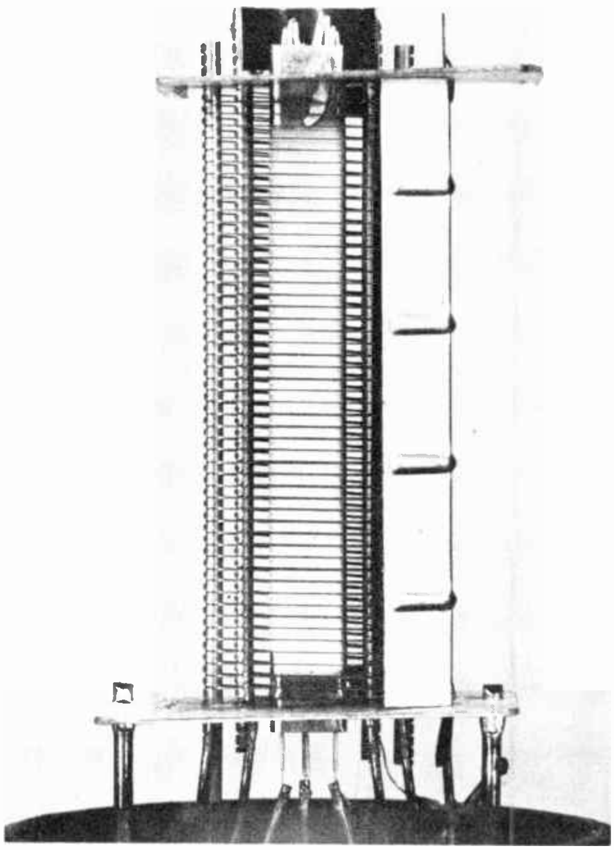


Fig. 11-25. Both beam forming electrodes are shown at the left. One has been removed in the picture at the right.

screen, grid, and cathode through the center opening between the two beam forming electrodes on either side. In the picture at the right one of the beam electrodes has been removed, showing that the screen and grid, with their supports, are much the same as in other tubes. The cathode is large, because all beam tubes are power types capable of handling large plate currents and large swings of plate current.

The beam electrodes are internally connected to the cathode, which places them at a potential highly negative with reference to the plate and screen. Negative electrons are repelled by the negative beam electrodes, forcing the electron flow between cathode and plate to travel in two concentrated or confined beams through the spaces between the opposite electrodes.

Screen voltage is made about the same as plate voltage, when there is no signal. When electrons which have been drawn from the cathode through the screen reach the space between screen and plate the electrons slow down because there isn't enough difference between plate and screen voltages to maintain the speed of these electrons. These slower traveling electrons accumulate in great density at a point between the screen and the plate, just as people who have been racing along would jam together if they all had to slow down at one spot.

The accumulation of negative electrons is a negative charge at the same place where a suppressor would be in an ordinary pentode. This charge repels secondary electrons trying to go back from plate to screen, and we have suppression of secondary emission effect.

Confining the entire electron flow from cathode to plate into two beams makes this flow quite dense all the way. The tendency would be for many electrons to enter the highly positive screen and form an unduly large screen current. This is prevented in a manner clearly evident from the right-hand picture of Fig. 25. The negative grid and the positive screen have the same number of turns, equally spaced, and with each screen turn directly outside of a grid turn. Electrons from the cathode are repelled by the turns of the negative grid, and thus are diverted away from the turns of the screen. Each negative grid wire forms a sort of electron shadow, within which is the adjacent screen wire.

PENTODES OPERATED AS TRIODES. A pentode tube may be operated as a triode type by connecting the screen and plate of the pentode together at the socket terminals. This is done with all types of pentodes; with sharp or remote cutoff voltage amplifiers, also with power amplifiers. With the triode connection the plate resistance is much less than with the pentode connection. Plate voltage and grid bias usually are chosen to allow about the same plate current with either connection. Transconductance ordinarily is about the same with the tube working either way. When a pentode is connected as a triode its plate characteristics are of the same general form as those for any other triode.

WHERE TUBE TYPES ARE USED. We have examined the performance of diodes, triodes, tetrodes, pentodes, and beam power tubes. All except the tetrode are to be found in almost any television receiver. The tetrode long ago was replaced by the pentode.

In Fig. 11-26 are shown symbols for eighteen tubes, exclusive of the picture tube, as found in a fairly typical television receiver. Beginning at the antenna we have the following types. First is the r-f amplifier, a twin triode. Undesirable feedback that could occur in R. F. amplifier triodes working at high frequencies is avoided here by special circuit design.

Next comes another twin triode, with one section used as the r-f oscillator and the other section used as the mixer tube. Note that here we have two separate cathodes, because the two sections perform functions which must be kept distinct. In the r-f amplifier there is only a single cathode for the two sections, with only the grids and plates separate in each section.

Following the mixer come three voltage amplifying pentodes acting as i-f amplifiers for video and sound signals. Following the third i-f amplifier is a twin diode with one section used as a video detector. In this particular receiver the other section is unused, although it might be used for an automatic gain control tube, for a restorer, or for some other purpose. Following the video detector is a pentode video amplifier.

In the sound section we have first a pentode i-f amplifier. Then comes a single tube containing two separate diodes which are used as the sound detector or sound demodulator, and also a triode which is used as a voltage amplifier for audio frequencies. Next we have a beam power tube delivering its audio signal to the speaker.

Between the video amplifier and the sweep section of the receiver is a pentode used as a sync separator. Signals from this separator go to the horizontal and vertical sweep oscillators, both of which are twin triodes with separate cathodes. The horizontal sweep oscillator operates into a beam power tube which

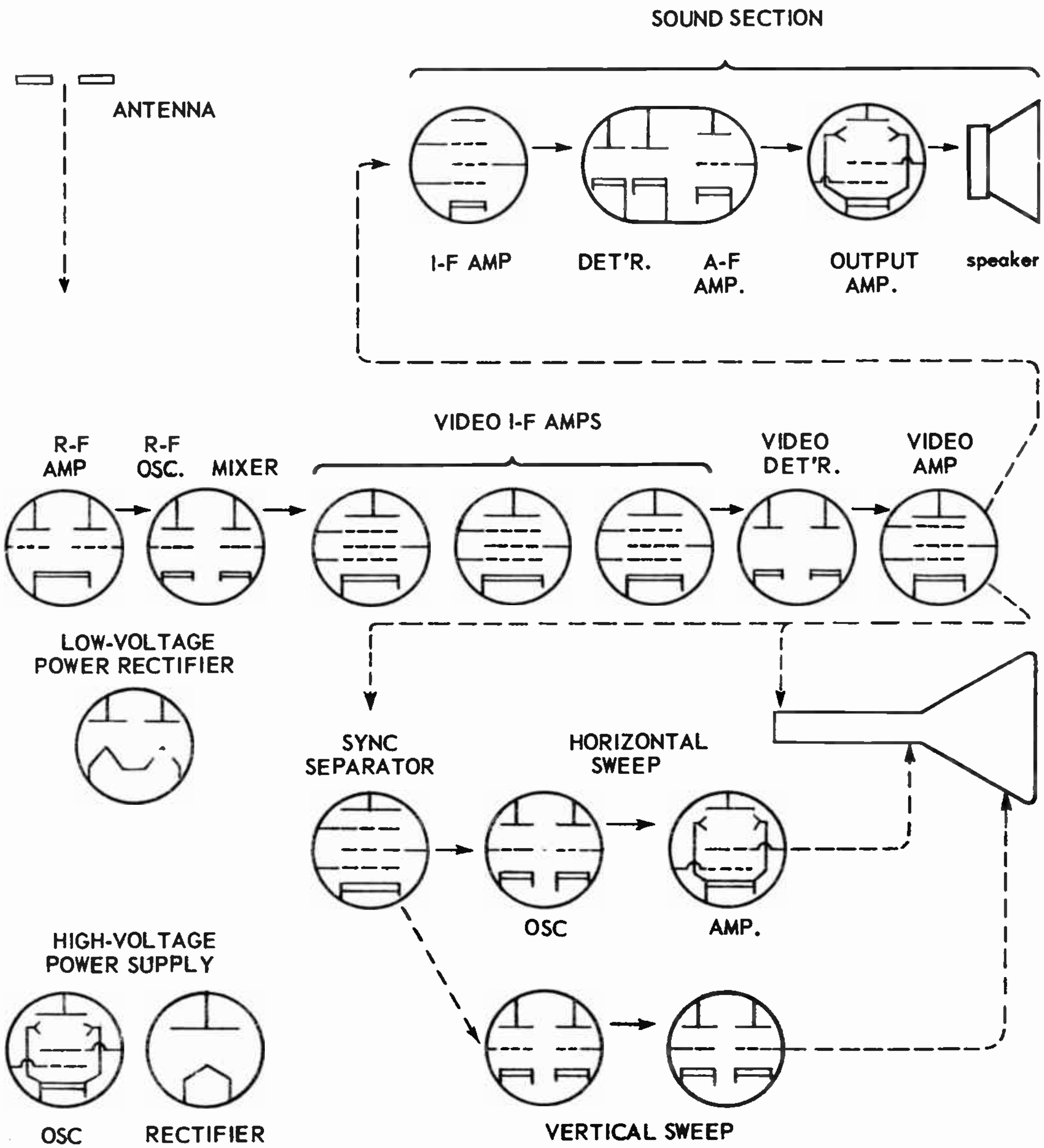


Fig. 11-26. The various types of tubes used in the circuits of a typical televisiou receiver.

is the horizontal sweep amplifier. The vertical sweep oscillator operates into another twin triode used as the vertical sweep amplifier.

At the lower left in Fig. 11-26 are the power supplies. In the low-voltage supply we have a rectifier consisting of a twin diode with two plates and a single filament-cathode. In the high-voltage power supply there is a beam power tube used as a high-frequency oscillator, also a single diode with filament-heater which is the high-voltage rectifier.

TELEVISION

LESSON NO. 12 CAPACITORS FOR TELEVISION AND RADIO

Training Your Mind
C. T. I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind

C.T.I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

LESSON NO. 12

CAPACITORS FOR TELEVISION AND RADIO

IT IS TIME TO STUDY CAPACITORS

We have just finished learning a good deal about electronic tubes. Among other things we have learned how a tube operates as an amplifier, how a signal voltage applied to the grid reappears in the plate circuit much strengthened. Now it is in order to learn how signals are brought to the grid, and how amplified signals are taken from the plate circuit and carried along toward the picture tube or speaker.

Were we to trace the paths through which signal voltages come to grid circuits and leave plate circuits, most of these paths would be found to include resistors. Many would include inductors or coils. But every single tube circuit would include capacitors — with which we are not yet familiar. Capacitors do an astonishing variety of jobs in modern television and radio receivers. It is difficult to explain the workings of practical circuits without having to talk about capacitors and how they contribute to performance.

Capacitors help select the channel in which reception is desired. Capacitors carry signals from one tube to another. Capacitors prevent waste of signal strength. Capacitors keep signal voltages from going where they should not go. Capacitors maintain correct grid biases. We might go much further with this recitation of what capacitors do, but would only add to the reasons why the time has come to get acquainted with these remarkable devices.

CAPACITORS FOR TELEVISION AND RADIO

A count of the number of parts of each principal kind in a popular model of television set shows seventeen tubes, forty-three inductors or coils, eighty-five resistors, and seventy-six capacitors. From this it is easy to see why we cannot progress much further without an understanding of capacitors. Inductors and coils may come later, for there are many circuits without these parts, but we must learn about capacitors right now.

Fig. 12-1 is a circuit diagram for the tuner of a television receiver. There is a pentode r-f amplifier tube, a twin triode used as r-f oscillator and mixer, and all the connections from the antenna terminals at the upper left through to the mixer output line at the upper right. In this tuner section are two tubes, seven inductors or coils, seven resistors, and twelve capacitors.

The capacitors are shown by their usual symbol, a straight line and a curved line with a space between them. These are the symbols marked with letters in the diagram. When an arrow is drawn across the two lines of the symbol it indicates that the capacitor is adjustable, that its value or electrical size may be altered by turning a shaft or screw. A symbol with no arrow indicates a fixed capacitor, one whose value cannot be altered.

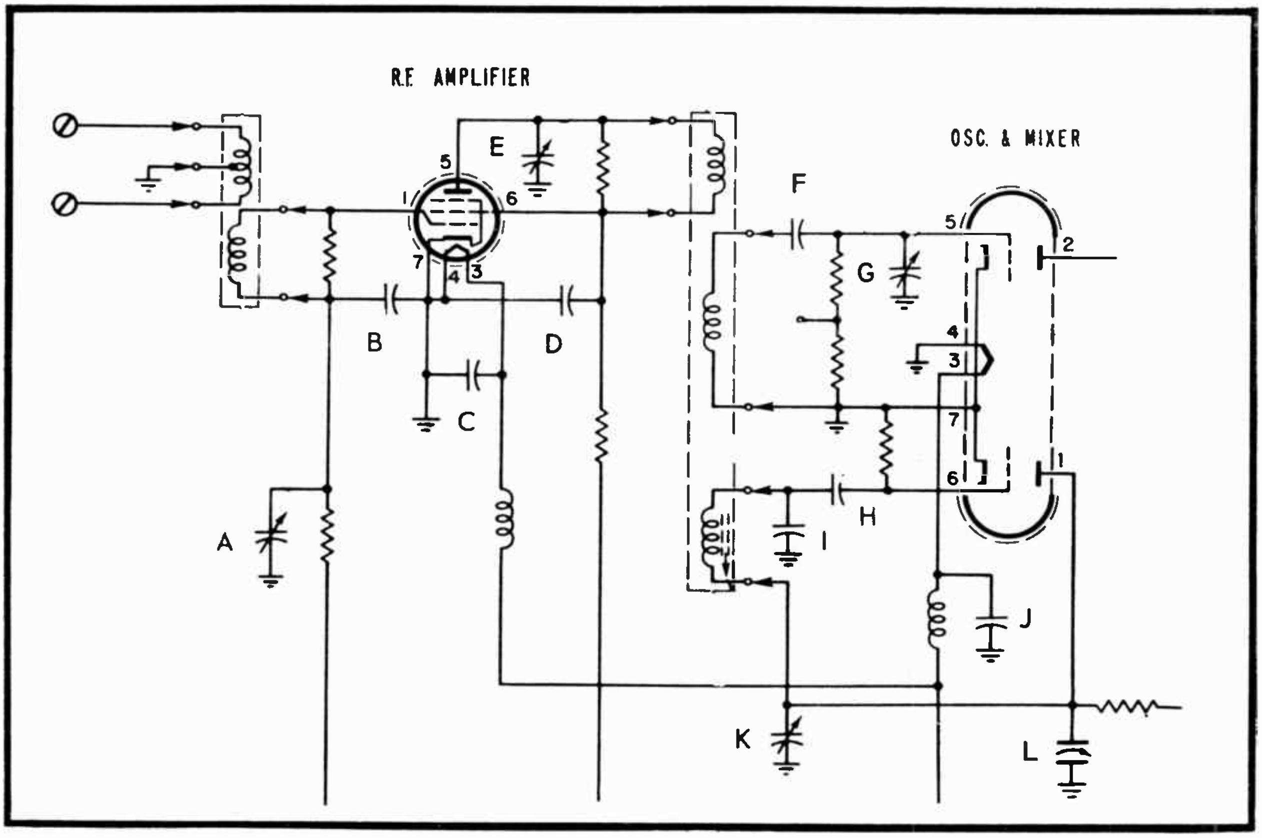


Fig. 12-1. The parts of a television tuner, as shown by symbols.

Fig. 12-2 is a picture of an adjustable capacitor such as used for tuning of superheterodyne sound receivers. In this unit there really are two capacitors adjusted simultaneously by the one shaft. One capacitor is for the r-f amplifier, the other is for the r-f oscillator. Fig. 12-3 shows several fixed capacitors of different sizes. Circuit connections are made to the wire "pigtails" which extend out from opposite ends of the capacitors.

The capacitor symbols are appropriate, because a capacitor consists of nothing more than two conductors, or two sets of conductors, separated by insulation. The conductors of a capacitor are called its plates. The insulation used to separate the plates is called the dielectric of the capacitor. Dielectrics of capacitors used in radio and television may be air, paper, mica, ceramic compounds, or thin films of oxides.

As we shall learn shortly, the usefulness of a capacitor is due to the fact that its plates may be charged and discharged. To one plate of a capacitor we may add electrons and give this plate a negative charge, at the same time taking away from the other plate an equal quantity of electrons to give the other plate a positive charge. The dielectric between the plates prevents electrons from flowing from plate to plate within the capacitor, and thus keeps the charges separated. The plates may be charged and discharged only through external connections.

Insulating materials used for dielectrics are no different from the same materials when used as insulation for confining electron flow to certain paths. The materials are called dielectrics when used between the plates of a capacitor, or when we are considering the materials as to their behavior when subjected to electrostatic fields such as exist between opposite charges in a capacitor and also at many other places in radio apparatus. The name dielectric merely tells us how or where the insulating material is used, or indicates what properties of the material are important in a particular application.

The devices which we call capacitors formerly were called condensers, and still are called condensers by many people. The name condenser was originally adopted because two plates separated by a dielectric seem to condense or collect electricity within the device. But there are many other kinds of condensers; some for steam and some for other vapors, including condensers used in refrigerators, and there are synchronous electric condensers which are a type of electric motor. If we use the name condenser as meaning a capacitor we really should say electrostatic condenser to distinguish it from all the other kinds. It is easier to use the name capacitor, which has only one meaning.

KINDS OF CAPACITORS. Capacitors may be classified or described in any of several ways. One classification is based on whether the capacitor is fixed or adjustable. Another classification is based on the kind of dielectric in the capacitor. Fig. 12-2 shows an adjustable air-dielectric capacitor. Fig. 12-3 shows

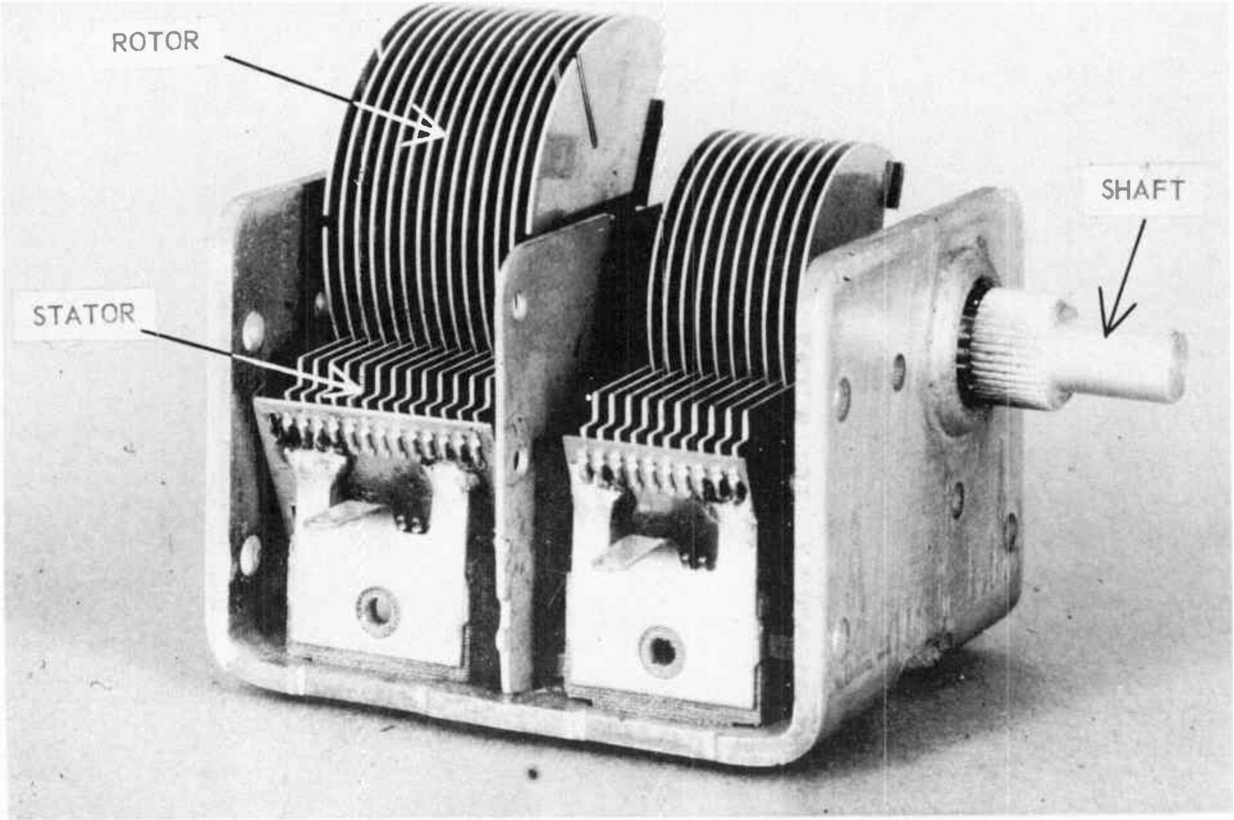


Fig. 12-2. A dual or tandem tuning capacitor for a superheterodyne sound receiver.

Fig. 12-3. Fixed capacitors made with paper dielectric.

fixed paper capacitors, in which the dielectric is thin sheets of paper. We may have also mica capacitors with mica dielectric, ceramic capacitors with ceramic compositions as dielectric, and electrolytic capacitors with oxide films as their dielectric.

Still another method of classification is based on the particular purposes served by the capacitors. According to their uses, the capacitors of Fig. 12-1 would be described as follows.

A, E and G. Adjustable trimmer capacitors which tune the r-f amplifier grid and plate circuits, also the mixer grid circuit, for frequencies of the channel to be received. All these tuning capacitors are service adjustments.

I, K and L. These tune the r-f oscillator circuit to produce the correct intermediate frequency for the i-f amplifier which follows the tuner. Unit I is a fixed capacitor, K is a service adjustment, and L is a "fine tuning" adjustment used by the operator of the receiver.

B. A bypass capacitor to complete a path for high-frequency signal voltages from grid to cathode of the r-f amplifier without forcing these voltages to act through other parts of the receiver.

H. A blocking capacitor to prevent the high positive voltage in the oscillator plate circuit from reaching the oscillator grid.

F. A coupling and blocking capacitor to carry signal voltage from the plate of the r-f amplifier to the grid of the mixer while preventing the positive voltage at the plate from reaching the grid.

D. A decoupling capacitor which lessens signal feedback through connections running from the same power supply to more than one tube.

C and J. Decoupling capacitors which prevent signal transfer from tube to tube through heater connections.

Evidently a capacitor is capable of doing many different things. In the preceding list are capacitors for tuning (A-E-G-I-K-L), others for bypassing a signal around parts in which it should not flow (B), for blocking a positive voltage so that it does not reach a negative grid (F-H), for coupling a signal from one circuit into another (F), and for decoupling to prevent signal feedback from one circuit to another (C-D-J). Even this imposing list does not include all the uses for capacitors in radio and television. Later we shall investigate highly important filter capacitors which help separate one frequency from others, and other capacitors for such special purposes as padding and peaking.

The capacitors shown in a wiring diagram may be represented by any of a wide variety of symbols. Across the top of Fig. 12-4 are symbols for fixed capacitors. The one most often used in radio and television circuit diagrams is at the left. If a capacitor is of such construction that one side is best adapted for connection to ground, and if this capacitor actually is connected to ground, the curved line is supposed to indicate the grounded side. Over at the right in the top row of symbols are represented multiple capacitors having a single plate for a charge of one polarity and more than one plate for opposite charges.

The lower row of symbols represent adjustable capacitors. Always there is an arrow. If the arrow is centered on one of the plates, that is the movable plate of the capacitor. At the right-hand end of the lower row of symbols is one for a capacitor with a single movable plate and two fixed plates.

The two plates of a capacitor may be of various shapes and sizes. In the adjustable capacitor of Fig. 12-2 there are sets or groups of movable plates which are called rotor plates because they rotate with the shaft

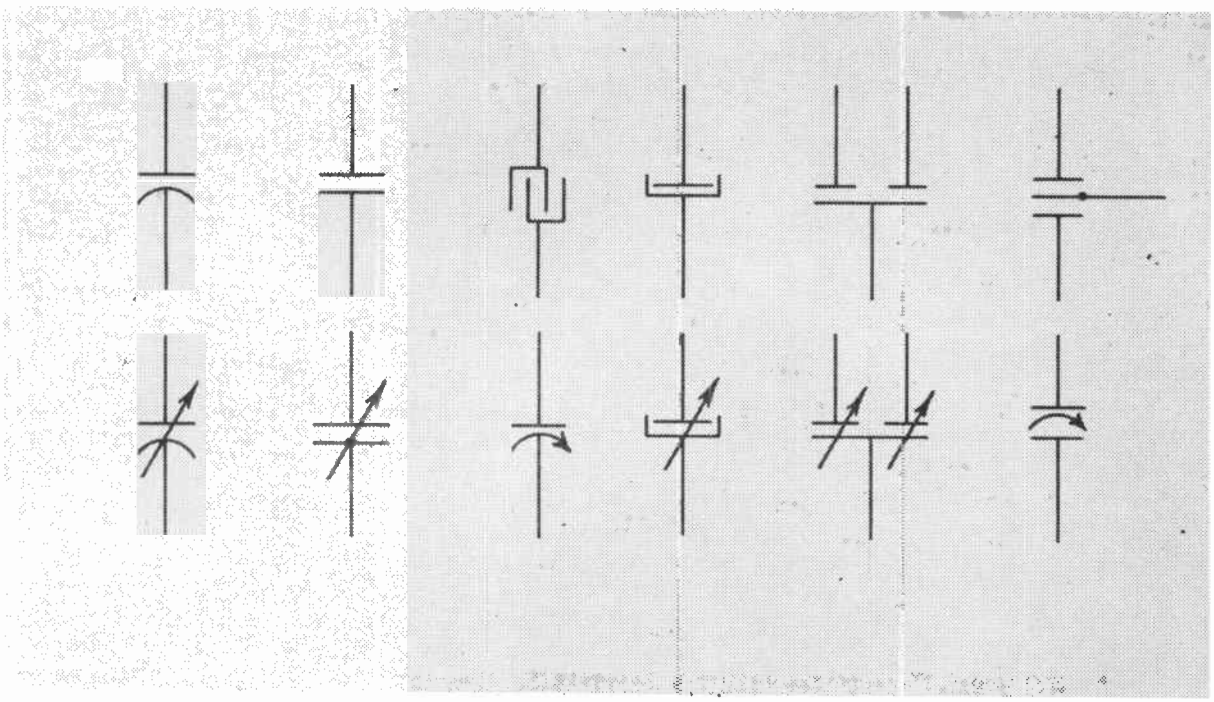


Fig. 12-4. Symbols for fixed capacitors (top row) and for adjustable capacitors (bottom row).

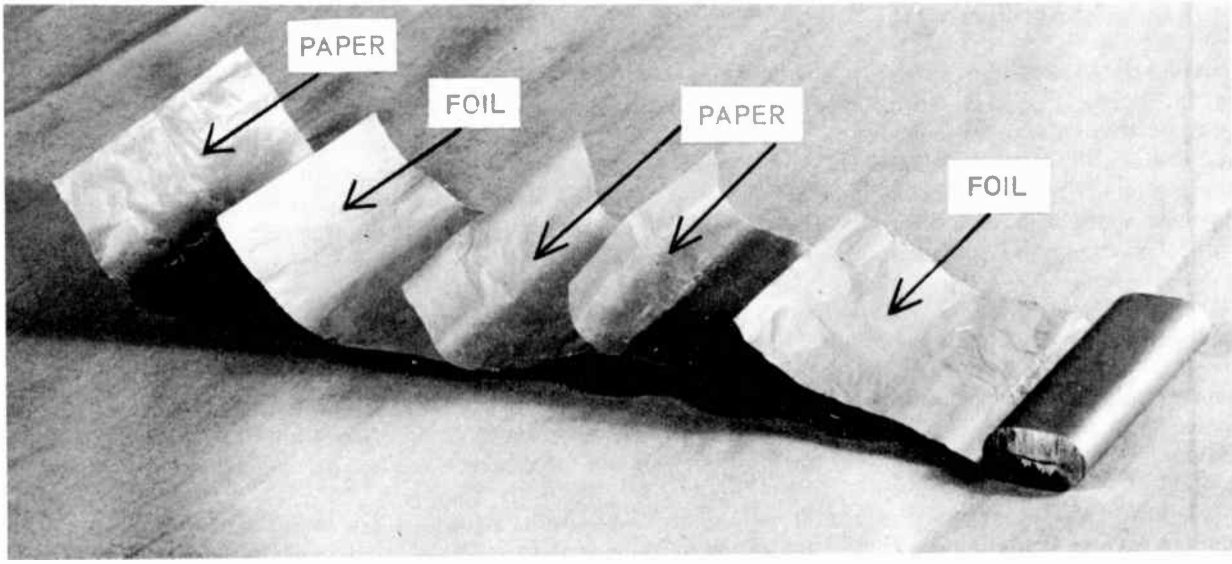


Fig. 12-5. The foil plates and paper dielectric of one kind of fixed capacitor.

through part of a turn. There are sets of stationary plates which are called stator plates. All the rotor plates are joined together to form one electrically continuous conductor, and all the stator plates are joined together to form the other continuous conductor. In the construction pictured, the rotors and stators alternate; first a rotor, then a stator, another rotor, and so on. Between each rotor or stator and plates of the opposite kind on both sides is air which acts as the dielectric in this capacitor.

Fig. 12-5 shows the internal construction of a fixed capacitor which has thin aluminum foil for plates and oiled or waxed paper for the dielectric. The capacitor has been unrolled to show how it is built. Extending out toward the extreme left is one sheet of paper dielectric. Then comes one of the metal foil plates, followed by two more layers of dielectric paper, and finally the second metal foil plate. When the foil and paper are rolled tightly together the foil plates are everywhere separated by the dielectric paper. A terminal, or pigtail connection, is provided for each foil plate, and the whole arrangement is protected with some form of insulating cover which often is a wrapping of heavy paper.

Later we shall look into still other kinds of capacitor construction, but always there will be two electrically continuous conductors, the plates, separated everywhere by the dielectric. Consequently, for discussing the behavior of capacitors we may represent any and all of them by two conductor plates separated by a dielectric.

CHARGING AND DISCHARGING. At the left in Fig. 12-6 a capacitor is represented by two plates with a space between them. In series with the capacitor are a battery, a switch, and a meter of the zero-center type. With the switch open, as shown by full lines, the meter pointer stands at zero because there is no electron flow. In the instant during which you close the switch, to its broken-line position, the pointer of the meter will swing momentarily away from zero and then immediately return to its zero position. There has been a momentary electron flow, in spite of the fact that the capacitor plates are separated by insulation (the dielectric) through which no electrons can flow.

What happens is this: As soon as the switch was closed, free electrons left the negative terminal of the battery, passed through the switch and meter, and entered the left-hand plate of the capacitor. During the same time an equal number of free electrons flowed out of the right-hand plate of the capacitor and entered the positive terminal of the battery.

The left-hand plate of the capacitor acquired a surplus of electrons and became negatively charged. The right-hand plate of the capacitor lost electrons and became positively charged. Electrons flowed into and out of the capacitor plates, but no electrons flowed through the insulating dielectric between the plates and no electrons flowed all the way through the capacitor.

We know that there is a difference of potential between positive and negative electric charges. In only the smallest fraction of a second after closing of the switch, the charges on the two capacitor plates became large enough to have a potential difference equal to that of the battery. As you can see by the positive and negative signs on the capacitor and battery, the potential of the capacitor opposes that of the battery – the two are connected positive to positive and negative to negative. With the two equal potential differences opposing each other there remains no force to continue the flow of electrons, the flow stops, and the pointer of the meter returns to zero.

By connecting the battery to the plates of the capacitor we charged the capacitor. The plate connected to the negative terminal of the battery acquired a negative charge, and the one connected to the positive terminal of the battery acquired a positive charge. Potential difference due to charges on the capacitor plates became equal to the potential difference or the terminal voltage of the battery.

If the switch is opened after the capacitor is charged, the charge must remain in the capacitor or the two charges must remain on the two plates. The charges have to remain because electrons cannot flow through the insulating dielectric, and they cannot flow through the circuit connections which are open at the switch.

While the switch is opened, and the capacitor charged, we shall remove the battery and make a direct connection as at the right in Fig. 12-6. Upon again closing the switch there is a momentary swing of the meter pointer. But now the pointer moves away from zero in a direction the opposite to that in which it moved during charging of the capacitor. Surplus electrons from the negatively charged plate of the capacitor have flowed through the meter and the closed switch over into the positively charged plate. The plate which was negative loses its surplus of electrons, while the one which was positive regains its lost electrons. Both plates become neutral, and the capacitor is discharged.

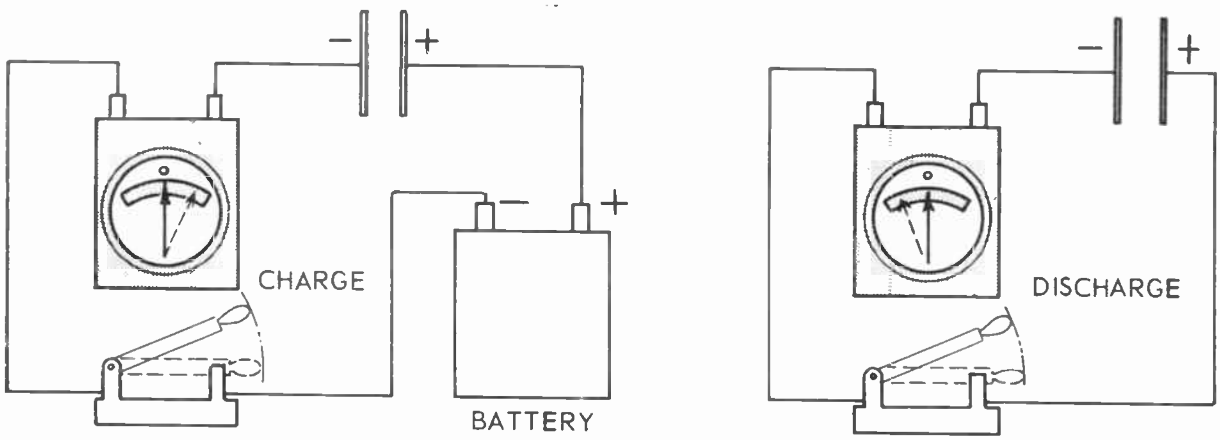


Fig. 12-6. Directions of electron flow are opposite during charge and discharge.

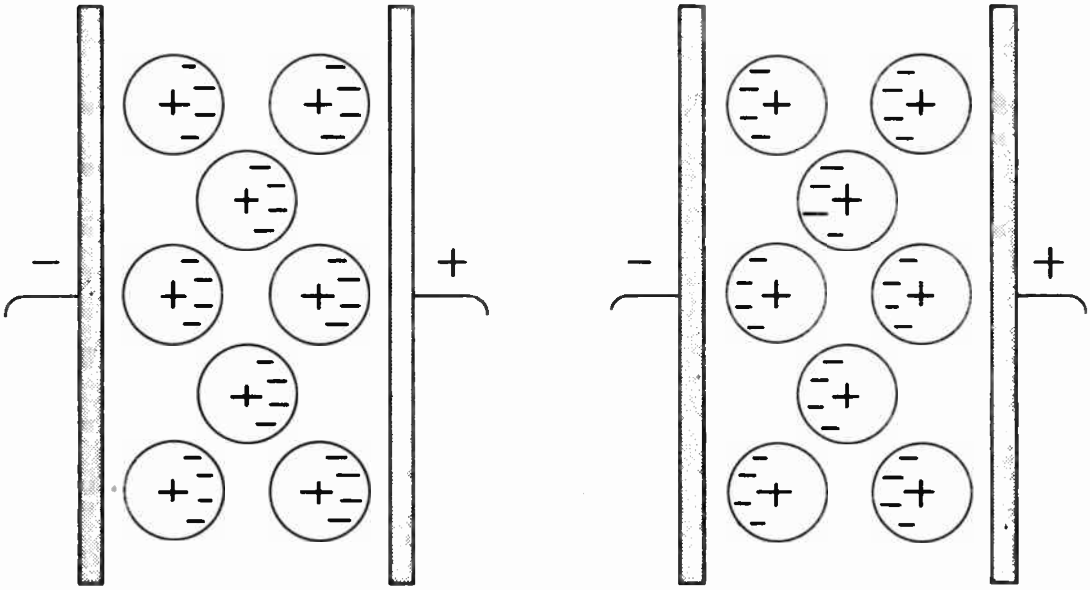


Fig. 12-7. Charges strain the dielectric to cause displacement of electrons and nucleus in atoms.

WHAT HAPPENS IN THE DIELECTRIC. While the capacitor plates retain their charges there are lines of electric force or electrostatic force extending in all directions around the plates, but being most intense in the space between the plates. In this space is a relatively strong electric field. The space between the plates contains whatever substance is acting as dielectric, and so the electric field produced by charges on the plates is acting in the dielectric.

In Fig. 12-7 a few of the atoms in the dielectric are represented by circles with positive signs for the central positive nucleus and negative signs for a few of the electrons in each atom. All the negative electrons in all the atoms of the dielectric are repelled by the negative charge of one capacitor plate while being attracted by the positive charge of the opposite plate. Free electrons will shift away from the negative plate and toward the positive plate. But in addition to this easy movement of the free electrons all the other electrons will undergo some slight displacement in their atoms. These latter are the bound electrons, which do not leave their atoms but do move within the atoms. When the polarity of the capacitor plates is reversed, as between left and right in Fig. 12-7, all the electrons will shift in a reversed direction — always trying to follow the plate attraction and repulsion.

The positive nucleus of every atom is repelled by the positive plate and attracted by the negative plate. Consequently, there is a slight shifting of all the nucleuses in a direction opposite to that of electron shift within the atoms. The atoms become strained and distorted, although they do not move away from their original relative positions in a solid dielectric material. The slight movement of electrons within the dielectric may be called a displacement current, because there is movement of electrons even though they are not free electrons such as cause ordinary conduction currents.

It takes work to shift the electrons and nucleuses within their atoms. If a capacitor is connected into a circuit having an alternating potential, the shifting back and forth will occur during every cycle. At high frequencies the shifting occurs so often that the necessary work produces heat in the dielectric as the electrical energy changes into heat energy. Incidentally, this is the principle utilized in one of the important processes of industrial electronics, the process which is called dielectric heating. By means of this method heat is produced at the same time and uniformly throughout the whole body of any non-conducting substance.

ANOTHER HYDRAULIC ANALOGY. In one of the earlier lessons we used a water circuit to show why electron flow must be the same everywhere in a series circuit, even when the flow is alternating. In that water circuit we had a reciprocating water pump and some piping. Now, in another closed water circuit completely filled with water and containing a reciprocating pump we shall insert a hydraulic arrangement which well illustrates the behavior of a capacitor in an electric circuit. This new water circuit is shown by Fig. 12-8.

At the top are two water chambers separated by a sheet or a diaphragm of thin, flexible rubber. The two water chambers correspond to the two plates of a capacitor. Water which flows into and out of the two chambers corresponds to electrons which flow into and out of the capacitor plates. The rubber diaphragm which may be stretched or strained or deformed corresponds to the dielectric of the capacitor. Down below is the reciprocating water pump which corresponds to a source of alternating potential and current.

When the pump piston is pushed toward the left, at A, water is forced into the left-hand chamber. Comparing the water particles with electrons, we would say that the left-hand chamber becomes negatively charged, for it has an excess of water particles (electrons). An equal quantity of water particles (electrons) must leave the opposite chamber (plate) and go to the pump (source).

It is quite necessary to realize and remember that, so long as the water circuit is completely filled with water, it is impossible for water to enter one chamber without having an exactly equal quantity leave the opposite chamber, and no water can leave one chamber unless an equal quantity enters the opposite chamber.

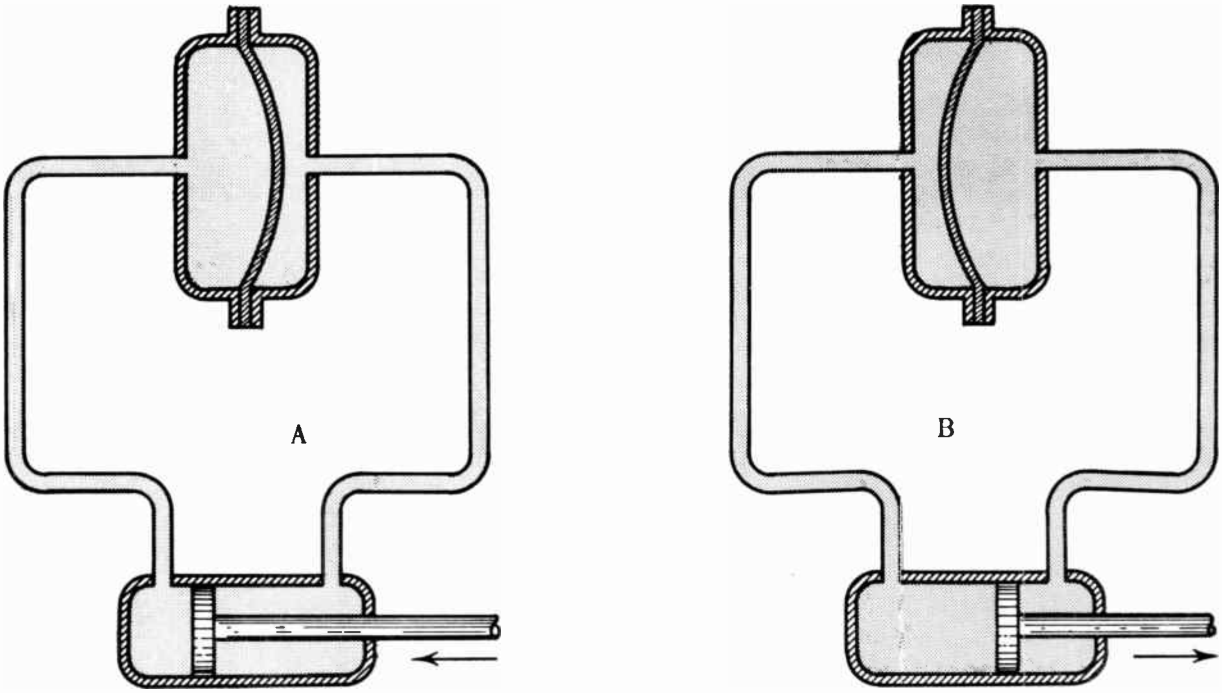


Fig. 12-8. A capacitor acts much like two water chambers with a flexible diaphragm between them.

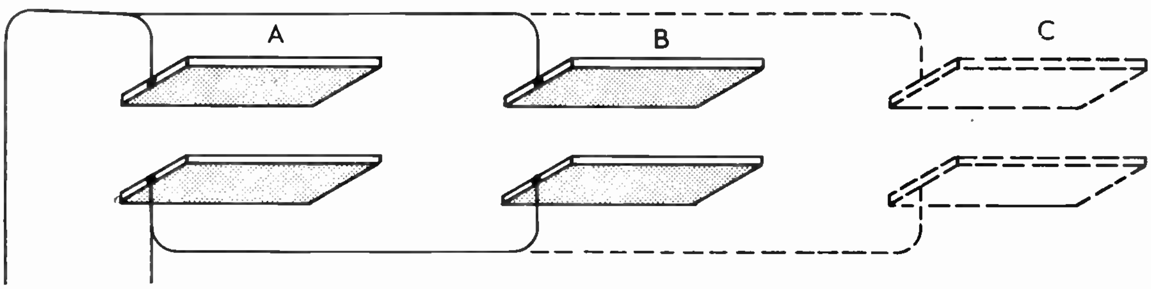


Fig. 12-9. Increasing the plate area increases the charge.

The real importance comes in remembering that charge and discharge must be equal in the two plates of a capacitor. When electrons flow into one plate from a circuit connected to the other plate through external conductors, just as many electrons must flow out of that other plate and into the circuit conductors.

When the pump piston is moved toward the right, as at B, all the actions are reversed. An excess of water particles (electrons) is forced into the right-hand chamber (plate) which now is given a "negative" charge. An equal quantity of water particles (electrons) must flow out of the left-hand chamber (plate), leaving here a deficiency which, for electrons, is a positive charge.

Were the pump piston moved to either end of its stroke and then released, the stretched diaphragm would return to its central position as water left one chamber and entered the other. Thus our water "capacitor" would discharge. Similarly, if a potential difference which has been maintaining charges in capacitor plates is removed, the capacitor will discharge through any complete conductive circuit which is between the two plates.

Were the pump piston moved rapidly back and forth there would be an alternating flow of water into and out of the two chambers, and the diaphragm would be flexed one way and then the opposite way as the flow reversed. But no water would flow all the way through the water "capacitor", because that would require puncturing the rubber diaphragm.

Similarly, in an electrical capacitor carrying alternating current, no current or no electrons flow all the way through the capacitor. Only were the potential difference applied across the plates and dielectric so great as to cause actual puncture of the insulating dielectric could there be electron flow right through the capacitor. With all potential differences less than that which will puncture the dielectric, the electron flow can be only into and out of the capacitor plates as the dielectric is strained first one way and then the other.

From what has just been said it becomes quite plain that alternating current may flow in a circuit which contains a capacitor or capacitors in series. The electrons surge back and forth in all parts of the circuit, including the capacitor plates, and we have alternating current. It is equally plain that a capacitor will not permit flow of direct current through a circuit with which the capacitor is in series. Direct current means electron flow always in the same direction. A flow in one direction might continue until the series capacitor became charged to a potential difference equal to that causing the flow, but then the flow would stop. It is absolutely essential that you learn to look on capacitors as permitting flow of alternating current, and as preventing flow of direct current other than during the instant of charging the capacitor.

If we wish to have a greater alternating flow of water in the water circuit, and if on the pump piston there can be exerted only some limited amount of pressure, there are several things which may be done. For one thing we might use larger water chambers. Then more water could leave and enter without distending the diaphragm so far and without developing so much diaphragm pressure for opposing the pump pressure. Another way would be to use a more flexible diaphragm, one which would stretch farther each way with a given applied pressure, and thus would allow more water to enter and leave the chambers.

There are corresponding ways of increasing the flow of alternating current in capacitors with any given voltage from the source. Using larger plates or more of them corresponds to using larger water chambers. Using a dielectric material of greater "dielectric constant" corresponds to using a more flexible diaphragm between the water chambers. We shall investigate these electrical methods.

ABILITY TO RECEIVE CHARGES. Increasing the electron flow or the rate of electron flow in a capacitor by using more or larger plates is not difficult to understand. If some certain charging voltage is applied to the pair of plates marked A in Fig. 12-9 some certain quantity of electrons will flow into one plate and out of the other plate. If the same charging voltage is applied at the same time to the pair of plates marked B, and if these plates and their dielectric are exactly like the plates and dielectric at A, the total flow of electrons from and to the charging source will be doubled. Adding a third pair of similar plates at C will allow a total charge and total electron flow three times as great as with only one pair connected to the charging source.

In Fig. 12-9 we have the equivalent of three separate capacitors with one plate of each unit connected to one side of the charging source and with the remaining plate of each unit connected to the other side of the source. The same effects on electron flow could be had by doubling or trebling the size of the plates in a single capacitor.

In Fig. 12-10 we have some multi-plate capacitors. At the left is the type with one positive plate and one negative plate, the type we have already considered. At the center is a type with a single negative plate in between two positive plates. Note that the addition of the one extra plate has produced one additional electrostatic field, because there is such a field in the space between any two charges of opposite polarity. Over at the right there has been added another negative plate outside one of the positive plates. This again increases the number of electric fields by one. The number of fields always is one less than the number of plates.

The only part of the capacitor plates which counts, so far as ability to receive a charge is concerned, is the surface area which is on one side of a field between opposite charges. Another way of saying the same thing is to state that only the plate surface area in contact with the dielectric has any effect on ability to

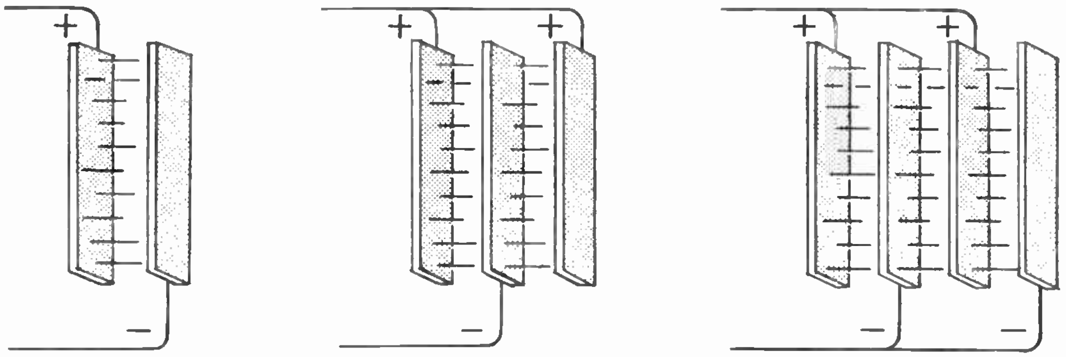


Fig. 12-10. Ability of a capacitor to receive a charge is affected by plate surface in contact with the dielectric, or on opposite sides of electric fields.

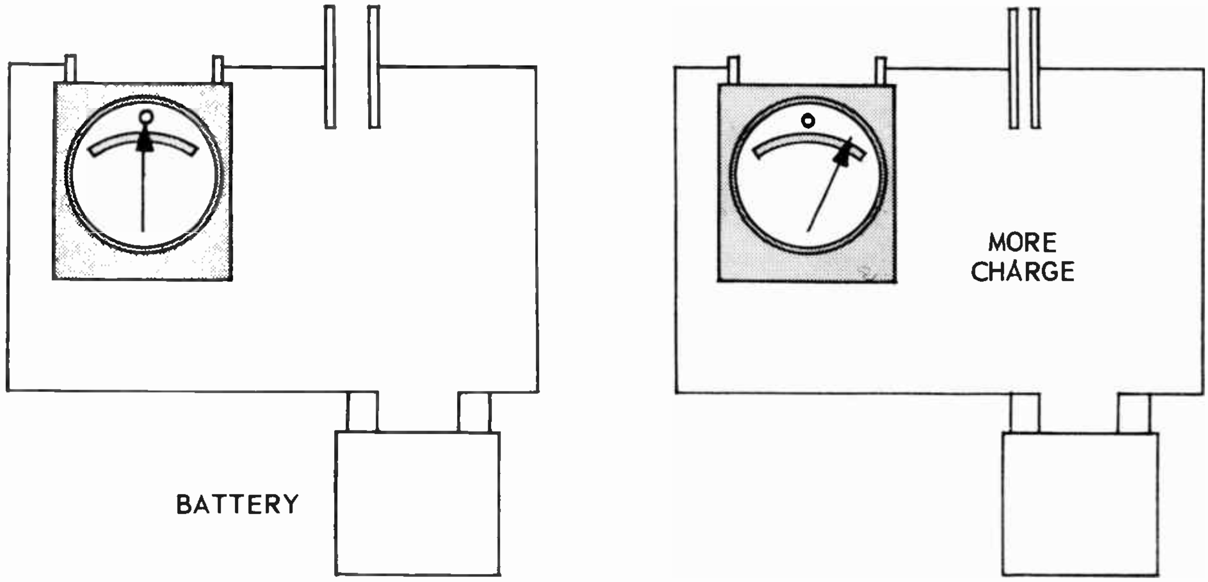


Fig. 12-11. Moving the plates closer together or using thinner dielectric allows the capacitor to take more charge.

receive a charge. The outside of the plates, which are not in contact with the dielectric, have no effect on charging.

The reason for all this becomes apparent if we look back at Fig. 12-7. The shifting of electrons in the atoms of the dielectric places an excess of these electrons (a negative charge) on the side of the dielectric which is toward the positive plate, and at the same time makes a deficiency of electrons (a positive charge) on the side of the dielectric toward the negative plate.

There really is attraction between the negative charge on one side of the dielectric and the adjoining positive charge of one plate, and between the positive charge on the opposite side of the dielectric and the adjoining negative charge of the other plate. The free electrons forming the charges on the plates are "bound" by the opposite charges of the dielectric. Only the electrons at the surfaces of the plates are thus bound by the dielectric charges. No matter how thick a plate is made, without increasing its surface area in contact with the dielectric, no additional charge can be held by the plate. A plate 1/1000 inch thick holds just as much charge as one an inch thick.

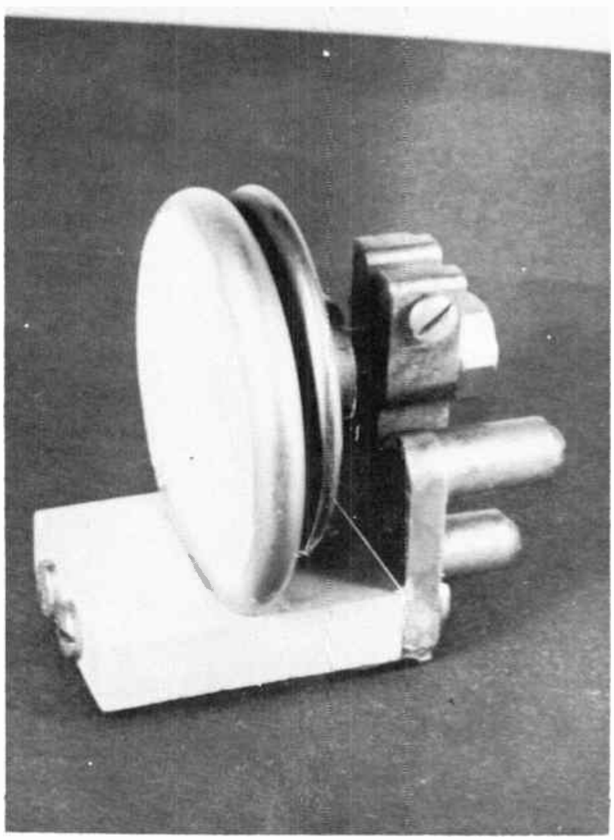
Now for another factor which affects the ability of the capacitor to receive and retain a charge, or the ability of the plates to receive and retain negative and positive charges which, together, are spoken of as the charge of the capacitor. At the left in Fig. 12-11 are the two plates of a capacitor in series with a zero-center current meter and a battery. The plates have taken their charges and the meter pointer has returned to zero. If the plates now are moved closer together, as at the right, there will be a momentary swing of the meter pointer in the direction which indicates charging. Merely moving the plates closer together, or decreasing the thickness of the dielectric, has enabled the capacitor to take an additional charge. There has been no change of surface area in contact with the dielectric, and no change of battery voltage, only a change in plate separation.

With all other factors unchanged, the charge that a capacitor will take is inversely proportional to the separation between plates or the thickness of dielectric. Half the separation enables the capacitor to take twice as much charge. Twice the original separation would cut the charge receiving ability in half. The reason is as follows.

The less the distance between the plates, or the thinner the dielectric, the more intense is the electric field between the plates and in the dielectric for any given battery voltage. The stronger or more intense this field, the greater is the displacement of electrons in atoms of the dielectric material. This increases the strength of charges at the surfaces of the dielectric, and these stronger dielectric charges attract and repel greater quantities of free electrons in the plates, or bind greater quantities of free electrons on the plates. Thus the plates and the capacitor are able to receive and retain greater charges.

Now we have two ways of altering the charge-receiving ability by changing the construction of a capacitor.

1. The charge is increased by more plate area in contact with the dielectric.
2. The charge is increased by less separation between plates, or by thinner dielectric.



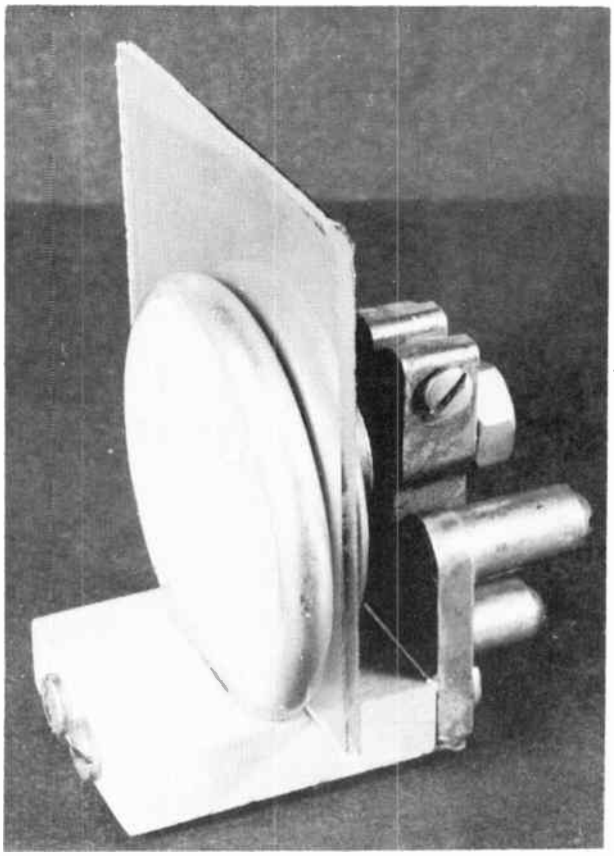


Fig. 12-12. Using glass instead of air as the dielectric increases the ability to receive and retain a charge.

There is one more thing which may be done with the construction of a capacitor to alter its ability to receive and retain a charge. This structural change may be illustrated by means of the small adjustable capacitor pictured at the left in Fig. 12-12. This is a type used for controlling energy feedback in some tube circuits. At the left there is only air acting as the dielectric between the two circular metallic plates. At the right there has been slipped into the space between the plates a sheet of glass to act as dielectric. The charge taken by the capacitor with this particular kind of glass dielectric proved to be more than four times as much as with air dielectric.

If mica were to be used instead of glass the charging ability of the capacitor would be increased about the same as with glass; much depending on the grade or kind of mica. Ordinary paper used for the dielectric would allow approximate doubling of the charge. Various kinds of ceramic dielectric materials would increase the charge more than one hundred times as compared with that for a capacitor similar in every way except for having air dielectric.

DIELECTRIC CONSTANTS. The number of times that the charging ability of a capacitor is increased by substituting for air dielectric some other dielectric substance, with everything else unchanged, is called the dielectric constant of that other substance. The dielectric constant of air usually is taken as 1.0. Only a few gases (hydrogen for one), and a complete vacuum, have dielectric constants less than that of air, and even then the difference is a very small fraction of one per cent. The accompanying table lists the dielectric constants of insulating and dielectric materials commonly used in radio and television.

DIELECTRIC CONSTANTS

Air	1.0	Polyethylene	2.3 to 2.4
Glass, flint	9.0	Polystyrene	2.4 to 2.8
Pyrex	4.0 to 5.0	Porcelain, unglazed	5.0 to 7.0
Mica	5.4 to 8.0	Quartz	4.7 to 5.1
Paper, plain	2.0 to 2.6	Steatite	4.8 to 6.5
wax impregnated	3.5	low loss	4.4
Phenolic compounds, mica filled	5.0 to 6.0	Titanium dioxide	90 to 170
low loss	5.3	Waxes	1.9 to 3.2

As shown by the table there is wide variation of dielectric constant in samples of materials of the same name, because there are differences in methods of manufacture and in the proportions and kinds of elementary substances put into various compounds or mixtures. Also, if a dielectric absorbs moisture the constant is increased. Charging with direct potential gives a different constant than charging with alternating potential, and the higher the alternating frequency the less is the effective dielectric constant. When charging is with direct potential the effective dielectric constant varies with the time the charging potential is continued. The apparent dielectric constant is altered also when other dielectric materials are close enough to be in the same electric field. Unless the dielectric constant of a particular batch of material is tested under laboratory conditions the specified value from tables is only approximate.

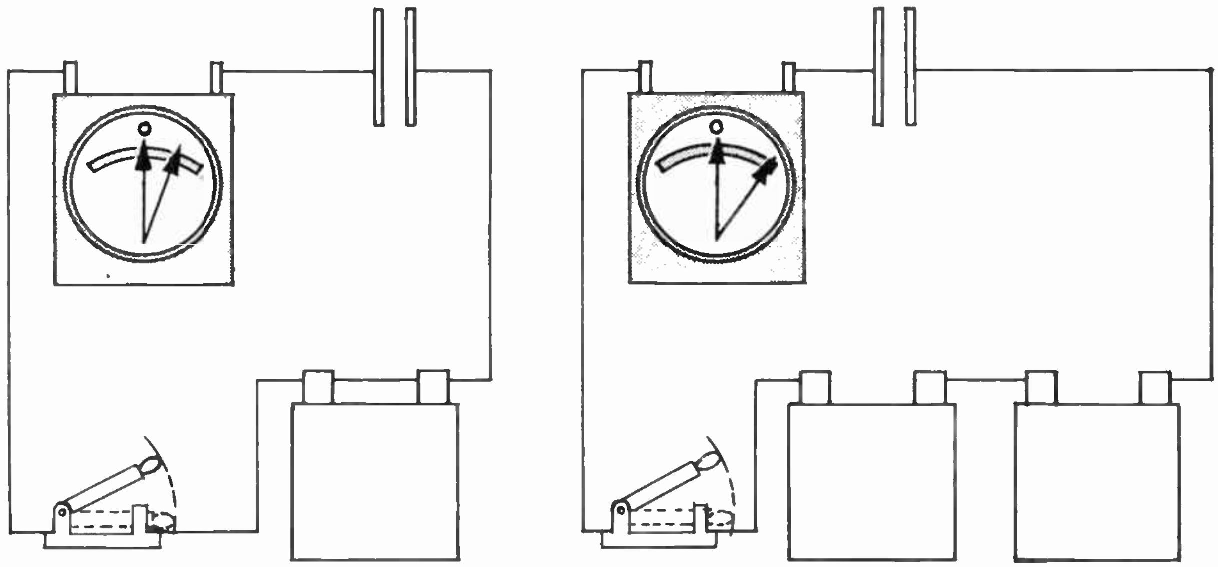


Fig. 12-13. The quantity of charge is directly proportional to applied voltage.

EFFECT OF APPLIED VOLTAGE. We have considered the three things which may be done in the capacitor itself to alter its ability to receive and retain a charge. These are:

1. A change of plate area in contact with the dielectric.
2. A change of plate separation, or of dielectric thickness.
3. A change of dielectric constant.

The only other thing which may be done to vary the charge of a capacitor has nothing to do with the manner in which the capacitor is constructed or adjusted. This other thing is the potential difference applied to the opposite plates of the capacitor. At the left in Fig. 12-13 the capacitor has been charged by closing the switch. The amount of charge, or the current producing the charge, has caused the meter pointer to swing about half the scale length toward the right before returning to zero with the charge completed.

If the capacitor now is discharged, and then, as at the right, an extra battery just like the first one is connected into the series circuit there will be twice the original potential difference applied to the capacitor when the switch again is closed. The meter pointer will swing about twice as far as before.

The quantity of charge taken by any capacitor is directly proportional to the applied difference of potential. Twice the potential difference will make the capacitor take twice as much charge, half the potential difference will cause only half the former charge. By using greater applied potential difference we may use a smaller capacitor and still have just as much charge, or we have more charge with the same capacitor.

CAPACITANCE. The ability of a capacitor to receive and retain a charge is called capacitance. If, when a potential difference of one volt is applied to the plates, the capacitor takes a charge of one coulomb of electricity then the capacitance is one farad. The farad is the basic unit of capacitance, just as the volt is the basic unit of potential difference and the ampere is the basic unit of rate of electron flow.

Fig. 12-14. High-voltage capacitors are larger than those of the same capacitance for use at lower voltage.

If you could construct a capacitor having a capacitance of one farad you would have difficulty in finding a place for your masterpiece, for if the plates were parallel and 1/16 inch apart, and the dielectric were air, this capacitor would cover a little less than 45,000 acres of ground when laid on its side. Needless to say there are no one-farad capacitors.

The common units of capacitance are the microfarad and micro-microfarad. One microfarad is the one-millionth part of a farad, and one micro-microfarad is the one-millionth part of a microfarad. Microfarad is abbreviated as mfd or sometimes as mf. Micro-microfarad is abbreviated as mmfd or as mmf. The symbol for capacitance in farads is the capital letter C. When this letter C is used for smaller units of capacitance the unit really should be specified, but this rule is not always followed.

When the rotor plates of the double capacitor of Fig. 12-2 are turned all the way in between the stator plates, for maximum area in contact with the dielectric, the capacitance of the larger section is 375 mmfd and of the smaller section is 150 mmfd. Capacitance of the two-plate capacitor of Fig. 12-12 without the glass is 7 mmfd, and with the glass it is 34 mmfd.

Fig. 12-14 shows two paper-dielectric capacitors, each having capacitance of 2 mfd. The dielectric paper in the smaller unit is thick enough to withstand maximum working voltage of 600 volts, and in the larger unit the dielectric is enough thicker to withstand 1500 volts. Using thicker dielectric reduces the capacitance per square inch of active plate surface, consequently the capacitor for higher voltage requires more plate surface and greater size for the same capacitance. The smaller of these two units is 2 7/8 inches high and the larger one is 4 1/8 inches high.

Formerly it was general practice to specify all capacitances in microfarads and decimal fractions of microfarads. Nowadays we usually specify small capacitances in micro-microfarads. Here are a few comparisons of equal capacitances shown in the two units.

Mfd.	.00001	.0001	.00025	.0022	.0068	.05
MMfd.	10	100	250	2200	6800	50000

Examination of these equal values discloses an easy way of translating from one unit to the other. To change the decimal fraction of mfd to the whole number of mmfd add to the fraction enough ciphers to make a total of six places to the right of the decimal point.Example: change .00025 into .000250. Then drop the decimal point and all ciphers to the left of the first numeral. To change a whole number of mmfd to a decimal fraction of a mfd add to the left of the whole number enough ciphers to make a total of six digits, then put on the decimal point at the left and drop all ciphers originally to the right-hand side of the whole number. Example: change 2200 into .002200, then drop the right-hand ciphers to make .0022.

The property of a capacitor which we call capacitance sometimes is called capacity. The correct name would be electrostatic capacity, to distinguish this kind from all the other kinds of capacity, from that of a quart measure on through a long list. It is easier to say capacitance, a word whose meaning cannot be mistaken.

TELEVISION

LESSON NO. 13

TYPES OF CAPACITORS AND THEIR USES

"Training Your Mind"

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

DO NOT TEAR — CUT ALONG THIS LINE AND SEND IN FOR GRADING.

Training Your Mind

C.T.I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

LESSON NO. 13

TYPES OF CAPACITORS AND THEIR USES

We have learned how capacitors charge and discharge, how they permit flow of alternating current while preventing flow of continual direct current, and we have learned about the features of a capacitor which determine its capacitance. All this will help in understanding the behavior of circuits which include capacitors, and will help in recognizing some of the troubles which are due to faulty capacitors. But in service work it is necessary to understand also the particular advantages and disadvantages of the different kinds of capacitors, and to be acquainted with such matters as voltage ratings, tolerances, and the effects of temperature change. These things we shall now proceed to discuss.

VOLTAGE RATINGS. One of the items always mentioned in descriptions of capacitors is working voltage, which is the maximum continual voltage at which the capacitor is designed to operate without danger of breakdown. This rating nearly always is specified as a d-c (direct-current) working voltage. A d-c working voltage is also the maximum or peak working voltage. Were the same limit to be expressed as an effective alternating voltage, this a-c voltage rating would be only 0.707 times the d-c rating. This is because, as you will recall, the effective value of an ordinary alternating voltage is only 0.707 times the peak value of the alternating voltage. It is true also that the peak or maximum value of an alternating voltage is approximately 1.4 times its effective value. Consequently, the d-c working voltage of a capacitor must be 1.4 times as great as the highest effective a-c voltage which will be applied to the capacitor. As an example, if a capacitor is to be subject to 200 effective alternating volts, the d-c working voltage should be no less than 1.4 times 200, or should be at least 280 volts.

In addition to working voltage you may find mention also of a flash voltage or a flash test. This refers to momentary application of a voltage which usually is about twice as high as the rated working voltage. A flash voltage test is a check on possible breakdown with brief voltage surges such as may occur in some applications.

TOLERANCE. The tolerance of a capacitor is the maximum difference which may exist between actual capacitance and the listed or rated capacitance. Tolerances are specified as percentages, plus or minus. For example, if the tolerance is $\pm$ 20%, the actual capacitance may be anything from 20% less than the rated or listed value up to 20% more than that value. Were a capacitor rated at 1000mmfd with tolerance of $\pm$ 20%, the actual capacitance might be anything between 800 mmfd (20% less) and 1200 mmfd (20% more). When no tolerance is mentioned, it usually is at least $\pm$ 20%, and for some capacitors may be as much as 50%.

PAPER CAPACITORS. Nearly always, it is true that in radio and television receivers more of the capacitors will be of the tubular paper-dielectric type having the general form pictured in Fig. 13-1 than of any other kind. The tubular capacitor at the top has an outer covering of cardboard or heavy paper. Down below is a tubular style with an outer covering of molded insulation.

Paper tubular capacitors most often are used in capacitances ranging from 0.0001 mfd (100 mmfd) up to 1.0 mfd. For smaller capacitances it is usual practice to use mica or ceramic capacitors, and for larger cap-

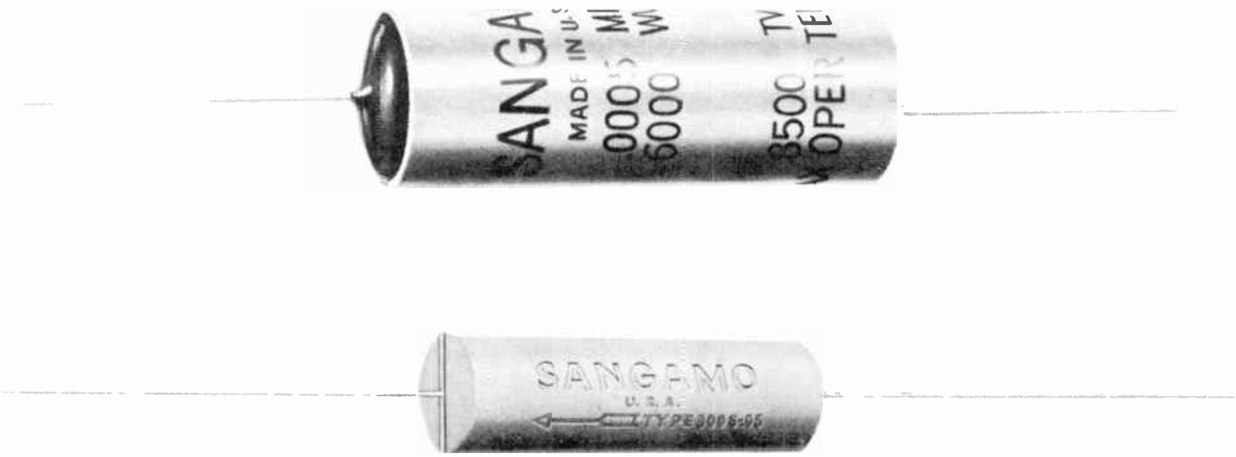


Fig. 13 – 1 Tubular capacitors with paper dielectric.

acitances to use electrolytic types. Paper capacitors are available with d-c voltage ratings all the way from 100 to 10,000 volts. Common ratings for types having wax impregnation include 100, 200, 400, 600, and 1000 volts.

For any voltage rating of 600 or more, paper capacitors often are of the oil filled type. These have dielectric of oil treated paper, and the entire capacitor is enclosed within a metal container filled with insulating oil. Oil filled capacitors are made in d-c voltage ratings from 600 to 10,000 volts. Capacitances range from 0.001 mfd (1000 mmfd) up to 4.0 mfd and even more. Several oil filled capacitors are shown in Fig. 13-2.

The internal construction of tubular paper capacitors is shown in Fig. 13-3. In the left illustration, layers of paper dielectric extend out beyond the metal foil plates on both sides, and at both ends on the final roll. When such an assembly of papers and foils is rolled into a cylinder the strips of metal foil become, in effect, long coiled conductors insulated from each other throughout their entire length by the paper dielectric. Any insulated coiled conductor will act as an inductor and will have the property called inductance. Inductance is not like capacitance and, for reasons which we shall discover later, is not desirable in a capacitor which is to be used at high frequencies.

The property of inductance in a rolled paper capacitor may be greatly reduced by using the internal construction shown at the right in Fig. 13-3. One strip of metal foil is so positioned as to extend beyond the paper dielectric all the way along one edge, while the other strip of foil is placed so it extends beyond the paper all along the other edge. When the foils and papers are rolled, the two foils are everywhere separated from each other by paper, but one foil sticks out at one end of the capacitor, while the other foil sticks out at the other end.

The rolled foil extending from each end then is pressed together into a fairly solid mass to form a terminal connection at that end. Thus all the turns of each foil are joined together, and instead of acting like long coiled conductors insulated throughout their lengths, there is the effect of two groups of short interleaved plates, one group formed by each of the foils. The construction produces what is called a non-inductive capacitor.

Non-inductive paper capacitors in values from 0.001 mfd to 0.5 mfd sometimes are made with molded insulated coverings of flat rectangular shape. In the smaller sizes some of these paper units look very much like mica capacitors, and may even have names which suggest the use of mica dielectric. However, the capacitance of these small paper-dielectric units in a housing of given dimensions usually is much greater than that of a mica capacitor of the same size. After handling a few such types you are not likely to be fooled.

When a paper capacitor constructed as at the left in Fig. 13-3 is rolled into its final form, one end of one foil will be around the outside of the remainder of the capacitor elements. In circuits which require that either end of the capacitor be connected to ground, it is desirable that the ground connection be to the outside foil which then acts to shield the capacitor from effects of external electrostatic fields. The terminal connected to the outside foil most often is identified by circular printed bands of contrasting color or shade such as you can see near the ends of the capacitors in Fig. 13-4. Sometimes you will find the words "Outside Foil". The terminal for the outside foil of the capacitor at the bottom of Fig. 13-1 is identified by a ridge across that end of the molded cover and by an arrow pointing toward the outside foil terminal.

MICA CAPACITORS. Fixed mica capacitors always are enclosed within a housing molded of some resin or phenolic insulating material such as Bakelite. Some typical units are pictured by Fig. 13-5. Mica capacitors of types in most common use are constructed with alternate thin sheets of mica dielectric and of metal foil plates. Every second metal plate sticks out on one side of the mica, and the intervening plates stick out on the opposite side. The extended foils are pressed together on each side, and each group is

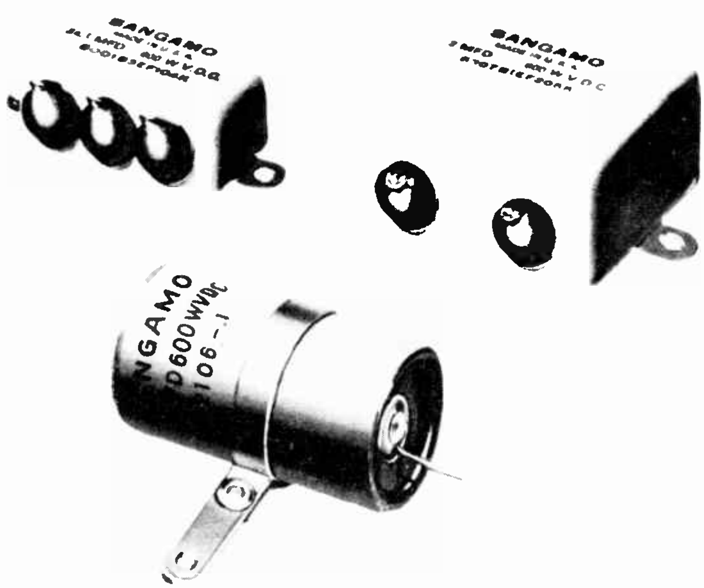


Fig. 13 – 2 Oil Filled paper capacitors.

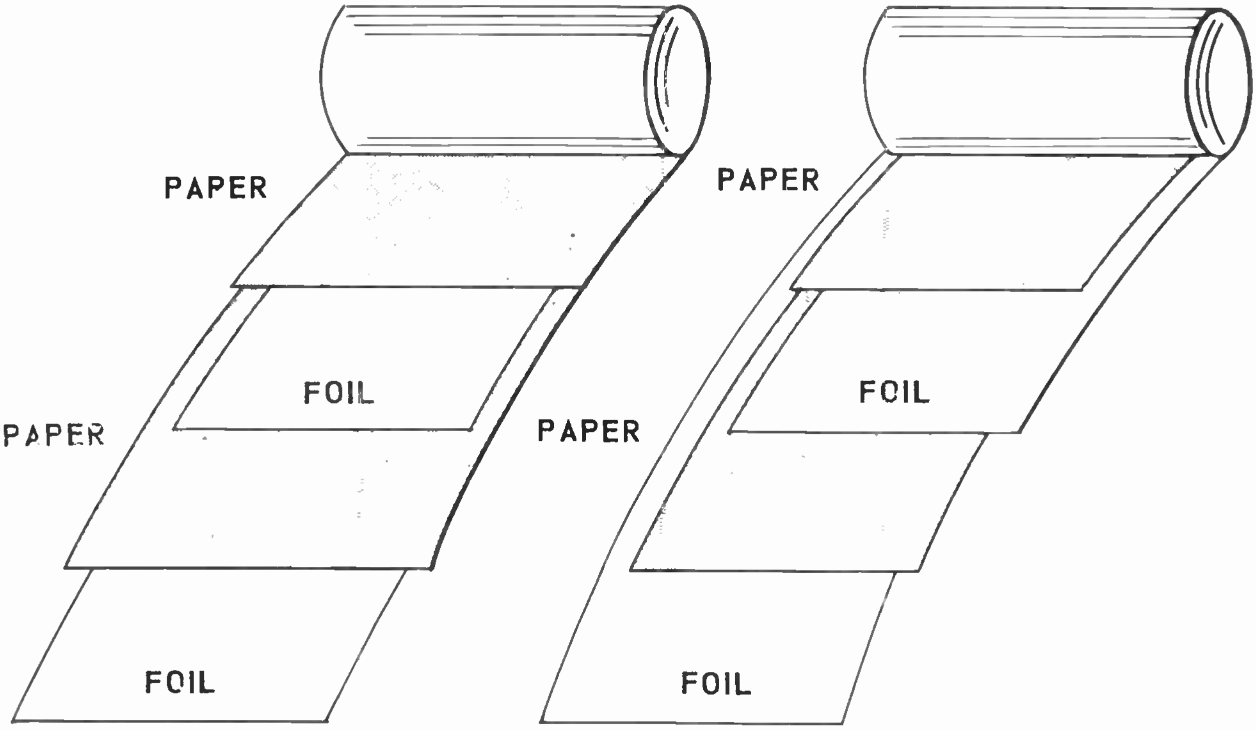


Fig. 13 – 3 Construction of inductive rolled paper capacitor (left) and of non-inductive type (right).

connected to its external terminal. Since there are no coiled conductors, mica capacitors are inherently non-inductive. Good quality mica capacitors have long life with minimum trouble, and generally give dependable performance.

The chief disadvantage of mica capacitors is that their cost is considerably more than that of paper and ceramic types in capacitances much greater than 1000 mmfd. Another possible disadvantage of micas used as coupling capacitors is their larger size compared with ceramics. Between the metal in any capacitor and the metal of a receiver chassis there is capacitance because the two metals act like plates which are separated by dielectric insulation of some kind or other. A large capacitor has more such capacitance to ground than has a smaller one, and since this external and uncontrolled capacitance usually is undesirable, the large capacitor may be less desirable.

Capacitances of commonly used foil-mica capacitors range from 5 mmfd to 0.01 mfd. D-c voltage ratings most often are either 300, 400, or 500 volts. Standard tolerance, when not otherwise specified, is $\pm$ 20%. Tolerances of 10% and 5% are available at extra cost. In types called transmitting mica capacitors the capacitances extend as high as 0.1 mfd, and d-c voltage ratings may extend to 12,000 volts or even higher in the smaller capacitance sizes. These high-voltage micas are used in some television circuits.

A kind of mica capacitors called silver micas are made with highest grade sheet mica coated with a compound which is changed to pure silver in high-temperature furnaces during manufacture. These capacitors retain their original characteristics very closely for long periods of time, and they have small energy losses even at very high frequencies. Capacitances range from 1 mmfd to 0.01 mfd or even higher. D-c working voltages are 300, 400, or 500 volts. Standard tolerance, when not otherwise stated, usually is ± 10%. Tolerances of 5%, 3%, 2%, and 1% are available at greater cost. The cost of silver micas is several times that of foil micas with the same capacitance, voltage rating, and tolerance.

CERAMIC CAPACITORS. Ceramic capacitors take their name from the kind of dielectric, which is a ceramic (porcelain-like) material with which is mixed a substance of exceedingly high dielectric constant. This substance is most often titanium dioxide. The resulting dielectric constant of the compound may be as much as 1500 to 2000, or even more in some cases. The ceramic compound may be formed into thin-walled cylinders which are plated or coated inside and outside with layers of silver which act as the plates for the resulting tubular or cylindrical capacitor. Ceramic capacitors are made also in the form of thin round discs wherein is a wafer of the ceramic compound coated with silver on opposite sides to form the plates. The combination of dielectric cylinder or wafer with the coated plates is protected with an outer cover of insulating compound or sometimes with an outer tubing of steatite.

Tubular or cylindrical ceramics may have radial leads which extend outward from one side, near the two ends, as at A in Fig. 13-6, or they may have axial leads from the two ends as at B. Diameters range from 3/16 to about 3/8 inch, with lengths from about 1/2 inch up to about 1 1/4 inches. The range of capacitances is all the way from 1/2 mmfd to 10,000 mmfd or 0.01 mfd. Tolerances most often are ± 20% or 10%, but may be much closer when greater accuracy is required. D-c working voltage usually is 500, but in some types may be anything between 100 and 2,000 volts, in steps.

Disc ceramics may contain a single capacitor unit with its two extending leads as shown at C in Fig. 13-6, or they may contain two units as at D wherein each of the two outside leads connects to one capacitor, with the center lead being common to both units. Capacitances of disc ceramics usually are 0.001 mfd or more, ranging to around 0.01 mfd in single units and about half this in dual types. D-c working voltages are 500 unless otherwise specified. These disc type units are little more than 1/8 inch thick, and their diameter is around 5/8 to 3/4 inch.

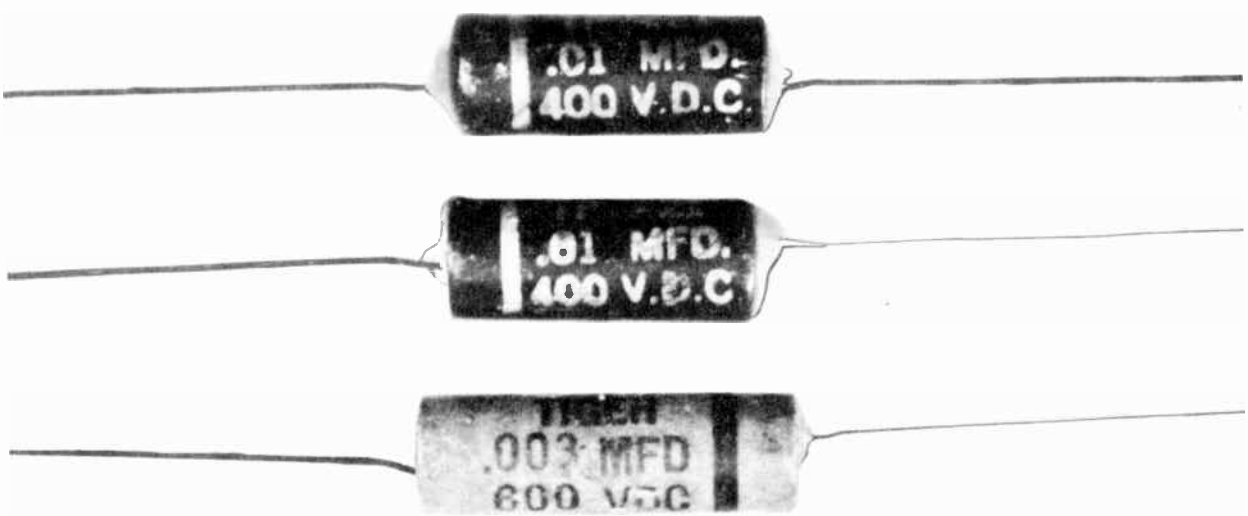


Fig. 13 – 4 Bands on these paper tubulars are near the leads to be used for grounding.

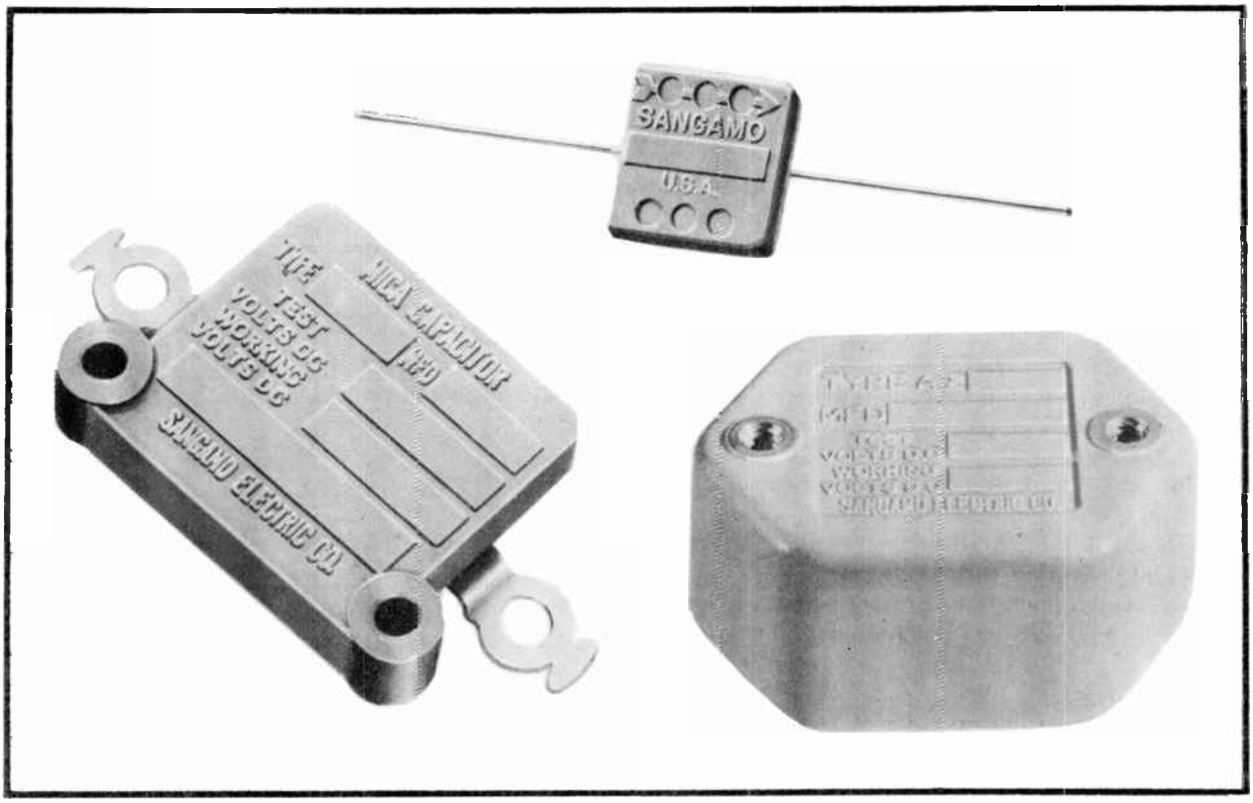


Fig. 13 – 5 Mica capacitors with leads, lugs, and screw terminals.

Because of the small dimensions of all ceramics, and their method of construction, the inductance is so very small as to make these capacitors well suited for high-frequency applications. Great numbers are found in television and f-m receivers. The chief uses are for by-pass capacitors and coupling capacitors in all kinds of circuits, and for temperature compensating capacitors in r-f oscillator circuits. Ceramics, as a class, seem less able to withstand high temperatures and momentary voltage surges than do the mica and paper types. This may be because ceramic capacitors too often have been put into circuits where their voltage ratings are exceeded. Great care must be used when soldering to the terminal leads that the capacitors are not overheated to a degree which will break the dielectric and cause either open or short circuits.

TEMPERATURE COMPENSATING CAPACITORS. When there are considerable changes of temperature in the coils, conductors, and supports for parts of amplifier and oscillator circuits there are resulting changes of dimensions and positions which cause the operating frequencies to vary. In nearly all cases the physical changes are such that, when there is a rise of temperature, the frequency becomes lower than it should be, or lower than the value for which adjustments originally were set. When temperature of the circuit parts later returns to a lower value there is an increase of operating frequency.

A lowering of frequency in the circuits of the tuner and video i-f amplifier of a television receiver, and in generally similar circuits of any other kind of receiver, is equivalent to an increase of capacitance in

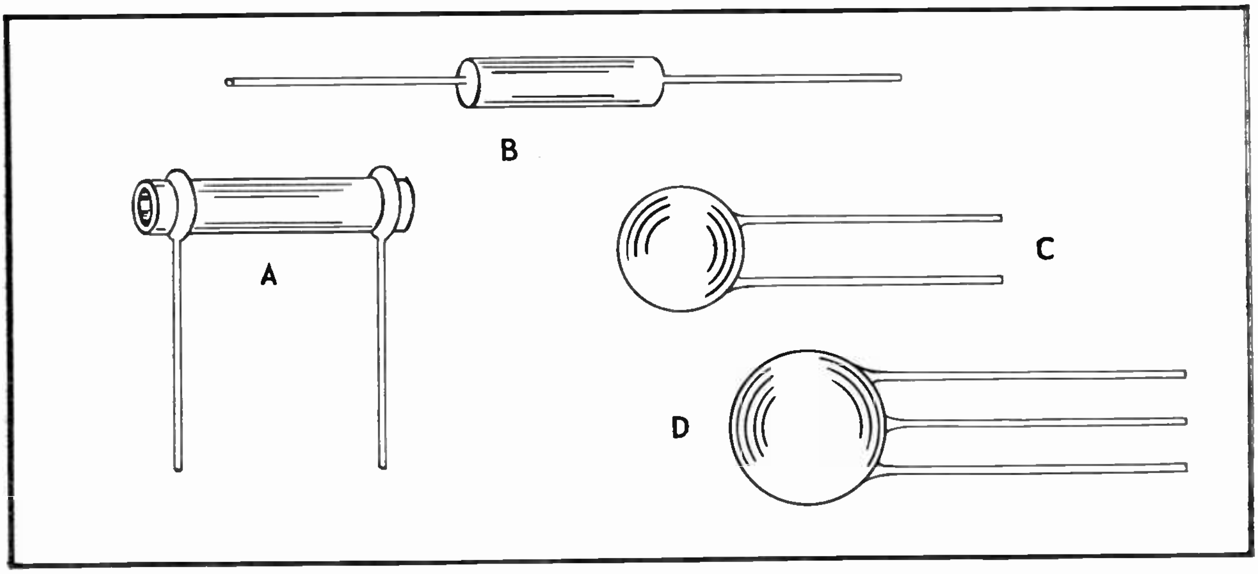


Fig. 13 – 6 Lead arrangements as used with tubular and disc ceramic capacitors.

those circuits. At least, it would be possible to cause the same lowering of frequency by using more capacitance. The undesirable lowering of frequency with rise of temperature may be counteracted by connecting into the affected circuit a special type of capacitor whose capacitance decreases when its temperature rises. The decrease of capacitance in this "temperature compensating" capacitor then tends to raise the operating frequency just as much as the rise of temperature tends to lower the frequency, and the result is a nearly constant frequency with ordinary changes of temperature.

A capacitor whose capacitance becomes less as its temperature rises is said to have a negative temperature coefficient of capacitance, or simply a negative temperature coefficient. Were the capacitance to increase with rise of temperature there would be a positive temperature coefficient. If the capacitor maintains a nearly constant capacitance, when there are changes of temperature that capacitor has a zero temperature coefficient. It happens that ceramic capacitors may be made to have any kind of temperature coefficient, simply by changing the composition of the dielectric. Consequently, nearly all temperature compensating capacitors are of the ceramic type. In order that changes of temperature may have full effect on the compensating capacitor, this unit must be mounted close to parts which otherwise would cause troublesome variation of frequency. When possible, the capacitor leads are soldered directly to the affected parts.

Temperature coefficient is specified in "parts per million per degree centigrade". This is the same as specifying a change of so many micro-microfarads for every microfarad of rated capacitance for each centigrade degree of temperature change. Here is an example. Assume we have a capacitor whose normal capacitance is 1000 mmfd or 0.001 mfd, and assume that this unit has a negative temperature coefficient of 750. If the capacitance were 1 mfd the change then would be 750 mmfd for every centigrade degree of temperature change. But the capacitance is only 0.001 or 1/1000 mfd, and so the change actually will be only 1/1000 of 750 per degree. This means a change of 750/1000 mmfd for every centigrade degree of temperature change.

Supposing we assume that adjustments will be made at 20° C (68° Fahrenheit) and that normal operating temperature will be 45°C (113° Fahrenheit). This is a change of 25 centigrade degrees. For each degree the capacitance will decrease by 750/1000 mmfd. Multiplying 750/1000 by 25 shows that the total decrease of capacitance will be 18.75 mmfd. The original 1000 mmfd of capacitance will decrease to 981.25 mmfd when temperature rises from 20° to 45° centigrade.

Sometimes it is important to know how much, or how little the capacitance may vary with changes of temperature even though the capacitor is not used for temperature compensation. For instance, we might wish to know how much a capacitor is likely to upset the operating frequency of a circuit when there are changes of temperature. Such information is given by specifying maximum negative and positive temperature coefficients. Where moderate change of capacitance will have little effect the coefficient might be something like 120 positive to 160 negative.

ELECTROLYTIC CAPACITORS. An electrolytic capacitor is radically different from all other types in both construction and action. One of the plates in the electrolytic capacitor is a thin sheet of aluminum

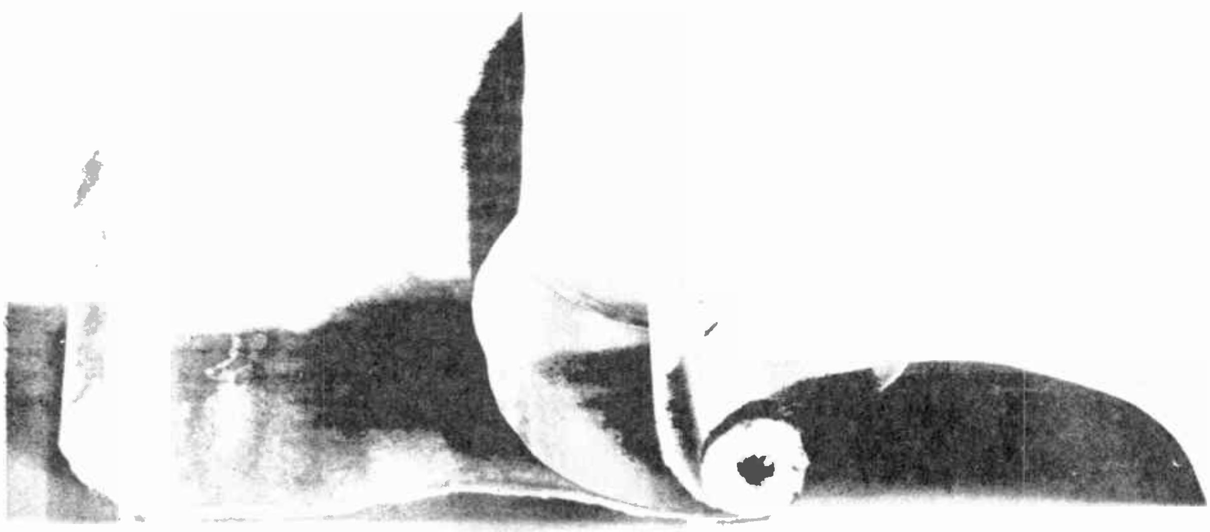


Fig. 13 – 7 Construction of electrolytic capacitor with gauze for holding the electrolyte.

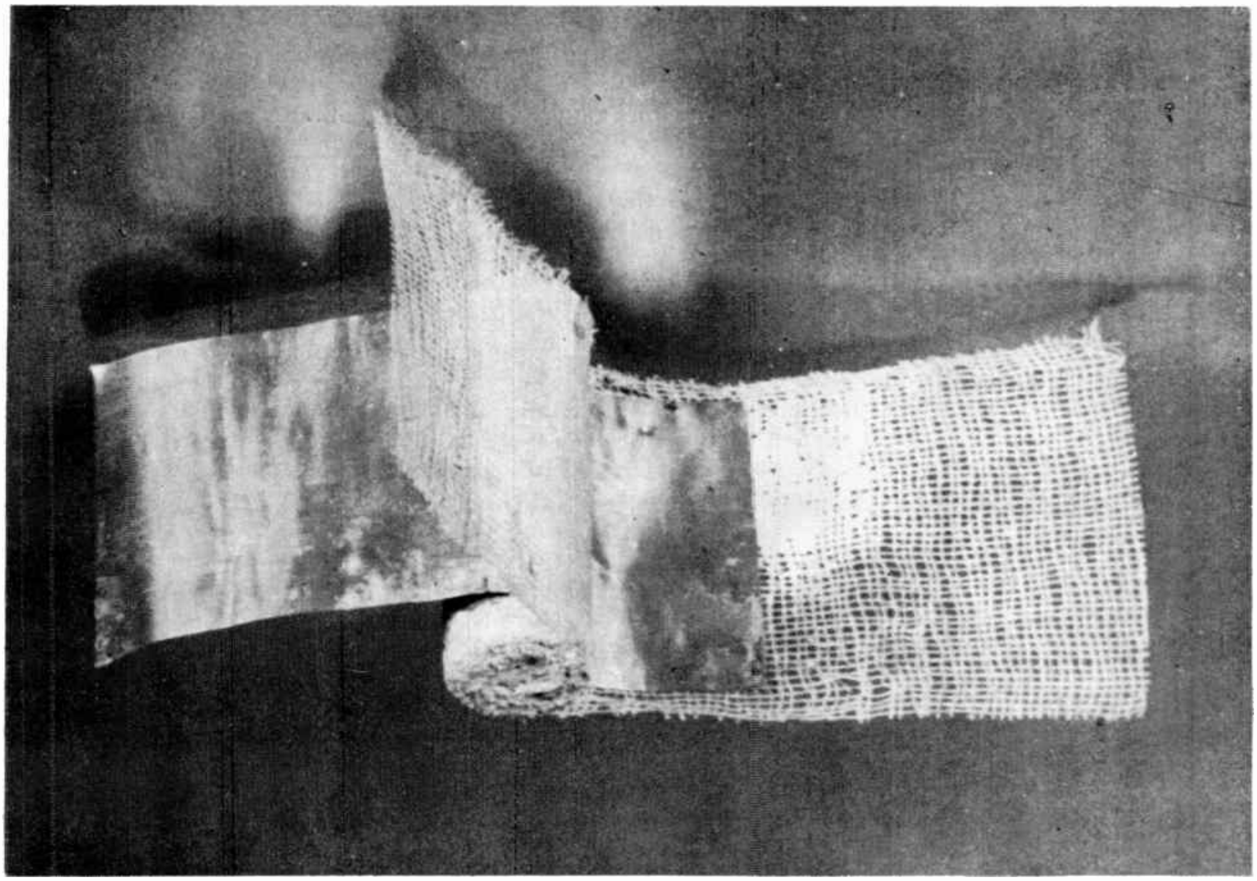


Fig. 13 – 8 A method for holding the paste inside a dry electrolytic capacitor.

which may be rolled or corrugated and which usually has the surface roughened or etched to increase the area. The dielectric is a thin film of oxide which is formed by electro chemical action right on the surface of the aluminum. The second plate is either a moist paste containing the electrolyte liquid, or else is the electrolyte liquid itself. The paste is used in what are called dry electrolytics, the liquid is used in wet electrolytics.

All the early electrolytic capacitors were of the wet type, while most of those used today are of the dry type. Dry electrolytics have the advantages of lighter weight and somewhat smaller sizes for very large capacitances, and are practically free from trouble resulting from escape of liquid. Wet electrolytics have the advantage that a breakdown due to application of excessive voltage is healed by the liquid flowing back into place, while a breakdown or puncture due to excessive voltage on a dry electrolytic causes permanent failure. Unless specifically stated otherwise, the remainder of our discussion will relate to the generally used dry electrolytic capacitors.

Dry electrolytic capacitors are of rolled construction such as shown for a fairly typical unit in Fig. 13-7. Extending toward the left is one layer of cellulose or cotton gauze saturated with electrolyte liquid. Then comes the active aluminum plate on whose surface is formed the dielectric film. Next is a second layer of the absorbent gauze, and extending over toward the right is another thin sheet of aluminum. This second aluminum sheet acts neither as an active plate nor as a support for the dielectric, it is merely a conductor whose purpose is to bring electron flow or current from the external terminal of the capacitor to the electrolyte which is an active plate.

There are many variations in construction details of electrolytic capacitors, but principles remain unchanged. For example, in the capacitor shown partly unrolled by Fig. 13-8 the electrolyte paste is held in place by an open mesh fabric which looks like a piece of cheesecloth.

When an assembly of aluminum sheets and electrolyte paste or liquid is first made up it is not a capacitor because there is no dielectric film. The film is formed when a direct electron flow is sent through the unit in such polarity that the flow passes first to the inert aluminum conductor, thence through the electrolyte to the aluminum on which the dielectric film is to be formed, and from this plate back to the external source. As the dielectric film is formed, there is increasing high resistance to electron flow in the direction mentioned. If applied voltage is made higher in order to maintain the flow, the film continues to become thicker and thicker and to increase its resistance.

The inert foil at which electron flow enters the capacitor is called the cathode, and the foil which becomes the active plate is called the anode. During the process of film formation the cathode is connected to the negative side of the source and the anode is connected to the positive side. If these external connections should be reversed, even after the film has been built up to have high resistance to the original flow, there will be a large current in the reverse direction. The electrolytic capacitor has dielectric resistance to electron flow in only the direction which originally formed the dielectric film. Consequently, this type of electrolytic capacitor may be used only in circuits where it is subjected to direct voltage or direct electron flow, not to alternating voltage or alternating flow.

One of the principal uses of electrolytic capacitors is in receiver power supply sections following the rectifier, where there is only direct voltage. These capacitors are used also as by-passes in plate, screen, and grid circuits, where again there is direct voltage. There are many other places where we shall find electrolytics, but always the circuits will be those operating at the relatively low audio frequencies, sync frequencies or sweep frequencies, and power line frequencies. These capacitors cannot be used in circuits operating at radio, video, or carrier frequencies because under such conditions the energy losses become excessive.

The electrolytic capacitor which may be used only where there are direct voltages and no alternating voltages is called a polarized type. One or both external terminals always are plainly identified for correct connection. Usually only the positive terminal is marked, either with the word positive, the abbreviation "pos", or a plus sign (+). If there are flexible insulated leads the insulation on the positive lead may be red in color. By the positive lead or terminal is meant the one which must be connected toward the positive side of whatever source is furnishing voltage to the capacitor, or toward the side of the circuit toward which there is to be electron flow.

There are non-polarized electrolytic capacitors in which there are two anodes and two dielectric films. These are in common use for a-c motor starters and other purposes in industrial and commercial electrical equipment, but not in radio or television receivers.

The principal advantages of electrolytic capacitors over all other types are the very large capacitances obtainable in small space and at fairly low cost. The large capacitance results from the extreme thinness of the dielectric film which, for a forming potential of 100 volts is only about 1/25000 inch. At higher form-

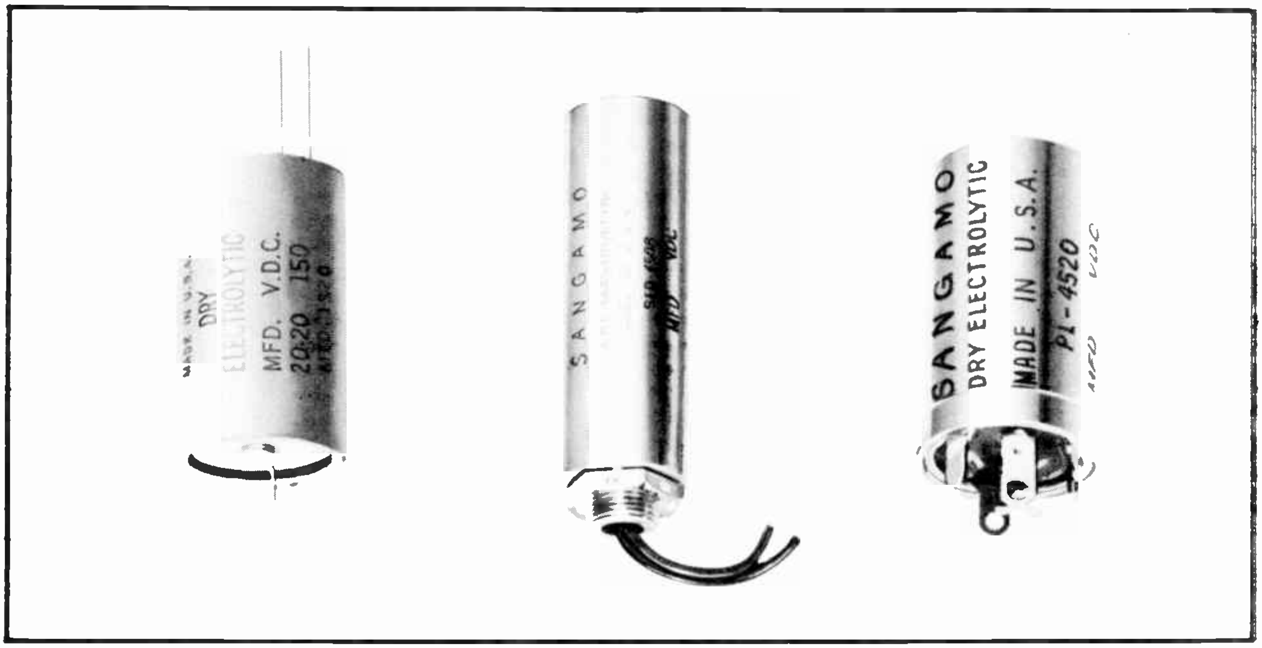


Fig. 13 – 9 Types of cases or housings used for dry electrolytic capacitors.

ing voltages and working voltages there is thicker film, which means less capacitance per square inch of plate area and a larger capacitor for any given capacitance. Capacitance seldom is less than 0.5 mfd, and in single capacitors may often be as high as 4000 mfd and sometimes as high as 6000 mfd.

Capacitance tolerances usually are on such a basis that the rated or listed capacitance is practically the minimum actual value, while the actual capacitance may be almost anything from one-half more than the rated value up to twice the rated value. When you buy an electrolytic capacitor of any reputable make you are practically certain of getting enough capacitance, and usually more than enough. Working voltages most often are in the range between 25 and 450 d-c volts, but may be as low as 6 volts and sometimes a little higher than 450 volts.

A few of the many styles of housings or casings used for electrolytic capacitors are illustrated in Fig. 13-9. At the left there is a cardboard sleeve around the outside of a metal can which encloses the capacitor. At the center the capacitor is contained in a metal can, the bottom end of which is a threaded extension that goes through a hole in the chassis of the receiver. A nut screwed on from below holds the capacitor in position. At the right is another metal can from the bottom edge of which extend thin metal lugs that pass through slots in the chassis or in a mounting plate fastened to the chassis. After the lugs are put through the mounting slots the lugs are twisted to hold the capacitor in place.

Other electrolytic capacitors are in rectangular cardboard housings of various proportions and sizes. Still others are within molded housings of insulating material much like the molded covers used for some tubular paper capacitors. Wet electrolytics nearly always are in cylindrical metal cans similar to the style at the center of Fig. 13-9, but having liquid-tight bushings in the extension that goes through the chassis. Wet electrolytics are designed for mounting only with the terminals or leads downward and with the cans vertical.

Electrolytic capacitors are available with one, two, three, or four capacitor units within a single housing. All the types pictured by Fig. 13-9 are dual, with two sections in each housing. The metal can generally is the negative terminal. The single lead on one end of the capacitor at the left is attached to the can, so is negative. The two leads from the other end are positive, one for each section. The two leads of the center unit are positive and the metal can which comes in contact with chassis metal is negative. Capacitors of this general style may be fitted with insulating bushings or washers for the extension that goes through the chassis hole, thus allowing the negative can to be insulated from the chassis metal. The negative can of the capacitor at the right connects to and grounds to the chassis when the mounting lugs are put through the slots and twisted.

Fig. 13-10 is a picture of a dry electrolytic capacitor removed from its metal can. The positive terminal lug at the left is mounted in an insulating disc which was fastened into one end of the can. The negative foil of the capacitor has attached to it a long strip of metal which will make close contact and an electrical connection to the inside of the metal can.

When an electrolytic capacitor is correctly connected into a circuit, with negative and positive of the capacitor respectively toward negative and positive of the voltage source, there will be a small current through the capacitor even though it is in perfect condition. This is leakage current. Normal leakage current increases with capacitance and increases with working voltage or applied voltage. In a 1-mfd unit operating at 100 d-c volts the leakage may be as much as 0.3 milliampere, and with 40 mfd at 450 volts the leakage may be 2.0 to 2.5 milliamperes without indicating a defective capacitor.

When voltage is applied to an electrolytic capacitor which has been out of use the initial leakage current is five or six times normal, but drops to normal within only a few seconds of time provided the capacitor has not been too long idle. The longer the capacitor has been idle the greater is the initial leakage and the longer the time before leakage comes down to normal. If one of these capacitors has been idle for as much

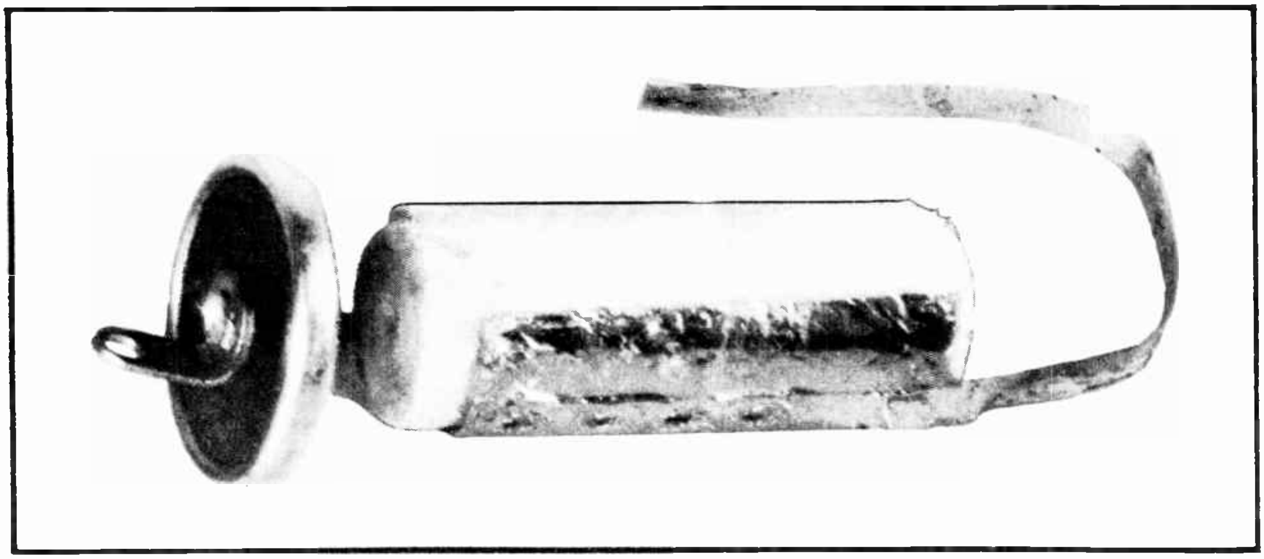


Fig. 13 – 10. Insulated positive terminal and negative connecting strip of a dry electrolytic.

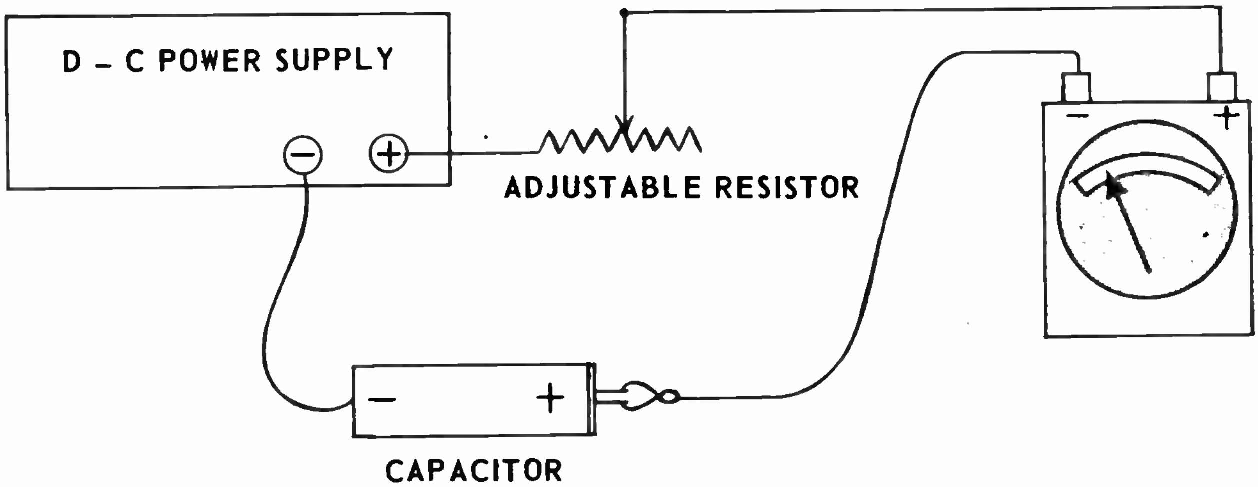


Fig. 13 – 11. Connections for re-forming dry electrolytic capacitors.

as a year and then is subjected to voltage such as through the rectifier in a power supply, the large leakage current is likely to so overload the rectifier as to ruin the electron-emitting surface of the rectifier cathode.

Where there is possibility of excessive leakage current the capacitor should be re-formed before it is used in the regular circuit Re-forming is done with a direct voltage from a source whose positive side is connected to the positive terminal of the capacitor and whose negative side is connected to negative of the capacitor. Three or four dry cells connected in series will furnish enough voltage to build up the dielectric film in the capacitor and thus bring the leakage down to a point where the capacitor may be connected into its regular circuit.

Any d-c power supply may be used for re-forming electrolytics as shown by Fig. 13-11. Current through the capacitor must not exceed 5 milliamperes, and is better held down to about 3 milliamperes. Maximum resistance of the adjustable resistor must be such that no more than the maximum allowable current can flow with full voltage from the power supply. This resistance is computed with your regular formula for ohms, volts, and milliamperes.

Electrolytic capacitors must not be subjected to excessively high temperature. The usual limit is about 140° F. or 60° C. unless the capacitor is especially designed for high-temperature operation. Capacitors located too close to tubes, especially rectifiers, or too close to large resistors or to power transformers are likely to overheat. This tends to evaporate the electrolyte in spite of good sealing, to increase leakage current, and to shorten the useful life of the capacitor. Excessive temperature or frequent voltage overloads may force electrolyte out through the seal, as illustrated by Fig. 13-12. Without overloading or overheating, the life of an electrolytic capacitor usually will be six to eight years in ordinary service.

CAPACITIVE REACTANCE. A capacitor connected in series with a circuit prevents any continued one-way electron flow in that circuit, because no flow can pass through the insulating dielectric unless the dielectric punctures. But this capacitor in series with a circuit may be continually charged and discharged, first in one polarity and then in the opposite polarity, provided electron flow surges back and forth in the circuit. This simply is a way of saying that a capacitor permits alternating electron flow or alternating current in itself and in the connected circuit.

A

Of course, the alternating flow in a capacitor will not be so free as in a continuous conductor, because, during each alternation the capacitor canbe charged with only a certain fraction of a coulomb of electricity, not with an infinitely large quantity. During the following alternation only that same quantity of electricity can flow in the opposite direction as the capacitor discharges. It is evident that the number of coulombs of electricity which can move during each alternation and during each cycle, with any given applied voltage, must be directly proportional to the capacitance of the capacitor. Doubling the capacitance will allow doubling the movement, as measured in coulombs. Halving the capacitance will halve the quantity of electrons moved during each cycle.

Supposing both the applied voltage and the capacitance remain unchanged while frequency is varied. Then there will be a certain quantity of electricity or a certain number of coulombs moved during any one cycle. During two cycles there will be twice the total quantity moved, during three cycles there will be three times the quantity, and so on. The coulombs of electricity moved always will be directly proportional to the number of cycles, and, of prime importance, the coulombs per second must be proportional to cycles per second. The name for cycles per second is frequency, and the name for coulombs per second is amperes. So we may say that current in amperes is directly proportional to frequency and is also directly proportional to capacitance. The greater the capacitance and the higher the frequency of applied voltage the greater will be the capacitor current in amperes.

From that last statement it follows that the smaller the capacitance and the lower the frequency, the less will be the current with any applied voltage. A reduction of current with no change of voltage means, in effect, that there is greater opposition to flow of current. Reducing capacitance and frequency effectively increases the opposition of the capacitor to flow of alternating current. This opposition of a capacitor to flow of alternating current is called capacitive reactance. The symbol for reactance is the capital letter X. To show that this reactance is that of a capacitor we add the symbol for capacitance, making the symbol X_C for capacitive reactance.

All kinds of opposition to all kinds of current are measured in the same unit, the ohm. Consequently, the unit for capacitive reactance is the ohm. One ohm of capacitive reactance has the same effect in limiting alternating current as has one ohm of resistance in limiting the same current.

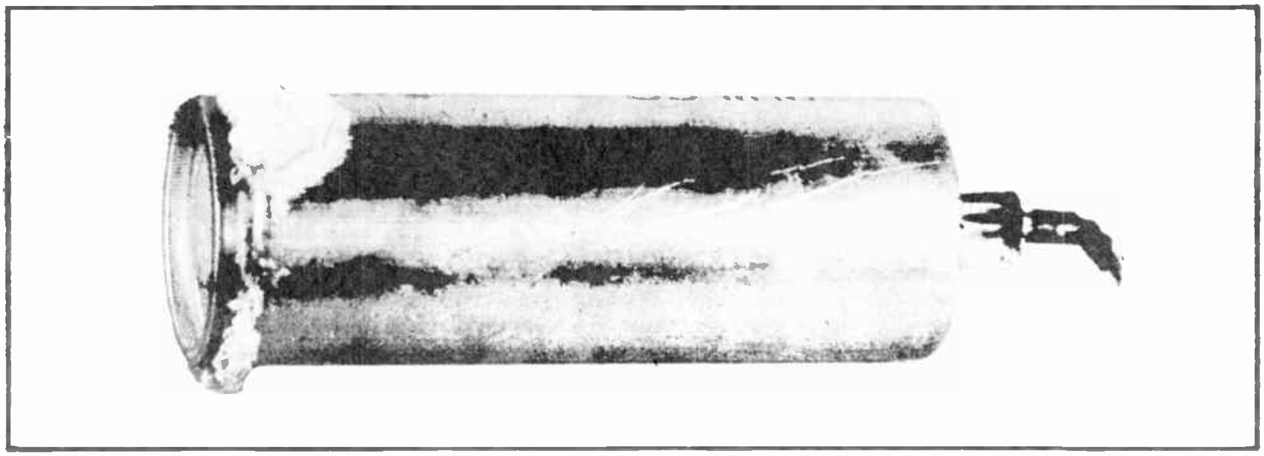


Fig. 13 — 12 How a capacitor may be ruined by overheating or excessive voltage.

The fundamental formula for capacitive reactance is as follows:

$$\text{Reactance, ohms} = \frac{1}{2 \times 3.1416 \times \text{frequency, cycles} \times \text{capacitance, farads}}$$

This formula is not particularly easy to use because all our frequencies in kilocycles and megacycles would have to be changed to equivalent cycles, and all the capacitances would have to be changed to the equivalent fraction of a farad. It is more convenient to use formulas based directly on capacitances in microfarads or micro-microfarads, and on frequency in cycles, kilocycles, or megacycles. Those which follow are among the relatively few formulas really useful when making replacements of capacitors and figuring out the causes for certain troubles in television and radio circuits. The factor 160,000, with its multiples and submultiples, will give answers accurate to within one-half of one per cent. For greater accuracy use 159, 155 and the corresponding multiples and submultiples.

$$X_C \text{ ohms} = \frac{160\ 000}{\text{cycles} \times \text{mfd}} \qquad X_C \text{ ohms} = \frac{160\ 000\ 000\ 000}{\text{cycles} \times \text{mmfd}}$$

$$X_C \text{ ohms} = \frac{160}{\text{kilocycles} \times \text{mfd}} \qquad X_C \text{ ohms} = \frac{160\ 000\ 000}{\text{kilocycles} \times \text{mmfd}}$$

$$X_C \text{ ohms} = \frac{0.16}{\text{megacycles} \times \text{mfd}} \qquad X_C \text{ ohms} = \frac{160\ 000}{\text{megacycles} \times \text{mmfd}}$$

TELEVISION

LESSON NO. 14

FIXED RESISTORS FROM THE SERVICE STANDPOINT

"Training Your Mind"
C. T. I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

COMMERCIAL TRADES INSTITUTE

Training Your Mind — C.T.I. — Perfecting Your Skill

LESSON NO. 14

FIXED RESISTORS FROM THE SERVICE STANDPOINT

The service technician never has a chance to forget the importance of capacitors and resistors, for every time he looks underneath a receiver chassis, as in Fig. 14-1, a large part of the view is taken up with these units. It is essential, of course, to understand the basic principles relating to how these circuits elements do their work. Otherwise the process of trouble shooting would be a hit-and-miss affair. But for service to be satisfactory to the customers, and profitable to you and your organization, it is almost as essential to know how to select and apply such units as are available for replacement and repair. In the preceding lesson we went far toward gaining such a practical understanding in relation to fixed capacitors. Now we shall do the same for fixed resistors, and add a few items which relate to both capacitors and resistors.

A small portion of a receiver service diagram is reproduced in Fig. 14-2. The resistors are shown by the familiar symbol of the zig-zag line. The capital letter R, as used in practically all service diagrams, identifies a replacement part number as that of a resistor. For instance, R109 means resistor number 109 in the parts list for the particular receiver. Near the resistor symbol usually is marked the value of resistance. When no letter follows this resistance value it is the number of ohms. Near resistor R109 at the upper left in Fig. 14-2 is the number 5600, which means 5600 ohms. Farther toward the right are resistors R6 and R7, with their values shown as 10K. The capital letter K means thousands of ohms, so 10K means 10,000 ohms. When the value of a resistance is shown on a wiring diagram by a number followed by the capital letter M it usually means the number of megohms. While this use of the letter M as a symbol for megohms is common practice, especially on the more recent diagrams, you will find some cases where the letter M means thousands of ohms. There is little likelihood of confusion, because you won't find both K and M meaning thousands of ohms on the same diagram. Resistances in megohms often are identified by the abbreviation "meg", which cannot be mistaken in its meaning.

ELECTRIC POWER. Were someone to ask you to name the most important thing about a resistor you would answer, quite correctly, that it is the value of resistance. But were that person to look in the catalogs and other descriptive literature issued by makers and suppliers of resistors he might easily conclude that the most important thing about resistors is "watts", for almost every description is headed by a statement that the resistors are rated at some number of watts or some fraction of a watt. Down below will be the ohms and megohms of resistances available for the specified wattage.

A watt, as undoubtedly you know, is the unit in which electric power is measured. Lamps, toasters, and nearly all other electrical things are marked with the number of watts of power required for their operation. Power is a rate of doing work. In an earlier lesson we talked about work and energy. We talked about how work is done on electrons when they are forced to move against electric fields, and about how electrons do work when they move with the fields. There we learned that mechanical work may be measured in a unit called the foot-pound, which is the amount of work done when a weight of one pound is lifted through a distance of one foot against the force of gravity.

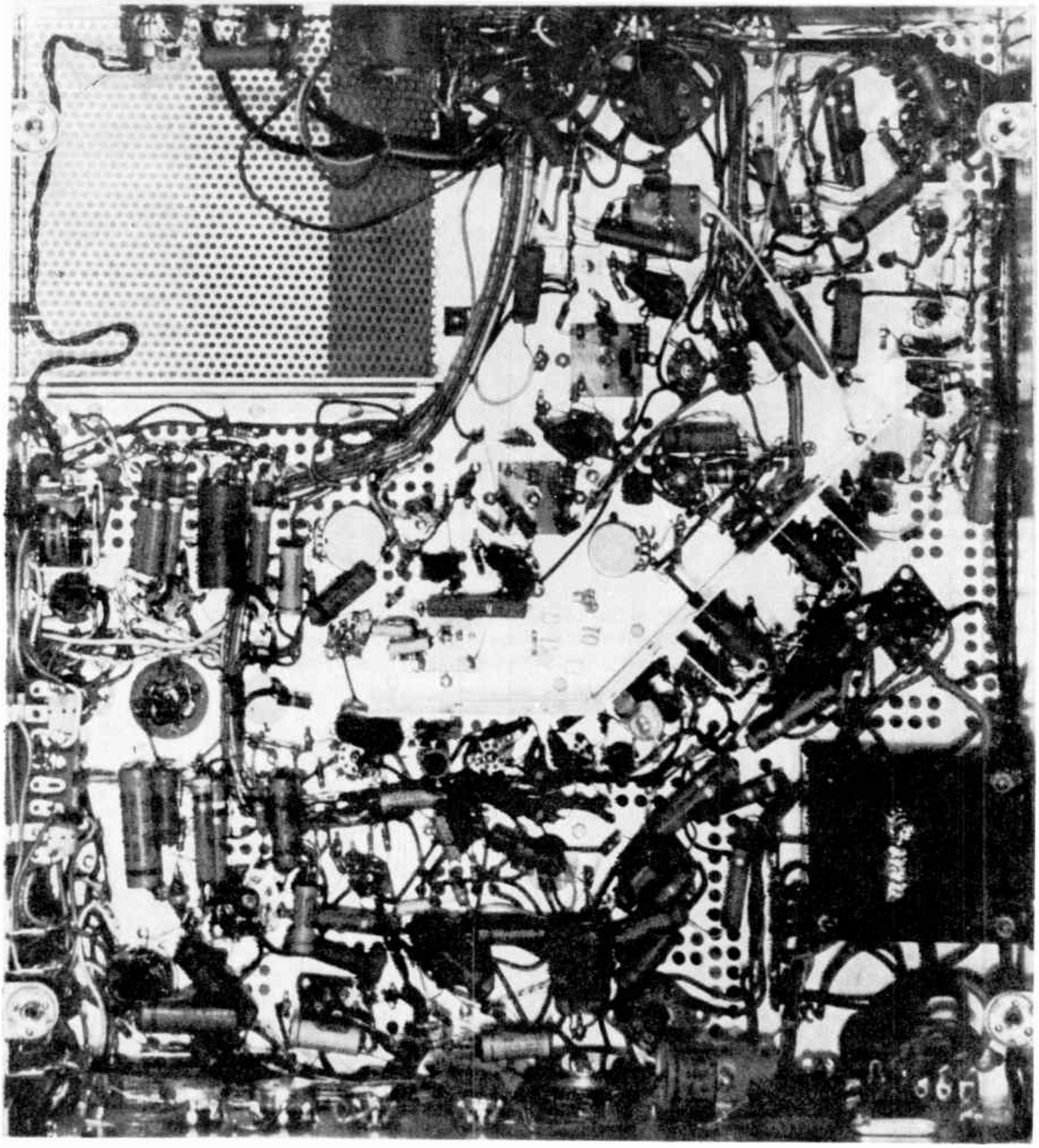


Fig. 14-1. Resistors and capacitors account for many of the problems in television and radio servicing.

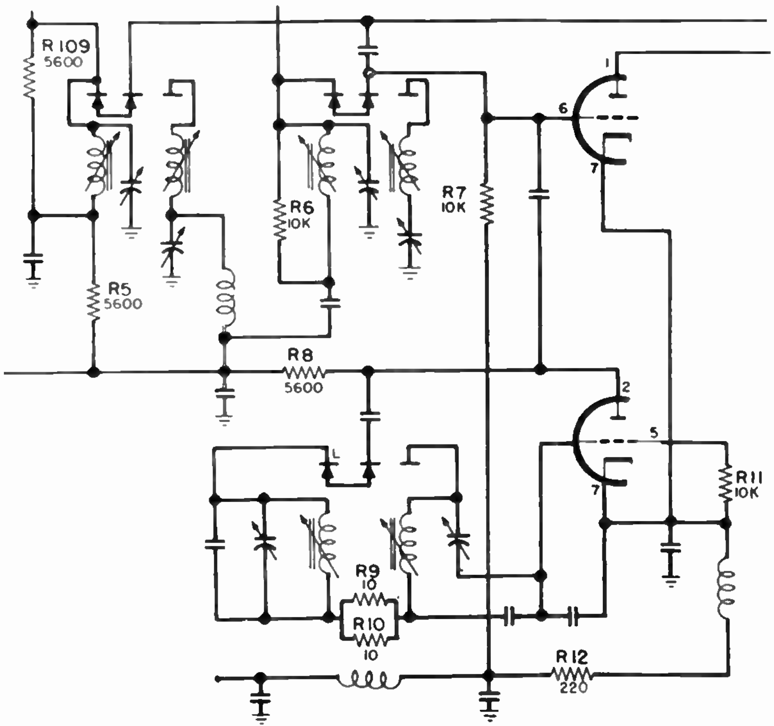


Fig. 14-2. How resistors may be shown in a service diagram.

Were it possible for you to work at the rate of 33,000 foot-pounds per minute you would be working at the rate of one horsepower. Were you to work at the rate of only about 44 foot-pounds per minute you would be working at the rate of one watt of electric power – if you were composed of free electrons rather than of flesh and blood. Only electrons in motion do work at rates measured in watts of electric power.

When electrons move, the work they do may cause mechanical motion, as in motors or in the vibrating cones of loud speakers. The electrons may work also to produce light, as in all kinds of electric lamps. The working electrons may cause chemical changes, or do electroplating, and may do many other things. One of the important jobs for moving electrons is production of heat, as in electric ranges. This heating effect interests us greatly, because every bit of electron work that does not produce motion or light or chemical changes or something else of that nature must produce heat.

In resistors used for television and radio the moving electrons do work. This work doesn't produce motion or chemical changes, and we hope it won't produce light by making the resistors red hot. But the electron work does produce heat in greater or less quantity. If electrons were to work at a rate of one watt of electric power in a resistor, and continue this rate of working for one hour, they would generate enough heat to raise the temperature of one cubic foot of air by 188°F, provided all the heat were retained for raising the air temperature.

Many television receivers require electric power at a rate of about 180 watts for their operation. It takes only a negligible fraction of this power to produce motion in the speaker and light at the picture tube. Prac-

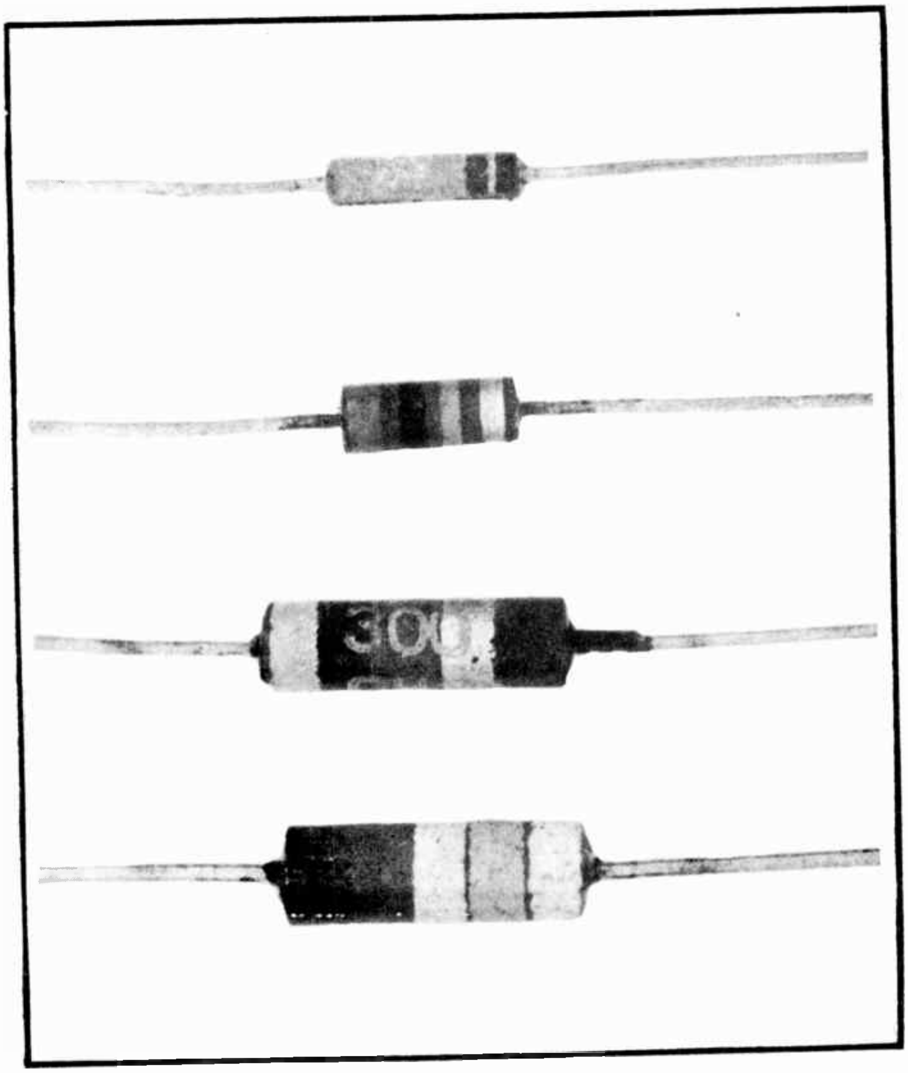


Fig. 14-3. Fixed resistors of composition and wire wound types.

tically all the power ends up by producing heat. If all this 180 watts of heating effect were put into 20 pounds of aluminum, and none escaped, the temperature of the metal would go up 140°F within an hour. Actually a great deal of the heat does appear either in the heaters of the tubes or else in the fixed resistors. Some typical fixed resistors of small size are pictured in Fig. 14-3.

Now let's inquire what is needed in the way of current and potential drop and resistance to produce the electric heat which is measured in watts. The relations are exceedingly simple. When current is one ampere and potential drop is one volt, the rate of power production is one watt. The product of the number of amperes of current and the number of volts potential drop or potential difference is equal to the number of watts of power. You know that current can be one ampere with a potential difference of one volt only when resistance is one ohm. So we may say that there is a power dissipation of one watt when a current of one ampere flows through a resistor of one ohm when the voltage across the terminals of the resistor is one volt.

We may compute the power rate in watts when knowing any two of the three factors, current, potential difference, and resistance. In power formulas we shall use the milliampere as the unit of current, the volt for potential difference, and the ohm for resistance. The simplest power formula requires that we know the current and the potential difference to begin with. Here is the formula.

$$\text{Watts} = \frac{\text{milliamperes x volts}}{1000}$$

By combining this power formula with our earlier formula for voltage, in terms of current and ohms, we can arrive at another power formula with which we need to know current and resistance.

$$\text{Watts} = \frac{(\text{milliamperes})^2 \text{ x ohms}}{1\ 000\ 000}$$

With this second power formula we square the number of milliamperes of current in the resistor, then multiply by the number of ohms resistance of the resistor, and divide by one million to find the number of watts of power being used.

Here, the power formula is combined with our earlier formula for current, in terms of volts and ohms.

$$\text{Watts} = \frac{(\text{volts})^2}{\text{ohms}}$$

In Fig. 14-4 is represented a 5000-ohm resistor in which the measured current is 20 milliamperes, and across which the measured potential difference is 100 volts. You might check this relation between ohms, milliamperes, and volts with some or all our original formulas for resistance, current, and potential difference. Now we shall use each of the three power formulas to determine the rate of watts at which power is being dissipated in this resistor. First, with milliamperes and volts.

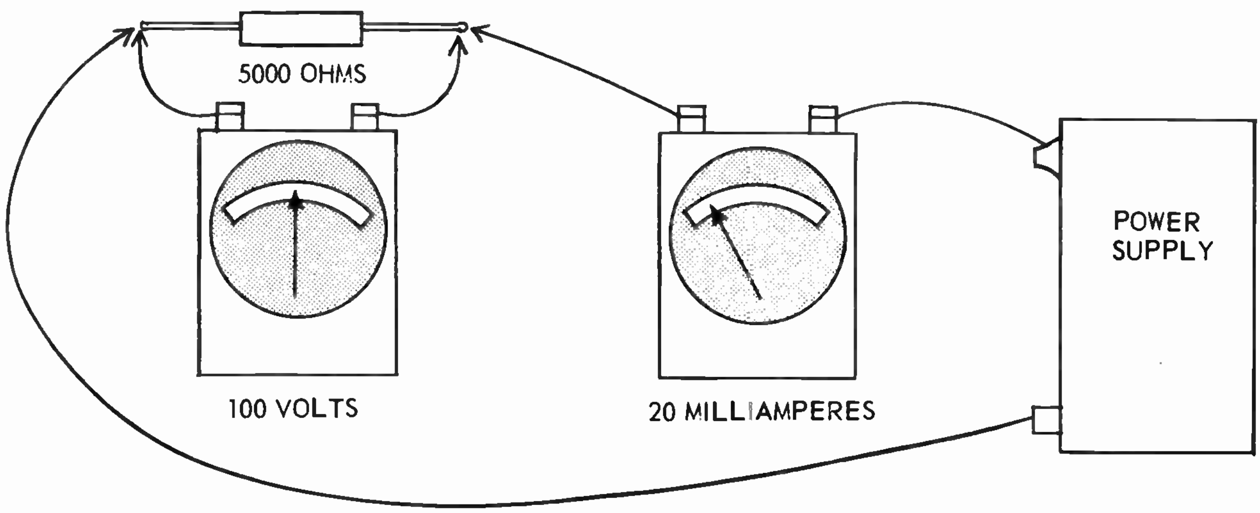


Fig. 14-4. Power in watts may be computed from values of potential difference, current, and resistance.

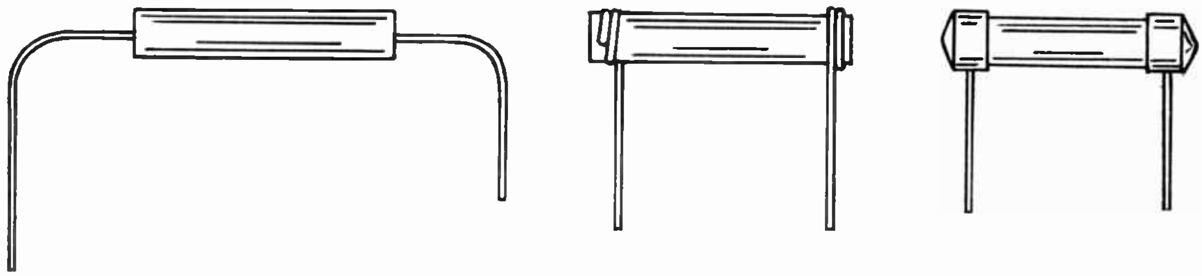


Fig. 14-5. Resistors with axial and radial leads.

$$\text{Watts} = \frac{\text{milliamperes x volts}}{1000} = \frac{20 \times 100}{1000} = \frac{2000}{1000} = 2 \text{ watts}$$

Next, the formula which uses milliamperes and ohms.

$$\text{Watts} = \frac{(\text{milliamperes})^2 \text{ x ohms}}{1\ 000\ 000} = \frac{400 \times 5000}{1\ 000\ 000} = \frac{2\ 000\ 000}{1\ 000\ 000} = 2 \text{ watts}$$

And finally, the formula with which we must know volts and ohms.

$$\text{Watts} = \frac{(\text{volts})^2}{\text{ohms}} = \frac{10\ 000}{5\ 000} = 2 \text{ watts}$$

HEAT DISSIPATION IN RESISTORS. Every bit of power used in a resistor goes into the production of heat. Unless this heat is dissipated or gotten rid of, the resistor eventually will burn up. Remember, power is a rate of doing work, it is a measure of work continued through time, and power in a resistor is a measure of the rate of heat production. A watt of power doesn't measure a definite quantity of heat, in the way that a gallon measures a certain quantity of water. A watt measures a rate of heat production, as gallons per minute measure a rate of water flow. Heat raises the temperature of any substance in which the heat is being produced, and so long as heat continues to be produced by continued use of power, the temperature will continue to go up unless heat is allowed to escape from that substance.

Fortunately, the higher the temperature of a resistor, and the greater the difference between its temperature and temperature of surrounding air and other bodies, the more rapidly heat will pass from that hot resistor into the cooler things around it. When the resistor produces heat due to power dissipation the temperature will rise and the rate of heat loss will increase until heat is passing out of that resistor at the same rate heat is produced within it. Only then will the temperature of the resistor become constant. Should the resistor become so hot as to melt or burn up before the condition of heat balance is reached, that's just too bad for the resistor and the apparatus in which it is connected.

Heat passes out of a hot resistor through its surface and through whatever terminal connections may be used. About the only practical way to increase the rate of heat loss in a given style of resistor is to increase its surface area. As a consequence, the greater the power and the heat which a resistor must dissipate, the larger must be the resistor.

RESISTOR RATINGS IN WATTS. The watts rating or power rating of a resistor is the maximum number of watts of power which may be dissipated in the resistor without raising its temperature to a point at which the resistor itself might be damaged or a point at which its resistance might vary beyond the percentage limits of tolerance. How much heat and temperature a resistor will stand depends on its construction and size. The rating is based on having the resistor supported in free air with at least a foot of clear space in every direction – a condition never realized in television and radio receivers.

As resistors actually are used and mounted they have almost no free air space and little ventilation. Then the temperature limit will be reached with power much less than the rating. Furthermore, the maximum temperature which the resistor will stand almost always is too high for capacitors, wiring, and other parts

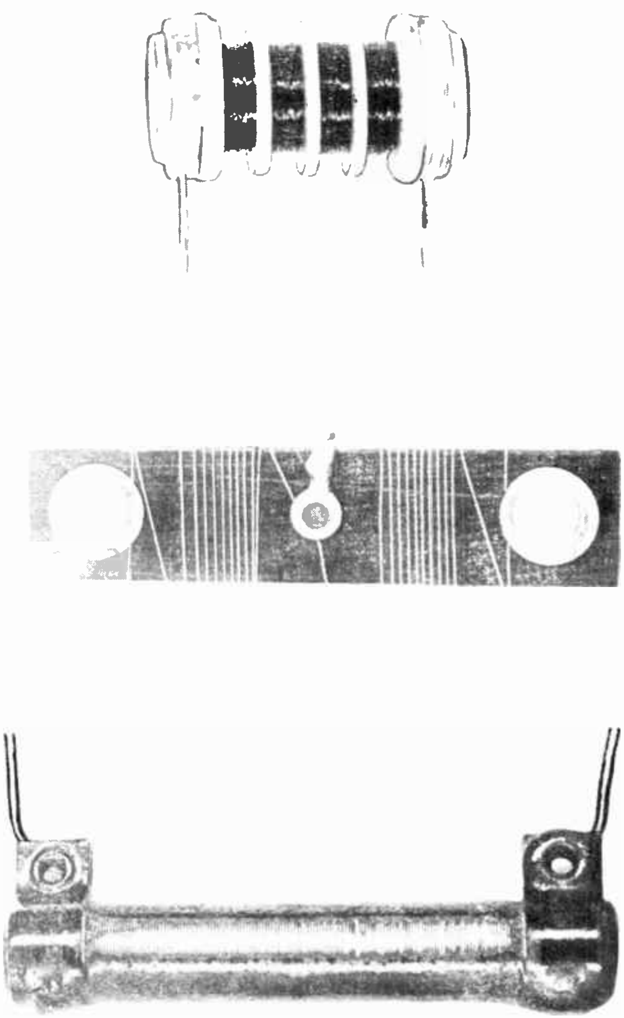


Fig. 14-6. Types of wire wound resistors.

which are near the resistor. These other parts might be charred, or their insulation might melt out. In practice, you never should use a resistor rated for less than double the actual power as computed from current, potential drop, and resistance. Even then the resistors will run very hot when mounted in the usual manner. It is preferable to use resistors rated at three, four, or five times the actual power that they will have to handle. The greater the rated watts in comparison with actual watts,the cooler the resistor will operate. The resistor of higher watts rating is larger, and the greater surface allows more rapid dissipation of heat.

RESISTANCE MATERIALS. Composition resistors, called also carbon resistors, have for their resistance elements a mixture of carbon or graphite with clay, ceramics, or phenolic compounds in such proportions as to give the desired combination of resistance and size. Nearly all composition resistors now manufactured are encased by an outer insulating tube sealed at both ends, and usually are impregnated with some moisture-proofing material.

Other resistors are of the metallic film type of the metallized type. The resistance element is a small-diameter glass tube on the surface of which is baked a thin coating of metal to provide the necessary conductivity and resistance. This tube is enclosed by an outer insulating tube or may be molded into a covering of phenolic insulating material.

The standard resistance tolerance for composition and metallized resistors most often is $\pm$ 10%. A tolerance of $\pm$ 5% is available at somewhat greater cost. Many resistors which sell for lower prices have standard tolerance of $\pm$ 20%. In the case of many resistors which sell at bargain prices the marked value of resistance may have little relation to the actual resistance. Such units have to be checked with an ohmmeter before being put into service.

Insulated resistors are provided with wire leads or pigtails extending straight out from the ends. These are called axial leads, because they are in line with the axis of the cylindrical resistor. With older types of uninsulated resistors the leads extend out at right angles from both ends of the cylindrical body. These are called radial leads. The differences are illustrated by Fig. 14-5. At the left is an insulated resistor with axial leads. On the uninsulated type at the center the leads are twisted around the ends. On the very old style at the right the leads are attached to two metal end caps. The central cylindrical bodies of the uninsulated types are almost always coated with or embedded in insulation, the uninsulated part consists of the exposed leads around both ends which easily may touch any nearby conductors to cause short circuiting or accidental grounding.

Resistors of the composition and metallic film type most often are used in power ratings of 1/2 watt, 1 watt, and 2 watts, but occasionally are found in 3-watt and 5-watt sizes. Resistors whose power ratings are 5 watts or more nearly always are of the wire wound type. Wire wound resistors are available also in the smaller ratings, all the way down to the 1/2-watt size.

There are many varieties of wire wound resistors. At the top of Fig. 14-6 is a style having many turns of enamel insulated wire wound into slots along the supporting form. At the center is a resistor formed by a few turns of exposed wire around a strip of insulation. This is a center-tapped resistor, with a connection at the center of the winding in addition to connections at both ends. Down below is a vitreous enameled wire wound resistor from which some of the enamel coating has been removed to expose the turns of wire. Other wire wound resistors are of the same general appearance as the units pictured in Fig. 14-3.

Vitreous enameled wire wound resistors are made with wire of high resistance per foot wound onto a tube or flat strip of heat-resistant insulating material. The assembly is coated with an enamel and is fired to red heat in furnaces. The enamel fuses into a glass-like (vitreous) coating. These resistors are available with power ratings from 5 watts to 200 watts and in resistances from about 0.2 ohm up to more than 200,000

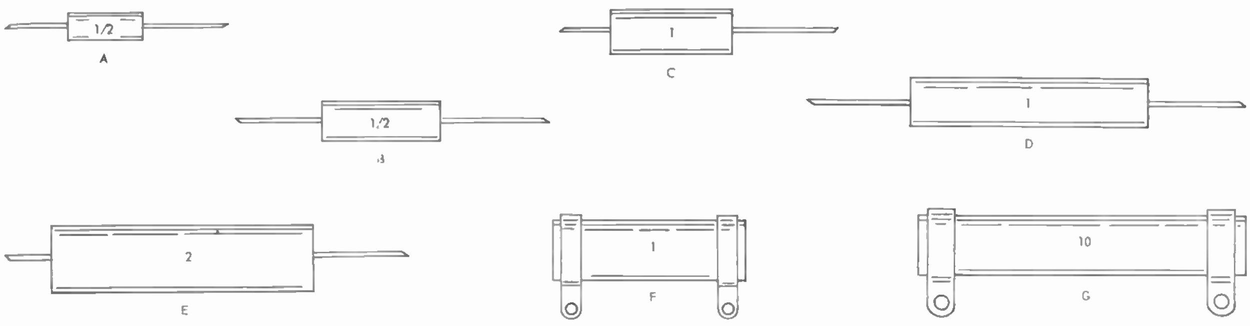


Fig. 14-7. Relative sizes of resistors having various wattage ratings.

ohms. There are also heat-resistant enamels other than the vitreous kind. Then too, there are so-called low-temperature enamel coatings which are not fired during their formation but are amply able to withstand the highest temperature at which a resistor may operate.

A wire wound resistor of any ordinary type consists of a coiled conductor on an insulating form. Coiled conductors are used in television and radio circuits to provide the property called inductance. Inductance is no more desirable in resistors than it is in capacitors, but all the common types of wire wound resistors are inductive and are likely to cause trouble at radio frequencies. There are several ways of winding wire to avoid most of the inductive effect, and for some applications it is necessary to use such non-inductive wire wound resistors.

RESISTOR DIMENSIONS. In Fig. 14-7 are outline drawings showing relative lengths and diameters of resistors in the various wattage ratings marked on the drawings. The 1/2 watt resistor at A measures approximately 3/8 inch long and 1/8 inch in diameter. This size formerly was rated at 1/4 watt until improvements in materials and design made the higher rating possible. The size at A is found in composition and metallic film units. Wire wound resistors of 1/2 watt rating have approximate dimensions shown at B, 5/8 inch long and about 3/16 inch diameter. At C is shown the relative size of 1-watt composition and metallic film resistors, about 5/8 inch long and 1/4 inch in diameter. The outline at D shows the relative size of 1-watt wire wound resistors having axial leads, the length being about 1 1/4 inches and the diameter about 1/4 inch. Dimensions of 2-watt composition, metallized, or wire wound resistors with axial leads are shown at E. The length is about 1 3/8 inch and the diameter about 3/8 inch.

At F in Fig. 14-7 is represented a wire wound resistor of 1-watt rating having radial lugs, and at G is shown the relative size of a 10-watt vitreous enameled wire wound resistor. In all the resistors whose relative dimensions have been shown, the size is determined by the power rating in watts and not by the resistance in ohms. The only time when resistance affects size is in the case of a wire wound unit of such high resistance as to require a form which will accommodate the extra length of wire.

There are many circuits in television and radio receivers which require the use of a number of resistances connected in series with one another. This need often is met by using tapped resistors of the wire wound type, like the two units pictured in Fig. 14-8. The resistance sections may be of equal or unequal values. At a point between each pair of adjacent sections along the wire winding is connected a terminal lug as shown, or sometimes a wire lead.

PRECISION RESISTORS. Fixed resistors whose tolerance in range of resistance is less than $\pm$ 5% commonly are called precision resistors. Such units are available in all the usual constructions and in a number of special forms. Carbon composition units often are used, as are also some special forms of metallic film construction, and wire wound types with various insulations. The resistor at the top of Fig. 14-6 is a precision type, with sections of wire wound in slots on a ceramic insulation form. This and other special methods of winding are used to lessen the inductive effective.

Precision resistors most often are used in meters and other service instruments where accuracy is important. Tolerance of $\pm$ 1% is the usual standard, although $\pm$ 2% may be used, and in some instruments we find high-precision tolerances such as 1/2%, 1/4%, or 1/10%, all plus or minus. The wire or other resistor material is of such kinds as will undergo relatively small change in its resistance value when there are changes of temperature such as ordinarily occur in occupied buildings. In the various makes and styles of precision resistors it is possible to obtain resistances all the way from about 0.1 ohm up to 15 megohms.

Power ratings of precision resistors are available in a range from 1/2 to 5 watts, with 1 watt probably the most common. Compared with ordinary types, the precision resistors nearly always are oversize for their power rating. That is, where full power dissipation in an ordinary resistor might raise its temperature by 250 to 450 degrees Fahrenheit, full rated power in a precision unit may raise the temperature by no more than 75 degrees. A 1-watt precision resistor may be as large as a 5-watt or 10-watt unit with commercial tolerance.

RESISTOR TROUBLES. When a resistor is found burned out, as evidenced by an open circuit, there must have been a reason for the damage. Every effort should be made to discover the reason, either before replacement or immediately after replacement and before leaving the job. In some early production runs of receivers there may be resistors of too low power rating. After installing a new resistor of the correct number of ohms, measure the voltage drop across this resistor. Then use the measured voltage drop and the number of ohms in the formula for watts. This will give the actual power dissipation. The resistor must have a rating several times as high.

The resistor may have been located too close to hot parts such as rectifier tubes and output amplifier tubes, or the resistor may have been pushed into a position where there is too little air circulation. If a new resistor appears to be over-loaded according to the check with the voltage drop and watts formula, make a careful examination for short circuits and accidental grounds in circuits connected to the resistor. Such troubles may allow excessive voltage and current.

A resistor may be cracked or its terminal leads may be loosened or broken by abuse during service operations. Enamel coatings may have been chipped enough to expose the wire and allow short circuits or accidental grounds. Some coatings may be damaged by overheating, the effects of which may show up as rough, bubbly, or dark spots on the insulating cover.

It is a well known fact that all pure metals increase their resistance with rise of temperature, while pure carbon and graphite lower their resistance when temperature goes up. Good quality resistors of all types are practically unaffected by "temperature coefficients of resistivity" which are measures of the effect of temperature on resistance. The resistor materials are so selected that whatever small change of resistance does occur with rise or fall of temperature will be well within the rated tolerance of the resistors.

PREFERRED NUMBERS OR VALUES. Now we are going to talk about something which applies equally to fixed resistors and fixed capacitors. It is the values of capacitance and of resistance which are becoming standardized and which are most generally available from radio supply houses. The object of standardizing on certain values is to reduce the stock of resistors and capacitors required when carrying on a service business and to reduce the varieties which have to be manufactured, which brings down the cost per unit.

Supposing you were going to use resistors, or capacitors, of 20% tolerance. How many values would you have to have on hand to cover the entire range from 8 to 80 ohms or 8 to 80 mmfd or any other range whose low and high limits have the ratio of one to ten? You would need only six different values. Here they are.

10 15 22 33 47 68

You cannot possibly name any number between 8 and 80 which is not within the range of 20% below to 20% above one of these six numbers. If you don't believe it, try it. Even though you pick some number such as 26.4 ohms or mmfd, it is between the preferred values of 22 and 33. If you add 20% of itself to 22 you get 26.4. If you deduct 20% of itself from 33 you get 26.4. If you wish to extend your stock to cover, with 20% tolerance, everything from 80 to 800 ohms of mmfds you will need to add only six more preferred values, which are these.

100 150 220 330 470 680

If you multiply this last set of preferred numbers by 10 you will have values going all the way from 800 to 8000, with 20% tolerance, and so you may keep on until with a stock of only 36 different values you can handle every job calling for resistance or capacitance between 8 and 8,000,000 ohms or mmfd, provided the tolerance is ± 20%.

If you want to handle every replacement requiring tolerance of ± 10% it is necessary to add only six more values in each range where the ratio of minimum to maximum is one-to-ten, or to add another 36 values for everything between 8 and 8,000,000 ohms or mmfd. To get down to a tolerance of ± 5% isn't quite so easy, but it requires adding only 12 more values in each one-to-ten range.

The accompanying list shows the preferred or standard values in three columns. Each column covers only one range with a low to high ratio of one-to-ten, but the values in each column may be multiplied or divided by 10, 100, 1000 or any other multiple of 10 to extend the list as far as you like. Here we have all the significant numbers which you will see so often as the values of resistors and capacitors on diagrams for radio and television service work.

Resistors and capacitors in values other than those of the preferred numbers are made in all styles of units. Wire wound resistors in power ratings of 5 watts and up nearly always have values in round numbers, such as 5, 10, 12, 15, 20, 25 ohms and so on. Values of precision resistors do not follow the preferred numbers. Fixed capacitors are more common in values such as 10, 20, 50, 100 mmfd, and for larger sizes in values like 1, 2, 4, 8, 12 mfd than in the preferred numbers. The principal use for the preferred values is in resistors whose power ratings are from 1/2 to 2 watts.

PREFERRED OR STANDARD VALUES

Tolerance 20%	Tolerance 10%	Tolerance 5%
10 ----------	----------	----------
		11
	12 ----------	----------
		13
15 ----------	----------	----------
		16

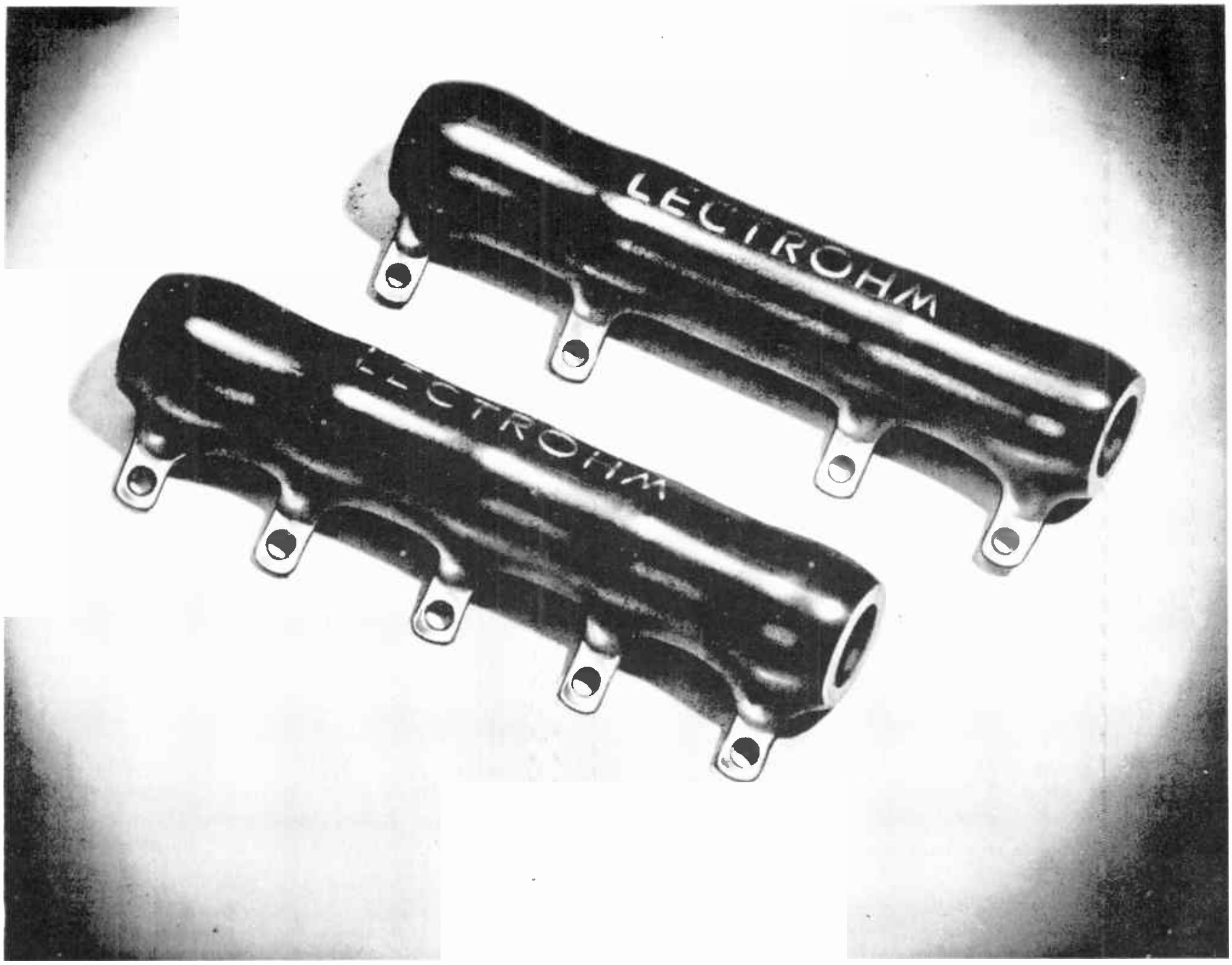


Fig. 14-8. Wire wound vitreous enameled resistors with tap connections.

18 ----------------

20

22 -------------------------------

24

27 ----------------

30

33 -------------------------------

39 – – – – – – – – – – – – – –

43

47 –

51

56 – – – – – – – – – – – – – –

62

68 –

75

82 – – – – – – – – – – – – – –

91

100 –

In theory it is permissible to replace a resistor with any other resistor whose actual value is within the tolerance limits of the one replaced. For example, a resistor rated at 47 ohms with 20% tolerance might have actual resistance anywhere from 37.6 to 56.4 ohms, and could be replaced with a unit of 40, 45, 50 or 55 actual ohms, or of any other actual resistance between the tolerance limits. The same principle applies to replacement of 10% tolerance units and of 5% tolerance units, and applies to capacitors as well as to resistors. The danger in making such replacements is that the original unit may have been selected for some resistance which must be accurate within closer limits than the marked tolerance. For instance, the original resistor might have been picked because it measured between 40 and 45 ohms, even though marked 47 ohms, and replacement with something like 50 or 55 ohms might lead to difficulties.

COLOR CODES. If you examine the composition and metallized film resistors shown by Fig. 14-3 you will notice a number of bands around the cylindrical bodies. These bands are of various colors. The colors and their relative positions indicate the resistance in ohms, and may show also the tolerance. A somewhat similar system of colored bands is used on small cylindrical capacitors to show the capacitance, the tolerance percentage, and in some cases the temperature coefficient. On flat rectangular mica capacitors or the similarly shaped non-inductive paper capacitors the identifying colors are applied in small round dots such as you can see on the units of Fig. 14-9.

Fortunately, the colors which indicate the numerical digits from 1 to 9, and which indicate how many ciphers are to be used, these colors are the same for both resistor and capacitor color codes. The colors and their meanings are important in the business of servicing, for oftentimes you will have no other means for learning the value of a small resistor or small capacitor. Here are the colors and the digits for which they stand.

Black	0	Green	5
Brown	1	Blue	6
Red	2	Violet	7
Orange	3	Gray	8
Yellow	4	White	9

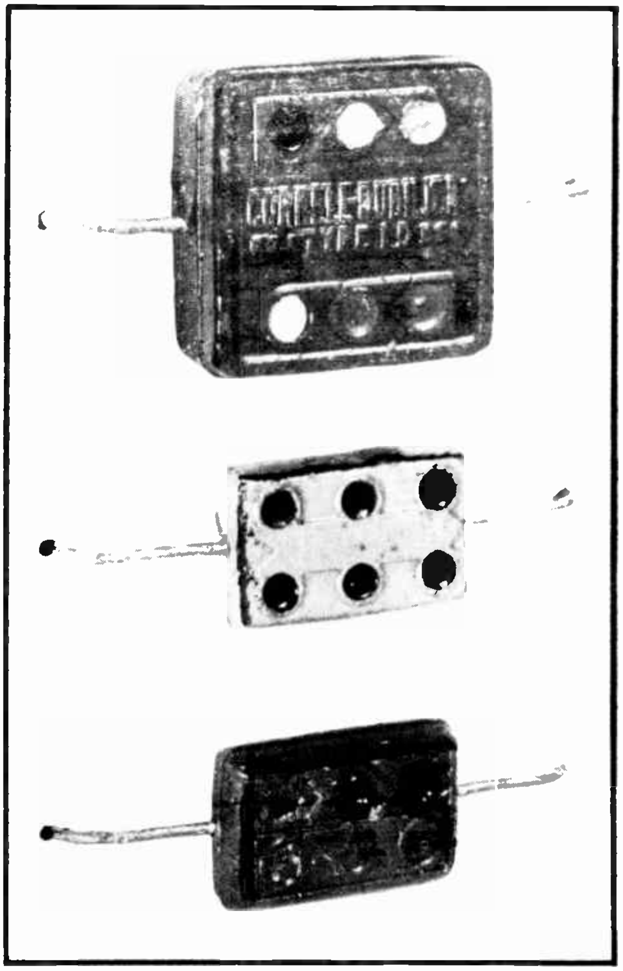


Fig. 14-9. Fixed mica capacitors with color code markings.

The colors from red through violet are arranged as they occur in the rainbow, but this does not help to remember the order of black, brown, gray, and white – and some of us don't remember the order of colors in the rainbow spectrum. To make it easier to remember all the colors of the code, most of our students memorize the following truthful statement in which the initial letters of successive words are the same as the initial letters of the names of colors in their correct order in the code.

C.T.I. training	Brings	Better	Rewards	Once	You	Gain	By	Very	Good	Work
	black	brown	red	orange	yellow	green	blue	violet	gray	white
	0	1	2	3	4	5	6	7	8	9

Or you may look at it this way.

"C. T. I. training	Brings	black	0
	Better	brown	1
	Rewards	red	2
	Once	orange	3
	You	yellow	4
	Gain	green	5
	By	blue	6
	Very	violet	7
	Good	gray	8
	Work	white	9

With this system of code colors, and nothing more, each digit in a number would require the use of a separate band or dot. A value of 1,000,000 ohms would call for one brown band or dot, followed by six black bands or dots to indicate the six ciphers. This is avoided by using each of the colors to indicate whether a number consisting of two digits is to be multiplied by 10, 100, 1000 or other multiple of 10, or is to be divided by 10 or 100. Then a value of 1,000,000 would be shown as the number 10 (two digits) multiplied by 100,000. To multiply by 100,000 we may add five ciphers. The color which represents 5 in our code is green, so we use green to indicate multiplication by 100,000 or to indicate the addition of 5 ciphers when the green band or dot is in a position which shows it to be a multiplier. All the other colors are used to indicate their appropriate number, or corresponding single digit. Following is a list showing for each color the corresponding single digit and also the number of added ciphers or the multiplier indicated by the color.

COLOR CODE FOR RESISTORS AND CAPACITORS

Color	*When Used In Digit Positions*	*When Used In Multiplier Position* Multiply By	Ciphers To Be Added
Black	0	1	none
Brown	1	10	one or 0
Red	2	100	00
Orange	3	1000	000
Yellow	4	10,000	0 000
Green	5	100,000	00 000
Blue	6	1,000,000	000 000
Violet	7	10,000,000	0 000 000
Gray	8	100,000,000	00 000 000
White	9	1,000,000,000	000 000 000
Gold		1/10	
Silver		1/100	

Fig. 14-10 shows the positions of the color bands on fixed resistors having axial leads. The present standard arrangement is shown at the left, where there are four bands. Colors in the first two bands, commencing always at the band nearest one end of the resistor, indicate the first two digits in the number. The color in the third band indicates the multiplier, or how many ciphers are to be added after the two digits. The fourth band shows the resistance tolerance, thus:

Gold	indicates tolerance of	± 5%
Silver	indicates tolerance of	± 10%
No color	indicates tolerance of	± 20%

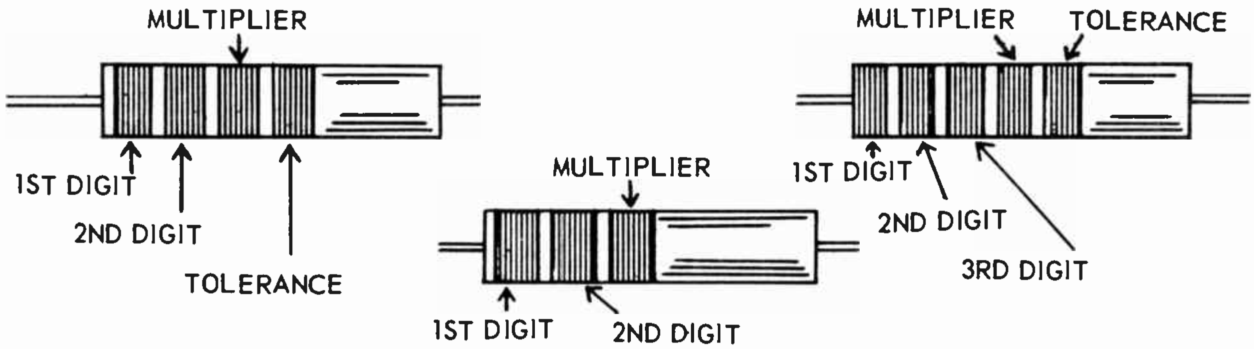


Fig. 14-10. Color code bands as used on fixed resistors having axial leads.

Some resistors will have only three color bands, as shown at the center of Fig. 14-10, when no tolerance is indicated or when there is no color indicating the tolerance. This is the case where tolerance is $\pm$ 20% .

Resistances to be shown within very close limits may require five bands, as at the right, with the first three bands indicating three digits or three significant figures in the value.

Following are some combinations of colors, together with resistances and tolerances which they represent. Examine each combination with care, making sure that you understand how the colors are used to indicate desired values.

OHMS -------	15	350	6800	75,000	400,000	2,200,000
TOLERANCE ----	5%	10%	20%	5%	10%	20%
1st Band	brown	orange	blue	violet	yellow	red
2nd Band	green	green	gray	green	black	red
3rd Band	black	brown	red	orange	yellow	green
4th Band	gold	silver	none	gold	silver	none

Here is another set of color code combinations in which the digits always are the same, with changes shown entirely by multipliers and tolerances.

OHMS -------	47	470	47,000	4,700,000	4.7	0.47
TOLERANCE ----	10%	10%	5%	5%	10%	5%
1st Band	yellow	yellow	yellow	yellow	yellow	yellow
2nd Band	violet	violet	violet	violet	violet	violet
3rd Band	black	brown	orange	green	gold	silver
4th Band	silver	silver	gold	gold	silver	gold

This business of color coding may seem a rather difficult part of the language of radio and television. But after you work with resistors for a while, a string of color bands will look just like a number. Three red bands will look just as much like twenty-two hundred as do the figures 2200.

At the left in Fig. 14-11 are color code arrangements used on fixed resistors having radial leads. The first digit is indicated by the color of the main body of the resistor. The second digit is shown by the tip color. The multiplier is shown by a dot or a band. All radio men used to remember that "body – tip – dot" meant first, second, and multiplier.

At the right are color band positions used on older fixed resistors having axial leads. Such markings will be found in a few older radio sets. Here again the color of the resistor body indicates the first digit.

CAPACITOR COLOR CODES. Fig. 14-12 shows positions for color code dots on small mica and paper dielectric capacitors encased in insulation of rectangular shape. At the upper left are positions for the old six dot RMA system frequently found, and still used, in radio and television receivers. This coding allows for three numerals, or digits, in the capacitance value; also, for a multiplier, a capacitance tolerance, and a voltage rating in D. C. volts. At the bottom of Fig. 14-12 is a three-dot coding showing only two digits and a multiplier. Capacitors thus coded are supposed to have tolerance of ± 20%, and to have D. C. working voltage rating of 500 volts.

At the upper right in Fig. 14-12 are shown positions for the new RMA and JAN (Joint Army-Navy) capacitor color code. Capacitors coded in this system are distinguished by the fact that the upper left-hand dot will generally be black or white. If silver is used in the upper left hand corner, it would indicate an AWS (American War Standards) code, and the capacitor would be a paper molded rather than a mica molded.

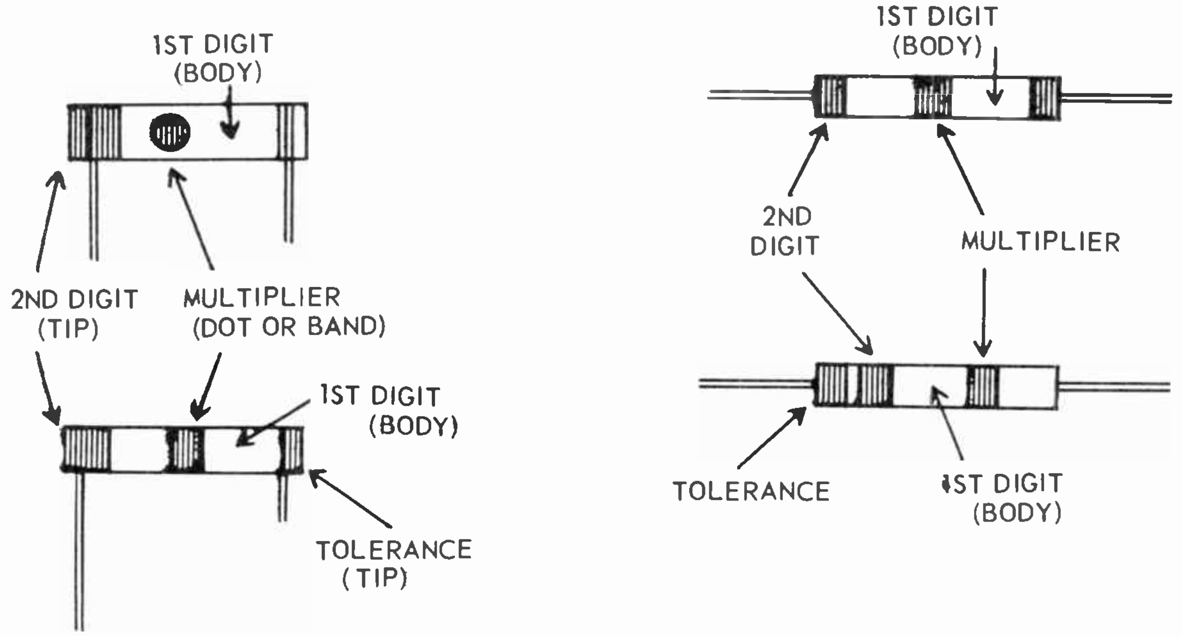


Fig. 14-11. Color code bands and dots on fixed resistors having radial leads.

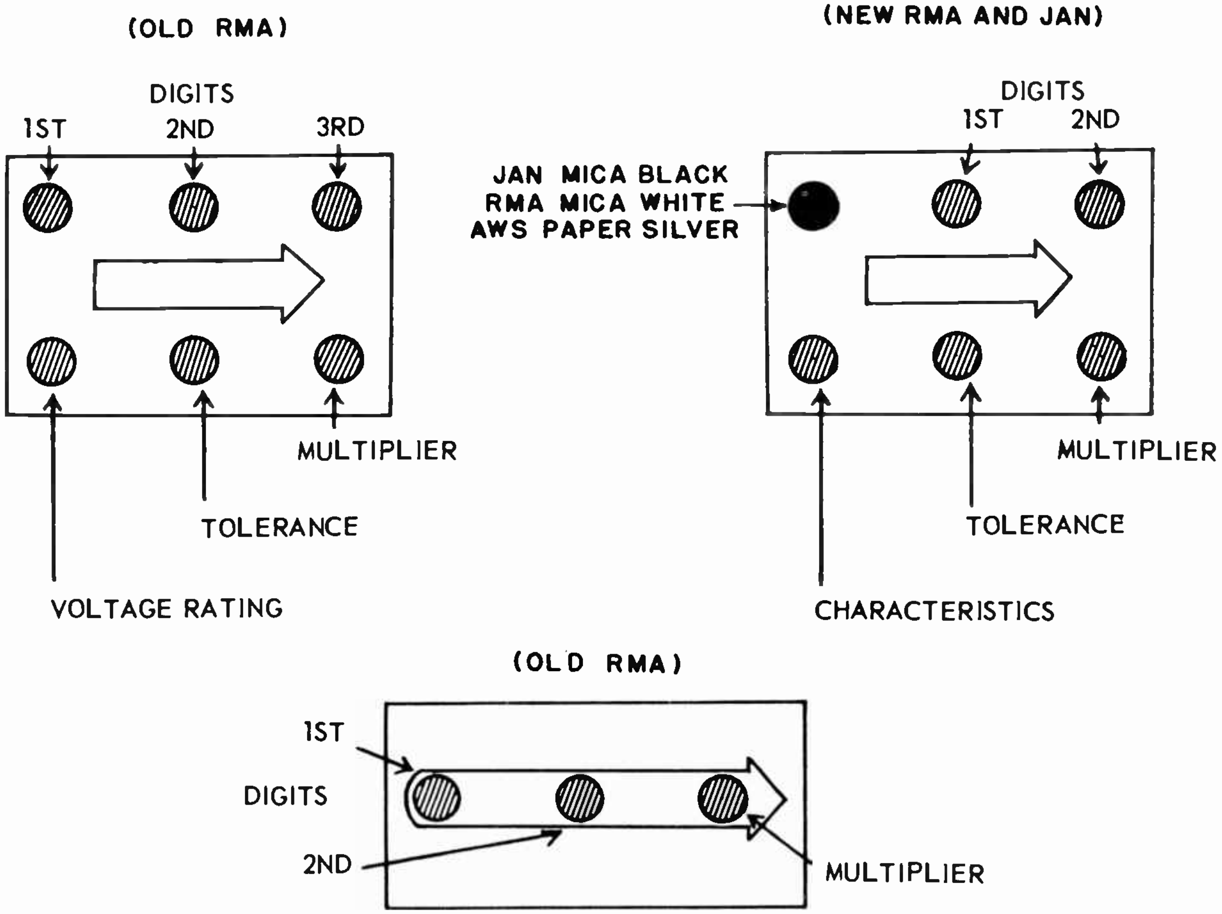


Fig. 14-12. Color code dot positions for mica and molded paper capacitors.

To read the dot indications in correct order, any of these capacitors must be held so that an arrow or anything resembling an arrow in form will point toward your right, or so that the name of the manufacturer or any other wording is right side up. Capacitance values are to be read in micro-microfarads.

One of the accompanying tables shows the meanings of all the colors in their various positions on mica and molded paper capacitors coded according to the RMA or JAN systems. Digit numerals and also multipliers are the same as for resistor coding. Note that code designations for tolerances are not the same in all cases for the RMA and JAN systems. Characteristics relate chiefly to temperature coefficients. These coefficients do not indicate suitablity of a capacitor for temperature compensation; they merely show how much or how little the capacitances may be expected to vary from rated and tolerance value when there are changes of temperature.

MICA AND MOLDED PAPER CAPACITORS COLOR CODE

COLOR	DIGIT NUMERAL	MULTIPLIER (OR CIPHERS)	TOLERANCE	WORKING VOLTAGE	CHARAC- TERISTIC*
Black	0	1	20%		A
Brown	1	10	1%	100	B
Red	2	100	2%	200	C
Orange	3	1,000	3%	300	D
Yellow	4	10,000	4%	400	E
Green	5	100,000	5% (RMA ONLY)	500	F
Blue	6	1,000,000	6%	600	G
Violet	7	10,000,000	7%	700	
Gray	8	100,000,000	8%	800	
White	9	1,000,000,000	9%	900	
Gold		1/10	5% (JAN ONLY)	1000	
Silver		1/100	10%	2000	
No color			20% (RMA OLD)	500	

* Characteristic	Temperature Coef. mmfd per mmf per °C.	Capacitance Change
A (black)	+ 100	– – –
B (brown)	Not specified	(may indicate a loss factor)
C (red)	– 200 to + 200	$\pm$ 0.5%
D (orange)	– 100 to + 100	0.2%
E (yellow)	– 20 to + 100	0.05%
F (green)	0 to + 50	0.025%
G (blue)	0 to – 50	0.025%

Fig. 14-13 shows color coding used on tubular and cylindrical ceramic capacitors. At one end, usually out at the extreme end, is the color which shows temperature coefficient applying to the use of these units for temperature compensation when they are of suitable characteristics. Then follow four spots of color, which may be round dots or may be bands, but often are best described as small "blobs" of color. The first two colors refer to the first and second digits in the capacitance value. Then comes the color for a multiplier, and finally the color for tolerance.

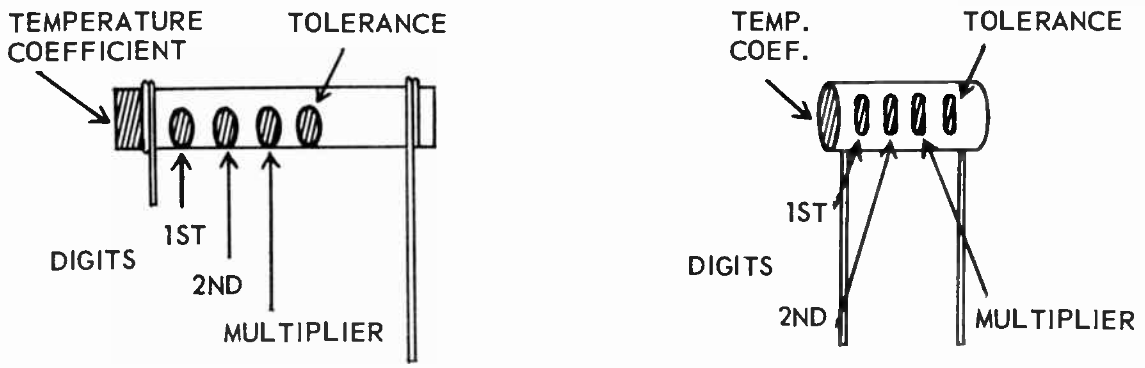


Fig. 14-13. Positions of color code markings on ceramic capacitors.

CERAMIC CAPACITOR COLOR CODE

COLOR	DIGIT NUMERAL	MULTIPLIER	TOLERANCE ± If more than 10 mmfd	If 10 mmfd or less	Temperature Coefficient
Black	0	1	20%	2.0 mmfd	Zero
Brown	1	10	1%	0.1 mmfd	– 30
Red	2	100	2%	0.2 mmfd	– 80
Orange	3	1000			– 150
Yellow	4				– 220
Green	5		5%	0.5 mmfd	– 330
Blue	6				– 470
Violet	7				– 750
Gray	8	1/100			+ 30
White	9	1/10	10%	1.0 mmfd	Wide variation

Note: No color in the tolerance position may mean that the capacitance is not less than the rated or marked value.

Temperature coefficient is the change in mmfd per mfd of capacitance per centigrade degree of temperature change.

Another of the accompanying tables shows how color coding is applied to ceramic capacitors of tubular or cylindrical shape. Note that tolerance here is in plus or minus percentage for capacitors whose capacitance is more than 10 mmfd, and is in plus or minus fractions of, or whole numbers of micro-microfarads for units whose capacitance is 10 mmfd or less.

Many capacitor color codings other than those shown were used before the RMA undertook standardization in this field, and some non-standard features still are in use. Such non-standard markings would be difficult to interpret without many additional tables or charts, and in any case the replacements would have to be made with newer units having standard markings or else plainly numbered with their values.

It should go without saying that much of this lesson consists of reference data. Until you are working with resistors and capacitors, and have almost hourly need for deciphering the color codes, it would be largely a waste of time to attempt committing the details to memory. It may be well to remember the sentence which helps tie the colors to their numbers, but tolerances, coefficients, and variations between coding systems will come without effort later on.

DO NOT TEAR — CUT ALONG THIS LINE AND SEND IN FOR GRADING.

TELEVISION

LESSON NO. 15 GRID CIRCUITS AND GRID BIASING

Training Your Mind

C.T.I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

COMMERCIAL

TRADES

INSTITUTE

Training Your Mind

C.T.I.

Perfecting Your Skill

Chicago, Ill.

LESSON NO. 15

GRID CIRCUITS AND GRID BIASING

In one of the earlier lessons we talked about grid voltage and about grid bias. There it was stated that grid bias is the potential difference which exists between control grid and cathode of a tube when there is no varying signal voltage applied to the grid. We learned that bias voltage may be such as to make the grid either negative or positive with reference to the cathode. We learned also that with no biasing voltage at all, the average potential of the grid remains equal to cathode potential. This is the condition of zero grid bias. Another important thing brought out in that earlier lesson is this: The potential of the grid, whatever it may be at any instant, always is measured with reference to the potential of the cathode. When we say that a grid is positive or is positively biased, we mean that the grid is positive with reference to the cathode in the same tube, or is more positive than that cathode. When we say that a grid is negative or is negatively biased, that grid is negative with reference to the cathode in the same tube, or is more negative than the cathode.

Now we must learn about practical methods of obtaining grid bias voltages in television and radio receivers. Many facts relating to grid voltage and grid bias are illustrated by Fig. 15-1. In diagram "A", we have a 2-volt alternating (a-c) signal voltage applied between grid and cathode. There is no bias voltage, or there is zero bias, which means that grid potential would be the same as cathode potential before the a-c signal is applied. We assume, as always, that cathode potential remains constant and that the entire variation of signal voltage occurs at the grid. Cathode potential is considered to be zero potential. Grid voltage will vary as shown at the right, becoming alternately positive and negative with reference to the zero cathode potential. Since we have a 2-volt alternating signal, the grid becomes alternately 2 volts positive and then 2 volts negative.

A direct-current voltmeter connected between grid and cathode sides of the tube circuit will read zero voltage, which is the bias voltage. So long as the frequency of an alternating voltage acting on a d-c voltmeter is 30 cycles or more, the meter pointer cannot vibrate rapidly enough to follow the changes of voltage. Even though the meter pointer could vibrate at such rates, your eye could not see the movements. Therefore, a d-c voltmeter connected between grid and cathode will read bias voltage in spite of the fact that an alternating signal voltage may be present at the same time.

Now let's proceed to diagram "B" of Fig. 15-1. A 3-volt bias source has been inserted between cathode and grid. The positive terminal of this source is toward the grid, thus providing a 3-volt positive bias. Alternating signal voltage always adds to and subtracts from the bias voltage, so that actual voltage of the grid at every instant is the sum or the difference of signal and bias voltages. The result is shown at the right. The sum of signal peaks and bias voltage is 5 volts, the difference is 1 volt. Grid voltage swings alternately from 5 volts positive to 1 volt positive, and back again. The d-c voltmeter reads 3 volts positive, which is the bias voltage.

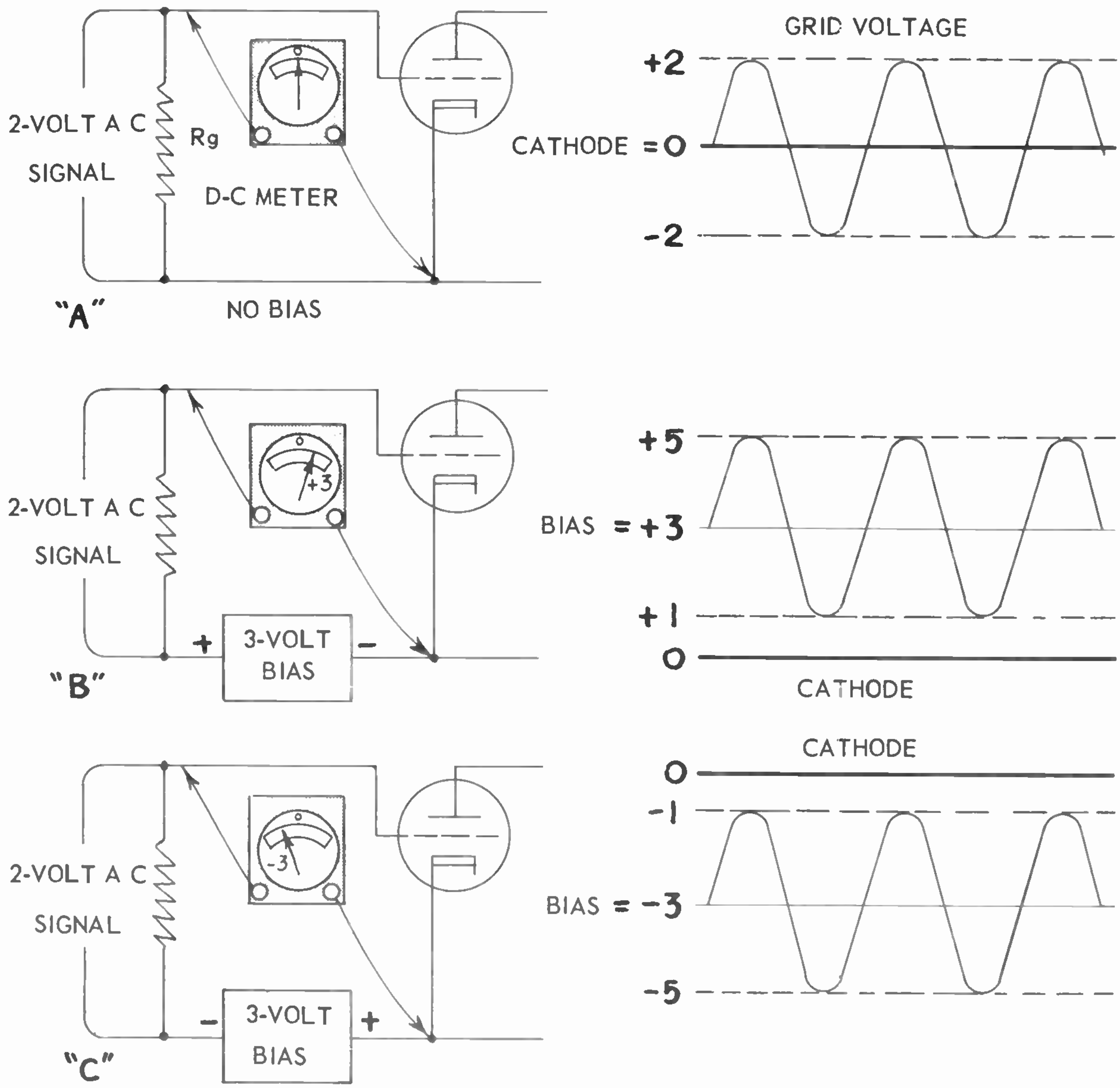


Fig. 15-1. How grid voltage is affected by biasing when the input signal remains unchanged.

In diagram "C" of Fig. 15-1 the bias source has been turned around so that its negative terminal is toward the grid. Now we have a 3-volt negative bias. Again the signal voltage adds to and subtracts from the bias voltage, with the result shown at the right. Adding 2 volts positive signal to 3 volts negative bias leaves the grid 1 volt negative. Adding 2 volts negative signal to 3 volts negative bias makes the grid 5 volts negative.

A

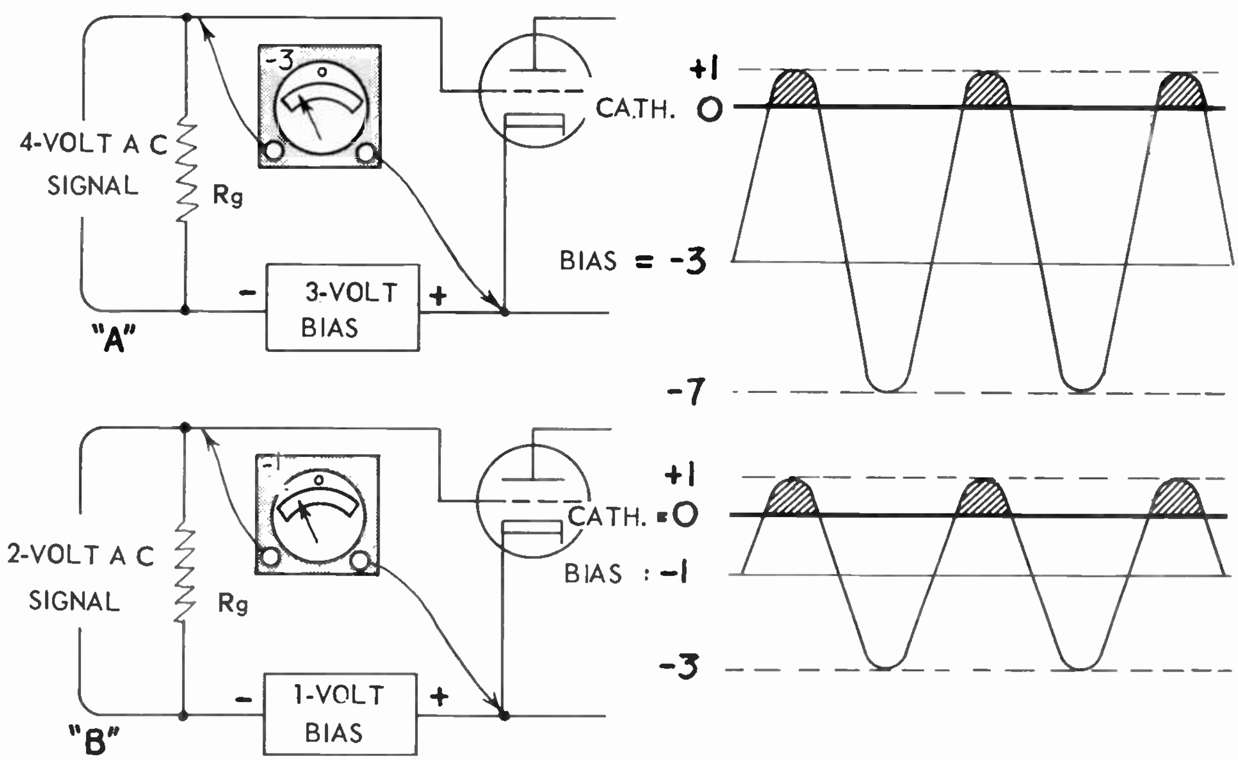


Fig. 15-2. Grid current flows when there are incorrect relations between bias and peak voltage of the input signal.

POSITIVE GRID POTENTIAL. In practically all amplifiers we wish to have what is called linear amplification this means that changes of signal voltage in the plate circuit of the amplifier tube remain exactly proportional to changes of signal voltage at the grid. Anything else is non-linear amplification, more commonly called distortion of the signals. There are many possible causes for non-linear amplification. One fairly common cause is operation of the tube in such a way that the grid becomes positive with reference to the cathode.

Incorrect relations between peak voltage of the signal and bias voltage may allow the grid to become positive as shown by Fig. 15-2. In diagram "A" we have the 3-volt negative bias used in diagrams "B" and "C" of Fig. 15-1, but now have a signal with 4-volt peaks instead of the former 2-volt peaks. As is clearly shown over at the right, every time the signal voltage goes through one of its 4-volt peaks it more than counteracts the 3-volt bias. At each peak of signal voltage the grid becomes momentarily 1 volt positive. This is a case of the signal being too strong for the existing bias.

In diagram "B" of Fig. 15-2 we have gone back to the original 2-volt signal, but the bias has dropped to 1 volt negative instead of the former 3 volts negative. Here again the grid is driven positive at every positive peak of the signal. The positive peaks more than counteract the bias voltage.

One of the reasons why non-linear amplification occurs when the grid goes positive is because there are flows of grid current. The positive grid acts like a positive plate. Electrons from the cathode are drawn into the grid. These electrons flow from the grid through grid-circuit resistor *Rg* and then back to the cathode. It takes only a small positive potential on the grid to attract a lot of electrons, for the grid is very close to the cathode and actually is right in the space charge which exists near the cathode.

This electron flow into the grid is grid current. The grid current is subtracted from the current that should flow only in the plate circuit. What happens is best shown by using a mutual characteristic curve, as in Fig. 15-3, to indicate relations between signals on the grid and in the plate circuit. Plate current should increase steadily and uniformly as grid voltage becomes less negative. Such an increase of plate current would be represented by the full-line curve. But when grid voltage goes over onto the positive side of the zero line, the current subtracted from the plate and diverted into the grid makes the curve bend over as shown by the broken line portion at the top.

Changes of plate current and of signal in the plate circuit have a linear relation to the grid signal so long as the grid is negative. But when the grid becomes positive, at positive peaks of the grid signal, the positive peaks of plate current are chopped off. The changes of plate current are not alike at the positive and negative peaks, and there is a distorted signal in the plate circuit.

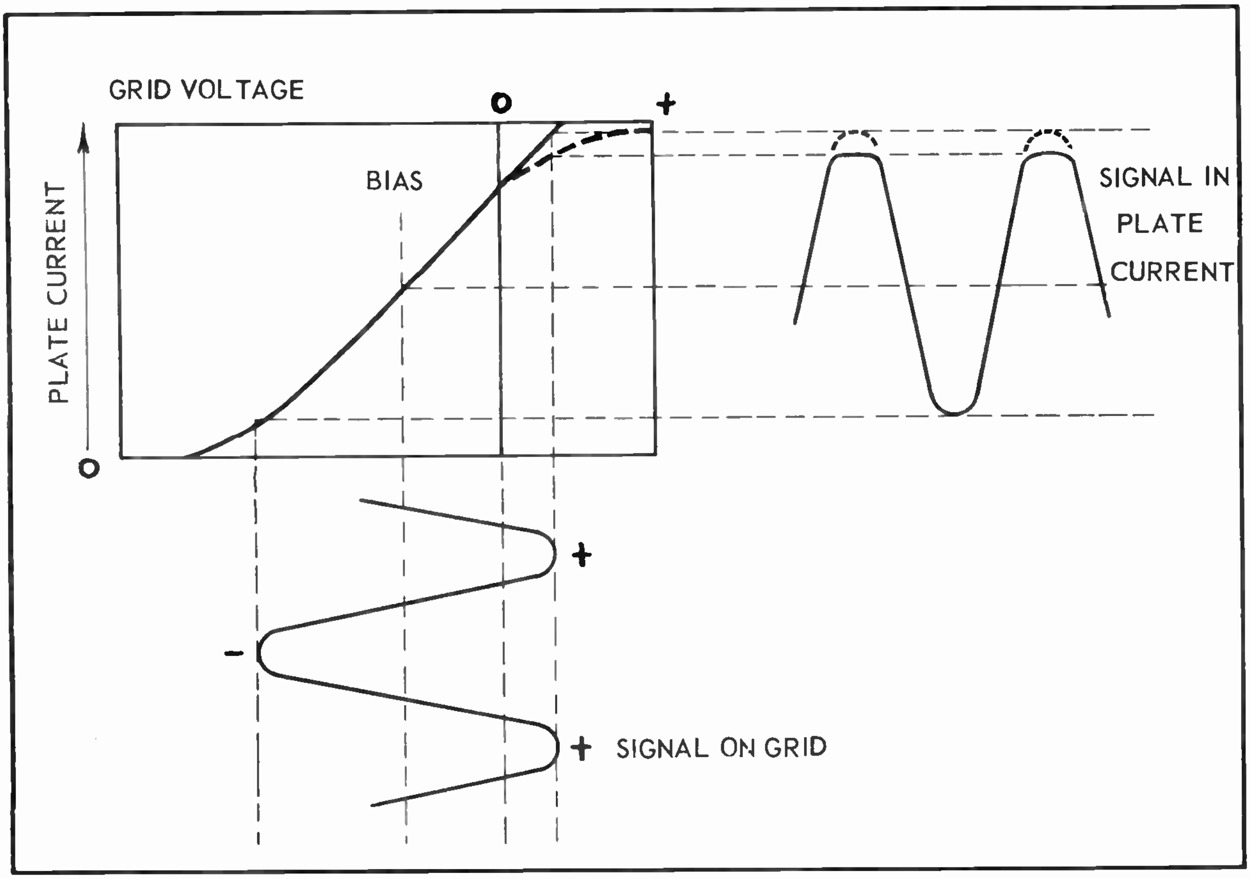


Fig. 15-3 Grid current causes cut-off at the tops of plate current alternations.

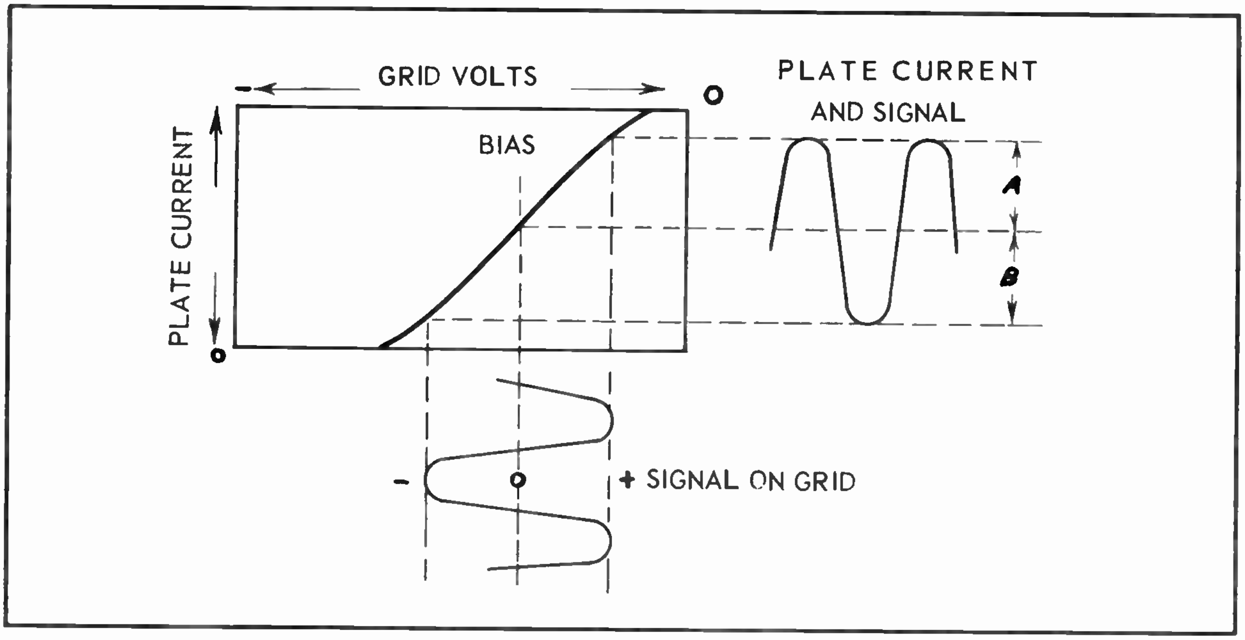


Fig. 15-4 There is linear amplification when signal voltage acts only on the straight portion of the mutual characteristic curve.

The flattening of plate current peaks in Fig. 15-3 is somewhat exaggerated so far as its being caused by current diversion is concerned, but it is not an exaggeration of what really happens to plate current. The contributing cause is this: Grid current which flows in grid circuit resistance *Rg* of Fig. 15-2 requires energy or power to cause the flow. This energy can come only from the source of grid signal voltage. Grid bias voltage comes from the bias source, and plate signal voltage and power come from the power supply, but signal voltage has to come from some other part or some other circuit which precedes the circuits of the amplifier tube. In nearly every case this source of signal voltage is something which cannot furnish any appreciable power without allowing the signal voltage to cease rising. Consequently, when the signal source is called on for power to cause grid current, this source simply lies down on the job and stops increasing its voltage at the instant in which grid current commences. The end result, so far as plate current and plate signal are concerned, is as shown by Fig. 15–3.

EXCESSIVE NEGATIVE BIAS. To have linear amplification or distortionless amplification the prime requirement is to keep the grid signal voltage on the straight portion of the mutual characteristic. Such performance is represented by Fig. 15-4. Where the characteristic curve is straight, or nearly so, changes of plate current will be proportional to changes of grid voltage, or nearly so. The bias must be sufficiently negative that the grid cannot be driven positive by peaks of the applied signal. The signal must not be so strong as to overshoot the straight portion of the characteristic curve. If a longer straight portion is required it may possibly be obtained by using higher plate and screen voltages, with suitable bias, or maybe a different type of tube will be needed.

In Fig. 15-4 the changes of grid signal voltage from zero to positive peaks are equal to the changes from zero to negative peaks. Resulting increases of plate current which have amplitude *a* are equal to the decreases which have amplitude *b*. Now look at Fig. 15-5. Here the bias has been made more negative. It is too far negative for the signal voltage at the grid to remain on the straight portion of the characteristic

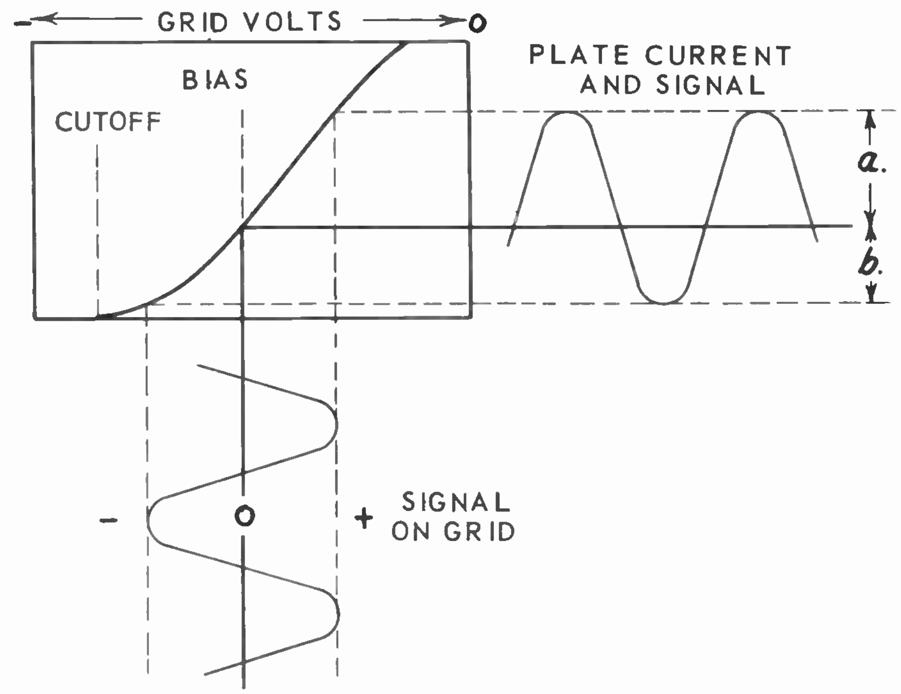


Fig. 15-5 Excessively negative bias causes distortion.

curve. Changes of grid signal voltage from zero to positive peaks cause proportional rises of plate current in the upward alternations *a*. But changes of plate current in downward alternations *b* are not proportional to negative changes of grid signal voltage, nor are they like the upward alternations of plate current.

Were the bias made even more negative than in Fig. 15-5 the negative alternations of grid signal voltage would make the grid more negative than the value for plate current cutoff. Then the downward peaks of plate current would be flattened out to a still greater extent. There are a number of television circuits in which the grid bias is intentionally made so strongly negative that there is plate current cutoff and flattening of the bottom loops of plate current. There also are television circuits in which a considerable grid current is intentionally caused, producing the effect shown by Fig. 15-3. For any one type of tube, with certain plate and screen voltages and a certain grid signal voltage, the mode of operation and the shape of the output signal voltage is a matter of choosing the required grid bias.

SOURCES OF BIAS VOLTAGE. All early receivers obtained voltages for plates, screens, and filament-cathode heating from batteries. Bias voltage for tubes in those sets was obtained from dry cell batteries called "C-batteries" connected as are the bias sources shown in Figs. 15-1C and 15-2. In modern battery-operated portable receivers the grid biases are obtained from voltage drops in the filament-cathode circuits and sometimes from voltage drops in plate-power supply circuits, to the negative ends of which are connected the grid circuits.

The only other type of bias source which at all resembles the C-battery is the bias cell. The cell consists of a black disc about a half-inch in diameter enclosed within an acorn shaped metal cup about 0.3 inch from front to back and 5/8 inch in diameter. The surface of the black disc is the positive terminal. The metal cup is the negative terminal. The bias cell will furnish no current, but has an emf or terminal potentiol difference of 1.25 volts, plus or minus 10 per cent. These cells can be used for biasing in grid circuits which carry alternating currents, but not where there will be direct current through the cell. For bias voltages in multiples of 1.25 volts, it is possible to connect two, three, or four cells in series. Life isindefinitely long if the cells are not abused or overloaded with direct current.

There are three principal methods of biasing in which the voltages are obtained entirely from the receiver circuits and power supply rather than from batteries or cells. Onemethodis called cathode bias or self-bias. Bias voltage is obtained from the potential drop across a resistor connected in series with the cathode of the tube. A second method is called fixed bias. Bias voltage is obtained from the drop across resistors which are part of the power supply system for plate and screen voltages. A third method is called grid-leak bias or grid rectification bias. The bias voltage is obtained from potential drop across a resistor in the grid circuit. Cathode bias and fixed bias are used in both television and radio receivers. Grid-leak bias is unimportant in radio receivers, but serves many important purposes in television sets. A tube may be biased in any desired manner whether the tube is a triode, a pentode, or a beam power type. Biasing principles illustrated for any one of these types will apply also to the other types.

CATHODE BIAS. The elementary principle of cathode bias or self-bias is illustrated by Fig. 15-6, where directions of electron flow in plate and screen circuits are shown by arrows. Electron flows for both plate and screen originate at the cathode inside the tube. The total flow then goes through the external circuit parts, which include the power supply, and returns to the cathode through resistor *Rk*. Direction of electron flow in *Rk* is such that we know the top end or cathode end of this resistor is positive with reference to the lower end, or we may say that the lower end is negative with reference to the cathode end.

The grid of the tube connects through grid resistor *Rg* to the lower end of resistor *Rk*. Thus the grid is connected to a point which is more negative than the cathode, and the grid is thus negatively biased. Because the grid is negative there is no electron flow from cathode into the grid and no flow through resistor *Rg*. With no electron flow in *Rg*. there can be no difference of potential across this resistor. Consequently the grid potential is the same as the potential at the bottom of resistor *Rk*, and the grid bias voltage is equal to the potential difference across *Rk*. This resistor *Rk* may be called the cathode resistor or the cathode-bias resistor.

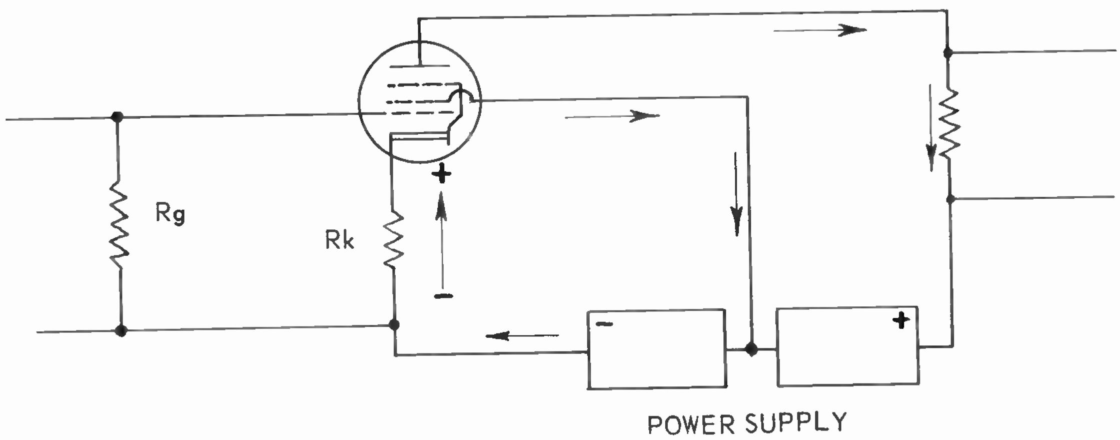


Fig. 15-6 Electron flows in circuits of a tube which has cathode bias.

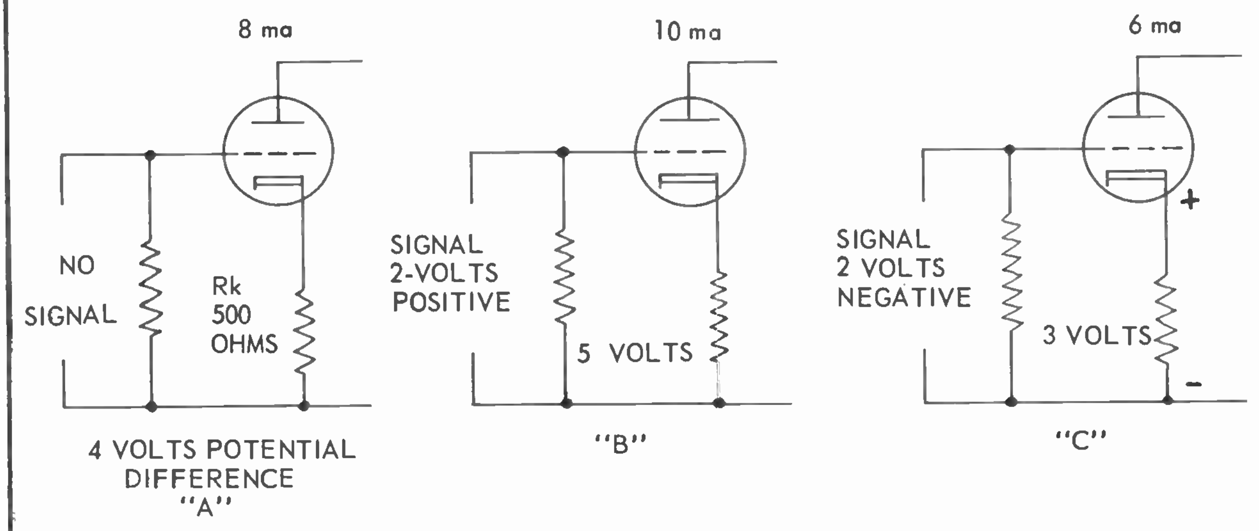


Fig. 15-7 Signal voltage changes plate current in a manner which causes cathode bias to oppose these changes.

To determine the number of ohms resistance in the bias resistor for any required number of volts of grid bias we use our regular formula for resistance in ohms. The number of volts to be used in the formula is the bias voltage. The number of milli-amperes is the total cathode current. This total current is the same as plate current in a triode, and is the sum of plate and screen currents in pentodes and beam power tubes. Then we have,

$$\frac{\text{Ohms in bias}}{\text{resistor}} = \frac{1000 \times \text{grid bias volts}}{\text{milliamperes, total cathode current}}$$

As an example, assume you are working with a tube having 2.9 ma plate current, 0.9 ma screen current, and you wish to have a 3-volt bias. Here is the solution.

$$\text{Ohms} = \frac{1000 \times 3}{2.9 + 0.9} = \frac{3000}{3.8} = 790 \text{ ohms (approximate resistance value)}$$

There are no standard resistors rated at 790 ohms. In preferred values of 5% tolerance there would be resistors of 750 or 820 ohms, which would be about 5% too little or 4% too large were the values exactly as marked, or 9% too little or too large were the resistors at their tolerance limits in one direction, or almost "on the head" with tolerance limits in the other direction. There are standard resistors rated at 800 ohms with 5% tolerance, which are not of the preferred value type. The chances are that performance would be satisfactory with any of these resistors.

Cathode bias has a rather peculiar effect on amplification. What happens, and why, is illustrated by Fig. 15-7. In diagram "A" there is no applied signal. Plate current is 8 ma. This current flows in resistor *Rk*, whose resistance is 500 ohms. Your regular formula for volts shows that potential difference across *Rk* must be 4 volts. Therefore, the grid bias and also the grid voltage are 4 volts negative.

In diagram "B" the signal is 2 volts positive, with the result that plate current becomes 10 ma. With 10 ma in the 500 ohms resistance of *Rk* the potential difference across this resistor must be 5 volts. This potential difference is grid bias, so the bias has been made 5 volts negative. The change of signal voltage is such as causes an increase of plate current. But this very increase of plate current makes the grid bias more negative, and the effect of an increasingly negative grid bias is to decrease the plate current. The bias changes in a way that opposes the effect of the signal. Amplification is less than it would be with some type of bias that does not vary with variations of signal voltage.

In diagram "C" the signal has become 2 volts negative, with the result that plate current drops to 6 ma. With 6 ma in the 500 ohms resistance of *Rk* the potential drop across this resistor must be only 3 volts. Since this drop forms the grid bias, grid bias must now be 3 volts negative. The bias has been made less negative than at diagram "B". Making a bias less negative tends to increase the plate current. But the thing which makes the bias less negative is the change of signal voltage in a direction which decreases plate current. Again the bias changes in a way which opposes the effects of the signal.

Cathode bias always changes in such manner as to oppose the effects of signal variations, and changes in such manner as to oppose all changes of plate current no matter what the reason for current change. This is an advantage when making tube replacements. Tubes of the same type do not have absolutely uniform characteristics. If a new tube permits greater plate current with the same voltages as applied to the original tube, the change of bias from a cathode resistor tends to lessen the plate current. If the new tube allows less current, the bias automatically changes to increase this current.

BYPASSING FOR CATHODE BIAS. In order to lessen the effect of signal voltage variations on grid bias, the bias resistor which is in series with the cathode may be "bypassed" with a capacitor as shown in Fig. 15-8. Incidentally, this diagram shows also that all connections to the negative side of the power supply are most often completed through ground or through the chassis metal.

The variations of cathode current which are the result of variations in signal voltage are really an alternating current. This alternating current in the cathode bias resistor may be thought of as combined with a

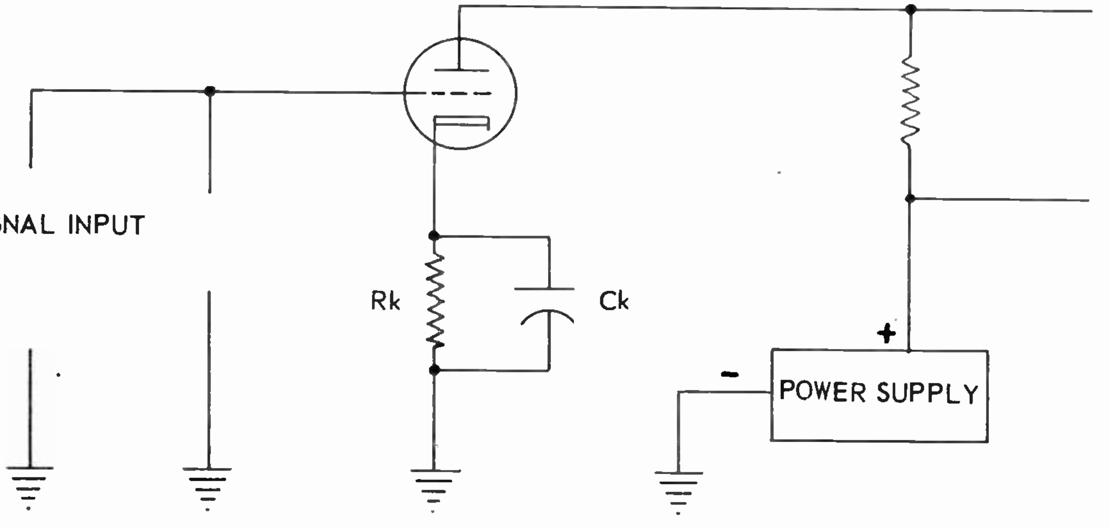


Fig. 15-8 Bypass capacitor on the cathode-bias resistor.

direct current. The cathode current flows always in the same direction, electrons always flow toward the cathode, and so this current is a direct current. But there are variations in the direct current, and the variations would be an alternating current corresponding to the alternating signal voltage were the current variations to exist by themselves.

Since variations of current in the cathode resistor are the equivalent of an alternating current, and since alternating current may flow in a capacitor, the signal variations of current may flow in bypass capacitor *Ck* of Fig. 15-8 instead of in resistor *Rk*.

Whether the alternations of current act in the resistor or in the capacitor depends entirely on the relative oppositions of these two units to flow of current in them. Opposition of the bypass capacitor is measured in ohms of capacitive reactance, while opposition of the bias resistor is measured in ohms of resistance. Obviously, if the opposition of the resistor is something like ten times as great as opposition of the capacitor, most of the current alternations will take the easier path through the capacitor. In fact, ten times as much of the alternating current would flow in the capacitor as in the resistor. The relatively small alternations or fluctuations of current remaining in the bias resistor would cause proportionately small variations of grid bias. Then the grid bias would remain quite constant, and would oppose in a less degree the changes of plate current and changes or effects of signal voltage.

Capacitive reactance varies with frequency, and the capacitance needed to provide any particular reactance in ohms must also vary with frequency. Supposing we have 500 ohms resistance in a biasing resistor and wish to have one-tenth as many ohms, or 50 ohms reactance in the bypass capacitor. At a frequency of 60 cycles we shall need about 53.3 microfarads of capacitance. At a frequency of 60 megacycles the same bypassing effect is obtained with capacitance of about 53.3 micro-microfarads. In later lessons we shall have more to do with the matter of correct bypassing.

Two or more tubes may be biased with a single cathode resistor as shown in Fig. 15-9. Current in the biasing resistor then is the sum of all the plate and screen currents of the tubes whose cathodes are connected

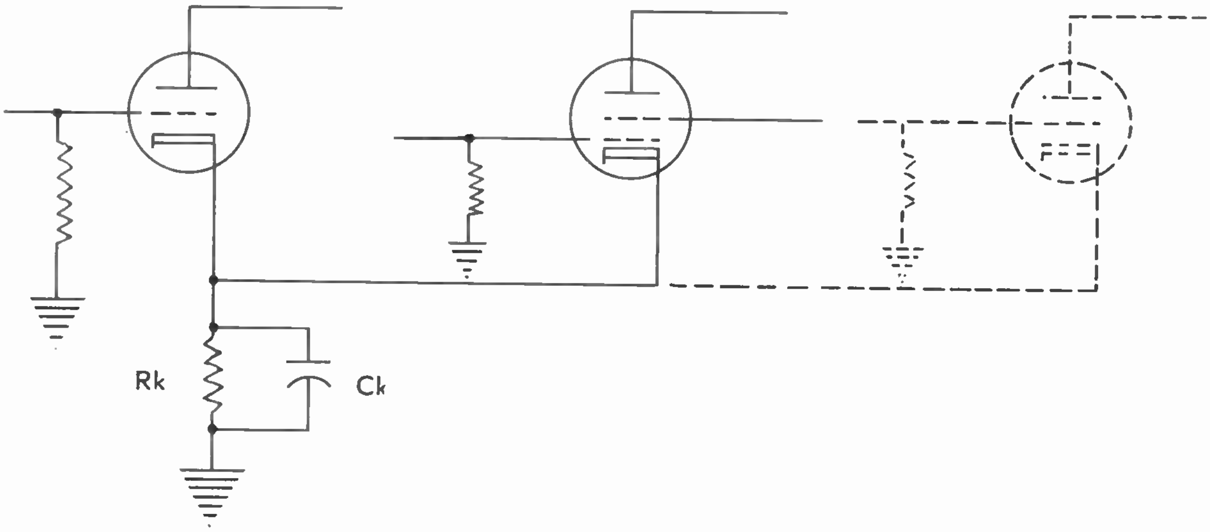


Fig. 15-9 Several tubes may be biased with a single cathode bias resistor.

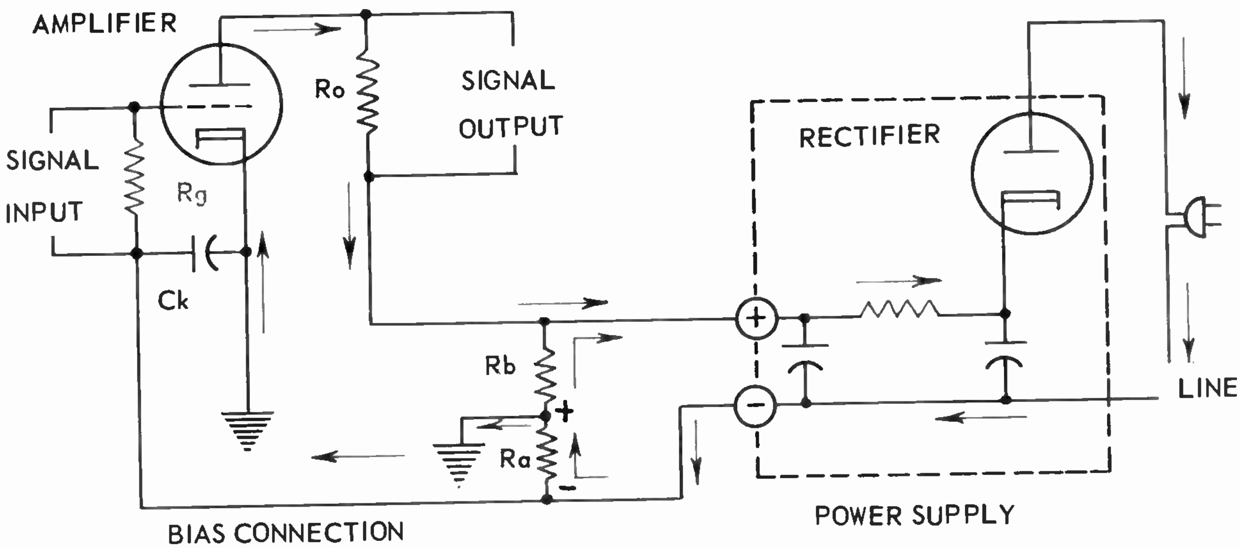


Fig. 15-10 Connections for fixed bias.

to this resistor. The grids in all the connected tubes will be biased with the same number of volts. Changes of plate current or of screen current in any one of the connected tubes will alter the bias, not only in this tube but also in all other tubes connected to the single bias resistor.

FIXED BIAS. The elementary principle of fixed bias is illustrated by Fig. 15-10. Although the diagram is simple, it will bear careful study, for this is the first time we have looked at a complete system of plate, grid, and cathode circuits with their connection through a power supply to the a-c light and power line. For the time being, we need pay little attention to parts and connections of the power supply which are within the broken line enclosure. It is enough to note the rectifier which produces pulses of direct current from the alternating voltage of the a-c line, and the two capacitors and a resistor which act as a "filter" for smoothing the pulses into a fairly steady direct current or electron flow. Directions of electron flow in all the conductors are shown by arrows.

Electron flow from the lower side of the line plug goes through the power supply to the negative terminal, thence upward through resistor *Ra*. Part of this electron flow from the top of *Ra* goes through the ground connections and metal of the chassis to the cathode of the amplifier tube. This portion of the total current goes through the amplifier, out by way of the plate, then through the plate load resistance *Ro* and to the positive terminal of the power supply. Were the amplifier to have a screen element, we would similarly trace the screen electron flow or current from the tube cathode to the screen and to the positive terminal of the power supply. A small portion of the total electron flow through resistor *Ra* goes on up through resistor *Rb* and from there to the positive terminal of the power supply. All electron flow coming back to the positive terminal of the power supply goes through the internal filter resistor and from cathode to plate in the rectifier tube on its way to the upper side of the line plug.

Grid bias in this circuit is the potential drop across resistor *Ra*. Electron flow is upward in *Ra*, so we know that the bottom of this resistor is negative with reference to the top. The amplifier grid is connected through resistor *Rg* and the bias connection to the negative end of bias resistor *Ra*, while the amplifier ca-

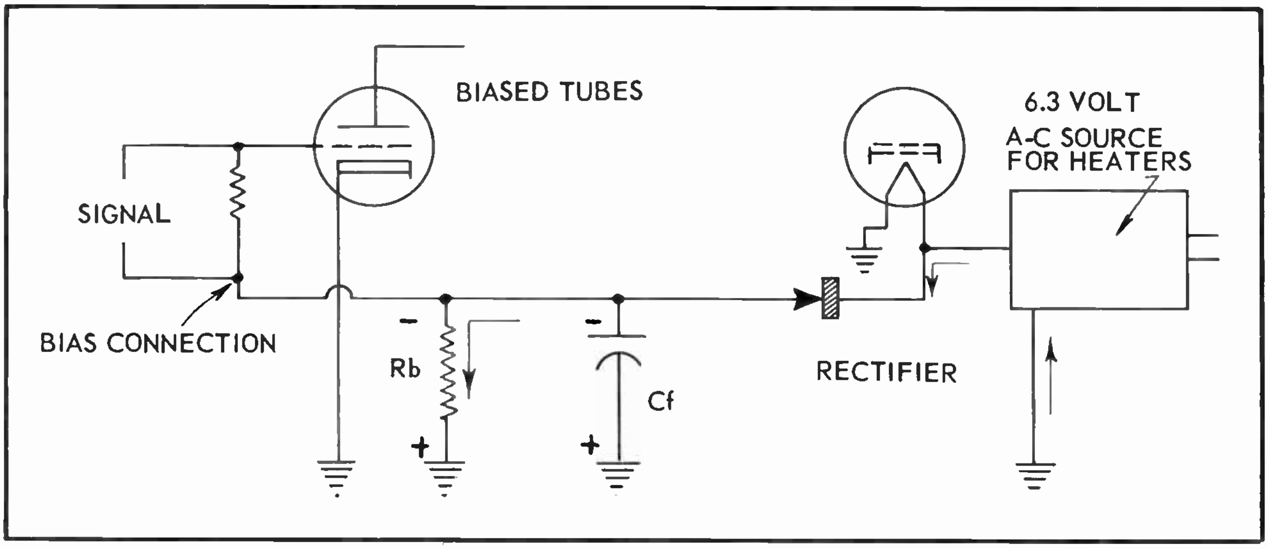


Fig. 15-11 Fixed bias with a rectifier used for this purpose alone.

thode is connected through ground to the top of *Ra*. Consequently, the amplifier grid is negative with reference to its cathode, and the grid is negatively biased.

With the very simple system of Fig. 15-10 the biasing current through resistor *Ra* is only the amplifier plate current plus the small additional current which goes on through resistor *Rb*. Here the grid bias for the amplifier would be affected by every change of plate current except for the bypassing of capacitor *Ck*. Do you see that this capacitor really is connected across resistor *Ra*? The left-hand end of the capacitor is directly connected to the bottom of resistor *Ra* through the bias connection The right-hand end of the capacitor is connected through ground to the top of *Ra*.

Resistor *Rb* is called a bleeder resistor, because it is continually bleeding some current or electron flow from the negative to the positive side of the power supply. Current through *Rb* is called bleeder current. Note that current in bias resistor *Ra* is the sum of amplifier plate current and bleeder current. If bleeder resistance is rather small, so that bleeder current is large in proportion to amplifier plate current, then most of the current in bias resistor *Ra* will be bleeder current. Large changes of plate current then will cause only small changes of total current in *Ra*, for amplifier plate current is only a small part of the total. This means that grid bias will be affected only slightly by changes of plate current in the amplifier, and we have what correctly may be called fixed bias. On the other hand, if bleeder current is small in comparison with amplifier plate current, every variation of plate current will have a large effect on potential drop in resistor *Ra* and on grid bias. Then we do not have real fixed bias.

By dividing the biasing resistance into several sections it becomes possible to obtain a number of different bias voltages. Bias connections from the grids of different tubes may be connected to whateverpoints along the biasing resistance give the desired number of negative bias volts.

In some receivers you will find a rectifier used exclusively for producing direct current and a direct potential difference employed for grid biasing. Such a method is illustrated by Fig. 15-11. The rectifier is not a tube, it is a contact rectifier or a crystal rectifier with which we shall become better acquainted later on. One side of the rectifier is connected to any point in the circuit which furnishes 6.3-volt alternating current for the heaters in the receiver tubes. Any other a-c source might be used, but this one is already installed

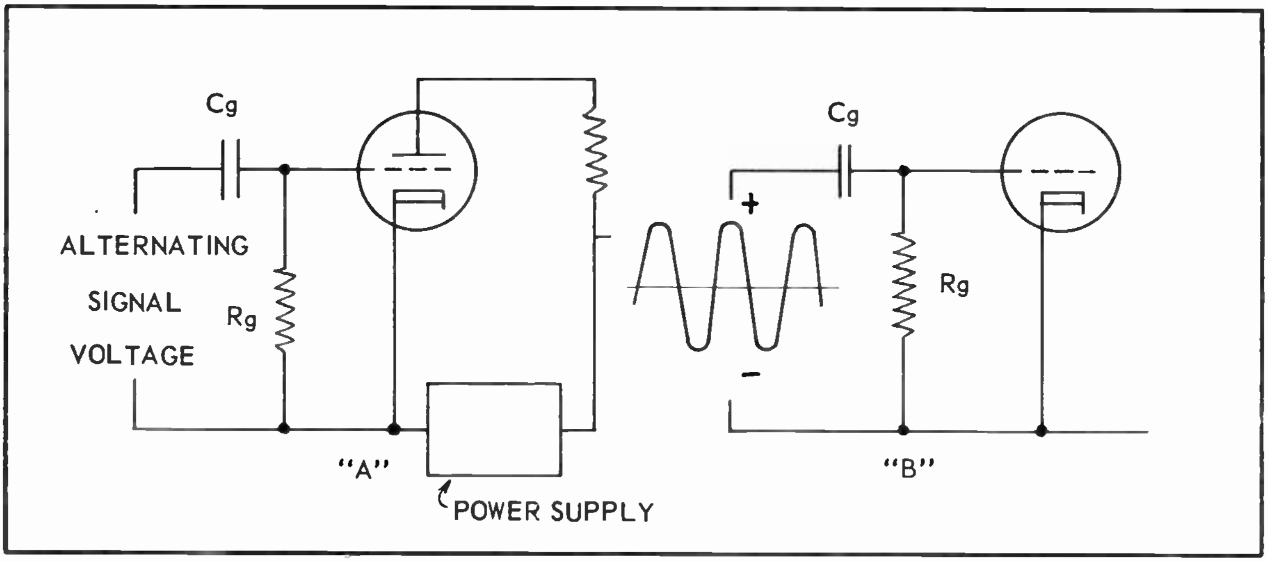


Fig. 15-12 Only the grid and the cathode are active in furnishing grid-leak bias.

and adds nothing to costs. The other side of the rectifier connects through bias resistor *Rb* to ground, and through ground back to the other side of the a-c source. Across resistor *Rb* is connected capacitor *Cf* which acts as a filter capacitor for the pulses of rectified direct current and helps change the pulses to a smooth flow in resistor *Rb*.

Directions of electron flow for biasing are indicated by arrows. The flow is downward in biasing resistor *Rb*, which means the top of *Rb* is negative with reference to the bottom. The grids of biased tubes are connected to the upper negative end of resistor *Rb*, and the cathodes connect through ground to the lower end. Thus the grids are negatively biased with reference to the cathodes.

GRID-LEAK BIAS. The third common method of biasing which we shall discuss is usually called grid-leak bias, although it may be called grid leak-capacitor bias or grid rectification bias. Connections for a tube with grid-leak bias are shown at diagram "A" of Fig. 15-12. In series between the grid and the source of alternating signal voltage is a grid capacitor *Cg*. A grid resistor, *Rg*, called the grid leak, is connected from the grid to the cathode either directly or through ground. Grid-leak bias may be used for triodes, pentodes, or beam power tubes, but no matter what the type of tube, the only elements which take any part in the biasing action are the grid and the cathode, as shown at "B". Grid leak bias is applied to oscillators, to mixers, to certain classes of high-power audio amplifiers, to television sync limiters, and to some sync amplifiers.

The beginning of the biasing action is shown in diagram "A" of Fig. 15-13. At the instant here represented the signal voltage is of positive polarity at the connection to the grid capacitor, and is negative at the connection for the tube cathode. Electrons are withdrawn from the side of the grid capacitor toward the signal source. Electrons then must flow into the opposite side of this capacitor. These other electrons flow as shown by arrows from the negative side of the signal source to *Rg* and the tube cathode, from cathode to grid within the tube, and from grid and *Rg* to the side of the capacitor which is receiving a negative charge. The cathode has been made negative by its direct connection to the signal source. The grid has been made positive because electrons which first flowed into the capacitor through *Rg* were taken from the grid to leave the grid deficient in electrons. The potential difference across *Rg* is positive bias for the grid.

During the following half-cycle of signal voltage we have the conditions shown at "B" in Fig. 15-13. Signal polarity has reversed. Electrons from the signal source are being forced into one side of the capacitor and forced out of the other side. The electrons leaving the side of the capacitor which is toward the grid cannot pass from grid back to cathode through the tube because there can be no electron emission from the cold grid with the small potèntial existing here, and electrons would not flow from the negative grid to the positive cathode. Consequently, these electrons flow downward through grid resistor *Rg*. This direction of flow means that the upper end of *Rg*, connected to the grid, is negative with reference to the lower end, connected to the cathode. The potential difference across *Rg* is negative bias for the grid.

During this first cycle of signal voltage the capacitor is charged during all the time that the grid is held positive, for during all this time there is flow of grid current through the tube and *Rg* into one side of the capacitor. There is discharge through the grid leak during the half-cycle in which the grid is negative and the cathode positive. Let's assume that we make the grid leak resistance so great that the capacitor can discharge only slowly. This means making the resistance so great that the potential of the charged capacitor can cause only a small current in the resistor. At the end of this half-cycle of discharge only part of the charge has leaked off the capacitor. Then begins another positive half-cycle, during which the capacitor again is charged. This is followed by another negative half-cycle during which there is slow discharge. Only a few cycles are required before the charging pulses have to do nothing more than replace the slight losses of charge which occur during alternate half-cycles.

At "A" of Fig. 15-14 are represented alternate charges and discharges of the grid capacitor. At "B" is represented the alternating signal voltage. As the capacitor accumulates more and more charge and more and more potential difference between its plates, the signal voltage has to rise higher and higher before it can overcome the capacitor voltage and add still more charge. Consequently, the charging times and the charging currents become smaller until the currents are merely brief pulses which replace the charge lost through the grid resistor.

The capacitor acts much like a tank for electrons. Electrons leak out of the tank through the grid resistor at a practically steady rate. This constant leakage is replaced by intermittent spurts of electrons forced into

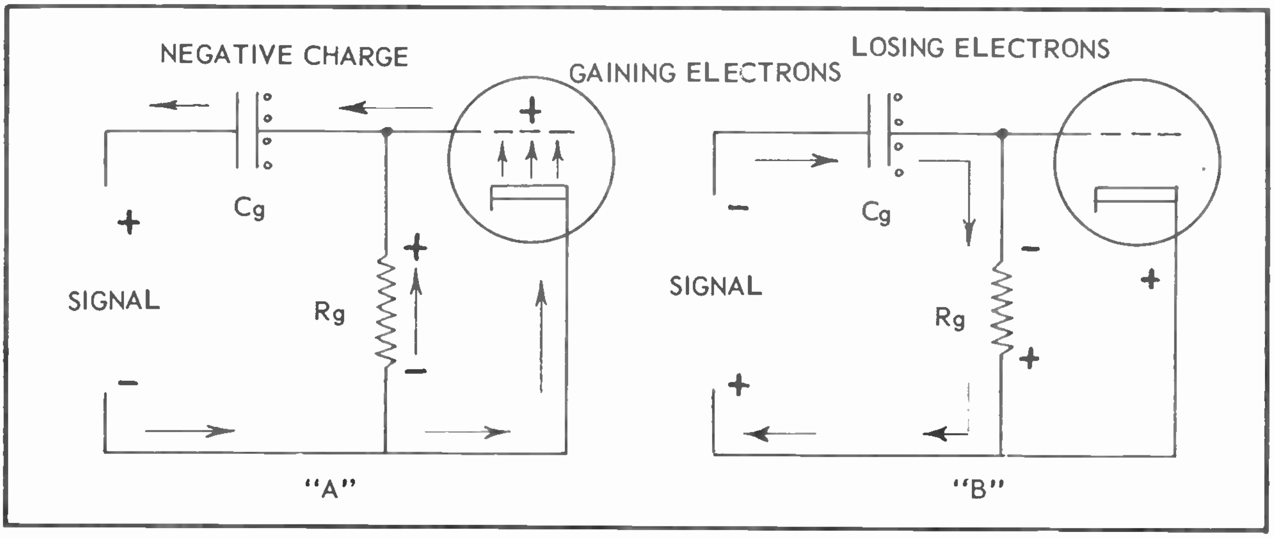


Fig. 15-13 Electron flows during alternate half-cycles of signal voltage when there is grid-leak bias.

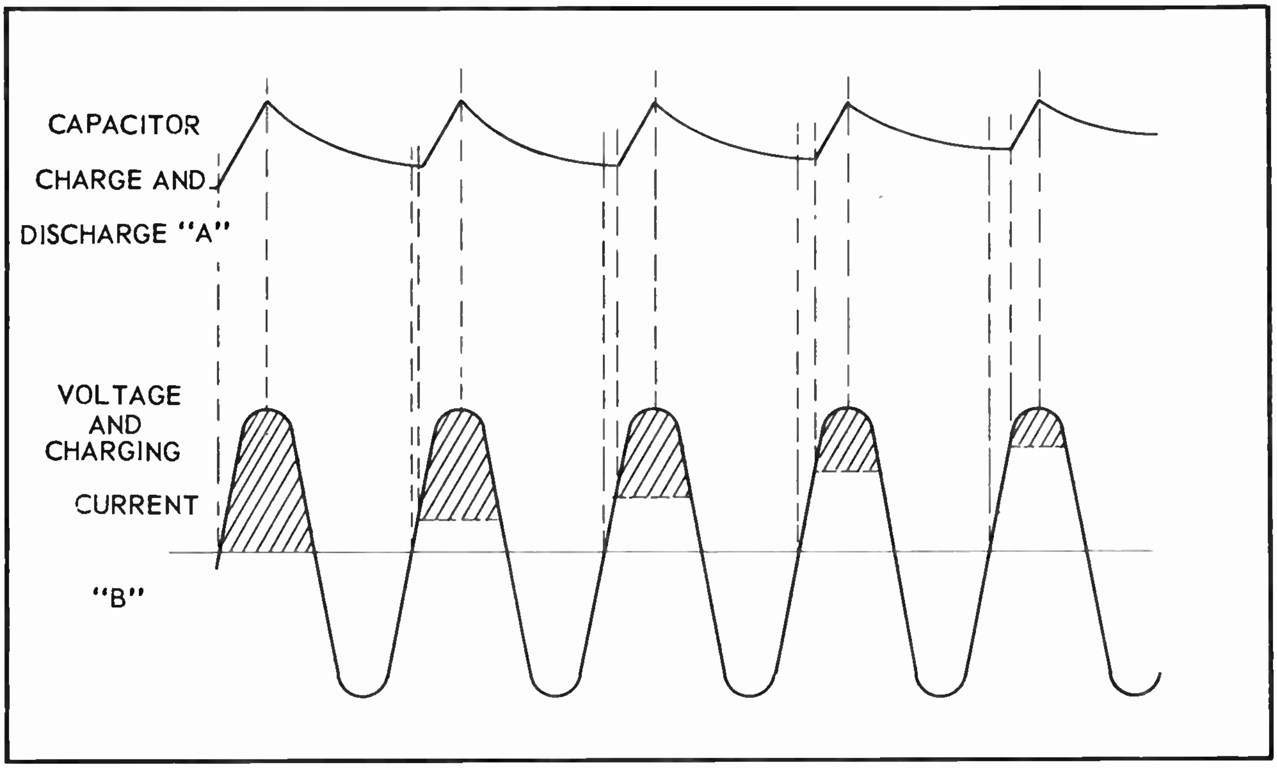


Fig. 15-14 With grid-leak bias only brief pulses of grid current are needed to keep the grid capacitor charged

the capacitor by the signal voltage peaks. The practically steady electron flow through the leak resistor means that there is a practically steady potential drop across this resistor. This potential drop is negative bias voltage for the grid.

The greater the resistance of the grid leak the more charge is retained in the capacitor and the higher becomes the average voltage of the capacitor. Thus the grid bias voltage is increased or the bias is made more negative by using more leak resistance. Increasing the capacitance of the grid capacitor causes some increase of bias voltage, but nowhere near as much as an increase of leak resistance.

Signal pulses such as shown at "B" of Fig. 15-14 act through the grid capacitor to vary the instantaneous voltage of the grid, while the average voltage or the bias is maintained by the action which has been described. The instantaneous variations of grid voltage act on the electron stream from cathode to plate, and cause signal variations of plate current and plate voltage in the usual manner.

Pulses of charging current for the grid capacitor come through *Rg* and the grid of the tube, which means grid current. Grid current causes distortion of the output signal. The very small pulses of grid current do not cause much distortion, but there is enough to limit the use of grid-leak bias to applications in which distortion does no harm or in which the distortion is compensated for by other means.

Grid-leak bias voltage is produced as a result of signal voltage. If no alternating signal voltage is applied to the grid circuit there can be no grid-leak bias; or rather there will be zero bias. Except with low

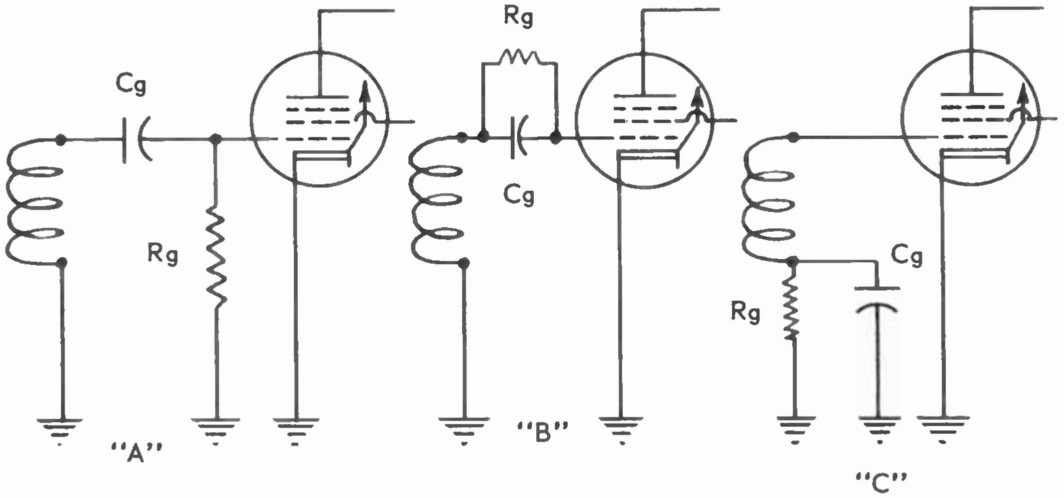


Fig. 15-15 Connections for capacitor and leak resistor with grid-leak bias.

values of plate and screen voltages, or with tubes whose plate resistance is very high, the plate current with no signal and zero bias may become so large as to damage the tube or other parts in the circuits. Cathode bias or else fixed bias often is used in addition to grid-leak bias. These other methods of biasing provide enough minimum negative bias to prevent excessive plate current with no signal. When a signal is applied, the bias is made still more negative by the grid-leak action.

Connections for the capacitor and leak resistor used with grid-leak biasing may take the form of any of those shown in Fig. 15-15. The coil or inductor in each diagram represents any source of signal voltage. At "A" is the same arrangement of parts used in Figs. 15-12 and 15-13. At "B" the leak resistor is connected across the capacitor. Electron flow in the resistor here goes through the signal source, through ground, and thence to the cathode of the tube. At "C" the leak resistor again is connected across the capacitor, but here both these units are on the ground side or cathode side of the signal source rather than on the grid side.

CATHODE RETURN CIRCUITS. In all the grid circuits which have been examined, and in every other grid circuit, it is possible to trace all the way from the grid to the cathode, outside the tube, without having to go through a capacitor. This is the same as saying that every grid circuit consists of conductive connections, no insulators, for its entire length. We speak of the grid connection to the cathode as the grid return. In Fig. 15-16 the grid return is from the grid through resistor *Rg*, then resistor *Ra*, and ground, back to the cathode.

A conductive grid return is necessary in order to maintain a grid bias or a grid potential, which is referred to the cathode. The bias is a steady voltage, not alternating in itself, and any such voltage or the current it might cause would be blocked by the insulating dielectric in any capacitor. Without a conductive grid return we have what is called a "free grid" or a "floating grid". A free grid collects negative free electrons from the space charge, and cannot get rid of them. Then the negative free grid blocks plate current, and causes erratic performance in general.

There must be also a conductive return path all the way from the plate to the cathode through external

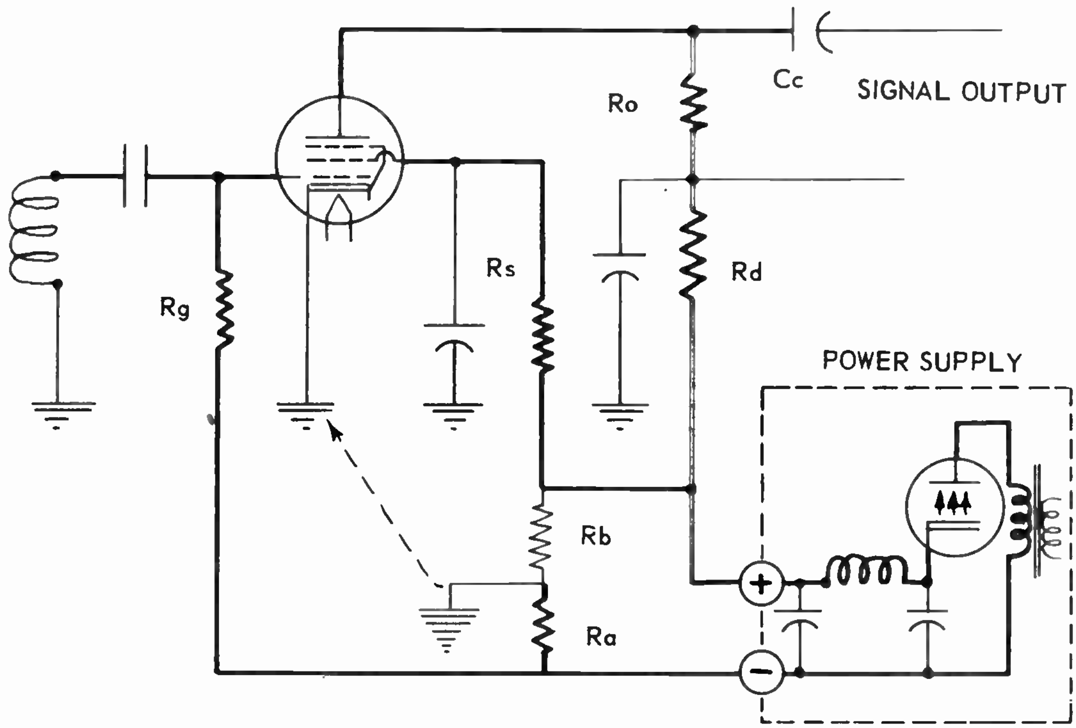


Fig. 15-16 Return circuits to the cathode from grid, plate, and screen.

connections. The plate return circuit of Fig. 15-16 goes from the plate through resistors *Ro* and *Rd* to the positive terminal of the power supply, through the power supply to its negative terminal, thence through resistor *Ra* and ground to the cathode.

There must be a conductive plate return in order that direct electron flow or direct current may flow in the plate circuit. It is this direct current that is varied for the alternating signal current and voltage. In Fig. 15-16 the output signal is taken from the plate through capacitor *Cc*. The one-way or direct portion of the plate current cannot get through capacitors, so must be provided with a conductive plate return circuit.

There is direct electron flow or direct current also between cathode and screen, and this direct current must be provided with a conductive path (no capacitors) from screen back to cathode outside the tube. The screen return of Fig. 15-16 is from the screen through resistor *Rs* to the positive terminal of the power supply, then through the same path as the plate return back to the cathode.

Fig. 15-17 shows cathode return points for battery operated filament-cathode tubes. Returns from the grid (G) always are made to the end of the filament-cathode which is connected to the negative side of the battery used for filament heating. This connection provides negative grid bias equal to half the potential difference across the filament, because the negative side of the filament is more negative than the average potential by half the potential difference. If the grid return is made to the end of the filament that connects to the positive of the heater battery, the bias will become more positive than the average poten-

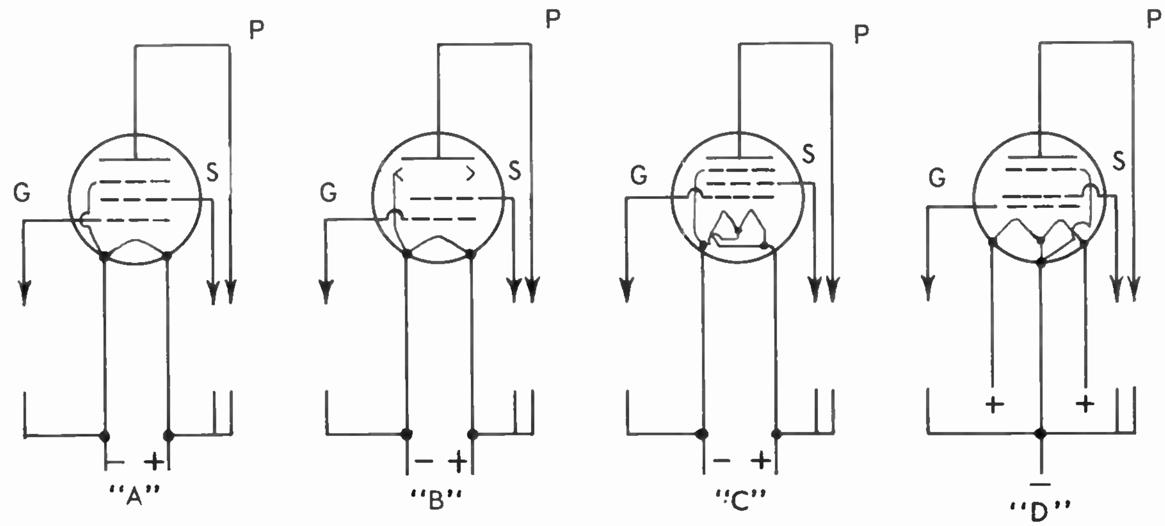


Fig. 15-17 Cathode returns for tubes with battery operated filament-cathodes.

tial of the filament cathode. The difference between negative bias with one connection and positive bias with the other is equal to the full potential difference across the filament. Additional bias may be provided by means of a fixed bias or by means of grid-leak bias.

Returns from screens (S) and from plates (P) may be made to the side of the filament that is connected to the positive of the heater battery. This connection adds to the plate and screen voltages an extra voltage equal to half the potential difference across the filament. If the extra plate and screen voltages are not particularly desired, the plate and screen returns may be made to the negative end of the filament. Such a connection is made in diagram "D".

A-, B-, AND C-VOLTAGES. As has been mentioned, all the early receivers obtained all their power from batteries. The battery used for heating the filament-cathodes was called the A-battery, and voltage from that battery sametimes was called A-voltage. When automatically charged batteries and other line-powered filament heating sources were used they were called A-power supplies. Such use of the letter "A" has all but disappeared.

Batteries furnishing voltage and current for plates and screens were called B-batteries. Supply voltages for plates and screens were called B-voltages. Power supplies for plates and screens were called B-power supplies. Today we still have B-batteries for portable battery-operated radio receivers. We still speak of B-power supplies. In fact, all the power supplies which have been shown and mentioned in all the lessons to and including this one are correctly called B-power supplies. These power supplies furnish B-voltages for plate and screen circuits. From the positive side of the B-power supplies we have B-plus or B+ voltages, and from the negative side have B minus B– voltages or potentials. The letter "B" is just as much alive as ever, and from now on we shall use it regularly.

Batteries furnishing grid bias voltage for early receivers were called C-batteries, and their voltages often were referred to as C-voltages. C-batteries still are used in a few applications, but bias voltage for battery-operated receivers more often is secured from the B-battery circuit in much the same manner as employed for fixed bias systems.

16 → 20

COMMERCIAL TRADES INSTITUTE

KNOWLEDGE + AMBITION = SUCCESS

LEARN . . .

Training Your Mind

C. T. I.

Perfecting Your Skill

ACT . . .

"The heights by great men reached and kept
Were not attained in sudden flight,
But they, while their companions slept,
Were climbing upward in the night."

ENJOY . . .

COMMERCIAL TRADES INSTITUTE

CHICAGO, ILLINOIS

TELEVISION

LESSON NO. 16 MAGNETS IN TELEVISION AND RADIO

"Training Your Mind"

C. T. I.

"Perfecting Your Skill"

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

COMMERCIAL TRADES INSTITUTE

Training Your Mind — C.T.I. — Perfecting Your Skill

LESSON NO. 16

MAGNETS IN TELEVISION AND RADIO

THE NEED FOR TUNED CIRCUITS

We have accumulated much information about the electrical behavior and practical applications of resistors, capacitors, and tubes. But no combination of these elements alone would allow tuning to a radio or television program. Even if we could get signals through the tuner we couldn't make the i-f amplifier select the one carrying a desired program. And if the i-f amplifier would work we still couldn't operate a sweep circuit for television, nor any kind of loud speaker except types popular twenty years ago. Of course, you know the reason for all these difficulties, we would have no inductors or coils.

8 The first thing we need in any receiver is tuned circuits, for without such circuits there is no chance of separating signals of one station or one channel from those of other stations or channels. To build tuned circuits of practical usefulness requires inductors. Next we need great amplification for limited bands of carrier frequencies and intermediate frequencies, but there can be no such amplification without resonant circuits, and resonance requires inductors. We should like also to use picture tubes with magnetic deflection and focusing, but here again we need inductors. Finally, all modern speakers use inductors, and most of them use magnets as well.

It happens that everything related to the basic principles of tuning and resonance, and everything about magnetic deflection and focusing, and most of the principles of loud speakers – all these have their beginnings in the principles of magnets and magnetism. We could devote several lessons to magnetism alone, and all would be interesting, but instead, the essential facts will be presented and limited to those which are related in one way or another to radio and television. The relation of some of the facts to practical applications may not be immediately apparent, but as we proceed it will become plain that each is necessary for full understanding of what is to come.

It is not so important to remember every little detail about magnetism and magnetic circuits as to gain a good overall comprehension of how so many processes important in reproduction of pictures and sound are tied together by the simple beginnings from which all have been developed.

All of us have played with or experimented with small toy magnets at one time or another. These are so-called "permanent" magnets, made of hard steel. A magnet will strongly attract and may pick up nails and other articles made of iron or steel, but will refuse to attract objects made of copper, brass, aluminum, or any metal other than iron or steel, and will refuse to pick up anything which is not a metal. The usual definition of a magnet says it is a body which attracts iron and steel.

For all practical purposes, magnets and all substances having magnetic properties contain iron in greater or less proportions. Steel is merely a form of iron which contains carbon. The metals nickel, cobalt, gadolinium, and some alloys are weakly magnetic, but these materials do not interest us at present.

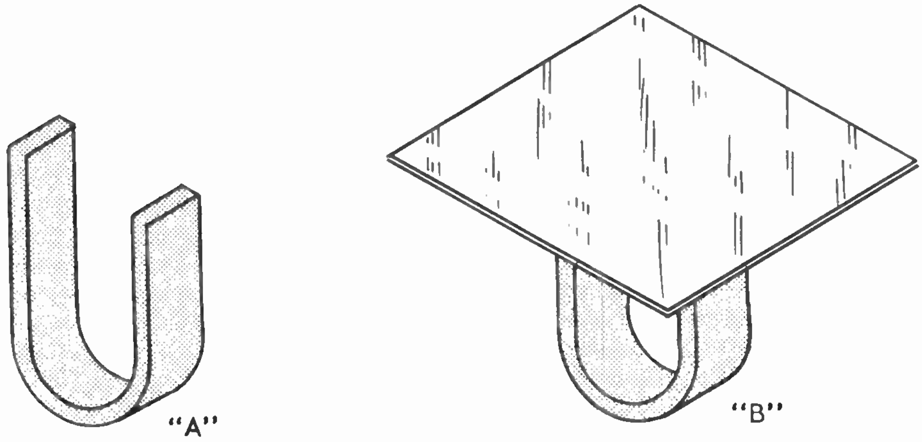


Fig. 16-1. How the magnet and paper are used for "mapping" magnetic fields.

The useful thing about a magnet is the force which extends through space around it. It is this force which exerts attraction, but in radio and television we are not at all interested in the ability of magnets to attract iron and steel. Instead we are interested in the ability of this magnetic force to move electrons, and are interested in changing the energy of moving electrons into magnetic force, and in using the magnetic force to move electrons.

It is quite easy to see the directions in which magnetic force is acting, or to see the lines along which it acts in the space around a magnet. The experiment requires a U-shaped or horseshoe permanent magnet supported with its two "poles" upward as at "A" in Fig. 16-1, and resting on the poles a sheet of stiff paper or light cardboard as at "B". When iron filings are sprinkled from a height of about a foot onto the paper or cardboard, the filings arrange themselves in a pattern of the general form shown by Fig. 16-2. This is an actual photograph.

It is plain to be seen that the magnetic force is acting along very definite lines between the poles of the magnet. These are lines of magnetic force. If a very small magnet could be so freely supported as to float in the space around the poles, that magnet would move from one pole to the other along the line of force on which the small magnet happened to be placed.

The space in which there is magnetic force and magnetic lines of force is called the magnetic field. The field around a magnet extends outward to an indefinite distance, but becomes rapidly weaker as we move away from the magnet. The intensity of the magnetic field and magnetic force is much greater near the magnet poles than far from them.

The magnetic field and its lines of force radiate in all directions from the poles, not only in a single plane as shown by Fig. 16-2. At "A" in Fig. 16-3 the paper has been supported a short distance from the poles and at "B" it has been supported still farther from the poles. The two patterns give an idea as to how the lines extend through space.

If any magnet is supported so that its two poles may move in a horizontal plane, the magnet will swing around until one pole is toward the geographical north and the other toward the geographical south. The north-pointing pole is called the north pole of the magnet, and the south-pointing pole is called the south pole. For convenience and uniformity in making explanations of magnetic actions, it is assumed that lines of force issue from the north pole, pass through curved paths between the poles, enter the south pole, and pass through the interior of the magnet to again come out at the north pole. It must be remembered that nothing actually is moving along these lines; their direction is purely imaginary. The assumed directions are shown for a U-shaped magnet and for a straight or bar magnet at "A". Fig. 16-4.

If the poles of two magnets are brought close together there may be either attraction or repulsion. As shown in "B" of Fig. 16-4, if the two poles are alike, both north or both south, there is repulsion. Unlike poles, north and south, have attraction for each other. Repulsion of like magnetic poles is similar to repulsion of like electric charges, and attraction of unlike poles is similar to attraction between unlike electric charges.

A north pole cannot exist without a south pole in the same magnet, nor can a south pole exist without a north pole. There may be more than one north pole or more than one south pole, but there must be at least one of the opposite kind. With two opposite poles, their field strengths or intensities must be equal. The total strengths of all north poles must equal the total strengths of all south poles.

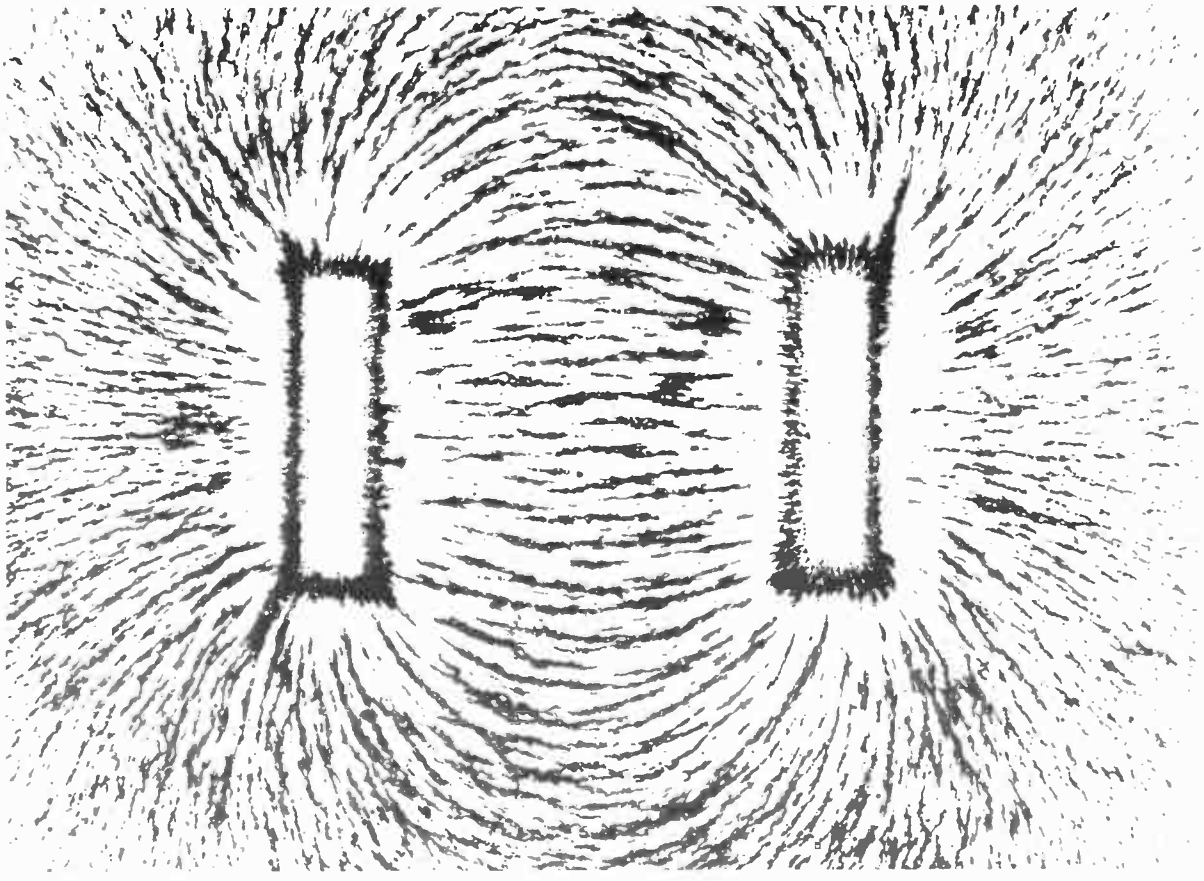


Fig. 16-2. Iron filings show directions of magnetic lines of force between the poles.

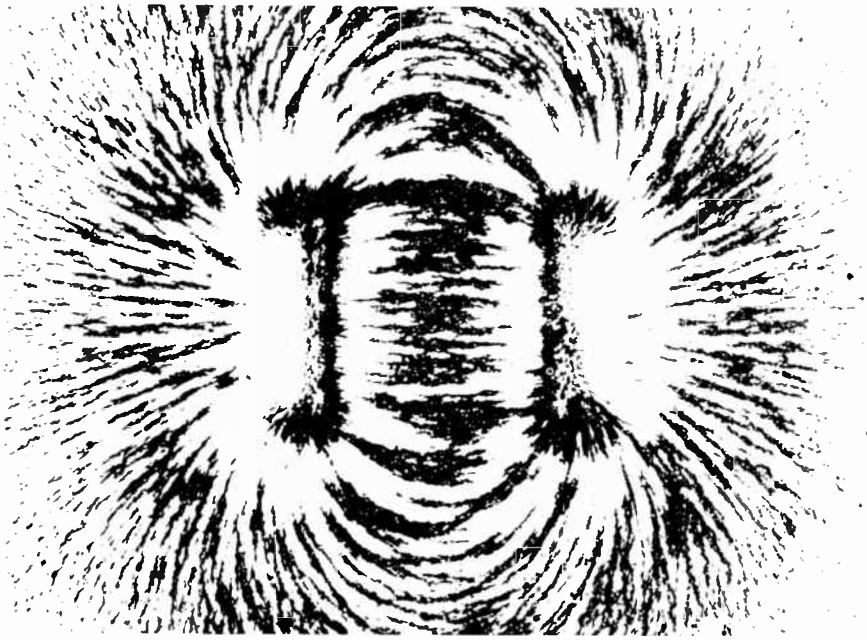


Fig. 16-3 A

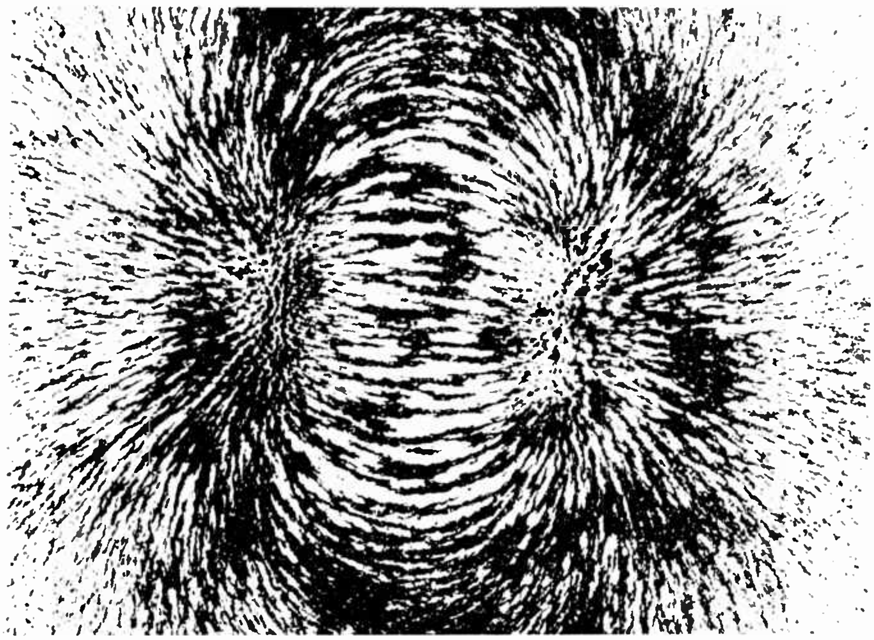


Fig. 16-3B Field lines extend above and around the poles as shown here.

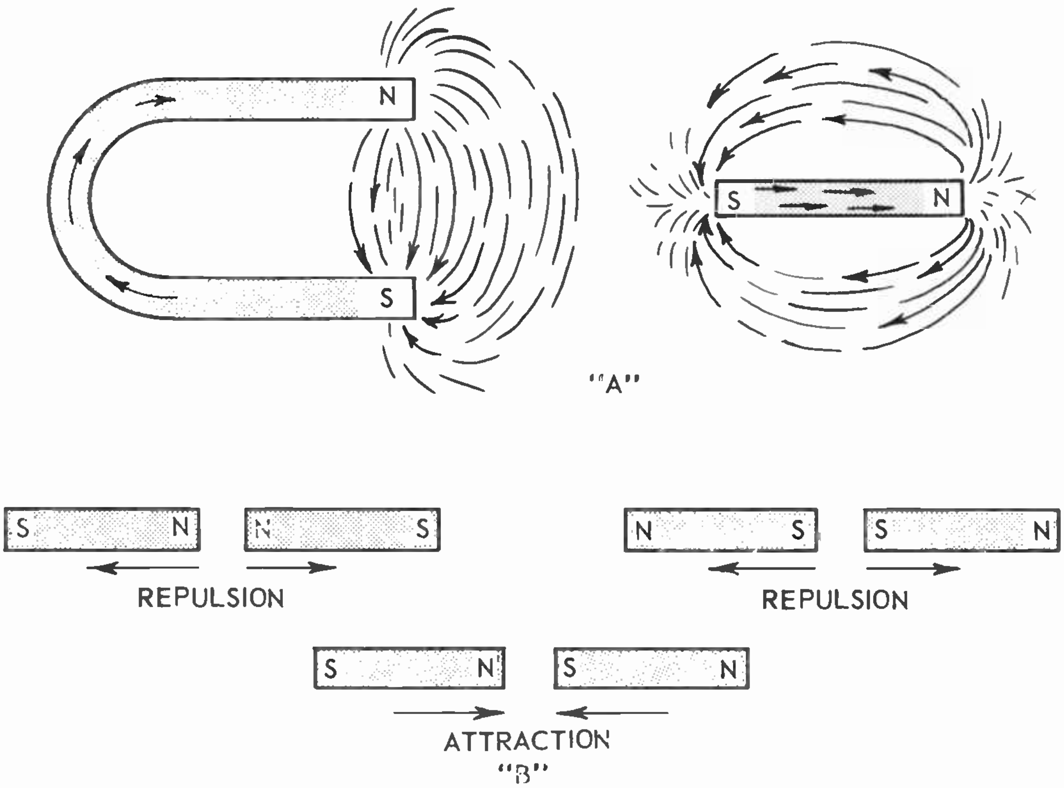


Fig. 16-4. Magnetic poles which are alike repel each other, but opposite poles attract.

When an unmagnetized piece of iron or steel is brought near the north pole of a magnet, the piece of iron or steel is magnetized and its end nearest the north pole becomes a south pole. If the end of the piece of iron or steel is brought near the south pole of a magnet, that end becomes a north pole. A magnetic pole always induces a pole of the opposite kind.

When one magnet is in the field of another magnet, as at "A" in Fig. 16-5, the first magnet tends to turn until the lines of force through it extend in the same direction as the field lines, as shown at "B". This turning effort is called magnetic torque.

The magnetic field and magnetic lines of force pass through all non-magnetic substances just as though those substances did not exist. Forces of attraction and repulsion at any given distance from the pole of a magnet have exactly the same strength whether the intervening space is a vacuum, or is filled with air, brass, paper, glass, aluminum, or anything else which is not iron or steel or which does not contain iron or steel, or is not a magnetic alloy. There is no such thing as an insulator for magnetism.

It is so much easier for the magnetic field, or force, or effect, to pass through iron or steel than through anything else that the field may be directed or confined as in Fig. 16-6. As at "A", any piece of iron or steel near the magnet will draw into that piece nearly all the field lines. Even though the path through pieces of iron and steel is relatively long and broken up, as at "B", most of the field lines still will fol-

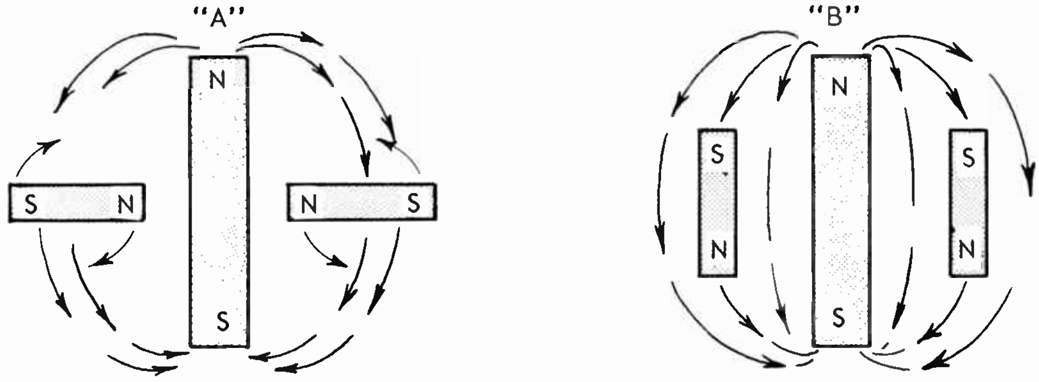


Fig. 16-5. A magnet turns to place its own lines in the same direction as the lines in a surrounding field.

low the longer but easier path. A complete enclosure of iron or steel, as at "C", forms a magnetic shield. The field outside the shield will be zero for all practical purposes.

Permanent magnets are used chiefly in the type of loud speaker called a permanent-magnet or "PM" speaker, in some forms of ion traps for magnetic picture tubes, in some types of magnetic focusing for picture tubes, and in all common types of voltmeters and current meters. Fig. 16-7 is a picture of a circular permanent magnet forming the largest single part in the "movement" of a current measuring meter.

Permanent magnets are used where the need is for a magnetic field of constant strength. After being correctly prepared or "aged" after manufacture, these magnets will maintain a uniform strength for an indefinite time unless subjected to abuse. Most permanent magnets are made of carbon steel, or of alloys containing chromium, tungsten, or cobalt, or alloys containing aluminum, nickel, and cobalt which go by the trade name Alnico. Alnico magnets are much used in radio and television because of their small size for any given field strength. For the same ability to maintain an external field, Alnico magnets of types, I, II, III, and IV need be only about one-sixth the weight of chrome or tungsten magnets, and a magnet of Alnico V may weigh only about one-eighteenth as much.

ELECTROMAGNETS. In Fig. 16-8 is a coil of insulated wire from which two connections go to a source of current. Through the center of the coil are rods of iron, and on opposite ends of these rods are blocks of

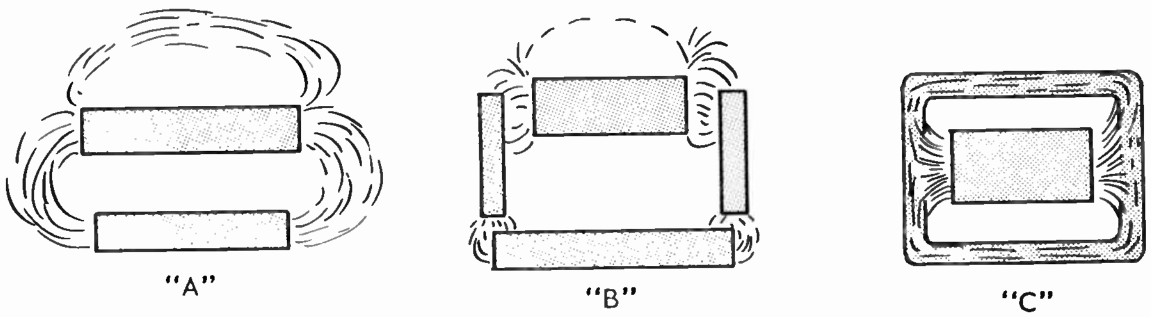


Fig. 16-6. Magnetic fields may be diverted or confined by iron or steel.

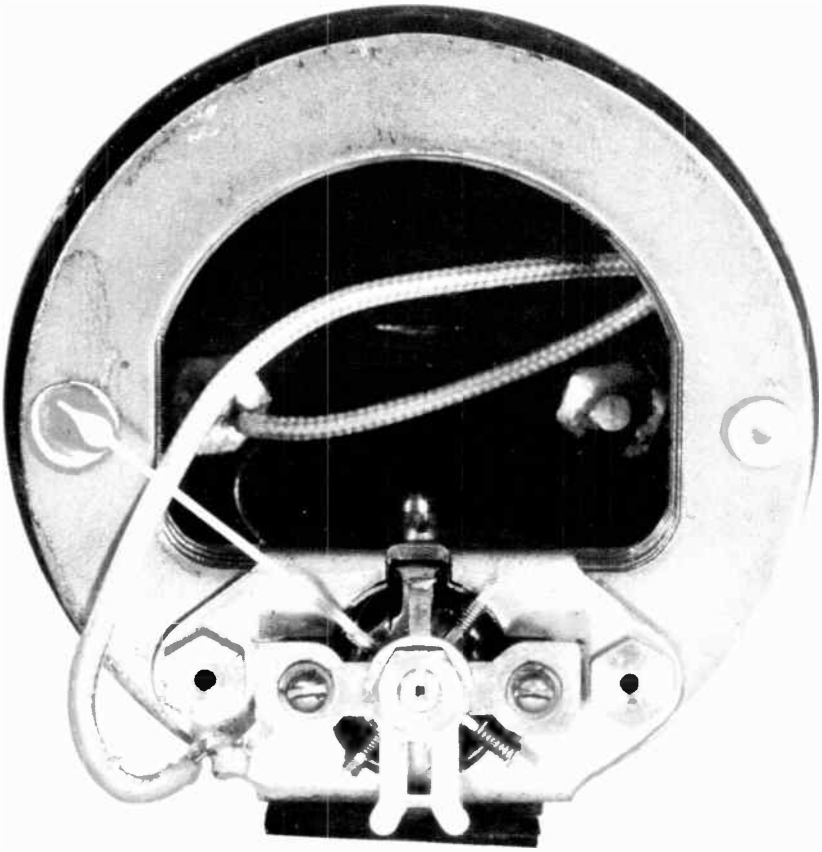


Fig. 16-7. In this "movement" of a current meter the large circular part is a permanent magnet.

soft steel. A sheet of light cardboard was laid over the steel blocks, just as over the poles of the permanent magnet in Fig. 16-1B, and iron filings were sprinkled onto the cardboard while current flowed in the coil. Fig. 16-9 is a photograph of the iron-filing pattern. Compare this picture with the one in Fig. 16-2.

Here we have seen the action of an electromagnet. An electromagnet consists of a center or "core" of iron or soft steel wholly or partially surrounded by a winding of insulated wire in which an electric current causes the iron to become a magnet. As soon as current ceases to flow, the iron core loses practically all its magnetic properties. Everything which has been said about magnetic lines of force, about their fields and assumed directions, about attraction, repulsion, and torque, and about north and south poles in general applies to electromagnets exactly as to permanent magnets.

In Fig. 16-10 we have the same coil of insulated wire as in Fig. 8, but now the iron core has been removed. Note that the field is weaker and not as concentrated as with the iron core; also, the north and south poles have been brought to the ends of the coil rather than the ends of the soft steel blocks which extended beyond the ends of the coil.

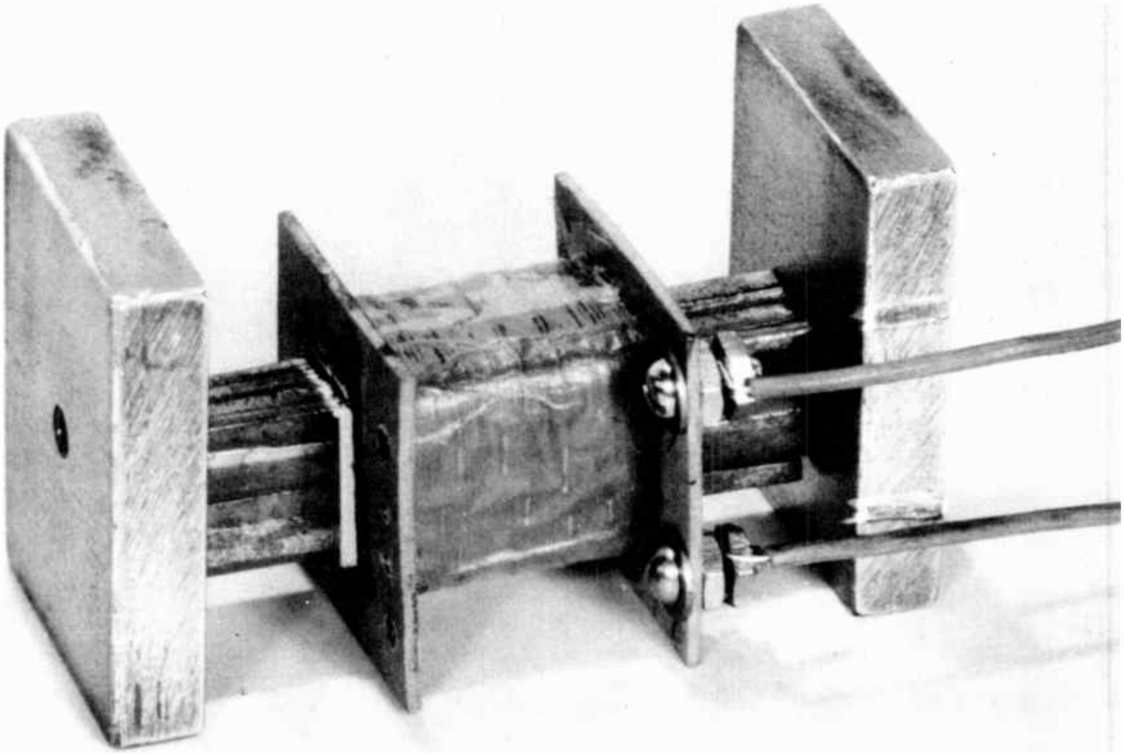


Fig. 16-8. The experimental electromagnet used for mapping field lines.

(10) Often it is convenient, sometimes necessary, to be able to remember the relations between direction of electron flow around the turns of a solenoid or electromagnet and which of the ends will be the north and south poles. These relations are pictured in Fig. 16-11. There are several rules which can help your memory. One is: With your left hand grasp the solenoid so the tips of your fingers point around the turns in the direction of electron flow and your thumb extends along the length of the winding. Your thumb then points toward the north pole, or points in the direction of lines of force passing through the center of the solenoid. Here is another: When looking at the north pole, electron flow is clockwise around the turns of the solenoid.

If we consider only a single turn of the solenoid at "A" in Fig. 16-11, with the correct direction of field lines through its center, this one turn may be shown as at "A" in Fig. 16-12. The lines of force which run all the way through the inside of the solenoid are the result of combining little circular lines of force which encircle every turn of the conductor forming the coil. These circular lines of force are shown around the conductor which forms the single turn. If the conductor is straightened out, as at "B", the relations between directions of the lines of force and of electron flow are as shown here.

There are rules which help you to remember the direction of lines around a straight conductor and electron flow through the conductor. If you grasp the conductor in your left hand, with your thumb pointing in the direction of electron flow, the tips of your fingers point in the direction of the lines of force. If you look at the positive end of the conductor, or the end toward which electrons are flowing, the lines of force encircle the conductor in a clockwise rotation.

MAGNETIC CIRCUITS. A magnetic circuit is a continuous path through which extend magnetic lines of force. There are magnetic circuits at "A" of Fig. 16-4, and in Fig. 16-6, where the sources of magnetic energy are permanent magnets. Fig. 16-13 shows the parts of a magnetic circuit in which the source of magnetic energy is the wire winding or coil around the top of the iron or steel core. Lines of force extend all the way through the core, including the part of the core surrounded by the winding, and the lines form the external magnetic field in the gap between the ends of the core.

Lines of force in a magnetic circuit are called the magnetic flux. Flux means a flow, but there is no real flow or movement – the idea of a flux or flow is merely a convenience for purposes of explanation and computation. We say that the flux follows a circuit because the lines always must be continuous, they must go all the way around from any starting point back to that same point. Magnetic flux is measured in the number of lines, or sometimes in the number of "maxwells," which is the same thing because one maxwell equals

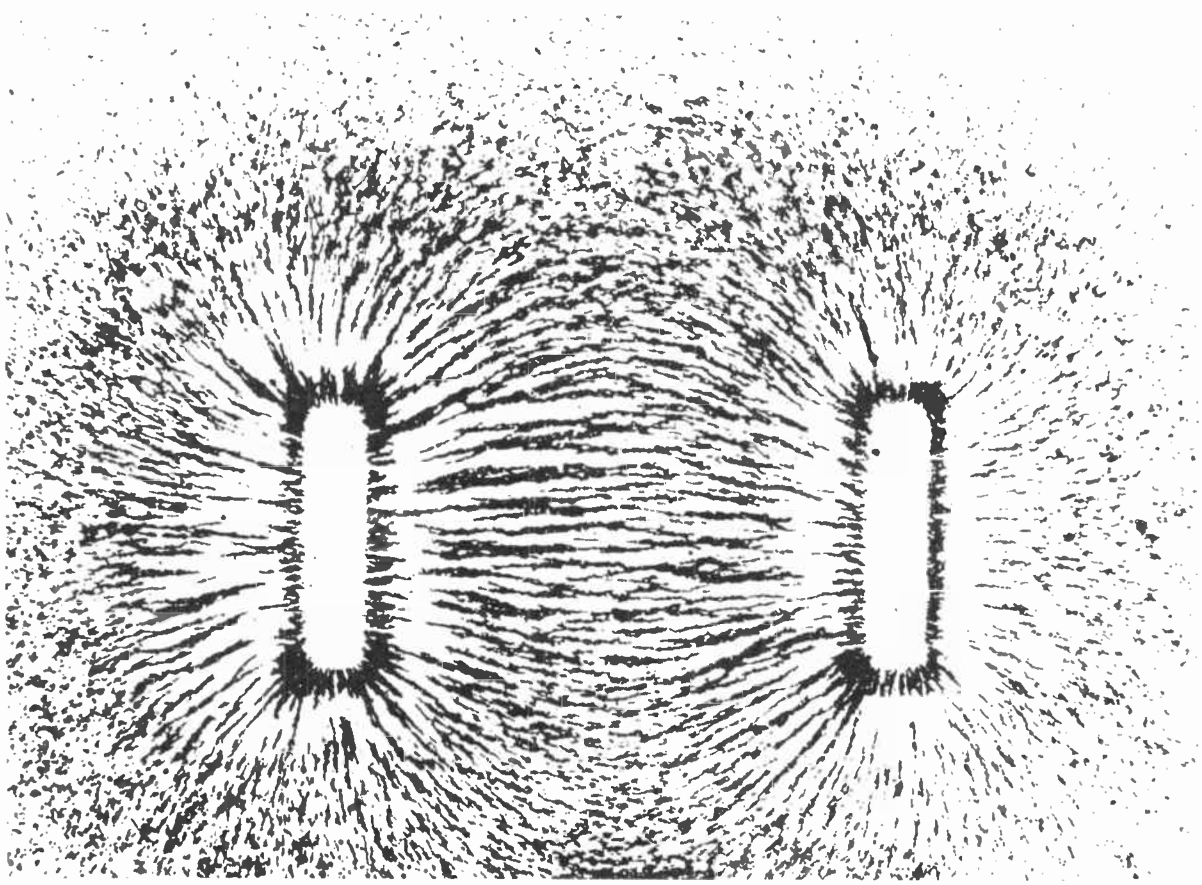


Fig. 16-9. The lines of force in the field of the electromagnet.

one line. Only rarely in radio and television will you encounter such magnetic words and units as flux, lines, maxwells, and others, but they will be mentioned in this lesson just in case you ever have need for the information.

The force which maintains the magnetic flux is called magnetomotive force, and is abbreviated mmf. Magnetomotive force maintains magnetic flux just as electromotive force maintains electron flow or current in electric circuits. Magnetomotive force is furnished by electron energy acting in the winding on a magnetic circuit. This force is measured in ampere-turns. The number of ampere turns is the product of the number of amperes of current and the number of turns in the winding. For example, ½ ampere and 100 turns gives an mmf of 50 ampere-turns. The same mmf of 50 ampere-turns would result from 1/10 ampere and 500 turns, or any other combination whose product is 50. Magnetomotive force sometimes is measured in "gilberts". One gilbert is equal to 0.796 ampere-turn.

Now we have lines of force of flux in the magnetic circuit which are comparable to electron flow or current in the electric circuit, and we have magnetomotive force in the magnetic circuit comparable to electromotive force in the electric circuit. What is comparable to resistance in the electric circuit? Reluctance

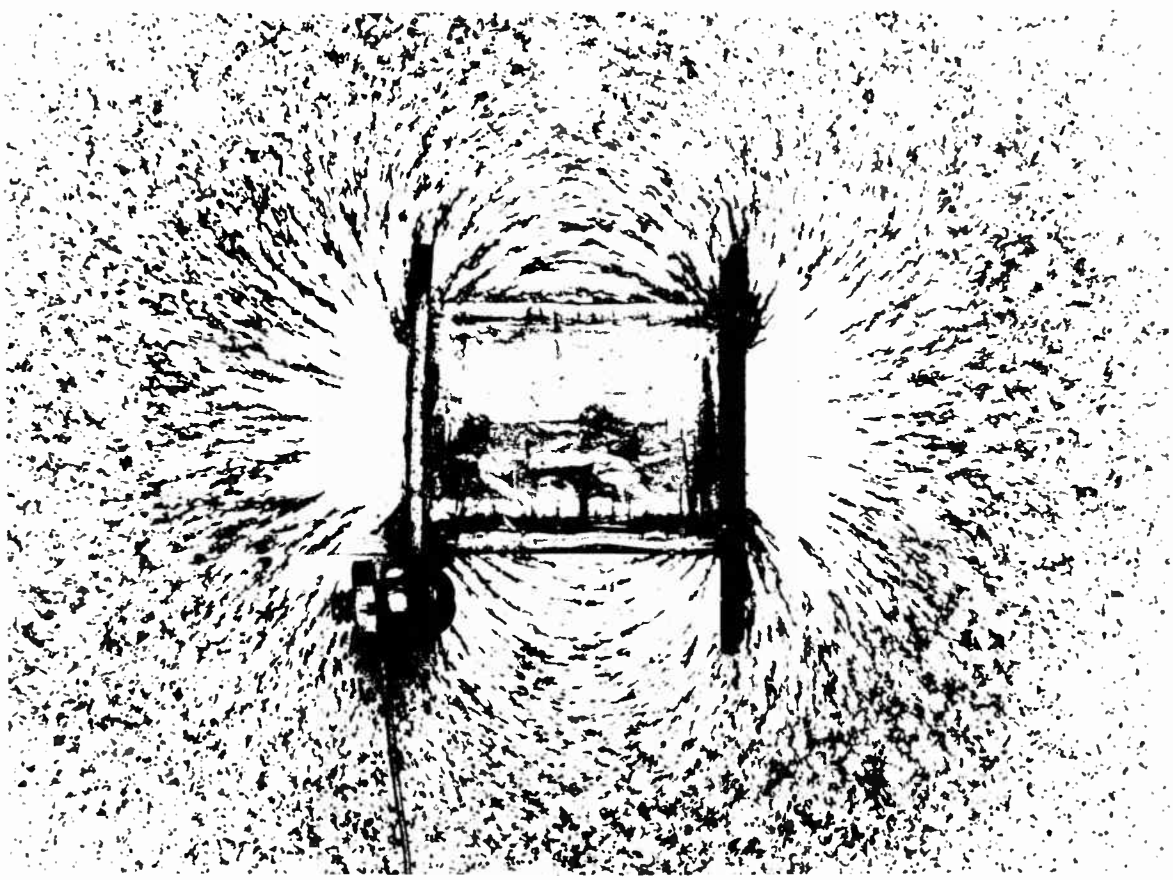


Fig. 16-10. The magnetic field of the coil or solenoid remaining when the core is removed from the electromagnet.

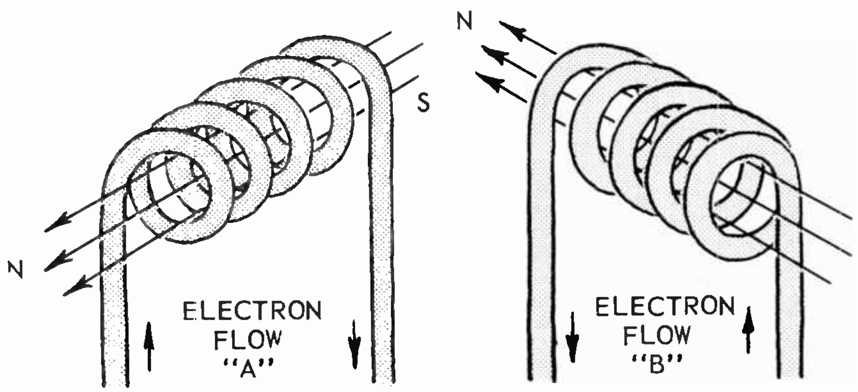


Fig. 16-11. Relations between direction of electron flow and positions of poles of a solenoid.

is the magnetic quality corresponding to electric resistance. Magnetic reluctance increases directly with length of the magnetic circuit, as resistance increases with length of the electric circuit. Reluctance decreases with an increase in the cross sectional area of the circuit in the same way as electrical resistance. Reluctance varies also with the kind of material in the magnetic circuit, as does resistance in the electric circuit, but reluctance varies too with the quantity of flux or with the flux density or concentration.

In the magnetic circuit we have a quantity or a property which has an effect on magnetic flux similar to the effect of dielectric constant in a capacitor on the rate of flow of alternating current in the capacitor circuit. This magnetic quantity is called permeability. The permeability of air is 1.0, just as the dielectric constant of air is 1.0. If you substitute iron for air in a magnetic circuit there will be an increase of flux with the same magnetomotive force, because the reluctance of iron is far less than that of air. The number of times the flux is increased is the permeability of the iron. If flux were 1000 lines with air, and increased to 1,500,000 lines with iron, the permeability of that iron would be 1,500.

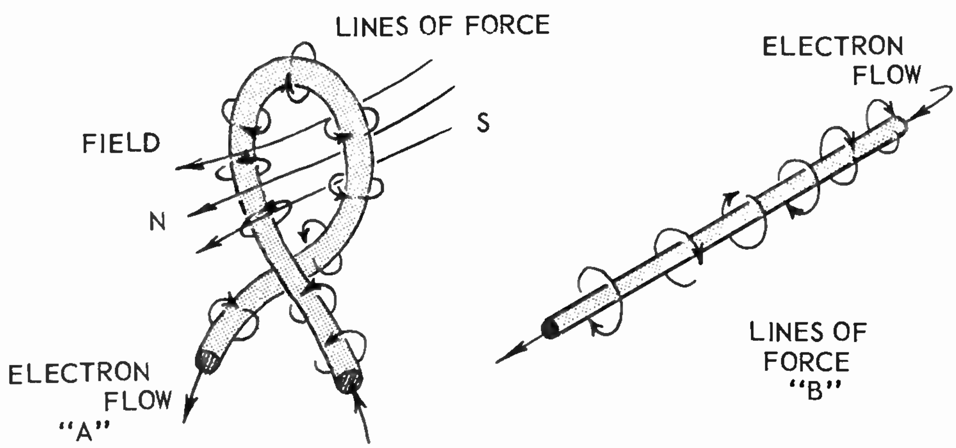


Fig. 16-12. The magnetic lines which encircle a conductor in which there is current.

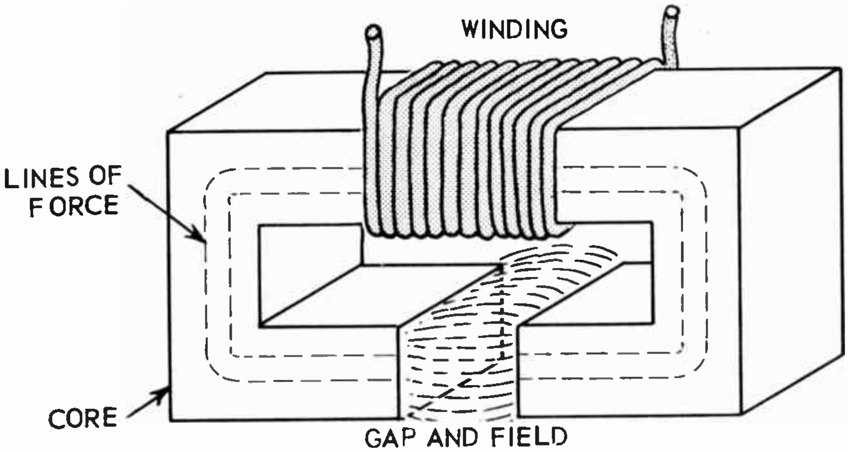


Fig. 16-13. The parts of a typical magnetic circuit.

Permeability of iron, steel, and other magnetic metals decreases as the flux increases. Permeability of air, and of all other non-magnetic materials, remains constant with its value of 1.0 when there are changes of flux. The permeability of iron, steel, and magnetic alloys is so very great in comparison with that of air that in a magnetic circuit containing an air gap the flux depends almost entirely on the width of the gap, not on the length of the metallic portion of the circuit. We may say that reluctance of an air gap will usually be hundreds or even thousands of times the reluctance of the entire remaining magnetic circuit in many common applications.

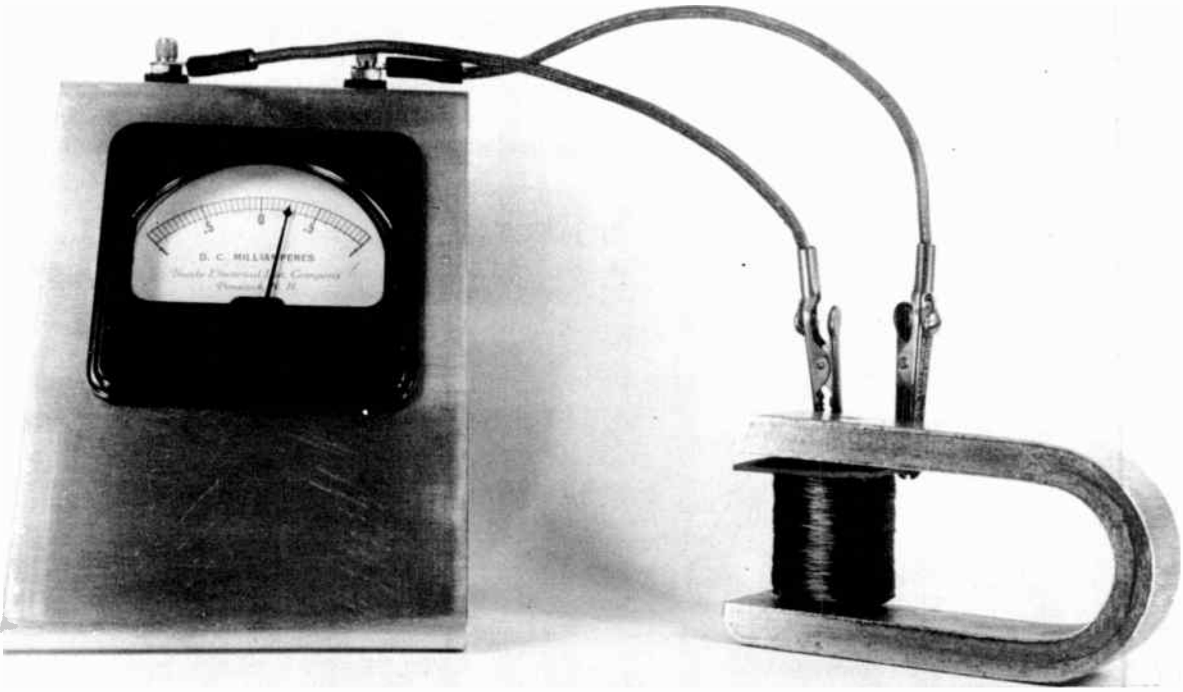


Fig. 16-14. Moving the coil into the magnetic field induces emf of one polarity.

The flux in any magnetic circuit increases directly with magnetomotive force and is inversely proportional to reluctance, just as current in any electric circuit increases directly with electromotive force and is inversely proportional to resistance. Magnetomotive force is used up in the reluctance of a magnetic circuit in much the same way that electromotive force of potential is used up in the resistance of an electric circuit. The required magnetomotive force depends on the desired flux and on the circuit reluctance, just as required electromotive force in volts depends on desired current in amperes and on circuit resistance in ohms.

ELECTROMAGNETIC INDUCTION

In preceding pages of this lesson we have seen how magnetic fields are produced by electric currents. Now we shall produce electric currents from magnetic fields. It took the world until the year 1830 to get as far as we shall get in this one lesson. Had men not learned how to produce electromotive force and electric current from magnetic fields we would have no television, no radio, no long-distance transmission of electric power, no electric generators, no important uses of alternating current, and not much of anything else in the way of electrical conveniences and necessities.

Until 1830 the most powerful magnet on earth would lift 9 pounds. Then Joseph Henry made an electromagnet that lifted 750 pounds when furnished with current from batteries. Later he invented a simple electric motor that worked with electromagnets. But Henry wanted to produce electric current from magnetic fields. He and another experimenter, Michael Faraday, both discovered how to do this at about the same

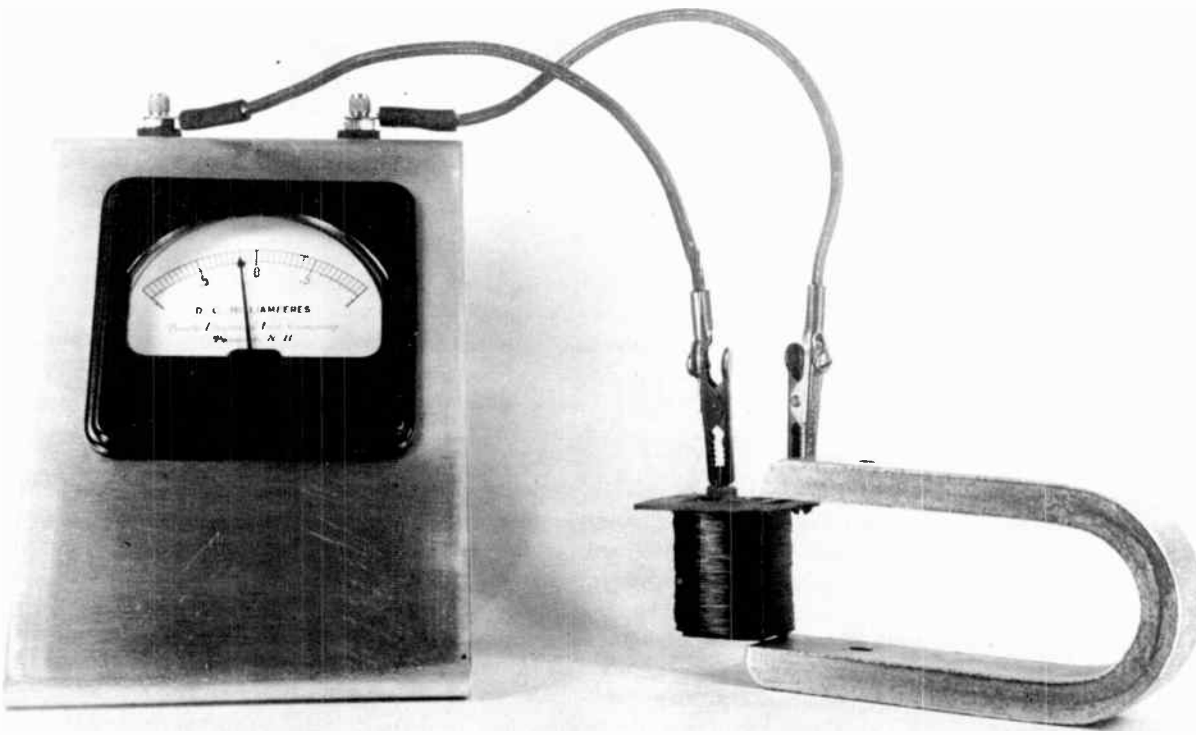


Fig. 16-15. Moving the coil out of the magnetic field induces emf of opposite polarity.

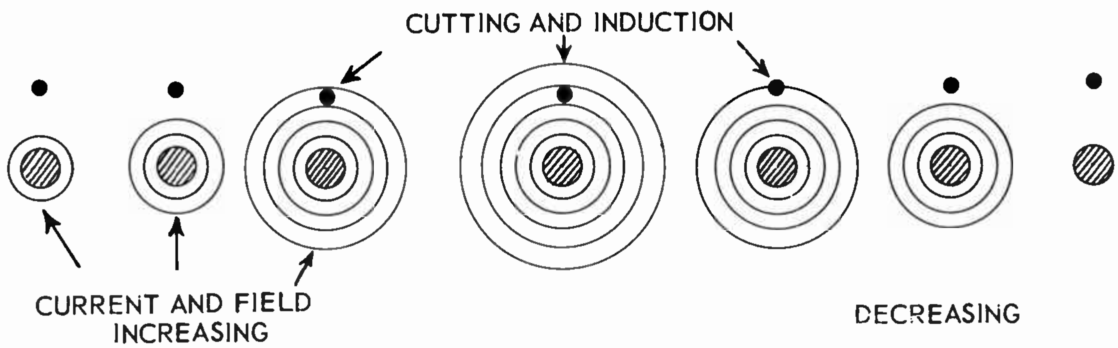


Fig. 16-16. Magnetic lines expand and contract around a conductor, and cut another conductor.

time and independently of each other, one here in America and the other in England. The production of electromotive force from magnetic fields is called electromagnetic induction, or more often is called simply by the name induction.

When producing magnetic fields from an electromagnet we must have moving electrons, in the form of electric current in the magnet winding. Stationary electrons can produce nothing in the way of magnetic fields. When it comes to producing emf and current from magnetic fields we again must have motion.

There are several ways of obtaining the necessary motion for electromagnetic induction. With one method we move a conductor, containing free electrons, into or out of the stationary field of either a permanent magnet or an electromagnet. Then an emf induced in the conductor will cause flow of the free electrons in the moving conductor.

With a second method the conductor which contains free electrons remains stationary while a permanent magnet or an electromagnet and its field are moved toward and away from the conductor. An emf is induced in the stationary conductor, and will cause electron flow in the conductor.

With a third method an electron-carrying conductor and a field-producing electromagnet are placed close together, but both remain stationary while current in the electromagnet is increased and decreased. This change of current causes the magnetic field lines to first increase and move outward from the magnet, then to decrease in number and contract back into the magnet. These moving lines of the magnetic field pass through the stationary conductor, first one way and then the other. Movement of the field induces an emf in the conductor, and the emf will cause current.

It is especially important to note that in all three methods there is relative movement of magnetic lines or magnetic field and a conductor. Either the field or the conductor may move, or both may move at the same time, but at least one of them must move in order to have induction of an electromotive force. Usually we say that the magnetic field or the magnetic lines cut the conductor, or that the conductor cuts the field.

The first of the methods of obtaining electromagnetic induction is illustrated by Figs. 16-14 and 16-15. In Fig. 16-14 a coil of many turns of wire is being moved into the field which is between the poles of a powerful magnet. The ends of the coil are connected to the terminals of a milliammeter. The emf induced in the conductor which is the coil is causing current to flow in the closed circuit consisting of coil, leads, and meter, and the pointer of the meter is deflected to the right. Deflection continues only so long as motion continues. When the coil comes to rest, the meter pointer will return to zero.

In Fig. 16-15 the coil is being rapidly withdrawn from the field between the poles of the magnet. Again there is deflection of the meter pointer, but now toward the left. Polarity of the induced emf and the direction of resulting electron flow depend on the direction of cutting, and reverse when there is reversal of the direction of cutting.

These two photographs might illustrate also the second method of causing electromagnetic induction, for it is easy to realize that holding the coil stationary while moving the magnet toward and away from the coil would produce exactly the same results as pictured.

We should take note of the fact that work is done and energy expended in moving either the coil or the magnet. The energy used for moving the coil or the magnet is mechanical energy, and it is changed into electric energy in the moving electrons of the induced electric current.

Fig. 16-16 illustrates the principle of the third method of obtaining induction, with both conductor and magnet stationary, and with movement of only the field lines. Here are represented end views of two conductors. The lower conductor is in the electromagnet, or might be a single current-carrying conductor. Emf is induced in the upper conductor. As current increases in the lower conductor, the magnetic field or the concentric field lines increase in number and move farther and farther out around this conductor. The outermost field lines cut through the upper conductor. While current decreases in the lower conductor, the field lines become fewer and shrink back to and into this conductor. In doing so, these lines again cut the upper conductor, but in a reversed direction.

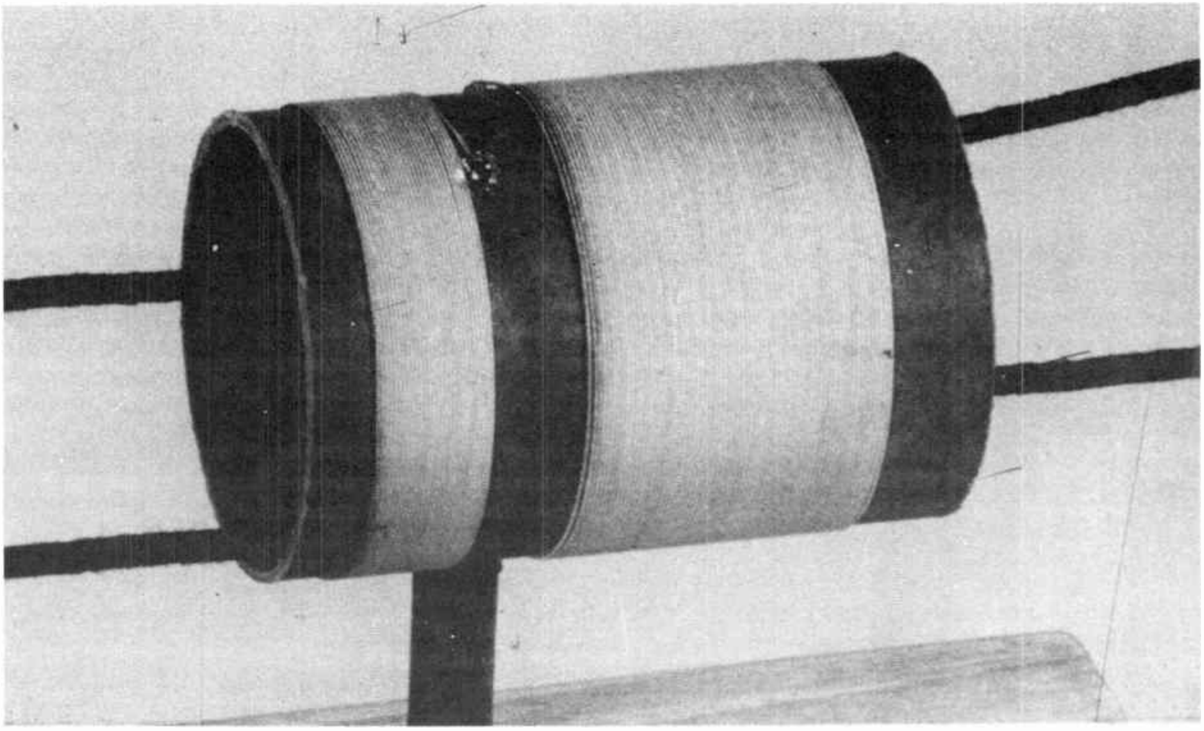


Fig. 16-17. These two windings form a transformer which might be used at radio frequencies.

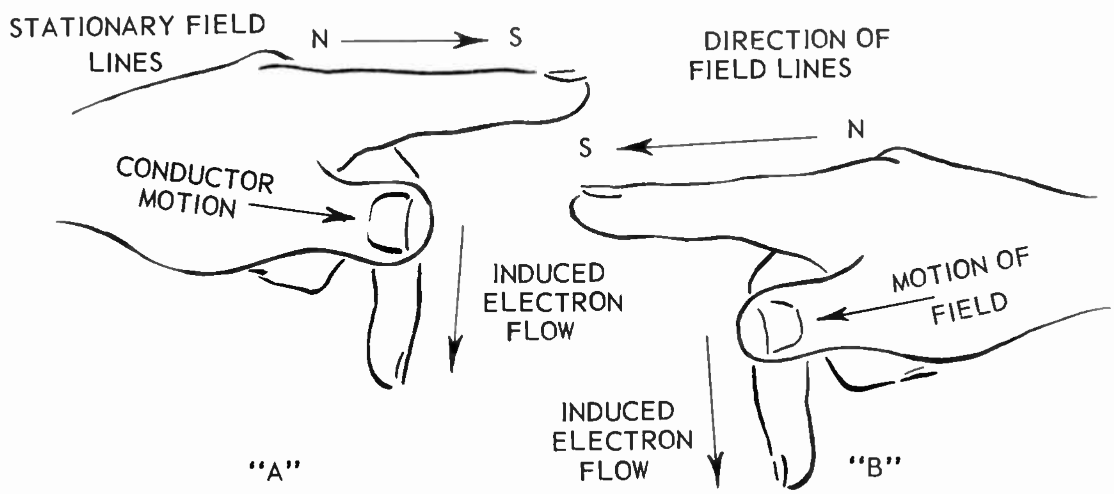


Fig. 16-18. Directions of the field lines, of motion, and of induced electron flow with electromagnetic induction.

The emf induced by outward moving field lines, with magnet current increasing, will be of one polarity. The polarity of the induced emf will be reversed while the field lines move inward. Doubtless you already have realized that if current is alternating in one of these conductors the magnetic field will move alternately out and back, and will induce in the other conductor an emf which alternates in polarity. This is the action discovered by Henry during his first successful experiment which is on record. It is the principle of the transformer. Of all important things in radio and television, all the way from antenna and power supply right through to the sweep circuits and the loud speaker, transformers certainly are among the most important. The reasons will appear as we employ transformers for more and more different kinds of work in lessons to come.

2 *MUTUAL INDUCTION.* A transformer utilizes the action called mutual induction. When there is mutual induction, every change of current in one conductor or one circuit causes emf's and changes of emf in another nearby conductor or circuit. If the nearby conductor is part of a complete or closed electric circuit, or if the nearby circuit is closed, the induced emf's will cause corresponding induced currents in the nearby conductor or circuit.

One of the simplest types of transformer is pictured by Fig. 16-17. There are two separate windings, each with its pair of terminal connections, on a form or tubing of insulating material. Either winding may be connected to a source of current. Whenever current starts in that winding, a magnetic field will arise and the field lines will cut through the other winding. This will induce an emf in the other winding, and if that other winding is part of a closed celectric circuit there will be induced current in the other winding and circuit. When current is stopped in the first winding, its magnetic lines will collapse or shrink back and disappear. In doing so the lines will cut back through the other winding and induce in that other winding an emf whose polarity is opposite to that of the first emf.

SELF-INDUCTION. If you give some thought to the cutting of magnetic lines of force through "nearby conductors" you will realize that these moving lines of force must cut also through the conductors from which the lines originate. Supposing, for instance, there is a change of current in either of the windings of Fig. 16-17. Since current must be the same everywhere in the series connected turns of the winding, there will be changes of current in every turn, and around every turn there will be lines of force which expand

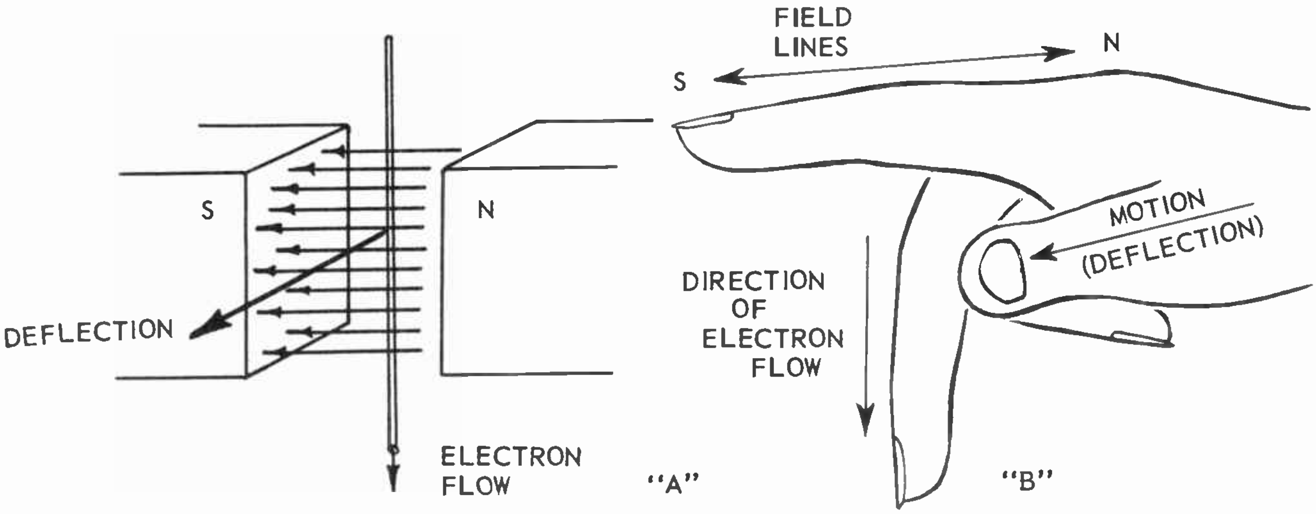


Fig. 16-19. Direction in which a conductor or the electron beam of a picture tube is deflected out of a magnetic field.

and contract with variations of the current. The moving lines from every turn must cut through the turns on both sides and also through turns farther away if these moving lines are ever to reach the other winding.

7 The result of this is that emf's are induced in the conductors wherein there are changes of current that produce movement of the magnetic lines. This is called self-induction. Every coil, circuit, or conductor in which there are changes of current, induces in itself emf's as a result of moving magnetic lines.

A most important feature about this self-induced emf is its polarity. If the self-induced emf were to be in such polarity as to assist the emf or voltage which is causing the current we might have perpetual motion, the current could keep itself flowing. Naturally, there is no such action. The polarity of the self-induced emf always is such as to oppose the change of current that is causing the movement of the field that induces the emf. That's something to think about.

5 The significance of the polarity of induced emf is easier to comprehend if we say this: The induced emf always is of such polarity as to try to keep the current flowing when current is decreasing, and to try to keep current from increasing when it actually is increasing. Self-induced emf sometimes is called counter-emf or back-emf. When current is increasing, and the magnetic field expanding, the polarity of the counter-emf is opposite to that of voltage or potential difference being applied to the circuit. When current is decreasing, and the magnetic field contracting, the polarity of the counter-emf is the same as that of the applied voltage or potential difference.

It is counter-emf that causes sparking or arcing at the contacts of a switch as the switch is opened to stop the current. The counter-emf is determined to keep the current flowing, and it builds up such a high voltage across the break as to maintain current in the form of an arc after the switch contacts have started to separate. When circuits have many coils or large coils, and lots of induction, the voltages of counter-emf may become so great as to puncture the insulation unless suitable precautions are taken.

DIRECTIONS OF MOTION, EMF, AND CURRENT. At "A" in Fig. 16-18 is illustrated a rule which helps us to remember the relative directions of magnetic lines in a stationary field, of conductor motion through the field, and of direction of electron flow induced in the conductor. Extend the thumb, forefinger, and middle finger of the left hand so each is at right angles to the other two. With the forefinger pointing in the direction of magnetic lines in the field (north to south pole) and the thumb pointing in the direction the conductor is moved through the field, the middle finger points in the direction of induced electron flow.

If the conductor is stationary and the field lines move, the rule may be followed by using the right hand as shown by "B". Then the forefinger must point in the direction of the field lines (north pole to south pole), the thumb in the direction that the field is moved, and the middle finger will point in the direction of induced electron flow in the stationary conductor.

DEFLECTION OF CONDUCTOR OR ELECTRON BEAM. Closely related to the matter of electromagnetic induction is that of deflection of a current-carrying conductor or of an electron stream in a magnetic field. The principle of conductor deflection is used in loud speakers, and in all electric motors. Electron beam deflection in a magnetic field is the principle of magnetic deflection for television picture tubes.

As shown by Fig. 16-19, the direction of field lines (north to south), the direction of electron flow in the conductor or beam, and the direction that the conductor or beam is deflected out of the field, all these are at right angles to one another. At "A" in Fig. 16-19 the field lines extend from right to left, the electron flow is downward, and the deflection is out of the paper.

A hand rule for electron beam or conductor deflection is illustrated in Fig. 16-19 "B". The forefinger, middle finger, and thumb of the right hand are held at right angles to one another. With the forefinger pointing in the direction of field lines (north to south) and the middle finger pointing in the direction of electron flow (in conductor or beam) the thumb points in the direction that the conductor or electron beam is deflected out of the field.

The force of deflection increases with increase of rate of electron flow and with field intensity, which is the number of lines per square inch of cross section of the field. Deflection force is increased also by having a greater length of conductor or electron beam within the magnetic field.

The reason for deflection is that two fields are brought together in the same space. One field consists of the magnetic lines between the magnet poles. The other field consists of the magnetic whirls around the conductor or around the electron beam, as shown by Fig. 16-12. When lines in these two fields are parallel a force is developed. If the two sets of lines are in the same direction the fields repel, and if the lines are in opposite directions the fields attract. There is no great need to go into long explanations about this matter of field reactions, the result is shown clearly and more simply by Fig. 16-19.

TELEVISION

LESSON NO. 17

INDUCTORS AND THEIR BEHAVIOR

"Training Your Mind"

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind

C.T.I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

LESSON NO. 17

INDUCTORS AND THEIR BEHAVIOR

THEORY OFTEN PAYS DIVIDENDS

This is another lesson in which a great number of important facts are compressed into small space, because we still are in a hurry to get to the direct practice of television and radio service operations. Probably the man already familiar with servicing, especially in television, would say right here that time constants, reactances, inductances, and impedances described in this lesson are directly and intimately related to all kinds of adjustments, replacements, and methods of trouble location. But until we get further along in our own work, these subjects are bound to seem somewhat theoretical. Later you will realize that it is the so-called theoretical information that lets you figure your own way out of difficult situations, where the ordinary methods and commonly available instructions won't solve the problems.

No matter how we look at things, it is essential to have at least a speaking acquaintance with all this groundwork. Whether you ever actually wind one of your own coils which is to furnish a certain number of microhenrys for a television sound trap makes little difference. The real point is, you should know where to get information that will let you compute the necessary number of turns if you ever are confronted with the necessity.

This line of reasoning applies in a general way to all the formulas in this lesson. If you don't like arithmetic you can skip the formulas, for now, but someday you will come looking for them. A shop mechanic in an auto service station can get along without a micrometer, but the boss uses one, and when the mechanic gets good enough he will do likewise.

INDUCTORS AND THEIR BEHAVIOR

When electrons move in a conductor or coil the result is a magnetic field in the surrounding space. That is electromagnetism. When there is relative movement between a magnetic field and a conductor the result is electromotive force and movement of electrons in the conductor. That is electromagnetic induction.

7 We may start with electrons in motion, let them produce a magnetic field, then let the field collapse to move electrons. Or we may start with a magnetic field, let the field collapse to produce emf which moves electrons, and let the moving electrons produce another magnetic field. This alternate changing of electron movement into a magnetic field and of the magnetic field into electron movement is almost all we need in order to have resonance. Resonance is the thing that makes tuning possible. Tuning lets us pick certain signals and certain frequencies from all the other signals and frequencies in existence. Without tuning there would be no radio or television.

ENERGY IN MAGNETIC FIELDS. What really happens with the magnetic field and the electron flow is an exchange between two forms of energy. Long ago we learned that electrons in motion possess energy,

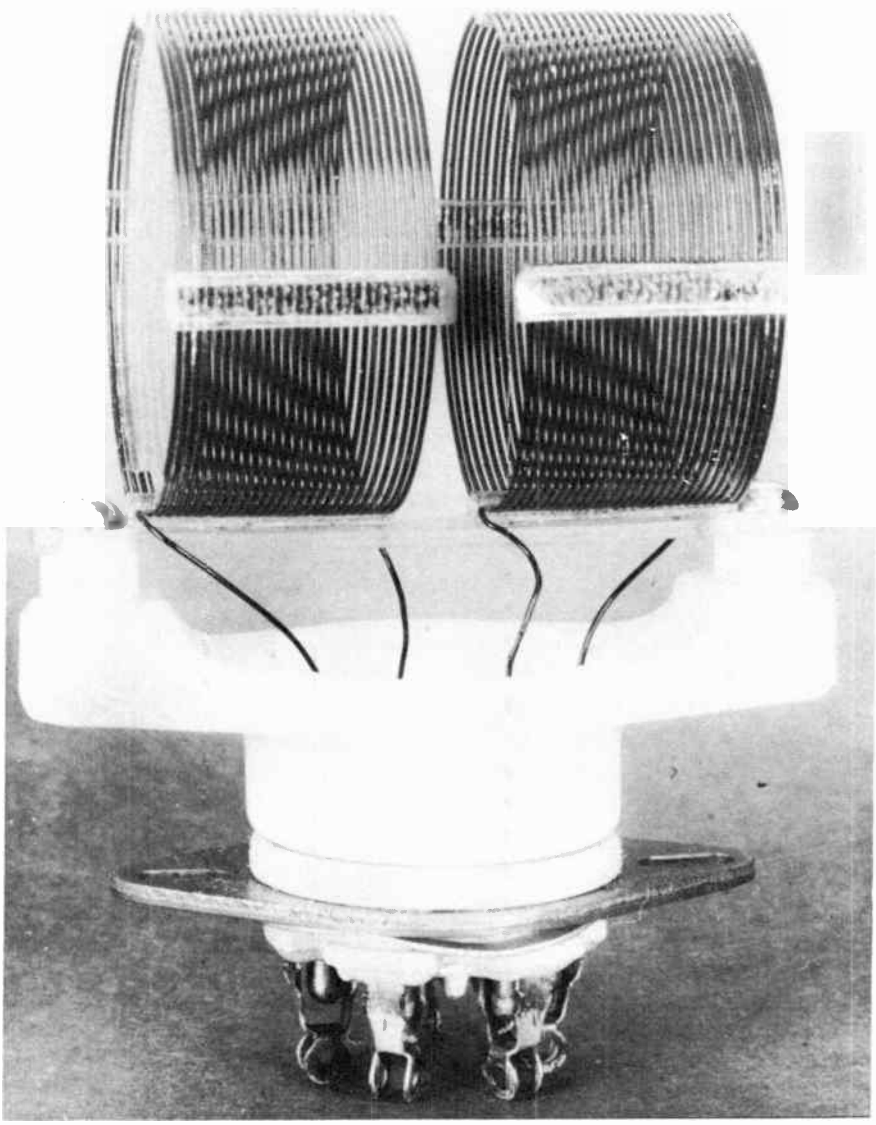


Fig. 17-1. These inductors are tuned to resonance in short-wave radio.

which is the ability to do work. Electrons which flow in a conductor can use their energy of movement to do two things; they can produce a magnetic field and they can produce heat. Part of the electron energy always must go into production of heat which results from work done in overcoming resistance. The remainder of the electron energy does the work of building up a magnetic field. There is energy in this magnetic field.

The work done by the moving electrons in building up a magnetic field is comparable to the work you do in lifting a weight. When you get the weight raised off the ground there is energy in the weight. When moving electrons force the magnetic field out into the surrounding space there is energy in the field. When your elevated weight falls back to earth it will do work. When the magnetic field collapses back into the conductor it will do the work of electromagnetic induction, which moves electrons.

The similarity between the weight and the magnetic field really is remarkable. It takes work to lift the weight and it takes work to form the field. But if the weight is not again moved after you once elevate it, no further work is being done on the weight (because, technically, there is no work unless there is motion). If the magnetic field is built up to just some certain strength and maintained there, the moving electrons will have to do no further work on the field. The moving electrons have to do work to form the field, but not while the field is maintained at a steady value or position in space.

So long as electron flow continues, work must be done to overcome resistance and produce heat. But no inductive work is done to maintain the magnetic field at a fixed strength. If the magnetic field has previously been built up to some certain strength, and no further electron work is being done on this field, it will require additional work to make the field stronger or to push it farther out into the surrounding space. These relations are illustrated by Fig. 17-2.

When electron flow or current is allowed to decrease, after having established a magnetic field of certain strength, the field will fall back to a value and position corresponding to the reduced current. As the field contracts it must give up some of its energy, because for any certain current in a particular conductor or

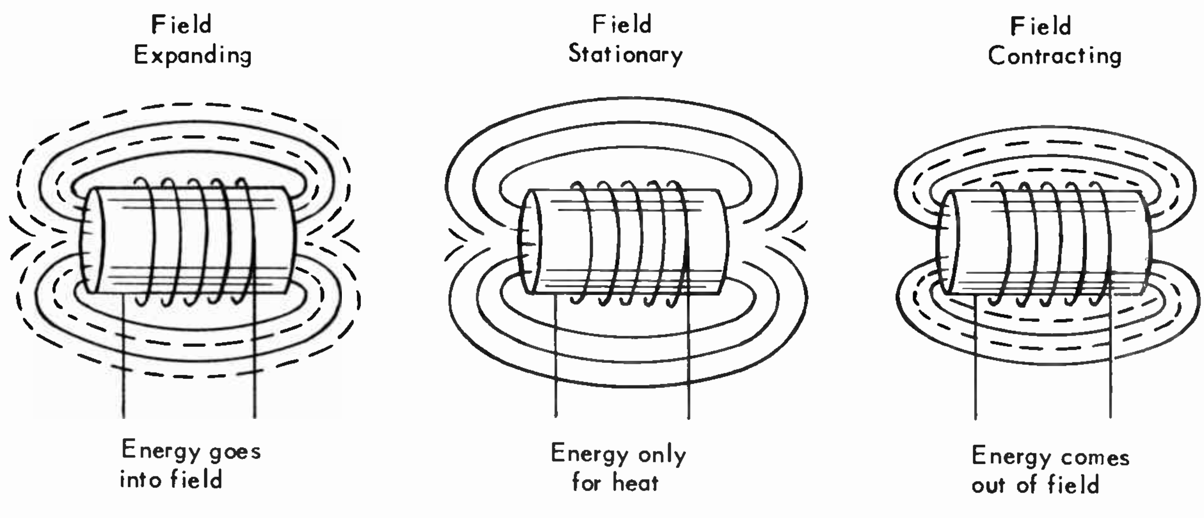


Fig. 17-2. Energy stored in a magnetic field is returned to the circuit.

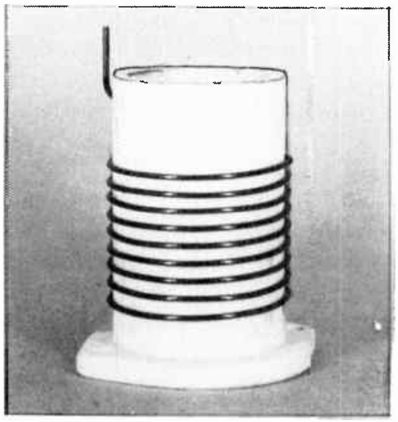


Fig. 17-3. A space-wound air-core coil for tuning at high frequencies.

circuit there is only one corresponding value of magnetic field energy. The energy given up by the field goes into the force of induction, producing in the conductor or circuit a counter–emf of such polarity as opposes the decrease of current. That is, the counter–emf is in the same direction as that of the external voltage being applied to the circuit, or of the voltage which was causing current in the first place.

9 We have been talking about work done in establishing a magnetic field, about energy temporarily stored in the field, and about work that the field will do when it collapses. The energy in a magnetic field is no theoretical thing, it can be measured just like the energy paid for on bills for electric light and power. The magnetic energy increases with increases of current and with ability of the circuit to utilize this current for production of magnetic flux. All circuits and coils are not alike in this flux-producing ability, which is called inductance.

A given current, in milliamperes, flowing in the coil of Fig. 17-1 will produce around that coil a magnetic field far stronger than will be produced around the coil of Fig. 17-3 by the same current. This is because the larger coil, with more turns, has much more inductance than the smaller one with fewer turns. It won't be long until we get well acquainted with inductance.

ENERGY IN ELECTRIC FIELDS. While we are on the subject of energy in magnetic fields is a good time to discuss some rather similar effects in the electric fields which are between the plates of a charged capacitor. We know that while electrons flow into and out of the plates of a capacitor the electrons possess energy, because they are in motion. But when the plates are fully charged, and remain with neither discharge nor further charge, the electrons are stationary. Where is their energy which was due to motion? That energy has gone into the electric field or into the dielectric which is between the plates. The quantity of energy temporarily held in the electric field increases with increase of potential difference between the plates, and increases also with greater capacitance of the capacitor.

When the charged capacitor is connected to an external circuit, the capacitor will discharge into the circuit. The result is an electron flow in the circuit as the excess of free electrons leave the negative plate of the capacitor and flow around to make up the electron deficiency in the positive plate.

9 In the process of capacitor charge and discharge there is an exchange between two forms of energy. First, an electron flow will charge a capacitor and build up an electric field between the capacitor plates. Energy which was in the moving electrons then is in the electric field. As the capacitor discharges, and the field

disappears, energy which was in the field produces electron movement. These actions are illustrated by Fig. 17-4. We may start with moving electrons, produce an electric field, then let the field disappear in forming a new movement of electrons. Or we may start with an electric field (a charged capacitor), let its energy cause movement of electrons, and let the moving electrons produce another electric field and another charge.

1 *EXCHANGES OF ENERGY.* We must realize, and remember, that no energy need be lost when moving electrons produce either magnetic fields or electric fields. All the energy that goes into the formation of a magnetic field as the field is pushed outward can return to the conductor or circuit when the field collapses and induces an emf in the circuit. All energy which goes into formation of an electric field in a capacitor may go right back into the circuit when the capacitor discharges and the electric field disappears. These things are true because either a magnetic field or an electric field is capable of causing electron flow. There are merely exchanges in the form of energy, the energy is either in the form of electron movement or else in the form of a magnetic or an electric field. The energy remains in the circuit or in certain parts of the circuit.

On the other hand, energy used for overcoming resistance ordinarily is lost. Only in a few special cases (thermocouples) can heat produced in resistance turn around and cause another electron flow. In all ordinary circuits the heat is dissipated into air and surrounding objects, and the energy which produced the heat is lost so far as the original circuit is concerned. In all our common circuits heat energy is a loss but field energy is not a loss.

5 *TIME CONSTANTS.* As you read television service manuals and descriptions of how various circuits should operate you will find many references to time constants. Time constants relate to time required for formation of magnetic and electric fields, and to the time required for these fields to disappear. No matter what kind of circuit you work with, it is impossible for a magnetic field to increase to its full strength in no time at all, and it is impossible for an electric field and capacitor charge to reach full value in no time at all after voltage is applied. Neither can a magnetic or electric field disappear instantly after current stops

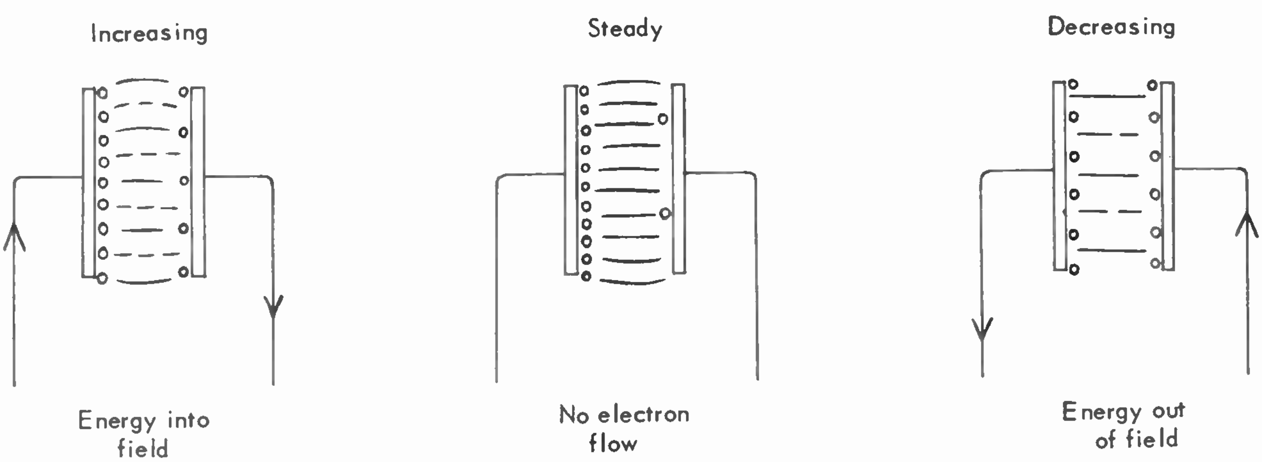


Fig. 17-4. Energy stored in an electric field returns to the circuit.

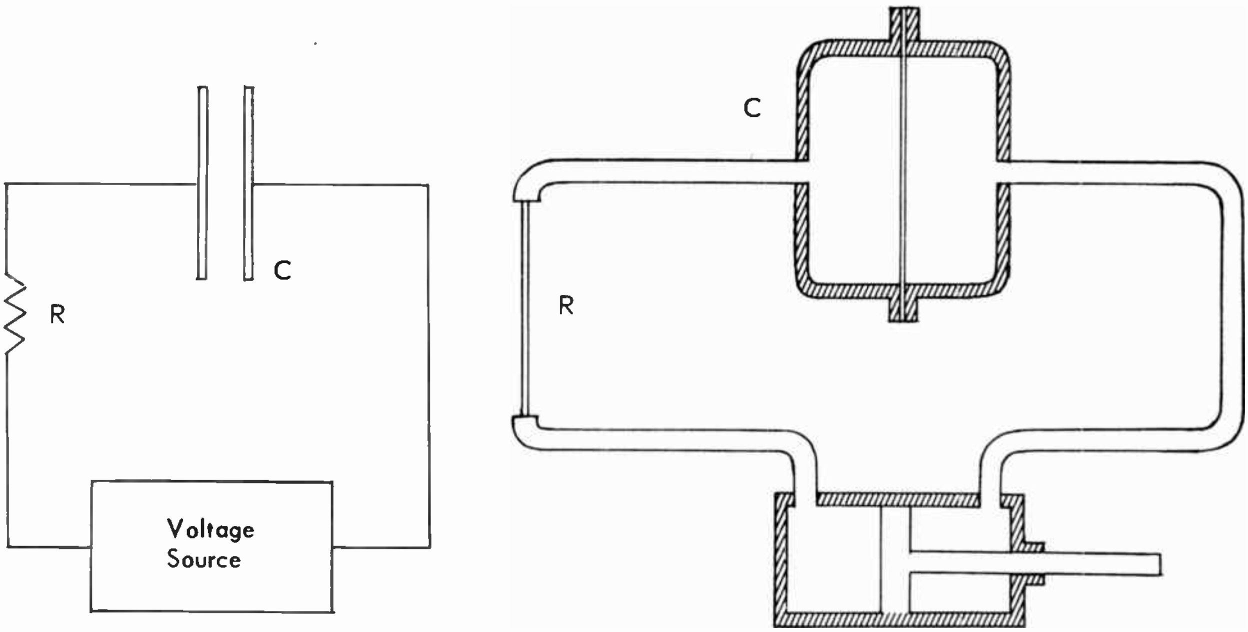


Fig. 17-5. The rate of capacitor charge and discharge depends largely on the series resistance.

or a discharge circuit is closed. Any of these actions might be completed in one or two millionths of a second, but there is a period of time – it is not instantaneous – and in many circuits the difference between one and two millionths of a second is all-important.

We shall talk first about capacitive time constants, the ones which have to do with capacitors. At the left in Fig. 17-5 is represented a capacitor *C* in series with resistance *R*. Electrons for charging and discharging the capacitor must pass through the resistance as they move between the voltage source and the capacitor. Always there is resistance in series with a capacitor. This resistance need not be in a resistor, it may be only the resistance of the connecting wires and of the metal in the capacitor plates, but it is there.

At the right in Fig. 17-5 the capacitor is represented by the two closed air chambers with a flexible diaphragm between them, and the source is a pump. Most of the piping between the chambers and the pump is large, but at *R* is a length of very small bore tubing. Supposing you suddenly apply a certain pressure at the pump, will the diaphragm between the chambers instantly deflect to a degree corresponding to that pump pressure? Of course not, for some time will be required for enough air to get through the small tubing to equalize pressures in the pump and in the air chambers. If you then release pressure on the pump piston, air will flow out of one chamber to the pump, and from the pump into the other chamber. But the diaphragm cannot return instantly to its neutral position, it can spring back only as air is forced slowly through the small tubing.

Similar things happen with the capacitor and the electrical resistance. Capacitor charge and discharge are measured in coulombs of electrons. Flow through the resistance is in coulombs per second (amperes). The rate of flow is directly proportional to potential difference in volts and inversely proportional to resistance in ohms. Remember, $I = E/R$, amperes equal volts divided by ohms. The capacitor cannot be instantly charged to a degree corresponding to applied voltage, nor can it instantly discharge, because the resistance slows down the coulombs per second that can flow into and out of the capacitor. If you wait long enough, a sufficient quantity of electrons will get through the resistor to make the capacitor charge proportional to voltage at the source, but there will be some delay.

The total quantity of electrons eventually forced into and out of the capacitor will depend on two things; on capacitance of the capacitor, and on voltage applied across the capacitor and resistance. The number of coulombs of charge is equal to the product of capacitance in farads and potential difference in volts.

It would take quite a while to completely charge a capacitor with a quantity of electrons proportional to applied voltage. The reason is shown by Fig. 17-6. At *A* there is zero potential difference between the capacitor plates, and the entire voltage of the source is effective in forcing electrons through the resistance. But as the capacitor gains more and more charge, the capacitor itself acquires a potential difference as at *B* and *C*. This potential difference or voltage of the capacitor is of such polarity as to oppose the electron flow for further charging. The charging rate becomes slower and slower as the capacitor voltage more nearly equals the source voltage. The last of the charge will flow through the resistance very slowly, but finally, as at *C*, the charge will be complete, with capacitor voltage equal to source voltage, and with no electron flow.

At *D* in the diagram the source has been taken out of the circuit and the capacitor plates are connected together through a conductor. Now the full voltage of the charged capacitor is effective in forcing electron flow through the resistance. But as the capacitor discharges, and potentials of its plates become more nearly equal, the capacitor voltage drops lower and lower. Then the electron flow rate through the resistance slows down, and discharge proceeds more and more slowly. Eventually there will be complete discharge, and zero potential difference between the capacitor plates.

Fig. 17-7 shows how the charging and discharging of every capacitor will vary with time. These two curves apply not only to paper capacitors, or to mica capacitors, or to electrolytic capacitors, but to every kind of capacitor and capacitance which can be charged and discharged. The curves apply to every value of capacitance, whether it is a fraction of a micro-microfarad, or hundreds of microfarads.

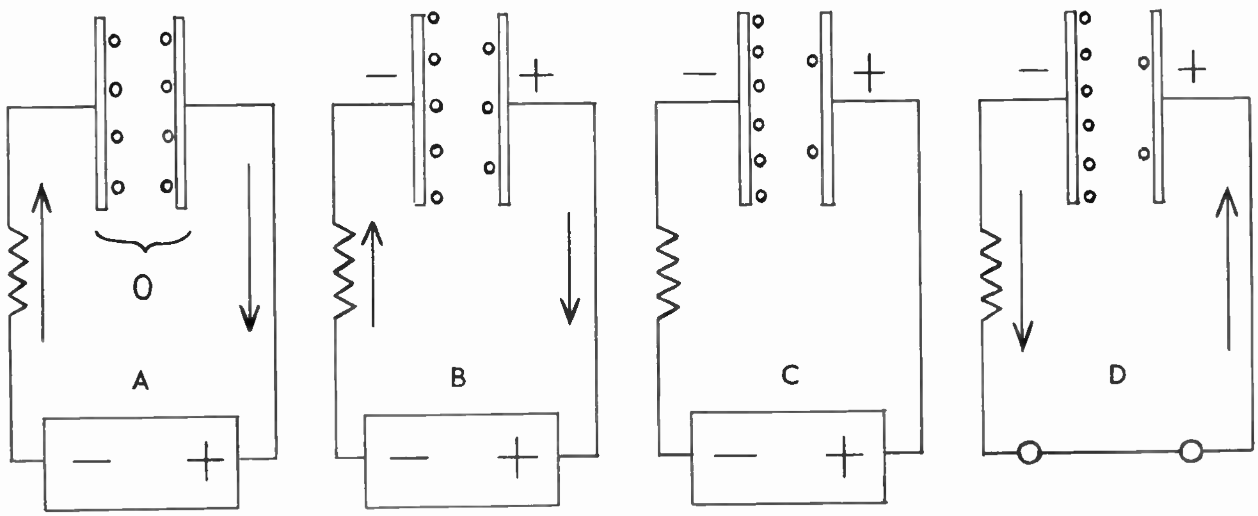


Fig. 17-6. Electron flow stops when capacitor voltage becomes equal to source voltage.

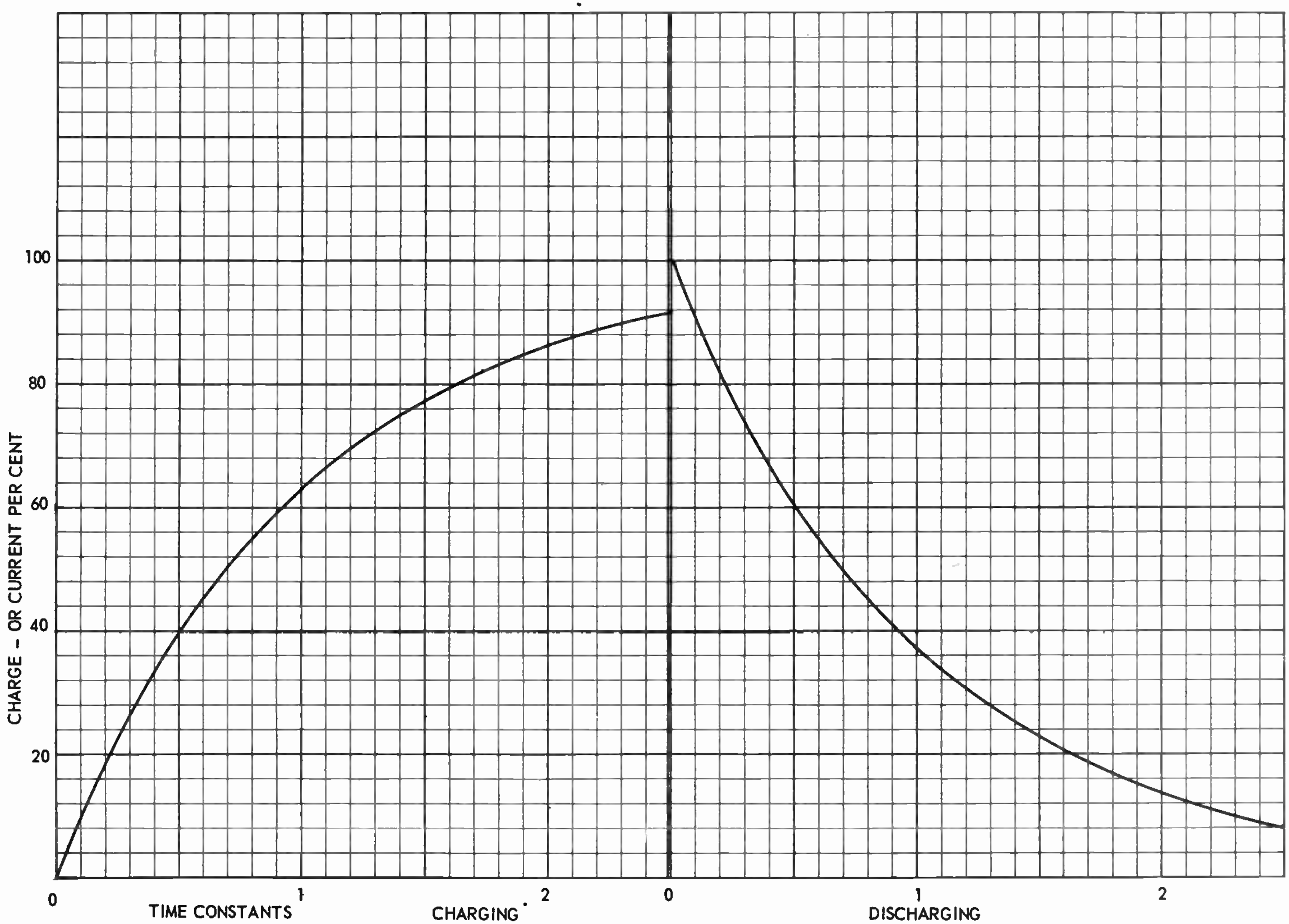


Fig. 17-7. Charge and discharge of a capacitor, or rise and fall of current in an inductor, in relation to time measured in time constants.

The left-hand vertical scale shows percentage of charge which has been acquired during charging or which remains during discharging. The bottom scale shows time in the unit called a time constant. The number of seconds or the fraction of a second in one time constant depends entirely on the capacitance of the capacitor and on the resistance which are in the circuit being considered. All you need do is multiply the number of microfarads capacitance by the number of megohms resistance, and the product is equal to one time constant in seconds.

Time constant, in seconds = microfarads × megohms

As an example, supposing you have a grid-leak bias circuit in which the capacitance is 0.00025 mfd and the leak resistance is 2 megohms. Multiplying 0.00025 by 2 gives 0.0005, which is the fraction of a second time constant (five ten-thousandths of one second).

Now look again at Fig. 17-7. During the time of one time constant the capacitor does not gain a full charge, the charge increases to only 63.2% of what it eventually will become if the charging voltage continues to be applied. During one time constant of discharge, shown on the right-hand curve, the capacitor has lost 63.2% of the original charge and has remaining 36.8% of that original charge. The original charge is considered to be 100%. Whether this original charge is 100% of all the charge that the capacitor could hold or is some smaller quantity makes no difference. It is all the charge there is when the discharging starts, consequently is 100% of all the electrons that can be discharged.

The curves show that at the end of two time constants the capacitor will have gained 86½% of its ultimate full charge, and at the end of two time constants of discharge will still retain 13½% of the charge with which it started. The discharge curve is merely the charging curve upside down. A capacitor would continue charging or discharging. At the end of three time constants the charge would have risen to about 95% of its ultimate value, and at the end of five time constants would be about 99-1/3% of the ultimate.

There are a number of important matters to be pointed out in connection with capacitive time constants. First of all, the time constant is not the time for complete charge or discharge, but only for 63.2% of the ultimate charge and only 63.2% of total discharge. Second, the time constant of a resistor-capacitor combination does not depend on absolute values of either capacitance or resistance, but only on their product. You can have the same time constant in fractions of a second with large capacitance and small resistance or with small capacitance and large resistance. Look at the following list of capacitances and resistances, all of which give the same time constant of 0.00024 second. Capacitances are given in both mfd and mmfd, and resistances in both megohms and ohms, because in order to use the formula you often have to convert other units into microfarads and megohms.

Microfarads	Micro-microfarads	Megohms	Ohms
0.008	8 000	0.03	30 000
.002	2 000	.12	120 000
.00024	240	1.0	1 000 000
.00003	30	8.0	8 000 000

Here is a third point. Applied voltage has no effect on time constant. If you increase the voltage on a given resistor-capacitor combination you increase the number of coulombs of ultimate 100% charge, because more voltage puts more charge into any given capacitance. Then, although the higher voltage does cause faster electron flow, it will take just as long as before to reach 63.2% of the greater ultimate charge.

A time constant may be anything from the most minute fraction of a second up to many minutes. The smaller the product of capacitance and resistance, and the shorter the time constant, the faster a circuit

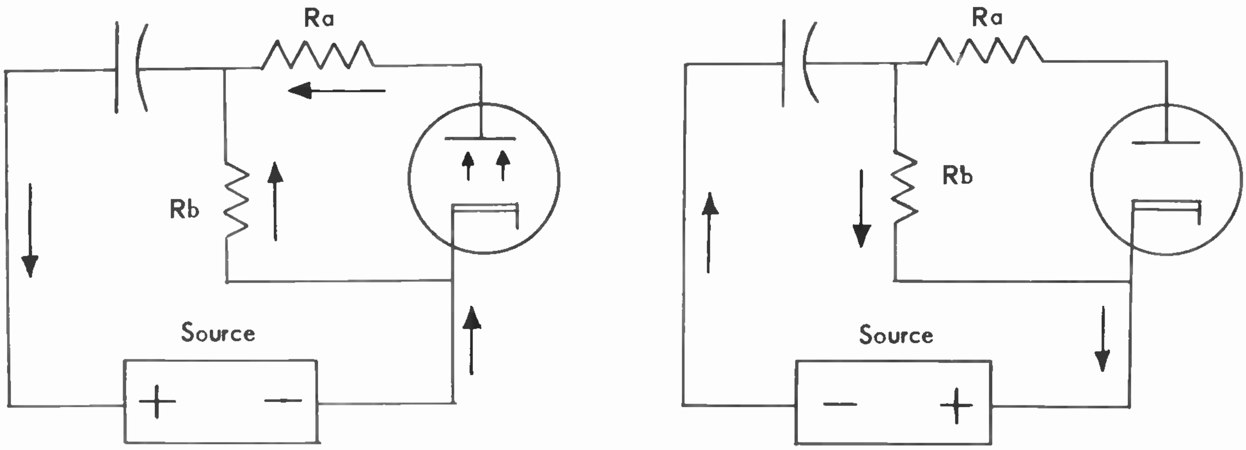


Fig. 17-8. The tube makes the time constant for charge shorter than for discharge.

will act and the more rapidly it will respond to any changes of voltage. At high frequencies of charging voltage there is time during one alternation for only a small portion of the ultimate possible charge unless the time constant of the circuit is very short.

The time constant for any given combination of capacitance and resistance is the same for both charge and discharge. But it is possible to switch resistances so that there is one time constant for charging and a different constant for discharging. One method is shown by Fig. 17-8. With source polarity as at the left the capacitor is charged by electron flow through the diode tube and resistor *Ra*, and at the same time through resistor *Rb*, When polarity of the source reverses, as at the right, the tube is made non-conductive (cathode positive and plate negative) and the capacitor discharges through resistor *Rb* alone.

In Fig. 17-9 the resistor *Ra* is connected across the capacitor. Potential difference across the capacitor and across resistor *Ra* always will be equal, but *Ra* does not control the time constant for charging. This charging time constant is controlled by series resistor *Rb*. When the switch *Sw* is opened, as at the right, the capacitor discharges through resistor *Ra*, and there is no electron flow through resistor *Rb*.

3 *INDUCTIVE TIME CONSTANTS.* If a direct voltage or steady voltage is applied to a circuit containing only resistance, with no capacitors or coils or inductors, current immediately reaches a value equal to E/R, or the number of amperes becomes equal to the quotient of dividing the number of volts by the number of ohms. If the circuit contains inductors the current in the circuit does not rise instantly to its final value but increases in exactly the same manner as shown by the left-hand curve of Fig. 17-7, which illustrates charging of a capacitor.

Things happen in this fashion. At the instant in which voltage is applied there is nothing except resistance to oppose flow of current, and there is a sudden surge of current. But this causes an equally sudden or rapid rise of a magnetic field out of the inductor. The lines of the field cut outwardly through the turns of the coil or inductor and induce a counter-emf. The polarity of this counter-emf is such as to oppose the applied voltage, and the effect is to slow down the increase of current. Then the magnetic lines expand less rapidly and induce less and less counter-emf to oppose the increase of current, but the current does actually continue to increase. The greater the current the more of the applied voltage must be used up in overcoming resistance. Finally we end up with current proportional to applied voltage and to circuit resistance.

In using the curves of Fig. 17-7 for inductive time constants we read the left-hand vertical scale as percentage of final current rather than as percentage of charge. The current in the circuit containing an inductor will rise to 63.2% of its final value in a time period of one time constant. The final 100% value of current depends only on applied voltage and resistance. If a circuit containing an inductor is short circuited while carrying some certain current, the current will not come to an instant stop but will decrease as shown by the right-hand curve. In a period of one time constant the current will have decreased to 36.8% of its original value. The circuit could be short circuited by disconnecting the source of external voltage and making a direct connection to replace the source.

To determine the value, in fractions of a second, of the inductive time constant we must know the inductance of the inductor or the circuit. Inductance is the ability of an inductor to produce a magnetic field or magnetic flux from current flowing in the inductor. We shall investigate this ability in the next few pages, but in connection with inductive time constants we need know only that inductance is measured in a unit called the henry. The inductive time constant, in seconds or fractions of a second, is equal to the ratio of inductance in henrys to resistance in ohms.

$$\text{Time constant, in seconds} = \frac{\text{inductance, in henrys}}{\text{resistance, in ohms}}$$

The resistance here considered is that of the circuit containing the inductor and its inductance. This resistance may be only that of the conductors in the coil and the circuit connections, or it may include also any type of resistor. We should note that the time constant depends not on any particular values of inductance and resistance, but on their ratio. Inductance of 10 henrys and resistance of 100 ohms would have a time constant of 1/10 second. Inductance of 1 henry and resistance of 10 ohms would have the same time constant of 1/10 second. If inductance and resistance increase and decrease together and in the same proportions, the time constant will not change.

SELF-INDUCTANCE. You will recall that self-induction is the action whereby a coil or a circuit induces in itself emf's which are due to changes of current in the coil. These emf's of self-induction result from

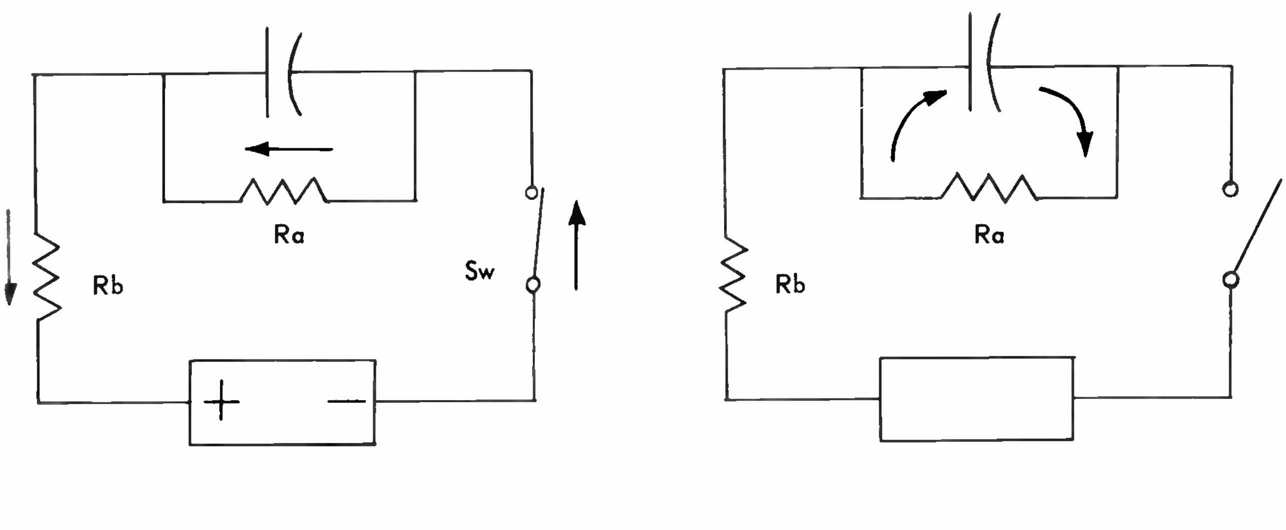


Fig. 17-9. Time constant for charge determined by one resistor, and for discharge by another resistor.

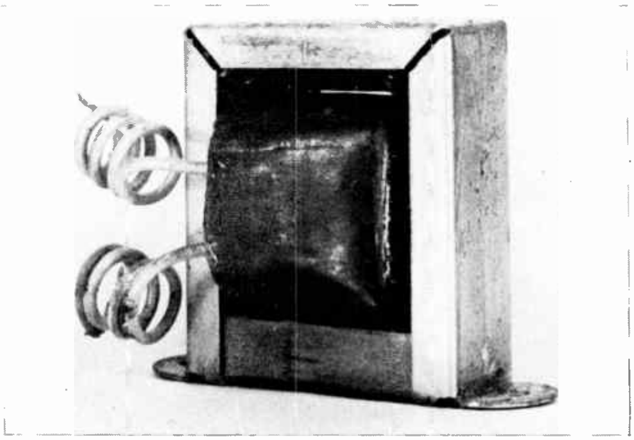


Fig. 17-10. An iron-core choke coil having large inductance.

magnetic lines expanding and contracting and cutting through the conductors. The expanding and contracting, or the motion of the magnetic lines, is due to increases and decreases of current.

Just what emf, in volts, is induced by any given change of current depends on the number of turns in the inductor or coil, on its size and shape, and on whether or not there is iron inside or around the coil. These are the things which affect the ability of the coil to generate an induced emf in itself. This ability is called inductance, and is measured in a unit called the henry, as mentioned before.

If a coil is of such design that current change at the rate of one ampere per second induces an emf of one volt, the inductance of the coil is one henry. A current change at the rate of one ampere per second would be an increase from zero to 1 ampere in 1 second, or from 1 to 2 amperes in one second, or from 100 to 101 amperes in one second. There would be the same rate of change were current to decrease in an amount of one ampere during a time of one second.

An inductance of one henry is quite a lot of inductance. Often we use smaller units. One is the millihenry, which is equal to one one-thousandth of a henry. Another is the microhenry, equal to one one-millionth of a henry. Fig. 17-10 is a picture of a type of inductor called a choke, because it chokes or tends to prevent rapid changes of current. In the coil itself, which you cannot see, are hundreds of turns of wire. Through the center and around the outside of the coil is a "core" of steel. The inductance is about 8 henrys. Any inductor whose inductance is great enough to be measured in henrys will have within and around its coil a core of steel or iron. It is so easy for magnetic lines or flux to form in steel or iron that a core of these materials permits hundreds of times as many lines as would be formed in air or any other substance by the same current. When this great number of magnetic lines cut through the turns of the coil they induce a strong counter-emf. Consequently, an inductor or coil with an iron or steel core possesses much inductance.

Fig. 17-11 is a picture of another kind of choke. Here there is no iron or steel core, but there are a great many turns of wire in a relatively small space. The inductance of this coil is about 3 millihenrys or 3/1000 henry. The iron-core choke of Fig. 17-10 has more than 2500 times as much inductance as this air-core choke. The total inductance of the coil pictured in Fig. 17-1 is about 85 microhenrys, and of the coil in Fig. 17-3 it is a little less than 2 microhenrys. The iron core choke has about 4 million times as much inductance as this latter coil. In all these cases we are talking about self-inductance, due to the action of self-induction in the same coil or circuit that possesses the self-inductance.

Fig. 17-11. An air-core choke coil used at radio frequencies.

There are three ways of determining the self-inductance of an inductor or coil. First, we may take the manufacturer's word for it, and be reasonably sure that the figure is correct provided the inductor is used according to recommendations. Second, we may measure the inductance with suitable laboratory apparatus. This is an excellent method if you have the equipment, know how to use it, and can afford to spend the necessary time. Third, you can make computations by means of arithmetic. Practically never will the result of any reasonably simple computation give the true value of inductance. But you can figure the inductance close enough to know whether or not the particular inductor is likely to work in the circuit where you intend to use it. Furthermore, you can figure the number of turns needed for a certain required inductance and come fairly close. Then, if you have to wind the coil, you can put on a few extra turns, place the coil in actual operation, and take off turns until things work to your satisfaction.

Here is a formula as good as any of the simple ones for approximate calculation of inductance of an air-core coil wound in a single layer as illustrated by Fig. 17-12.

$$\text{Inductance, microhenrys} = \frac{N^2 \times D}{40 \times (B/D + 0.45)}$$

The letters stand for the following values.

N, number of turns of wire on the coil

D, diameter of winding, in inches

B, length of winding, in inches

Note that the length of winding is from end to end of the conductor turns; it is not the length of the form on which the coil is wound. This formula is fairly accurate for coils whose minimum length is not less than one-third the diameter of the coil.

As an example, assume a coil of 50 turns on a 2 inch coil form, extending over a length of 1.5 inches. Its inductance is determined by using these values in the formula shown above.

$$\frac{50^2 \times 2}{40 \times (1.5/2 + 0.45)} = \frac{2500 \times 2}{40 \times (0.75 + 0.45)} = \frac{5000}{40 \times 1.20}$$

$$= \frac{5000}{48} = 104 \text{ microhenrys inductance, approximately.}$$

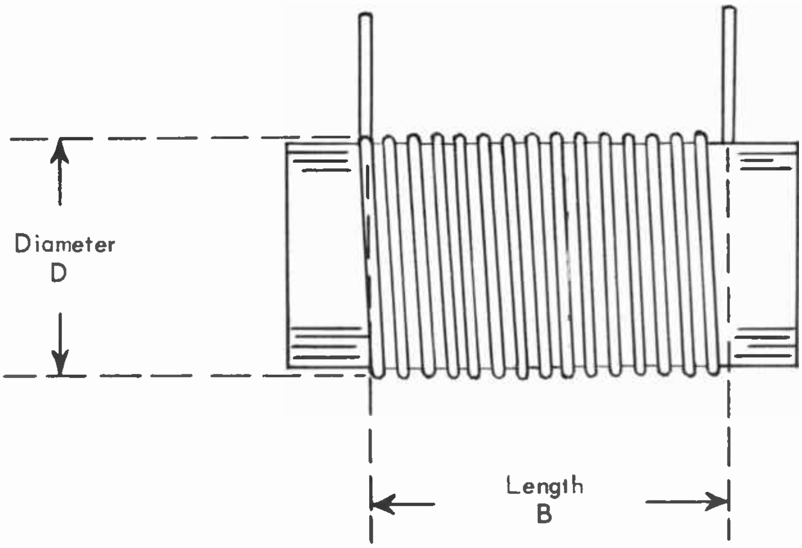


Fig. 17-12. Coil dimensions used when computing inductance.

The formula and method just explained allow computing the inductance of a coil already constructed. More often we wish to learn the number of turns required with a certain diameter and length of winding to produce some specified inductance. Here is the formula for number of turns.

$$\text{Number of turns} = \sqrt{\frac{40 \times L \times (B/D + 0.45)}{D}}$$

The meanings of the letters are the same as before. Here there is a new letter symbol, the capital letter L. The capital letter L is the symbol for inductance. In this formula the inductance is in microhenrys. The formula for number of turns is derived from the earlier formula for inductance, and computations with the two formulas should be in general agreement. To check this point, assume we wish to have inductance of 104 microhenrys on a 2-inch diameter with winding length of 1.5 inches. Putting the known values in the formula gives

$$\sqrt{\frac{40 \times 104 \times (1.5/2 + 0.45)}{2}} = \sqrt{\frac{4160 \times (0.75 + 0.45)}{2}}$$

$$= \sqrt{\frac{4160 \times 1.20}{2}} = \sqrt{\frac{4992}{2}} = \sqrt{2496} = 50 \text{ turns}$$

We should note the following points with reference to effects of coil design on inductance. With any given diameter and number of turns, increasing the length lessens the inductance. Increasing the diameter increases the inductance. Inductance also increases as the square of the number of turns. Thus, if a coil of 4 microhenrys has its number of turns doubled, the inductance will have increased to 4 x 4, or 16 microhenrys. Fig. 17-13 illustrates these things.

If iron or steel could be substituted for all the air which is in the field of the coil winding, the inductance would be multiplied by the permeability of the steel or iron. All field lines will not be confined to the steel or iron, so inductance will not increase directly as permeability but there always is a very great increase in comparison with a coil otherwise similar except for having an air core.

6 *INDUCTANCE REACTANCE.* While studying capacitors and their uses we learned that any opposition to flow of alternating current is called reactance, and that such opposition in a capacitor is called capacitive reactance. Opposition to flow of alternating current in an inductor or coil is called inductive reactance. The symbol for inductive reactance is X_L, wherein the X stands for reactance and the subscript L shows that the reactance is caused by inductance. Inductive reactance, like all other oppositions to current or electron flow, is measured in ohms.

We learned that capacitive reactance is altered by changes of capacitance and by changes of frequency. Inductive reactance is altered by changes of inductance and by changes of frequency.

Inductive reactance really is the result of counter-emf. You recall that counter-emf always is of such polarity as to oppose every change of current, it opposes every increase and it opposes every decrease of current. This is the same as saying that counter-emf opposes alternating current, because alternating current is continual increase and decrease of flow rate, and if all the increases and decreases of alternating current were flattened out, you no longer would have any alternating current at all. Consequently, if we check on the effects of changes of frequency and of inductance on the production of counter-emf we shall learn how frequency and inductance affect inductive reactance.

First we shall check on the effect of frequency by assuming the frequency to be doubled, with nothing else changed. There will be twice as many changes or alternations of current during every second of time. This means that the magnetic field will expand out of and contract back into the coil twice as many times during every second. In other words, the field lines will cut through the conductors of the coil twice as often as before, and, of course, will induce twice the original counter-emf.

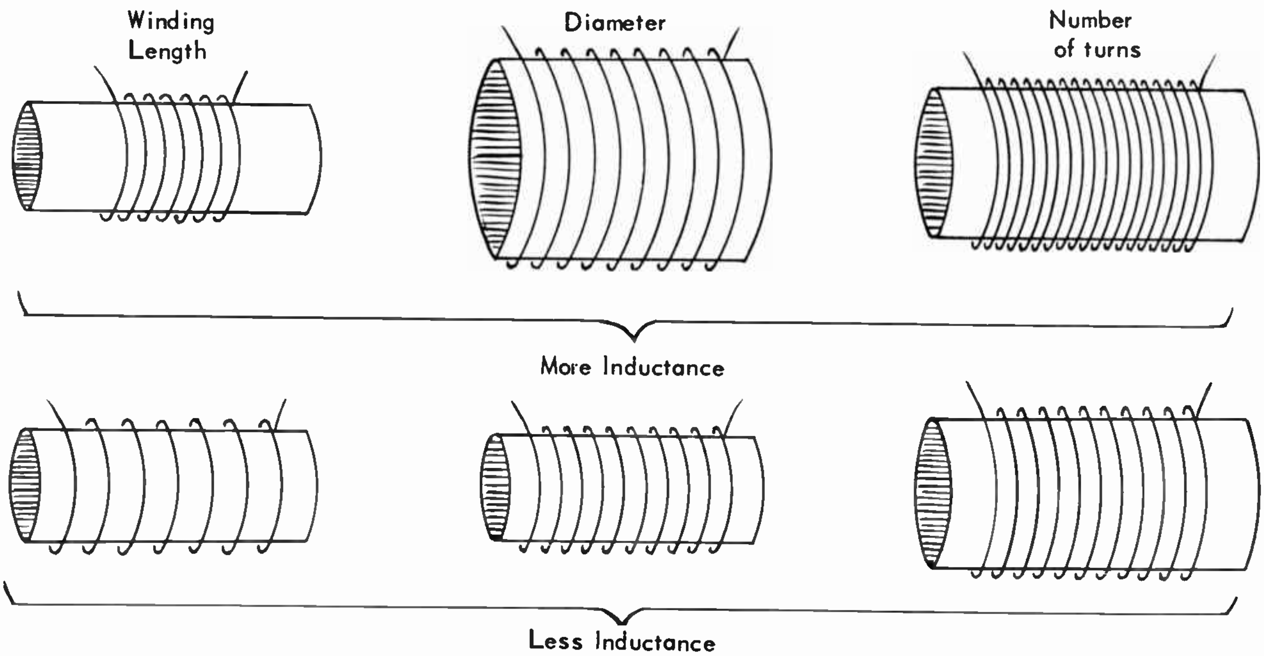


Fig. 17-13. The factors which affect the inductance of a coil.

From this we must conclude that the opposition to alternating current which is called inductive reactance is directly proportional to frequency. Twice the frequency will mean twice the reactance, half the frequency will mean half the reactance, and so on.

Now what about the effect of inductance on inductive reactance? This question is easy to answer, for we know that inductance means ability of the coil or circuit to produce magnetic lines or a magnetic field. Twice as much inductance means production of twice as many magnetic lines. Twice as many magnetic lines will cut through the coil conductors twice as many times during every second, and that means twice as much counter-emf. Since the opposition to alternating current which is called inductive reactance is the effect of counter-emf, we conclude that inductive reactance is directly proportional to inductance as well as being directly proportional to frequency.

The effects of inductors and capacitors or of inductance and capacitance are opposite. Some of the differences are shown in the accompanying tabulation.

When We Have The Things Listed Below	This Happens In Inductors	This Happens In Capacitors
Direct Current	Free Flow	No flow
Higher frequency	More reactance	Less reactance
Greater inductance	More reactance	
Greater capacitance		Less reactance

Because of these opposite effects, inductive reactance sometimes is called positive reactance while capacitive reactance is called negative reactance.

The fundamental formula for inductive reactance is as follows.

$$\text{Inductive reactance, in ohms} = 6.2832 \times \text{frequency, cycles} \times \text{inductance, henrys}$$

Other formulas, based on this one, allow using different units. The formulas given here employ cycles, kilocycles, or megacycles for frequency, in combination with henrys, millihenrys, or microhenrys for inductance.

$$X_L \text{ ohms} = 6.28 \times \text{cycles} \times \text{henrys} \qquad X_L \text{ ohms} = 6280 \times \text{kilocycles} \times \text{henrys}$$

$$X_L \text{ ohms} = \frac{\text{cycles} \times \text{millihenrys}}{160} \qquad X_L \text{ ohms} = 6.28 \times \text{kilocycles} \times \text{millihenrys}$$

$$X_L \text{ ohms} = \frac{\text{cycles} \times \text{microhenrys}}{160\,000} \qquad X_L \text{ ohms} = \frac{\text{kilocycles} \times \text{microhenrys}}{160}$$

$$X_L \text{ ohms} = 6\,280\,000 \times \text{megacycles} \times \text{henrys}$$

$$X_L \text{ ohms} = 6280 \times \text{megacycles} \times \text{millihenrys}$$

$$X_L \text{ ohms} = 6.28 \times \text{megacycles} \times \text{microhenrys}$$

IMPEDANCE. Supposing you have an inductive reactance (a coil) in series with a capacitive reactance (a capacitor) as at the left in Fig. 17-14. Assume that you know the inductive reactance at the frequency in

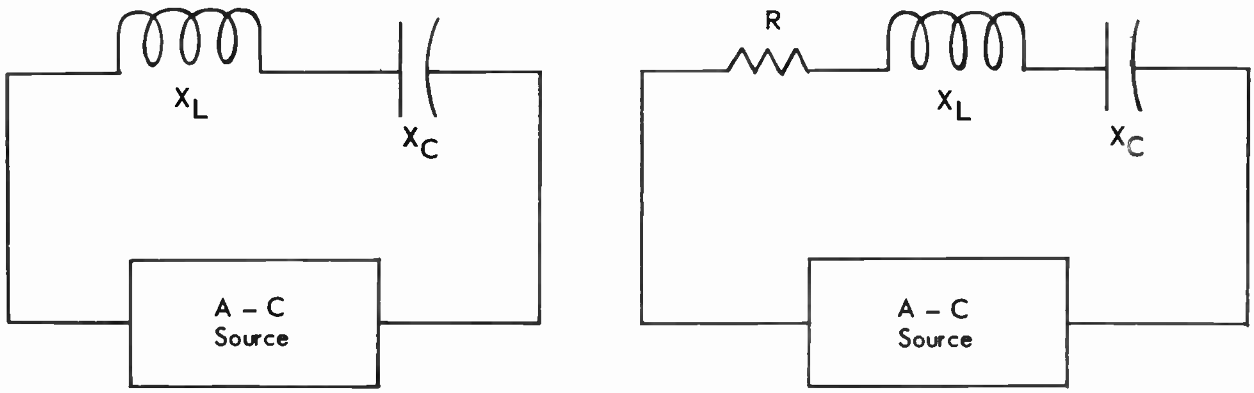


Fig. 17-14. Inductive and capacitive reactances oppose their effects when in a series circuit.

use is 1,000 ohms, and that the capacitive reactance at this same frequency is 600 ohms. What is the combined or total reactance or opposition to flow of alternating current from the source through these two parts? Quite likely your answer was wrong, because the combined reactance is 400 ohms – which is the difference between the two kinds of reactance in the series circuit. Furthermore, this combined reactance, which we more often call net reactance, is inductive reactance.

All this comes about because inductive and capacitive reactances act oppositely. In the example just considered, the 600 ohms of capacitive reactance cancels the effect of 600 ohms of inductive reactance, and leaves the remaining 400 ohms of inductive reactance to oppose current in this series circuit. To determine the net reactance in any circuit we subtract the smaller from the larger of the two different kinds of reactance. The difference is net reactance. If the larger reactance is inductive, then this net reactance is inductive. If the larger reactance is capacitive, the net reactance is capacitive.

(2) It would be impossible to construct a circuit with only inductance and capacitance, and no resistance – because all conductors have resistance. Therefore, in any circuit containing inductance and capacitance we must consider also the resistance if it is of a value comparable with either of the reactances. That is, we usually must consider a circuit as represented at the right in Fig. 17-14. We determine the net reactance as usual, by subtracting the smaller from the greater of the two reactances. Supposing that you have found the net reactance to be 400 ohms, and have measured the resistance of the circuit as 300 ohms. What is the total opposition due to this combination of reactance and resistance? Again you are quite likely to be wrong, for the total effective opposition is equal to the square root of the sum of the squares of resistance and net reactance. This opposition due to both resistance and reactance is called impedance. Like every other opposition, impedance is measured in ohms. The symbol for impedance is the capital letter Z. Here is the impedance formula.

$$Z\text{, in ohms} = \sqrt{R^2 + X^2}$$

In this formula R is resistance in ohms, X is net reactance in ohms. With 300 ohms resistance and 400 ohms net reactance put into the formula we would have,

$$Z = \sqrt{300^2 + 400^2} = \sqrt{90\,000 + 160\,000} = \sqrt{250\,000} = 500 \text{ ohms impedance.}$$

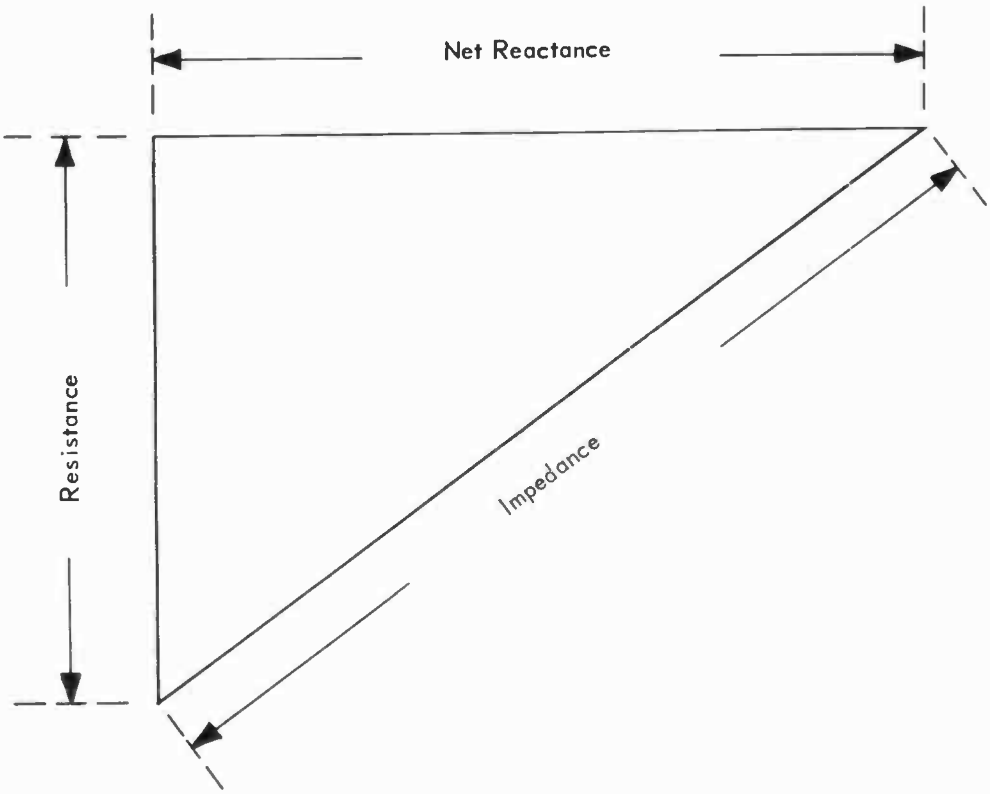


Fig. 17-15. The impedance triangle.

If you don't have to determine the exact number of ohms impedance you may arrive at a close approximation by drawing the impedance triangle of Fig. 17-15 – provided the two kinds of reactance and the resistance are in series. Make the length of one leg proportional to ohms of resistance. Make the other leg, at right angles, of length proportional to net reactance. Then the length of the hypotenuse, the diagonal, is proportional to ohms of impedance. Of course, all lengths must be measured in the same unit, inches, quarters, or anything convenient, but the same for all three sides. The proportions of the triangle of Fig. 17-15 are such as you would draw for 300 ohms resistance and 400 ohms net reactance, with 500 ohms measured on the impedance side.

Impedance always is greater than either the net reactance or the resistance, as is evident from the triangle. However, either the inductive reactance or the capacitive reactance may be greater than the impedance. Remember, in the examples just worked out, the inductive reactance is 1000 ohms, yet the impedance is only 500 ohms.

In one of the earliest lessons we worked with rules for computing either resistance, current, or voltage when the other two of these three quantities are known. Those rules, all of which include resistance, apply only to circuits in which inductive reactance and capacitive reactance are of negligible value in comparison with the resistance. But if you know the impedance of a circuit you can substitute impedance for resistance in all three of our rules, and solve problems relating to circuits containing inductance, capacitance, and resistance in any combination and with the circuit carrying alternating current. Here are the revised rules.

$$\text{Volts} = \frac{\text{milliamperes} \times \text{impedance, in ohms}}{1000} \qquad \text{Milliamperes} = \frac{1000 \times \text{volts}}{\text{impedance, in ohms}}$$

$$\text{Impedance, in ohms} = \frac{1000 \times \text{volts}}{\text{milliamperes}}$$

Milliamperes in these formulas are alternating current values. Volts are alternating, and are effective values, not peak values.

TELEVISION

DO NOT TEAR — CUT ALONG THIS LINE AND SEND IN FOR GRADING.

LESSON NO. 18 COUPLING ONE CIRCUIT TO ANOTHER

"Training Your Mind"

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind

C.T.I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

LESSON NO. 18

COUPLING ONE CIRCUIT TO ANOTHER

Whenever two circuits are in such positions or have such connections that changes of current or voltage in one circuit can cause changes of voltage in the other, the two circuits are coupled. Coupling is needed at many places in receivers. First of all, to get antenna signals or voltages into a receiver we need coupling between the antenna and the grid-cathode circuit of the first amplifier tube. Fig. 18-1 is a picture of an antenna coupler for a radio receiver. Tuners for television receivers contain antenna couplers.

Coupling is needed also between the r-f amplifier and the mixer, and between the r-f oscillator and the mixer, and from the mixer to the first amplifier in the i-f section. Then there must be more couplings between each amplifier and the one following, couplings into and out of detectors or demodulators, couplings to the loud speaker and to the grid-cathode circuit of the picture tube, and couplings all the way through sync and sweep sections as far as the deflection plates or coils at the picture tube. All these couplings are means for transferring signals from one circuit to another or from one tube to another. Fig. 18-2 illustrates some i-f coupling transformers.

There are almost as many varieties of couplings as there are places in which couplings are used. This makes it rather difficult to give any simple definition that describes every type of coupling. We may say, however, that if any resistance, reactance, or impedance is connected into two circuits, those circuits are coupled. This definition does not cover all couplings, but it does account for a large percentage of them. Now let's see how the definition applies to some practical circuits.

COUPLING WITH COMMON ELEMENTS. A common element is a resistance, reactance, or impedance that is common to two circuits or that is part of each of two circuits. Fig. 18-3 represents the simplest possible couplings between antennas and r-f amplifiers. An antenna for television or for frequency-modulation (FM) radio consists of two elevated horizontal conductors (usually pieces of metal tubing) from between which a two-conductor "transmission line" goes to the receiver. Radio waves in space induce signal voltages across the two halves of the antenna, and these voltages act through the transmission line.

In diagram 1 the resistance *R* is connected across the transmission line, therefore is in the antenna circuit. This resistance is connected also from grid to cathode of the amplifier tube, so is in the grid circuit. Thus we have the same resistance connected into two circuits, and the two circuits are coupled.

There is transfer of signal voltage in this manner. Whatever voltages are induced in the antenna by signal waves in space must appear across resistance *R*. Also, whatever voltages are across resistor *R* will act between grid and cathode of the tube. Thus the antenna voltages are coupled into the grid circuit of the amplifier. You ask, why use the resistor, why not connect the ends of the transmission line only to the grid and cathode of the tube? In this particular case such connection would leave us with nonconductive grid return to the tube cathode. The coupling resistor provides such a return.

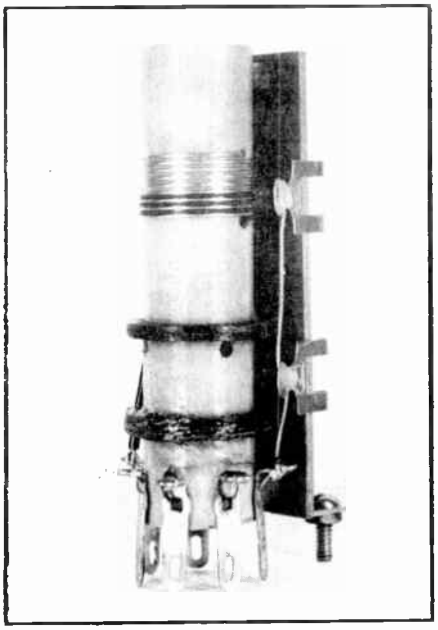


Fig. 18-1. An antenna coupler used in a standard broadcast radio receiver.

At 2 in Fig. 18-3 we have substituted a coupling coil for the coupling resistor. The coil is in the antenna circuit and also in the amplifier grid circuit. Antenna signal voltages appear across the inductive reactance of the coil, and thus are coupled into the grid circuit of the amplifier. The inductance of the coil may be of a value providing high reactance, in ohms, at the received frequencies. This will provide effective coupling. Many television and f-m receivers have inductance coils in their antenna couplings. Very few use resistance coupling for the antenna, which is shown here chiefly to illustrate the principle.

Instead of the inductive reactance for coupling we might use the capacitive reactance of a capacitor, as in diagram 3. In this antenna circuit the capacitor alone would provide no conductive grid return to the tube cathode. A resistor, in addition to the capacitor, would have to be connected from grid to cathode. Such a combination of coupling capacitance and resistance would work, but it is not a type in general use.

Next, let's see what we can do about transferring a signal voltage from one tube to another by means of a resistance common to the first plate circuit and to the second grid circuit. We may commence with diagram 1 of Fig. 18-4. Resistance *Ro* is in the plate circuit of the left-hand tube and in the grid-cathode circuit of the tube at the right. This necessitates connecting the plate of the first tube directly to the grid of the second tube, which places the plate and grid at the same potential. It is shown that the power supply furnishes 100 volts (B+) to the plate circuit. Allowing for potential difference across resistor *Ro*, the plate may be at something like 90 volts positive with reference to the cathode in the left-hand tube. Grid bias for the right-hand tube is secured by connecting the cathode of that tube at a point on the power supply which is 110 volts positive. With the grid 90 volts positive and the cathode 110 volts positive, the grid is 20 volts less positive than the cathode, or is effectively 20 volts more negative, and we have a 20-volt negative grid bias for the right-hand tube.

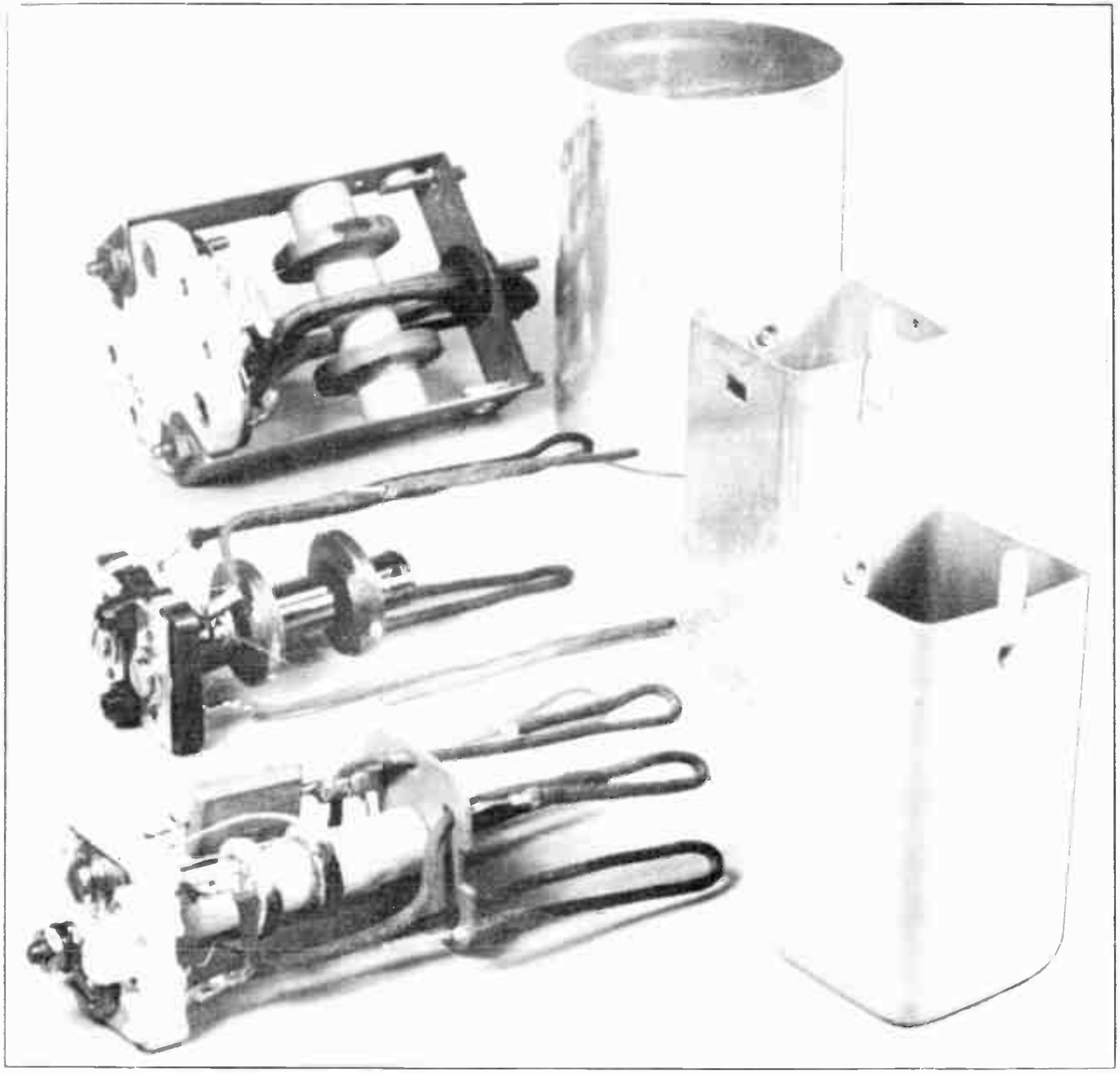


Fig. 18-2. Intermediate-frequency transformers removed from their metallic shields.

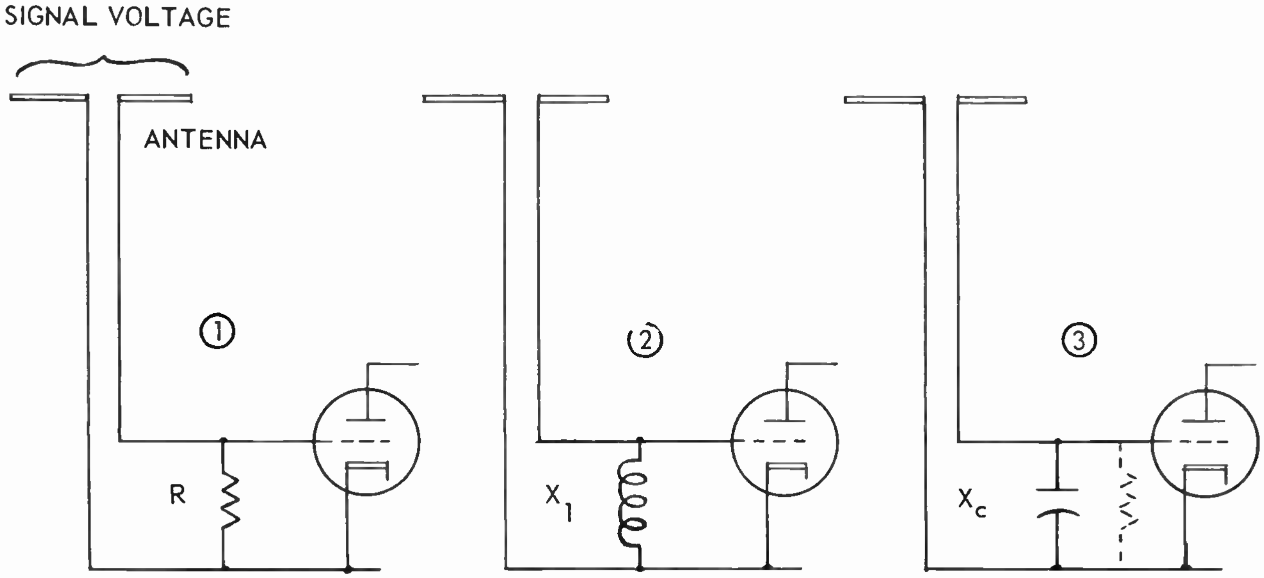


Fig. 18-3. Coupling elements common to the antenna circuit and the grid circuit.

Successful audio-frequency amplifiers have been built and still are built to use the coupling method just explained, but it is not a method in general use. Later you will find the same principle employed for coupling the last video amplifier to the grid circuit of television picture tubes.

In working toward a more popular method of coupling we shall go on to diagram 2 of Fig. 18-4. Here we have made it possible that grid bias for the right-hand tube may be entirely independent of plate voltage and B+ voltage on the left-hand tube. This is done by inserting capacitors *Cc* and *Cd* which block the high direct voltages of the plate supply circuit out of the grid circuit. Now we may use any desired voltages in plate and grid circuits, whereas with the method of diagram 1 these voltages would have to increase with every added tube or stage, and would become impracticably high after two or three stages of amplification.

In diagram 2 we still have the coupling resistor *Ro* in the plate circuit of the first tube and in the grid circuit of the second tube. The grid circuit now extends through capacitor *Cc*, then through resistor *Ro*, and capacitor *Cd* back to the cathode. Since this grid circuit need carry only the alternating voltages of the signal taken from across resistor *Ro* there is no objection to having the capacitors in series with the signal path. Between the grid and cathode of the right-hand tube must be some provision for grid bias voltage and for a conductive grid return path.

Diagram 3 of Fig. 18-4 shows the circuit of diagram 2 rearranged in a practical manner, or in the form which you will find used to show resistance-coupling in circuit diagrams for radio and television. Changes include ground connections at the tube cathodes and for capacitor *Cd*, also the usual grid resistor *Rg* leading to the bias source and to the grid return connection.

As shown by Fig. 18-5 we may substitute inductive reactances for one or both the resistances in the coupling system. At the left, inductor or coil *Lo* has been put in place of resistor *Ro*, and at the right there has been substitution also of inductor *Lg* in the place of grid resistor *Rg*. Coupling inductor *Lo* is in both the plate circuit of the first tube and the grid circuit of the second tube, just as coupling resistor *Ro* was in both circuits.

The method of coupling illustrated by Fig. 18-5 was first employed in audio-frequency amplifiers. Nowadays you will find this general principle employed chiefly for television intermediate-frequency amplifiers and in quite a few radio-frequency amplifiers of television receivers. The usual name for this system is impedance coupling, which refers to the impedance due to combined inductance and resistance in the coil. Another name which is technically correct is direct inductive coupling, but this use of the word inductive is likely to lead to confusion with another type of coupling soon to be examined.

Any attempt to substitute capacitive reactance for the coupling resistance, or coupling inductance and inductive reactance, results in the same difficulty encountered in Fig. 18-3 diagram 3. That is, the coupling capacitor forms an open circuit for both the plate return to its cathode and for the grid return to the other cathode. A capacitor in this position would have to have a resistor connected across it to provide the necessary conductive return paths, and such coupling would have most of the characteristics of resistance coupling. There would, of course, be reduction of capacitive reactance at higher frequencies and increase of this reactance at lower frequencies in the band being handled by the amplifier.

Coupling by means of a common capacitive reactance connected as shown by Fig. 18-6 is found in a few television amplifiers. Capacitor *Cb* and its capacitive reactance are common to the first plate circuit and the second grid circuit, and are considered to provide the coupling. Capacitors *Ca* and *Cc* block or isolate the high B+ voltage for the first plate and the biasing voltage for the second grid, and at the same time have

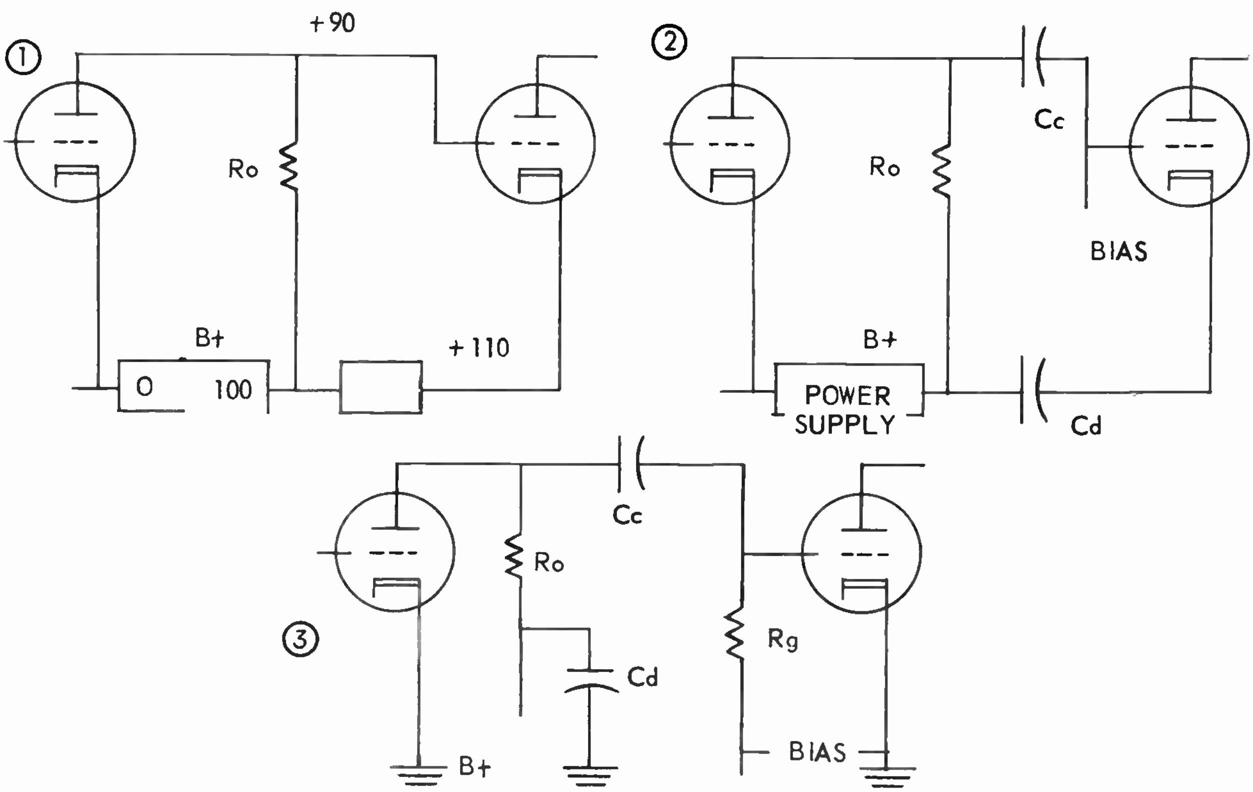


Fig. 18-4. Resistance coupling from the plate of one tube to the grid of a following tube.

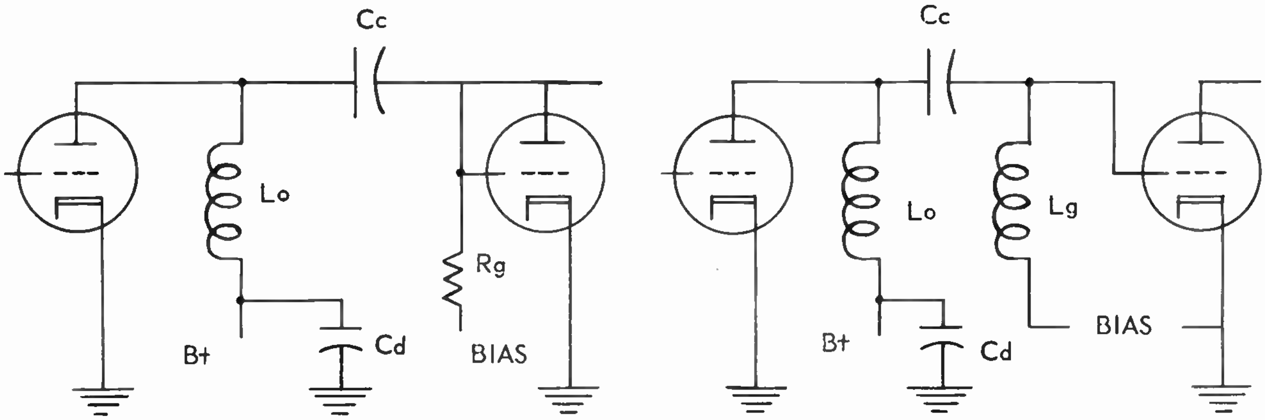


Fig. 18-5. Impedance couplings which employ a coil or inductor as the common element.

considerable effect on coupling. Increasing the capacitance at *Ca* and *Cc* allows greater transfer of signal energy, while increasing the capacitance at *Cb* lessens the signal transfer.

There are numerous other arrangements whereby a single resistance or reactance or impedance is included as a part of two circuits, and sometimes as a part of more than two circuits. All the circuits then are coupled, and energy will pass from one to another. We shall not go into the details of all these coupling modifications right now. It will be better to give them consideration as we come to the parts of receivers in which such coupling methods are used.

5 *INDUCTIVE COUPLING.* There is one highly important type of coupling in which the element common to the coupled circuits is not a resistance or any kind of impedance, but instead is a common magnetic field. If two conductors, two coils, or two circuits are close together, and if there is a changing current and the accompanying expanding and contracting magnetic field in one, the lines of that field are going to cut through the conductors of the other coil or circuit. This cutting will induce an emf in the other circuit, and if the other circuit is closed or complete there will be a current in it. This is the action called mutual induction.

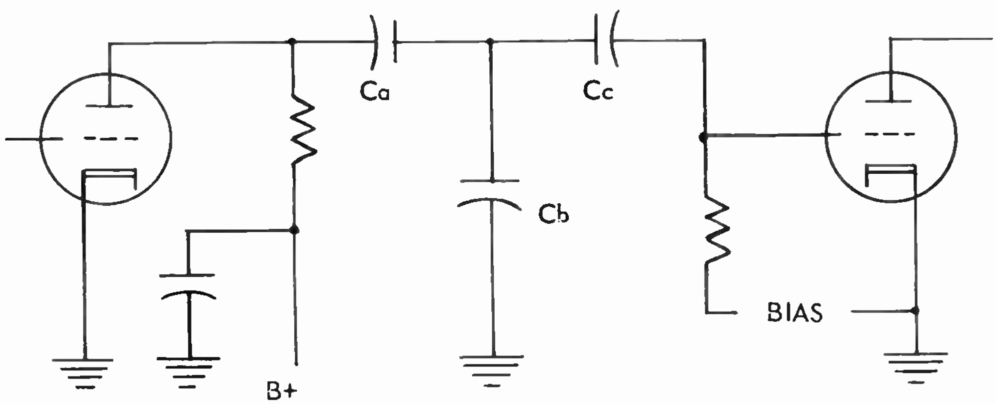


Fig. 18-6. One type of capacitance coupling from plate to grid.

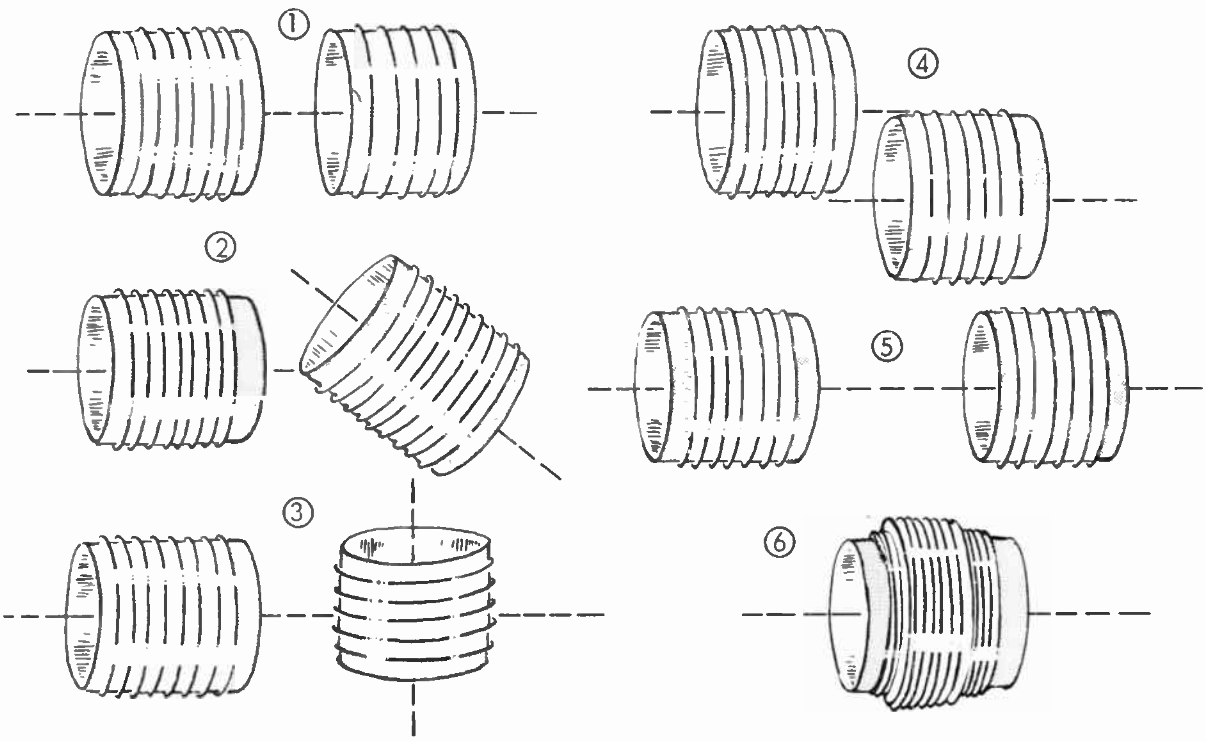


Fig. 18-7. Changes in relative positions of coupled coils varies the degree of coupling and the transfer of energy.

Energy is expended in the first circuit to provide its magnetic field. When an emf and current are induced in the other nearby circuit there will be energy in the moving electrons of that current. This energy must come from the first circuit, the one in which occured the original change of current and movement of magnetic lines. This transfer of energy from one circuit to another occurs through a form of coupling. The two circuits are coupled by mutual induction. Such an arrangement is spoken of as an inductive coupling.

Two coils, inductively coupled, are shown in Fig. 18-7. At 1 the axes of the two coils are in line with each other and the coils are quite close together. There would be considerable energy transfer, or, as we usually would say, there is a close coupling or a tight coupling. There is reduced energy transfer when the two axes are at an angle with each other as at 2, although each axis still passes through the other one. When there is reduced energy transfer we say that the coupling is loose or weak. If the two axes are at right angles and still intersecting, as at 3, there is still weaker coupling and less energy transfer. In fact, so long as the axes continue to intersect and there is no change in distance between coil centers, this position provides the least or loosest coupling.

If the coil axes are parallel but not in line, as at 4, the coupling is weaker and the energy transfer less than at 1, although there has been no change in the distance separating the coils. The coupling is made looser and energy transfer less by moving the coils apart with their axes still in line, as at 5. There would be the maximum coupling if both coils were of exactly the same shape, size, and construction, and if both could occupy the same space. Since this is impossible, the closest coupling of two coils is shown at 6, where one coil is wound directly over the other one with the axes and centers in line.

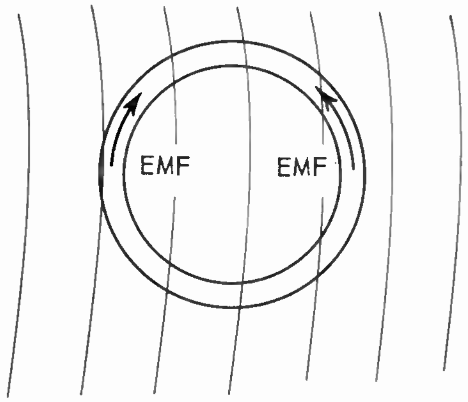


Fig. 18-8. Opposing emf's are induced in opposite sides of a coil turn.

Inductive coupling is due wholly to cutting of conductors in one coil by magnetic lines produced in another coil. Whatever increases the number of cuttings increases the coupling and the energy transfer, and whatever decreases the cuttings lessens the coupling and energy transfer.Three changes illustrated in Fig. 18-7 may be explained on this basis. At 4 and 5 the second coil is moved into a position where the magnetic field of the first coil is less concentrated or has fewer magnetic lines in a given area. At 6 the second coil is in a position where every magnetic line moving out of and back into the first coil has to cut through the second one, while at 1 a great many of the lines curve around and re-enter the first coil before cutting the turns of the second coil.

At 3 in Fig. 18-7 there is no decrease in the number of cuttings as magnetic lines from either coil pass through the turns of the other coil, yet there is a great decrease of coupling. For the explanation we must remember that energy transfer results from emf's and currents induced in one coil by movement of the magnetic field from the other coil. Then we may look at Fig. 18-8 to see what happens when the axis of one coil is at right angles to the axis of the other coil. If we consider the magnetic lines as moving from left to right there will be emf's induced in the two sides of each coil turn as shown by arrows. Since the magnetic lines move the same direction through both sides of the coil, the induced emf's must be of the same polar-

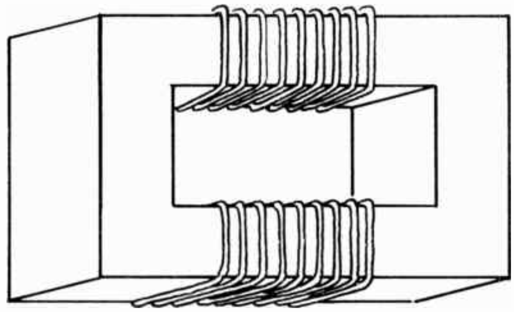


Fig. 18-9. Two windings coupled by means of common iron or steel core.

ities. The diagram shows both emf's acting upward. But in the coil turn these two emf's are acting oppositely, one is trying to produce electron flow in a clockwise direction and the other in a counter-clockwise direction around the coil turn. The net effect of the opposed emf's is practically no current, no energy transfer, and no coupling.

The coupling in diagram 2 of Fig. 18-7 is reduced for the reason just explained with reference to diagram 3, except that the reduction is not so great because the coil axes are not at right angles to each other.

There is something else we may do to increase the number of cuttings; we may increase the inductance of the coil in which the magnetic field is produced. You will recall that inductance is a measure of the ability to produce magnetic lines or magnetic flux. More inductance then must mean more lines, and more lines moving together must mean more cuttings, more induced emf, more energy transfer, and more coupling. We may go back to all the factors that increase the inductance, and consider each of them as a means for increasing the coupling.

Fig. 18-9 illustrates still another way of increasing the coupling between two coils. Here the two coils are wound on a single core of iron or steel. The permeability of iron or steel is so very much greater than of any other substances that the presence of the core within each coil greatly increases the inductance and the number of lines produced by any given current. But fully as important, maybe more important, is the fact that the continuous path of iron or steel insures that almost every magnetic line produced by one coil will go through the other coil. It is so much easier for the magnetic force to act through the iron than through any other substance around the coils that hardly any of the force or the lines will escape from this magnetic circuit.

COEFFICIENT OF COUPLING. We have been talking about greater or less degrees of energy transfer with couplings which are tight or loose, but nothing has been mentioned about any unit in which to measure the couplings. There is such a measure; it is called the coefficient of coupling. If there is maximum possible coupling the coefficient of coupling would be 1.0. All lesser couplings have coefficients less than 1.0, and no coupling at all would have a coefficient of 0.0 or zero. In service operations, and even in experimental design work, it is practically never necessary to either compute or measure coefficients of coupling, or to have anything else to do with them in a technical way. When you need more energy transfer you increase the coupling until satisfactory results are obtained, and if you want less energy transfer you lessen the coupling – without worrying about the coefficient as a fraction in either case.

In order to compute the coefficient of coupling for circuits having a common resistance or impedance it would be necessary to know the value of that common element and also the values of resistances or impedances in each of the coupled circuits. Then we multiply these latter resistances or impedances together and extract the square root of the produce. This square root is divided into the resistance or impedance which is common to both circuits, and the quotient is the coefficient of coupling as a fraction.

Coefficients of coupling sometimes are expressed as per cents. Multiplying the fraction by 100 changes it to the equivalent per cent. As an example, measurements of antenna couplers for radio receivers would usually show coefficients of coupling between 0.02 and 0.10, which are equivalent to couplings between 2% and 10%.

MUTUAL INDUCTANCE. It was just explained, in connection with coefficient of coupling, that the degree of coupling depends on the resistance or impedance of an element that is common to both the coupled circuits. There is no common resistance, inductance, or capacitance with the inductive couplings shown by Figs. 18-7 and 18-9. The only thing common to both circuits is the magnetic field, and there is no such thing as resistance or impedance of a magnetic field. There is, however, one other common thing which does not show up in pictures or diagrams; it is the mutual inductance of the two couples circuits.

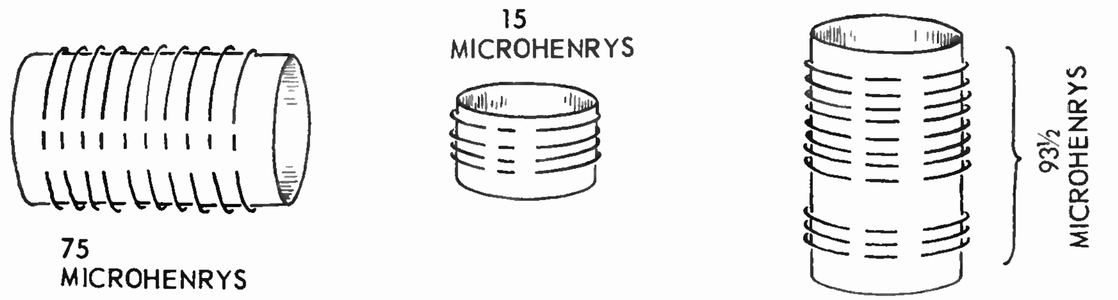


Fig. 18-10. Self-inductances of coils have added to them the mutual inductance when the coils are coupled.

Mutual inductance is inductance added to the sum of the inductances of two coupled circuits, and added because of the fact that the circuits are coupled inductively. At the left in Fig. 18-10 is represented a coil whose self-inductance is 75 micro-henrys, and next to it is another coil whose self-inductance is 15 microhenrys. The sum of the self-inductances is 90 microhenrys. If the two coils are inductively coupled, as at the right, with a coupling coefficient of about 10 per cent, the total inductance will be about 93-1/3 microhenrys. The added 3-1/3 microhenrys is due to mutual inductance. It is due to the induction (production of emf's) that results from the magnetic field of one coil cutting the turns of the other coil, whereas the self-induction in each coil is due to the magnetic field cutting the turns of the coil in which the field is being produced.

Mutual inductance is measured in henrys, millihenrys, and microhenrys, just as is self-inductance. The symbol for mutual inductance is the capital letter M.

If you consider the fact that a magnetic field may spread to a great distance from the circuit in which the field is produced it is easy to realize that circuits and coils which are well separated in space still may have inductive coupling. This coupling at a distance may be undesired, and it may cause many difficulties because signals get into circuits where they don't belong.

There may be similar unwanted couplings by means of resistances or impedances common to two or more circuits. For example, resistances in a power supply may be in the plate, screen, or grid circuits of two or more tubes, and stages containing those tubes may be coupled when their operations should be entirely independent.In this lesson we are talking about methods of obtaining couplings. Sometime later we shall have to learn how to prevent couplings.

COUPLING WITH TRANSFORMERS. Fig. 18-9 shows one type of the device which we call a transformer. The coupled coils at the right in Fig. 18-10 is a transformer. The antenna coupler of Fig. 18-1 is a transformer, and i-f transformers are pictured by Fig. 18-2. Radio and television receivers have lots of transformers. We may define a transformer as consisting of two or more windings or coils having a common magnetic circuit or magnetic field, whereby a varying current in one winding induces a varying emf in the other winding. A transformer is a coupling device in which energy transfer occurs because of mutual induction.

Fig. 18-11 shows, by means of symbols, how transformers may be used to couple an antenna to the grid circuit of the first amplifier tube and to couple the plate circuit of the first tube to the grid circuit of a second amplifier. The transformer coupling or inductive coupling between the two tubes would serve the same purpose as the resistance coupling of diagram 3 in Fig. 18-4.

Transformer symbols such as found in circuit diagrams are shown by Fig. 18-12. The symbol at 1 represents any air-core transformer in which the two windings are in fixed positions, not adjustable. If the coupling of the air-core transformer may be varied or adjusted, this feature may be shown by the symbol at 2. An iron or steel core of the general style represented in Fig. 18-9 is shown by the symbol at 3. The iron core thus indicated would be made up from thin sheets held tightly together, and would be suitable for use at audio frequencies and at power line frequency, but not at intermediate and radio frequencies. For these higher frequencies the transformer may have a core of finely divided or powdered iron cemented into suitable shape. A powdered iron core may be indicated by the transformer symbol at 4.

If an entire powdered iron core or part of such a core is movable as a means for varying the inductance or coupling, this fact may be shown in a symbol as at 5 or 6 in Fig. 18-12. Sometimes there are connections, called taps, between the outer ends of a winding or coil. These are indicated as in the symbol at 7. Various features of all the symbols may be combined. For example, at 8 is represented an air core transformer with both windings tapped. Most transformer symbols show the two windings with the same number of loops, and side by side, but occasionally you will find other arrangements, such as the one at 9. Any transformer symbol is easily recognized, for whenever two or more coils are drawn in such positions as to indicate a close relation between them, the complete symbol practically always means some type of transformer.

7 The transformer winding in which changes of current produce a varying magnetic field is called the primary winding of the transformer. The winding in which an emf is induced by the varying magnetic field is called the secondary. Power or energy enters the transformer through its primary, and leaves by way of the secondary. The primary is connected to a source of power or energy, the secondary is connected to a load

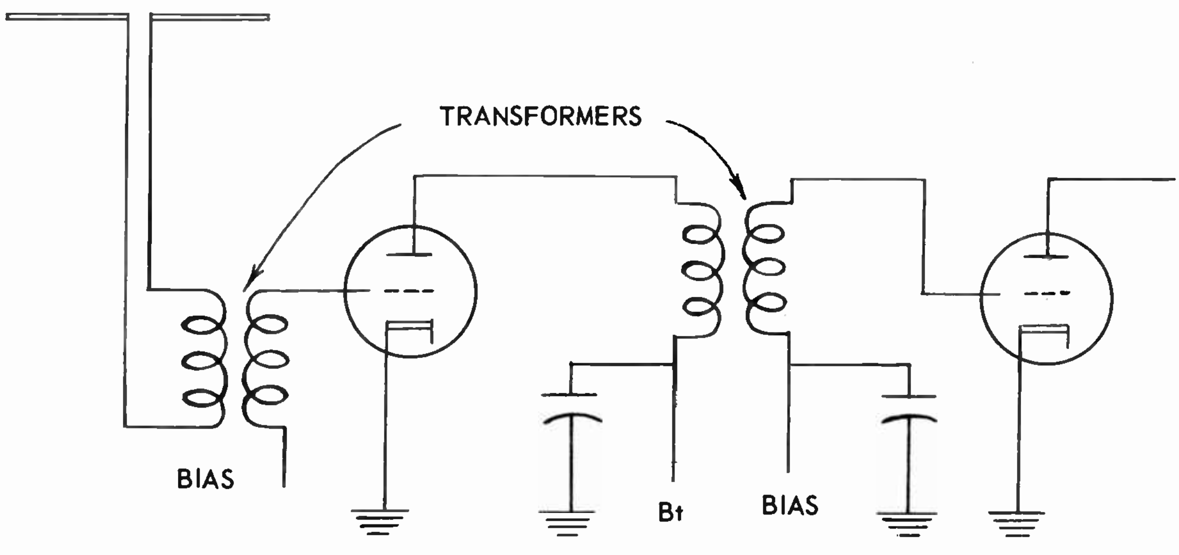


Fig. 18-11. Antenna coupling and interstage coupling by means of transformers.

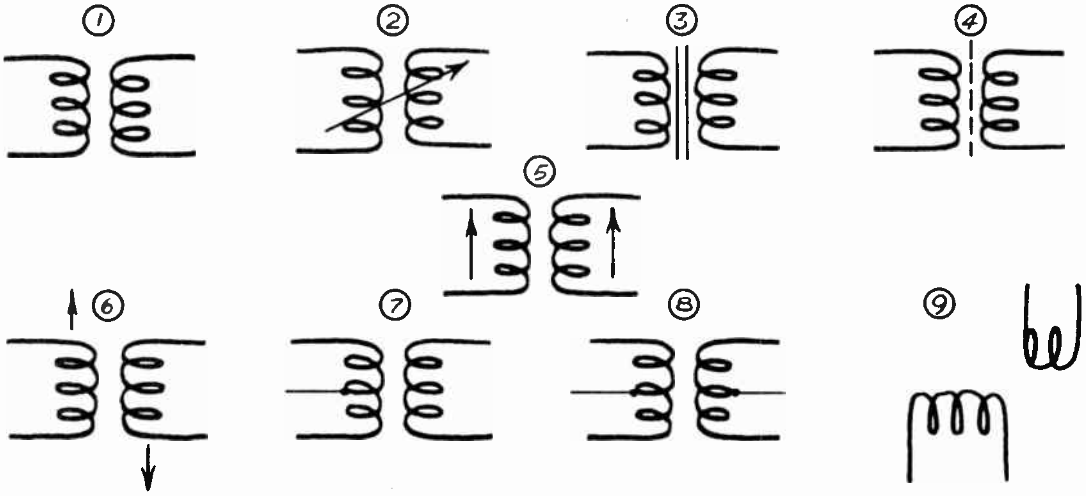


Fig. 18-12. Symbols used to indicate transformers in circuit diagrams.

in which the energy is used or changed to another form. Symbols do not indicate which winding is the primary and which is the secondary. These things are evident from the manner in which the transformer is connected to other circuits.

LINK COUPLING. Two coils or two circuits which are too far apart or otherwise positioned to prevent a single magnetic field from extending through both may be inductively coupled with a link connection as shown by Fig. 18-13. This method is suitable where the coils to be coupled are of the air-core type or any other type working at radio frequencies. Link couplings are used for some measurement operations in service work.

The two link coils should be alike in diameter, length, and number of turns. They need have only a few turns, maybe four or five for use at very high frequencies and up to something like twenty for standard broadcast frequencies. The two-conductor line between the link coils may be of any insulated wire. Its length may be as much as two or three feet without much loss of coupling ability.

DIRECT CURRENT BLOCKING IN COUPLINGS. Always it is necessary to have direct current in the plate circuit of all tubes working as amplifiers, oscillators, detectors, and for other purposes. This direct current is the electron flow which is made to increase and decrease by the signals. The variations of direct current represent the signal, and by themselves would be an alternating current. The plate current of a tube which is carrying a signal is really an alternating current combined with a direct current. We may call this combination current "a direct current with an alternating component" or we may say that the alternating current is superimposed on the direct current.

We may represent the direct plate current, with no signal, as at the left in Fig. 18-14, and may show the alternating signal current by itself as at the center. Then the plate current would be represented at the right, This plate current still is direct current, for it always is in the same direction or always is in the direction which we are calling positive. But the direct plate current becomes alternately more positive and less positive. Its peaks have a voltage equal to the sum of the original direct voltage plus the alternating peak voltage. Remember, alternating peak voltage is measured from zero to either the positive or negative peak of

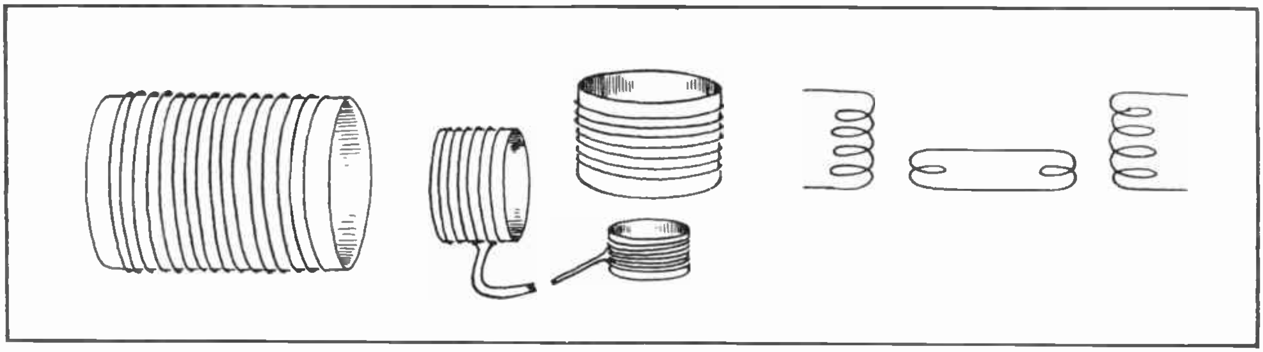


Fig. 18-13. A link coupling.

the alternating wave. The minimum voltage at the dips in the combination direct current is the difference between the original direct voltage and the peak alternating voltage of the added signal.

Only the alternating component of the plate current is passed on to the grid of the following tube with types of coupling in general use. The direct current portion must be left behind. Incidentally, when we speak of the alternating component in the plate current we may speak also of the direct component.The direct component is the original direct current with no signal, and it is also the direct current that will remain after the alternating component is removed and sent on to the following grid circuit.

The alternating component of plate current is separated from the direct component as illustrated in Fig. 18-15. At the left we have a resistance coupled arrangement like the one at 3 in Fig. 18-4. At the right we have transformer coupling, as in Fig. 18-11. The paths followed by direct plate current or the direct component of plate current are shown by heavy lines. With resistance coupling this d-c path extends through the tube, the power supply from ground to B+, and through coupling resistor *Ro*. With transformer coupling the d-c path extends through the tube and power supply, and through the primary winding of the coupling transformer.

3 Direct current and its accompanying highly positive voltage are blocked from the following grid circuit in the resistance coupled stage by capacitors *Cc* and *Cd*. Direct current cannot pass through the insulating dielectric of a capacitor. With transformer coupling it is impossible for the direct current and voltage to get

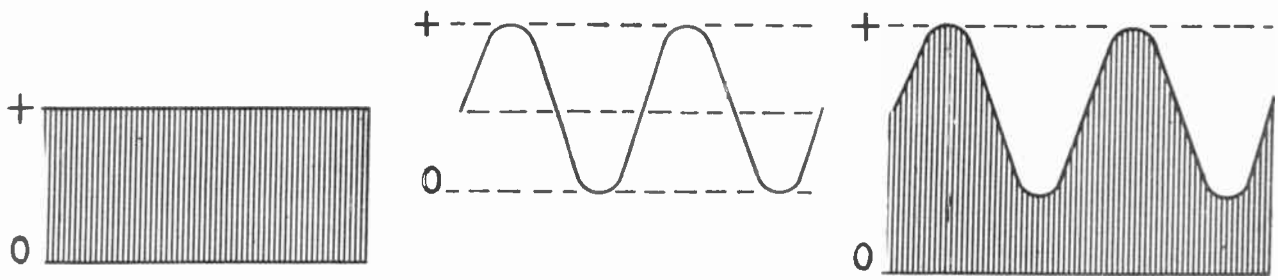


Fig. 18-14. Direct current (left) with addition of alternating current (center) becomes direct current with an alternating component (right).

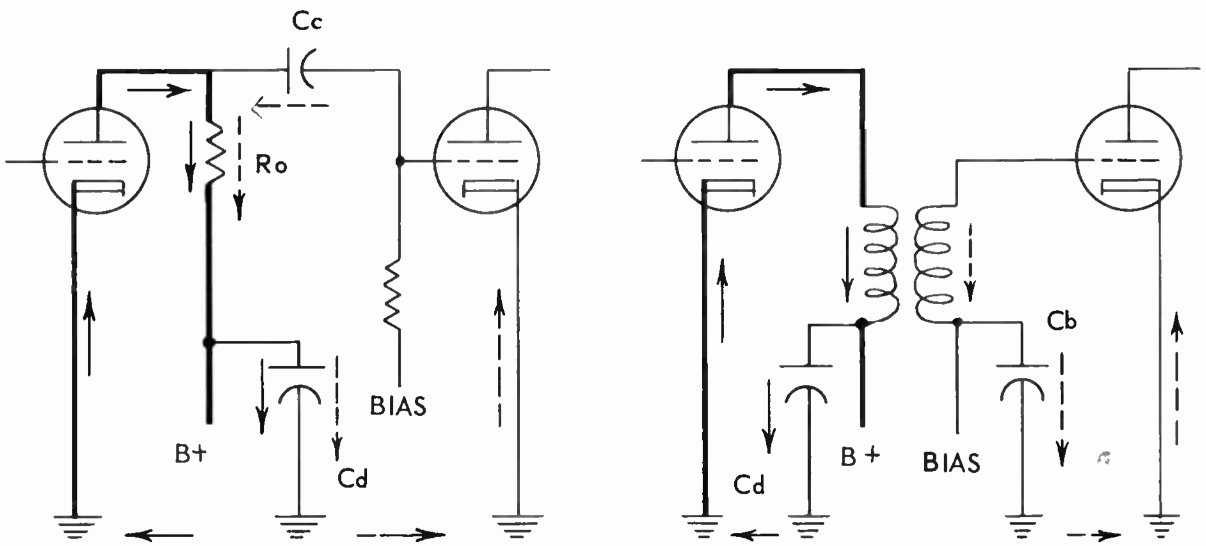


Fig. 18-15. How alternating signal currents are separated from direct plate currents with resistance coupling and transformer coupling.

through the transformer, because the only connection between primary and secondary windings is the magnetic field. The two windings are completely insulated from each other.

The path of alternating signal currents is shown by full-line arrows in both diagrams of Fig. 18-15. This path is the same as that for direct current, with one exception. Bypass capacitors *Cd* provide a path of low capacitive reactance through which the alternating signal voltage and current may get from the B+ end of the plate circuit to ground and back to the tube cathode without having to go through the power supply. By using enough capacitance at *Cd* the reactance of these capacitors may be made so small as to offer much less opposition to signal currents than the path through all the parts of the power supply.

The alternating signal voltage in resistor *Ro* of the resistance coupled system is also in the following grid circuit, it is the source of signal voltage for the grid circuit. The path of signal voltage in the grid circuit is shown by broken-line arrows. The alternating signal voltage acts easily through capacitor *Cc* on the high side and through capacitor *Cd* on the low side or ground-cathode side of the grid circuit.

With transformer coupling the alternating component of the plate current produces around the primary winding a magnetic field which expands and contracts in unison with signal changes. The moving field cuts through the turns of the secondary winding and in this winding induces emf's which correspond to the signal. These signal emf's cause alternating signal current to follow the grid circuit path shown by broken-line arrows. Here we have a bypass capacitor *Cb* which lets the signal voltages act through ground to the tube cathode without having to go through whatever circuits are furnishing the grid bias voltage.

2 Other methods of separating alternating signal voltages from direct currents in plate circuits provide advantages desired in some applications. For example, with iron-core coupling transformers it is desirable to keep direct current out of the primary winding, because the greater the current in the transformer the less becomes the permeability of the iron, and the less the inductance of the windings. Also, when very high voltages are used in the plate circuit and in the transformer primary, the entire transformer must be insulated to withstand these voltages. With only the voltages used in grid circuits much less insulation would be needed.

Requirements such as just outlined are met by using a style of connection called shunt feed or parallel feed, as illustrated by Fig. 18-16. Direct plate current flows only in the parts shown by heavy lines. In series with plate resistor *R* is a choke coil or inductor whose inductive reactance and opposition to flow of alternating current are high at the signal frequencies being handled. This reactance and the resistance at *R* force alternating signal currents in the plate circuit to take the easier path shown by full-line arrows, through capacitor *Cc*, the transformer primary, and ground connections back to the tube cathode. Signal voltages induced in the transformer secondary act in the following grid circuit as shown by broken line arrows. There is no direct current and no plate voltage in either side of the transformer. As a means for distinguishing the two methods of "feeding" direct current to the plate and taking it from the plate, the one shown at the right in Fig. 18-15 may be called a series feed.

In Fig. 18-16 the grid bias voltage for the second tube is secured from a resistor in series with the cathode. This cathode bias might be used with any of the other circuits shown earlier without altering the principles of coupling. All the tubes shown in diagrams of interstage couplings have been triodes. Any or all of them might be changed to pentodes or to beam power tubes without altering the coupling methods.

ALTERNATING VOLTAGES AND CURRENTS. We seem to be working more and more with alternating voltages and currents, largely because all signals for radio and television are essentially of these torms. Soon we shall investigate the performance of power supply systems, and there be dealing with alternating voltages and currents which come from the supply lines. In view of all this it may be well to briefly review some of the fundamental facts relating to such voltages and currents.

Many times you will come across the term "sine wave," which refers to the manner in which an ideal alternating current or voltage varies with reference to time. The meaning of a sine wave is illustrated by Fig. 18-17. On the left is a circle around which a point is rotating at uniform and constant speed. If you could stand off to one side and look toward the edge of the circle, that point would appear to move up and down, repeatedly. Were this up and down motion represented as a wavy line, progressing the same horizontal distance during every second of time, the line would appear as at the right. It would be a sine wave. This is the manner of variation of alternating voltage or current assumed in all simple computations involving these quantities. Unless specifically stated otherwise, we always are assuming a sine wave voltage or current.

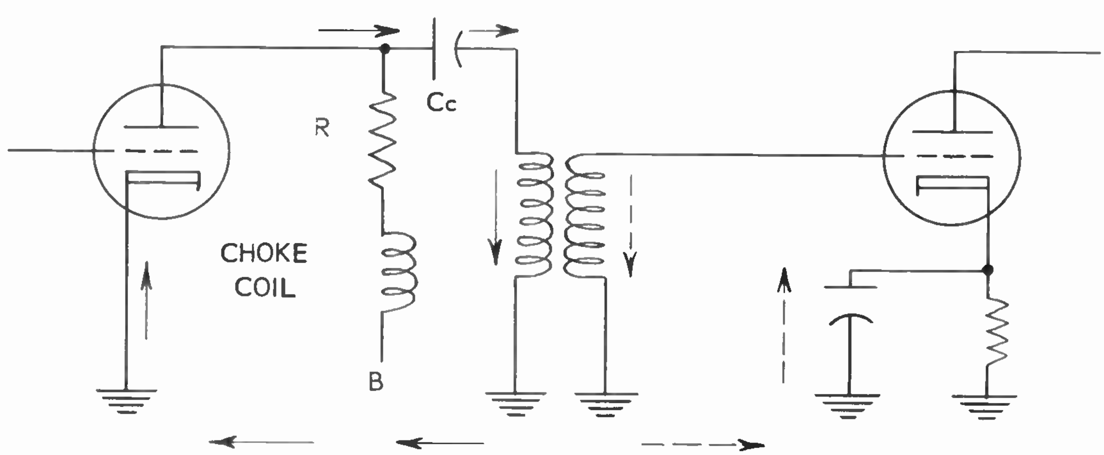


Fig. 18-16. Shunt feed or parallel feed as used with transformer coupling.

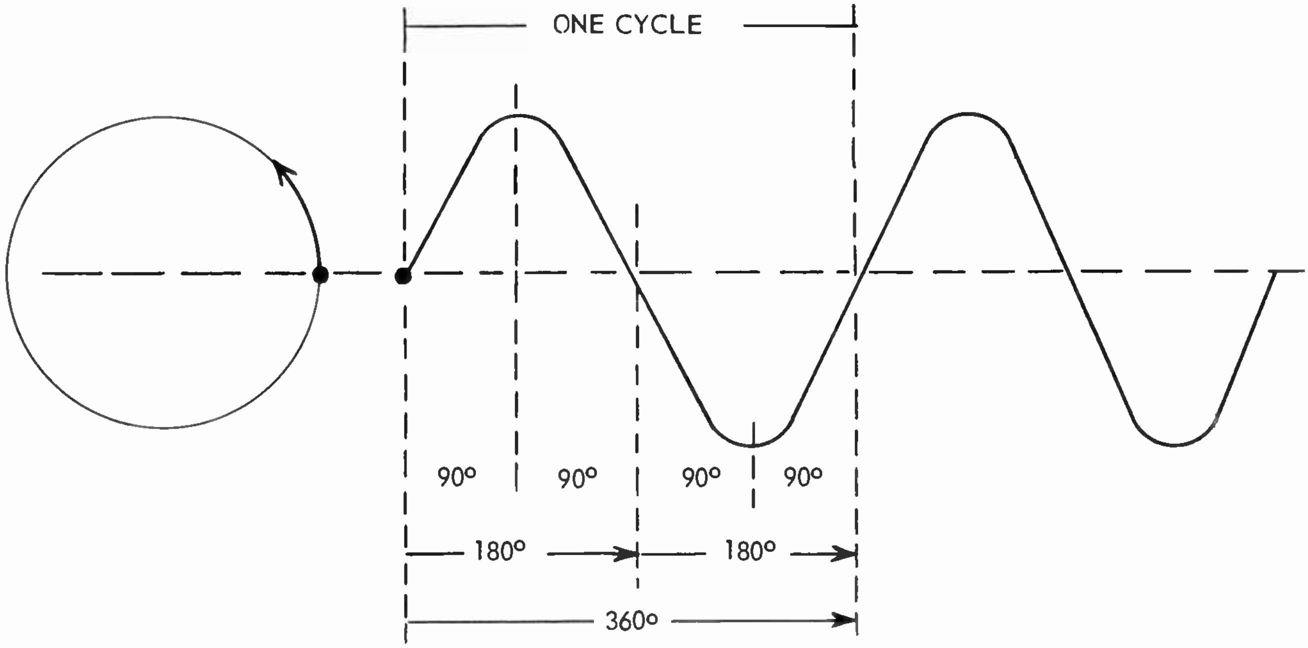


Fig. 18-17. The development of a sine wave, the ideal waveform for alternating voltage and current.

One complete series of changes of voltage or current is called one cycle. For some purposes it will be convenient to think of a cycle as consisting of 360 electrical degrees, to correspond with the 360 angular degrees through which the point moves in going around one complete circle. A half-cycle then is 180 electrical degrees, a quarter-cycle is 90 electrical degrees, and so on. A half-cycle may be called an alternation.

Fig. 18-8 shows several of the values which apply to sine wave voltages or currents. The maximum value with reference to zero, in either direction, is the peak voltage or current. As mentioned in an early lesson, the effective value of an alternating voltage or current is the number of volts or the number of amperes which, in a direct current, would cause the same heating effect as would be produced by this alternating current. When we speak of an effective alternating voltage or current we simply are saying that this alternating voltage or current would do the same heating as a direct current of the specified number of volts or amperes.

The effective value of a sine wave voltage or current is 0.707 times the peak value. The peak value in a sine wave is 1.414 times the effective value. These ratios hold only for sine wave voltages and currents, not exactly for any other waveforms. The average value during any number of complete cycles is zero, because there is the same voltage and current both ways from zero.

The amplitude of an alternating voltage or current is its greatest departure from zero in either direction. When the two peaks are equal, the amplitude is the same as one of the peaks (as the peak value) but only then. Peak-to-peak voltage, which we often hear of in television, is the sum of the two opposite peak voltages. It is not the same as amplitude, although in a sine wave it would be twice the amplitude. As we continue with our work you will use these values so much as to become well acquainted with all of them, but it is well to have the whole collection together in this one place as a sort of reference.

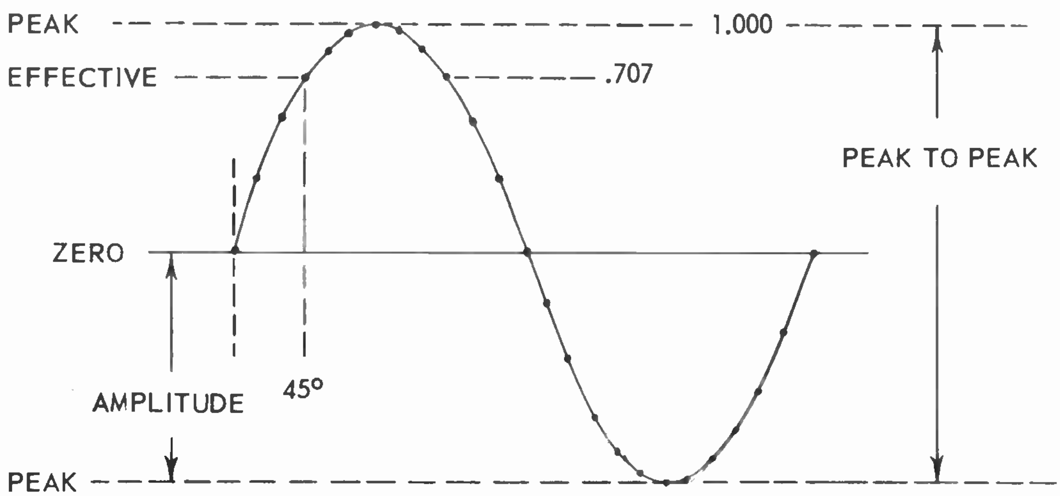


Fig. 18-18. Values in a sine wave alternating current or voltages.

Do Not Tear — Cut Along This Line and Send In For Grading.

TELEVISION

LESSON NO. 19 RECTIFIERS FOR POWER SUPPLIES

"Training Your Mind"

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind

C.T.I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

LESSON NO. 19

RECTIFIERS FOR POWER SUPPLIES

All the tubes in every receiver require direct current and voltage for their plates and also for screens where screens are used. Many tubes require also a direct voltage for grid bias. All picture tubes require high values of direct voltage for acceleration of beam electrons, while electrostatic tubes require other high voltages for their deflection plates. Finally, all tubes must be supplied with voltages and currents suitable for heating their cathodes.

The high voltages and the low ones, the large currents and the small ones, all must come from the power supply sections of the receiver. In any ordinary radio or television receiver there are more wires and connections in the power supply circuits than in all others put together. Except in the relatively few portable radios which are battery powered, and the still fewer sets made to operate on direct-current lines, all power comes originally from the same alternating-current lines that furnish power for lamps, appliances, and electrical machinery in general.

In every television receiver there are at least two principal divisions of the power supply system. In Fig. 19-1 these are shown connected to a picture tube and to a single pentode tube whose plate, screen, and grid circuits may represent such circuits for all the other tubes in a receiver. The section called the high-voltage power supply furnishes direct potential differences ranging from 5,000 to 15,000 volts or more for beam acceleration in all types of picture tubes and for deflection plates of electrostatic tubes. This high-voltage power supply obtains its energy from the low-voltage power supply.

The low-voltage power supply delivers direct potential differences up to a few hundred volts for the high-voltage supply section, also for plate, screen, and grid bias circuits of all tubes other than the picture tube, and usually to any elements of the picture tube which operate at voltages within this range. In addition, the low-voltage supply furnishes alternating voltages and currents for the heater-cathodes and for any filament-cathodes which may be used in the tubes.

The power supply sections of a receiver must do these four things:

1. Produce pulsating direct voltages and currents from the alternating voltage furnished by the line.
2. Change the pulsating voltages and currents to steady or smooth forms.
3. Raise or lower the line voltage as may be required to suit the needs of plate, screen, and grid circuits.
4. Lower the alternating voltage of the power line to values suitable for heater-cathodes and filament-cathodes in the various tubes.

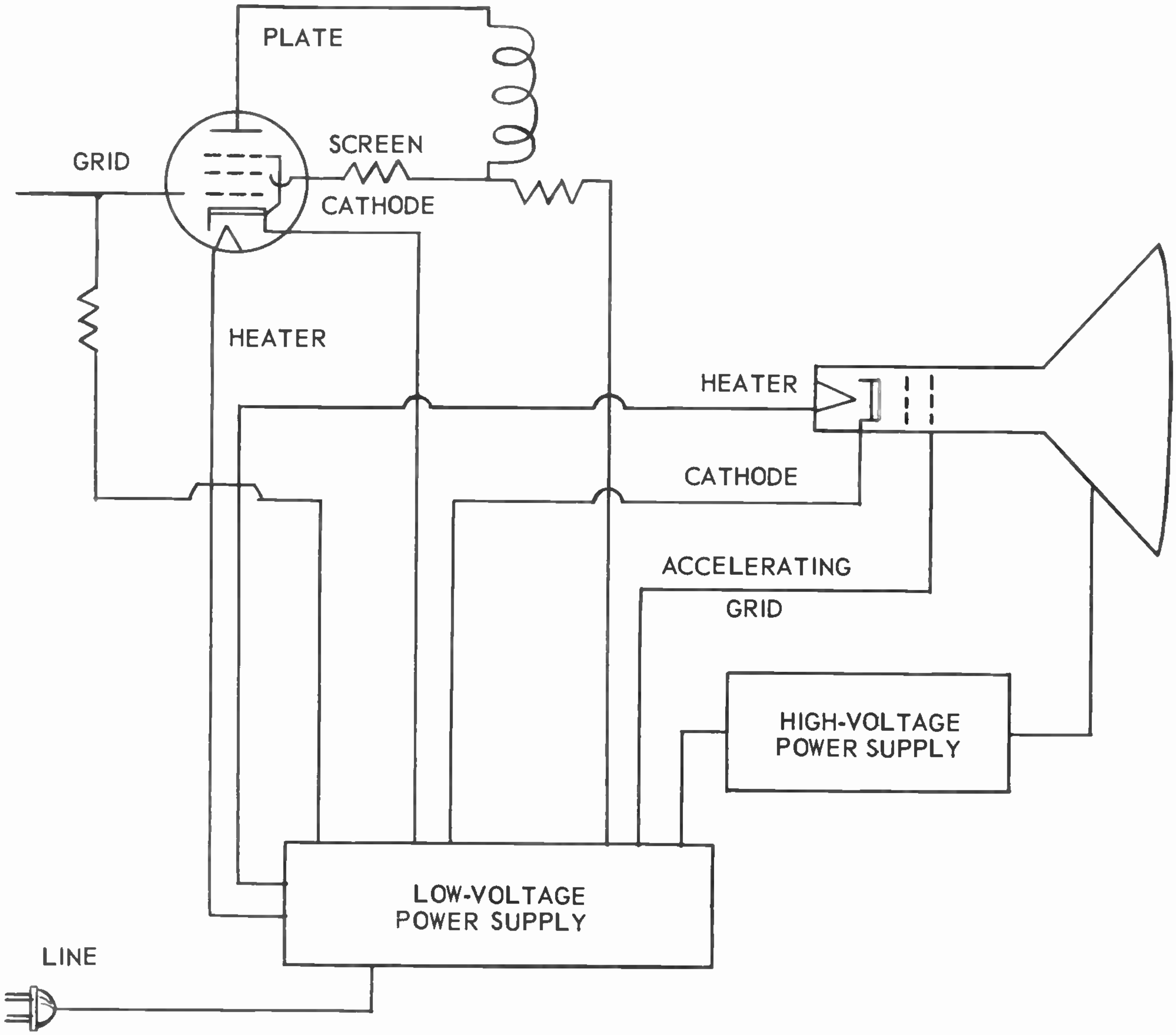


Fig. 19-1. Connections for low-voltage and high-voltage power supplies of a television receiver.

(6) Let's consider these four requirements, one at a time, and see what is needed to satisfy them. The first requirement is for direct voltages and currents. To produce direct voltages and currents from alternating line voltage it is necessary to have one or more rectifiers. As we have learned, when alternating voltage is applied to a rectifier there is electron flow in one direction but not in both directions. Electron flow in one direction is direct current.

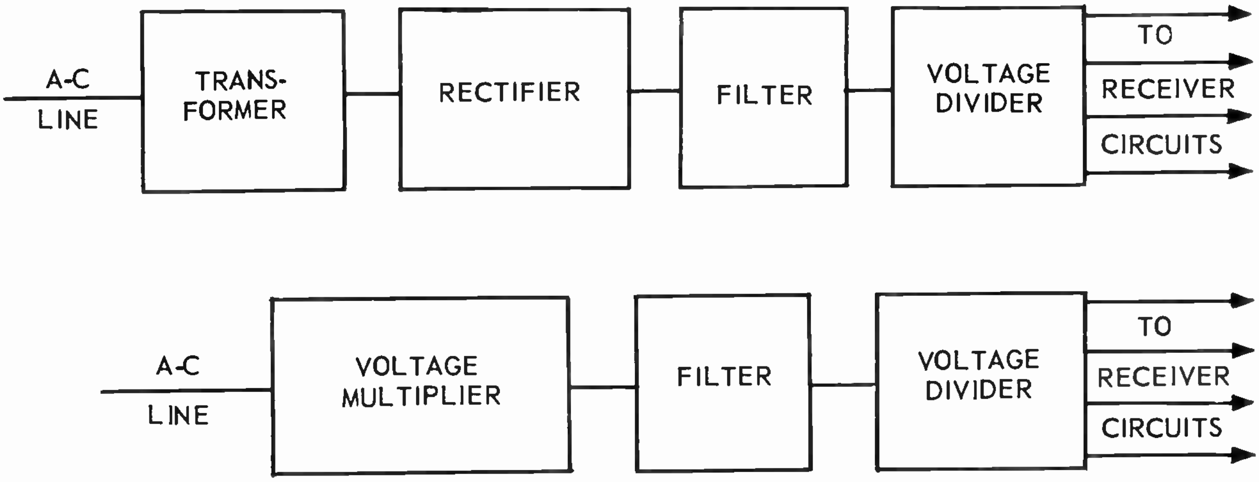


Fig. 19-2. The power transformer and rectifier sometimes are replaced by a voltage multiplier.

The second requirement is for smooth or steady direct voltages and currents. The direct current from a rectifier consists of intermittent pulses rather than of the steady current required by plate and screen circuits and by circuits in which grid bias voltage is developed. The pulses are smoothed out in a part of the power supply called the filter. Every power supply must have one or more filters.

The third requirement is for voltages of values suited to the various tube elements and receiver circuits. In some of the smaller and less costly sound radio receivers all plates and screens are operated at voltages in the neighborhood of 100 to 115 volts, which are easily obtained from the power line without any step-up. Any lower voltages are obtained by employing resistors to use up or drop a portion of the higher voltages.

In most radio receivers and in all television receivers it is necessary to have voltages much higher than can be obtained directly from the line. There are two distinctly different ways of raising the line voltage, or of raising voltage taken from the line. One way is to use a transformer. The dictionary says a transformer is an apparatus for transforming electric current from a high to a low potential or vice versa. The original use of transformers, and still their most important use in the electrical industry, is for raising and lowering voltages.

The other way of raising the voltage obtained from the line is to employ voltage multiplier circuits. These are combinations of rectifiers and capacitors so connected that two or more capacitors are individually charged to line voltage, then discharged in series either with one another or the line so the several voltages add together. This method is used in the numerous smaller and less costly radio and television sets often referred to as transformerless receivers. Fig. 19-2 illustrates in a simple manner the relations between parts of power supplies with and without transformers.

Whenever the need is for a voltage lower than that furnished from the line, from a transformer, or from a voltage multiplier, part of the original high voltage is used up or dropped in one or more resistors. Such resistors sometimes are connected into a circuit arrangement called a voltage divider.

The fourth of our original requirements calls for alternating voltages lower than line voltage, as needed for cathode heating. There are two different ways of supplying these lower voltages. First, if a transformer is used for raising the line voltage the same transformer may be used for lowering the line voltage to values suited for heaters and filaments. A single transformer connected to the power line may be designed to furnish any number of higher and lower voltages to the receiver circuits. The power supply transformer of Fig. 19-3 has leads or connections for furnishing many different voltages.

The other way of operating heaters or filaments at voltages lower than line voltage is to connect a number of them in series across the line. For example, if we connect in series with one another the heaters of enough tubes to have a total voltage requirement of 117 volts, that series "string" of heaters may be connected directly across a 117-volt line. Usually one or more series resistors must be inserted to have the necessary voltage match.

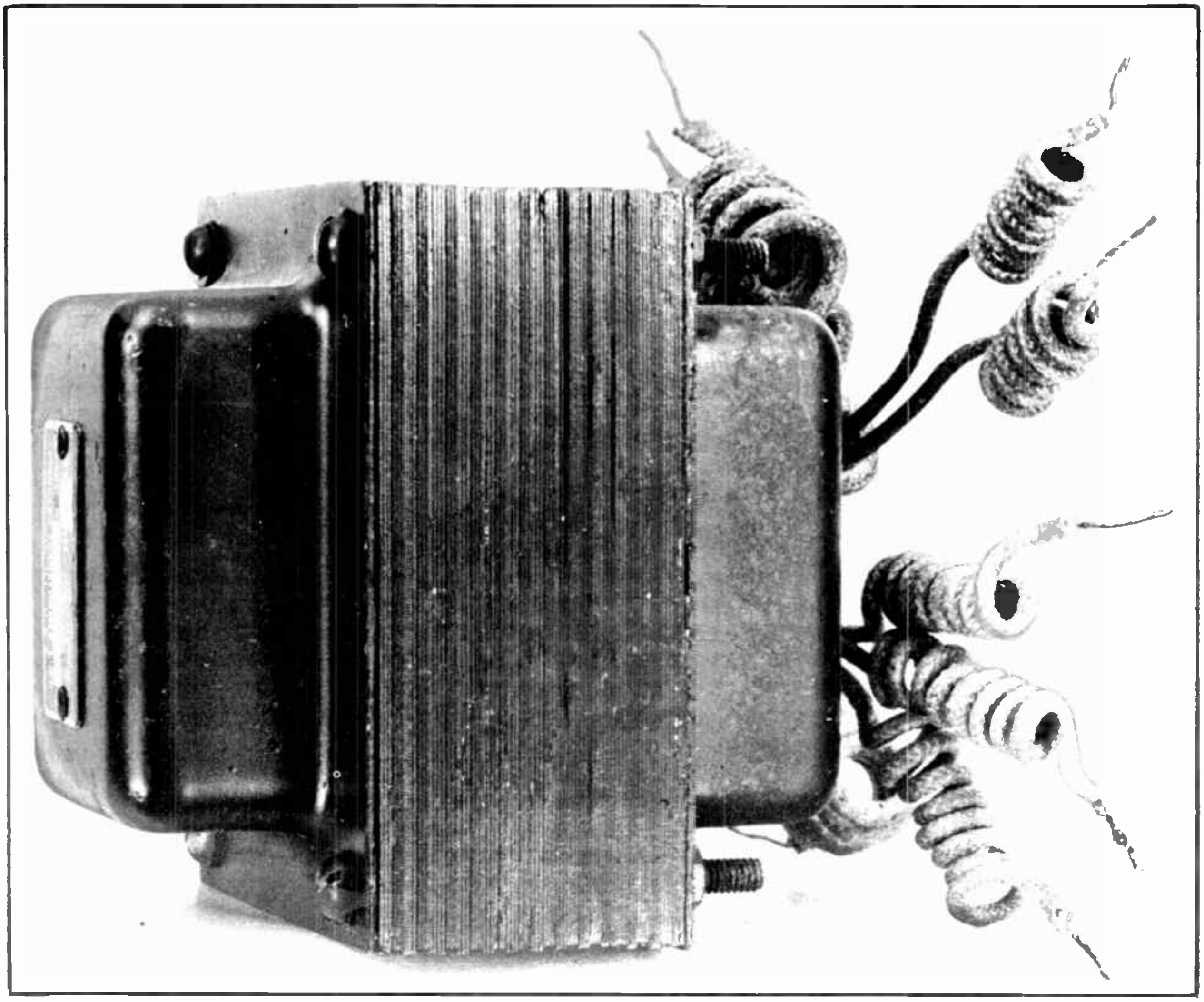


Fig. 19-3. A typical radio power transformer.

The accompanying chart or table summarizes the four requirements for power supplies and lists the apparatus or methods employed for meeting each of the requirements. The right-hand column of this summary forms a list of things which we shall examine while getting acquainted with low-voltage power supplies in general.

LOW-VOLTAGE POWER SUPPLIES FOR RECEIVERS

Requirements		*Apparatus and Methods*
1 Direct voltages and currents from alternating voltage	→	Rectifiers
2 Smooth direct voltages and currents	→	Filters
3 Direct voltages which differ from line voltage	Higher →	Transformers Voltage multipliers
	Lower →	Resistors Voltage dividers
4 Alternating voltages lower than line voltage	→	Transformers Series heaters and filaments

RECTIFIERS. An electronic tube employed as a power supply rectifier needs for its elements only a heated cathode and a relatively cool plate. Rectifier tubes are made with either filament-cathodes or heater-cathodes. As a general rule a rectifier designed to handle large current or for use at very high voltages will have a filament-cathode, while one for smaller currents and lower voltages usually has a heater-cathode.

All electronic tube rectifiers used in radio and television receivers are of the vacuum type, in which nearly all air and other gases have been pumped out of the glass or metal envelope to leave a nearly perfect vacuum. In some other branches of electronics, high-power rectifier tubes are of types whose envelopes contain mercury vapor or a gas such as argon or neon. Those gaseous rectifiers have less internal resistance between their cathodes and plates than have vacuum types, which is an advantage, but they are quite easily damaged by overloading. In addition, some gaseous rectifiers tend to cause severe interference with radio-frequency signals.

When a rectifier is connected in series with a source of alternating voltage and any kind of load, as at the left in Fig. 19-4, the rectifier plate will be made alternately positive and negative with reference to the cathode. The plate will be positive during half of each voltage cycle and negative during the other half. While the plate is positive there will be electron flow from the source through the load, thence from cathode to plate inside the rectifier tube, and back to the source. But while the plate is negative there will be no electron flow in this path. During each half-cycle of alternating voltage in which the rectifier plate is positive there will be a pulse of electron flow or current. These intermittent pulses may be represented as shown near the load.

In the left-hand diagram of Fig. 19-4 the rectifier tube has a heater-cathode. The heater for this cathode would be provided with voltage and current for heating, taken either from a transformer or from the line. Were the rectifier tube to have a filament-cathode, as at the right, the load circuit would be connected to one side of the cathode. Again the cathode would require voltage and current for heating the electron-emitting

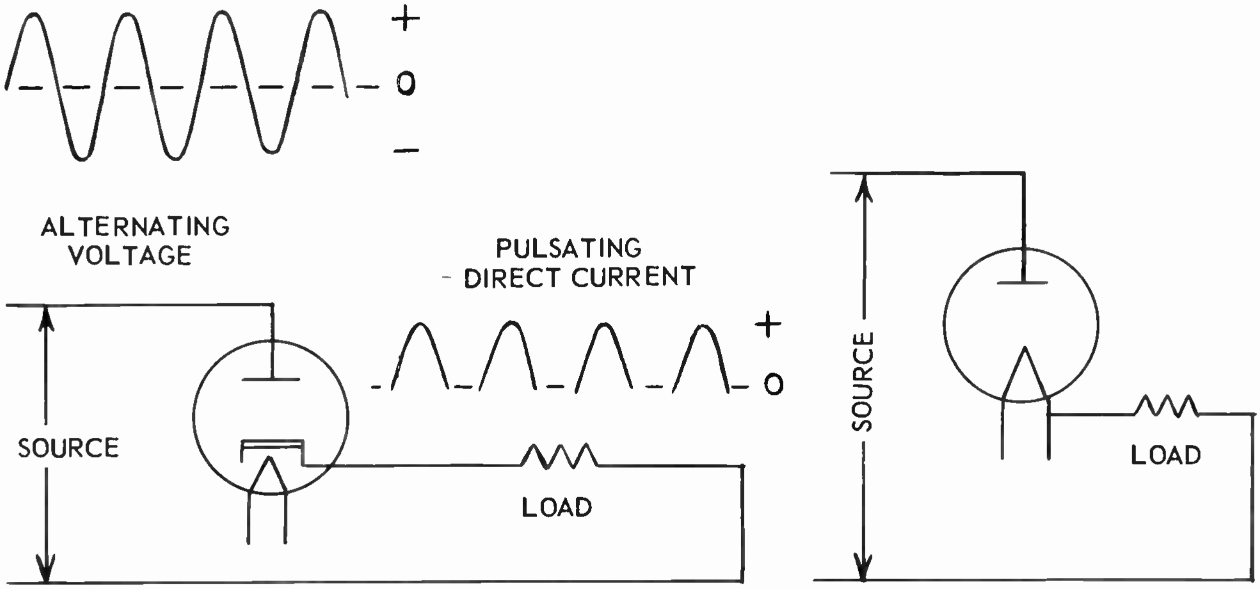


Fig. 19-4. Action of a half-wave rectifier.

surface. With filament-cathode rectifiers the heating current usually is furnished by a transformer, for if taken from the line there would have to be a direct connection between the high-voltage load circuit and the line.

In Fig. 19-5 voltage and current from a transformer are applied to a rectifier having a filament-cathode. The primary winding of the transformer is connected to the a-c line, with a switch in series with one lead so that the apparatus may be turned on and off. There are two secondary windings, a plate winding and a filament winding. Energy from the filament winding serves only to heat the filament-cathode of the rectifier tube, taking no further part in the rectifier action.

One end of the plate winding is connected directly to the plate of the rectifier tube. The other end of this winding is connected through the load circuit to the cathode of the rectifier. During half-cycles of alternating voltages which make the rectifier plate positive there will be pulses of electron flow in the direction shown by arrows. This flow direction is determined by the rectifier tube, because electrons can pass only from cathode to plate within the rectifier.

7 So far as the load circuit is concerned, the rectifier and the plate winding of the transformer make np the source of direct voltage and current. Since electron flow always is from negative to positive, the positive terminal of this rectifier-transformer source is the cathode of the rectifier. Always remember that the cathode of a rectifier is positive when considered with reference to the load circuit. By remembering this one fact you always can determine the direction of electron flow is connected circuits and in every part used in these circuits.

HALF-WAVE RECTIFIERS. The circuits illustrated by Figs. 19-4 and 19-5 are of the type called half-wave rectifiers. Only half of each voltage wave or voltage cycle is rectified, or only half of each cycle is effective in producing flow of direct current.

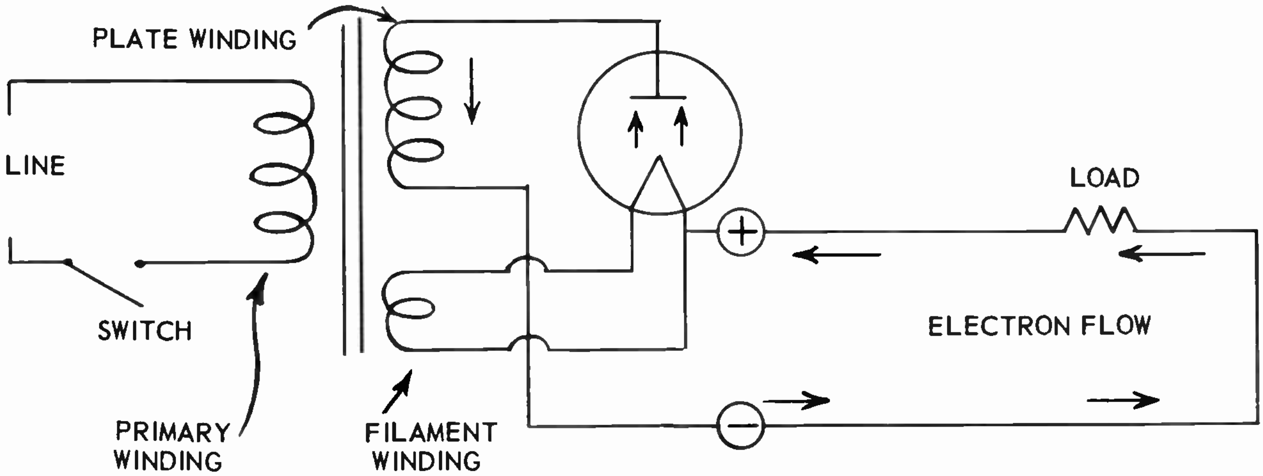


Fig. 19-5. Half-wave rectifier connected to a power transformer.

Fig. 19-6 shows at the left four cycles of alternating voltage applied to a half-wave rectifier, and at the right shows the resulting four cycles of direct current or the four pulses of direct current. We may speak of cycles of current because a cycle is any series of regularly recurring events. Each current cycle consists of a rise of current, then a drop of current, and an interval of zero current. Then these events repeat.

The frequency of current from a half-wave rectifier is the same as the frequency of the applied alternating voltage, or output frequency is the same as line frequency. If line frequency is 60 cycles per second, output frequency will be 60 cycles per second. A sound produced by the output current in a loud speaker would have the same fundamental pitch as a sound produced by the line voltage and current.

FULL-WAVE RECTIFIERS. Now we shall proceed to rectify both half-cycles of the alternating voltage, or to produce direct current pulses in the same direction from both the positive and the negative alternations of voltage. First as we shall examine the potentials obtained at the same instants from opposite ends of a secondary winding on a transformer. Both these potentials will produce rectified current pulses.

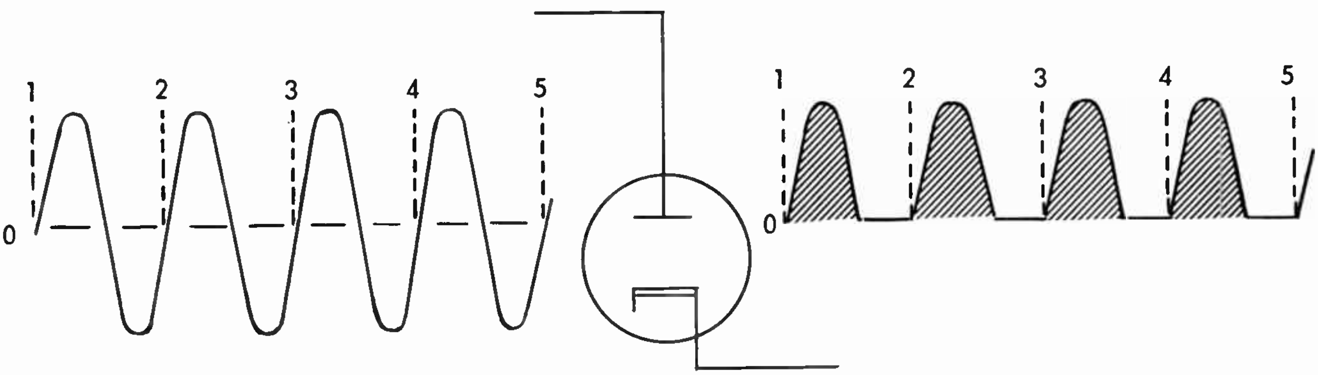


Fig. 19-6. Current pulses in the output from a half-wave rectifier.

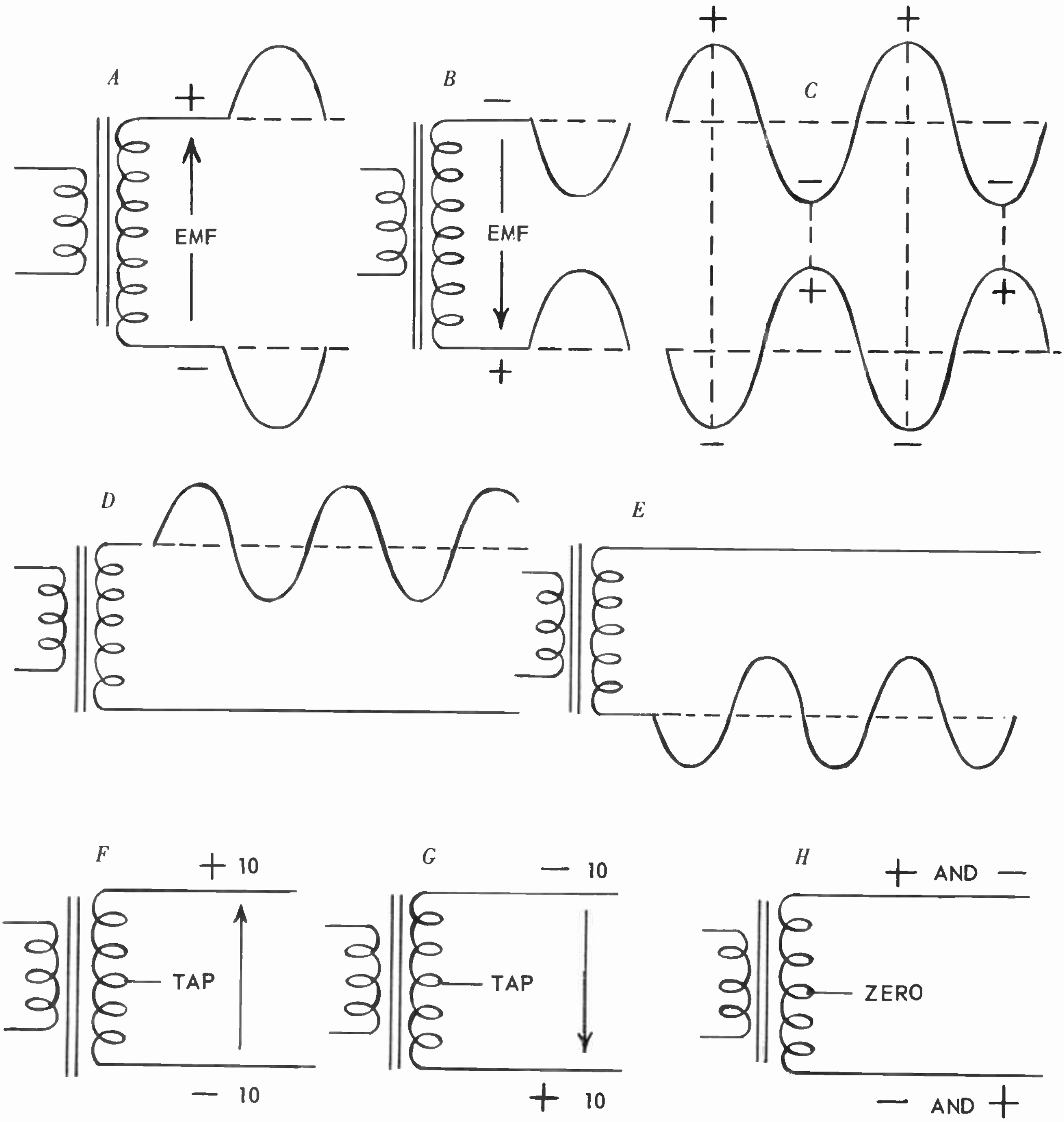


Fig. 19-7. Potential changes at opposite ends of a secondary winding.

Fig. 19-7 illustrates the changes of potentials in the secondary windings of transformers. As you know, secondary potentials result from emf's induced in the secondary winding when there are changes of current in the primary winding. When primary current is alternating, the induced secondary emf's and potentials likewise must be alternating.

During some one half-cycle the secondary emf will be acting in the direction shown by diagram *A*. This makes the upper end of the secondary positive with reference to the lower end, and makes the lower end negative with reference to the upper end, or we may say that the upper end becomes positive while the lower end becomes negative at the same time. Changes of potential at the two ends of the secondary during a half-cycle are as shown by the curves. These curves would show also the electron flows were the ends of the secondary connected to a closed circuit of any kind.

During the following half cycle there is reversal of the emf induced in the secondary winding, as shown by diagram B. Now the upper end of this winding is made negative with reference to the lower end, and the lower end is made positive with reference to the upper end. Again the changes of potential during the half cycle are as shown by the curves. These curves would show electron flows to and from the ends of the winding were there a closed circuit.

Since cycles of alternating current in the primary winding are continual, the corresponding alternations of emf or electron flow in the secondary winding may be shown by the continued curves at C. Note that opposite ends of the secondary always are of opposite polarity except during instants in which both pass through the zero value.

In many discussions of transformer action it is assumed, as in diagrams *D* and *E*, that either one end or the other of the secondary winding remains at zero potential while all alternations of potential and electron flow take place at the other end. Such an assumption cannot be correct, because electrons could not leave or enter either end of the secondary unless at the same time electrons entered or left the other end, and electrons could flow neither toward nor away from any point which remains always at zero potential.

There is a way of having a point at constant zero potential on a transformer secondary. It is illustrated by diagrams F, G, and H of Fig. 19-7. Here there is a connection to the exact electrical center of the secondary winding. This connection is called a center tap. We may assume, in diagram F, that the upper end of secondary is momentarily 10 volts positive and the lower end 10 volts negative. Midway between these equal positive and negative potentials the potential must be zero – just as midway between temperatures of 10° above zero and 10° below zero on the thermometer the temperature must be zero.

In diagram G the direction of induced emf has reversed, making the upper end of the winding 10 volts negative and the lower end 10 volts positive. Potential at the center tap still must be zero, for the same reason as explained for diagram F. No matter what the end potentials they always are equal and opposite, as shown by diagram C, and potential at the center tap always will remain zero.

At the right in Fig. 19-7 a tap connection has been added at the center of the transformer secondary. The potential at this center tap remains zero with reference to potentials at the outer ends or at the top and bottom of the secondary.

Now we go to the left-hand diagram of Fig. 19-8 where our center-tapped secondary winding is connected at its outer ends to the plates of two rectifiers, with its center tap connected to both cathodes through a resistor which may represent any load. During the half-cycle here considered the top of the transformer secondary is positive and the bottom is negative. The plate of the upper rectifier now is positive with reference to its cathode, and this rectifier will conduct. The plate of the lower rectifier is negative with reference to

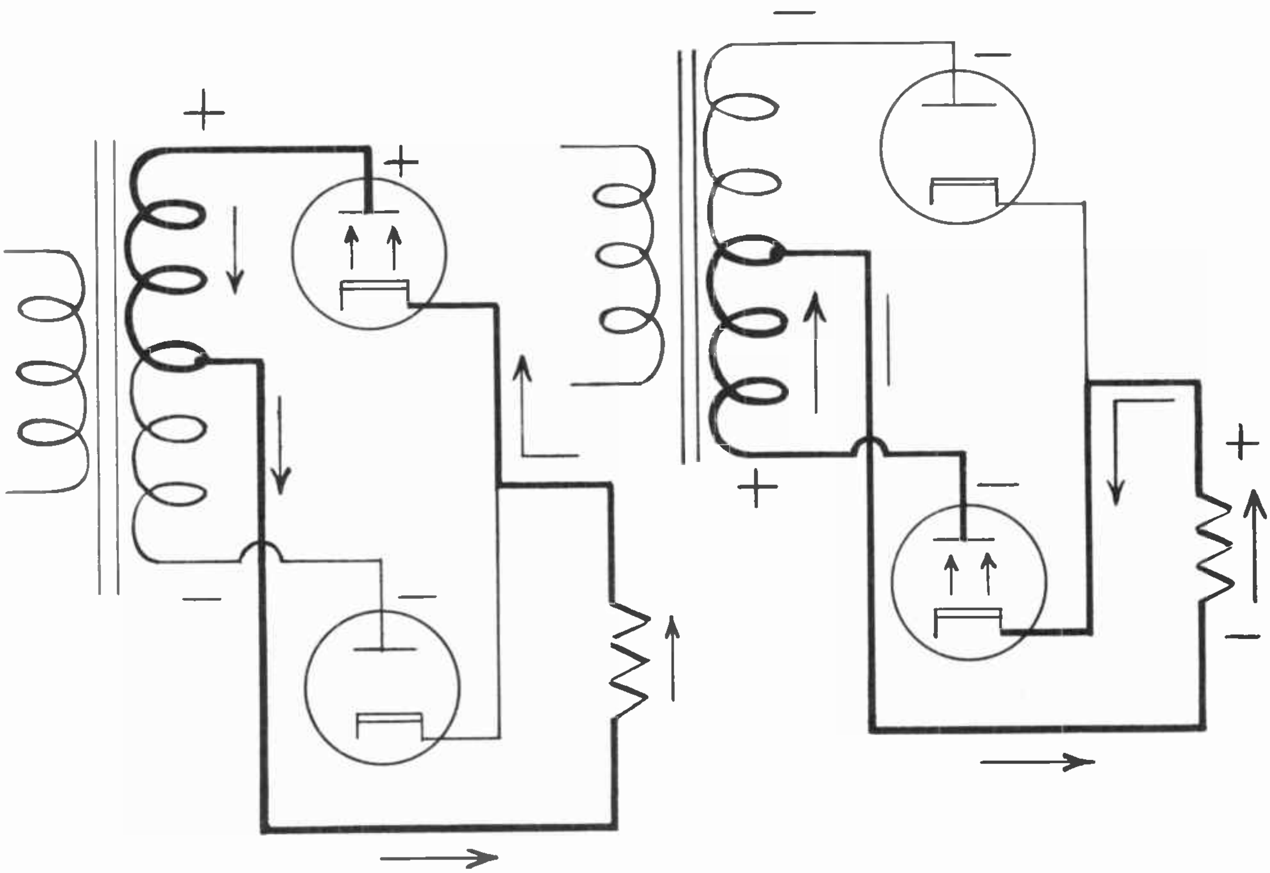


Fig. 19-8. Action of a full-wave rectifier using two tubes.

its cathode, so this rectifier will not conduct. Electron flow is through the path shown by heavy lines, and is in the direction indicated by arrows. Remember, the manner in which a rectifier is connected determines the direction of electron flow in the circuit.

Over at the right in Fig. 19-8 we are considering a half-cycle of voltage during which the top of the secondary winding is negative and the bottom positive. Now the plate of the upper rectifier is negative with reference to its cathode, and this rectifier cannot conduct. But the plate of the lower rectifier is positive with reference to the cathode in this tube, so there is conduction. The changed path for electron flow again is shown by heavy lines, and the direction of flow is indicated by arrows.

With the arrangement of Fig. 19-8 we are using both half-cycles of the alternating voltage induced in the transformer secondary. Note the direction of electron flow in the load, it is from bottom to top during both half-cycles of voltage. This is full-wave rectification.

(8) Full-wave rectification circuits ordinarily employ not two separate tubes but a single tube containing two plates and a single cathode which serves the same purpose as the two directly connected cathodes in Fig. 19-8. The first three symbols in Fig. 19-9 show the cathode and plate arrangements for generally used full-

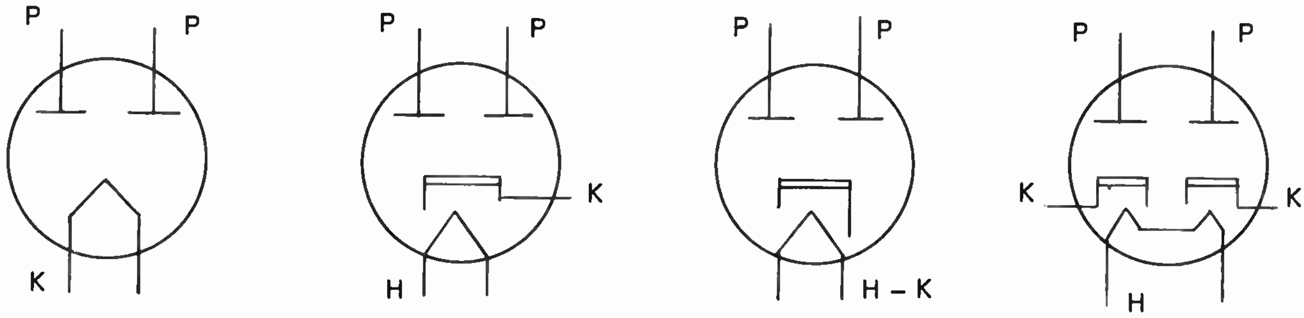


Fig. 19-9. Symbols for rectifier tubes.

wave rectifier tubes. The first symbol is for a filament-cathode type. The second symbol shows a cathode completely insulated from its heater and having a separate terminal or base pin, while the third symbol shows an indirectly heated cathode connected internally to one end of the heater so that one of the base pins serves for both cathode and heater. The symbol at the right shows a tube with a separate cathode for each of the plates. This is a type used in voltage multiplier circuits.

At the left in Fig. 19-10 are represented the alternating potentials applied to the two plates of a full-wave rectifier tube. The positive alternations are rectified, each positive alternation of both waves producing a direct-current pulse from the rectifier output. These pulses are shown at the right. During the period of each complete cycle of input voltage there are two positive alternations, one at each end of the center-tapped transformer winding. During this same period of time there are two current pulses in the output of the rectifier. Consequently, the frequency in the output of a full-wave rectifier is twice the frequency of the input voltage or is twice the line frequency. If line frequency is 60 cycles per second, the output frequency of a full-wave rectifier will be 120 cycles per second.

Looking back at Fig. 19-8 makes it plain that only half the plate winding or secondary winding of the transformer is working at any one time. Then the voltage accompanying the d-c output current of the full-wave

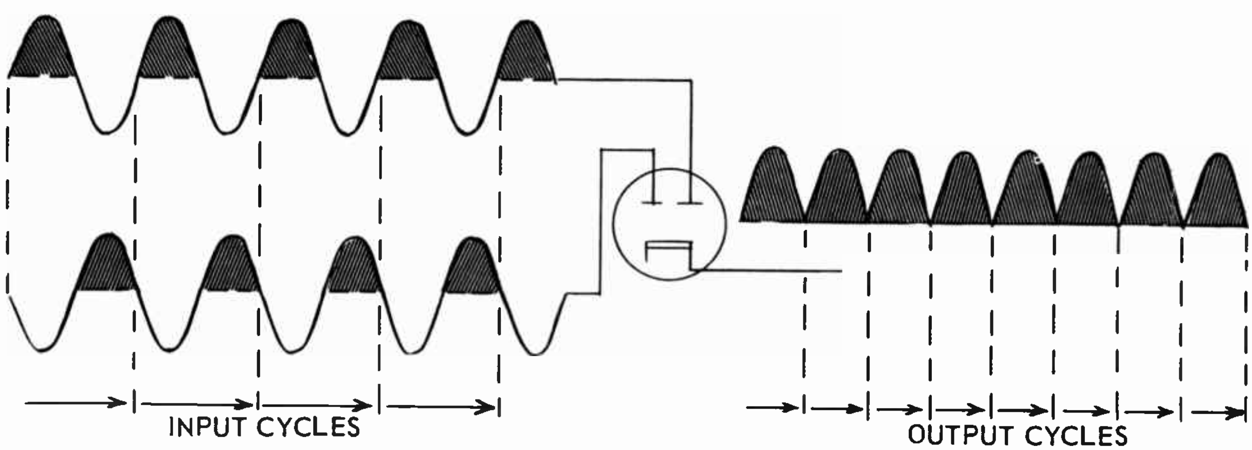


Fig. 19-10. Current pulses in the output from a full-wave rectifier.

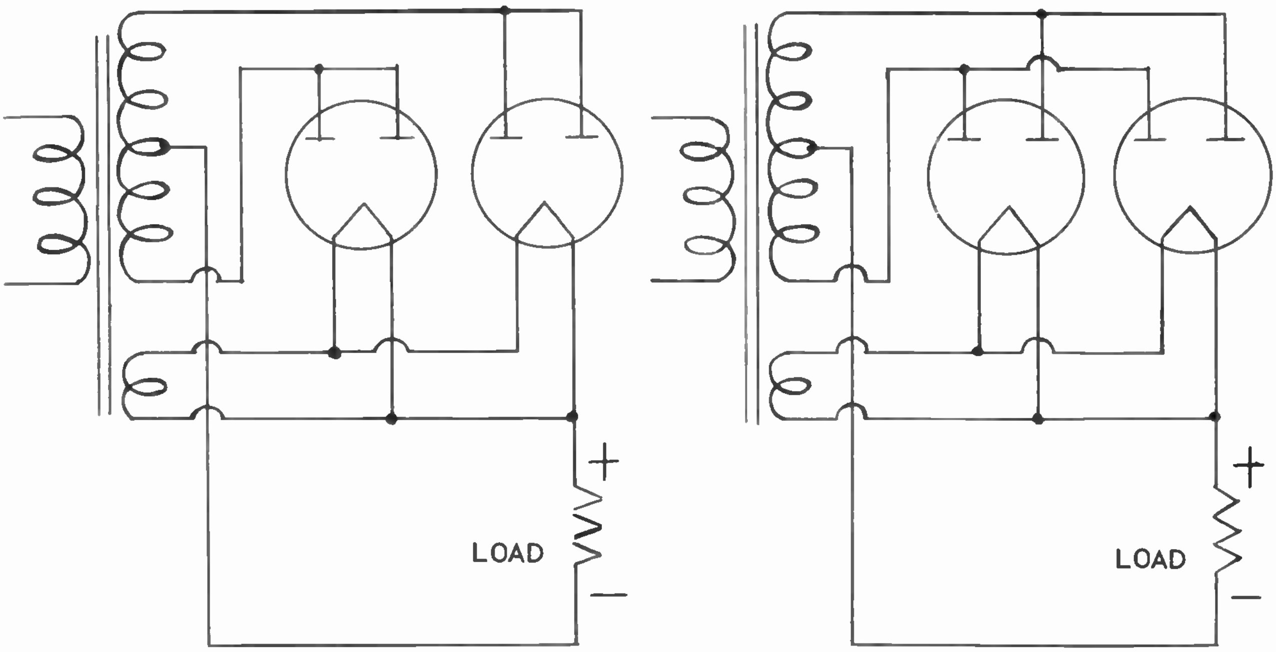


Fig. 19-11. Rectifier tubes and plates connected in parallel.

rectifier is only that which corresponds to the voltage across one half the secondary winding. If alternating voltage from end to end of the secondary winding is, for example, 800 volts we have only 400 volts effective at any one time for production of rectified current and voltage in the output. The a-c voltage of this center-tapped secondary winding usually would be specified as 400-0-400 a-c volts, or as 400 volts each side of center tap, or as 800 volts center-tapped, or in some other way making it clear that the overall voltage is divided between the two parts of the full-wave rectifier.

RECTIFIERS IN PARALLEL. Standard types of rectifier tubes are well able to handle the highest voltages required in low-voltage power supply systems, but quite often the total current required for plate, screen, and grid bias circuits is more than can be handled by a single tube. The obvious remedy is to use two or more rectifier tubes, and this is exactly what is done in many television receivers and in some of the larger radio receivers and sound reproduction systems. In some receivers there are two complete low-voltage power supplies, each with its own transformer, rectifier, and filter. Part of the amplifiers and other tubes receive plate and screen current from one rectifier, while remaining tubes are connected to the second rectifier.

Another method of obtaining increased rectifier capacity is shown at the left in Fig. 19-11. The circuit connections are the same as in Fig. 19-8, but instead of using rectifier tubes with only a single plate in each we now use full-wave rectifiers with the two plates of each tube connected together to form the equivalent of one large plate. The plates are said to be connected in parallel. The filament-cathodes of both tubes are similarly connected together, in parallel, and are operated from a single filament winding on the transformer. The two cathodes thus connected are the equivalent of a single large cathode.

Sometimes you will find full-wave rectifier tubes connected in parallel with the method shown at the right in Fig. 19-11. Here the two plates in one tube are not tied together, but one plate in each tube is connected

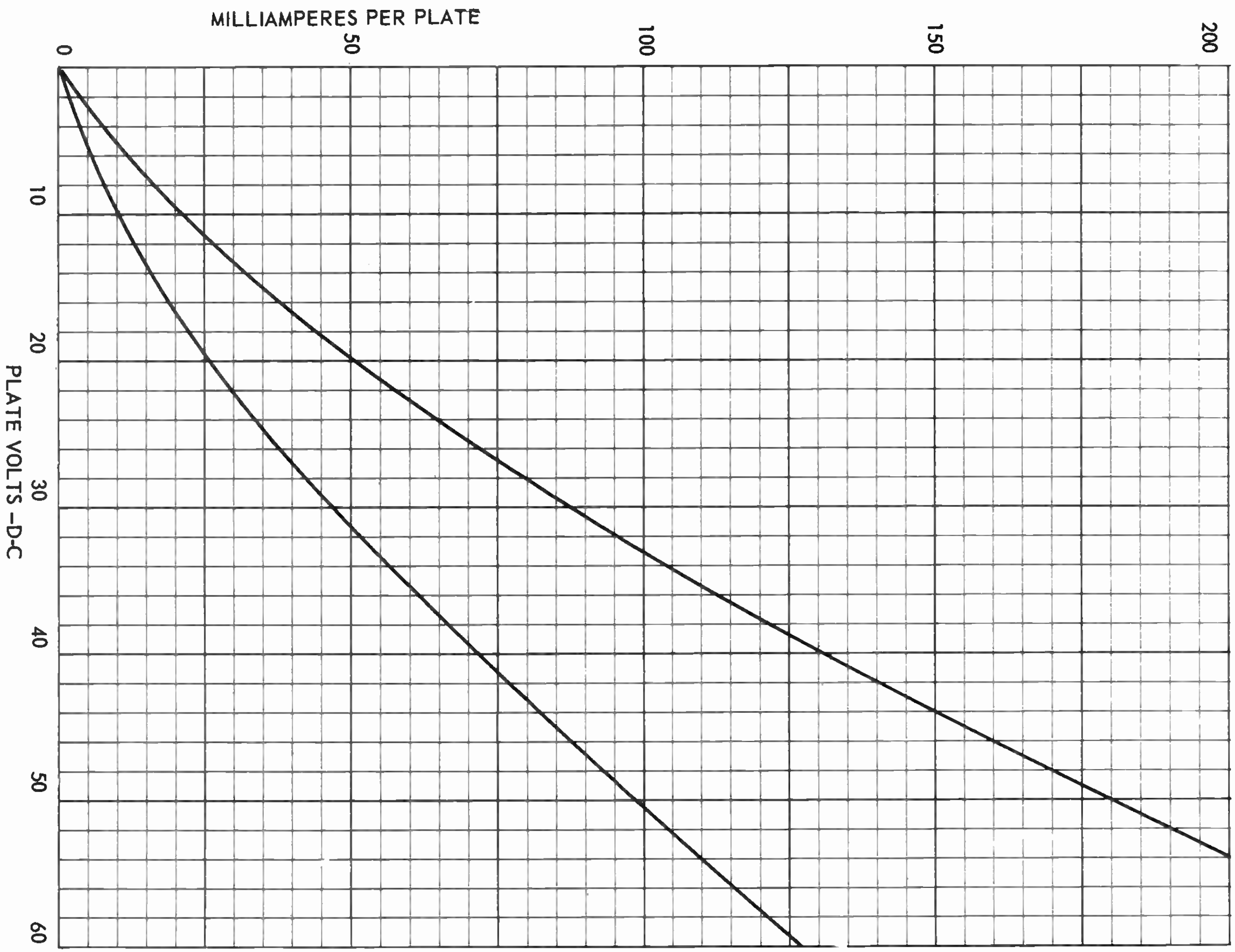


Fig. 19-12. Plate characteristics of two rectifier tubes. Upper curve is for a 5U4-G tube. Lower curve is for a 5Y3-GT tube.

to one plate in the other tube. With this connection there is half the total current in each cathode at all times, since both cathodes are working during both half-cycles or both alternations of the applied alternating voltage. With connections as at the left, the cathodes work alternately and each carries the total current during one half-cycle.

5 *RECTIFIER CHARACTERISTICS AND RATINGS.* Rectifier tubes, like all other tubes, have plate characteristics showing relations between plate voltage and plate current. Two typical characteristic curves are shown in Fig. 19-12. The upper curve applies to the 5U4-G tube and the lower curve applies to the 5Y3-GT tube. Both are full-wave rectifiers with filament-cathodes, and both are widely used in radio receivers and television receivers. The curves show currents in milliamperes which would flow in the tubes with various direct voltages applied between plate and cathode, or they show the voltage drop or loss in the tube for various values of current.

We could compute the apparent resistance between cathode and plate of the rectifiers by multiplying any of the voltages by 1,000 and dividing that product by the number of milliamperes of current. Such computations for the 5U4-G tube (upper curve) would show the apparent internal resistance decreasing from about 475 ohms at 10 d-c volts to about 280 ohms at 50 d-c volts. With the 5Y3-GT tube (lower curve) the apparent internal resistance drops from about 1,000 ohms to approximately 500 ohms between the same applied d-c voltages. The 5Y3-GT has about the highest internal resistance of any power rectifier used in radio and television receivers. There are a number of commonly used power rectifiers having apparent internal resistances only about one-fourth that of the 5Y3-GT. All the rectifiers having low internal resistances or impedances are of types having heater-cathodes rather than filament-cathodes, but a few of the high-impedance rectifiers also have heater-cathodes.

2 In manufacturers' listings of rectifier ratings there always is one called "peak inverse voltage". This is the highest instantaneous voltage that the tube will withstand in a polarity opposite to that with which the tube is designed to pass current, or with cathode positive and plate negative. With sine wave alternating voltage applied to the tube, the inverse peak would be approximately 1.4 times the effective voltage.

As an example, peak inverse voltage for the 5U4-G tube is 1,550 volts. This would be the peak value in a sine wave voltage having an effective value of about 1,100 volts. Then applying more than 1,100 effective a-c volts to the 5U4-G ordinarily would cause its breakdown. Peak inverse voltages occur during every half-cycle in which there is no regular conduction, or during every half-cycle in which the plate is negative and the cathode positive. Some of the miniature tube rectifiers have peak inverse voltages as low as 330 to 350 volts, corresponding to effective a-c voltages of about 240 volts.

Sometimes there is a rating also for maximum d-c milliamperes delivered to the load. This maximum load current from the rectifier varies with a number of working conditions, among which are the kind of filter following the rectifier and the effective value of alternating voltage applied between each plate and the cathode of the rectifier tube. Maximum allowable d-c current from a 5Y3-GT rectifier usually is on the order of 125 milliamperes, and from a 5U4-G rectifier is no more than 225 milliamperes.

SELENIUM RECTIFIERS. There are many devices other than electronic tubes which are capable of rectifying alternating voltages or of allowing only a pulsating direct current to flow in a circuit to which alternating voltage is applied. Among them are contact rectifiers which employ various combinations of metallic elements and compounds that permit current to flow quite freely through the contacting surfaces in one direction while offering great opposition to reverse flow. Of the several types of contact rectifiers the one using selenium as one of the elements is most suitable for the rather high voltages found in plate and screen circuits. Accordingly, we find many of the smaller radio and television receivers built with power supplies which include selenium rectifiers instead of tube rectifiers.

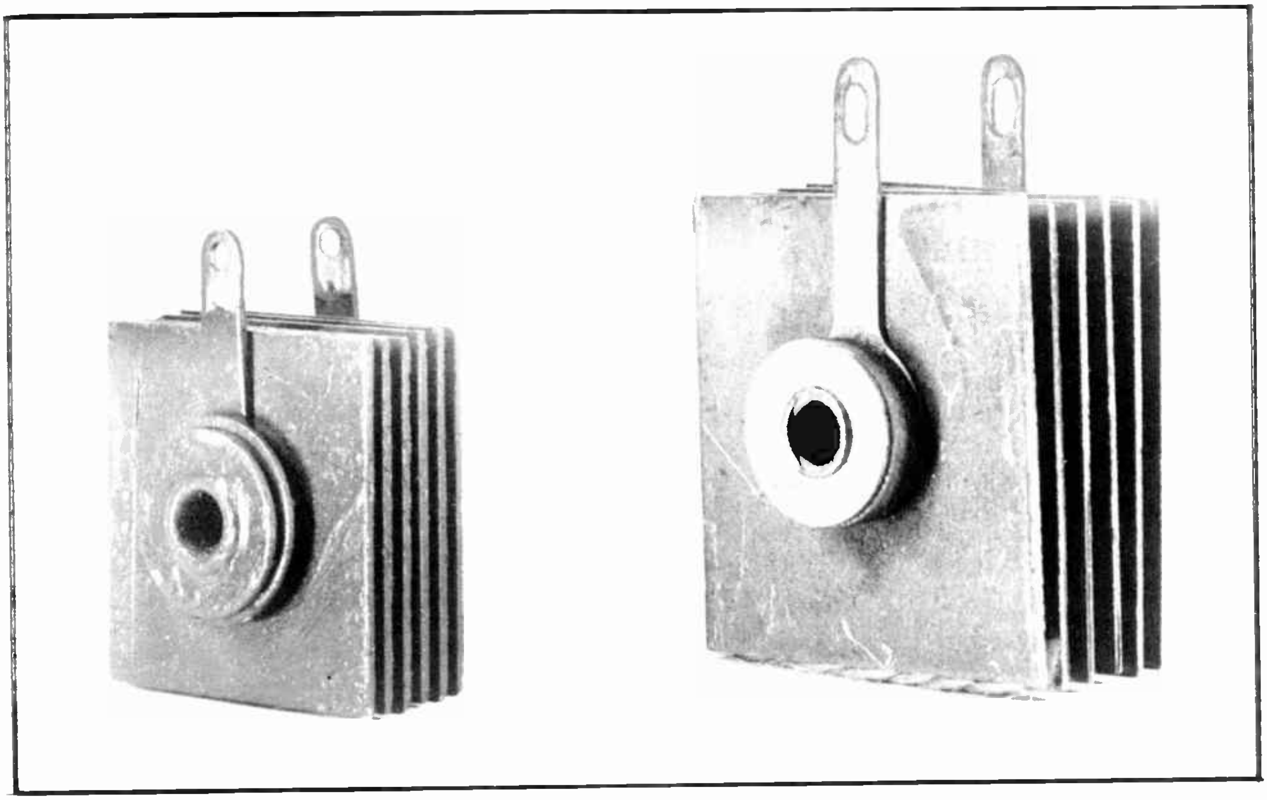


Fig. 19-13. Selenium rectifiers.

Fig. 19-13 is a picture of selenium rectifiers capable of furnishing direct currents of 75 and 150 milliamperes. Such units, in the several common sizes, measure one to two inches from side to side and up to an inch from front to back. Selenium rectifiers, in common with other contact rectifiers, require no heating of their cathodes, consequently take no power for this purpose and have no heater wiring. These units are mounted on the chassis or any convenient bracket with a screw or bolt which passes through the center hole in the rectifier. Thus the usual tube socket is eliminated.

Types of selenium rectifiers used in radio and television receivers are available in capacities for maximum direct-current outputs from 50 to 500 milliamperes. Maximum applied alternating voltage is 130 or 160 volts effective or rms value. Peak inverse voltages are 380 or 440 volts maximum. These low limits of voltage impose no particular difficulties, since selenium rectifiers seldom are used other than in transformerless receivers or power supplies, so are not subjected to anything higher than the alternating line voltage from building light and power circuits.

The usual symbol for a selenium rectifier is shown by Fig. 19-14. It looks somewhat like an arrowhead pointed against a bar or line. Direction of rectified electron flow through the rectifier is shown by an arrow. While examining wiring diagrams it is convenient to keep in mind that electron flow is *against* the arrowhead – assuming, of course, that the symbol is used or drawn correctly in the diagram.

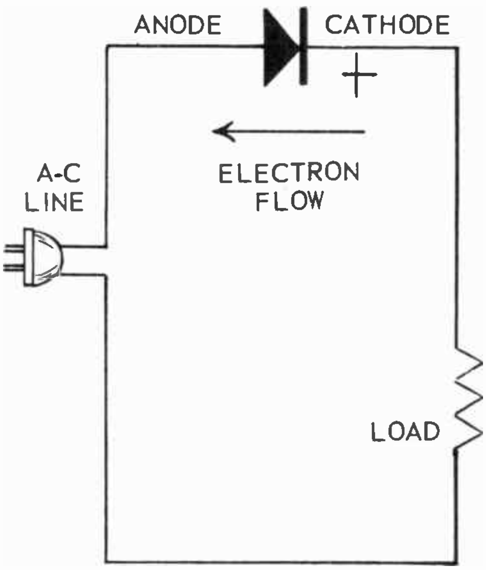


Fig. 19-14. Selenium rectifier in a load circuit.

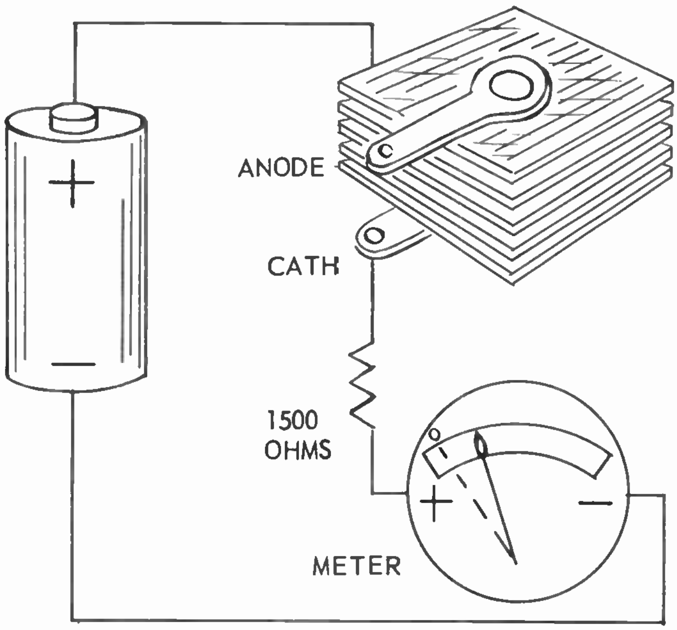


Fig. 19-15. Identifying the polarity of a selenium rectifier.

The terminal at which electrons enter the selenium rectifier is called the cathode, just as the element of an electronic tube at which electrons enter is called the cathode of the tube. The terminal of the selenium rectifier from which electrons leave the unit is called the anode. It corresponds to the plate of a tube. In fact, the name anode is entirely correct also for the plate of any tube. In nearly every electronic field except radio and television the element from which electrons leave any kind of tube is called the anode, not the plate.

The cathode terminal on some selenium rectifiers is marked CATH, on others it is marked with a positive (+) sign, and on still others is marked with a red dot. The anode ordinarily is not marked or specially identified in any way. If no cathode identification is visible you can check the rectifier polarity with an ohmmeter whose positive and negative terminals are known or marked. Resistance of the rectifier in the direction of normal electron flow will be only a few thousand ohms, but in the reverse direction will be almost infinite on most low-range ohmmeters.

Another way of checking the polarity of any contact rectifier is illustrated by Fig. 19-15. The rectifier is connected in series with one dry cell, a resistor of not less than 1,500 ohms resistance, and a sensitive current meter reading not less than one milliampere at full scale. The resistor will protect the meter in case the rectifier should turn out to be short circuited. The rectifier should be connected into this circuit first one way and then the reverse way and meter readings noted. With the greater of the two meter readings (more current) the rectifier anode is toward the positive terminal of the dry cell and the cathode is toward the negative terminal or zinc can of the cell.

Selenium rectifiers for plate and screen power supplies usually are connected into half-wave rectification circuits. An elementary half-wave circuit with a selenium or other contact rectifier is shown by Fig. 19-14. These rectifiers can be connected also into full-wave circuits as shown by Fig. 19-16, but then a center-tapped transformer is needed. Inasmuch as common types of selenium rectifier units will not withstand alternating voltages higher than line voltage, the only purpose of this transformer is to provide a center tap. It must not furnish a secondary voltage higher than could be secured directly from the line. The slight advantage of full-wave rectification seldom is worth the extra cost of the transformer.

The bolt or screw for mounting a selenium rectifier should preferably be insulated from chassis metal in spite of the fact that the bushing through the center hole of the rectifier usually is insulated from the active elements. The primary purpose of the thin metal fins on the rectifier is to assist the transfer of heat from the elements into surrounding air. Consequently, the rectifier should be mounted where there is opportunity for reasonably rapid circulation of air, and preferably with the fins vertical, as in Fig. 19-13. The location should not be close to any large tubes or large resistors which could radiate their heat to the rectifier. Maximum safe operating temperature for rectifier elements is 75° C or 167° F.

The metal fins of selenium rectifiers are electrically alive, they are conductively connected to the cathode and anode terminals. Precautions must be taken to avoid shortcircuiting the fins on other live parts. Rectifiers carry currents and voltages at power line frequency, which is within the operating range of audio-frequency amplifiers. For this reason the rectifiers should not be located close to audio-frequency circuits in which there might be hum pickup, rather the rectifiers should be near r-f or i-f parts of the receiver where the amplifying circuits are not particularly responsive to line frequency. This precaution with reference to rectifier mounting position applies also to tube rectifiers of all kinds.

Any circuit containing a selenium rectifier must be so arranged that the rectifier cannot under any circumstances become connected directly across the line without additional resistance in series. With a direct connection the rectifier would pass enough current to heat it to destruction in a very short time. The minimum series resistance usually is on the order of 20 ohms. Most of the failures of selenium rectifiers are due to current overloads, which result from short circuiting of parts in the load circuit, and from overheating which may accompany an overload or which may be caused by lack of ventilation.

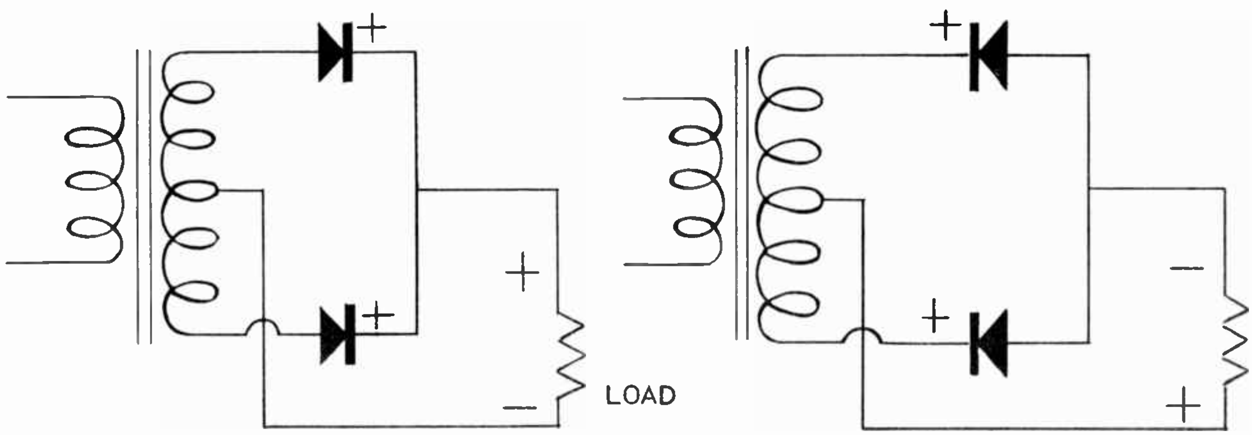


Fig. 19-16. Full-wave rectifier circuits with selenium units.

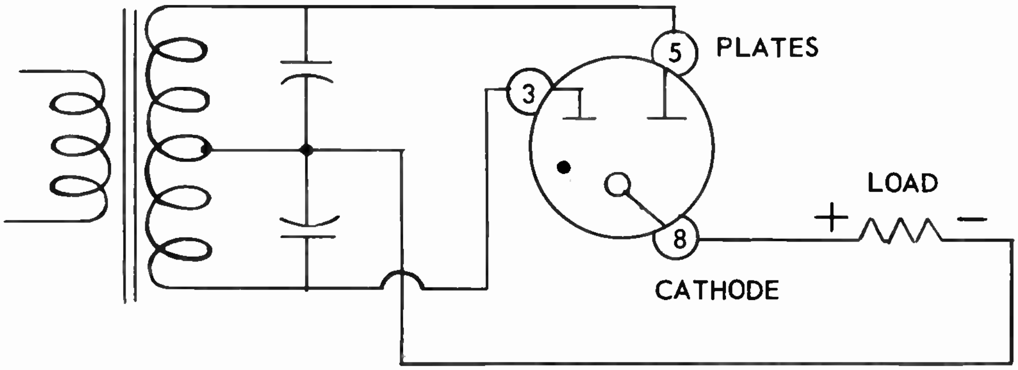


Fig. 19-17. Connections for a full-wave cold-cathode rectifier tube.

COLD-CATHODE RECTIFIER TUBES. Designers always have tried to avoid the need for hot cathodes, with their consumption of heating power and added wiring for heater or filament circuits. In many early radio receivers this desire lead to the use of cold-cathode rectifier tubes, among the several types of which you still may run across the 0Z4 in old equipment. This rectifier is used in a full-wave circuit shown by Fig. 19-17, where the tube is represented by its usual symbol. The two capacitors across the transformer secondary winding are "buffer" capacitors whose purpose is to bypass high-frequency currents caused by rectifier action and thus lessen the radio-frequency interference due to this class of tube.

The cold-cathode tube is not a vacuum type, it is gas filled. In a gas filled tube the change of potential in the space between anode and cathode is not uniform, as it is in a vacuum tube. For example, were the potential difference between plate and cathode of a gaseous tube to be 300 volts, only about 60 volts would be used up in 95 per cent of the distance between tube elements. The remaining 240 volts would act in the last 5 per cent of the distance, at the cathode end, so energetically as literally to pull electrons out of the cold surface of the cathode material.

To start the cold-cathode rectifier into action requires a minimum "breakdown" potential difference of 300 volts between plates and cathode. Then conduction commences, there is a blue glow near the cathode, and the potential drop through the tube decreases to about 24 volts. A current overload will cause a white-hot spot to appear on the cathode, which is soon destroyed by the intense localized heating. Maximum d-c output current of the 0Z4 tube is 75 milliamperes. Current through the tube must not drop to less than 30 milliamperes at any time or else conduction will cease entirely.

MERCURY-VAPOR RECTIFIERS. Another rectifier formerly used more than at present is the mercury-vapor type which is shown connected into a full-wave circuit by Fig. 19-18. This tube has a hot cathode of the filament type. The bulb is filled with the vapor of mercury and there are a few small drops of the liquid metal within the tube. In the vapor-filled space is a blue-green glow while the tube is operating. The small black dot in the circle of the tube symbol indicates that the envelope contains a vapor or a gas rather than a nearly complete vacuum. The same dot appears on the tube symbol of Fig. 19-17.

The two mercury-vapor tubes which have been used in power supplies, chiefly for public address and other heavy-duty sound systems, are types 82 and 83. Type 82 is now practically obsolete, but type 83 still is in use. This latter type is the equivalent of the 5U4-G vacuum rectifier in current handling capacity and voltage limits.

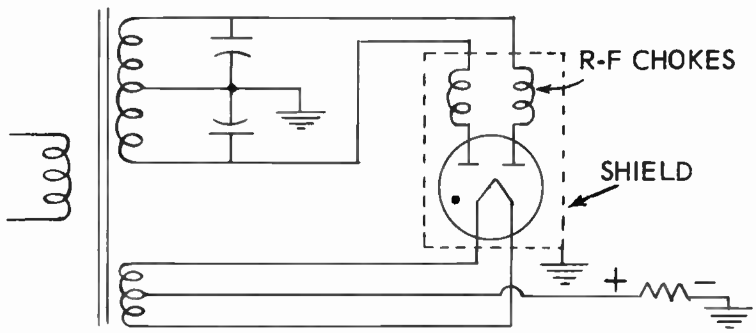


Fig. 19-18. Connections for a full-wave mercury-vapor rectifier.

The chief advantage of gas-filled and vapor-filled rectifiers is an internal voltage drop only a small fraction of the drop in vacuum rectifiers. This helps to maintain fairly steady voltages across the loads even with rather large variations of current.

Mercury-vapor rectifiers, like cold-cathode types, produce radio-frequency current changes which cause serious interference in receivers. To lessen the interference there are bypass capacitors across the transformer secondary winding in Fig. 19-18, also high-reactance inductors called r-f chokes in the plate leads, and a grounded metal shield around the entire tube, including its base and socket. Another disadvantage of the vapor rectifiers is that their maximum operating temperature should not exceed 140°F, and, to maintain vaporization of the mercury, must not fall below 68°F. Mounting must be with the bulb up and the base down in order that vapor condensing on the glass will run down into the bottom of the envelope.

TELEVISION

LESSON NO. 20

POWER TRANSFORMERS AND VOLTAGE MULTIPLIERS

Training Your Mind

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

DO NOT TEAR — CUT ALONG THIS LINE AND SEND IN FOR GRADING.

COMMERCIAL TRADES INSTITUTE

LESSON NO. 20

POWER TRANSFORMERS AND VOLTAGE MULTIPLIERS

When first talking about power supplies we looked at a picture of a typical power transformer from which extended a number of wire leads or connections for supplying various secondary voltages and for connection to the power line. Now we shall take that transformer apart, thus learning how such units are constructed while discussing the purposes of the various parts.

As mounted in a receiver, the power transformer is wholly or partially enclosed by sheet metal covers. With these covers removed, in Fig. 20-1, we can see the steel core and all the wire leads extending out from the windings or coil. Always there is a primary winding and one or more secondary windings which, considered all together, usually are spoken of as the "coil". Most power transformers are provided with leads, like those shown here, for connection into receiver circuits, but some have lugs to which may be soldered the receiver wiring.

The core of the transformer is made up of many thin sheets, called laminations. The laminations can be slipped apart and removed from the coil. In Fig. 20-2 the core, after having been removed from the coil, has been reassembled to show how it is made. Although the core actually is of a special grade of steel, we usually speak of it as the core iron. In the core are openings called windows, through which pass the sides of the coil. The core iron extends through the center of the coil and also around on opposite sides. Thus there is a complete circuit of iron all the way around the paths followed by the magnetic lines or flux.

7 The core is made of thin laminations instead of one or two solid pieces in order to make it more difficult for eddy currents to form in the iron. Eddy currents are electric currents induced in the conductive core iron by the varying magnetic field, just as currents are induced in the wire of the windings. These eddy currents heat the core and cause loss of energy. By using thin laminations insulated from one another the eddy currents can progress only a little ways in their natural direction of flow, and are kept relatively weak. Cores of various thicknesses for transformers of different power handling capacities may be assembled from a single shape of lamination. We often speak of a transformer as having a stack (core) of various thicknesses.

The coil with all its windings appears as in Fig. 20-3. Take note of the little wire sticking out all by itself on one side; later we shall see what it is for. In spite of the fact that the coil windings consist of flexible copper wire, we find the unit to be very solid and rigid. This is because the spaces in the windings have been impregnated with insulating varnish or waxes of suitable compounds, then baked to drive out all moisture and leave the completed coil assembly quite resistant to entrance of moisture later on.

Upon removing the outer covering of heavy kraft paper we come to the winding for the rectifier filament or heater, which is pictured in Fig. 20-4. When a rectifier has a filament-cathode or when its cathode is connected to one side of the heater, the filament or heater will be at the highest positive potential in the

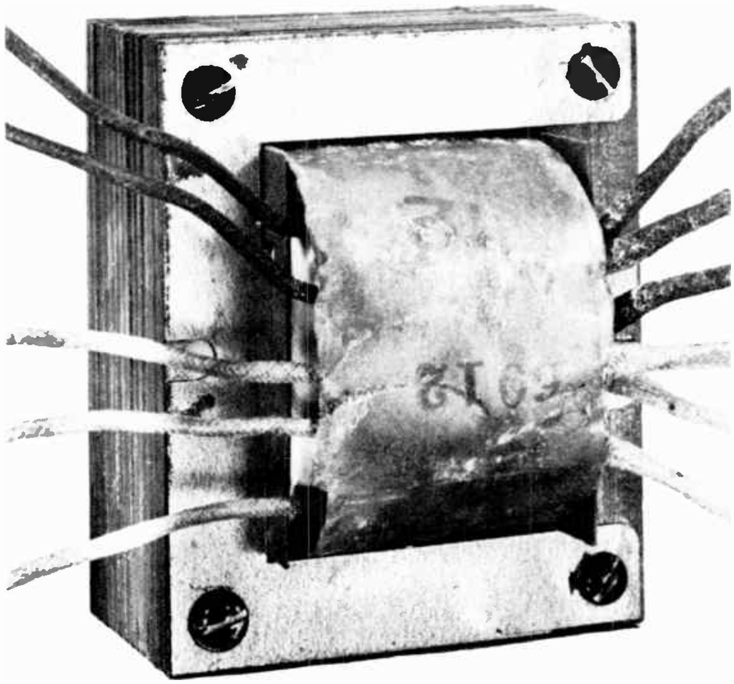


Fig. 20-1. The core and the "coil" of the power transformer, with all the connection leads from windings.

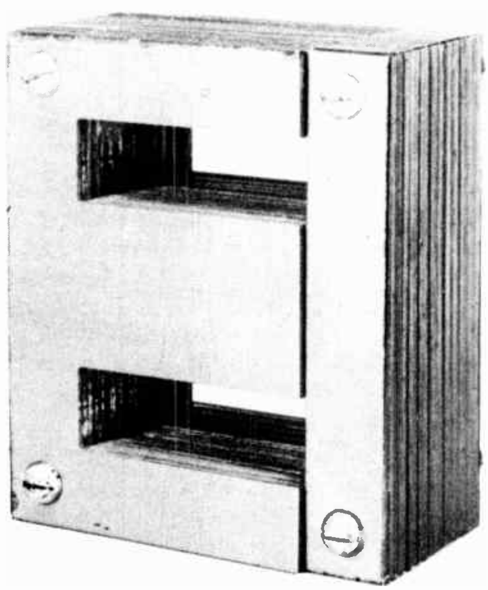


Fig. 20-2. The core after being taken out of the coil and reassembled.

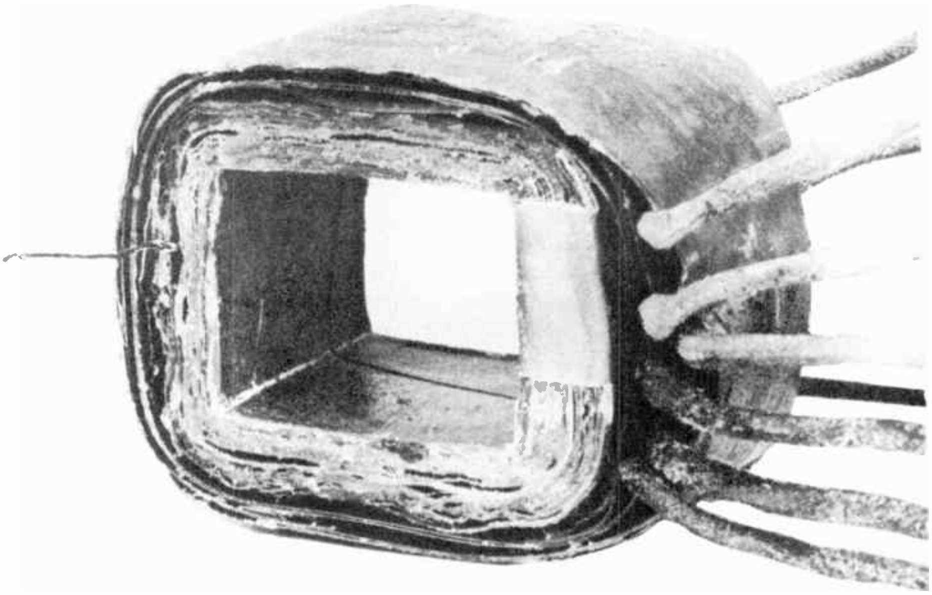


Fig. 20-3. The coil unit in which are all the windings.

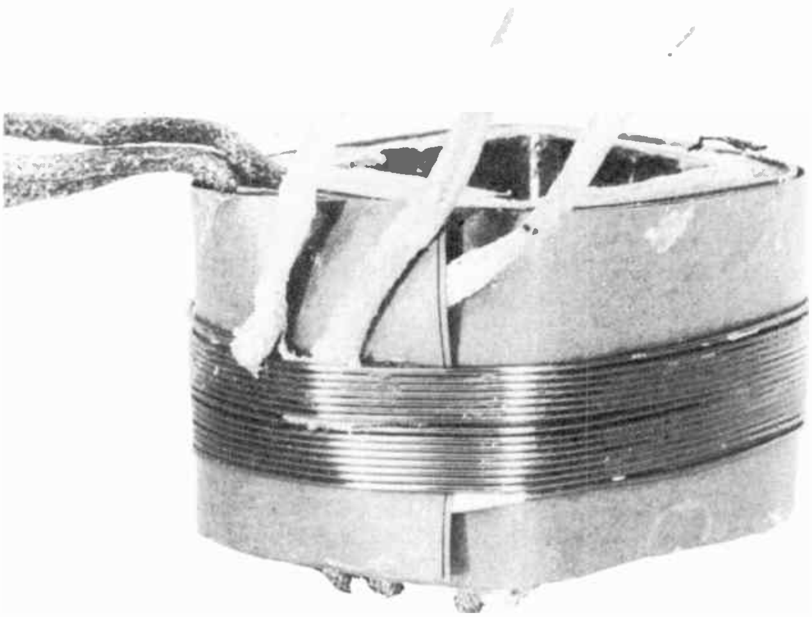


Fig. 20-4. Secondary winding for the rectifier filament.

d-c power circuit. Because of this possible high voltage with certain rectifiers the rectifier filament winding is separate from the winding which provides heater current for amplifiers and other tubes. In addition, the rectifier filament winding must be very well insulated because of its high potential.

After taking off the rectifier filament winding we come to the amplifier heater winding of Fig. 20-5 which furnishes current for heaters of tubes other than the rectifier. The amplifier heater winding and the rectifier filament winding provide only low voltages, but they must furnish currents of several amperes in most cases. To carry the heavy currents these windings are made of large wire, as is plainly apparent in the pictures, but they have only a few turns of this wire.

VOLTAGE AND TURNS RATIOS. The relation between winding turns and voltage calls for some explanation. Each turn of wire in the primary winding and in every secondary winding is cut by the same lines of force when the magnetic field expands and contracts. The emf induced in any one turn of every winding is the same as in every turn of every other winding. The total emf induced in any one complete winding is equal to this emf per turn multiplied by the number of turns. Consequently, there is a close relation between the number of turns in any winding and the voltage which is induced in and which appears across the ends of that winding.

Were you to measure the resistance of the primary winding in a typical radio power transformer it probably would be no more than 3 to 4 ohms. Were the primary to be connected to a direct-current power line furnishing 120 volts the small resistance would allow current of 30 to 40 amperes. Power dissipation would be about 4,000 watts. The resulting heat would burn out the transformer coil. But if connected to an a-c line furnishing 120 alternating volts the primary winding would take only about 1/5 ampere of current provided no power were being taken from the secondary windings, and even when handling full rated power the heating would be only moderate.

A transformer must never be connected to a power line furnishing smooth direct current nor to any other direct-current source capable of furnishing more than a small fraction of an ampere. The result of connection to a strong d-c source is invariably a ruined transformer. Often the wire of the primary will melt apart, and in any case the heat is certain to burn away the insulation and allow short circuited turns.

4 The instant you connect the primary winding to a source of alternating voltage furnishing the frequency for which the transformer is designed, a counter-emf is induced in the winding. If no current is being taken from the secondary windings this primary counter-emf is nearly equal to the applied alternating voltage. It is the small difference between applied voltage and primary counter-emf that causes the current of about 1/5 ampere. This current is just enough to supply power for overcoming resistance of the winding wire and for eddy current and hysteresis energy losses in the core iron.

In the primary winding of the transformer which we are taking apart there are 340 turns. With 120 applied alternating volts the induced counter-emf, in our particular transformer, is about 118.6 volts. Then the induced counter-emf per turn must be very close to 0.35 volt (118.6 volts divided by 340 turns). This induced emf per turn results from cutting of the magnetic field lines through the turns as the field expands and contracts.

As we shall see when getting back to disassembly of the transformer, the high-voltage secondary winding is wound right over the primary winding. Consequently, except for a small amount of "leakage flux" which escapes, every moving magnetic line which cuts through the primary turns cuts also through every secondary turn. This being the case, there must be induced in each turn of every secondary winding an emf of approximately 0.35 volt, the same as induced emf per turn in the primary.

On the high-voltage secondary winding of our transformer there are 2140 turns. Since induced emf is 0.35 volt per turn, the total secondary emf must be 749 volts (2140 x 0.35). This secondary emf or voltage exists

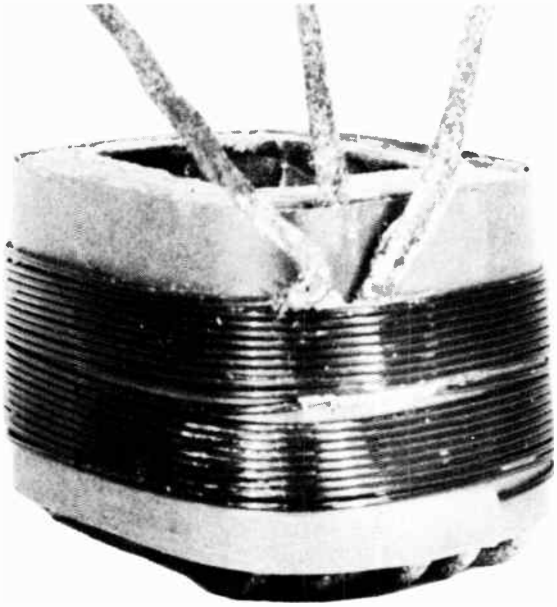


Fig. 20-5. Secondary winding for heaters of amplifiers and other tubes.

when no current is being taken from the secondary winding, or when this winding is open circuited. Incidentally, the high-voltage secondary often is called the plate winding, because its principal function is to furnish voltage and current for the plates of amplifiers and other tubes in the receiver.

Now we shall compare the ratio of secondary and primary voltages with the ratio of secondary and primary turns. A ratio is the result of dividing one number by another. To compute the voltage ratio we proceed thus:

749 (secondary volts) ÷ 118.6 (primary volts) = 6.3, approximately.
To compute the turns ratio we make another division, thus:
2140 (secondary turns) ÷ 340 (primary turns) = 6.3, approximately.

Had our original computation of voltage per turn been carried out to a more exact fraction, which would be 0.3488 volt, and were the divisions for ratios made with similar exactness, we should discover an important fact relating to transformers having iron cores. The fact is: The ratio of induced voltages or emf's is the same as the ratio of turns when no currents are being taken from the secondary windings. The ratios do not remain equal when secondary windings are delivering currents.

(1) When we connect the secondary windings to filaments, heaters, plate circuits, and screen circuits, the power used in all these circuits must come originally from the line into the primary and thence to the secondary windings of the transformer. Primary current will increase to provide for whatever power is being delivered from the secondary windings, and also to compensate for additional losses due to voltage drops in the windings and to stronger magnetization of the core iron. Primary voltage will remain equal to line voltage, but secondary voltages will drop. The voltage drop in the secondaries is due largely to an effect by which secondary currents tend to produce magnetic flux in opposition to flux produced by primary current, thus lessening total effective flux and induction. With all secondary windings delivering their full rated currents, the voltage from the plate winding of our particular transformer will drop from 749 to about 700, and other secondary voltages will drop as explained below.

The same induced emf or voltage per turn exists in all windings. On the rectifier filament winding of Fig. 20-4 there are 16 turns. With the induced emf of 0.35 volt per turn we have,

16 (turns) × 0.35 (volt) = 5.6 volts.

This is the voltage with no current drawn from the rectifier filament winding. When this secondary winding furnishes 3.0 amperes for the filament-cathode of a 5U4-G rectifier tube the voltage will drop to 5.0, which is correct for this tube.

On the amplifier heater winding of Fig. 20-5 there are 20 turns. Then,

20 (turns) × 0.35 (volt) = 7.0 volts, with no current from the winding.

Should this secondary winding furnish a total of 3.5 amperes for heaters of various tubes in the receiver its output voltage will drop to 6.3, which is standard heater voltage for a great many types of tubes.

The transformer with which we are working furnishes secondary voltages both higher and lower than voltage from the power line to the primary. A transformer which supplies secondary voltage higher than primary voltage is a step-up transformer. One which supplies secondary voltage lower than primary voltage is a step-down transformer. Our transformer furnishes secondary voltages which are stepped up and stepped down, it is a combination type.

VOLT-AMPERES. We have learned that power is measured in watts. This is true in circuits having only resistance, and neither inductance nor capacitance producing any appreciable reactance to alternating currents. It is true also in circuits carrying only smooth direct current, for there can be no reactance unless the current varies. In circuits such as just mentioned we may use this formula.

Watts = volts potential difference × amperes of current

Heaters and filaments in tubes have no appreciable reactance. A rectifier filament taking 3 amperes at 5 volts is using power at the rate of 15 watts.

In circuits which do contain appreciable reactance, due either to inductance or capacitance, a great deal of the input current (as primary current in a transformer) merely builds up magnetic fields. Much of the power or energy from this current is returned to the line when magnetic fields collapse. The input current may merely charge capacitors, which return most of the energy when they discharge. These currents furnish no useful power in the output of the circuit, they furnish no power which is correctly measured in watts.

Were you to measure the volts and amperes out into a transformer primary from the line, the product of volts and amperes would be considerably greater than the number of useful watts of power being taken from the secondaries. Yet the portion of the input current which only builds up magnetic fields, or a portion which would charge capacitors, should be given consideration because this current flows in the primary and the line connections and causes heating and loss of energy. To make allowance for the energy wasted by the "wattless" currents we measure power input to a transformer in volt-amperes, not in watts. Volt-amperes of power are equal to the product of input volts and input amperes. We might say that volt-amperes measure total power consumed, even though part of it is wasted, while watts measure total power which does something useful or desired.

In many cases the output power from a transformer secondary goes to a circuit containing considerable reactance. This happens when the plate winding feeds a filter, for in filters there is a great deal of capacitive reactance, in capacitors, and often inductive reactance in choke coils. Then we have wattless currents in the filter and in the plate winding. For such reasons, it is desirable, and is common practice, to rate transformer output power as well as input power in volt-amperes.

This brings us to a second fundamental fact relating to transformers: Except for losses in the winding conductors and in the core iron, output volt-amperes will be equal to input volt-amperes.

The power transformer with which we are working has output ratings shown by the accompanying list, which includes also the corresponding volt-amperes of power. Although the plate winding ordinarily would be rated as furnishing 700 volts at normal current, it is a center-tapped winding of which only half works at one time when connected to a full-wave rectifier. Therefore, for volt-ampere computations, we assume only half the total voltage, or 350 volts. Note that the total output volt-ampere power for all the secondaries together is the sum of their separate powers.

SECONDARY WINDINGS	VOLTS	CURRENT	VOLT-AMPERES
Plate	350	90 ma or 0.09 amp	31.5
Rectifier filament	5.0	3.0 amperes	15
Amplifier heaters	6.3	3.5 amperes	22
Total power output			68.5

Primary volt-amperes at full load then should be, theoretically, 68.5 volt-amperes, the sum of all secondary volt-amperes. Actually the primary volt-amperes will be more than this, in order to care for various energy losses. Measurement with a-c meters on this transformer shows primary power of 84.5 volt-amperes, nearly one-fourth more than the output volt-amperes.

Fig. 20-6. To the terminal strip are fastened leads and ends of windings for primary and plate.

A

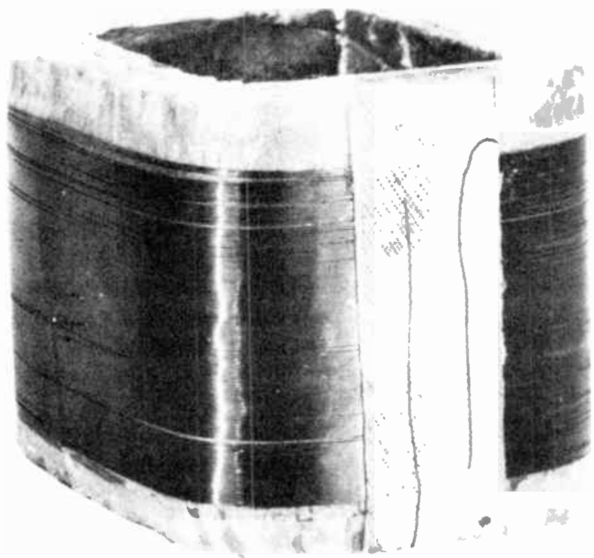


Fig. 20-7. The high-voltage plate winding.

The ratio of output power to input power is the power efficiency of the transformer. Dividing output power, 68.5 volt-amperes, by input power, 84.5 volt-amperes, gives the ratio as about 0.81. This may be called a power efficiency of 81%, which is fairly typical of high-quality radio and television power transformers.

TRANSFORMER WINDINGS. Now we may get back to disassembly of the transformer. Upon removing the amplifier heater winding we find the terminal strip pictured in Fig. 20-6. Here are the leads of stout wire for the plate winding and the primary winding. Wire in the primary winding itself is only about 1/50 inch in diameter, and in the plate winding is only about 1/140 inch in diameter. These small wires are soldered to the heavier and stronger leads at the terminal strip.

With the terminal strip removed we see the plate winding as in Fig. 20-7. The ends of the very small wire show up clearly against the binding tape. Most of the insulating wrappings which have been removed to expose successive windings are of strong kraft paper about 1/100 inch thick, but in some plates are of varnished or oiled linen cloth where still more insulation is required. Were you to unwind the wire from the plate winding you would find each layer separated from other layers by high-quality paper about 1/300 inch thick.

After the plate winding is out of the way, and just before coming to the primary winding, we uncover an electrostatic shield which can be seen in Fig. 20-8. This is a strip of very thin copper wrapped all the way around, but not joined at its ends. To one end is attached the small wire which we noted back in Fig. 20-3. This wire makes contact with the core iron of the assembled transformer, so that grounding of the core will also ground this electrostatic shield.

10 The purpose of the electrostatic shield is to prevent radio-frequency signal voltages being transferred from the primary winding into the plate winding and going from there to the plate and screen circuits of amplifiers. Such signal voltages come through all power lines into the primary winding, which is underneath the electrostatic shield. The shield metal might be thin aluminum or any other non-magnetic metal. In transformers

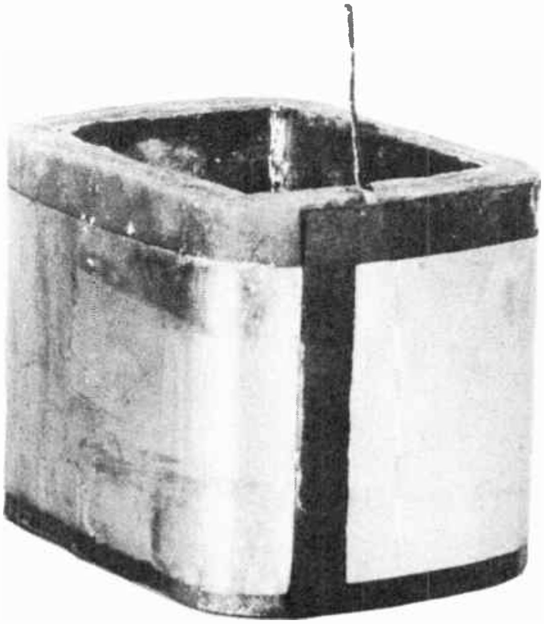


Fig. 20-8. The electrostatic shield which is between the primary and all secondaries.

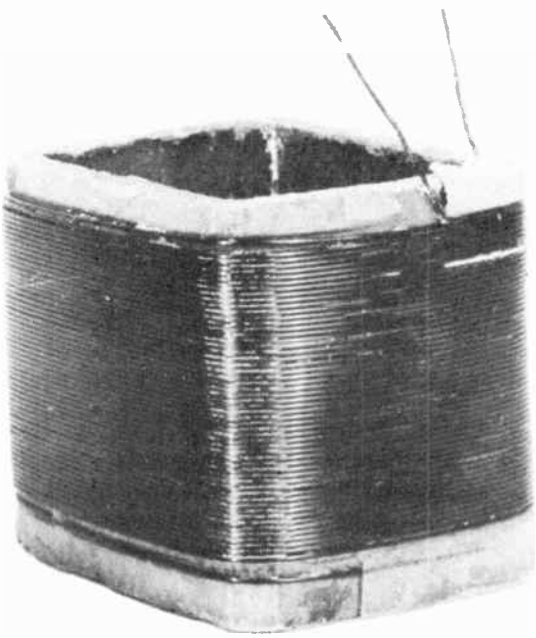


Fig. 20-9. The primary winding, which is next to the core.

having more than one plate winding there may be additional shields. This shielding against r-f interference is common for power transformers used in television receivers and in high-quality sound radios.

While on the subject of transformer shielding it may be mentioned that many power transformers for television receivers have a wrapping of thin copper about two inches wide around the outside of the core iron and the coil. The ends of this band are soldered together to make a single closed turn. The purpose is to prevent the alternating magnetic field of the transformer from spreading so far as to reach the deflection yoke on magnetic-deflection picture tubes. If this field does reach the deflection yoke it is likely to cause continual expansion and contraction of pictures.

The result of the final step in disassembly of the power transformer is shown by Fig. 20-9, where the primary winding is exposed. The primary is wound on a rectangular cardboard form called the bobbin. The bobbin slips over the center leg of the core.

5 The transformer which has been taken apart is a type designed for operation on 50-cycle or 60-cycle a-c power lines. Somewhat different proportions are required for operation on 25-cycle power lines found in some localities. This is because there is less induced emf per turn at 25 cycles than at 60 cycles, since induction depends on the rapidity of cutting of magnetic lines through conductors.

The lessened induction at 25 cycles is compensated for by using more turns in the windings and larger cores to work most effectively with the larger windings. Transformers for use on 25-cycle lines have about 2.2 times as much core iron and about 1.7 times as much wire in the windings as do equivalent 60-cycle types. Naturally, the cost of 25-cycle transformers is more than of 60-cycle types.

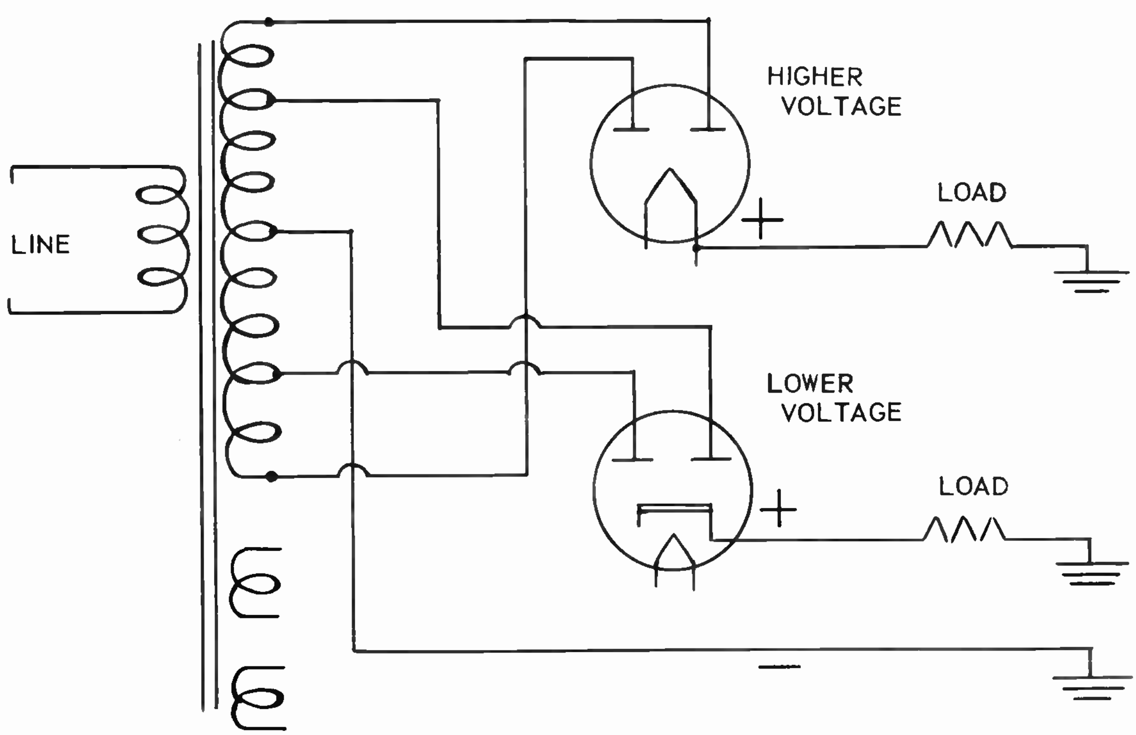


Fig. 20-10. A single secondary winding furnishes higher and lower voltages for two full-wave rectifiers.

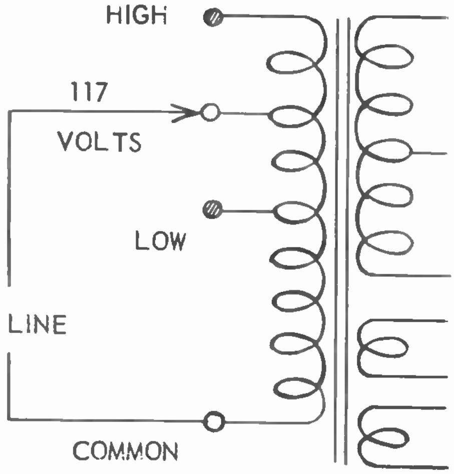


Fig. 20-11. A tapped primary winding.

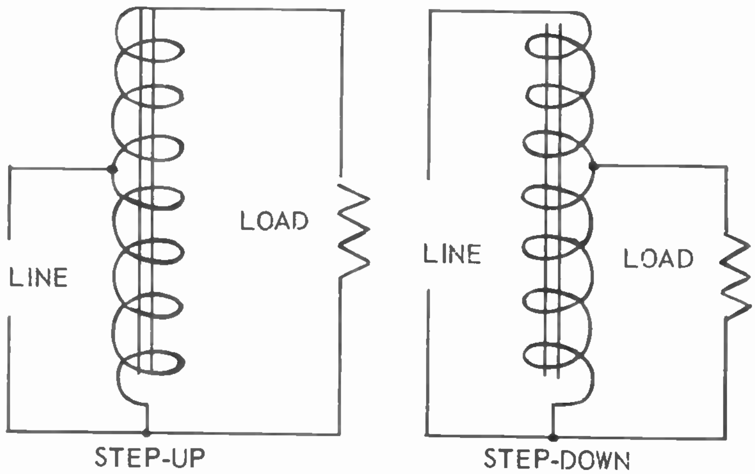


Fig. 20-12. Connections for an auto-transformer.

TAPPED WINDINGS. We have become fairly familiar with center-tapped secondary windings used with full-wave rectifiers, but sometimes there are other tapping arrangements for secondaries and there also are tapped primary windings on some power transformers, especially on replacement types.

2 In many power supplies for television there are two full-wave rectifier tubes, one operating at higher voltage than the other. Typical circuit connections are shown by Fig. 20-10. The plate winding of the power transformer has the usual center tap, which is the negative end of the output circuit. Between the outer ends of this winding and the center tap are two additional equally spaced taps. The outer ends of the plate winding are connected to the plates of the higher voltage rectifier, while the intermediate taps are connected to the plates of the lower voltage rectifier.

Because there are more winding turns between the outer ends than between the intermediate taps there is higher output voltage between the ends than between the intermediate taps, since volts per turn will be the same throughout the winding. Each rectifier supplies voltage and current for its separate load. Receiver tubes taking relatively high plate and screen voltages would form the load for the higher voltage unit, while tubes requiring somewhat lower voltages would form the load for the lower voltage rectifier. Filament and heater windings on the transformer of Fig. 20-10 are not shown connected to the tubes, this merely to simplify the diagram.

Connections to a tapped primary winding are shown by Fig. 20-11. The purpose of such tappings is to allow compensation for supply line voltages which are lower or higher than the one for which the power supply system is designed. A-c power line design voltage is standardized at 117 volts. In some localities the actual line voltage may be down around 105 to 110, while in other places the line voltage may be up around 125 or 130, and in still others the line voltage may sometimes be low and again may be high. Sometimes there are only two taps, one for normal line voltage and one for low line voltage. In some cases there are more than three taps.

In the diagram the power line is connected to the common lead at the bottom of the primary and to the middle one of the three taps at the upper end. This middle tap is used when the line voltage is between 115 and 120 volts. If line voltage is too high we make connection to the high tap, thus including more turns in

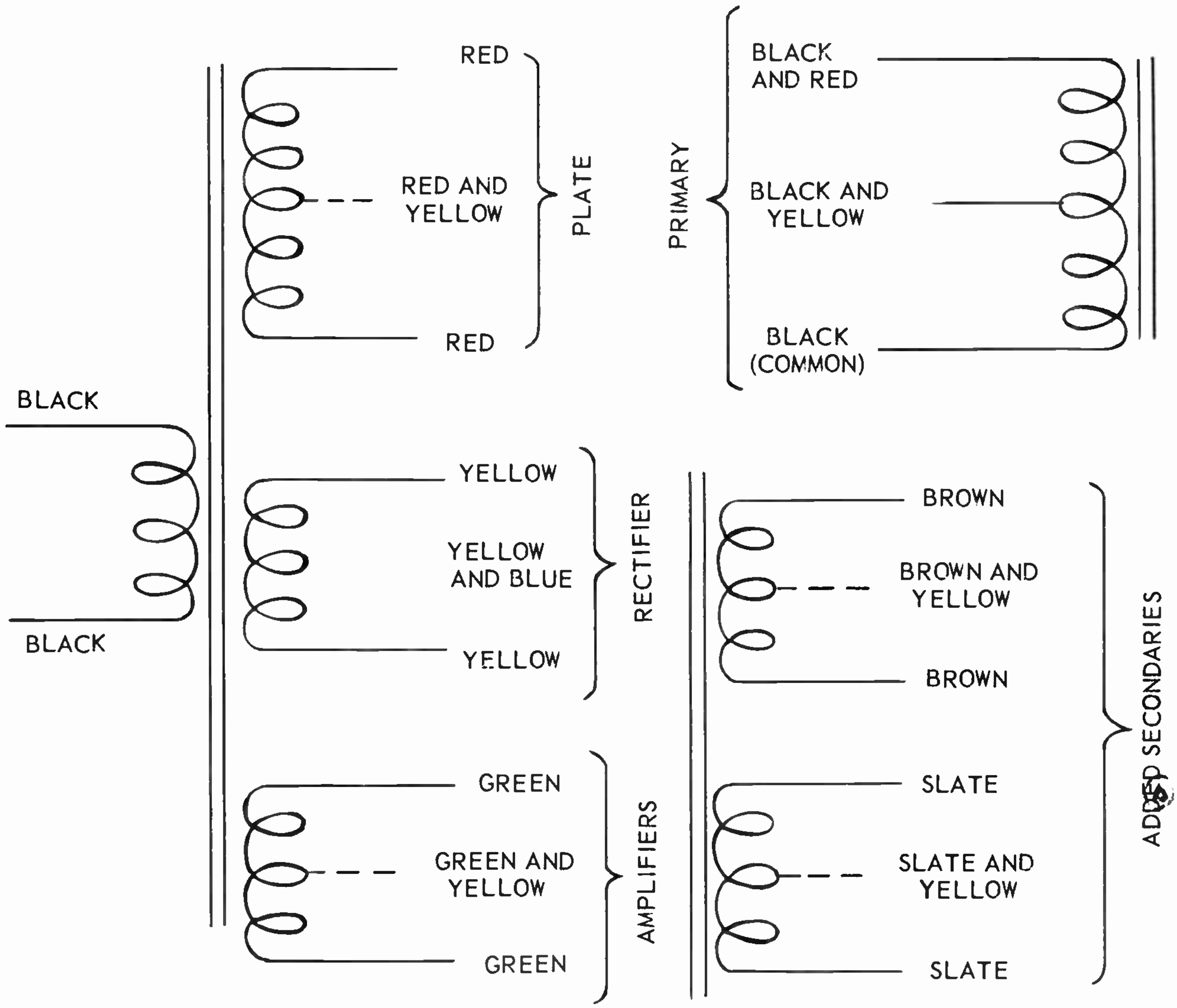


Fig. 20-13. Standard color coding for leads on power transformers.

the primary winding. With more turns there is less voltage per turn, not only in the primary but also in the secondaries, and secondary voltages are brought down where they belong in spite of the abnormally high line voltage. If line voltage is too low we connect to the low tap. Then the line voltage is divided among fewer turns, and voltage per turn is increased. This brings secondary voltages up to correct values, even with the low line voltage.

AUTO-TRANSFORMERS. Where voltage from the power line is permanently low or high the condition may be compensated for by using what is called an auto-transformer. As shown by Fig. 20-12, an auto-transformer has only a single continuous winding which serves as both primary and secondary, or has its two windings conductively connected together to form a continuous electric circuit.

The action of an auto-transformer is easily understood if we keep in mind that the voltage per turn is the same in all the turns and all the windings of any transformer. The left-hand diagram of the figure represents an auto-transformer connected for step-up of line voltage. Only part of the turns are across the line. This determines the voltage per turn. All the turns are across the load. Then in the greater number of winding turns across the load there must be more total voltage than across the smaller number across the line, and load voltage or secondary voltage will be higher than line voltage or primary voltage.

The right-hand diagram shows auto-transformer connections for step-down of line voltage. All the turns are across the line and only part of the turns are across the load. Then voltage applied to the load must be less than voltage from the line. There is no insulation between primary and secondary windings of an auto-transformer, and the load circuits are not insulated from the line. The load side of the auto-transformer may be tapped at two or more places to provide different output voltages which are higher or lower than line voltage. The line side may be tapped to allow using the transformer on line voltages higher or lower than standard.

Although auto-transformers most often are used for correction of line voltage, they sometimes are added to receivers which have been incorrectly designed and which need higher voltages than furnished by a standard line. These transformers do not take the place of the regular power transformers, but are connected between the line and the primary of the power transformer. Auto-transformers sometimes are used in the service shop to provide various voltages for testing, but the fact that apparatus being tested then is conductively connected to the power line makes these units generally undesirable for such purposes.

TRANSFORMER COLOR CODING. The leads of power transformers are supposed to have various colors of insulation which indicate the windings or taps to which connection is made. At the left in Fig. 20-13 are shown lead colors for a transformer having an untapped primary, a tapped plate winding at the top, and down below a rectifier filament winding and an amplifier heater winding. Some rectifier and amplifier windings have center taps, which then are color coded as shown. With no taps, only the end colors appear.

(a) At the upper right in Fig. 20-13 is the color coding for a tapped primary winding. One side of the line always is connected to the black (common) lead, and the other side is connected to whichever of the other leads allows desired voltage output. At the lower right are color codings for any additional secondary windings which may be used. Power transformers ordinarily have no identification for correct connections other than the color coding of leads. Remember where to find this diagram of the colors. Unfortunately it is necessary to remember also that not all manufacturers use this coding, and colors may not always mean what they are here shown as meaning.

AC-DC RECEIVERS. The abbreviation ac-dc means "alternating-current or direct-current", and when used to describe receivers it means that they will operate on power lines furnishing either alternating or direct current of suitable voltage. An ac-dc receiver cannot have a power transformer with insulated primary and secondary windings, for then the receiver would not operate on direct-current power. With smooth direct current there can be no varying magnetic fields and no induction, and there would be no transfer of power through a transformer – although the transformer windings would burn out. Ac-dc sets may be called also universal receivers or transformerless receivers. An ac-dc set must include a power rectifier in order that it may operate on a-c lines by rectifying the line voltage.

Fig. 20-14 is a circuit diagram for the high-voltage power supply portion of a small ac-dc receiver. The rectifier often is a type having two plates and two cathodes, with plates connected together and cathodes connected together. This type of rectifier and method of connection provide relatively small voltage drop in the rectifier, and preserve most of the line voltage for the receiver circuits. With 117 alternating line volts the direct voltage at the rectifier output usually is about 105 volts. The filter and the plate and screen circuits are represented in the diagram by a simple resistance for the load. Resistor *R*, in series with the rec-

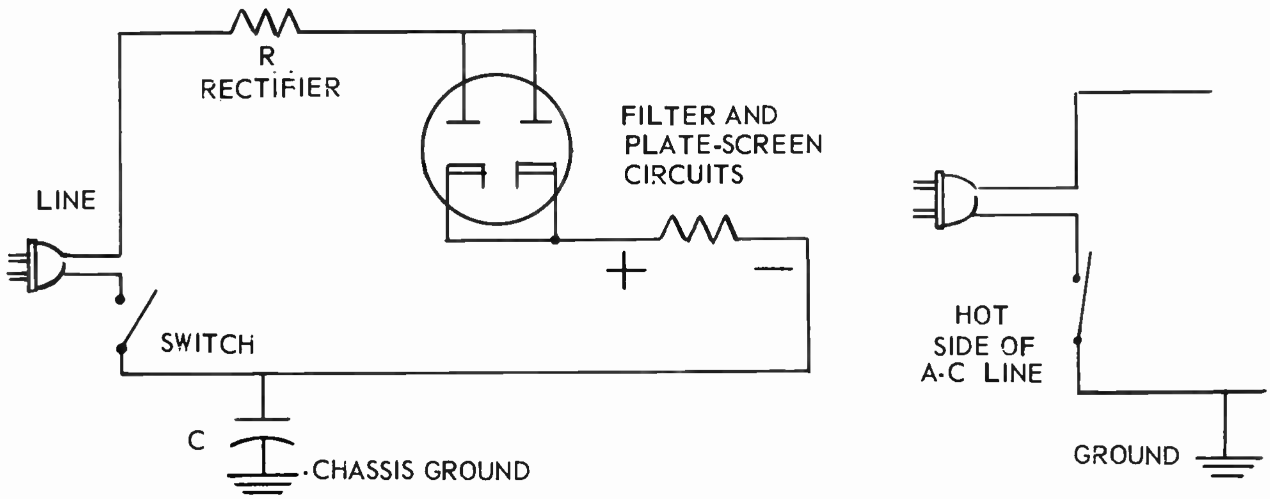


Fig. 20-14. Ac-dc power supply circuit with "insulated" chassis.

tifier, prevents excessive current when the line switch first is closed, also in case of a short circuit in the output load circuits. No heater connections are shown for the rectifier, because the rectifier heater nearly always is made part of a series heater string which includes all or part of the other tubes.

In some ac-dc receiver power supplies there is a regular half-wave rectifier tube with a single plate and single cathode, rather than the twin type of the diagram. In other sets of this class there is a half-wave selenium rectifier instead of the tube rectifier.

When the ac-dc set is connected to an a-c power line the rectifier delivers rectified direct voltage and current to the load. If the set is connected to a d-c power line the line plug must be inserted into the receptacle in the position that makes the rectifier plate or anode positive and the cathode negative. Then direct current will flow through the rectifier and to the load circuits. If the plug is inserted the wrong way around there can be no current through the rectifier. The receiver will not work until the plug is reversed, but no harm will result.

In the circuit of Fig. 20-14 there is no direct conductive connection from the rectifier or load circuits to "chassis ground", which is the metal of the chassis. There is a connection through capacitor C, which usually is of 0.05 mfd capacitance. This capacitor allows flow of r-f currents in circuits connected to chassis ground, but has reactance of more than 50,000 ohms at the 60-cycle line frequency. This high reactance would allow current of only about 2 milliamperes at a line voltage of 117.

Were there a direct connection to chassis ground, as in the sketch at the right, the power line could be short circuited with the plug inserted in one of the two possible positions. One side of every a-c line in all building wiring is required to be connected to ground, usually at a cold water pipe in the plumbing system. The ungrounded side of the building line is called the hot side. Were the plug inserted in the position connecting the hot side of the power line to the receiver chassis, and were the chassis to be in any way connected to the building ground, there would be a direct conductive connection from the hot side to the ground; ed side of the line. This would blow a fuse in the building circuit.

Not all ac-dc receivers have the power supply and the high-voltage circuits designed in such a way that the only connection from the negative side to ground is through a capacitor. Some, as shown by Fig. 20-15, have one side of the power line connected directly to chassis ground, which thus becomes the negative side of all plate, screen, and grid bias circuits. With this method of connection the receiver chassis must not be allowed to come in contact with any building ground unless precautions are taken to insert the line plug in a position that will not connect the hot side to ground.

3 *VOLTAGE MULTIPLIERS.* Voltage multiplier circuits furnish to the output load direct voltages which are higher than alternating line voltages, and do so without using step-up transformers. Consequently, the multiplier circuits are found in transformerless receivers. There are voltage doubler circuits furnishing direct output voltages approximately twice as high as applied alternating line voltage. There are voltage triplers furnishing approximately three times line voltage, also voltage quadruplers furnishing about four times line voltage, and in some applications (not radio or television receivers) there are multipliers delivering ten or more times the line voltage, all without transformers.

HALF-WAVE VOLTAGE DOUBLERS. Fig. 20-16 shows connections for a half-wave voltage doubler using two selenium rectifiers, *Da* and *Db*, and two capacitors, *Ca* and *Cb*. Diagram 1 shows part of what happens during half-cycles of alternating line voltage in which the upper side of the line is negative and the lower side positive. The remainder of the action during these half-cycles will be shown in diagram 3. Keep in mind that the direction of electron flow can be only against the arrowhead of the rectifier symbols, as shown by arrows alongside the conductors. Flow through rectifier *Da* charges capacitor *Ca* in the polarity marked. Charging raises the voltage of *Ca* to nearly the peak value of alternating line voltage. With line polarity now existing there can be no flow through rectifier *Db*, because the cathode of *Db* is positive and its anode is negative.

Diagram 2 shows electron flows during the alternate half-cycles of line voltage, in which the upper side of the line is positive and the lower side negative. With this reversed polarity there can be no electron flow through rectifier *Da*, but now there is flow through rectifier *Db* as shown by arrows. This electron flow from the line through *Db* not only goes through the load but also charges capacitor *Cb* in the marked polarity.

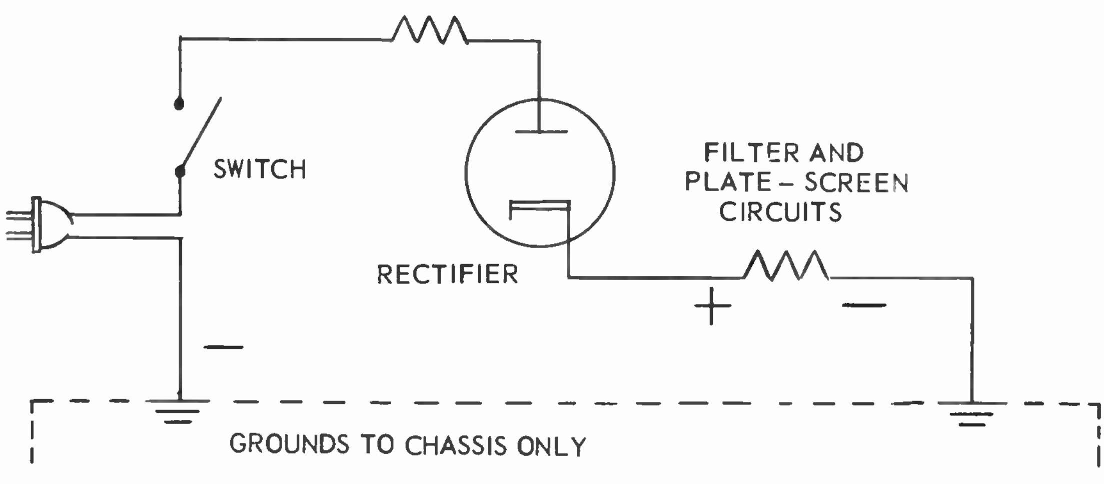


Fig. 20-15. Ac-dc power supply circuit with "hot" chassis.

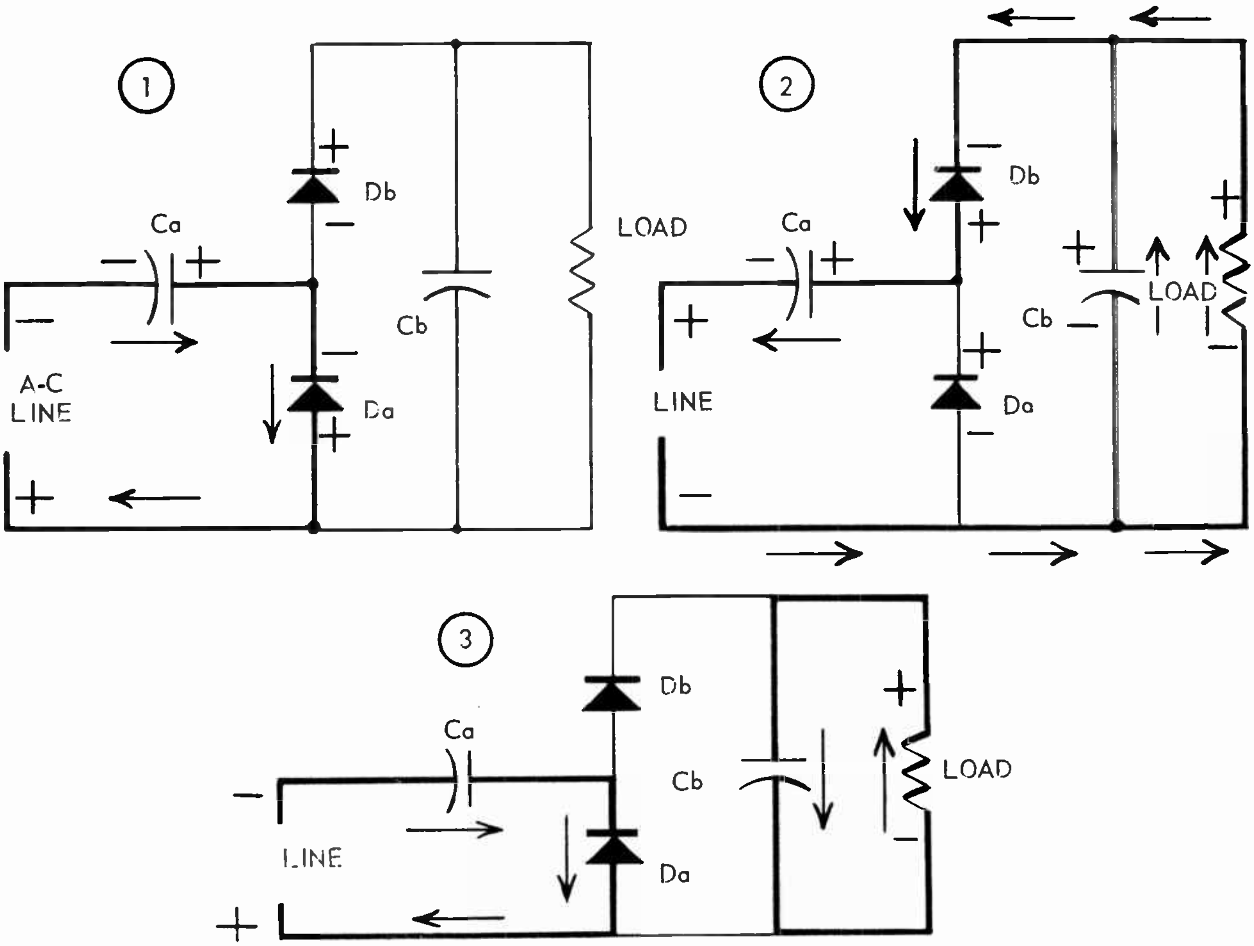


Fig. 20-16. The operation of a half-wave voltage doubler with selenium rectifiers.

Electron flow from the line now is going through the load, through rectifier *Db*, and through capacitor *Ca*, all in series. Capacitor *Ca* previously was charged to nearly peak line voltage. Now this voltage of *Ca* adds to the line voltage, since the two are in series, with the result that voltage across the load is approximately twice as high as the alternating line voltage. In this way we obtain voltage doubling.

In diagram 1 we found no line current through the load. But during the half-cycle represented by diagram 2 there was charging of capacitor *Cb*. Now, in diagram 3, capacitor *Cb* discharges through the load. This diagram shows the complete action during half-cycles in which the upper end of the line is negative and the lower end positive; it completes the action shown in part by diagram 1. Electron flow through the load is in the same direction during both half-cycles of applied alternating voltage. The direct electron flow or current in the load is not smooth or uniform, but has a large alternating component. Current is maximum at about the middle of the half-cycle shown by diagram 2, and is minimum near the end of the half-cycle shown by diagrams 1 and 3. The alternating component will be smoothed out in the filter.

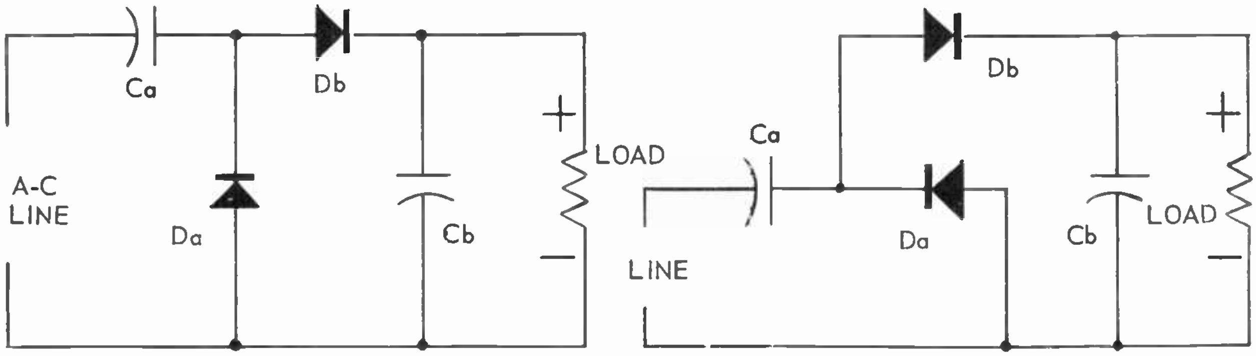


Fig. 20-17. Other ways of drawing the half-wave voltage doubler circuit.

You will find circuit diagrams for voltage doublers and other multipliers drawn in many ways, but all will have the same essential circuit connections. Two variations of the doubler circuit are shown by Fig. 20-17. The circuit is exactly the same as in Fig. 20-16, as you will discover by checking the connections between parts which are similarly lettered in all the diagrams.

Rectifier tubes instead of selenium rectifiers often are used for voltage multipliers. Fig. 20-18 shows connections for a half-wave doubler using rectifier tubes designed especially for this purpose. The special feature is a separate cathode to go with each of the plates. Such tubes include types 25Z6, 50X6, 50Y7, 117Z6, and others.

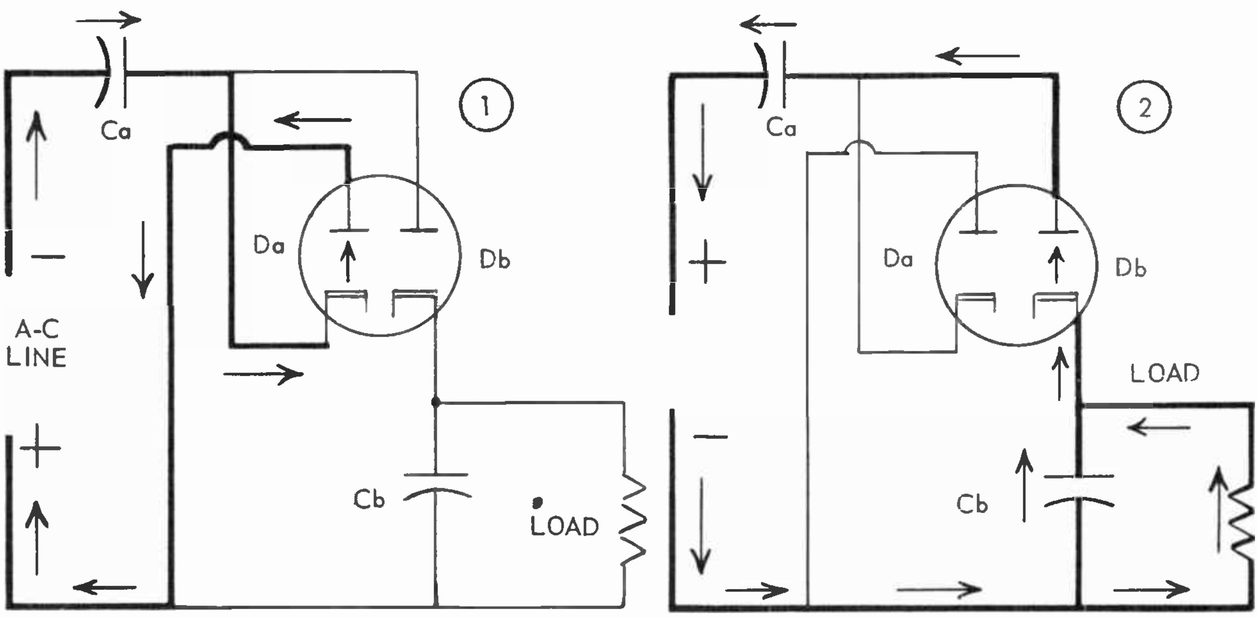


Fig. 20-18. Operation of a half-wave voltage doubler with a tube rectifier.

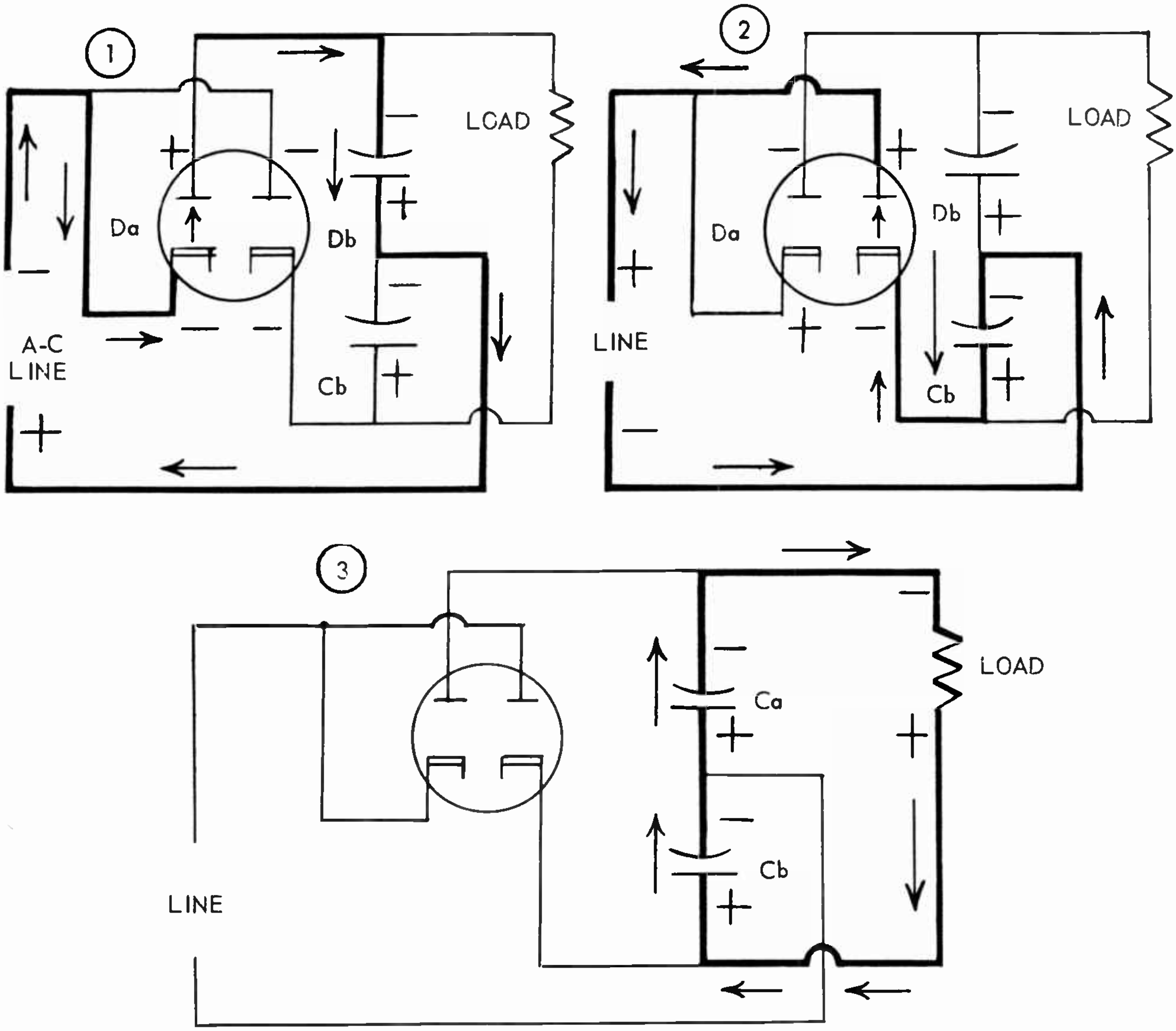


Fig. 20-19. Operation of a full-wave voltage doubler with a tube rectifier.

Except for the use of tube rectifiers instead of selenium units, the circuit of Fig. 20-18 is exactly the same as that of Fig. 20-16. If you go back and read the explanation for diagrams 1 and 2 of the circuit with selenium rectifiers, but look at the diagram having tube rectifiers, you will have an explanation of the half-wave doubler using tube rectifiers. Although no third diagram has been included in Fig. 20-18, the earlier explanation still applies. That is, during the action shown by diagram 1 there is also discharge capacitor *Cb* through the load, this capacitor having been charged during the action shown in diagram 2.

FULL-WAVE VOLTAGE DOUBLERS. Fig. 20-19 shows circuit connections for a full-wave type of voltage doubler employing a twin rectifier tube. Diagram 1 shows electron flow during half-cycles of alternating line voltage in which the upper end of the line is negative and the lower end positive. The cathode of rectifier *Da* now is negative and its plate is positive. Electron flow from the negative side of the line goes through this rectifier and charges capacitor *Ca* in the polarity marked, then returns to the positive side of the line.

In diagram 2 the line polarity has reversed, with the upper end now positive and the lower end negative. During this half-cycle, and all others like it, the cathode of rectifier *Da* is positive and its plate is negative, so there can be no flow through this rectifier. But now the cathode of rectifier *Db* is negative and its plate is positive. Consequently there is electron flow from the negative side of the line through capacitor *Cb*, thus charging this capacitor in the marked polarity, and flow is from the capacitor through rectifier *Db* back to the positive side of the line. The two capacitors are charged alternately, one during each half-cycle of alternating line voltage.

These two capacitors which are alternately charged are in series with each other and the load, as emphasized by diagram 3. The capacitor voltages add together, because their charging polarities are in the same directions. Then this double voltage is applied across the load to provide the doubling action. Electron flow from the capacitors through the load is shown by arrows in diagram 3. Discharge of the capacitors through the load continues at all times, while the charges are intermittently restored by the charging pulses occuring during actions shown in diagrams 1 and 2.

Full-wave doublers often employ selenium rectifiers instead of the tube rectifiers shown by Fig. 20-19. It would be excellent practice for you to make a pencil sketch of the circuit diagram from this figure, then erase the tube rectifier symbol and insert in its place the symbols for two selenium rectifiers.

VOLTAGE TRIPLERS. A voltage tripler consists of a half-wave doubler in series with a straight half-wave rectifier. The double voltage of the doubler adds itself to the single voltage of the straight rectifier, which delivers approximately line voltage by itself, and the result is nearly three times the line voltage.

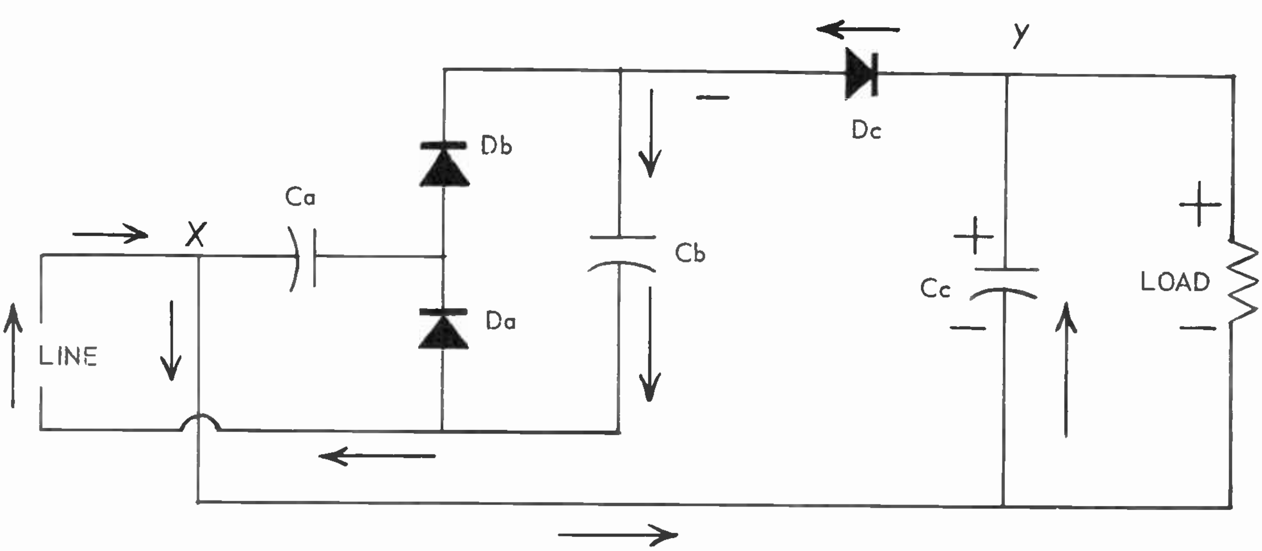


Fig. 20-20. A voltage tripler with selenium rectifiers.

A tripler circuit using three selenium rectifiers and three capacitors is shown by Fig. 20-20. The half-wave doubler portion, drawn with heavy lines, has purposely been made exactly the same as the circuit shown back in Fig. 20-16. We know the output voltage of this doubler circuit is approximately twice the applied line voltage. Instead of showing the output of the doubler circuit connected to a load, as in Fig. 20-16, it now is shown connected effectively in series with a simple half-wave rectifier circuit which includes the additional rectifier *Dc* and the additional capacitor *Cc*.

When the upper side of the line is negative there is electron flow from point *x* through the added connection to the negative side of capacitor *Cc* and the negative side of the load. The positive sides of capacitor *Cc* and the load are connected to the positive side of the doubler circuit at point *y*. Thus the load is subjected to the sum of the double voltage at point *y* and the line voltage from point *x* as rectified by added rectifier *Dc*. Capacitor *Cc* charges to approximately three times line voltage, which becomes the voltage applied across the load.

Voltage tripler circuits may be drawn in so many different ways as to become rather difficult to identify. The easiest identification is to note that the three rectifiers all are placed to carry electron flow in the same direction if you consider them to be in series. Commencing at the bottom of rectifier *Da* we may go in the direction of arrowheads in the rectifier symbols through *Db* and then through *Dc*. Another way of drawing the same tripler circuit is shown in Fig. 20-21. Here it becomes quite clear that the three rectifiers act in the same direction. Tripler circuits may be made up with a twin rectifier tube for units *Da* and *Db*, with a separate half-wave rectifier tube for unit *Dc*. Triplers sometimes include both tube and selenium rectifiers in various combinations.

Voltage quadruplers seldom are used in radio and television receivers. A quadrupler consists of two full-wave doublers with the output of the first one connected as input to the second one. Then the output of the second doubler is approximately four times the line voltage.

A more common way of obtaining four times line voltage is shown by Fig. 20-22. Here are two half-wave voltage doublers one above the other and connected in series with each other. To make it easier to follow the action, the rectifier and capacitors of each doubler circuit are lettered the same as in Fig. 20-16. Both doubler inputs connect through capacitors *Ca* to one side of the line. The negative end of the top doubler

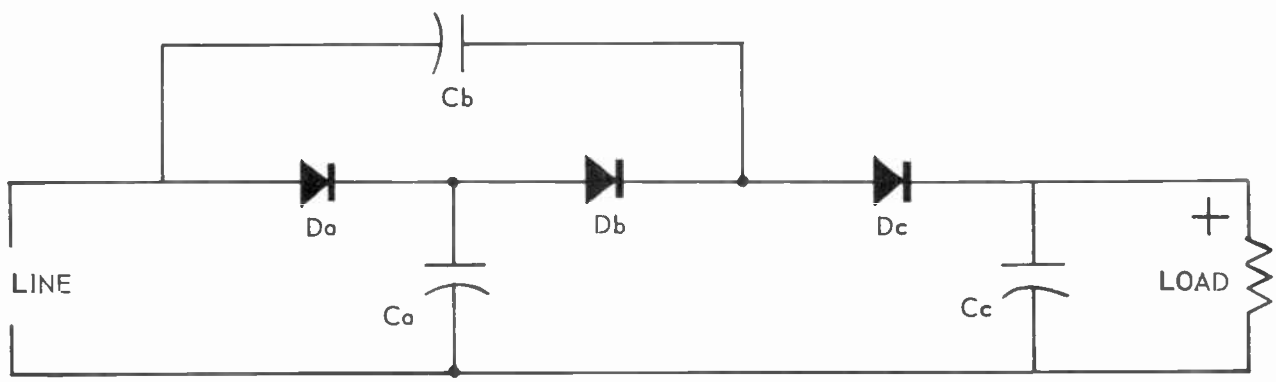


Fig. 20-21. Another way of drawing the voltage tripler circuit.

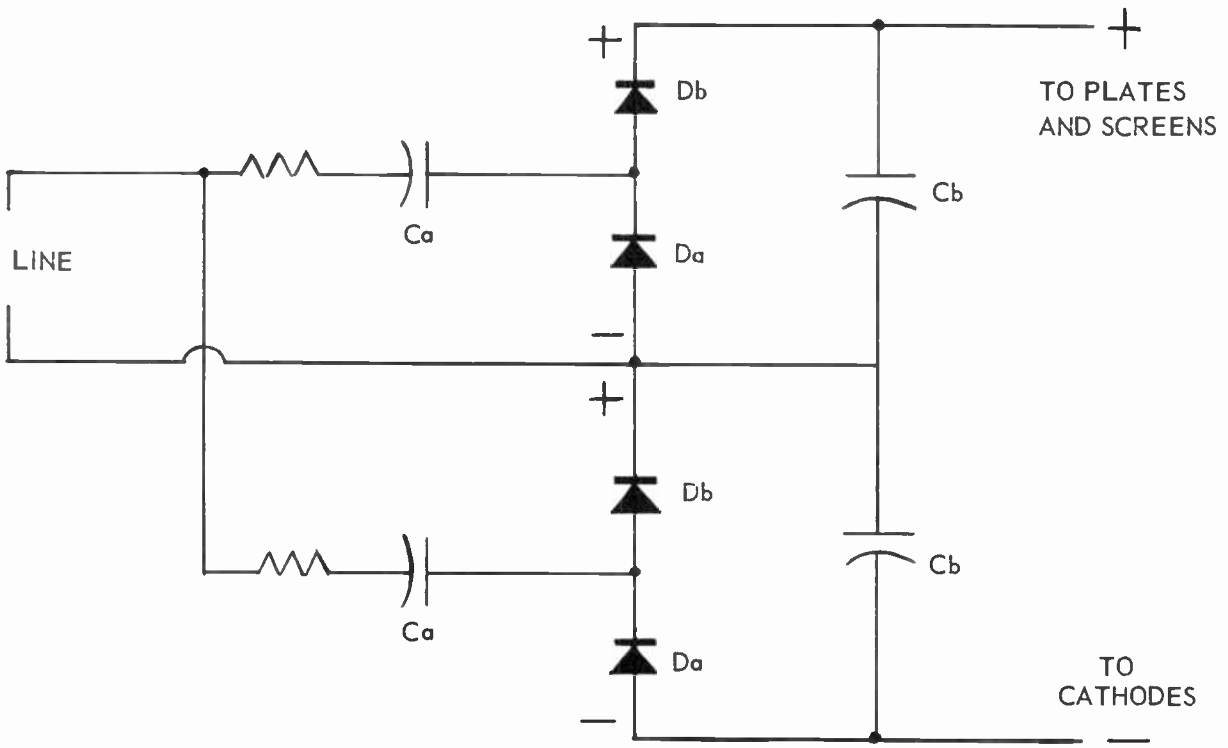


Fig. 20-22. Obtaining approximately four times line voltage from two voltage doublers.

and the positive end of the bottom one are connected together and to the other side of the line. Then the output of both doublers will be from positive on the top one to negative on the bottom one, and will be twice the output of one doubler. The positive of the output connects through a filter, not shown here, to the plates and screens of receiver tubes. The negative of the output connects through another filter to the cathodes of the receiver tubes. In this way the entire output voltage is applied between plates and cathodes and between screens and cathodes.

TELEVISION

LESSON NO. 21

FILTERS FOR POWER SUPPLIES

"Training Your Mind"

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

COMMERCIAL TRADES INSTITUTE

Training Your Mind... C.T.I. Perfecting Your Skill

HIGH SCHOOL DIVISION Chicago, Ill.

LESSON NO. 21

FILTERS FOR POWER SUPPLIES

Now we are ready to change the pulsating direct current and voltage coming from the power supply rectifier into smooth direct voltage and current by means of a filter system. First we shall examine the output voltage of the full-wave rectifier of Fig. 21-1. Connected across the rectifier output is a resistance to represent the load. The actual load in a receiver would be plate, screen, and grid biasing circuits of various tubes. Here we have as yet no filter system, only a rectifier connected directly to a load.

The output voltage of the rectifier, as it appears across the load, is shown at the right. This output is a direct voltage with a very large alternating component. In a test on certain equipment the direct potential difference measured 280 volts and the alternating component about 175 volts. The pulse frequency is 120 cycles, because we have 60-cycle power line frequency and a full-wave rectifier.

To commence construction of the filter we shall connect a capacitor across the rectifier output and across the load, as in Fig. 21-2. The voltage which now appears across the load, and also across the capacitor, is shown by the full-line curve at the right. In broken lines are shown the relations of the earlier voltage pulses to this new voltage curve.

If we assume the action to begin with rise of voltage during one of the pulses from the rectifier, the capacitor will be charged along the voltage line from *a* to *b*. Maximum voltage of charge will be equal to peak pulse voltage. When voltage from the rectifier commences to drop, the capacitor no longer receives additional charge but instead commences to discharge through the load resistance.

With 20 microfarads capacitance and 5,000 ohms (0.005 megohm) resistance the discharge time constant is 0.1 second. During the 1/120 second interval before another charging pulse occurs there is only about 1/12 of one time constant, and the capacitor has time to lose only about 8 per cent of its peak charge and voltage. This discharge and drop of capacitor voltage are shown along the line from *b* to *c*. At instant *c* the *rising* voltage in the next pulse from the rectifier equals the falling voltage of the capacitor. Then there is recharging from *c* to *d*. These discharges and recharges continue.

*　　　*　　　*

NOTE: These tests were carried out with a typical power supply of the following description. Transformer, 350 alternating volts each side of center tap. Rectifier, 5Y3-G, full-wave. Capacitors, electrolytic, 20 microfarads each. Choke, 10 henrys inductance, 160 ohms resistance. Load resistance 5,000 ohms. Load current from completed filter 72 milliamperes. Waveforms observed with service oscilloscope.

The fluctuations of voltage across the capacitor and across the load resistance now are as shown by the upper curve which is drawn with a full line. We still have an alternating component mixed with the direct

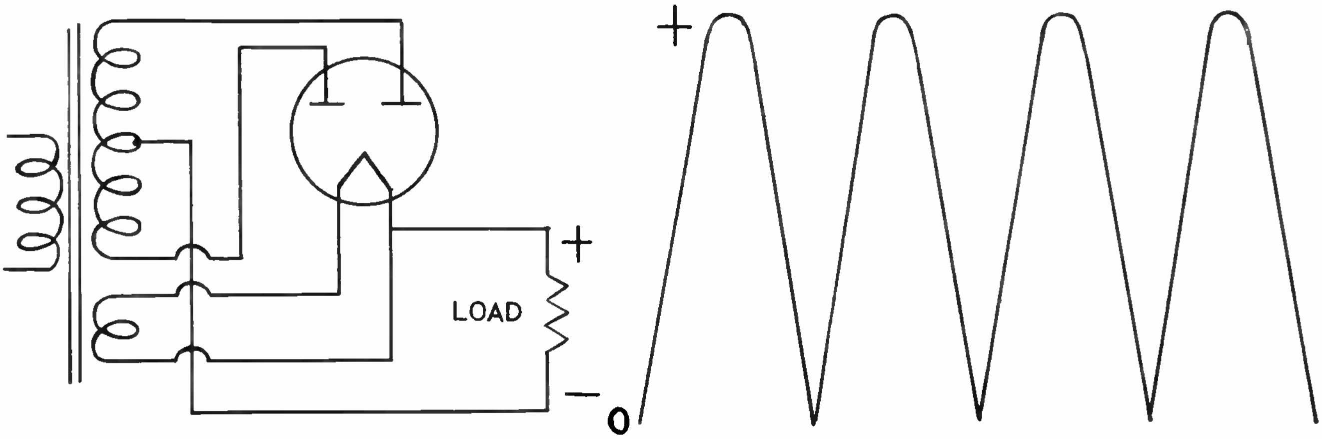


Fig. 21-1. Output voltage pulses from a full-wave rectifier.

voltage, but the alternating component has become relatively small. We are coming closer to smooth direct current and voltage.

The top curve shows alternating voltage which would remain were the direct voltage removed. This wavy line of alternating voltage looks something like ripples on the surface of water, and we call the alternating voltage a ripple voltage. If any appreciable ripple voltage should get to the plate and screen circuits of sound amplifier tubes it would cause audible hum at the ripple frequency. Were the ripple to get into television video and sweep circuits it could cause jumpy pictures with wavy edges and other undesirable effects.

Supposing we were to use more capacitance, what then? The greater capacitance would hold a greater charge, and the capacitor voltage would suffer less drop between charging peaks. That would mean less ripple effect, or smaller amplitude of ripple voltage. It would mean also a somewhat higher average direct voltage at the output, because ripples would not extend so far down. Conversely, smaller capacitance would mean a greater ripple voltage and a lower average direct voltage.

Ripple voltage at the output of the completed filter will be inversely proportional to the capacitance of the capacitor which is connected across the output of the rectifier. Doubling this capacitance will cut the ripple voltage amplitude in half. Halving the capacitance will double the amplitude of the ripple. This will be true no matter what other parts we add during completion of the filter system.

2

What about load resistance and load current? Supposing we were to use less than 5,000 ohms for the load, what would happen? The time constant would be shorter, and there would be longer discharges between peaks of recharging. There would be greater ripple voltage, also lower average direct voltage across the load. The smaller load resistance would allow greater load current, therefore we may say that greater load current or output current will increase the ripple voltage. The greater current will increase the hum in sound reproduction and may make bad pictures – provided enough of the ripple gets through to the amplifiers. Naturally, were we to use more load resistance, and have less load current, there would be smaller ripple voltage and higher average direct voltage.

As a matter of fact, ripple voltage is inversely proportional to resistance in whatever load is connected across the output of a completed filter. Doubling the load resistance will cut the ripple voltage in half, while half as much load resistance will allow double the ripple voltage when other things remain unchanged.

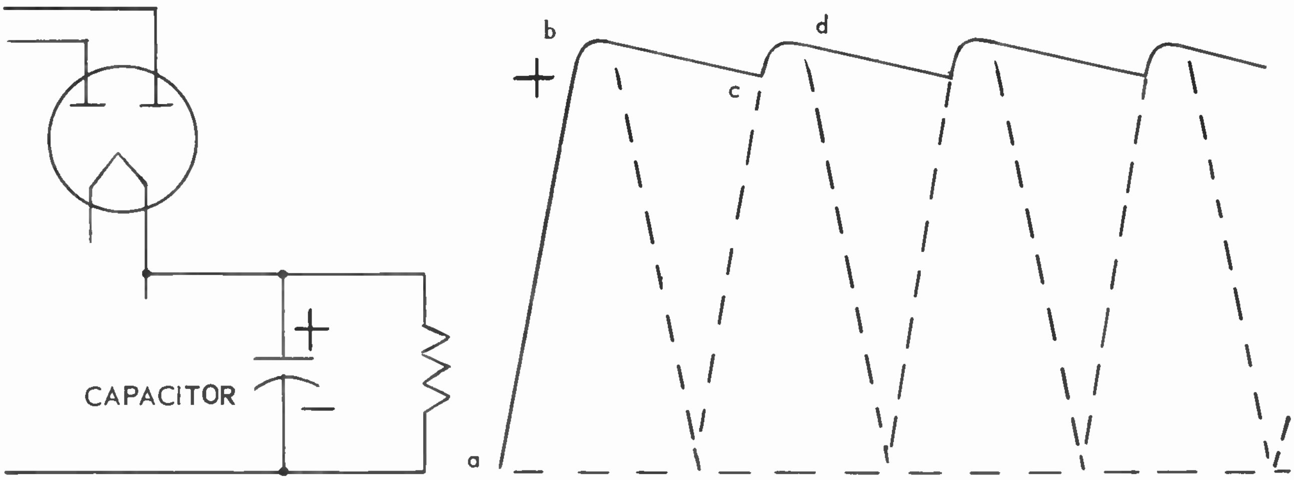


Fig. 21-2. Ripple voltage at a capacitor connected across a full-wave rectifier.

Ripple voltage most often is expressed not in volts, but as a percentage of the d-c output voltage from the completed filter system. For example, were the d-c output 250 volts and the alternating ripple voltage 5 volts effective or rms, we should have a 2% ripple voltage, because 5 is 2% of 250. In high grade sound amplifiers the ripple may be as little as one-fourth of one per cent.

HALF-WAVE RECTIFIERS. We have looked at what happens with a full-wave rectifier. Now, in Fig. 21-3, let's look at the effects of half-wave rectification. The interval between capacitor charging pulses will be twice as long as with full-wave rectification, and the capacitor will have twice as long in which to discharge and lose voltage. There will not be quite twice the drop of voltage, because capacitors do not discharge at a uniform rate through resistance – the discharge slows down as time increases. But with the

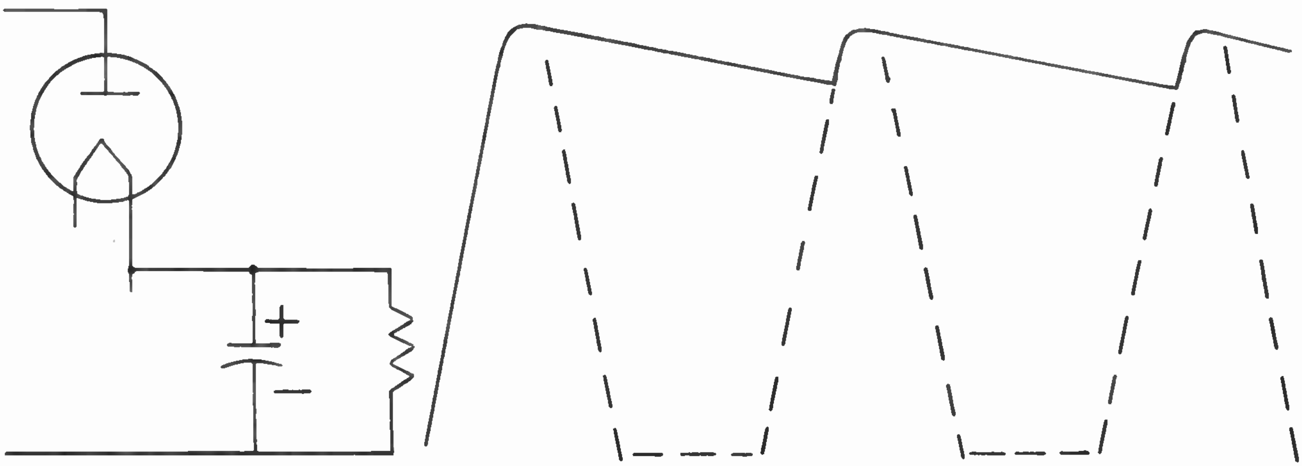


Fig. 21-3. Ripple voltage at a capacitor connected across a half-wave rectifier.

same time constant as before, meaning the same capacitance and resistance, the 1/60 second intervals with a half-wave rectifier will allow loss of about 14 per cent of peak voltage instead of about 8 per cent.

It follows that using the same capacitance and same load resistance on a half-wave rectifier as on a full-wave type will allow much more ripple voltage, and average direct output voltage will drop. In order to have no more ripple and no less output voltage it would be necessary to use nearly twice as much capacitance with half-wave rectification as with full-wave, assuming there is no change of load resistance. Of course, more load resistance and less load current would help to lessen the ripple and raise the direct voltage. You will find half-wave rectifiers in rather common use where load currents are relatively small.

Pulse frequency from a half-wave rectifier is 60 cycles per second, whereas with a full-wave rectifier it is 120 cycles per second on a 60-cycle power line. Ripple output frequencies from filters connected to the two kinds of rectifiers likewise will be 60 and 120 cycles per second. With the same degree of filtering the ripple voltage at 60 cycles will be reduced only about half as much as at 120 cycles. In spite of this, the greater low-frequency ripple voltage in a sound amplifier may be no more objectionable or may be even less objectionable than the lesser ripple voltage at the higher frequency. The reason is that a 60-cycle sound, must be about twice as strong as one at 120 cycles to give the same loudness effect on the average human ear. In a television amplifier, however, the troublesome effects will be proportional to strength of ripple voltage.

FILTER CHOKES. The sharp peaks of ripple voltage shown by Figs. 21-2 and 21-3 would produce sound effects worse than might be expected from the voltage value of this alternating component. Instead of a rather soft hum such as results from a low-frequency sine wave voltage there would be more of a buzzing sound at the ripple frequency. Ill effects on television pictures would be emphasized by the sudden changes of voltage.

To prevent these sudden changes in the ripple voltage we shall add a choke to our filter, as shown by Fig. 21-4. A choke such as used in power supply filters is an inductor with an iron core. Construction of the core is somewhat like that in a power transformer, but there is only a single continuous winding.

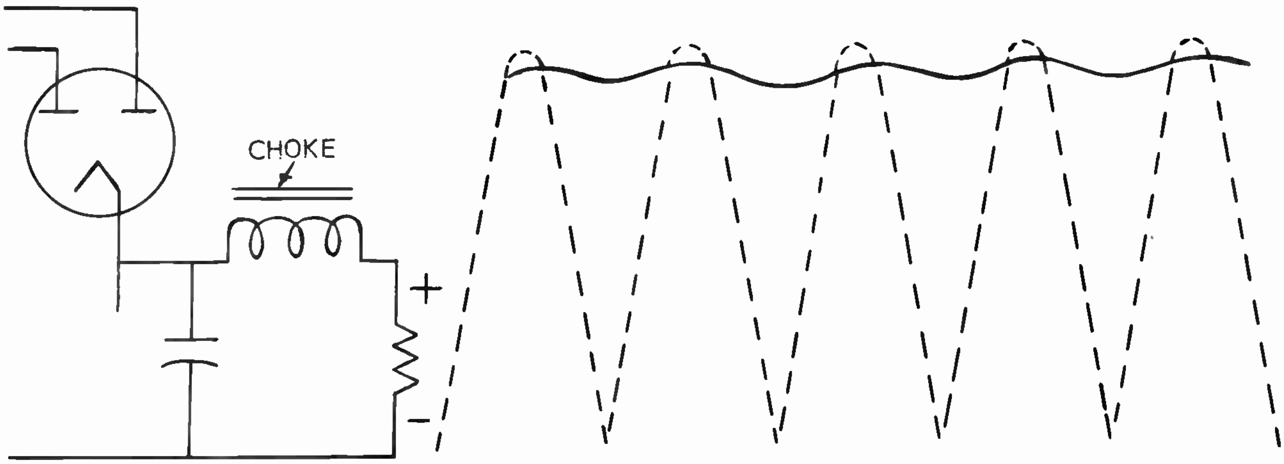


Fig. 21-4. How a choke smooths out the peaks of ripple voltage.

Changes of voltage across the filter capacitor are applied to the choke and the load resistance, since these latter two are connected across the capacitor. The changes of voltage cause changes of current and of magnetic field which induce counter-emf's in the choke. These counter-emf's oppose the voltage which is causing the changes. The more sudden the change of voltage applied to the choke, and the faster the current tries to rise, the greater is the opposition to this rise. The result, so far as load voltage or ripple voltage is concerned, is shown at the right in Fig. 21-4. Here we have a relatively smooth or gradually changing ripple voltage rather than one having such abrupt changes as occur with no choke.

Ripple voltage decreases as capacitance increases. With all else unchanged, 40 microfarads capacity will allow only half as much ripple voltage as 20 microfarads, but 10 microfarads will allow twice as much ripple voltage as 20 microfarads.

Since ripple voltage is inversely proportional to inductance, also inversely proportional to capacitance in the filter, the ripple voltage must be inversely proportional to inductance multiplied by capacitance. This being the case, we may have the same reduction of ripple voltage by using large capacitance with small inductance as by using small capacitance with large inductance. Practically speaking, 20 microfarads capacitance and 10 henrys inductance, whose product is 20 times 10 or 200, will be as effective as 10 microfarads capacitance and 20 henrys inductance, whose product again is 200.

The ratings of a filter choke include three values. One is the full-load direct current in milliamperes which the choke is supposed to carry. A second is inductance in henrys when the choke is carrying its rated direct current. Third is the resistance of the winding, in ohms. Resistances of various models of chokes range from 100 to 600 ohms. Low resistance is desirable because there is loss of voltage in the winding due to forcing current through its resistance. Consequently, the direct voltage at the output end of the choke is less than at its input end.

5 Actual inductances of commercially available filter chokes range from 5 to as much as 20 henrys when rated current is flowing. Inductance measured with no current will be two to three times as great. Choke inductance is caused to remain fairly constant with ordinary variations of current by placing a gap in the iron core. This means that, at some point in the magnetic circuit formed by core iron, the iron is not continuous. The gap is only a few thousandths of an inch. It is filled in with hard fibre or almost any material that is non-magnetic, that is not iron or steel. Although filled with solid non-magnetic material, the space is called an air gap. The gap reduces maximum inductance of the choke, but maintains a nearly constant inductance.

A certain choke with an air gap of 18 thousandths of an inch decreased its actual inductance from 10.7 to 10.3 henrys when current was increased from 10 to 30 milliamperes. With the gap closed, or with the core "butted," and with the same change of current, the inductance decreased from 25 to 10 henrys.

SECOND CAPACITOR. To complete the filter system we now add another capacitor beyond the choke, as in Fig. 21-5. To this capacitor comes the direct voltage and the ripple voltage from the choke. If this added capacitance is large enough to hold most of its charge during intervals between peaks of ripple voltage, there will be additional reduction of ripple. The action will be like that explained in connection with Fig. 21-2, although now the incoming alternations are only those of the ripple voltage rather than the pulses from the rectifier. In this diagram the two filter capacitors are connected together on their negative sides. This would indicate a dual electrolytic capacitor with a common negative terminal.

An important purpose of the second filter capacitor is to act as a reservoir of electrons which will care for sudden demands for more current in the load. There are many picture signals, sync signals, and audio signals having sudden changes of voltage. The necessary changes of plate currents would be delayed if they had to come through the filter choke, and the amplifiers would not give instant response to the signal voltages. But the second filter capacitor stores a large charge or large quantity of electrons, and is capable of delivering great quantities to the load without excessive drop of voltage at the capacitor and across the load. Thus there will be relatively small changes of voltages for plates and screens, even with considerable fluctuations of their currents.

CAPACITOR RATINGS. The first filter capacitor, connected across the rectifier, must withstand peak voltage coming from the rectifier. This peak voltage would be difficult to measure accurately with usual test equipment, but we may arrive at the approximate value by considering the relations between peak pulse voltage and direct output voltage from the filter system.

At *A* in Fig. 21-6 is represented alternating voltage across the secondary winding of a power transformer. Positive and negative alternations are rectified by a full-wave rectifier and we have the voltage pulses at B, which are the same as those in Fig. 21-1. The peak pulse voltage is lower than peak voltage from the transformer because there is considerable voltage drop in the rectifier tube. The effect of the first filter capacitor is to change the pulses to the ripple shown at *C*, which is like that of Fig. 21-2 and has the same maximum voltage as the rectified peaks. Average voltage of the ripple wave is shown at *D*. This is the average direct voltage from the first filter capacitor.

There is some voltage drop in the filter choke, so the final direct output voltage from the entire filter will be possibly 10 to 20 volts lower than from the first capacitor. This final direct output voltage is not a whole lot lower than the peak voltages at *B* and *C*. Therefore, by using a first filter capacitor having a d-c working voltage rating at least 20 per cent higher than actual direct output voltage from the filter we have a safe rating.

The second filter capacitor, which is across the load, is normally subject to somewhat lower maximum voltage because of the drop in the filter choke. However, were the load to be disconnected, as by some accidental open circuit, there would be no current flowing through the resistance of the choke, no voltage drop in these parts, and voltage on the second capacitor would be the same as on the first one. For this reason, both filter capacitors usually have the same voltage rating.

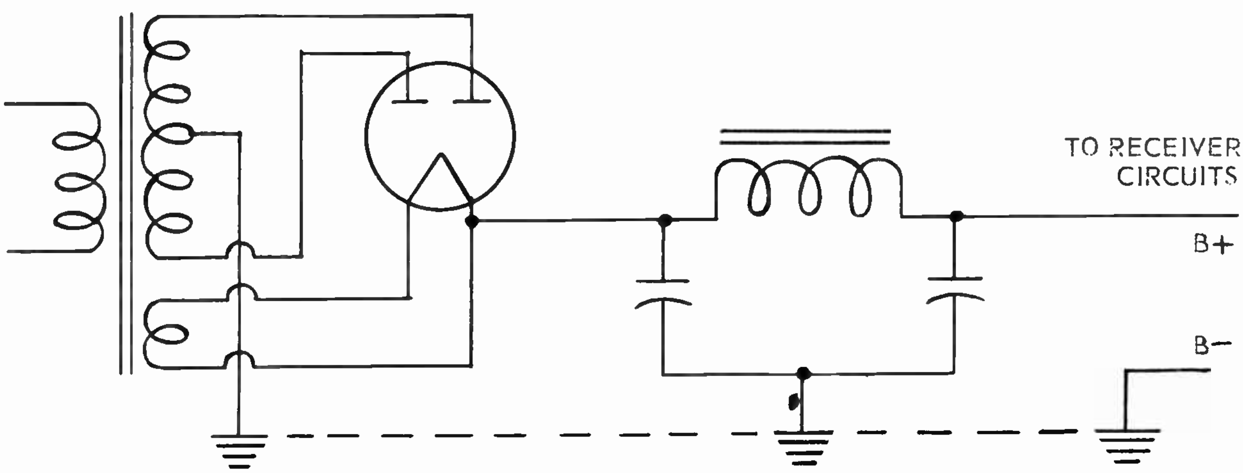


Fig. 21-5. Complete filter system with two capacitors and one choke.

RESISTOR-CAPACITOR FILTERS. The filters which have been examined may be classed as choke-capacitor types, because they are made up of chokes and capacitors. In another type, called a resistor-capacitor filter, the choke is replaced by a resistor.

Fig. 21-7 shows resistor-capacitor filters connected to a half-wave rectifier and to a full-wave rectifier. Action of both capacitors is same as in choke-capacitor filters. The resistor does not oppose *changes* of current, as does a choke, but it introduces a time constant effect which slows the discharge of the first filter capacitor. This helps retain the capacitor charge between pulses from the rectifier, and the effect on ripple voltage is similar to that shown back in Fig. 21-2 where we had only a resistance load following the capacitor.

Ripple voltage is reduced by increasing the filter resistance. That is, 1,000 ohms resistance allows half as much ripple voltage as 500 ohms. Since this is true, we may say that ripple voltage is inversely proportional to the product of resistance and capacitance, just as in a choke-capacitor filter the ripple voltage is inversely proportional to the product of inductance and capacitance.

There is a limit to reduction of ripple by increase of filter resistance, because the entire load current goes through this resistance and there is considerable drop in voltage as well as loss of power in heating if the resistance is made very great. Resistor-capacitor filters are used where load currents are rather small. By using large filter capacitances with moderate values of filter resistance there is satisfactory reduction of ripple.

Filter resistors in television receiver power supplies may have resistances between 150 and 1,000 ohms, with filter capacitances of 20 to 40 microfarads or sometimes more. In sound radio receivers of small and

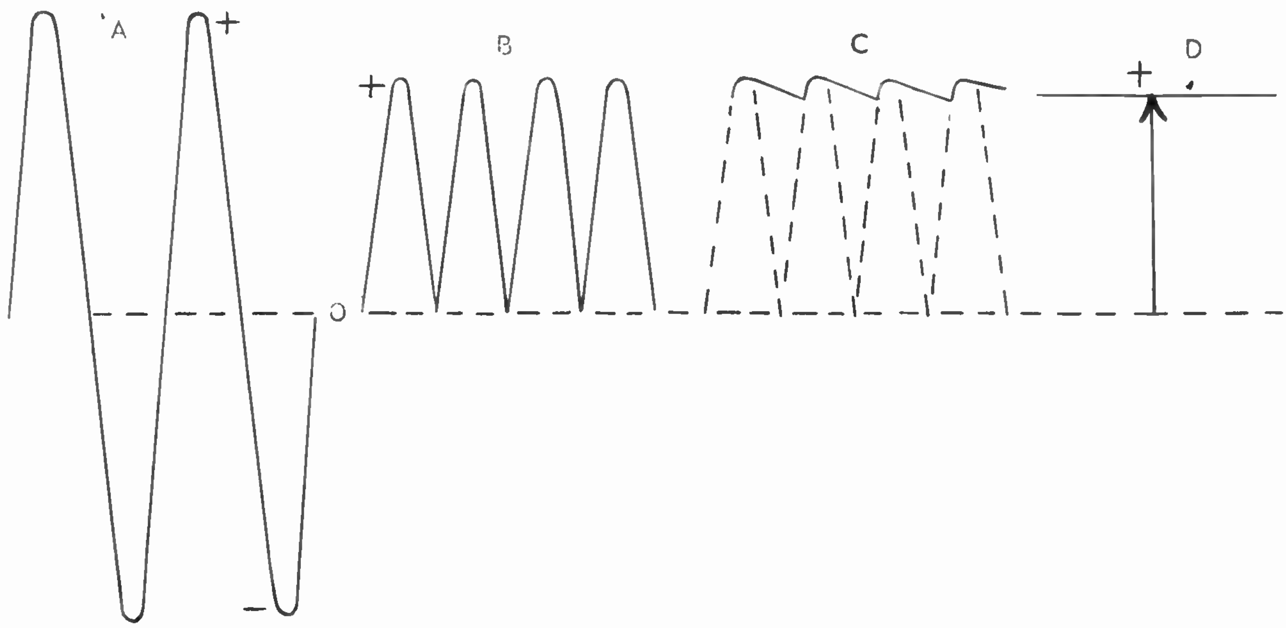


Fig. 21-6. Changes of voltage between the power transformer and the filter output.

COMMERCIAL TRADES INSTITUTE

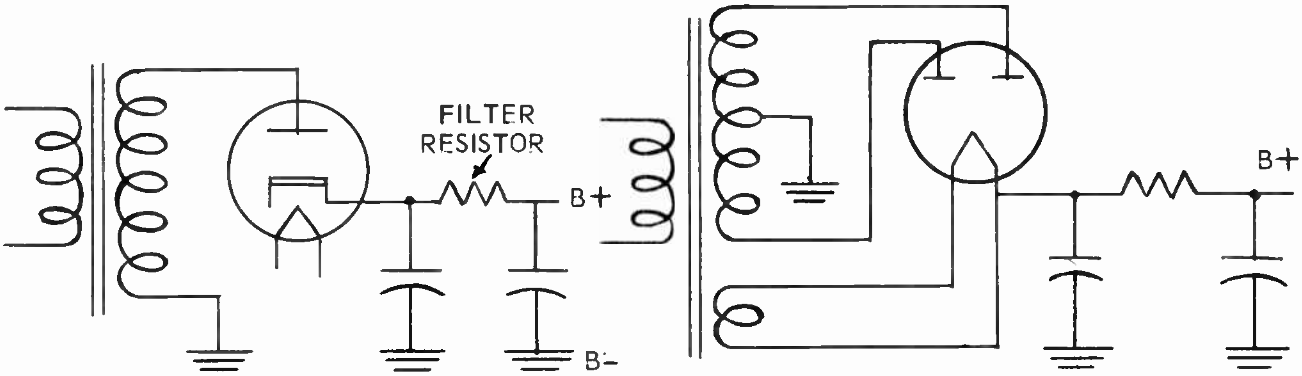


Fig. 21-7. Connections of resistor-capacitor filters.

medium size the filter resistance may be on the order of 1,500 to 4,000 ohms, with capacitance of 40 to 80 microfarads in most cases.

With a resistor-capacitor filter the voltage waveform at the rectifier output and at the first capacitor is like that of Figs. 21-2 and 21-3. At the filter resistor output and at the second capacitor the waveform is somewhat like that of Fig. 21-4, except that the dips are sharper and narrower. The filter resistor has a decided effect in smoothing out the waveform, for if the resistor is removed, and the two capacitors are connected together to add their capacitances, the waveform becomes merely a smaller amplitude version of the shape in Figs. 21-2 or 21-3.

When we come to study the subject of decoupling you will find combinations of resistors and capacitors at the plate and screen connections of many tubes. The primary purpose of these combinations is to prevent signals from going where they cause trouble, but an incidental effect is further filtering and further reduction of ripple voltage.

MULTI-SECTION FILTERS. Assume that we have added to the first filter capacitor at the left in Fig. 21-8 the choke and second capacitor drawn with heavy lines, and that we find this addition decreases the output ripple voltage to 1/20 or 5% of the ripple when using only the first capacitor. But supposing the ripple voltage must be reduced to 1/20 of 5%, which is 1/4 of 1%. This can be done by increasing either of the capacitances 20 times. For example, were we using capacitances of 8 microfarads, increasing one of them to 160 microfarads (8 times 20) would bring the ripple down where we want it.

There is another way of obtaining the desired reduction of ripple voltage, we may add a second section to the filter, as in the right-hand diagram. The second section consists of another choke and another capacitor. In this second section we may use a choke and a capacitor just like those in the first section. Previously we have found that this combination drops its output ripple voltage to 1/20 of the input ripple. The ripple input to this second filter section is our original 5%. The second section will drop this ripple to 1/20 of 5%, just what we want. By adding one 8-microfarad capacitor and a choke we have the same result as with an added 160-microfarad capacitor and no extra choke. Of course, the exact reduction of ripple in any filter section would depend on values of capacitance and inductance. We have used the fraction 1/20 merely to illustrate the principle.

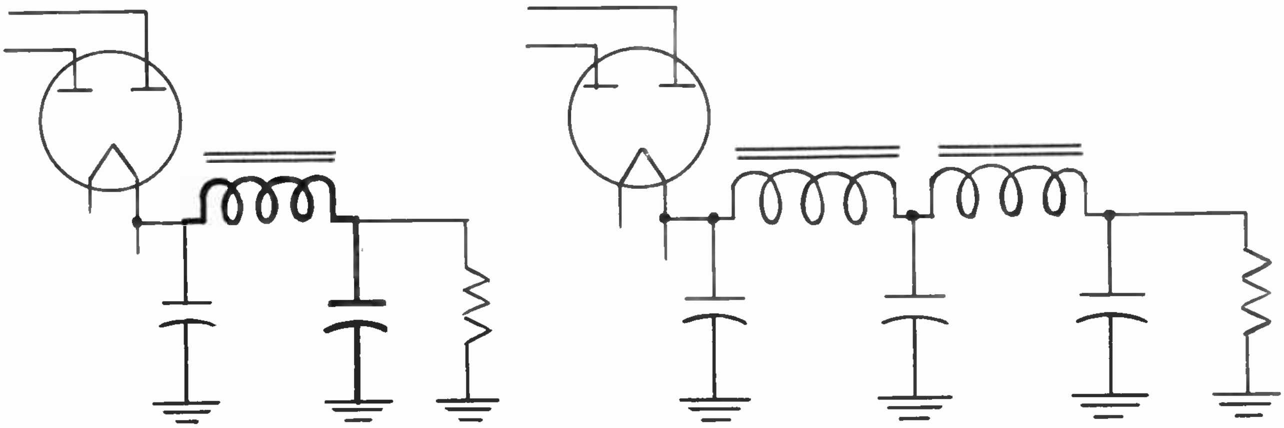


Fig. 21-8. Adding a second section to the filter.

Adding an 8 microfarad capacitor would cost much less than adding a 160-microfarad unit, but the extra choke would be quite costly. Were we using a resistor-capacitor filter system, adding an extra resistor and capacitor would cost much less than adding one big capacitor, because resistors cost relatively little. In modern receivers you seldom will find two-section filters using chokes, but there are many two-section and quite a few three-section filters of the resistor-capacitor type. At the output of every added section the ripple is a small fraction of its value at the input to that section.

FILTER CONNECTIONS. In the filter diagrams of this lesson have been shown full-wave and half-wave rectifier tubes. The rectifier tube is not a part of the filter, consequently any of the tube rectifiers might be replaced with a full-wave or half-wave selenium rectifier and the explanations of filter action would be unchanged.

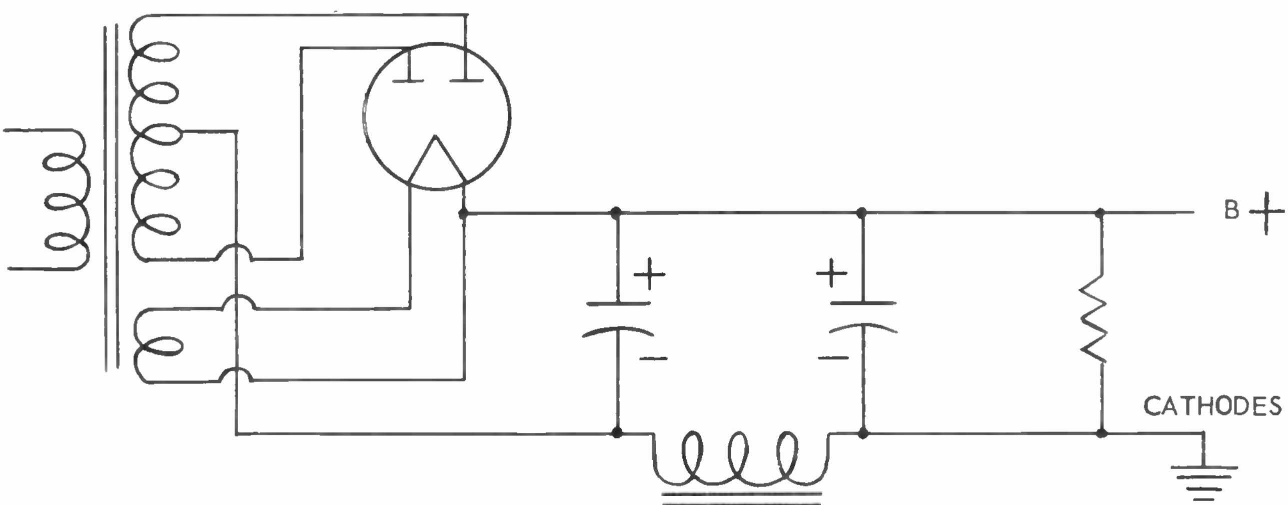


Fig. 21-9. The choke may be in the negative side of the filter system.

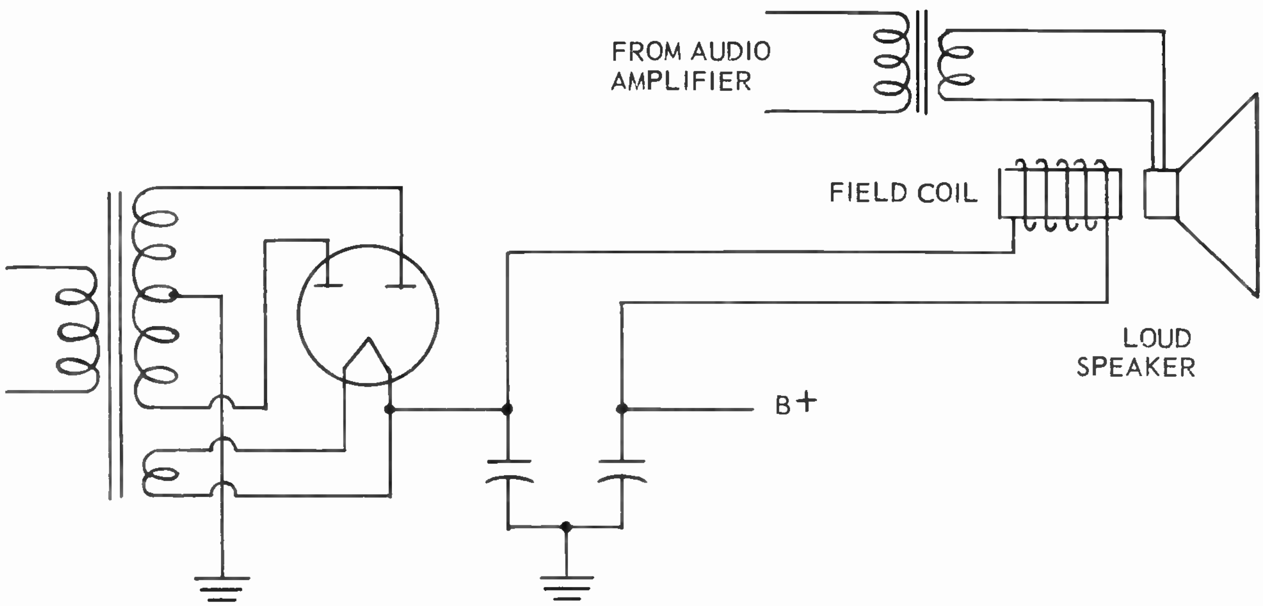


Fig. 21-10. Loud speaker field coil used as a filter choke.

The filter choke most often is in the positive side of the power circuit, as shown in preceding diagrams, but sometimes it is in the negative side as in Fig. 21-9. A choke provides the same smoothing effect when in one side of the circuit as in the other, because there must be the same changes of voltage and current in both sides. With the choke in the negative side of the filter system it is impossible to use a dual electrolytic filter capacitor having a common negative terminal for both elements.

In some receivers, both radio and television, the loud speaker may be of a type containing a "field coil" which is suitable for use as a filter choke in the power supply while also performing its regular duty in the speaker. The field coil consists of a many-turn winding on an iron core. The purpose of this coil, in the speaker, is to provide a strong electromagnetic field of constant polarity. This requires flow of direct current in the winding. This direct current is that passing through the power supply when the field coil is used as a choke. Typical connections are shown by Fig. 21-10.

Quite often you will find plate and screen voltages for certain tubes taken from one point on the filter system, while voltage for other tubes is taken from a different point. Such connections are shown by Fig. 21-11. Tube *A* is a voltage amplifier, a type which takes only a moderate amount of plate current. Tube *B* is an output amplifier or power amplifier, taking much greater plate current. These two tubes might represent any number of low-current and high-current tubes in a receiver.

In order to deal with actual figures in an example of what happens we may assume there is a gain of 20 times between the output or plate circuit of tube *A* and the output or plate circuit of tube *B*. Then any signal voltage coming from the plate of *A* to the grid of *B* will be amplified 20 times, and in the output from *B* will be 20 times stronger than in the output from *A*. Were plate voltages for both tubes to come from the same place in the power supply filter, both would contain the same ripple voltage. This ripple in the plate circuit of tube *A* would be made 20 times as strong in the plate circuit of tube *B*, because ripple voltage will be amplified along with signals.

But note that we take plate voltage for amplifier *A* from a second section of the power supply filter, and plate voltage for amplifier *B* from the first section. If the second section of the filter reduces ripple voltage to 1/20 its value at the output of the first section, the ripple voltage to amplifier *A* will be only 1/20 as strong as to amplifier *B*. When this ripple at tube *A* is amplified 20 times it still will be no stronger than the ripple put into the plate circuit of tube *B*.

This method of connection has still other advantages. Since the second section of the power supply filter furnishes only relatively small load currents, this section may be of the resistor-capacitor type, to save the cost of an added choke. Furthermore, plate voltage taken from the first filter section is higher than from the second section because it avoids the drop in the filter resistor. Output tubes or power tubes ordinarily require higher plate voltages than voltage amplifier tubes.

To furnish plate voltages for amplifiers which are followed by still greater gains we might add a third filter section, using in it a rather high resistance to obtain maximum reduction of ripple voltage. Variations in connections to filter systems are almost without number, but they employ the principles which have been explained and you should have little difficulty in determining how they are supposed to work.

VOLTAGE REGULATION. With every source of voltage and current there is a drop of output voltage when current increases. The reason for this voltage drop is simple. Let's say our source is the secondary winding of a power transformer which furnishes 200 volts when delivering no current. This 200 volts really is the emf induced in the winding. We shall assume that resistance of the wire in this winding is 100 ohms. If the winding delivers current of 50 milliamperes this current flows in the resistance of the winding. The drop of potential in 100 ohms carrying 50 milliamperes is 5 volts. Then the effect of the 100 ohms internal

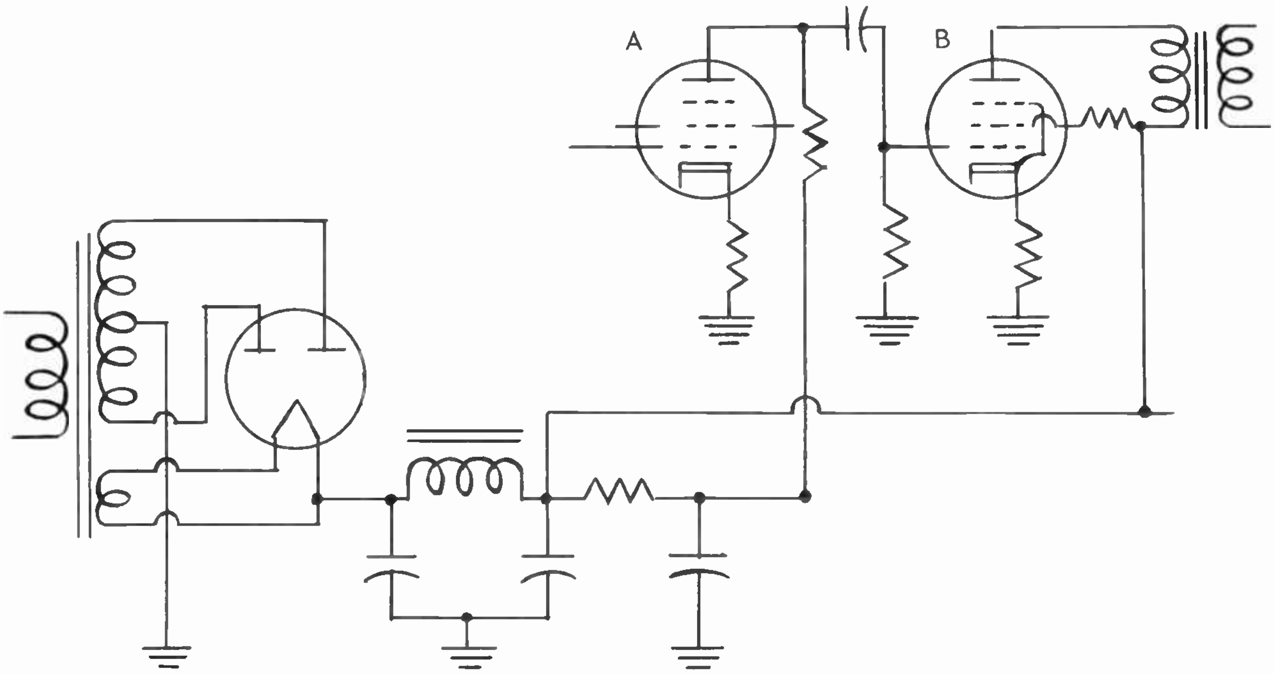


Fig. 21-11. Amplifier plate voltages from different points on the filter.

resistance of the winding is to drop the 200-volt induced emf down to 195 volts at the winding terminals, with the missing 5 volts used up in resistance of the winding. Induced emf still is 200 volts, but only 195 volts remain at the winding terminals. Were current to become 100 milliamperes the internal drop would increase to 10 volts, and the terminal voltage would go down to 190.

Every source, whether it be a power supply of the kinds we have examined, or a battery, an electric generator, or what not, every one of them must have more or less internal resistance. Sources cannot be built without conductors, and all conductors have resistance. Consequently, the output voltage or terminal voltage always will drop when current is drawn, and the drop will increase as more current is drawn.

If there is only small drop of output voltage with rather large increases of current we say that the regulation is good. If there is a large drop of voltage the regulation is poor. How good or how poor the regulation may be is usually expressed as a fraction or as a percentage. To determine the percentage of regulation we measure the higher voltage which exists with no current, then the lower voltage when current flows, and divide the difference by the higher voltage. This gives regulation as a fraction. Multiplying by 100 changes the fraction to a percentage.

For example, assume that a power supply provides 200 volts potential difference when delivering no current, but only 190 volts when some certain current flows. The difference is 10 volts. Dividing 10 by the higher voltage, 200, gives the fraction 0.05 or 5/100. Multiplying by 100 changes the fraction to 5%, which is the voltage regulation with this particular current.

Now what about the voltage regulation of an entire power supply system? If we consider the power supply as a complete unit or source it contains three internal resistances. First, there is resistance of the transformer winding if a transformer is used. Second, there is resistance of the rectifier, which is plate-cathode resistance of a tube or anode-cathode resistance in a selenium rectifier. Third, there is resistance of the filter choke or resistor, or both, or of more than one of either. All these resistances are in series, so their voltage drops add together. There also is resistance in the wiring connections, but it is negligible. The greater the sum of the resistances the greater is the voltage drop for any given current, and the poorer is the voltage regulation.

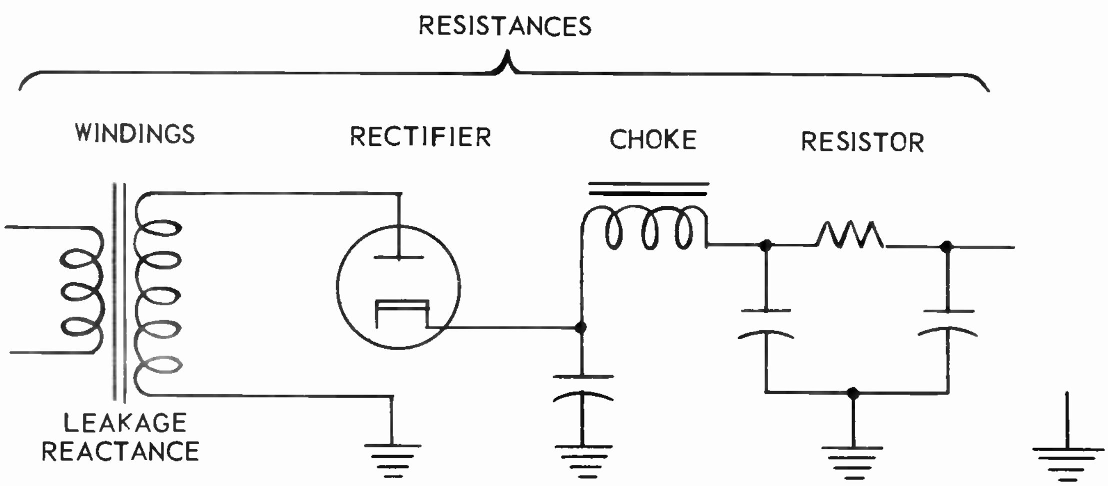


Fig. 21-12. Factors which affect voltage regulation.

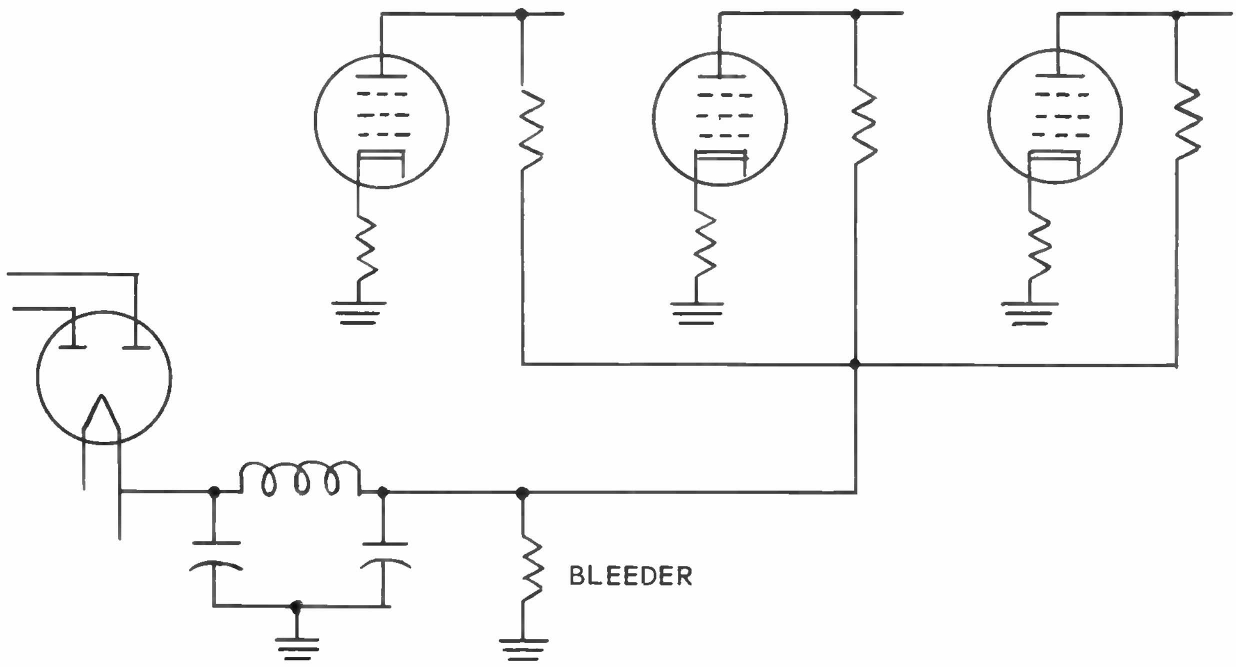


Fig. 21-13. A bleeder resistance on the filter output.

Resistances which affect voltage regulation are shown by Fig. 21-12, and the transformer is marked with another factor called leakage reactance. In a transformer of poor design or construction a great many magnetic lines produced by primary current do not cut through turns of the secondary windings, these lines "leak" away and do nothing useful in induction of secondary emf and voltage. The effect is like that due to increased reactance or resistance. When secondary current increases, the induced emf cannot keep up with the increase because leakage of magnetic lines prevents having enough induction or enough coupling between primary and secondary. As a consequence, secondary emf and voltage suffer a severe drop. Cheap transformers don't pay, especially in service replacements.

Poor voltage regulation, whatever its cause, is a bad thing, for when amplifiers and other tubes make demands for more current they get less voltage along with the greater current, and performance suffers.

There are considerable differences between the plate-cathode resistances of various rectifier tubes. A 5Y3-G tube has more internal resistance than the bigger 5U4-G, and would cause poorer regulation with the same current. However, the regulation might be no poorer with the smaller current which the 5Y3-G is supposed to carry. Some of the small heater-cathode rectifier tubes used in many transformerless receivers have rather low resistance, and do not cause poor regulation when carrying their normally small currents.

9 Filter chokes which have large cores do not require so many turns of wire in their windings to provide a given inductance. Also, windings of liberal size wire have less resistance. Such structural features allow good regulation. Resistors in resistor-capacitor filters tend to cause poor regulation when used for large currents.

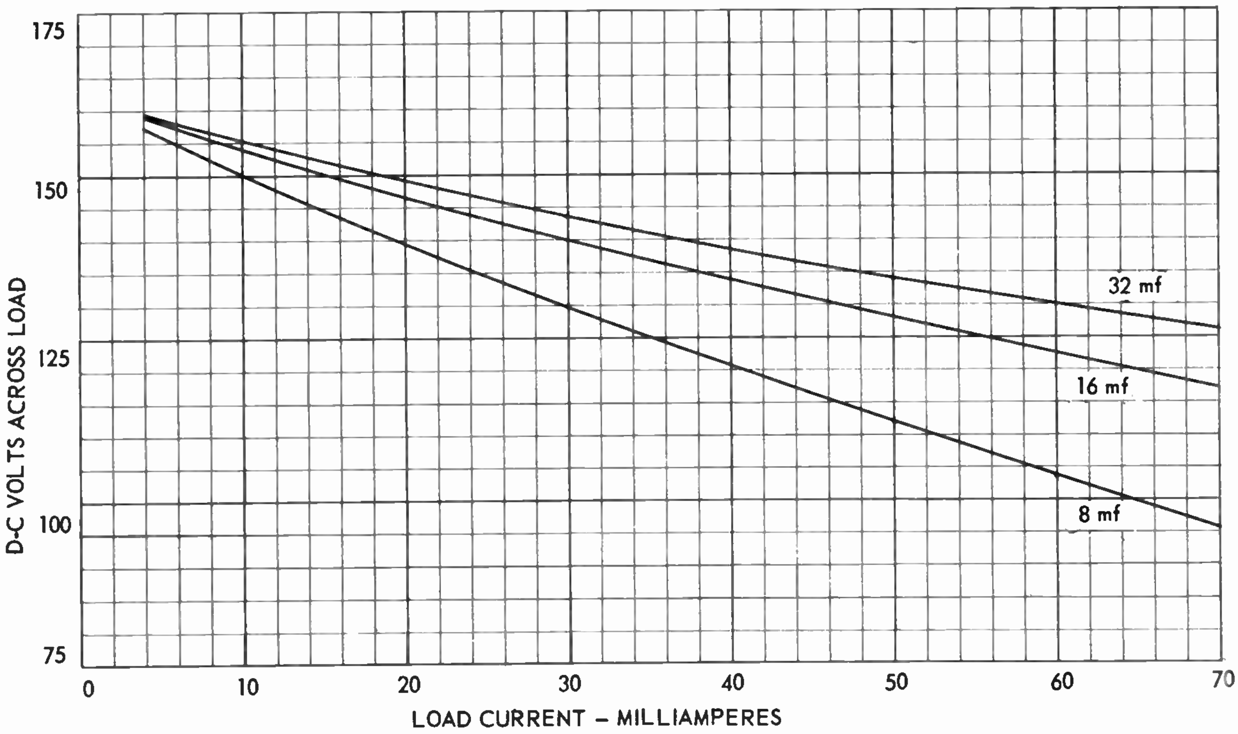


Fig. 21-14. How change of capacitance affects the relation between load current and load voltage.

Using bleeder resistance across the output of the power supply, as in Fig. 21-13, improves voltage regulation. A bleeder is any resistance so connected that through it there is steady flow of current from the power supply, in addition to current furnished to the receiver tubes. For an explanation of bleeder effect we shall assume that total plate current to all the tubes is 50 milliamperes, and that bleeder resistance is of such value as to allow in it a steady current of 10 milliamperes. Then total current from the power supply is 50 plus 10, or is 60 milliamperes.

Assume now that tube current momentarily increases to 65 milliamperes, making a total current of 75 milliamperes for tubes and bleeder. This is a current increase of 25 per cent, from the former 60-milliampere total. Without the bleeder the increase of tube current from 50 to 65 milliamperes would be 30 per cent of the initial 50 milliamperes. This greater percentage change of current would cause a greater percentage drop of tube voltage, or would cause poorer regulation.

Enough bleeder current to make worth while improvement of regulation imposes a heavy drain on the power supply. The greater total current increases the ripple voltage, as we learned earlier. Because of greater ripple and waste of energy in both bleeder resistance and power supply resistances, designers usually avoid the use of bleeders.

The value of filter capacitance affects power supply regulation materially. Fig. 21-14 shows how voltage furnished to the load is changed by load current when using three different values of capacitance in a capacitor connected across one of the smaller rectifier tubes, a 25Z6. For this test there is only a capacitor and a load resistance, as back in Fig. 21-2. There is no complete filter, because we wish to observe only the effect of capacitance, The lower curve applies with 8 microfarads capacitance, the middle one with 16 microfarads, and the top one with 32 microfarads across the rectifier.

For any given change of load current there is much greater change of voltage when using the small capacitance than when using a larger one. With 8 microfarads a change of current from 20 to 40 milliamperes drops the voltage from 139 to 121, a difference of 18 volts. With 16 microfarads the same change of current drops the voltage from about 147 to 134, a difference of only 13 volts. Here we may see also how more capacitance raises the voltage to the load. Input to the rectifier for this experiment was 117 alternating volts from a power line, with no transformer. With fairly small output currents the direct voltage to the load is higher than the effective alternating voltage to the rectifier.

Any given rectifier operated at high voltage allows greater change of voltage than when operated at lower voltage. We may see this from Fig. 21-15. The lower curve here is the same as the middle curve of Fig.

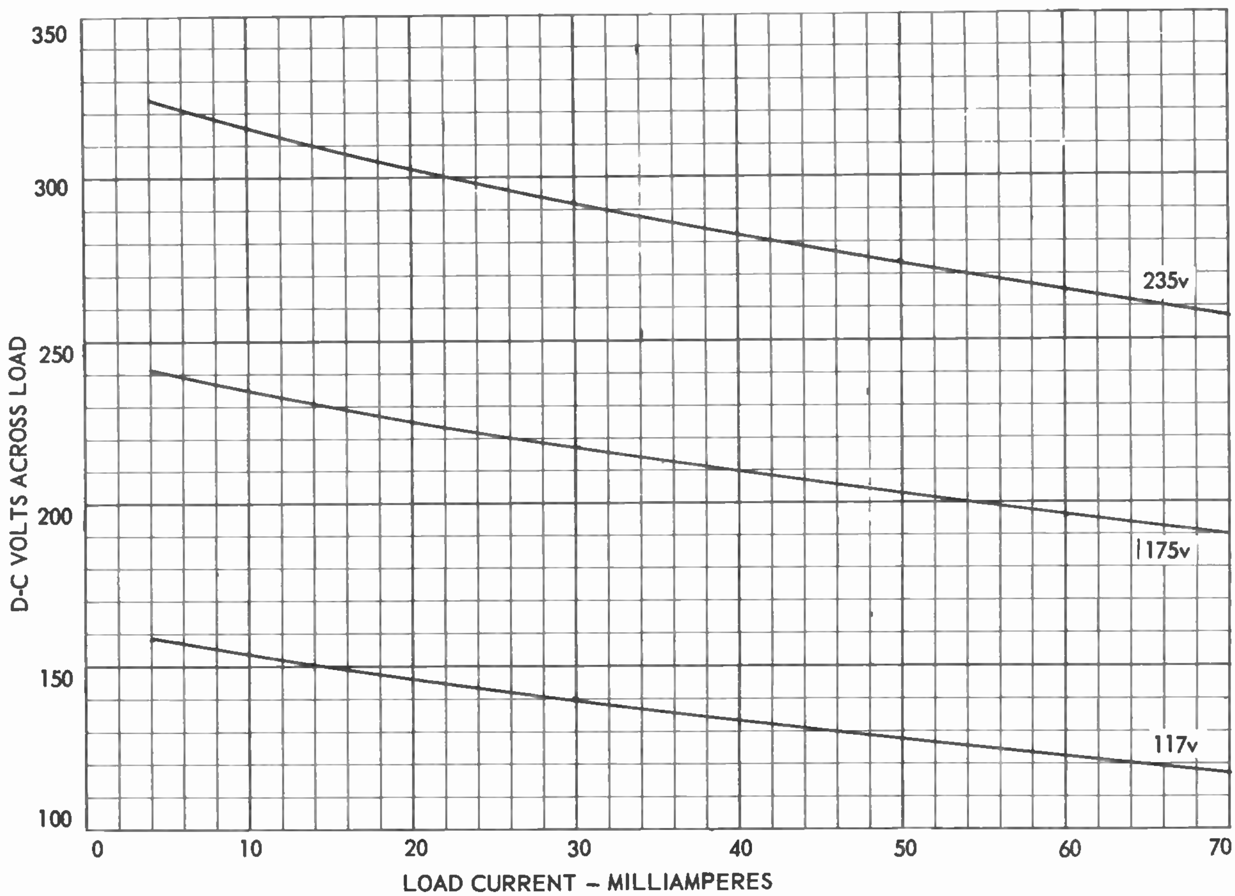


Fig. 21-15. How change of alternating voltage to the rectifier changes the relation between load current and load voltage.

(10)

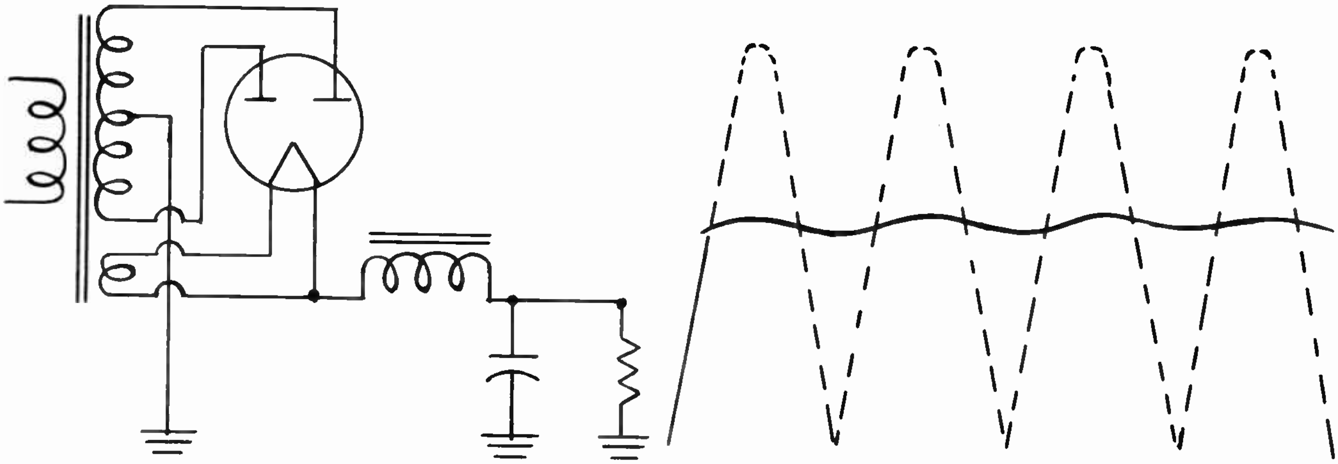


Fig. 21-16. A choke input filter and its performance.

21-14, it applies to the 25Z6 rectifier connected to 16 microfarads capacitance, with rectifier input 117 alternating volts from the power line. The curve appears flatter because it is drawn on a different voltage scale.

The middle curve shows output current and voltage relations when input to the rectifier is 175 alternating volts, and the top curve applies with 235 alternating volts input, always with the same 16 microfarads across the rectifier output. For any given change of current there is slightly greater drop of voltage along the top curve than along those below. Although there is a greater drop along the top curve when measured in volts, this drop is not so high a percentage as it is along the lower curves. Voltage regulation is not measured in volts, it is measured as a percentage. Therefore, we have better regulation at the higher operating voltages in Fig. 21-15 than at lower operating voltages.

If we compare a selenium half-wave rectifier with a tube half-wave rectifier, when both have similar ratings for maximum current-carrying capacity, most of the newer type selenium rectifiers will allow somewhat better voltage regulation. When comparing a half-wave voltage doubler with a straight half-wave rectifier, both using either the same type tube or the same type selenium unit, regulation of the doubler is poorer than of the straight rectifier. Also, the regulation of a voltage tripler is still poorer than that of a doubler. None of the differences, however, are great enough that they cannot be compensated for by using large capacitors or an extra filter section.

CHOKE-INPUT FILTERS. In all the filters which have been shown there is a capacitor across the output of the rectifier. All these are called capacitor-input filters. In Fig. 21-16 the output of the rectifier is connected directly to a choke, and the first capacitor follows the choke. This is a choke-input filter. At the right is shown, by the solid line curve, the voltage waveform which appears across the capacitor and the resistance being used for a load. In broken lines are shown the pulses of rectified voltage which exist at the rectifier output, between the rectifier cathode and B-minus or ground. These pulses act on the choke.

Compare the circuit and the output voltage of this choke-input filter with those of the capacitor-input filter in Fig. 21-4. In both places we have the same rectifier, the same choke, the same capacitor, and the same load. The only difference is in the arrangement of these parts in relation to one another.

Output voltage from the choke-input filter is only 65 to 70 per cent of that from the capacitor-input type. To have equal direct voltages at the outputs would require much higher alternating voltage at the plates of the rectifier for the choke-input system. Reduced voltage output results from applying to the choke the strong alternating component of the rectifier pulses. The choke has high reactance to this alternating component.

1 Voltage regulation with a choke-input filter usually is about twice as good, or is of about half the percentage of regulation with a capacitor type when both use the same inductance and capacitance and when both deliver the same output voltage. Also, the ripple voltage from a choke-input filter is less than that from a capacitor-input type when both have the same inductance and capacitance. The big objection to the choke-input filter is its low output voltage in comparison with the value of alternating voltage applied at the rectifier. Choke-input filters are not used in ordinary television or sound radio receivers. They are used with sound amplifiers from which is desired the best possible reproduction.

TELEVISION

LESSON NO. 22

VOLTAGE DIVISION IN THE POWER CIRCUITS

Training Your Mind
C.T.I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

DO NOT TEAR — CUT ALONG THIS LINE AND SEND IN FOR GRADI

Training Your Mind
C.T.I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON NO. 22

VOLTAGE DIVISION IN THE POWER CIRCUITS

In these lessons dealing with power supplies we have progressed from the a-c line through the power transformer, the rectifier, and the filter system. At the output of the filter we have smooth direct voltage for plates, screens, and grid biasing connections of the receiver tubes. A few of the tube circuits may require the full positive potential from the power supply, but many other circuits must be operated with various lower potentials. It is fairly apparent that a rather elaborate system of voltage division will be needed to handle the many tube circuits of even a simple receiver.

To get acquainted with practical methods of voltage division, we shall follow the power circuits which start from a low-voltage B-power supply like the one of the receiver pictured in Fig. 22-1. At the rear of the television receiver chassis is the power transformer. Behind the transformer is the full wave rectifier tube. Next to the rectifier tube is the filter choke and next to the choke is the filter capacitor. These four components represent the major parts of most low voltage full wave type power supplies, and are responsible for delivering B-power to every tube except detectors of some kind and certain tubes used for automatic controls.

Fig. 22-2 is a diagram of all the wiring between a power supply rectifier-filter system and plates and screens in the tubes of one of the smaller and simpler television sets. This diagram has been copied from a manufacturer's service diagram, but the wiring is shown only so far as its connections to plate loads and screens for the tubes requiring B-power. None of this wiring carries signal voltages or currents. It does nothing but furnish the direct voltages and currents to which are added the signal voltages and currents and from which are taken the signal voltages and currents as may be required during operation of the receiver.

At the ends of each of the wiring lines are the names of tubes whose plates or plates and screens are there connected. Where two lines are shown running to the same tube, the line extending vertically is for
2 the plate and the one extending horizontally is for the screen. Voltages which exist at plates and screens with the set in operation are marked on the diagram. All these are positive or B-plus voltages measured with reference to chassis ground as zero or B-minus. Nearly all the wiring lines terminate at resistors whose free ends would connect to a plate load or a screen in a complete diagram.

We may commence tracing at the filament-cathode of the rectifier tube in the low-voltage power supply. At this point there is a potential difference to ground of 255 volts. Directly from the rectifier output, ahead of the power supply filter, a line running downward to a tube in the high-voltage power supply furnishes 255 volts to the plate of that tube and 230 volts to its screen. The lower voltage on the screen is due to voltage drop as screen current flows in the resistor leading to the screen.

Doubtless you wonder why unfiltered direct voltage is furnished through this line. It is because the tube here connected is an oscillator used to generate voltages at frequencies in the neighborhood of 200,000

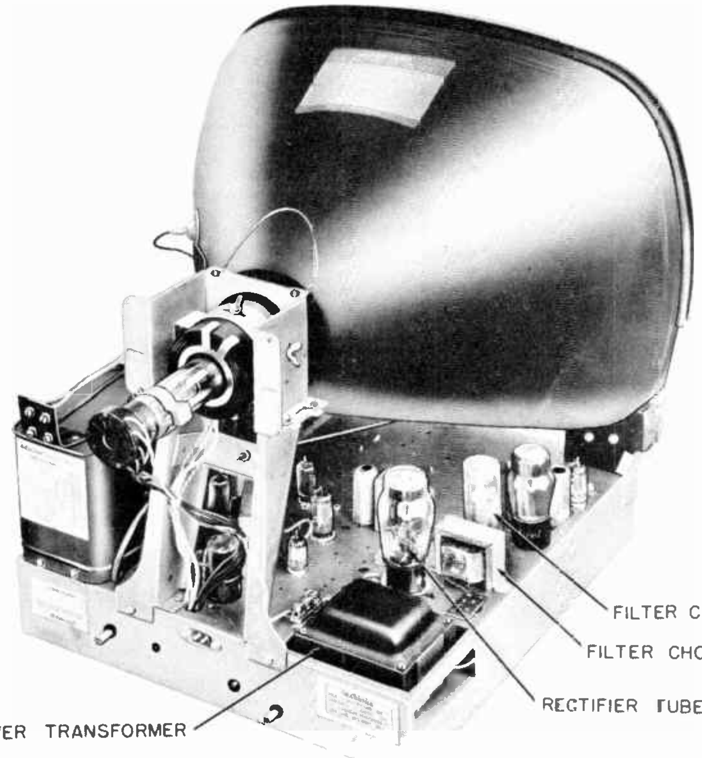


Fig. 22-1. Power supply parts are mounted at the rear of this television chassis.

cycles per second, which then are stepped up to about 6,000 volts, rectified, and filtered before being used. Preliminary filtering in the low-voltage power supply is unnecessary, so current for the high-voltage oscillator does not flow through the low-voltage filter. As you know, the less the current in the low-voltage filter the better is its smoothing effect.

At the output of the low-voltage power supply filter in Fig. 22-2 the voltage is down from 255 to 240 volts because of a 15-volt drop in resistance of the filter choke. The conductors directly connected to this 240-volt point are drawn with heavy lines until coming to a resistor or a tube element. Everywhere along these conductors the voltage will be 240. Following straight out to the right along the heavy-line conductor we come to the plate circuit of the horizontal sweep oscillator. There is a drop of 40 volts in the resistor, leaving 200 volts at the oscillator plate.

Another heavy line runs downward to the vertical sweep oscillator, at whose plate there remains a potential of only 55 volts. This indicates a loss of 185 volts (240 minus 55) in the resistance between the line and the tube plate. Total resistance here is about 5 megohms. In this much resistance a current of less than 1/25 milliampere will flow when the potential difference is 185 volts. You might figure this out for yourself by using one of the formulas you learned in an earlier lesson.

Next we follow upward and to the right along the heavy-line 240-volt conductor. The first connection to this line is at the audio intermediate-frequency amplifier, whose plate and screen each get 160 volts. This means a drop of 80 volts in the resistor leading to the plate and screen. The next connection delivers 85 volts, through a resistor, to the plate of the audio amplifier tube. The voltage drop in this resistor must be 155 volts.

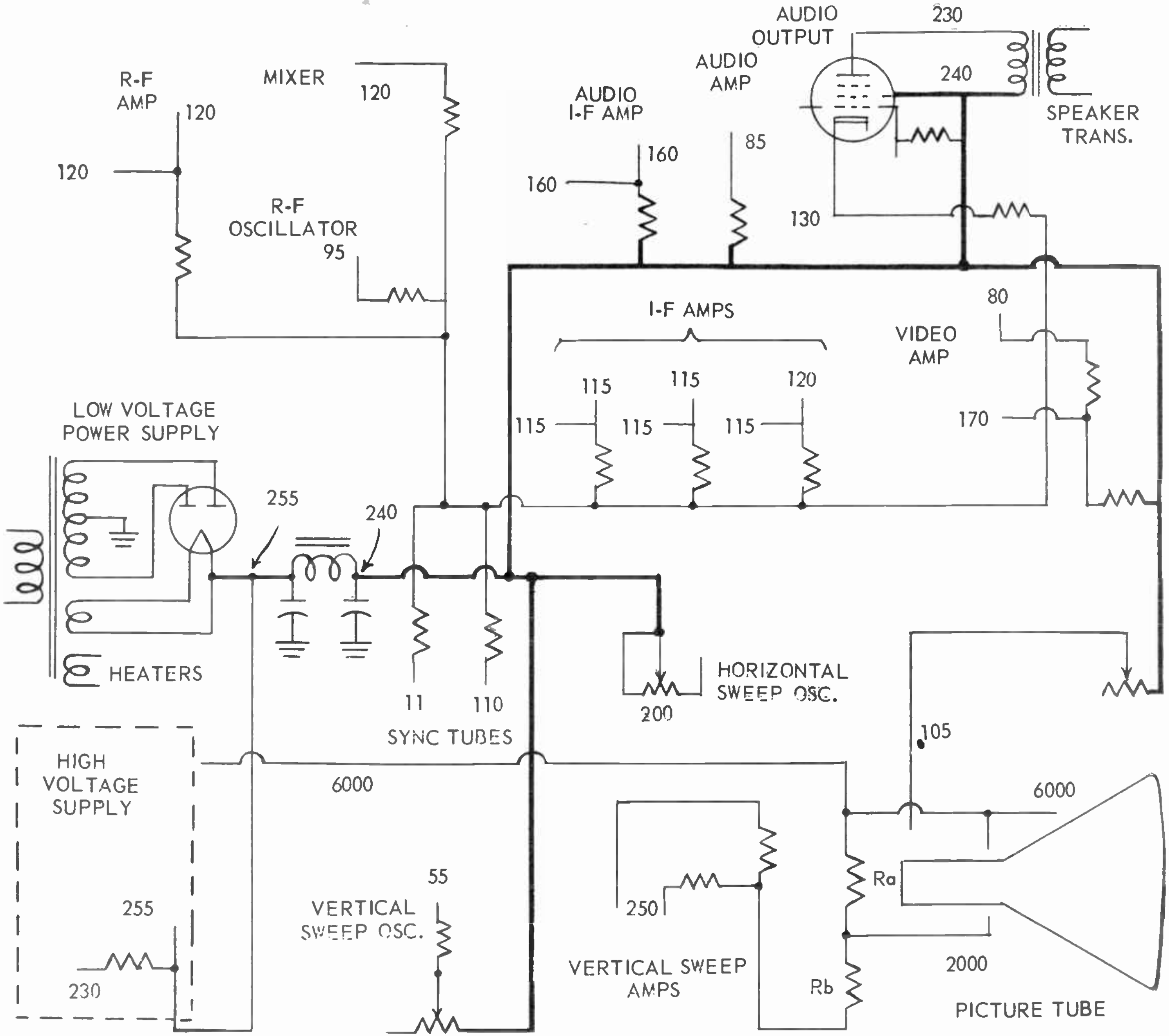


Fig. 22-2. The B-plus wiring of a television receiver, showing connections to plates and screens.

A little farther toward the right a branch line leads upward to the screen of the audio output tube, and to the primary winding of a speaker coupling transformer which connects to the plate of the output tube. Voltage at the screen of the tube is 240 and at the plate is 230. Now note that voltage at the cathode of this tube is shown as 130 volts. The difference between voltages at plate and screen, and at the cathode, is accounted for by potential drop within the tube. The 130 volts remaining at the cathode of the audio output tube becomes the plate and screen supply voltage for many other tubes.

If you follow along the line from the cathode of the audio output tube it will lead to plates or to plates and screens in the three intermediate-frequency amplifier tubes, in the two sync tubes, and in the r-f amplifier, oscillator, and mixer tubes which are in the tuner of the receiver. The cathodes of all these tubes handled through the audio output tube connect directly or through resistors to ground. Plate and screen voltages on all these tubes are somewhat less than the 130 volts from the audio output tube. Note that we get all the way down to 11 volts at the plate of one of the tubes in the sync section.

Again continuing to trace along the 240-volt line we find it feeding the plate and screen of the video amplifier tube with, respectively, 80 and 170 volts, and finally furnishing 105 volts to one of the elements of the picture tube. This element is the cathode.

Now we may go back to the high-voltage power supply, which receives 255 volts and 230 volts from the low-voltage supply. At the output of the high-voltage power supply the potential difference to ground is 6,000 volts. This very high potential is applied to plates and accelerating electrodes in the picture tube. Then there is a drop through some control resistors which are indicated in the diagram by a single resistor marked *Ra*. The remaining 2,000 volts go to a focusing electrode in the picture tube. Then there is a further drop through more resistance at *Rb* and we have 250 volts remaining for the plates of the vertical sweep amplifiers.

BIAS FROM POWER SUPPLY. With a power supply circuit such as in Fig. 22-2 the voltages for grid biasing are obtained by cathode bias or by means of the grid-leak method. With another type of power circuit there is provision for obtaining fixed grid bias from a voltage divider. The principle is illustrated by Fig. 22-3. Between the positive output of the filter and the negative center tap on the power transformer secondary are connected resistors *Ra* and *Rb* in series. These are the voltage divider resistors. From between these resistors there is a connection to ground.

Also connected between the positive and negative sides of the power supply is a triode tube. The plate of the triode is connected through resistor *Ro* to the positive side of the power supply, at the filter. The triode grid is connected through resistor *Rg* to the negative side of the power supply. The cathode is connected to ground, and through ground to the grounded point on the voltage divider.

Signal input to the triode grid is through a capacitor, and signal output from the plate is through another capacitor. These capacitors prevent passage through them of the direct electron flows in the power supply circuit. Consequently, the alternating signal voltages and resulting alternating components of current in the triode need not be considered in our present dealings with direct voltages and currents in the power supply circuits.

There are two electron flows, whose directions are shown by arrows. The flow shown by broken-line arrows starts from the negative center tap of the power transformer, passes upward through voltage divider resistors *Ra* and *Rb*, and then goes back through the power supply filter choke and rectifier tube to the transformer winding.

The other electron flow, shown by full-line arrows, starts from the negative center tap of the transformer

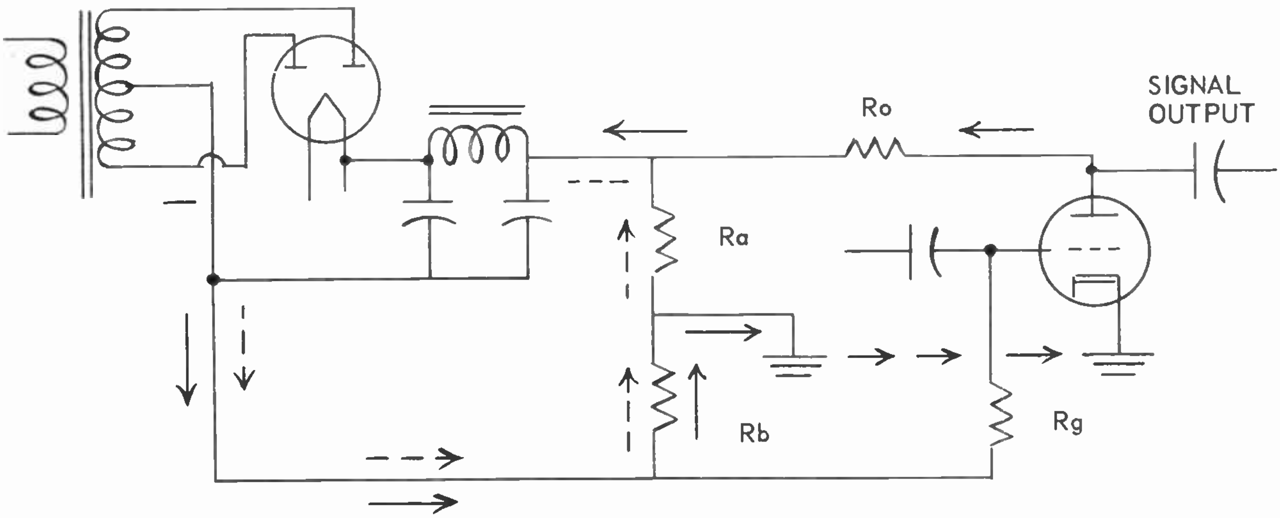


Fig. 22-3. Electron flows in a power supply voltage divider.

and passes upward through divider resistor *Rb* to ground without going on through resistor *Ra*. This flow goes through chassis ground to the cathode of the triode tube. Going through the tube from cathode to plate, the flow continues back through resistor *Ro*, the filter choke, and the rectifier to the transformer.

We wish to make the triode plate 205 volts positive and its grid bias 10 volts negative with reference to the cathode. You will recall that effective plate voltage and grid bias always are measured with reference to the cathode of the same tube. To obtain these voltages will require certain values of resistance at *Ra* and *Rb*, which we shall compute to show how it may be done. We need not concern ourselves with the value of resistance at *Rg*, because with no signal and the triode grid negative there will be no electron flow and no potential difference across this resistor. Then grid voltage will be the same as voltage at the bottom of *Rg*.

3 The cathode of the triode is connected through ground to the mid-tap on the voltage divider. The plate is connected to a point which is positive with reference to ground and the cathode. The grid is connected to a point which is negative with reference to ground and the cathode. With these correct polarity relations of cathode, plate, and grid it remains only to compute resistances which will allow the desired voltages.

To begin with we shall assume that there is a total potential difference of 260 volts between the positive output of the power supply filter and the negative center tap of the transformer, and that this potential difference is maintained during operation of the triode. We shall assume also that plate-cathode current in this particular triode will be 15 milliamperes with 205 volts on the plate and 10 volts negative bias on the grid. This current value would be found from characteristics applying to whatever tube is in use.

Now we may go to Fig. 22-4 where are shown currents and voltages from which may be computed the required resistances with the help of our regular formula for resistance. The steps are as follows.

5 1. We know that grid bias is to be 10 volts. Bias is the difference of potential between grid and cathode.

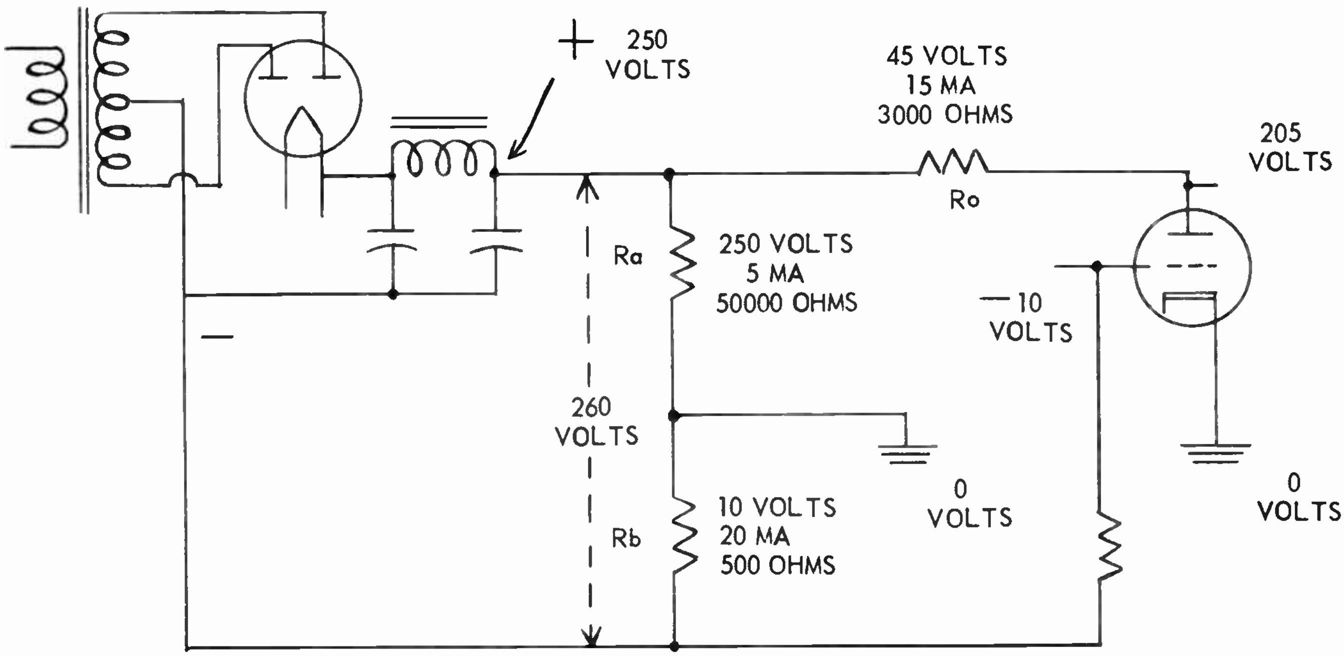


Fig. 22-4. Voltages, currents, and resistances in circuits connected to the voltage divider.

The grid and cathode are connected across divider resistor Rb. Therefore, potential difference across Rb must be 10 volts.

2. Total potential difference is 260 volts from positive filter output to negative on the power supply. This must be the potential difference across Ra and Rb together. Since 10 volts of this total must be across Rb, the remainder of 250 volts must be across Ra. The bottom of Ra goes to ground and the top goes to the filter output. Therefore, at the filter output the potential must be 250 volts positive with reference to ground.

3. Triode plate voltage is to be 205, which is 45 volts less than the 250 volts at the filter output. This 45 volts must be dropped in resistor Ro. We know that current in Ro is the 15-milliampere plate-cathode current. Then, using the formula,

$$\text{Ohms} = \frac{1000 \times 45 \text{ (volts)}}{15 \text{ (ma)}} = \frac{45\,000}{15} = 3000 \text{ ohms at } Ro.$$

4. Now we shall determine the resistance required at Ra. This resistor is merely a bleeder whose purpose is to improve voltage regulation. We may allow Ra to carry any current which seems suitable for this purpose. We might decide to allow 5 milliamperes. Then we compute the resistance which allows 5 milliamperes current with a potential difference of 250 volts.

$$\text{Ohms} = \frac{1000 \times 250 \text{ (volts)}}{5 \text{(ma)}} = \frac{250\,000}{5} = 50{,}000 \text{ ohms at } Ra.$$

5. In computing the resistance for Rb we note that this resistor is carrying the 15-milliampere plate-cathode

current to the cathode, also the 5-milliampere current which goes on upward through *Ra*, making a total of 20 milliamperes in *Rb*. Potential difference across *Rb* is to be 10 volts. Now we use the formula for resistance.

$$\text{Ohms} = \frac{1000 \times 10 \text{ (volts)}}{20 \text{ (ma)}} = \frac{10\,000}{20} = 500 \text{ ohms at } Rb.$$

Resistances for any simple voltage divider may be computed with this general method.

NEGATIVE CATHODE CONNECTIONS. Fig. 22-5 shows another way of employing a voltage divider. The principle here illustrated is used in a great many television sets. The divider consists of four resistors, from *Ra* down through *Rd*, with a ground connection from between *Rb* and *Rc*. We shall assume a total potential difference of 380 volts from the power supply. From ground to the top of resistor *Ra* the potential is 280 volts positive, and from ground to the bottom of resistor *Rd* the potential is 100 volts negative. There is an intermediate positive potential of 180 volts from between resistors *Ra* and *Rb*, and an intermediate negative potential of 15 volts from between *Rc* and *Rd*.

Potentials listed for this divider system are chosen merely to have some definite figures with which to show what happens. We could have any desired number and values of potentials from a voltage divider by using the necessary number and values of resistors, with taps between them. As many as six or eight volt-

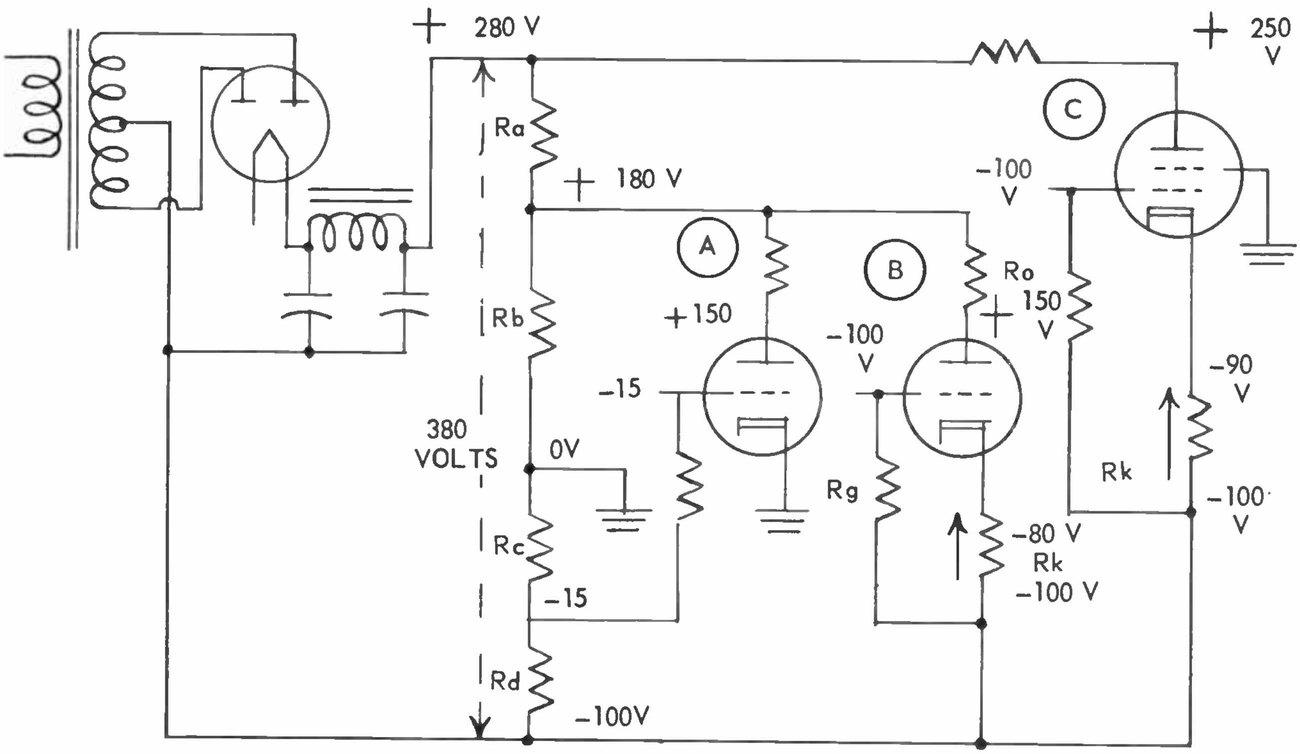


Fig. 22-5. Tube cathodes which are highly negative with reference to ground.

age taps, in addition to a ground tap, are not at all unusual in television receivers. How much of the total supply voltage is positive and how much is negative with reference to ground depends on where the ground connection is made along the divider.

Triode *A* of Fig. 22-5 is operated with fixed bias obtained by the same general method as in Figs. 22-3 and 22-4. The bias on this new triode is 15 volts negative and the plate is 150 volts positive with reference to ground, and the cathode. These or other plate and grid voltages may be secured by suitable choice of divider resistances.

Triode *B* is operated with cathode bias supplied by voltage drop across cathode resistor *Rk*. The bottom of *Rk* is connected to 100 volts negative on the voltage divider. Plate-cathode current flowing upward through *Rk* is accompanied by a drop from 100 volts to 80 volts negative. The grid is connected through resistor *Rg* to 100 volts negative, so grid potential is 100 volts negative because there is no current and no potential difference across resistor *Rg*.

But grid bias in triode *B* is not 100 volts negative. You must remember that grid bias always is the difference of potential between grid and cathode of the same tube. The grid of this tube is 100 volts negative and its cathode is 80 volts negative. This makes the grid 20 volts more negative than the cathode, so there is a 20-volt negative grid bias in spite of the fact that a voltmeter connected between grid and ground would show the grid to be 100 volts negative.

The plate of triode *B* is connected through resistor *Ro* to the 180-volt positive point on the voltage divider. An assumed 30-volt drop in resistor *Ro* leaves the plate 150 volts positive, with reference to ground. With the plate 150 volts positive and the cathode 80 volts negative, the difference of potential between plate and cathode is 230 volts, which is the sum of 150 and 80. Consequently, although a voltmeter measurement from plate to ground would show only 150 volts, the same as at the plate of triode *A*, the actual plate voltage or plate to cathode voltage on triode *B* is 230 volts.

To determine a difference of potential when one potential is positive and the other negative we add the two potentials. It is like determining a temperature difference when one temperature is above zero and the other below. If the temperature changes from 30 degrees above zero to 10 degrees below zero it has become 40 degrees colder, which we find by adding 30 (positive) and 10 (negative).

Pentode *C* of Fig. 22-5 is also operated with cathode bias. Here the potential difference across cathode resistor *Rk* is the 10-volt difference between 100 volts negative and 90 volts negative. The grid of this tube is connected to 100 volts negative on the divider, so grid potential to ground is 100 volts. But grid bias is 10 volts negative, because the grid is 10 volts more negative than the cathode. Grid potential, to ground, is the same on tubes *B* and *C*, yet there is a 20-volt bias on *B* and only a 10-volt bias on C.

The plate of pentode *C* is connected through a resistor to the 280-volt positive point on the divider. An assumed drop of 30 volts in the resistor leaves 250 volts at the plate. Cathode potential is 90 volts negative. Then plate voltage or plate-cathode voltage is the sum of 250 volts and 90 volts, or is 340 volts.

The screen of the pentode is shown connected to ground. The zero reference voltage of ground is 90 volts more positive than cathode potential on this tube. So the screen is operating at 90 volts positive with reference to the cathode, although the screen is grounded. The screen is shown grounded only to illustrate what strange things may happen. Screens usually are operated at about the same voltages as plates in the same tubes or at a slightly lower or higher voltage.

As has been mentioned several times before, voltage measurements during service operations ordinarily are made with reference to chassis ground in the absence of specific instructions to the contrary. Most service diagrams show voltages with reference to ground. But unless you have a circuit diagram, or know the circuit, you can be sure of measuring effective plate, screen, and grid voltages only by measuring from the tube cathode. With one terminal of your voltmeter connected to the cathode of a tube you can measure actual effective voltages on the other elements, because all element voltages are considered with reference to the cathode so far as operation of the tube is concerned.

BYPASSING FOR VOLTAGE DIVIDERS. Fig. 22-6 shows paths of electron flows or currents for the same three tubes as connected to a voltage divider in Fig. 22-5. All the currents are considered to start from the negative center tap on the power transformer, and all come back through the filter choke and rectifier tube to the transformer winding.

Electrons for tube A pass upward through divider resistors *Rd* and *Rc*, thence through chassis ground to the cathode of the tube. This current goes from cathode to plate, and from the plate it returns through divider resistor *Ra*, as shown by arrows. The grid of tube *A* is biased by potential drop in divider resistor *Rc*, which is carrying plate-cathode current for the same tube.

Electron flow for tube *B* does not go through the lower part of the voltage divider, but proceeds along the bottom conductor to the cathode of this tube. From the plate of tube *B* the current returns to the filter by way of resistor *Ra*, in which we already have current from tube *A*. It is only the current for tube *C* that does not go through any parts of the voltage divider.

When currents for different tubes flow together in the same resistor, signals may travel in the wrong direction between these tubes. There may be backward signal travel also when plate and grid circuits are in series

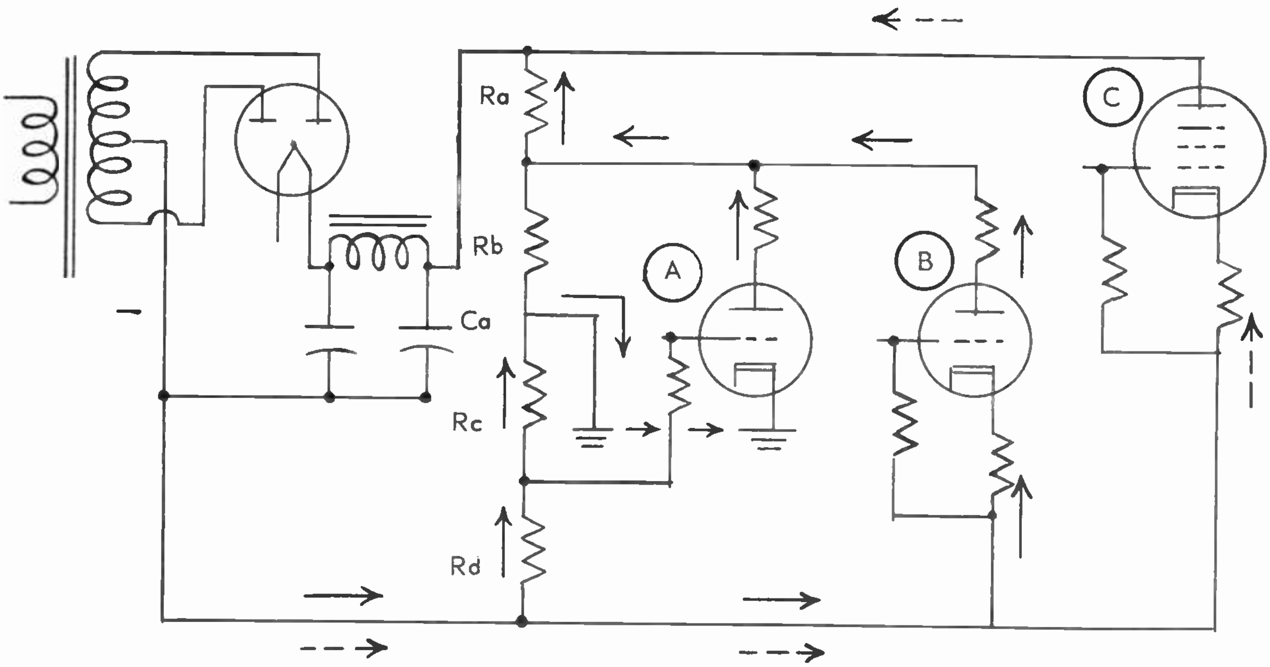


Fig. 22-6. Electron flows in tube circuits connected to the B-minus line.

with the same resistor. In Fig. 22-6 there is danger of back tracking and mixup of signals which are in tubes *A* and *B*. But the circuit for tube *C* does not go through any resistors that carry currents for the other tubes, so signals in tube *C* would be unlikely to cause trouble.

It is true that the circuit for tube *C* is connected across the entire voltage divider, and in this divider are currents for the other tubes. There is some possibility of signal mixup due to this connection, but the possibility is reduced practically to zero by the fact that across this tube circuit is connected capacitor *Ca* in the power supply filter. You will recall that a capacitor in this position maintains good voltage regulation. Good voltage regulation means reduction of alternating ripple voltage. This capacitor is equally effective in reducing or preventing alternating signal voltages in any lines connected to it. The capacitor doesn't care whether the alternations are ripple or signal, it gets rid of either or both.

To prevent trouble from signals in tubes *A* and *B* we must add filter capacitors across their circuits, but instead of calling the added units filter capacitors we call them bypass capacitors. These bypass capacitors must be connected to all resistors which may carry currents or voltages for more than one tube, and to all resistors which may carry both plate signal voltages and grid signal voltages. The by-passes must be connected from the resistors to the common negative side of the circuit, just like a filter capacitor.

The necessary bypass capacitors have been added in Fig. 22-7. Bypass *Cb* is connected from between resistors *Ra* and *Rb* to the negative side of the power circuit, which we may call the B-minus side. Bypass *Cc* goes from between resistors *Rb* and *Rc* to B-minus, with the connection completed through the two grounds

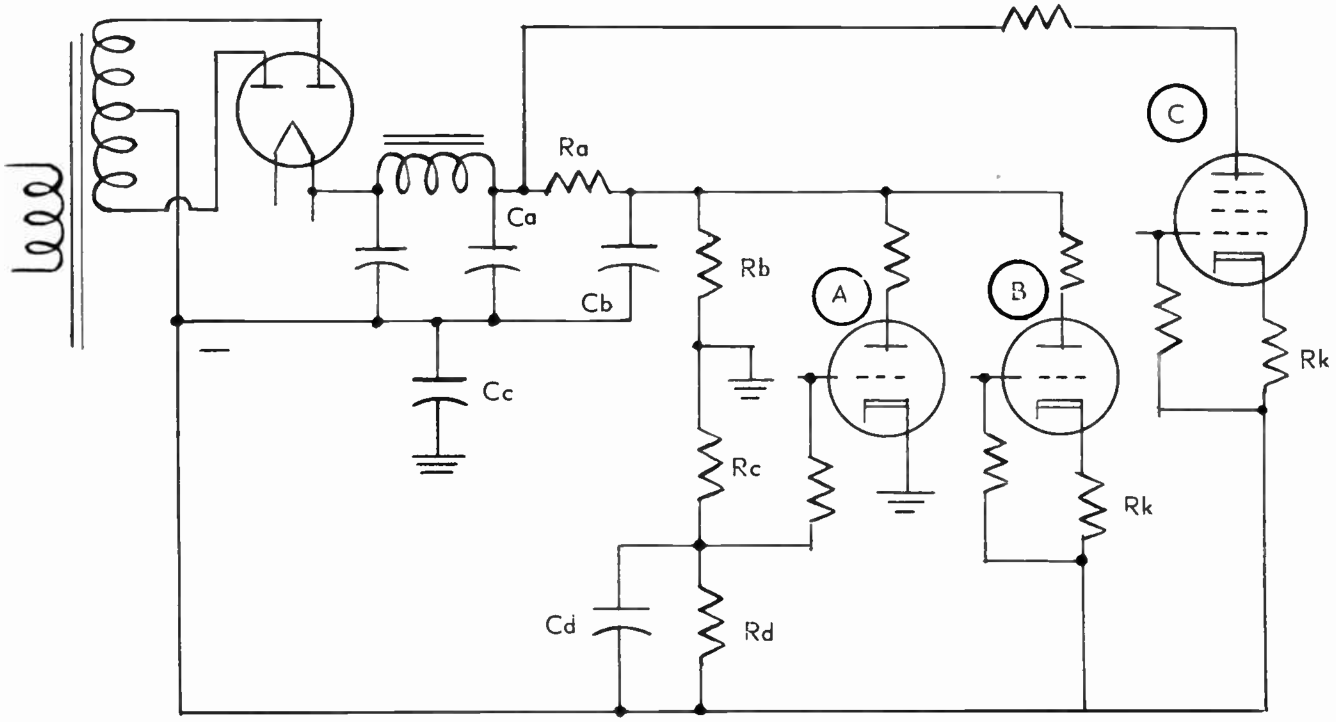


Fig. 22-7. Bypass capacitors added to the power supply circuits.

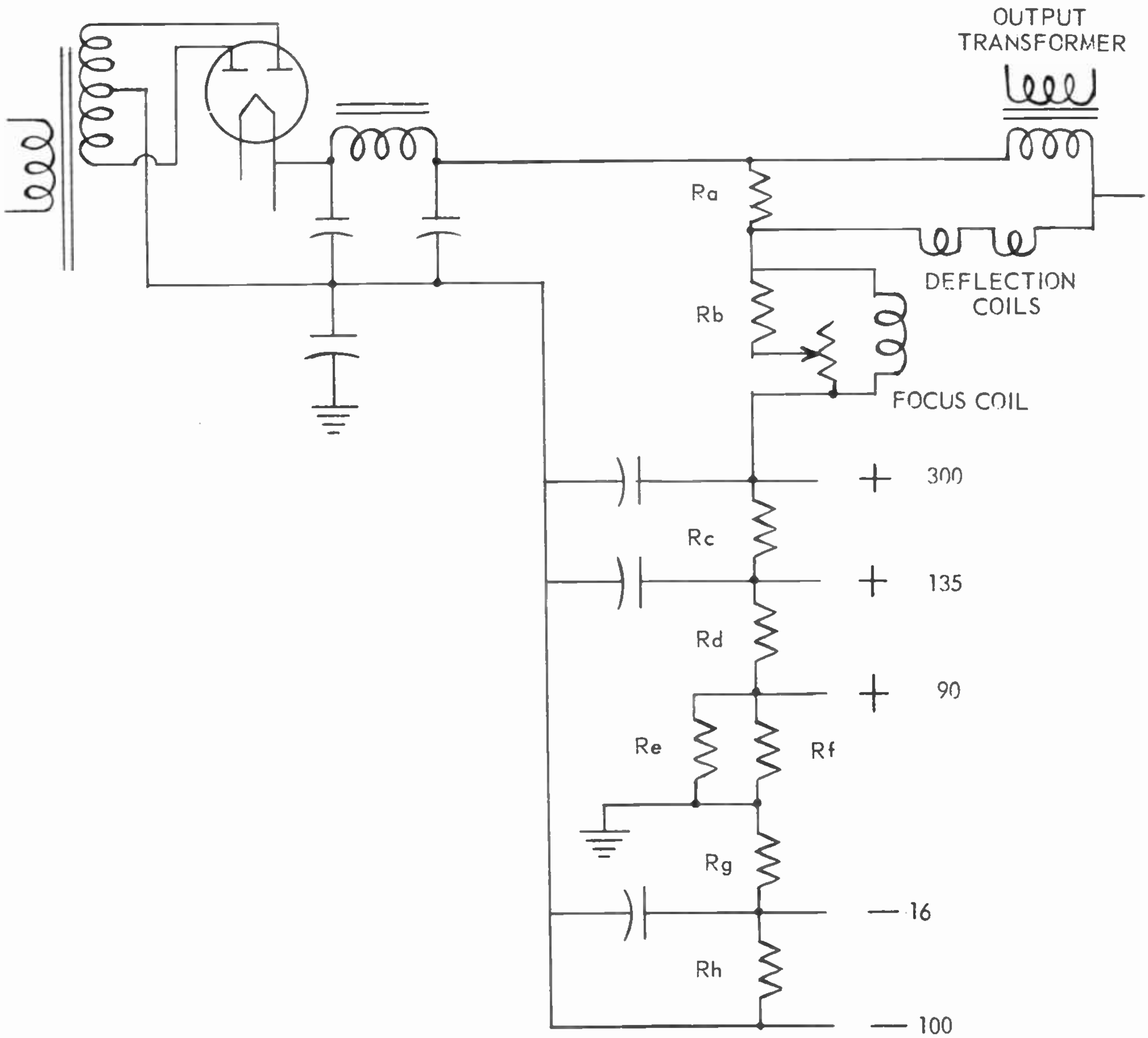


Fig. 22-8. Connections to a voltage divider system in a television receiver.

which are on the resistor tap and on capacitor *Cc*. Bypass *Cd* is connected from the top of resistor *Rd* to B-minus. If all these bypasses have sufficient capacitance they will absorb the variations of signal voltages and currents which might cause trouble, acting just as filter capacitor *Ca* acts for ripple.

In Fig. 22-7 resistor *Ra* is drawn in a position different than in earlier diagrams. With *Ra* in this new position, and with capacitor *Cb* in place, it is apparent that we have here what amounts to a resistor-capacitor filter section following the regular choke-capacitor section of the power supply. Almost any bypassed resistor in a voltage divider acts like an added resistor-capacitor filter section.

If *Ra* and *Cb* are considered as an added filter section, the plate of tube *C* is connected to a point between the first and second sections of the filter system. We talked about this method of connection to different filter sections in the lesson dealing with filters. More ripple would go to tube *C* than to tubes *A* and *B*, but this is all right in case tube *C* since it works from the amplified output of either of the other tubes. It might be a good idea to look back in the lesson on filters and examine the diagram showing this type of connection.

To a voltage divider system may be connected not only plate, screen, and grid biasing circuits, but also many of the circuits required for picture tube operation. Among the picture tube circuits so connected may be those containing deflection coils for magnetic deflection tubes, also focusing and centering controls for either magnetic or electrostatic deflection tubes. Fig. 22-8 shows a fairly typical divider system with deflection coils connected across resistor *Ra*, and with a focusing coil connected across resistor *Rb* and another resistor which is adjustable.

It is the practice of many manufacturers to furnish power supply service diagrams much like that of Fig. 22-8, showing each tap on the divider marked with its voltage to ground or else to B-minus, but omitting the connections which lead from the voltage divider taps to the various tubes. At each point on a tube circuit where, in the receiver itself, a connection would be made to one of these divider taps there will be marked on a complete service diagram the voltage of that particular tap. Omission of the many B-power connections from the diagrams allows you to follow the signal circuits with less danger of confusion.

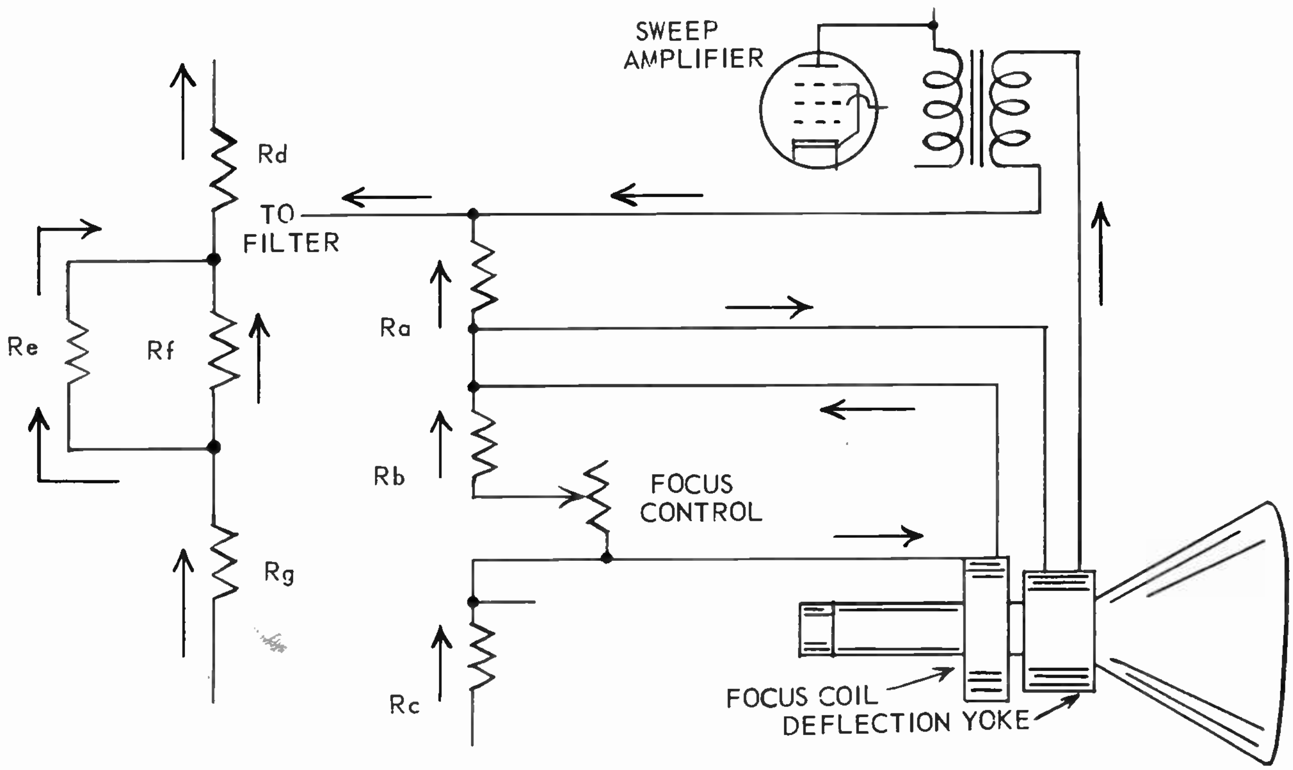


Fig. 22-9. Parallel connections in the voltage divider system.

Voltage values which are listed or marked on circuit diagrams are only approximate. Performance usually will be satisfactory if actual voltages are as much as 10 to 20 per cent higher or lower than the marked values.

8 *PARALLEL CIRCUITS.* By following the path of electron flow upward from the 100-volt negative line of Fig. 22-8 we find that this flow issuing from the top of resistor *Rg* divides into two parts. As shown separately at the left in Fig. 22-9, one part of the flow goes through resistor *Re* and the other part through resistor *Rf.* These two resistors are connected in parallel with each other. A parallel connection or parallel circuit is one in which the total electron flow or current divides between two or more paths.

Continuing on upward in Fig. 22-8 we find the two currents from resistors *Re* and *Rf* uniting and flowing through resistors *Rd* and *Rc*, which are in series. But the flow issuing from the top of resistor *Rc* again divides, with part of the flow taking a path through the focus coil and the other part following a path which goes through an adjustable resistor and the fixed resistor *Rb*. The focus coil and these resistors are connected in parallel.

At the right in Fig. 22-9 is shown the entire upper part of our voltage divider system, from the top of resistor *Rc* all the way to the connection for the power filter. The focus coil in a receiver would be mounted around the neck of the picture tube. The adjustable resistor connected across this coil, in series with resistor *Rb*, is the focus control whose adjustment allows obtaining good detail in picture reproduction. This control acts to send more or less of the total current through the focus coil.

Currents which have gone through the focus coil and the resistors in parallel with that coil come back together at the top of resistor *Rb*, but immediately divide again. Part of the total current now goes upward through resistor *Ra* while the other part goes through a deflection circuit for the picture tube. Resistor *Ra* is in parallel with this deflection circuit, or the resistor and the circuit are in parallel with each other.

The deflection circuit contains two deflection coils indicated by symbols in Fig. 22-8. These coils are part of the deflection yoke shown around the neck of the picture tube in Fig. 22-9. In series with the deflection coils in the yoke is the secondary winding of a coupling transformer whose primary winding is in the plate circuit of the sweep amplifier tube. Later we shall examine all these circuits which relate to picture tube operation. Just now we are interested only in their relations to the voltage divider system of this particular power supply.

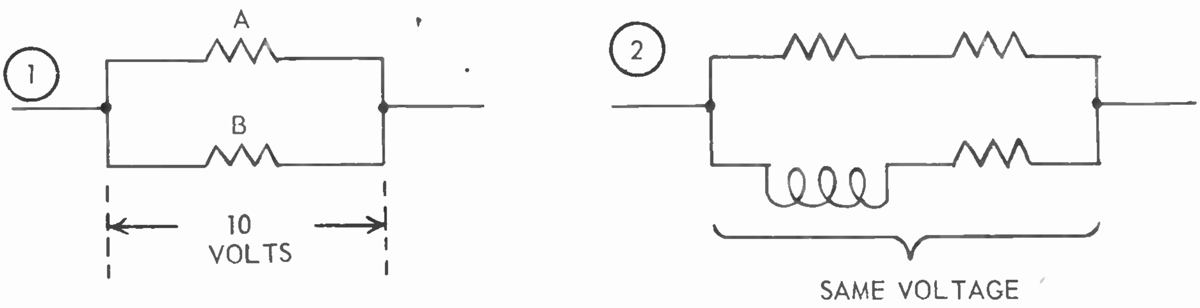


Fig. 22-10. Voltages across paths in parallel.

In Fig. 22-9 we have looked at only three of the many parallel connections found in all receivers, not only in power supply sections but everywhere else, and especially wherever there is some kind of adjustable control. It is just as necessary to understand the relations of voltage, resistance, and current in parallel circuits as in series circuits. Perhaps it is even more necessary, because service problems involving parallel connections often are more puzzling than those related to series circuits.

There are two possible ways of learning about parallel circuits. One way would be to commence with a detailed study of the fundamental principles relating to such circuits. This is a good way, but it takes a great deal of time and is not particularly interesting, because we have to learn and remember all the principles before watching them at work in television and radio receivers.

The other way is to make the briefest possible listing of the facts relating to parallel circuits, then keep this list handy as we later come to all the applications of parallel circuits in actual practice. Then, if it isn't already quite apparent how things are supposed to work, we may refer back to the list for help. This second way is the one we shall adopt. In the remainder of this lesson are the facts, and some handy formulas to go with them.

1. Voltage is the same across each of any number of paths in parallel.

Explanation: In diagram *1* of Fig. 22-10, if potential difference across resistor *A* is 10 volts it must also be 10 volts across resistor *B*, and across any other resistors in parallel, because both or all the resistors are connected directly together at their ends.

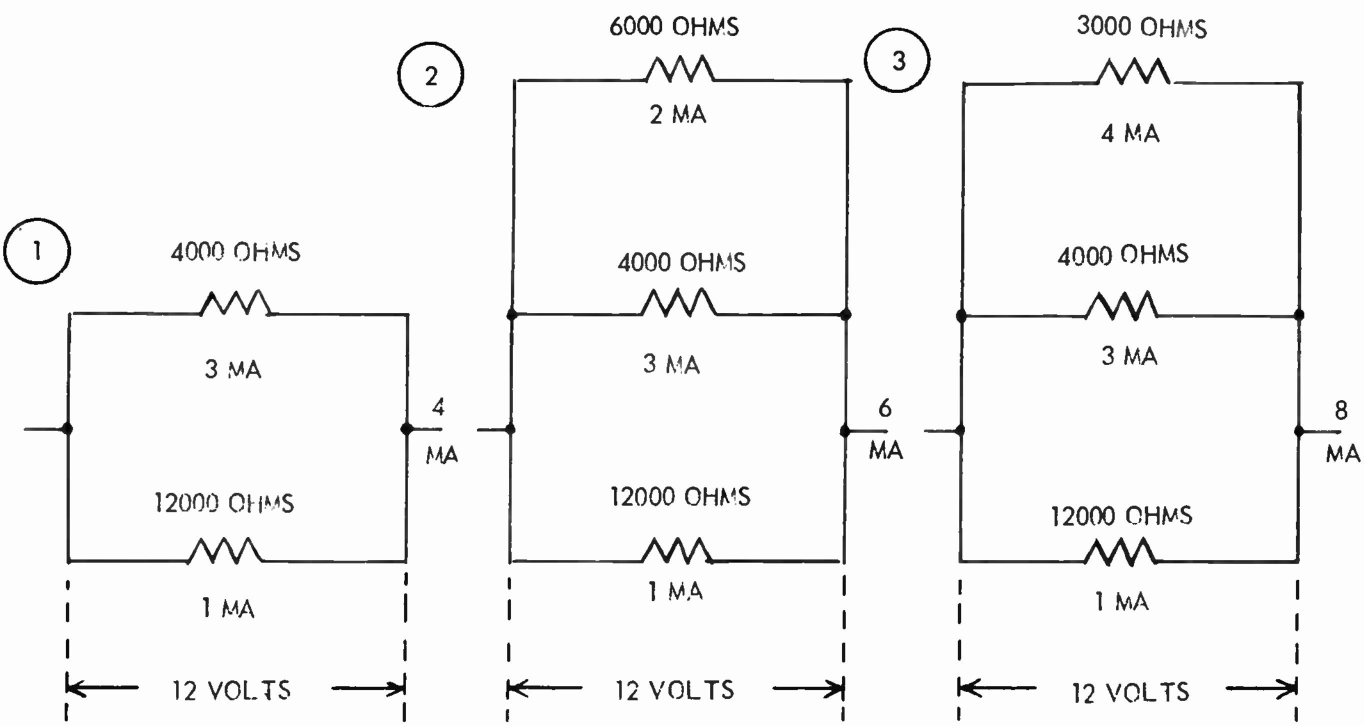


Fig. 22-11. Relations of voltage, resistance, and current in parallel paths.

2. Voltage, resistance, and current in any one parallel path have the same relations to one another as in a series circuit.

Explanation: At *2* in Fig. 22-10 the potentials at the ends of either path must be the same as at the ends of the other path. Then current in either path is equal to voltage divided by resistance, and resistance is equal to voltage divided by current, just as in a separate series circuit.

RESISTANCES IN PARALLEL. We shall continue with numbered facts relating to resistances in parallel with one another. What we shall call "combined resistance" is the opposition of all the parallel paths to the total current which will flow through all of them. In following explanations we shall be using our regular formulas for resistance, current, and potential difference – the same ones we use when working with series circuits.

3. Combined resistance always is less than the least resistance in any one of the paralleled paths.

Explanation: In diagram *1* of Fig. 22-11 the upper paralleled resistance is 4,000 ohms and the lower one is 12,000 ohms. Potential difference (across each) is 12 volts. Current in the upper resistance must be 3 ma and in the lower one 1 ma. Total current is 4 ma, with a potential difference of 12 volts. Our regular formula for resistance shows that 4 ma with 12 volts corresponds to 3,000 ohms resistance. This is the combined resistance of 4,000 ohms and 12,000 ohms in parallel. It is less than either of the separate resistances.

4. Adding an extra path in parallel reduces the combined resistance.

Explanation: In diagram 2 of Fig. 22-11 there has been added a third parallel path with resistance of 6,000 ohms. Potential difference across the circuit still is 12 volts. Then this added path will carry current of 2 ma. Total current in the three paths now is 6 ma. This current, with 12 volts, corresponds to 2,000 ohms –

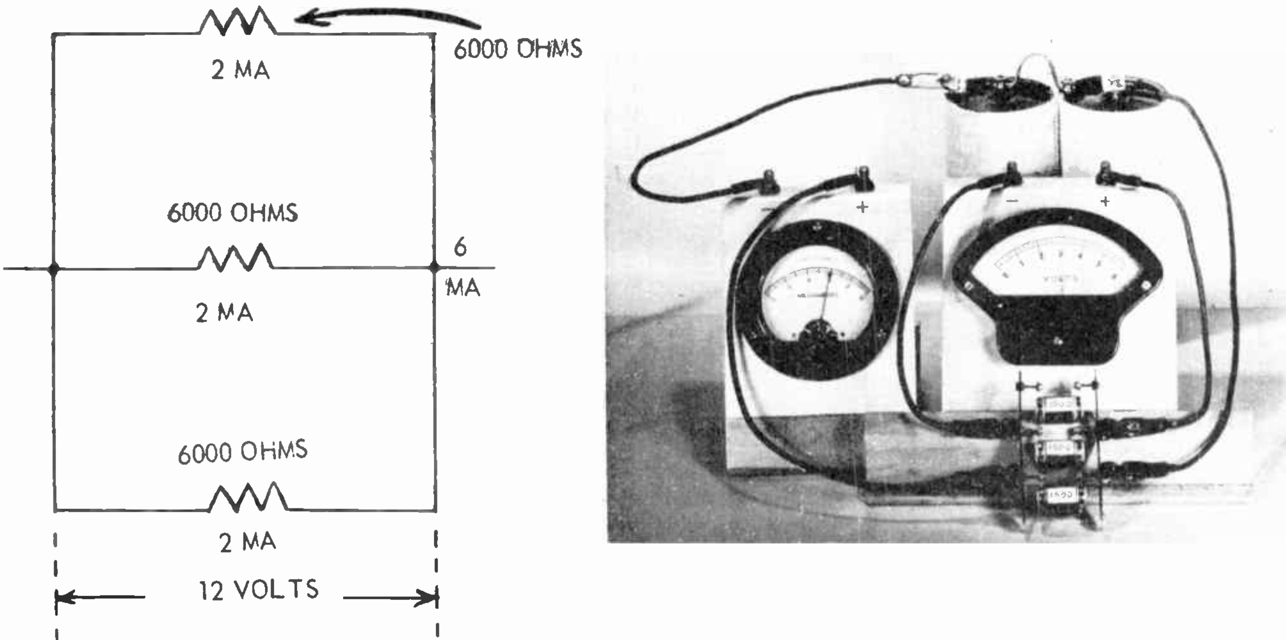


Fig. 22-12. Measuring combined resistance of equal resistances in parallel.

which is the combined resistance. Combined resistance has been reduced from 3,000 ohms to 2,000 ohms by adding a paralleled resistance.

5. *Combined resistance is altered by changing the resistance in any one path.*

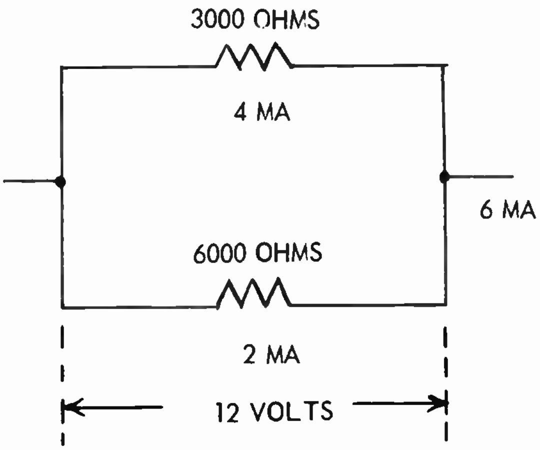
Explanation: In diagram *3* of Fig. 22-11 the upper resistance has been changed from 6,000 ohms to 3,000 ohms. With the 12-volt potential difference this upper path now carries current of 4 ma. Total current in the three paths is 8 ma. This current, with 12 volts potential difference, corresponds to 1,500 ohms combined resistance, whereas in diagram *2* the combined resistance is 2,000 ohms.

COMPUTING COMBINED RESISTANCES. Often it is necessary to compute the combined resistance of two or more paralleled resistances without going through the long reasoning processes used with Fig. 22-11. Following formulas help solve many service problems, especially when making replacements where you are unable to duplicate original combinations of resistances, but must have an effective substitute.

6. *Combined resistance of any number of paths whose resistances are alike is equal to the resistance of one path divided by the number of paths.*

Explanation: At the left in Fig. 22-12 are three paralleled resistances of 6,000 ohms each. Potential difference is 12 volts. Current in each path must be 2 ma. Total current through all three paths is 6 ma. This current, at 12 volts, corresponds to a combined resistance of 2,000 ohms, which is exactly one-third the resistance in any one of the paths.

At the right in Fig. 22-12 is pictured an actual measurement demonstrating the combined resistance of equal resistances in parallel. Three 1,500-ohm resistors are connected together at their ends by the vertical



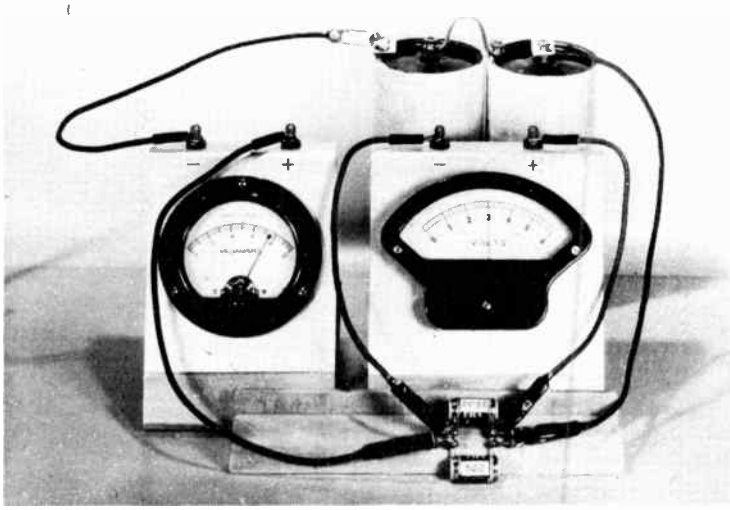
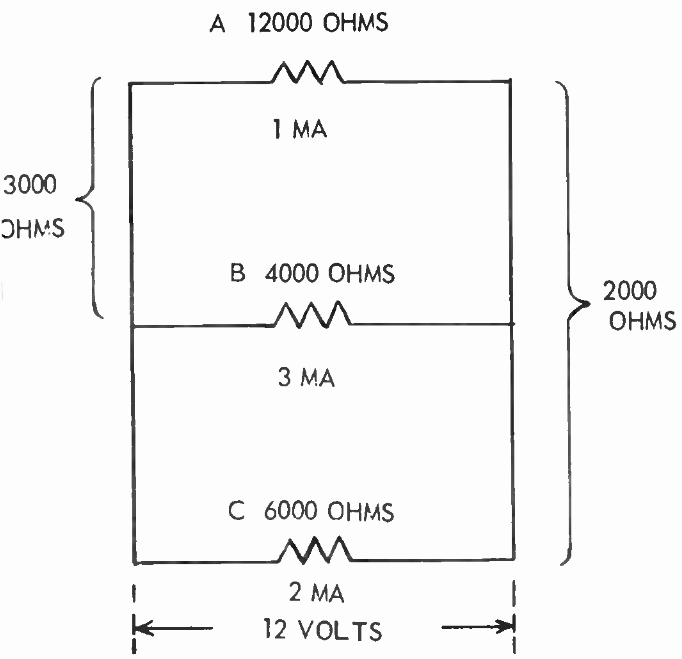


Fig. 22-13. Measuring combined resistance of two unlike resistances in parallel.



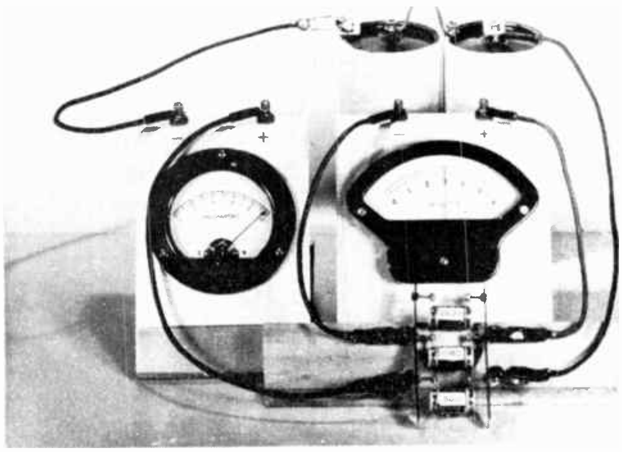


Fig. 22-14. Measuring combined resistance of three unlike resistances in parallel.

metal strips. To the parallel circuit is applied a 3-volt potential difference from two dry cells. The voltmeter indicates this potential difference. In series with the dry cells and paralleled resistors is a milliammeter indicating current of 6 ma. This current, with 3 volts, will give the combined resistance as 500 ohms, one-third the resistance of each unit.

7. *Combined resistance of two paths of equal or unequal resistances is found from dividing the product by the sum of the resistances.*

Calling the two resistances Ra and Rb, we have this formula.

$$\text{Combined resistance} = \frac{Ra \times Rb}{Ra + Rb}$$

At the left in Fig. 22-13 are paralleled resistances of 3,000 and 6,000 ohms. With 12 volts potential difference the currents are respectively 4 ma and 2 ma, for a total of 6 ma. This current, with 12 volts, indicates a combined resistance of 2,000 ohms. Using the separate resistance values in the formula gives,

$$\text{Combined resistance} = \frac{3000 \times 6000}{3000 + 6000} = \frac{18\,000\,000}{9\,000} = 2000 \text{ ohms}$$

Try the formula with resistances shown at *1* in Fig. 22-11, where we found combined resistance to be 3,000 ohms.

At the right in Fig. 22-13 the meters are used for computing the combined resistance of unequal resistances in parallel. The two resistances are 2,000 ohms and 500 ohms. Potential difference is 3 volts, as read on the voltmeter. The milliammeter reads 7.5 milliamperes. Using our regular formula for resistance (as employed for series circuits) shows that resistance corresponding to 3 volts and 7.5 ma is 400 ohms. Check this value of combined resistance by using the new formula for product divided by sum.

8. Combined resistance of more than two unlike resistances may be determined by first computing the combined resistance of any two, then using this value with a third resistance, and so on.

Explanation; At the left in Fig. 22-14 are paralleled resistances of 12,000 ohms, 4,000 ohms, and 6,000 ohms marked respectively *A*, *B*, and *C*. Using the product and sum method to determine combine resistance of *A* and *B* we have,

$$\text{Combined resistance} = \frac{12000 \times 4000}{12000 + 4000} = \frac{48\,000\,000}{16\,000} = 3000 \text{ ohms}$$

This combined resistance of 3,000 ohms was found correct with diagram *1* of Fig. 22-11, where we have the same two separate resistances.

Now we consider *A* and *B* as a single resistance of 3,000 ohms. Using the product and sum formula with this 3,000-ohm resistance and the third resistance of 6,000 ohms at *C* gives,

$$\text{Combined resistance} = \frac{3000 \times 6000}{3000 + 6000} = \frac{18\,000\,000}{9\,000} = 2000 \text{ ohms.}$$

This final combined resistance was found correct with diagram *2* of Fig. 22-11 where we have the same three separate resistances as in Fig. 22-14.

The meters are used at the right in Fig. 22-14 for computation of combined resistance of 2,000, 1,500, and 500 ohms in parallel. The voltmeter shows applied potential difference as 3 volts. The milliammeter indicates 9.5 ma, as nearly as can be read on its scale. This current, with 3 volts, corresponds to approximately 316 ohms combined resistance.

Using the product and sum formula to first compute the combined resistance of 2,000 ohms and 500 ohms, your answer should be 400 ohms. Then using the same formula with this 400 ohms combined resistance and the third separate resistance of 1,500 ohms should give a total combined resistance of approximately 316 ohms. You should use the formula to work out the resistance values for yourself.

CURRENTS IN PARALLEL CIRCUITS. Following are facts relating to current in parallel circuits. We continue our consecutive numbering.

9. Current is least in the greatest resistance, and is greatest in the least resistance.

This fact is easily verified by looking back at currents and resistances shown by Figs. 22-11 through 22-14.

10. Total current is equal to the sum of currents in the separate paths.

This fact is quite obvious, since the entire electron flow enters the paralleled paths at one end and leaves them at the other end of the circuit.

11. Total current is greater than current in any one path.

Another obvious fact. Naturally, no matter how little the current in any path, adding it to the other currents must increase the total.

12. Total current in paralleled paths is equal to applied voltage divided by combined resistance.

This is simply an application of our regular rule for current, as used with series circuits, stated in a way that applies to parallel circuits. We may write,

$$\text{Current, ma} = \frac{1000 \times \text{applied voltage}}{\text{combined resistance, ohms}}$$

EFFECTS OF RESISTANCE ON CURRENTS. Current in one parallel path is not altered by changing the resistance in another parallel path provided the overall voltage is not changed. In diagram *1* of Fig. 22-15 are parallel resistances of 3,000 ohms and 6,000 ohms, with 12 volts potential difference. Currents are respectively 4 ma and 2 ma. In diagram *2* the 6,000-ohm resistance has been replaced with 2,000 ohms, increasing the current in this lower path to 6 ma. But current in the 3,000-ohm resistance remains unchanged, because applied potential difference is not changed.

In diagram *3* of Fig. 22-15 the parallel resistances from diagram *1* are in series with 400 ohms. Combined

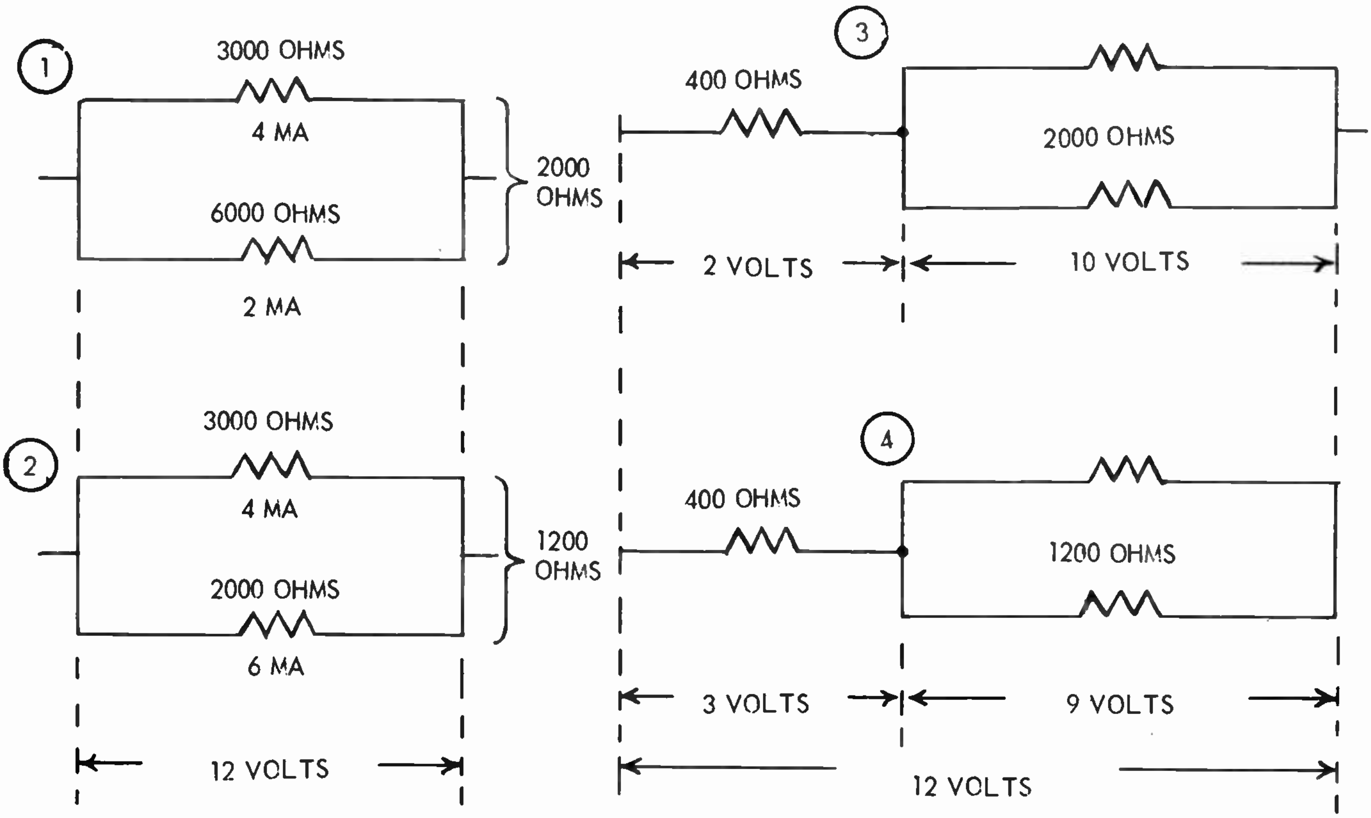


Fig. 22-15. The effects of resistance in series with a parallel circuit.

resistance of the two paralleled resistance is 2,000 ohms. The 12-volt potential difference now is applied across the entire circuit consisting of the 400-ohm resistance in series with the combined parallel resistance of 2,000 ohms. Potential difference divides proportionately to resistance, as we learned long ago, so there must be 2 volts across the added 400 ohms and 10 volts across the paralleled resistances.

In diagram *4* the parallel resistances from diagram *2* are in series with the 400-ohm added resistance. Combined resistance of the two paralleled units is 1,200 ohms. The 12-volt potential difference again is across the entire "series-parallel" circuit. Now the proportional division of potential difference gives 3 volts across the 400-ohm resistance and only 9 volts across the paralleled resistances.

1 Since potential difference across the paralleled resistances of diagram *4* is different than in diagram *3*, currents in the paralleled resistances will be different in the two diagrams. Altering the resistance from 6,000 to 2,000 ohms in the lower paralleled path has changed the currents in both paths, but only because there has been a change of the potential difference applied across the paralleled resistances.

DO NOT TEAR — CUT ALONG THIS LINE AND SEND IN FOR GRADING.

TELEVISION

LESSON NO. 23

VOLTAGE DIVIDERS AND SERIES HEATERS

Training Your Mind...
C. T. I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind
C.T.I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON NO. 23

VOLTAGE DIVIDERS AND SERIES HEATERS

In the first method of voltage division which we examined in the preceding lesson the high voltage from the power supply is dropped in the audio output amplifier tube, and the remaining lower voltage is applied to plates and screens of other tubes in the receiver. This particular method of voltage division is a fairly recent development. The chief advantage is a reduction of power supply current. With the same current or same electron flow passing through two groups of tubes, the demand on the power supply is only as great as for one of the groups alone. This allows using a smaller power supply for a given number of tubes, or allows reduction of ripple voltage in any case.

This method of voltage division involves the principles of parallel circuits. Since we now are acquainted with these principles we may make a more complete examination of how this system operates.

Power supply circuits as used in several television receivers are shown by Fig. 23-1. The positive terminal of the power supply filter connects through conductors drawn with heavy lines to all three audio tubes, also to the video and sweep amplifiers, and to the r-f oscillator which is in the tuner section of the receiver. From the cathodes of the three audio tubes are obtained plate and screen voltages and currents for the three video i-f amplifiers, for a sync amplifier, and also for the r-f amplifier and mixer tubes in the tuner section.

Tubes not shown on this diagram are supplied with plate and screen voltages and currents from the high-voltage power supply. The high-voltage supply, in addition to caring for the needs of the picture tube, furnishes plate and screen voltages and currents for two sweep oscillators, another sweep amplifier, and a sync amplifier.

The general scheme of voltage and current division from the low-voltage power supply is shown in Fig. 23-2. Here each of the tubes is represented by a circle. Positive and negative terminals of the power supply are at the left. On the simplified connections between the tubes, and to the power supply, are arrows indicating directions of electron flow.

Electrons are considered to start from the negative terminal of the power supply and to flow into chassis metal through the ground connection. From ground connections shown along the bottom of Fig. 23-2 there is electron flow to the cathodes of all tubes whose cathodes are connected directly or through a resistor to ground. All these connections may be seen on the circuit diagram of Fig. 23-1.

Electron flows which have passed through six tubes come together at point *a* on the diagram. Part of this combined electron flow goes through resistor *Ra* and returns to positive of the power supply along the top line of the diagram. The remainder of the combined flow goes to the cathodes of the three audio tubes, and from plates and screens of these tubes returns to the positive of the power supply along the top line. Electron flows for the r-f oscillator, video amplifier, and sweep amplifier go from ground to the cathodes of these

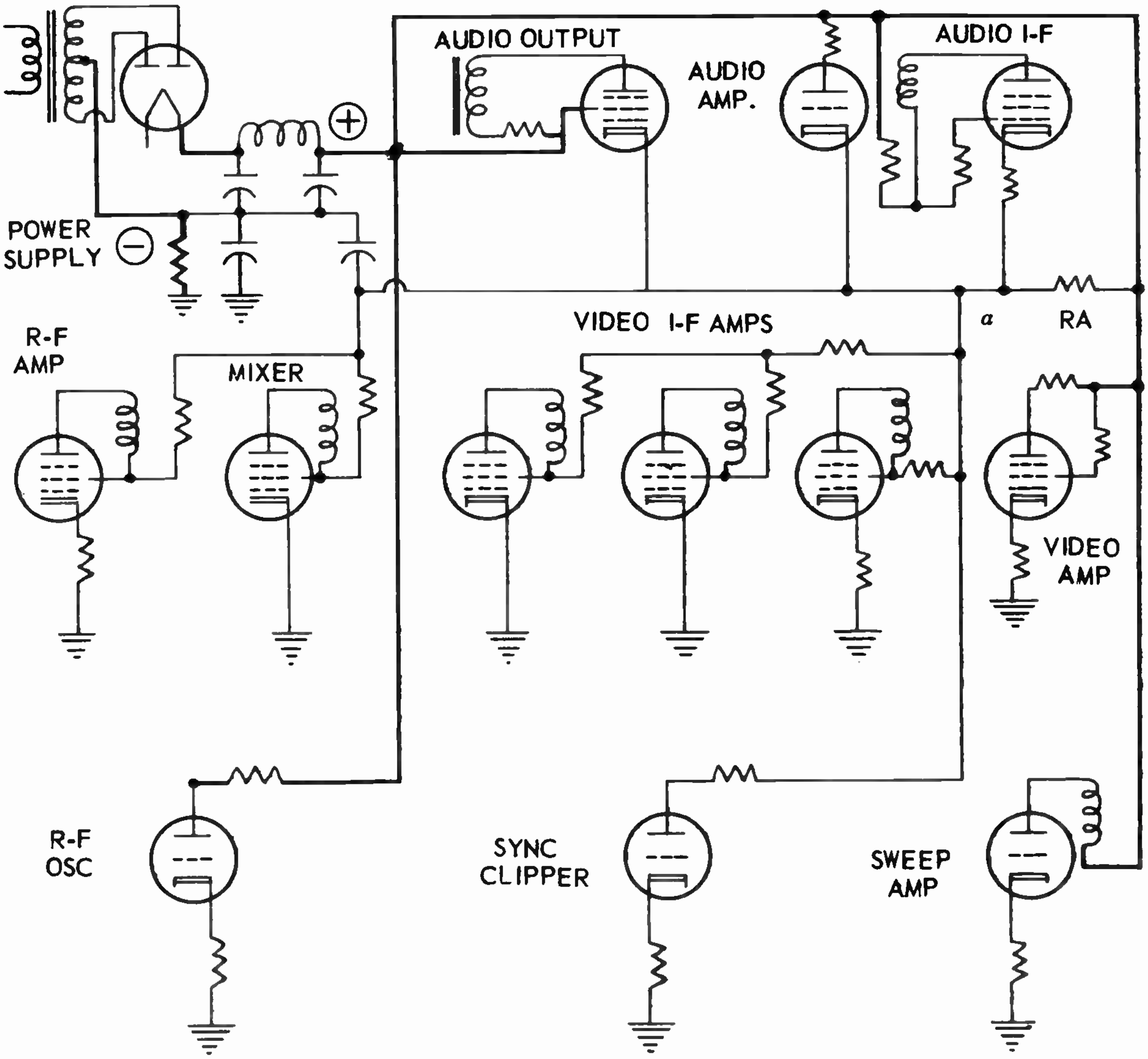


Fig. 23-1. Television power supply system in which audio-tube plate circuits are in series with plate circuits for video and r-f tubes.

tubes, and from their plates and screens the electrons return to positive of the power supply through the top line.

The six tubes whose electron flows join at point *a* are in parallel with one another. Current coming to point *a* then is the sum of plate and screen currents from all six tubes. Although the potential difference between point *a* and ground is applied to the circuits of all six tubes, these tubes may be operated at various voltages by utilizing potential drops in resistors shown in the diagram of Fig. 23-1.

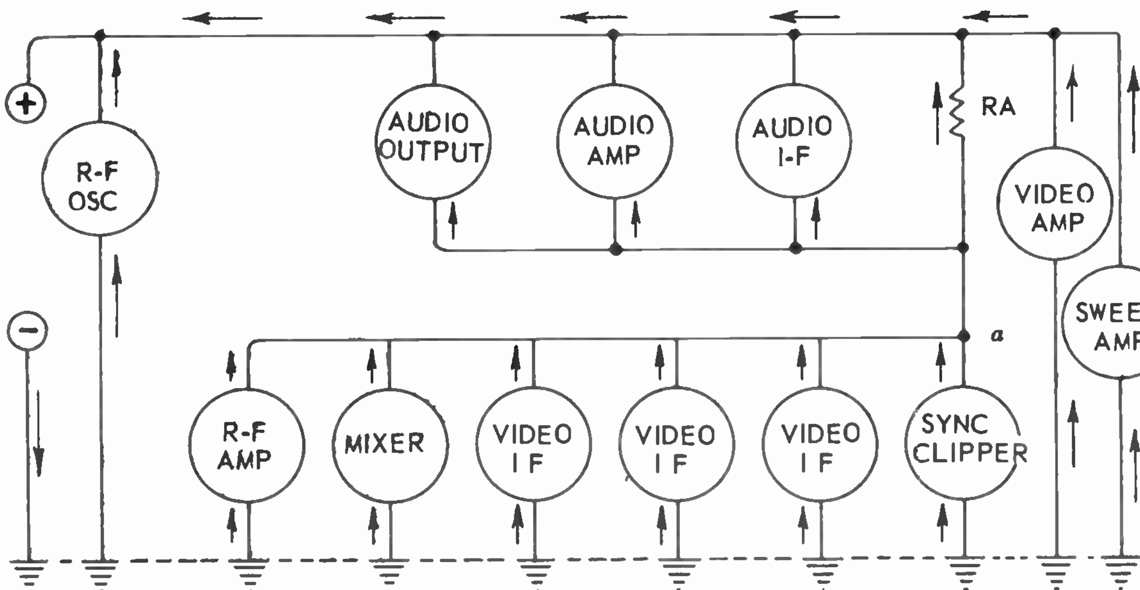


Fig. 23-2. Electron flows for tubes whose plate circuits are connected in series and parallel.

The three audio tubes are in parallel with one another and with resistor *Ra*. The total current in the three audio tubes and in *Ra* must be the same as the total current coming to point *a* from the six tubes down below. The potential difference which is across resistor *Ra* must be the same as applied to the circuits of each of the audio tubes, but plate and screen voltages for these tubes may be varied by means of resistors connected to each tube circuit.

Since the r-f oscillator, video amplifier, and sweep amplifier are each connected between positive and negative of the power supply the circuits of these tubes are subjected to the full supply voltage. Plate and screen voltages suitable for each of these three tubes are secured by voltage dropping resistors in their circuits. All this may be seen on the diagram of Fig. 23-1.

In the voltage division system which we are examining the high potential from the positive of the power supply to ground is 360 volts. The drop in the audio tube circuits is 220 volts. This leaves the potential on the cathode line for these three tubes at 140 volts positive with reference to ground. Grid voltages must be negative with reference to this cathode potential in order to have negative grid biases, but grid voltages still will be highly positive with reference to ground.

Methods of grid biasing are illustrated by Fig. 23-3. At the audio output tube there is a connection from the 360-volt supply line through resistors *Rb* and *Rc* to ground. These two resistors form a voltage divider across the top and bottom of which is the entire potential difference from 360 volts to ground. Part of this total voltage is dropped in resistor *Rb* and the remainder in *Rc*. The part dropped in resistor *Rb* is 235 volts, which leaves 125 volts at the top of *Rc*. This remaining 125 volts is dropped in resistor *Rc*. The resistances of the two voltage divider resistors are so proportioned as to divide the total voltage in this manner.

The 125-volt point on the voltage divider *Rb-Rc* is connected through resistor *Rg* to the grid of the audio output amplifier. The cathode of this tube connects directly to the 140-volt line. Consequently, the grid is 15 volts less positive than the cathode, and is effectively 15 volts negative with reference to the cathode.

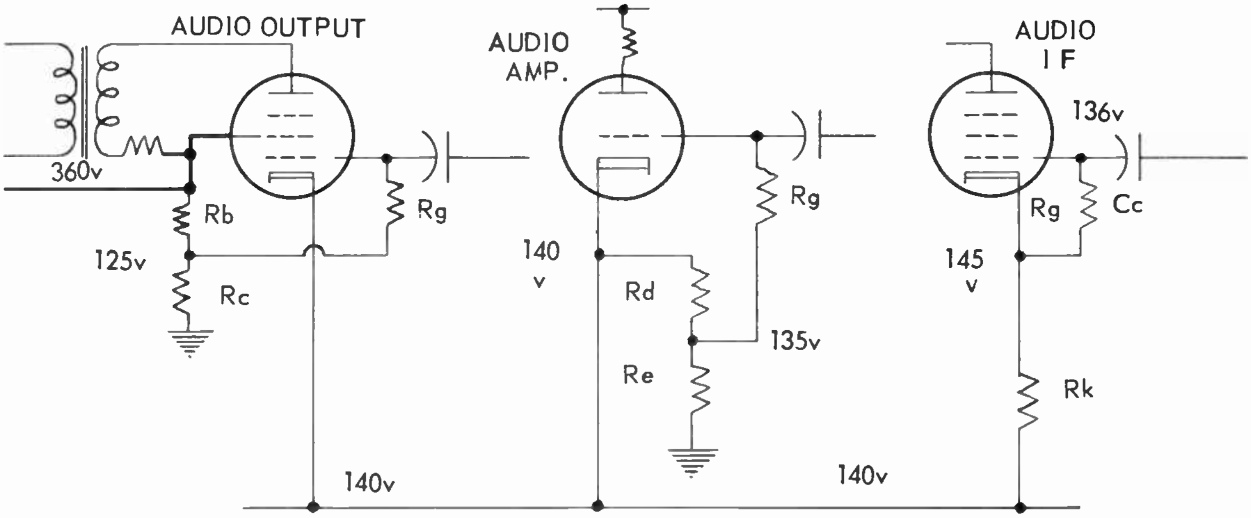


Fig. 23-3. Grid biasing for audio tubes whose cathodes are connected to plate and screen circuits of other tubes.

This is a 15-volt negative grid bias for this amplifier. Since the grid is always negative with reference to the cathode there is no current in resistor *Rg*. With no current there is no difference of potential across *Rg*, and at the end connected to the grid the potential is the same as at the end connected to the voltage divider.

The audio amplifier tube is biased by a similar method. Here the voltage divider resistors are *Rd* and *Re*. They are connected between the 140-volt line and ground. Only 5 volts out of the total 140 volts to ground is dropped in resistor *Rd*, which leaves 135 volts at the top of *Re*. The grid of the audio amplifier tube is connected through its grid resistor *Rg* to the 135-volt point on the voltage divider. The cathode is connected directly to the 140-volt line. Consequently, the grid is 5 volts less positive than the cathode, or is effectively 5 volts negative with reference to the cathode, and this tube is provided with a 5-volt negative grid bias. Again there is no current and no difference of potential in resistor *Rg*, because the grid always is negative with reference to the cathode.

The audio i-f amplifier tube, at the right in Fig. 23-3, is biased by the grid-leak method. The cathode of this tube is connected through resistor *Rk* to the 140-volt line. Accompanying the plate and screen currents in this resistor is a potential difference of 5 volts, which places the cathode at 145 volts positive. Grid-leak bias by means of grid resistor *Rg* and coupling capacitor *Cc* will make the grid more or less negative with reference to the cathode as signal strength increases and decreases. Grid potential will average about 136 volts with reference to ground. This is 9 volts less positive than the cathode, so we have a 9-volt negative grid bias.

Although resistor *Rk* on the audio i-f amplifier is in series with the cathode it is not a cathode-bias resistor because the grid return is not to the low end of this resistor. It is evident that bias is by means of grid leak action because the leak resistor *Rg* is connected directly to the cathode of this tube.

VOLTAGE REGULATION. When total voltage from the power supply divides between two groups of tubes, as in the system we are examining, the audio output tube acts to maintain nearly constant plate and screen voltages for the tubes down below. This voltage regulating action is explained as follows.

9 To begin with, practically all receivers have what is called automatic gain control, which makes the r-f and video i-f amplifiers more sensitive to weak signals than to strong ones and thus avoids the need for frequent manipulation of the contrast control. This automatic gain control operates to increase the plate currents when signals are weak. The increase of plate current tends to lower the plate voltage, as we learned when studying the behavior of triodes and pentodes.

Any tubes which, as at *1* in Fig. 23-4, are in series across the practically constant output voltage of the power supply are equivalent to resistances in series, as at *2*. So far as electron flow is concerned the tubes really are resistances, they are plate-cathode resistances. Now, with the tubes considered as resistances, it is plain that we have a voltage divider across the power supply. Division of voltage is such as to provide 140 volts at the cathode line.

Should a change of signal strength make the lower tubes decrease their plate voltage to 135 volts we would have the condition shown by diagram *3*. Our original voltage division has been upset because, in effect, the lower tubes have decreased their plate cathode resistances. What can be done to restore the former division of voltage?

To answer this question let's first find the real reason for lessened voltage on the cathode line. The reason is that much of the greater current being taken by the r-f and video i-f amplifier tubes, or the bottom resistance, goes through the resistance of the audio output amplifier. More current increases the voltage drop across this upper resistance, or the audio output tube. This greater drop across the upper resistance or tube leaves less voltage at the cathode line and at the plates of the r-f and video i-f amplifiers.

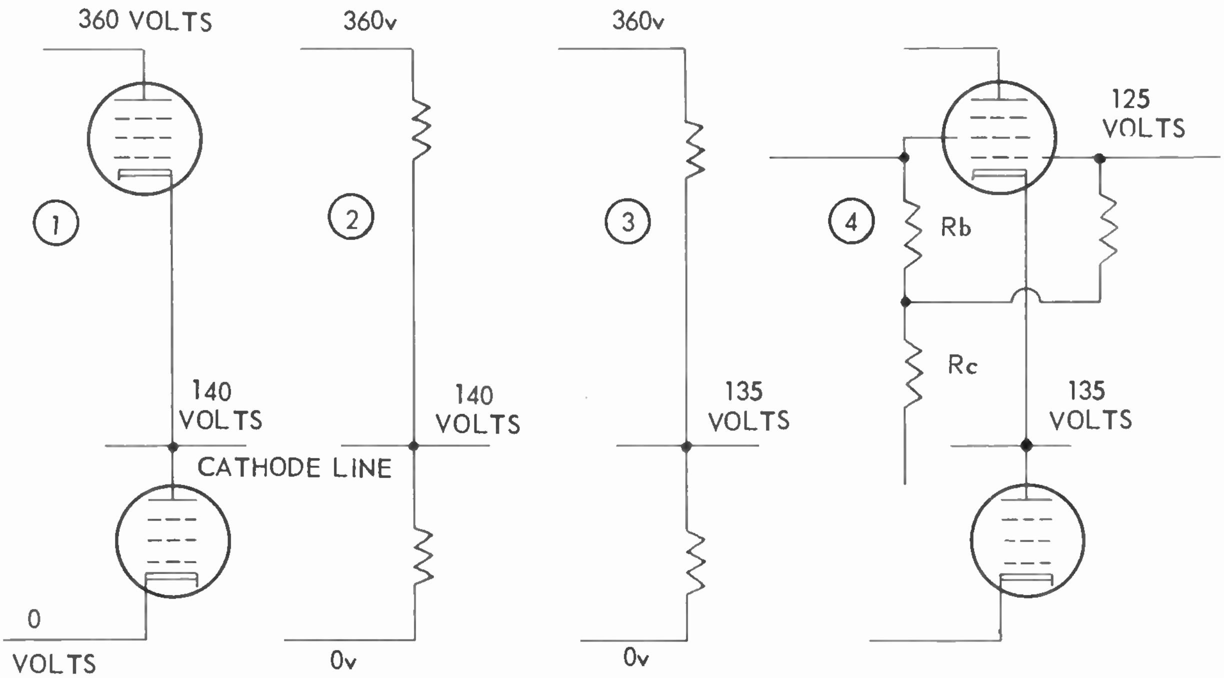


Fig. 23-4. An audio tube may provide automatic voltage regulation for tubes whose plates are fed from the audio tube cathode.

To correct the condition we must lessen the upper resistance, or the resistance of the audio output tube, so that there will be less voltage drop even with the greater current in this tube. To lessen the tube resistance we may make its grid bias less negative. This will allow more plate-cathode current and will be equivalent, in effect, to less tube resistance.

Grid bias on the audio output amplifier is changed as illustrated at *4* in Fig. 23-4. Voltage on the grid remains constant, because it is being taken from the constant voltage of the power supply at resistors *Rb* and *Rc* of Fig. 23-3. But, in our example, cathode voltage on this tube has decreased from 140 volts to 135 volts. With the grid remaining at 125 volts there will be a negative bias of only 10 volts instead of the former 15 volts. This allows more current through the audio output tube.

The extra current required in the tubes down below now flows through the lessened resistance of the audio output tube. We have lowered the resistance of the audio output tube to match the lessened resistance in the r-f and video i-f amplifiers. This has restored the proportions of our voltage divider resistances to bring the cathode line back to 140 volts.

All the actions which have been described occur together, and voltage regulation is almost instantaneous. The potential on the cathode line and on the plates of the lower tubes won't change more than a volt or two when switching from a strong signal to a weak one.

HEATER CIRCUITS. In voltage division systems for B-power supplies we have come across a number of parallel circuits, also many paralleled parts which are in series with other parts. Parallel circuits are used not only in B-power distribution systems but also nearly everywhere else in television and radio receivers. A particular case relating to power supplies in general is that of the circuits for tube heaters. Here we find parallel arrangements, also many series heater circuits and circuits where two or more series heater "strings" are in parallel with one another across the power line. All these combinations will be examined.

HEATERS IN PARALLEL. Heaters for the tubes in the majority of television and radio receivers are

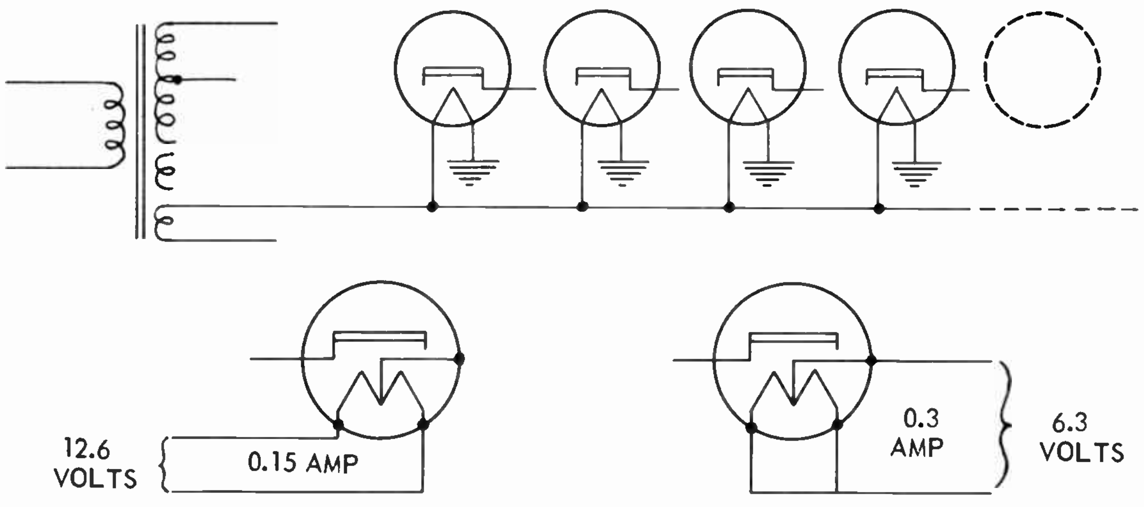


Fig. 23-5. Tube heaters in parallel.

connected in parallel with one another across a secondary winding provided for this purpose on the power transformer. The connection scheme is shown by Fig. 23-5. One side of each tube heater is connected through insulated wire to one end of the transformer winding. The other sides of all the heaters and the other end of the transformer winding are grounded to chassis metal, through which is completed the heater circuit to each tube.

5 All the paralleled heaters are subjected to the same voltage, which is that furnished by the transformer winding. The total current in this winding is the sum of the currents in all the heaters connected to it. All heaters connected in parallel must be rated for operation at the same voltage, which most often is 6.3 volts. The heaters need not be rated for equal currents, since each heater may take from the transformer whatever current that heater is designed to use at the applied voltage.

Wiring connections for parallel heaters seldom are drawn on service circuit diagrams. The ungrounded end of the heater winding at the transformer will be marked either with its voltage or with some symbol such as the letter "X". A similar marking is placed at the ungrounded heater terminal of each tube symbol, or, if no heater terminals are shown, there will be a note mentioning how connections are made in the receiver wiring.

Some tubes have heaters designed for operation on either 12.6 volts or 6.3 volts, as shown at the bottom of Fig. 23-5. The heater inside the tube is made in two sections with a tap from the mid-point connected to

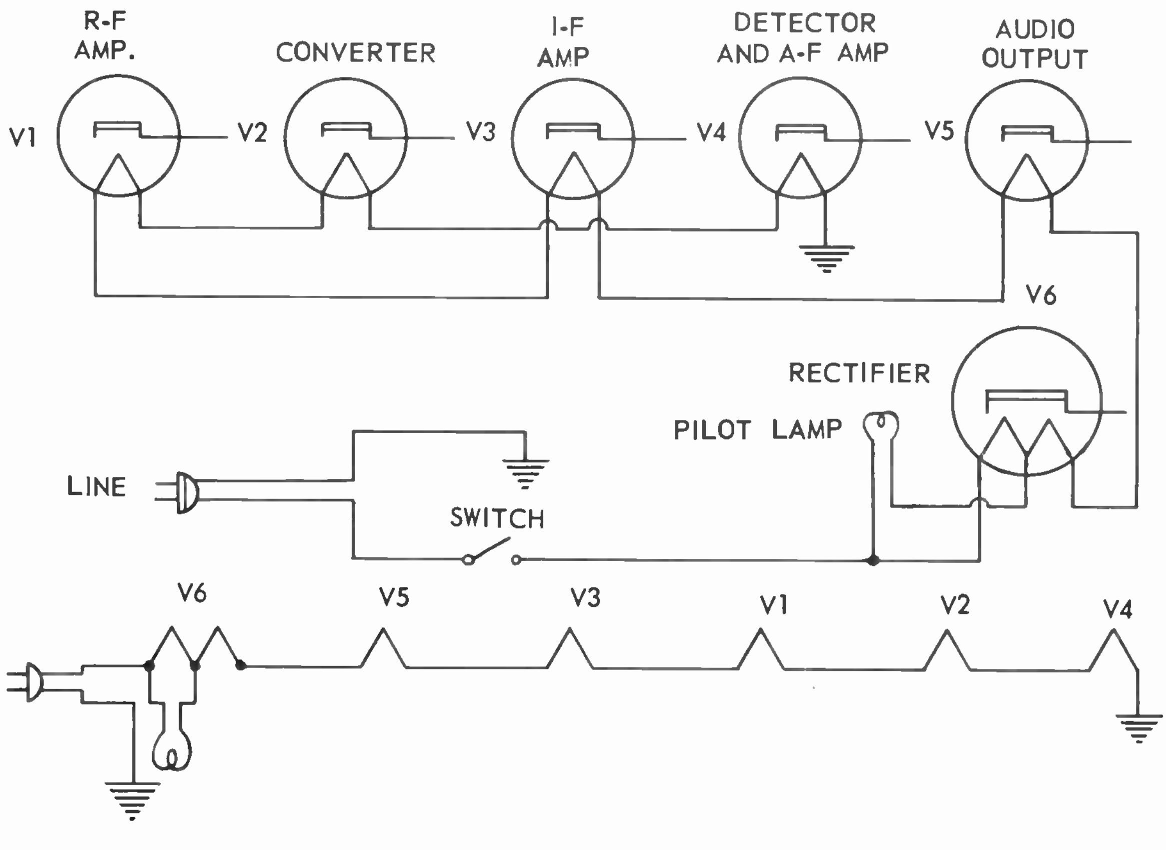


Fig. 23-6. Series heaters in a sound radio receiver.

a base pin. For operation on 12.6 volts only the outer ends of the heater are connected to the supply, so the two sections are in series. Current through the heater then is 0.15 ampere. For operation of 6.3 volts the heater sections are connected in parallel with each other, as at the right. The outer ends are connected together and to one side of the supply, and the center tap is connected to the other side of the supply. Current then is 0.3 ampere, with half this current, or 0.15 ampere, flowing in each section of the heater.

SERIES HEATERS. Transformerless receivers or ac-dc receivers, both sound radio and television, often have the tube heaters connected in series with one another across the a-c or d-c power line. Heaters connected in a series circuit are selected from such rated operating voltages that the sum is equal to or nearly equal to the voltage from the power line. Any excess of line voltage is used up by a resistor connected in series with the heaters and the line.

Fig. 23-6 shows series heater connections for a six tube sound radio receiver. This is a fairly typical heater circuit for radio receivers. From one side of the power line there is a connection through the on-off switch to the heater of the rectifier tube. In parallel with part of this heater is the pilot lamp. From the far side of the rectifier heater the circuit goes, in order, through the audio output tube, the i-f amplifier, the r-f amplifier, the converter tube, the combined detector and a-f amplifier, thence to ground. The side of the power line which does not go to the switch is grounded, thus completing the heater circuit through chassis ground.

Heater connections such as shown at the top of Fig. 23-6 ordinarily are omitted from service diagrams. The order of connections then is shown by a small separate diagram of the general style shown below. On this small diagram the heater "string" is drawn along a straight line, and the tubes are identified either by name or else with the same tube numbers found on the main diagram showing all other circuits.

The order in which the heaters are connected along the series circuit from power line to ground in sound radio receivers is chosen to reduce the likelihood of audible hum at the power line frequency. The tube most likely to pick up hum voltage is the detector, so the detector is nearly always at the grounded end of the heater line. The rectifier, whose plate-cathode circuit always carries voltage at line frequency, has its heater at the line end of the heater circuit. Remaining tubes are connected in various orders, according to the ideas of the set designer.

In the circuit of Fig. 23-6 might be tubes whose heaters are designed for the following voltages and currents.

R-f amplifier	12.6 volts	0.15 ampere
Converter	12.6 volts	0.15 ampere
I-f amplifier	12.6 volts	0.15 ampere
Detector and a-f amplifier	12.6 volts	0.15 ampere
Audio output	35.0 volts	0.15 ampere
Rectifier	35.0 volts	0.15 ampere
Totals	120.4 volts	0.15 ampere

In a series circuit the overall voltage is the sum of the separate voltages in all the parts, and the total current is the same as the current in each part. The overall voltage, 120.4, is within 3 per cent of standard line voltage, which is 117.0. On this standard line voltage, each heater would operate at about 3 per cent under its rated voltage. In some receivers you will find heaters operating at anything from 15 per cent under to 15 per cent over their rated voltages, although limits of 10 per cent under to 5 per cent over are considered much better practice. Tubes are designed to operate satisfactorily within these latter limits of heater voltage.

All parts in a series circuit must carry the same current. In Fig. 23-6 all tube heaters are of types designed to operate with current of 0.15 ampere. All the tubes might be of types having 0.3-ampere heaters, and the requirement of equal currents would be satisfied. Should you place a 0.15-ampere heater in a string designed for 0.3-ampere operation the low-current heater would burn out. With a 0.3-ampere heater in a string designed for 0.15-ampere operation the high-current heater would remain too cool and the tube would not operate correctly, if at all.

Fig. 23-7 is a circuit diagram for series heaters in a television receiver. The line current is alternating, but for purposes of explanation we may consider the current as flowing momentarily from the top lead in the line plug at the left. Total current divides at point *a*. Part of the current flows on upward through the 130-ohm fixed resistor, thence to the right through the heaters of eight tubes in series. Up above the V-shaped symbol for each tube heater is the name of the tube, the rated voltage of that heater, and the resistance of the heater. All these heaters are designed to operate with current of 0.3 ampere. Their resistances are determined by using our regular resistance formula, with this value of current and with the voltages shown for each separate heater.

The other part of the current leaving point *a* flows to the right through a 75-ohm fixed resistor and through the heaters of seven tubes in series. Below the heater symbols are the tube names, also the heater voltages and resistances.

Currents from the upper and lower series strings come together at their right-hand ends, in the conductor marked *b*. With 0.3 ampere from the upper string and another 0.3 ampere from the lower string of series heaters the total current coming to *b* is 0.6 ampere. This total current flows through the picture tube heater, which is rated to operate with 0.6 ampere and 6.3 volts.

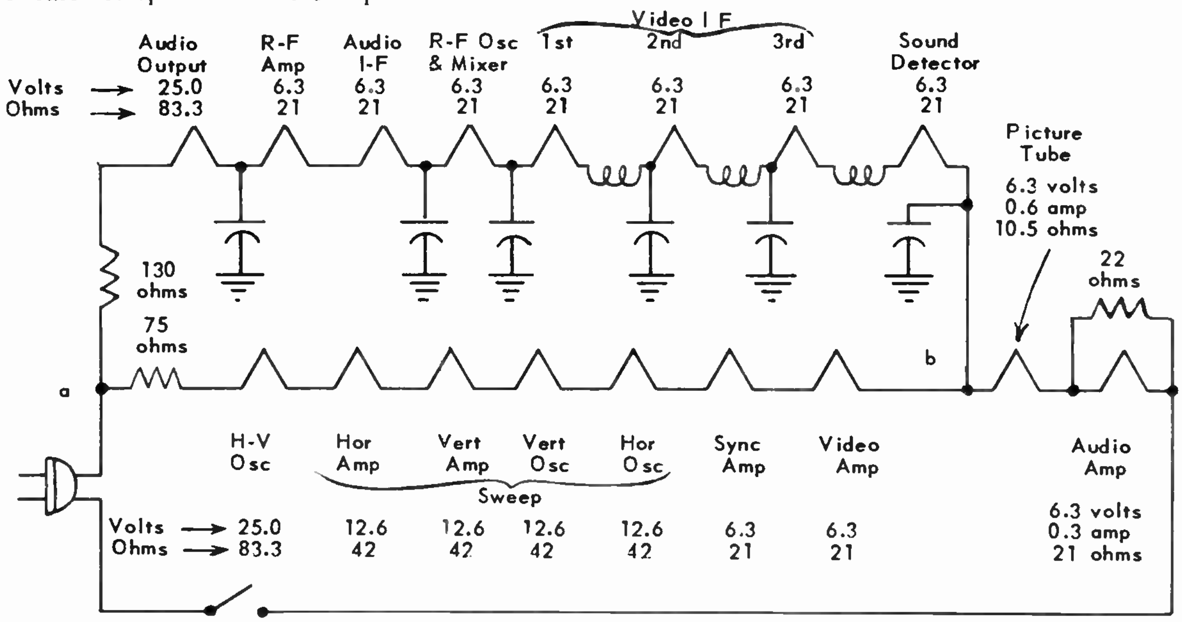


Fig. 23-7. Two series heater strings in parallel with each other, and in series with heaters for two tubes.

The next heater toward the right, which is the final one, is in the audio amplifier. The heater of this amplifier is designed for 0.3 ampere at 6.3 volts. Since 0.6 ampere is coming from the picture tube heater, it is necessary to connect in parallel with the audio amplifier heater a resistance to carry the extra current. This resistance is shown as 22 ohms, which is one of the standard resistance values. The 0.6 ampere current will divide between the 22 ohms of resistance and the 21 ohms in the tube heater. Since the two resistances are approximately equal, the currents in them will be approximately equal, and there will be 0.3 ampere in the audio amplifier heater.

The total current of 0.6 ampere returns to the power line through the conductor at the bottom of the diagram and through the on-off switch.

From various points along the upper string of heaters in Fig. 23-7 are connected capacitors to ground. These are bypass capacitors whose purpose is to prevent signal voltages which are in one tube from getting into other tubes by way of the heater circuit. Although the heater is separated from the cathode by insulation which prevents leakage of direct currents and potentials, this insulation combined with the conductive heater and cathode on opposite sides may provide enough capacitance to allow some leakage of high-frequency signals. Between heaters of the video i-f tubes, and between the third video i-f and the sound detector heaters are small inductors called radio-frequency chokes. These r-f chokes offer high inductive reactance to the high-frequency signal currents and thus provide additional safeguard against signal leakage.

Now we shall make a check of the total resistance offered by the heater strings to flow of current from the power line.

Series Resistor	130 ohms		Series Resistor	75 ohms
Audio Output	83.3 "		High Voltage Osc.	83.3 "
R.F. Amp.	21 "		Horiz. Sweep Amp.	42 "
Audio I.F. Amp.	21 "		Vert. Sweep Amp.	42 "
R.F. osc. and mixer	21 "		Vert. Sweep Osc.	42 "
1st Video I.F.	21 "		Horiz. Sweep Osc.	42 "
2nd Video I.F.	21 "		Sync Amplifier	21 "
3rd Video I.F.	21 "		Video Amp.	21 "
Second detector	21 "		Total Resistance	368.3 ohms
Total Resistance	360.3 ohms			

To find the equivalent resistance we use the product over the sum formula.

$$\frac{360.3 \times 368.3}{360.3 + 368.3} = \frac{132698.49}{728.6} = 182.1 \text{ approx.}$$

The combined resistance of the Audio Amplifier heater and its parallel resistor is about 10.7 ohms, as you may check by using the same formula mentioned above. Since we have the equivalent resistances of both the parallel groups we changed the circuit to a simple series circuit. The total resistance as the source sees it then would be the sum of these series resistances.

$$182.1 + 10.7 + 10.5 \text{ (picture tube resistance)} = 203.3 \text{ ohms}$$

If we use our regular formula for current with total resistances and the standard line voltage, 117 volts, the total current in the circuit would be about .5755 amperes. This current would flow with exactly 117 volts and with all tube heater resistances of exactly their rated values, which seldom happens in practice. Resis-

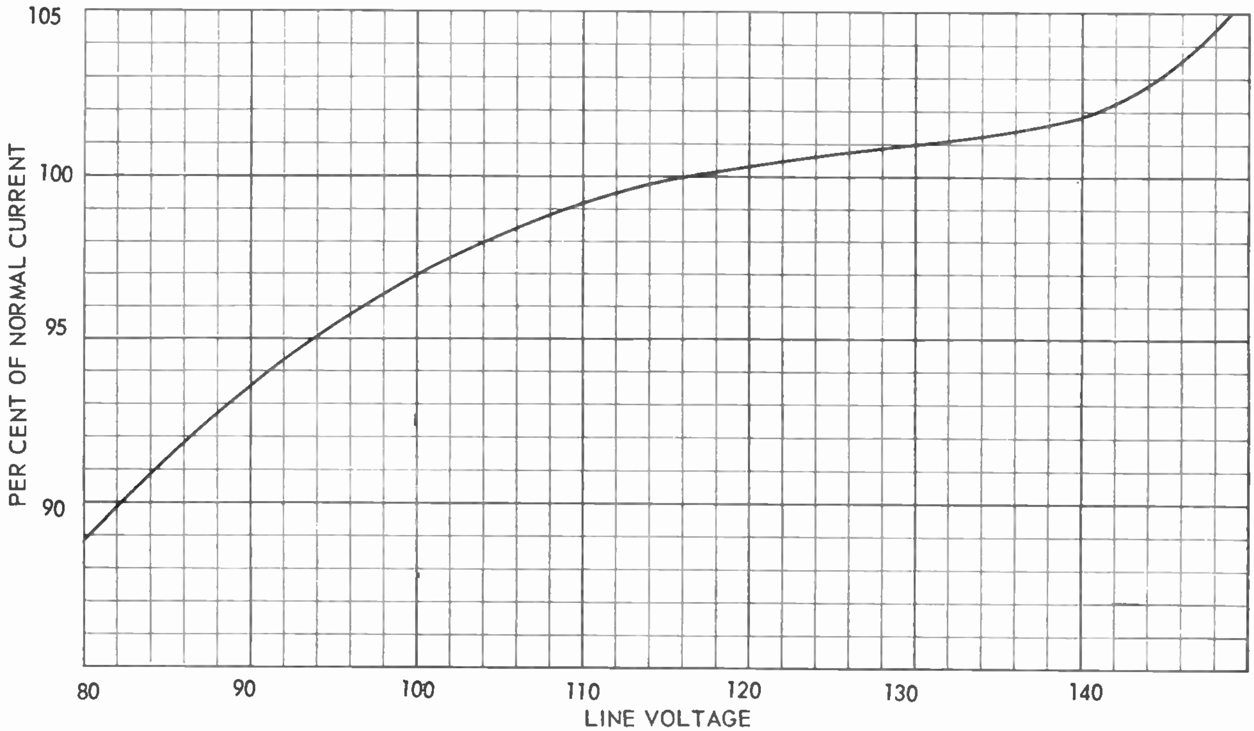


Fig. 23-8. How the current varies in a ballast when there are changes of line voltage.

tances of the r-f chokes in the heater string are so small that we may neglect them.

The accompanying table gives resistances in ohms of tube heaters rated for various voltages and currents. There is a column for each common voltage. Currents, in amperes are listed at the left.

TUBE HEATER RESISTANCES, OHMS

Current, amperes	Rated Voltages 6.3	12.6	25	35	45	50
0.15	42	84	167	233	300	333
0.3	21	42	83.3	117	150	
0.45	14	28				
0.6	10.5	21				

7 It is not possible to measure the effective resistance of tube heaters with an ohmmeter. The effective resistance or working resistance of the heater when at its normal operating temperature is much higher than its cold resistance, which would be the resistance measured with an ohmmeter. As examples, the cold resistance of a 12.6-volt 0.15-ampere heater will measure 12 to 13 ohms as against its hot resistance of 84 ohms, and cold resistance of a 35-volt 0.15-ampere heater will measure about 33 to 35 ohms as against 233 ohms hot resistance.

Whenever a heater rated for relatively small current is connected in series with a circuit carrying greater current, or with other heaters carrying greater current, the low-current heater must be parallel with a fixed resistor. This happened with the heater for the audio amplifier in series with the picture tube heater of Fig. 23-7. A parallel resistor used to lessen the current through some other part usually is called a shunt. The required shunt resistance is computed thus:

$$\text{Shunt resistance} = \frac{\text{rated volts for tube heater}}{\text{total amperes} - \text{amperes for tube heater}}$$

As an example, assume that we wish to connect a 12.6-volt 0.15-ampere heater into a string where the current is 0.3 ampere. Then we have,

$$\text{Shunt resistance} = \frac{12.6}{0.30 - 0.15} = \frac{12.60}{0.15} = \text{84 ohms, connected in parallel}$$

BALLASTS AND SERIES RESISTORS. At the left-hand ends of the heater strings in Fig. 23-7 are fixed resistors of 130 ohms and 75 ohms which use up enough of the line voltage to leave remainders equal to the sum of the voltage drops required by tube heaters in each string. Certain types of series resistors used with tube heaters and with other loads in some receivers may act also to counteract changes of line voltage, at least to a considerable extent. These voltage compensating or voltage regulating resistors are called ballasts.

The wire in the ballast resistance is of iron or iron alloy. Current in the ballast heats the iron wire. At first the resistance of the iron increases quite uniformly with increase of current and rise of temperature, but at or near a dull red heat the resistance of the iron commences to increase much faster than current. Then there is relatively great increase of resistance with small additional increase of current. The result is to allow current to rise to a certain value, then to strongly oppose further increase of current.

This action is illustrated by Fig. 23-8. Rise of line voltage causes rapid rise of current until resistance of the iron wire commences its rapid increase. Then the increasing resistance of the iron holds current almost constant with further rise of line voltage over a considerable range. If line voltage increases beyond this range the ballast wire becomes too hot to retain its regulating property, and current rises as fast as before the action commenced. With tube heater current thus maintained fairly constant there is correspondingly constant voltage drop in the tube heaters, so we may say that the ballast regulates heater voltages.

Ballasts usually are constructed on 8-pin (octal) or 4-pin bases of the kind used for radio and television tubes. This construction is pictured in Fig. 23-9. The ballast resistances may be sealed in glass or metal envelopes similar to tube envelopes, or they may be protected within cylinders of perforated steel without sealing.

Sealed envelopes often contain hydrogen gas. This gas is a poor conductor of heat, with the result that heat from the ballast wire passes only slowly to surrounding air and temperature of the ballast depends chiefly on current and change of current. A cracked envelope may admit air to form a weakly explosive mixture with the hydrogen. If the mixture is ignited by the red hot wire the glass of an envelope will be blown outward, whereas in an ordinary vacuum tube the glass collapses inwardly.

When ballast resistances are visible through glass or a perforated metal envelope you may see a dull red glow during normal operation. All ballasts, as well as plain series resistors, run quite hot. There must be free circulation of air around these units in order to carry away the heat.

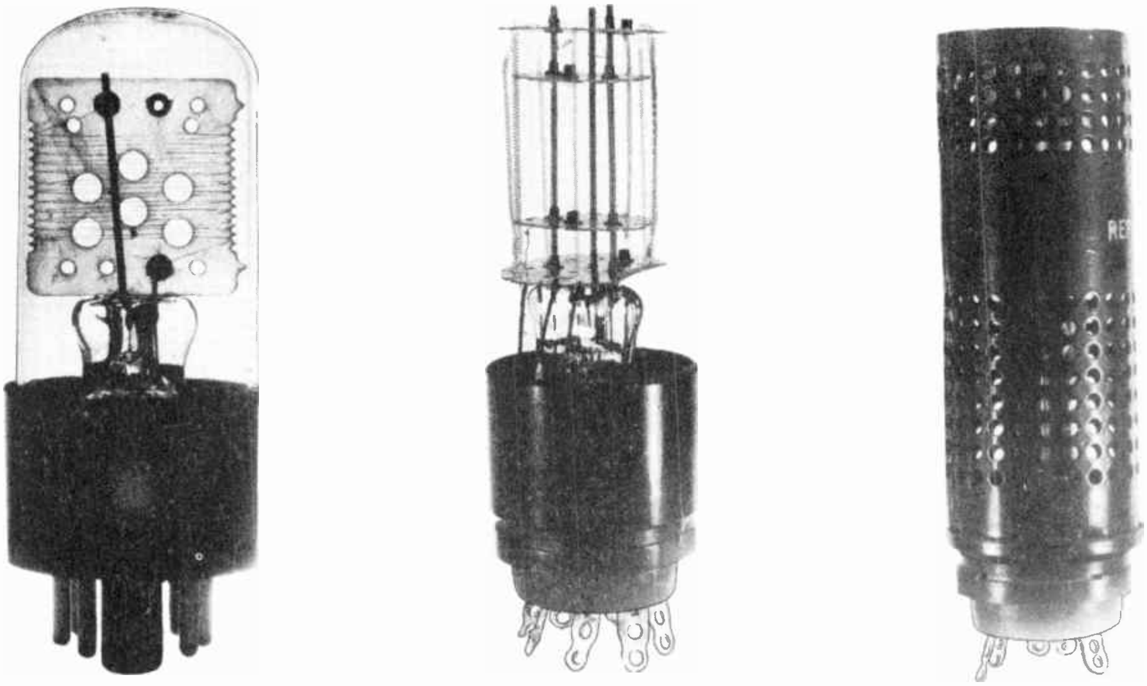


Fig. 23-9. Ballasts for television and radio receivers.
Left: Ballast wire on mica insulator enclosed by glass envelope.
Center: Ballast with burned out sections, shown with envelope removed.
Right: Ballast enclosed within perforated steel cylinder.

A short circuit or accidental ground in parts or wiring connected to a ballast almost always burns out the ballast wire, which necessitates replacement of the unit. If a new ballast is put in before locating and correcting the cause for burnout, the new unit will go the way of the original one. Ballasts cost about as much as radio tubes, so it pays to correct the trouble before burning up new ballasts.

All the ballast resistances or plain series resistances for a receiver usually are contained in a single unit which mounts in a tube socket to whose prongs the various circuits are connected. As an example, in Fig. 23-7, a single ballast unit would contain both resistances which are in series with the heater strings, and also the shunt resistance for the audio amplifier. The single ballast unit ordinarily will contain in addition one or more of the resistances used in filter sections of the B-power supply. In receivers having parallel-connected tube heaters run from a transformer there may be a ballast unit containing only filter resistors.

Pilot lamps in transformerless radio receivers often are connected across a portion of the series resistance or ballast resistance which is in the tube heater circuit. Fig. 23-10 illustrates a typical arrangement of this kind. The ballast, on an 8-pin base, has resistances between pins 3 and 8, also between 8 and 7. The entire resistance, from pin 3 to pin 7, is in series with the tube heaters. The pilot lamp is in parallel with the resistance between pins 8 and 7.

The most commonly used radio pilot or dial lamp is the type 47, designed to operate on 6.3 volts and 0.15 ampere, which means a hot resistance of 42 ohms. If tube heaters of Fig. 23-10 require current of 0.3 ampere

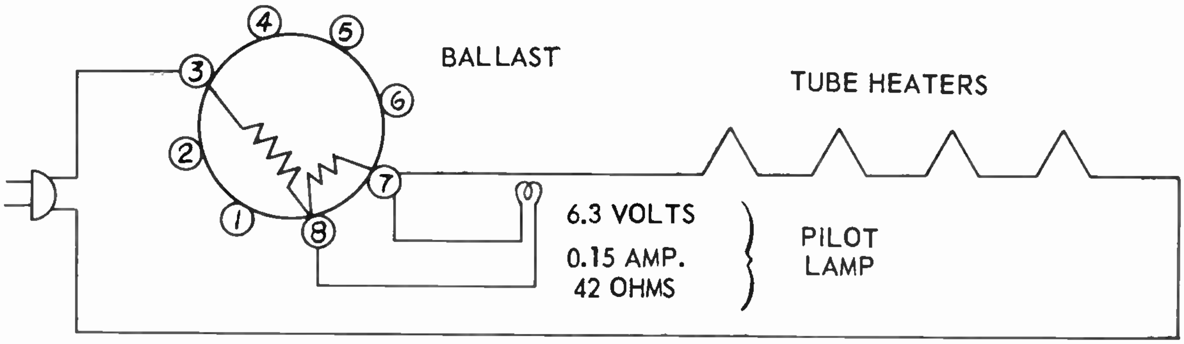


Fig. 23-10. Pilot lamp connected across a section of the ballast.

it is necessary to send half of this total current, 0.15 ampere, through the pilot lamp and to send the other half through the resistance between ballast pins 8 and 7. For equal division of current, the resistance of this ballast section must be equal to the hot resistance of the pilot lamp, or must be 42 ohms.

If resistance between pins 8 and 7 is 42 ohms, and resistance of the pilot lamp is 42 ohms, their combined parallel resistance is 21 ohms. Then, so far as the heater circuit is concerned, there is resistance of 21 ohms between pins 8 and 7. The remainder of whatever resistance is required in the heater circuit will be between pins 3 and 8 of this particular ballast unit.

Some of the rectifier tubes used in transformerless receivers are designed for connection of a pilot lamp or dial lamp across part of the tube heater resistance. The heater connections for a rectifier tube of this style are shown in Fig. 23-6. A more complete circuit is shown by Fig. 23-11. The pilot lamp is in parallel with the left-hand section of the rectifier heater. Voltage drop across this section of the heater is the same

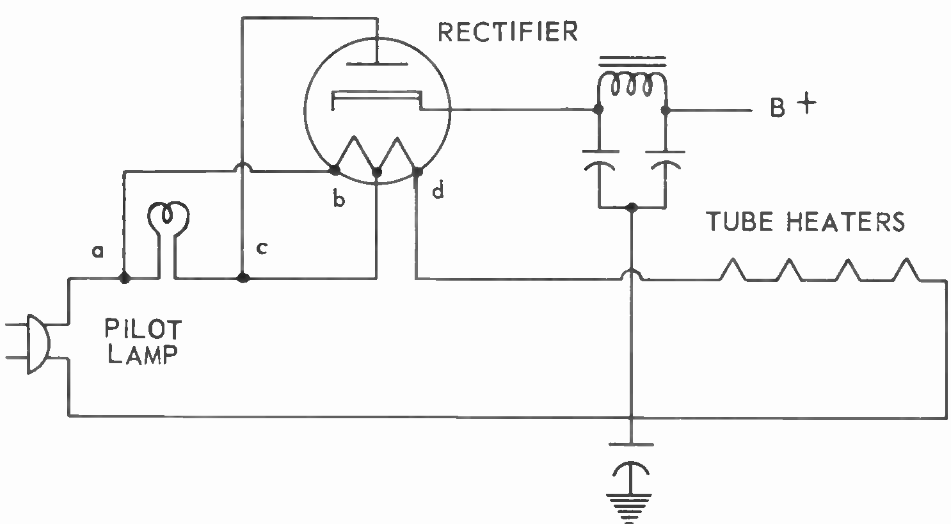


Fig. 23-11. Pilot or dial lamp connected across part of a rectifier heater.

as required by the pilot lamp. (Part of the current for the rectifier filament goes from *a* to *b*, through part of the heater to *d*, and to the other heaters. The remainder goes through the pilot lamp to *c*, and to the rectifier filament to *d.)*

Part of the line current for tube heaters other than in the rectifier goes through the pilot lamp to *c*, thence through the right-hand part of the rectifier to *d*, and to the remaining heaters. The remainder of the heater current goes from *a* to *b*, through both parts of the rectifier heater to *d*, and to the other heaters. Total drop in the rectifier heater is 35, 40 or 45 volts, depending on the type rectifier in use. If this is not enough drop for the heater circuit, an additional resistance will be connected in series with the heaters.

Resistance to be added in series with any string of tube heaters which are themselves connected in series may be computed from this formula.

$$\text{Added series resistance} = \frac{\text{line volts} - \text{sum of all tube heater voltages}}{\text{current in amperes, for one heater}}$$

As an example, in Fig. 23-11 we might have a 35-volt rectifier heater, three other tubes each having 12.6 volt heaters, and one audio output tube with a 25-volt heater. This makes a total of 97.8 volts for all the heaters. We shall assume that current, in each and all the heaters, is to be 0.15 ampere. Then the formula gives,

$$\text{Added series} = \frac{117\ (\text{standard line volts}) - 97.8}{0.15} = \frac{19.20}{0.15} = 128\ \text{ohms}$$

There is such great variety in ballast units and resistor units for all the radio and television receivers using them that the safe way to make a replacement is with the original part as furnished by the manufacturer, or with an "exact replacement" as specified for the particular receiver. When you become really proficient in using the rules and formulas for series and parallel resistances it will be safe to select values "on your own", but not until then.

5 *SERVICING SERIES HEATERS.* There are certain features about series heater circuits which require special consideration. First, a burned out heater in any one tube of a series string will put out the heaters of all other tubes in that string. Each tube may be removed from its socket and the heater checked with an ohmmeter or any circuit tester until the defective one is located. Another way is to use an a-c voltmeter capable of reading line voltage. With this meter connected across the heater terminals of each tube in the string there will be zero voltage until reaching the burned out heater, and there the meter will read practically full line voltage.

6 Not all tubes have their heaters connected to the same base pins. If you put a tube in a socket where it doesn't belong, in a series heater string, the heater of this tube may not light – because it is not connected in the heater circuit.

Oftentimes you will wish to remove one or more tubes from a series heater circuit and have the remaining tubes operate. This may happen with the picture tube, or you may wish to remove the r-f oscillator during alignment. The heater holes or prongs of the socket from which the tube is removed must be connected together with a resistor. Required resistance values are given in the table of tube heater resistances, or you may divide the rated number of volts by the rated current in amperes for the tube and thus determine the ohms of resistance needed.

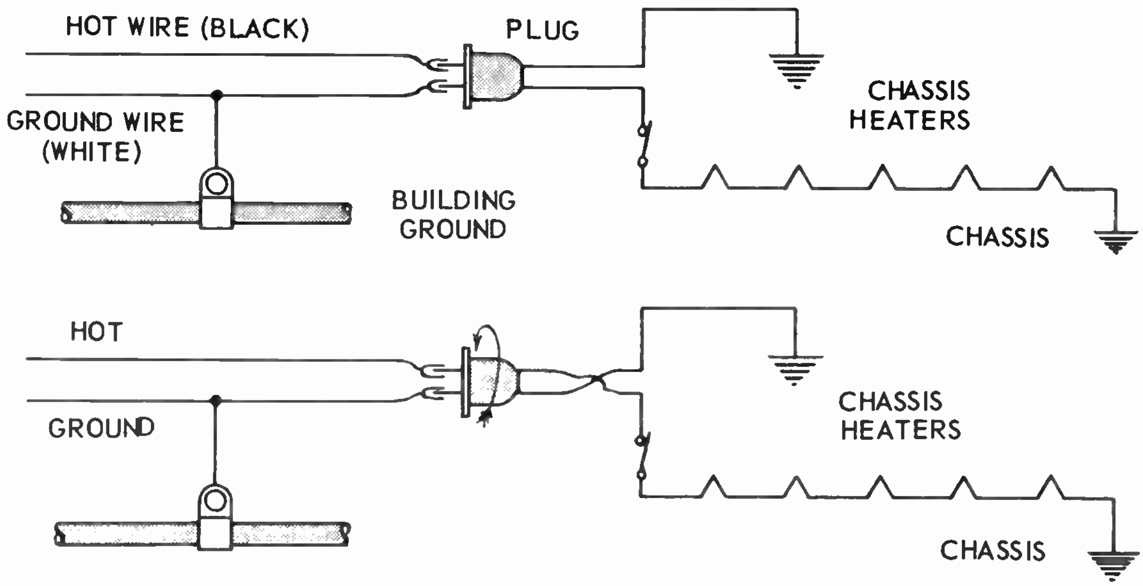


Fig. 23-12. Upper diagram shows hot chassis; lower diagram,building ground grounded chassis.

When jumping a heater in this way you must use a resistor which will safely dissipate the power in watts. Compute the number of watts used by the heater by multiplying the rated heater volts by the rated heater amperes, then double this product to determine the resistance for the resistor. For example, to jump the heater for a picture tube taking 6.3 volts at 0.6 ampere you will need 10 or 11 ohms resistance in a resistor which will dissipate a number of watts equal to twice the product of 6.3 (volts) and 0.6 (ampere). This comes to twice 3.78 watts, or 7.56 watts. A 10-watt resistor would be satisfactory.

With series heaters, and with transformerless receivers in general, the chassis metal may be "hot". This means that the chassis is electrically connected to the hot wire or the ungrounded wire of the building power circuit. This connection is shown by Fig. 23-12. With the line plug inserted in the power line receptacle as in the upper diagram, the receiver chassis is directly connected to the hot wire of the line. This is the wire with black or other color of insulation, not white, in the building wiring. With the plug reversed, as in the lower diagram, there is a connection from the hot wire of the building circuit through the tube heaters to chassis metal so long as the receiver switch is turned on.

If you touch the hot chassis with the plug inserted as in the above diagram and at the same time touch a water pipe, a gas pipe, an electric fixture, or anything else connected to the building ground you will get a severe shock because effectively you are in parallel with the tube heaters. You are, in effect, touching both sides of the building power line with nothing between you and the line. With the plug reversed, nothing can happen because there is no difference in potential between chassis and the building ground.

In Fig. 23-11 the heater circuit is completed on both sides with insulated wire. There is no conductive connection from either side of the line cord to chassis metal. Between the heater circuit and chassis metal

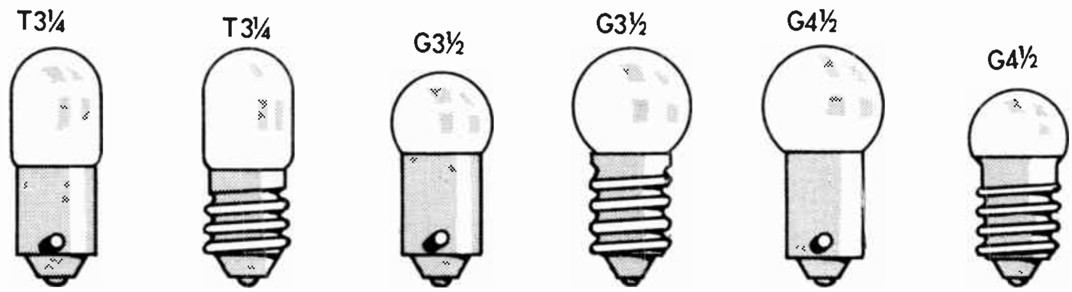


Fig. 23-13. Types of bulbs and bases used on pilot, panel, or dial lamps.

is a capacitor which completes some of the circuits for alternating signal voltages and currents. Capacitance here is usually 0.05 microfarad. Capacitive reactance at 60 cycles is more than 53,000 ohms, which would allow current of about 2¼ milliamperes with 117 alternating line volts. If your only connection to the line is through this capacitance you cannot feel even a tingle, and there is not the slightest danger of shock. There may, however, be other connections in the same receiver, either intentional or accidental, through which you will get enough voltage and current for a real shock.

3 When heaters are fed from a transformer secondary, as in Fig. 23-5, and the B-power supply is similarly operated from a secondary winding, everything in the receiver is insulated from the line. Then we have a cold chassis. There will be more to say about handling hot and cold chassis when learning to work with test instruments.

PARALLELED RESISTORS FOR REPLACEMENTS. Before leaving the subject of paralleled resistances, at least for the time being, we should get acquainted with a rule of formula that gets service technicians out of many difficulties when making replacements from a limited stock of resistors. Here is a typical difficulty. Assume that you wish to replace a 39-ohm resistor, a value which is not in stock. But you have quite a few 160-ohm units and some others of greater resistance than 39 ohms. Which of these others may be paralleled with 160 ohms to provide combined resistance of 39 ohms? Here is the solution for the problem.

1. Divide the resistance you have on hand by the number of ohms you want.

2. Subtract 1 from the above quotient.

3. Divide the resistance on hand by the difference found in step two.

Let's try it. Step 1: Dividing 160 by 39 gives approximately 4.1. Step 2: Subtracting 1 leaves 3.1. Step 3: Dividing 160 by 3.1 gives about 51.6 as the required parallel resistance. There is a standard resistor value of 51 ohms. If you have one, and connect it in parallel with a 160-ohm unit, their combined resistance is about 38.7 ohms. Other standard resistance values are 47 ohms and 56 ohms, which, paralleled with your 160-ohm unit, will give combined resistances of about 36.3 and 41.5 ohms. Both these combined resistances are within better than 8 per cent of the desired 39 ohms.

PILOT LAMPS. The accompanying table lists information relating to the types and sizes of pilot lamps or dial lamps in general use. The same lamp numbers are used for equivalent or like lamps by the various manufacturers. Shapes, sizes, and type numbers of the different glass bulbs are shown by Fig. 23-13. Bases are either bayonet or screw type, with each type of base for each of the bulbs. You may identify each style of lamp by noting its bulb, its base, and the color of a little glass bead at the bottom of the filament supports inside the lamp. Nominal voltages are those marked on the box in which the lamps are received, and sometimes on the bases also. Design voltages are normal actual voltages which you should use when making any computations. Amperes are the rated currents for the lamps. Resistances are hot resistances, as computed from rated voltages and currents. Keep this table for reference.

PILOT, PANEL, OR DIAL LAMPS

LAMP No.	BULB	BASE	BEAD COLOR	NOMINAL VOLTAGE	DESIGN VOLTS	DESIGN AMPS	RES. OHMS
40	T 3¼	Screw	Brown	6-8	6.3	0.15	42
40A	T 3¼	Bayonet	Brown	6-8	6.3	.15	42
41	T 3¼	Screw	White	2.5	2.5	.50	5
42	T 3¼	Screw	Green	3.2	3.2	.50	6.4
43	T 3¼	Bayonet	White	2.5	2.5	.50	5
44	T 3¼	Bayonet	Blue	6-8	6.3	.25	25
45	T 3¼	Bayonet	Green	3.2	3.2	.50	6.4
45	T 3¼	Bayonet	White	3.2	3.2	.35	9.1
46	T 3¼	Screw	Blue	6-8	6.3	.25	25
47	T 3¼	Bayonet	Brown	6-8	6.3	.15	42
48	T 3¼	Screw	Pink	2.0	2.0	.06	33.3
49	T 3¼	Bayonet	Pink	2.0	2.0	.06	33.3
49A	T 3¼	Bayonet	White	2.1	2.1	.12	17.5
50	G 3½	Screw	White	6-8	7.5	.20	37.5
51	G 3½	Bayonet	White	6-8	7.5	.20	37.5
55	G 4½	Bayonet	White	6-8	6.5	.40	16.3
291	T 3¼	Bayonet	White	2.9	2.9	.17	17
292	T 3¼	Screw	White	2.9	2.9	.17	17
292A	T 3¼	Bayonet	White	2.9	2.9	.17	17

DO NOT TEAR — CUT ALONG THIS LINE AND SEND IN FOR GRADING.

TELEVISION

LESSON NO. 24

CONTROL RESISTORS

Training Your Mind

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON NO. 24

CONTROL RESISTORS

Circuit diagrams issued by set manufacturers for use by service technicians show all the parts of a receiver by means of symbols. When an experienced technician looks at a symbol he sees, in his mind's eye, the part itself. Looking at the symbol for a tube he sees the grid, plate, cathode, and other elements. He sees the pins on the base, the socket, and connections to the socket lugs — all in his mind's eye. You have looked at so many tube symbols in so many diagrams that you are doing just this, whether or not you realize it.

It is easy for you to "see" a tube when looking at its symbol, because you are familiar with the construction of tubes. Now we are going to get acquainted with the construction of control resistors. Then their symbols will commence to look like the real thing when you are working from a circuit diagram.

ADJUSTABLE SLIDER RESISTORS. In Fig. 24-1 is a picture and a symbol of what may be called a slider type adjustable resistor, or an adjustable voltage divider. This is a wire-wound resistance element with the wire exposed or bared along one side. A movable contact member or slider presses against and makes good electrical contact with the wire at one point. The slider is built like a clamp, which may be loosened by turning its screw, then moved anywhere along the exposed wire and tightened to remain there. This resistor is connected into a circuit by means of wires soldered to the lugs on one or both ends and to the terminal which is formed on the slider down beyond the clamping screw.

Adjustable slider type resistors are available with resistances from less than 1 ohm all the way to 100,000 ohms, and with power ratings from 5 watts to as much as 200 watts. Units of lowest power rating are less than 2 inches long and little more 1/4 inch in diameter. Those of greatest power rating are more than 10 inches long and more than an inch in diameter. The size depends on the power rating. Any resistance may be had in any of the power ratings, and corresponding sizes.

More than one slider may be used on a single resistance unit. Fig. 24-2 shows an experimental setup in which is a long resistor with three slider clamps in addition to the two end connections. The resistor is shown as it would be connected for a voltage divider on a power supply. Two of the sliders are set for tap connections furnishing plate and screen voltages to a tube, with a cathode connection at the right. A third slider is unconnected.

POTENTIOMETERS. The resistors of Figs. 24-1 and 24-2 are suited only for service adjustments that are altered but rarely after the apparatus is first installed. They are not suitable for adjustments that may need altering by the operator of a receiver, nor are they very well suited for service controls that are adjusted during reception or while working with signal generators. For all the many operating and service controls in television and radio receivers it is usual to provide what are called potentiometers, with which the resistance in the circuit may be altered by turning a knob, pointer, or shaft.

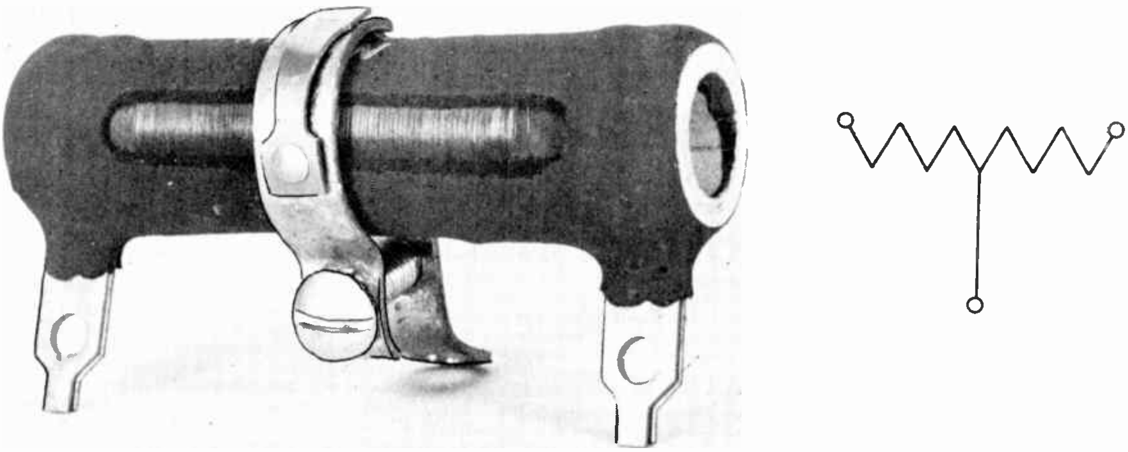


Fig. 24-1. A slider type adjustable resistor.

Fairly typical construction of potentiometers is illustrated by Fig. 24-3. Each unit has a cylindrical outer case or housing within which is carried a resistance element extending around the greater part of a circle. In the left-hand picture the resistance element consists of an insulating support on which are wound many turns of wire. This wire is of some kind having high resistance per inch or foot of its length.

The ends of the resistance wire are connected to the outside two of the three insulated terminals which extend out from the housing. The center terminal is electrically connected to a rotor member which may be turned by a shaft or a shaft and knob attached to the rotor. On the rotor is a tongue or extension that presses against the wire at one point along the resistance element. Turning the rotor shaft moves this tongue around the resistance element and thus makes contact from the center terminal to any point along the resistance.

In other potentiometers the resistance element is of carbon or a carbon-graphite composition supported on or embedded in insulation. The right-hand picture in Fig. 24-3 shows one style of carbon or composition potentiometer.

Potentiometers usually are represented in circuit diagrams by symbols like the one at the right in Fig. 24-1. In a few cases the symbol may be drawn as at the right in Fig. 24-3. It might be mentioned here that service technicians don't always use the long name potentiometer, they call these units "pots". You will speak of a carbon pot or of a wire-wound pot.

9 *WIRE-WOUND POTENTIOMETERS.* Wire-wound potentiometers, compared with carbon types, are capable of handling relatively large currents without overheating the resistance elements. Power ratings of commonly used wire-wound controls for receivers range from 2 to 5 watts. Total resistance between the two outside terminals ranges from 2 ohms to as high as 100,000 ohms in some styles, although resistances greater than 20,000 ohms are not in general use. Outside diameters of the housings run from slightly over an inch up to about $2\frac{1}{2}$ inches.

Fig. 24-4 is a picture of a disassembled 4-watt wire-wound control whose outside diameter is about $1\frac{1}{2}$ inches. At the upper left is a complete potentiometer with its three protruding terminals. At the lower left

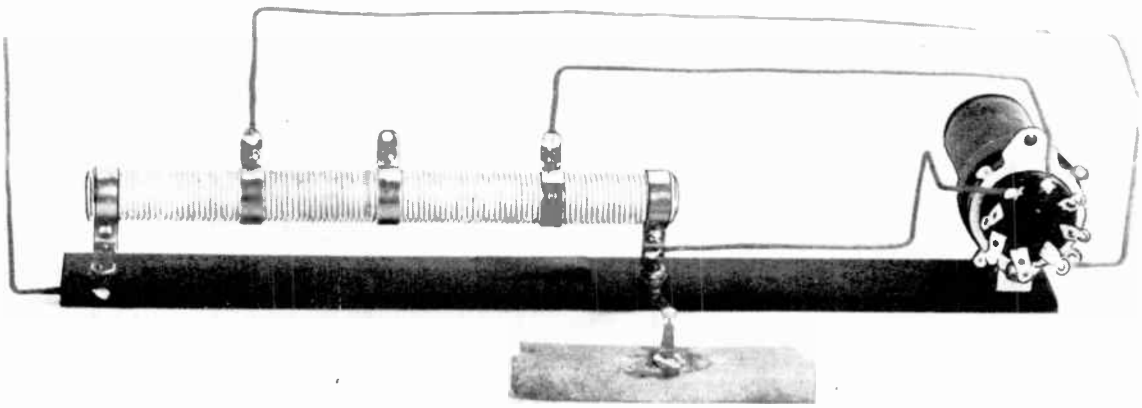


Fig. 24-2. Slider type adjustable resistor used as a voltage divider.

is the wire-wound resistance element In some potentiometers of this general style it is not too difficult to replace a burned out or otherwise damaged resistance element, but as a general rule a complete new unit will cost less than the labor and time for disassembly, reassembly, and obtaining of parts.

In the center of Fig. 24-4 is the rotor member with its attached shaft. Note that the part of the rotor that makes contact on the resistance element is insulated from the shaft. The insulation appears black in the picture. There is a sliding spring contact making electrical connection from the contact member of the rotor to the center terminal. At the right is the housing. A strip of insulating material around the inside of the housing prevents electrical contact between housing and resistance element. The tips of the spring contact for the rotor are visible in the bottom of the housing.

For heavy-duty testing apparatus in shop and laboratory there are available wire-wound potentiometers having power ratings of 25, 50, 100, 150, and even 200 watts, with resistances from less than one ohm to 10,000 ohms or more. Diameters of the higher power units may be close to 4 inches.

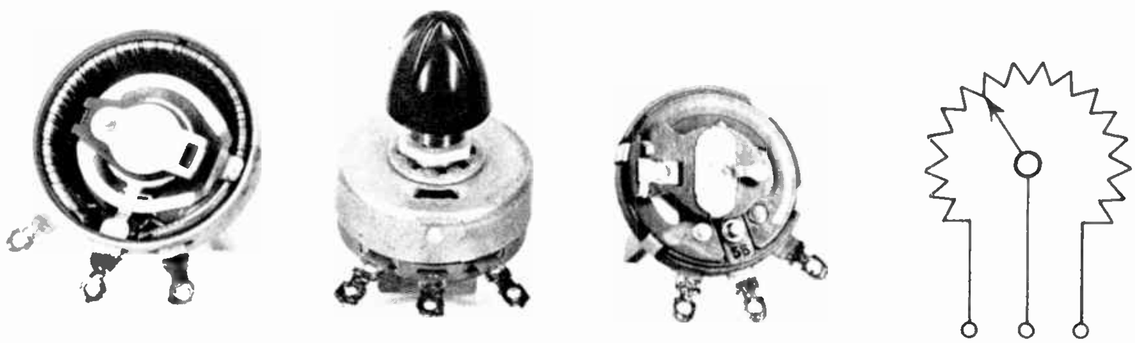


Fig. 24-3. Internal and external construction of some potentiometers.

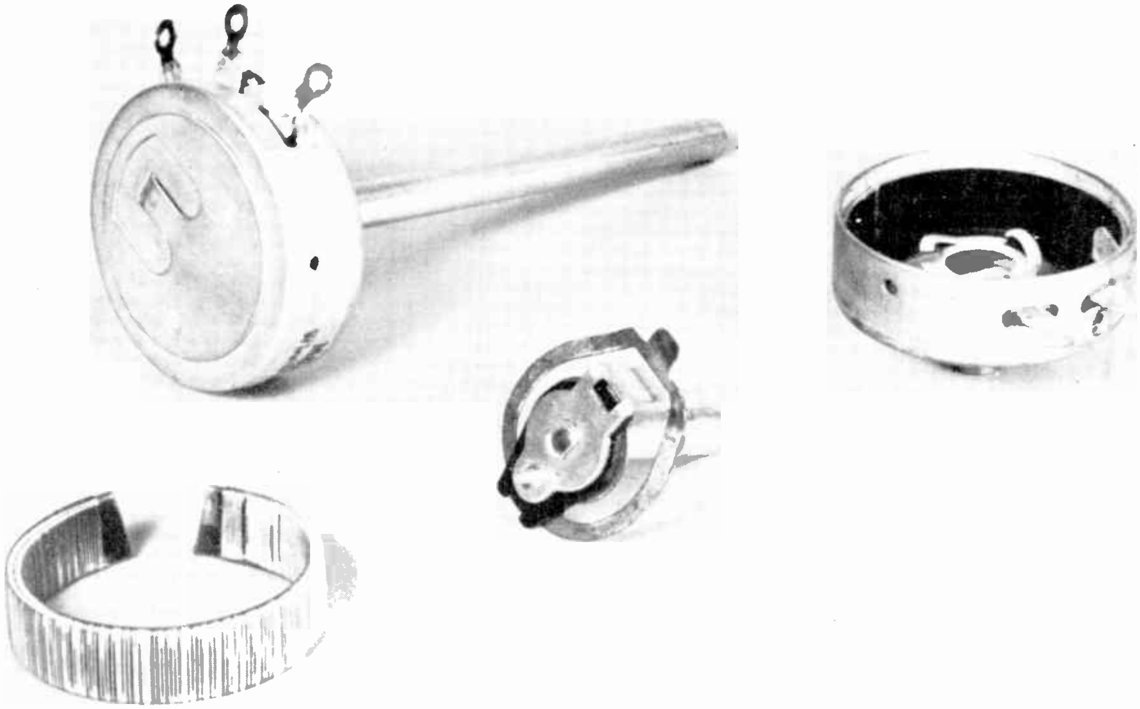


Fig. 24-4. Wire-wound potentiometer complete at upper left and disassembled below.

1 *CARBON POTENTIOMETERS.* Potentiometers employed for furnishing adjustable voltages, and required to carry no current or only small currents, often have resistance elements of carbon or carbon and graphite compositions. The construction of one such control is pictures by Fig. 21-5. At the left is the insulating base to which are attached the three terminals. The carbon resistance element, forming a nearly closed circle, is embedded in insulation. The two outside terminals connect to the two ends of this resistance element. The center lug connects to the circular metallic ring near the center.

At the right is the rotor. The small graphite contact brush is carried by a metallic member on which is a horseshow-shaped contact spring that bears on the metal ring in the base when the potentiometer is assembled. Thus there is electrical connection from the contact brush to the center terminal. As the rotor is turned, the brush rides around on the surface of the carbon resistance element.

In other carbon type potentiometers the resistance element is a flat strip formed into part of a circle around the inside of the housing. Such a style is pictured by Fig. 24-6. In this particular design the contact on the resistance element is made by a flexible ring of thin springy metal which is electrically connected to the center terminal. This flexible ring, which does not rotate, is pressed outwardly against the resistance strip by a small brush mounted on and turned by the rotor. This brush may be seen at the top center of the picture. Construction of this kind avoids having any sliding contact on the resistance element.

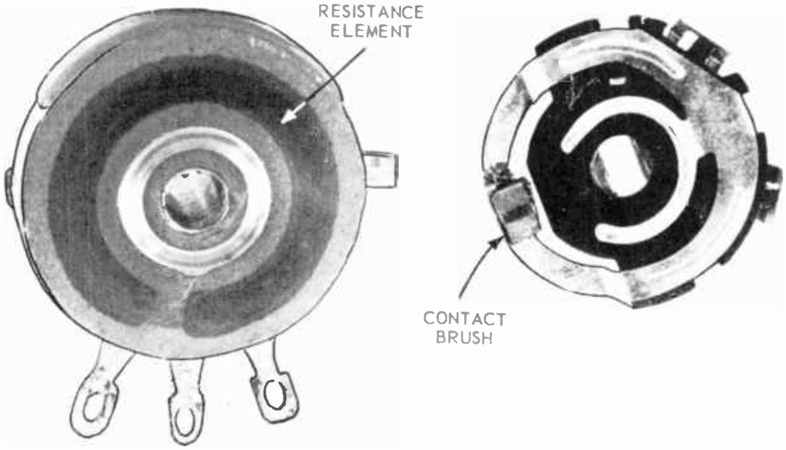


Fig. 24-5. Carbon potentiometer with flat disc resistance element.

Small carbon potentiometers often are called "volume controls" whether they actually are used as controls for sound volume or for some other purpose. The earliest uses of these units were for volume controls in sound receivers, and the name sticks. Resistances range from as low as 50 ohms up to 10 megohms. The smallest units have diameters slightly less than an inch, while larger ones have diameters of around $1\frac{1}{2}$ inches. Power ratings usually are $\frac{1}{2}$ watt or 1 watt, but some carbon controls are rated for as high as 2 or 3 watts of power dissipation.

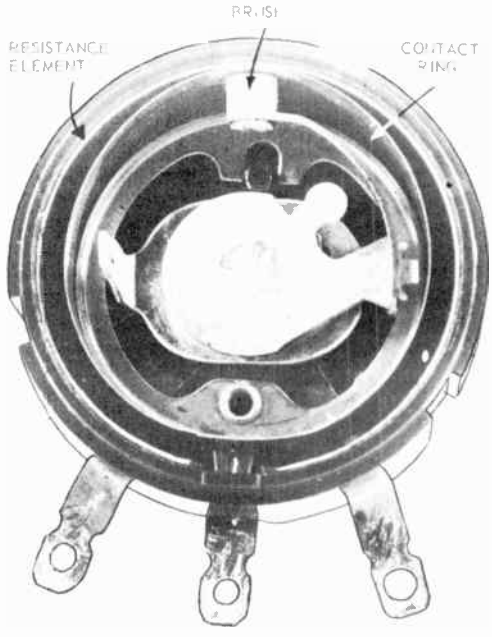


Fig. 24-6. Carbon potentiometer with cylindrical or wall type resistance element.

MOUNTINGS. Potentiometers of all types in common use usually are mounted as shown at the left in Fig. 24-7. On the back of the housing or case is a threaded metallic extension that will pass freely through a hole 3/8 inch in diameter cut through the panel or some part of the chassis of a receiver. On this extension is placed a lock washer. Then the threaded extension, with the lock washer in position, is put through the mounting hole. On the outside of the panel or chassis metal a plain washer is slipped onto the extension and the whole assembly is fastened securely by turning down a thin hexagon-shaped nut.

When you wish to have the exposed end of the threaded extension come flush with the outside of the hexagon nut, or nearly so, any extra length of the extension may be taken up with shim washers or else with one or more extra lock washers between the potentiometer body and the chassis metal.

The lock washer usually will prevent the housing of the potentiometer from turning on the chassis metal, even when the control knob or pointer is twisted rather hard at either end of its travel. There are stops which limit travel of the rotor, and any further pressure on the knob is exerted on the potentiometer housing. For more positive locking in position, some potentiometers have a small metal tab sticking out from the back of the housing. This tab fits into a small hole drilled in the chassis metal.

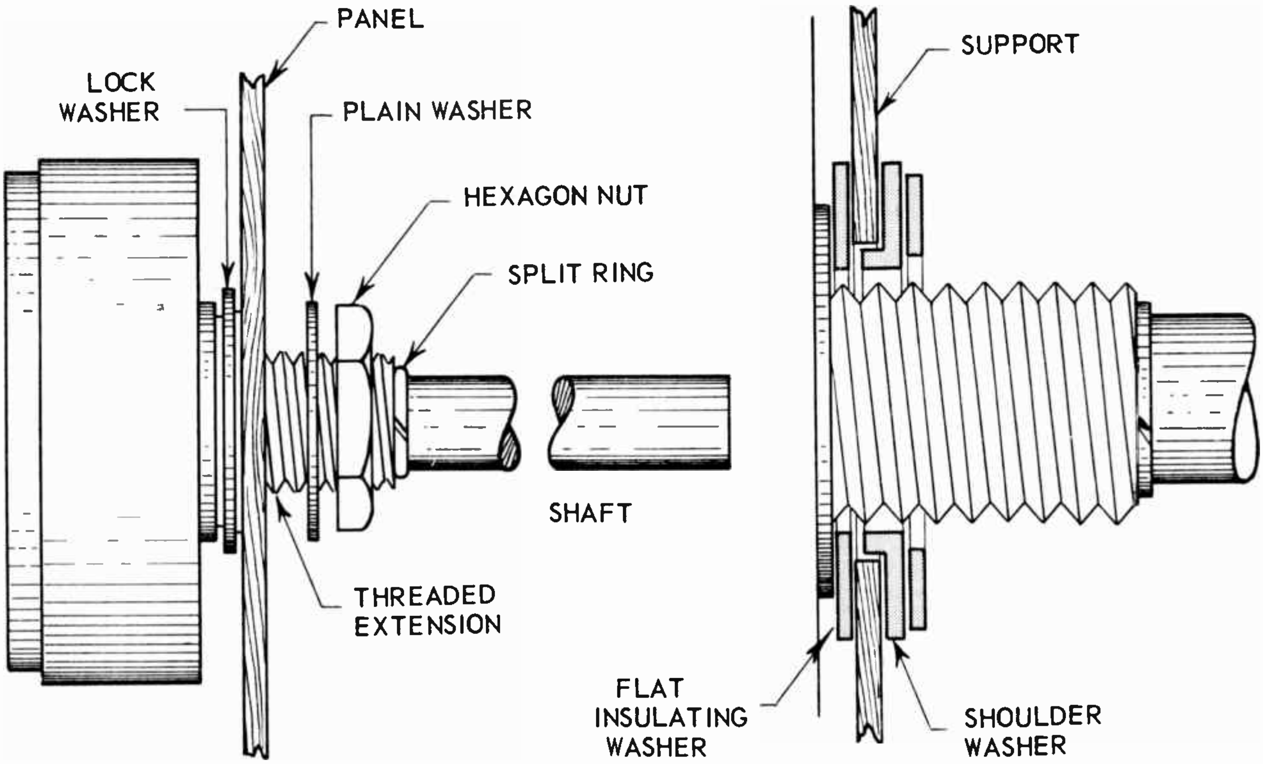


Fig. 24-7. Methods of mounting potentiometers.

The shaft that turns the rotor inside the potentiometer comes out through the threaded extension, and most often is held against endwise movement by a split ring that snaps into a groove around the shaft. To disassemble the potentiometer this split ring must be lifted out of the groove and slipped off the end of the shaft.

Replacement type potentiometers usually come with shafts about 3 inches long, which you may cut off to any length required for the job. If still greater length is required it may be provided by an additional rod or shaft fastened to the original shaft with a coupler having set screws, clamp bushings, or other fastening device. The shafts of practically all control potentiometers are ¼ inch in diameter.

For service controls the shaft may have a screw driver slot across its exposed end. A plain round shaft is used when a knob or pointer will be held in place with a set screw that digs into the side of the shaft. Other types of knobs or pointers simply push onto the end of the shaft and are held there by a tension spring carried inside the knob. Push-on knobs or pointers require that the exposed end of the shaft be flattened on one or more sides, or that it be fluted or grooved, or made with a narrow extending tongue, or otherwise suitably formed to take the knob.

Some replacement potentiometers are designed for use with separable shafts. The rotor shaft of the potentiometer comes out only to the exposed end of the threaded extension that mounts the unit. Any of various types of operating shafts may be attached to the rotor shaft with a fastening consisting of threaded, tongued, grooved, or other shaped ends held together by some suitable clamping or locking device.

A potentiometer shaft and the threaded extension through which it passes usually are insulated from the current-carrying parts of the rotor. Otherwise, when the metal extension is mounted in a metal panel or chassis, the live parts of the rotor will be electrically connected to the support or will be grounded to the support. Potentiometers used in high-voltage circuits of some television receivers may have the additional protection of a shaft housing, and sometimes a coupled-on shaft, made of some hard insulating material.

If the shaft of a potentiometer is not insulated from the rotor, and if the rotor must not have electrical connection with the supporting metal, insulating washers are placed around the mounting extension on both sides of the supporting metal, as shown enlarged at the right in Fig. 24-7. A flat insulating washer is placed between the body of the potentiometer and the supporting metal. On the other side of the support is placed a shoulder type insulating washer on which is a protruding ring that fits into the panel hold around the threaded extension of the potentiometer. The hole in the support has to be somewhat larger than the extension to accomodate the shoulder of the washer. The shoulder makes it impossible for supportingmetal to come in contact with metal of the potentiometer. Outside the shoulder washer is a plain metal washer against which the hexagon nut is screwed. The positions of the plain and shouldered insulating washers may be reversed.

Insulating washers do not grip the supporting metal with enough friction to prevent the potentiometer housing from turning when a knob or pointer is given a hard twist against the stop that limits rotor travel. Consequently, it is necessary to provide other means, such as an extending tab, to hold the potentiometer in place.

TAPPED POTENTIOMETERS. In some centering controls for television receivers and in certain other control circuits it is necessary that a potential or a current be varied in either polarity from zero or reference value. This is most easily handled by a tapped potentiometer, which is a type having an additional fixed connection at some point between the ends of the resistance element.

The tap connection is brought out to an additional terminal which, as in Fig. 24-8, usually is on the side

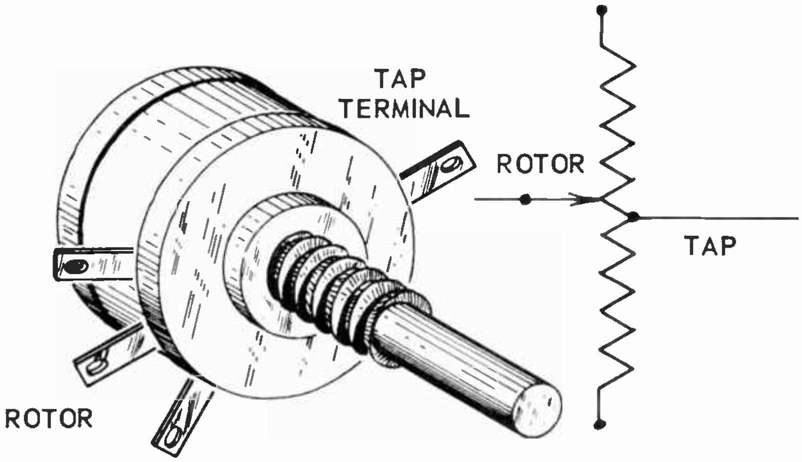


Fig. 24-8. Appearance of a tapped potentiometer and its symbol.

of the housing opposite the three terminals for the ends of the resistance element and the rotor. At the right is shown the symbol most often used to indicate a tapped potentiometer. In a few cases there may be two or even three taps connected to various points between the ends of the resistance element. There will be an extended insulated terminal for each tap. The single or multiple taps may be connected almost anywhere along the resistance, depending on circuit requirements. Any tap may be at a point as little as one-fourth of one per cent of the total resistance,from either end, or it may be anywhere else all the way to the center, measuring from either of the ends.

DUAL AND DOUBLE POTENTIOMETERS. A dual or twin potentiometer consists of two electrically separate units mounted end to end and operated together by a single shaft that extends all the way through one housing and into the other one. Controls of this kind are illustrated by Fig. 24-9. These combination units may be used where it is desired to simultaneously control an antenna circuit and a grid circuit or bias circuit, also in other applications. Resistances of the two sections of the dual unit may be alike or different. One or both may be of the tapped type.

A double potentiometer consists of two electrically separate potentiometers mounted end to end, but arranged for independent operation. Fig. 24-10 illustrates the principle. The shaft is a concentric type, with a central solid part passing through an outer tubular shaft. The rotor of the front potentiometer, nearer the panel or other support, is operated by the tubular shaft. The center shaft extends on through and operates the rotor of the rear potentiometer, farther from the support.

The inner solid shaft extends beyond the outer tubular shaft at their exposed ends.The knob for the front potentiometer fastens onto the tubular shaft.A smaller knob or pointer fastens onto the extended end of the central solid shaft and operates the rear potentiometer.Either knob or pointer may be turned without affecting the other one.

Concentric shaft potentiometers are used in a great many television receivers in order to lessen the apparent number of controls and to make for easier assembly.The two knobs or pointers take up no more front space than one and appear much like a single control. One of these double potentiometers might be used

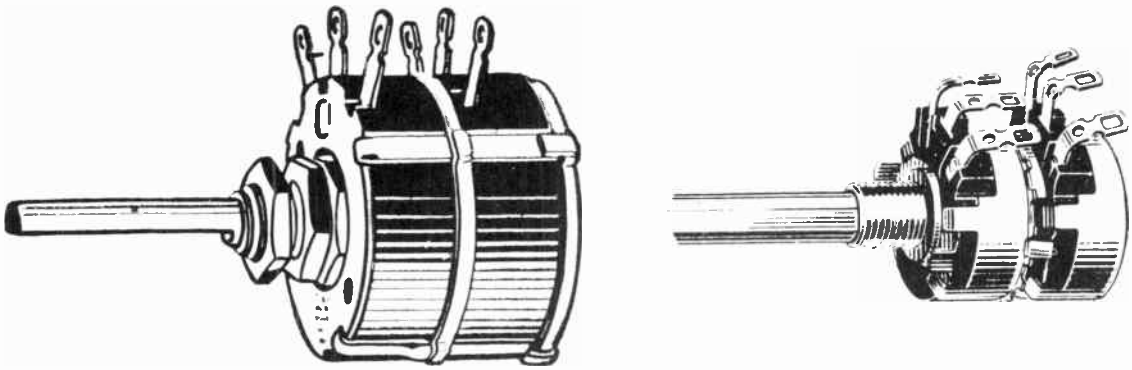


Fig. 24-9. Dual or twin potentiometers.

for the vertical and horizontal hold controls, another for vertical and horizontal centering controls, a third for the contrast and brightness controls, and so on.

RESISTANCE TAPERS OR CURVES. There are many control potentiometers in which the resistance does not change uniformly between one end and the other of the element. As the rotor is turned at a uniform rate from one end to the other of its travel there may at first be a slow change of resistance, followed by a rapid change at the far end of the travel, or there may be first a rapid change followed by a slower change.

When change of resistance is not uniform with rotation of the rotor contact the potentiometer is said to be tapered. To get acquainted with various tapers and their effects we shall begin by examining the performance of a potentiometer with no taper at all, with uniform change of resistance as the rotor is turned all the way around its travel.

Assume that we make circuit connections to the left-hand terminal and to the rotor terminal of a taperless potentiometer, as along the top of Fig. 24-11. At *A* the rotor shaft is turned all the way to the left, or counter-clockwise. There is zero resistance inserted in the circuit, because the rotor is in direct contact with the left-hand terminal.

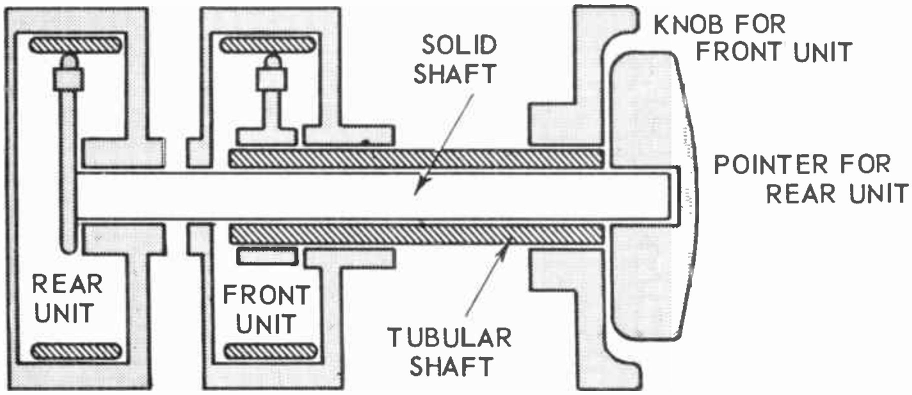


Fig. 24-10. Construction principle of a double potentiometer with independent control knobs.

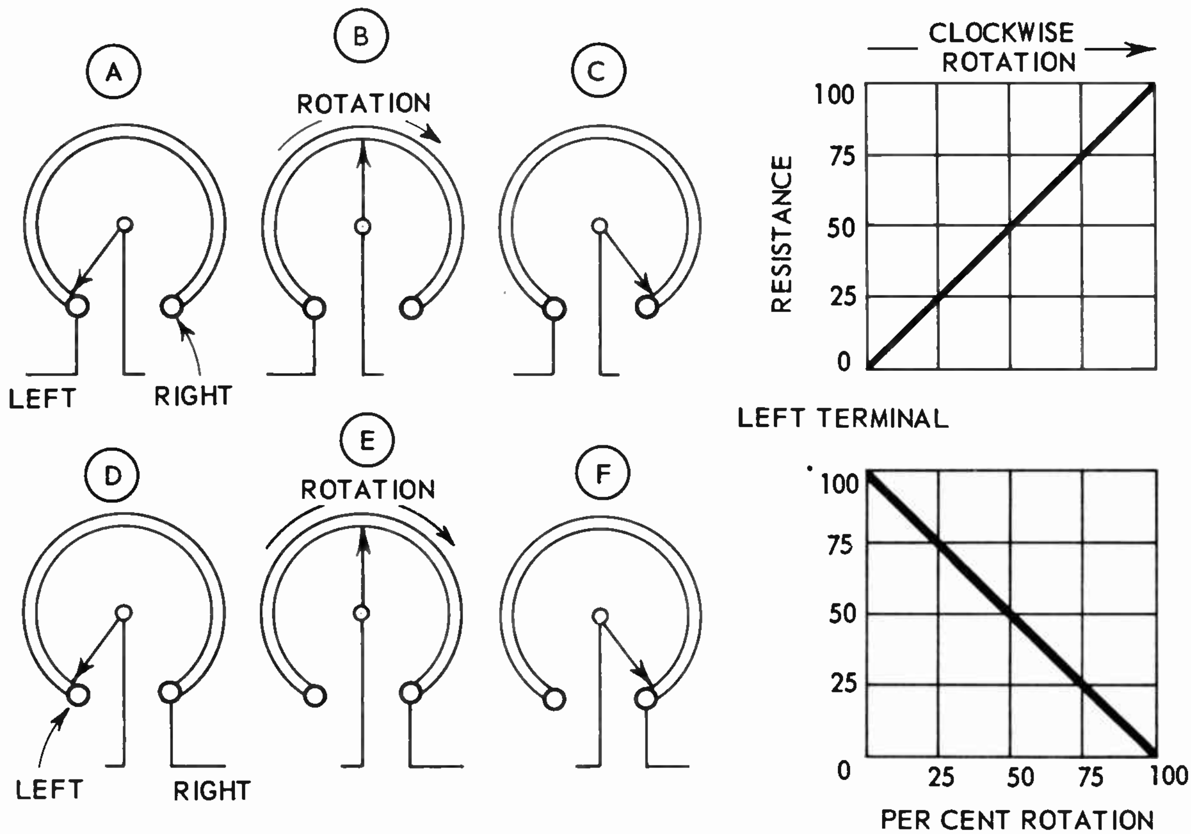


Fig. 24-11. Resistance curves for a linear or taperless potentiometer.

If the rotor is turned clockwise to its center position, at *B* there is half the total potentiometer resistance in the circuit. With the rotor turned fully clockwise, as at *C*, there is maximum resistance in the circuit because the entire resistance element now is between the rotor contact and the left-hand terminal. These changes of resistance may be shown by the curve at the upper right.

Assume next that the circuit is connected to the right-hand terminal and to the rotor terminal, as down below. With the rotor fully counter-clockwise, at *D*, the entire potentiometer resistance is in the circuit, because the whole resistance is between the rotor contact and the right-hand terminal. Turning the rotor half-way, as at *E*, leaves half the total resistance in circuit. With the rotor turned fully clockwise, at *F*, all the resistance is out of circuit, or there is zero resistance. These changes of resistance with rotation are shown by the curve at the lower right.

The curve which shows change of potentiometer resistance with movement of the rotor has been inverted by changing one of the circuit connections from the left-hand terminal to the right-hand terminal, with the other circuit connection remaining on the center terminal or rotor terminal. Any resistance curve may be similarly inverted by a change of circuit connections.

In Fig. 24-12 are some curves showing resistance tapers. These are not curves for any particular make or model of potentiometer, but they are typical of those available in nearly all makes. No two manufacturers show exactly the same curves for their various potentiometers. In addition to the curves shown by the graph, all kinds of potentiometers are available with the uniform change of resistance shown by Fig. 24-11. A uniform change sometimes is called a linear taper or a straight taper.

The curves on our graph show how resistance varies as the rotor is turned from either the left-hand or right-hand terminal all the way to the other end of its travel. Whether the start of a curve is at zero resistance or at 100 per cent resistance depends on how connections are made to the potentiometer terminals, as was shown by Fig. 24-11. Any resistance curve may be inverted by changing a connection from one outside terminal to the other. Drawing some of the curves one way and the remainder inverted avoids confusion which might result from crossings of the lines with all drawn the same way.

Tapers such as marked *K*, *L*, and *M* on the graph are used for adjustment or control of amplifier grid biases and for reducing the signal from an antenna by connecting the resistor in parallel with an antenna coil or coupler. Taper *N* is used as a bias control, *O* may be used for tone control, and *P* as an antenna signal control. Tapers *O* and *P* are used also for volume controls in audio amplifier circuits.

7 *POWER RATINGS AND CURRENTS.* Adjustable resistors may become defective and require replacement for various reasons. After a long period of service the surface of a carbon resistance element may become worn or roughened, or a wire winding may be worn through. Then, if the unit is in a television video or sweep circuit, the pictures will become jumpy or will flicker or will come and go as the control is oper-

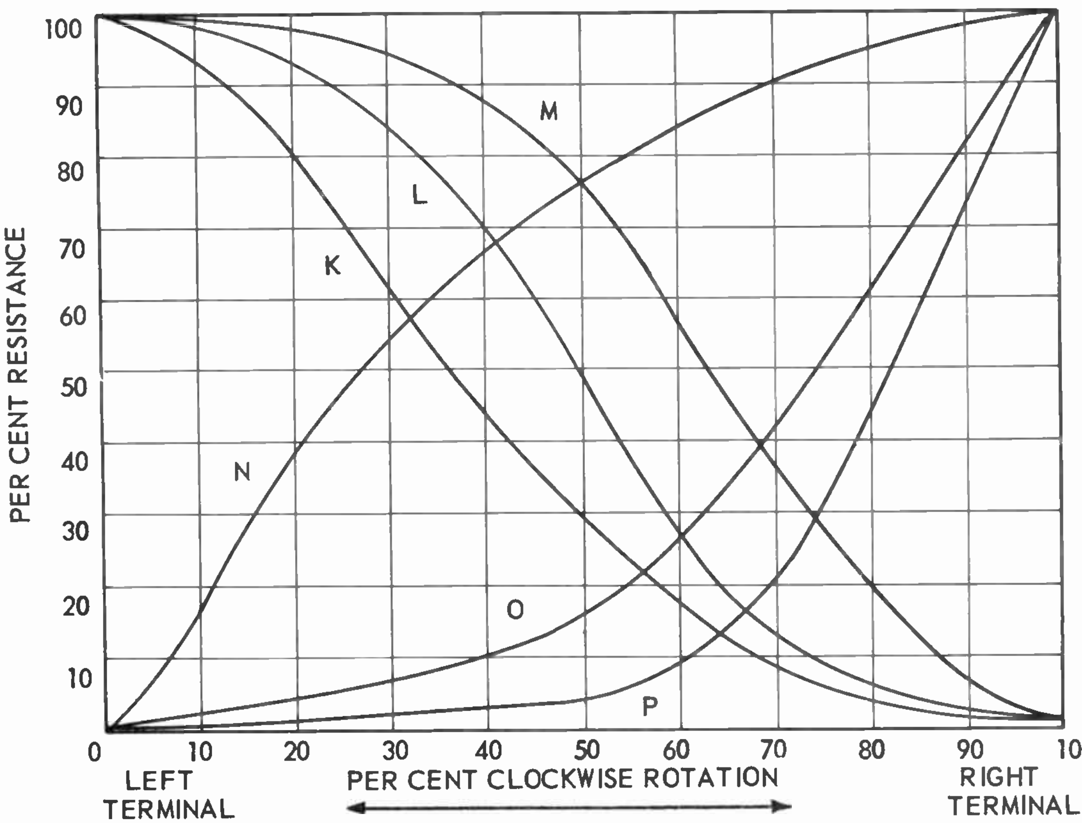
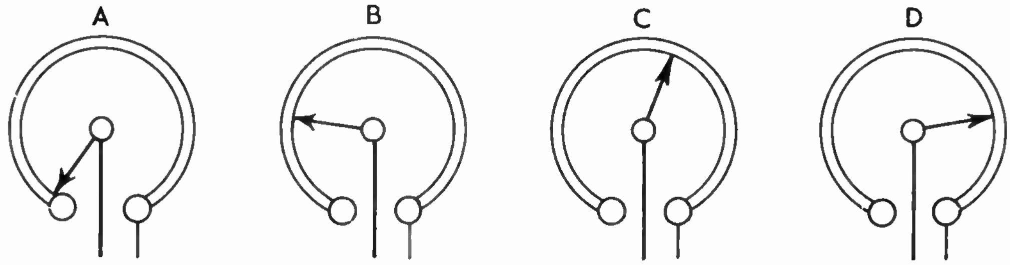


Fig. 24-12. Typical resistance curves or tapers for control potentiometers.



	A	B	C	D
VOLTS	100	100	100	100
OHMS	2500	2000	1000	500
MILLIAMPERES–	40	50	100	200
WATTS	4	5	10	20

Fig. 24-13. Changes of current and power dissipation as a potentiometer rotor is turned.

ated. If the control is in a sound amplifier circuit there will be much noise as the unit is operated. Good quality resistors in well designed circuits should not develop trouble of this kind during the normal life of a receiver.

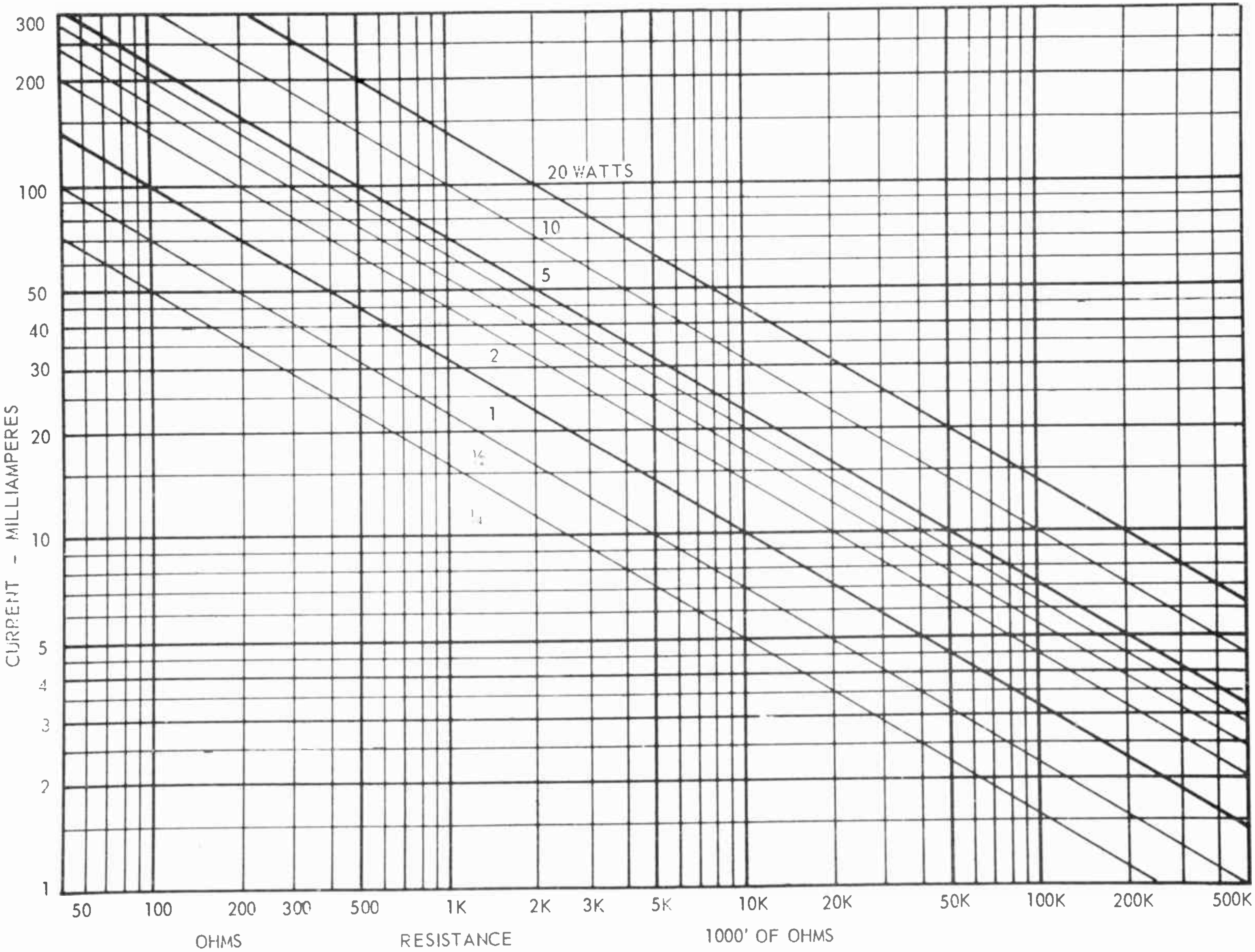
Most replacements are required because the resistors are overloaded with excessive current. Trouble then may develop because of arcing or sparking wherever the rotor makes contact on the resistance element, and the element will be burned at some places. Excessive current may so heat the resistance element as to burn it out or warp the supporting insulation and prevent good contact by the rotor.

2 6 Adjustable resistors sometimes are rated for both resistance and maximum current they will carry without becoming damaged. Most resistors are rated for resistance and power in watts they will dissipate as heat without being damaged. Wattage ratings are based on having a uniform current in the entire resistance element, all the way from one outside terminal to the other, or are based on having the entire resistance in the circuit. This also is the maximum current that may flow in any part of the resistance, since excessive current in any small portion of the total resistance will burn out that small part just as surely as would the same current in the entire length of the resistance element.

As an example of how wattage ratings affect the use of a resistor consider a unit rated at 1 watt. If the rotor is turned to a position where only half the total resistance is carrying current, then only ½ watt of power may be safely dissipated in that portion of the resistance. If only one-fourth the total resistance is in the circuit only ¼ watt of the power could be dissipated without danger of burnout. But the limit of current, measured in milliamperes, would be the same no matter how much or how little of the resistance element is carrying the current.

In Fig. 24-13 is represented a potentiometer whose total resistance is 2,500 ohms. To the center and right-hand terminals is connected a potential difference of 100 volts. In diagram *A* the rotor is in such position that the entire 2,500 ohms of resistance is in circuit. Using your regular formula for current shows that the potentiometer must be carrying 40 milliamperes. We may compute the power dissipation with a formula used in an earlier lesson, like this.

$$\text{Power, watts} = \frac{\text{volts} \times \text{milliamperes}}{1000} = \frac{100 \times 40}{1000} = \frac{4000}{1000} = 4 \text{ watts}$$



21-11. Service chart for resistance, current, and watts of power dissipation.

Provided the entire resistance always were to remain in the circuit, and the potential difference never would exceed 100 volts, we could use a potentiometer rated at 4 watts. In diagram *B* the rotor is turned to leave 2,000 ohms in circuit, still with 100 volts potential difference. Current now becomes 50 milliamperes, and power dissipation 5 watts. With the rotor turned still farther, in diagrams *C* and D, the resistances drop still lower while currents and power dissipations rise. At *D* we should need a 20-watt potentiometer. Movement of the rotor beyond this position would allow current and power dissipation to go so high as to almost certainly burn out the resistance element at its right-hand end – assuming that the voltage source could maintain the potential difference.

4 It becomes evident that potentiometer current must be limited to a value which will not exceed the current corresponding to the wattage rating of the unit. Usually there will be enough resistance in series with the potentiometer, somewhere in the external circuit, to satisfy this requirement. Potentiometers have to be selected on the basis of the worst that possibly can happen in the way of high current, not on the basis of normal working conditions.

Since it is excessive current that burns out potentiometers, and other resistors for that matter, it is desirable that we be able to determine maximum safe currents for units of any of the common wattage ratings. If we know the rating in watts and the maximum applied potential difference in volts it is possible to compute the maximum current by a slight change in the preceding formula for power in watts. The altered formula looks like this.

$$\text{Current, milliamperes} = \frac{1000 \times \text{watts}}{\text{volts}}$$

We don't have to know the resistance, only the power rating and the applied voltage. For example, supposing you have a 4-watt resistor on which the potential difference will be 100 volts, what is the maximum current in milliamperes? Putting the known values into the new formula gives,

$$\text{Current, ma} = \frac{1000 \times 4}{100} = \frac{4000}{100} = 40 \text{ milliamperes}$$

If we know the power rating in watts and the resistance in ohms there is another formula that will tell the maximum current in milliamperes. Here it is.

$$\text{Current, milliamperes} = 1000 \times \sqrt{\frac{\text{watts}}{\text{ohms}}}$$

This formula is not easy to use, because most of us don't extract square roots with any great speed. To make determination of maximum currents really easy the chart of Fig. 24-14 has been prepared.

Along the bottom horizontal scale are shown resistances from 50 ohms to 500,000 ohms, a range including most of the values commonly found in television and radio receivers. The left-hand vertical scale lists currents between 1 and 300 milliamperes, again including most of the usual values. The diagonal lines are for power ratings in watts. There is one diagonal line for each of the wattage ratings generally available, from ¼ watt up to 20 watts.

To determine maximum current with the help of the chart first find the resistance on the bottom scale. From this resistance follow straight upward to the diagonal line for the power rating of the unit in question. From the intersection of the lines for resistance and power follow straight over to the left-hand scale and there read the maximum current in milliamperes.

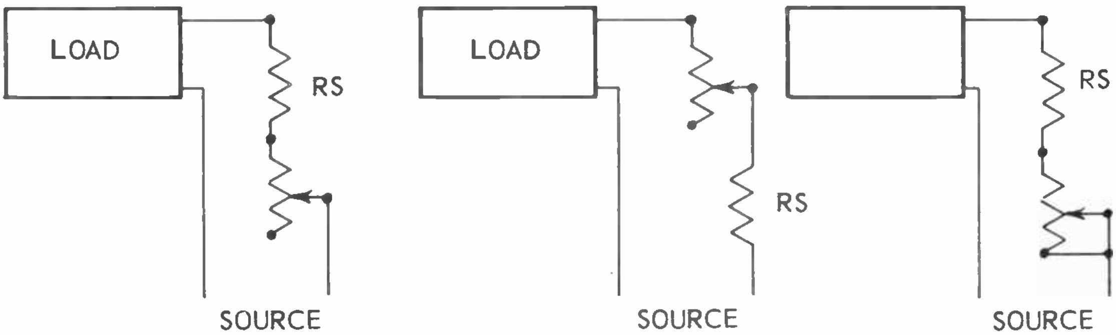


Fig. 24-15. Controls with external resistances in series.

Here is an example. You have a 2000-ohm 5-watt resistor, what is the maximum current it should carry? Follow upward from the 2000-ohm (2K) point on the bottom scale to the diagonal line which applies for 5 watts. Then, tracing to the left, you come to 50 milliamperes as the maximum current.

The scales of the chart are graduated logarithmically. Equal differences in ohms get closer and closer together as you go from left to right, and equal differences in currents get closer together as you go from bottom to top. The advantage of logarithmic scales over linear scales is that a great range of values may be covered with equal percentage of accuracy in reading anywhere on the chart. Spacing is the same between 100 and 200 ohms as between 1,000 and 2,000, and between 10,000 and 20,000, and between 100,000 and 200,000 ohms. In each of these cases the second value is exactly double the first one, or is 200 per cent of the first one. Then if you should make an error of something like an eighth of an inch in tracing from one point to another, this eighth of an inch will be the same percentage difference between true and assumed values anywhere on the chart.

This chart comes in very handy for selecting resistors of correct power rating when you know the ohms and milliamperes, as usually you do. As an example, assume that you are going to use 7,000 (7K) ohms resistance to carry a current of 15 milliamperes, how many watts must be allowed for? Find the intersection of the vertical line for 7,000 ohms, which is the second one to the right of the 5K line, and the horizontal line for 15 milliamperes, which is between the lines for 10 and 20 milliamperes.

This intersection is about half way between the diagonal lines for 1 watt and 2 watts. Therefore, a 1-watt resistor would be too small, and doubtless would overheat. A 2-watt unit will be large enough. Any greater wattage rating might be used.

The chart which we are using here for selection of necessary wattage ratings and for determination of maximum currents for adjustable resistors is just as useful when you are working with fixed resistors of any kind. Selection of suitable resistors is a problem continually recurring in service operations of all kinds.

POTENTIOMETER CONNECTIONS. Now we shall look at a few of the numerous ways in which potentiometers are connected into control circuits. Some of the ways are common and others not so common. Circuits will be shown in their simplest possible forms, consisting only of loads (which might be anything), of sources of current and voltage, and of the control units. As you trace current paths in service diagrams these connections will appear mixed in with others which might cause confusion were you not familiar with the basic operating principles.

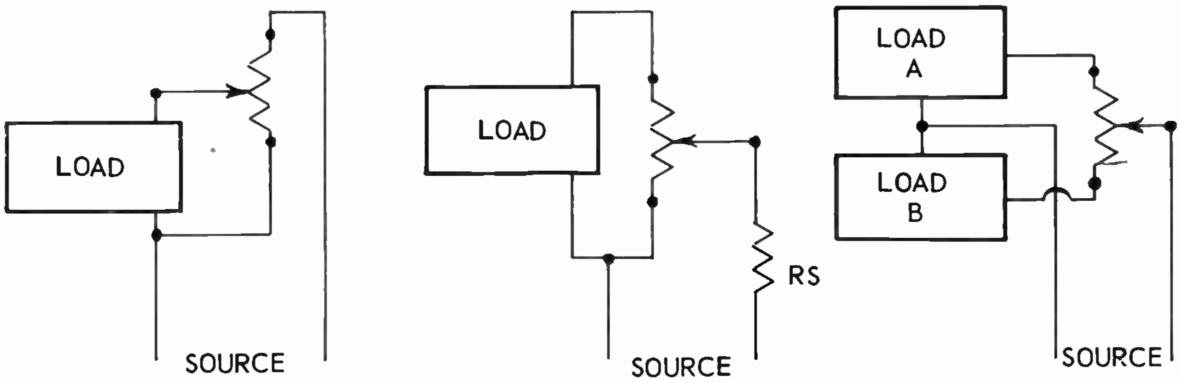


Fig. 24-16. Control circuits in which potentiometers are paralleled with loads.

In all three diagrams of Fig. 24-15 there is a fixed resistor, *Rs*, in series with the control unit. At the left the series resistor is between the control unit and the load, while at the center this resistor is between the control unit and the source. A series resistance may be of such value that current in the potentiometer will be limited to a safe value even when nearly all the potentiometer resistance is out of circuit.

The series resistor has further usefulness in making for less critical adjustment of the control unit. Supposing, for instance, we need circuit resistance of 30,000 ohms but that variation of resistance never need be more than about 3,000 ohms one way or the other. A 35,000-ohm adjustable unit would allow all necessary control, but rotor movement for the entire range of adjustment would occur in about one-sixth of the total travel. This would require careful setting with such a limited distance for rotor movement. By using a 10,000-ohm adjustable unit in series with 25,000 ohms fixed resistance it becomes possible to move the adjustable rotor through more than half its total travel to obtain the 3,000-ohm change in either direction. It will be much easier to make an accurate setting.

In the right-hand diagram of Fig. 24-15 the end of the potentiometer which is free in the diagram at the left is connected to the rotor. Then all resistance between this end and the rotor is short circuited, leaving the remainder active in the circuit. Should the rotor fail to make good contact with the resistance element there would not be total interruption of current, for then current from the source would go to the load through the entire resistance element. Current would be reduced, but not stopped.

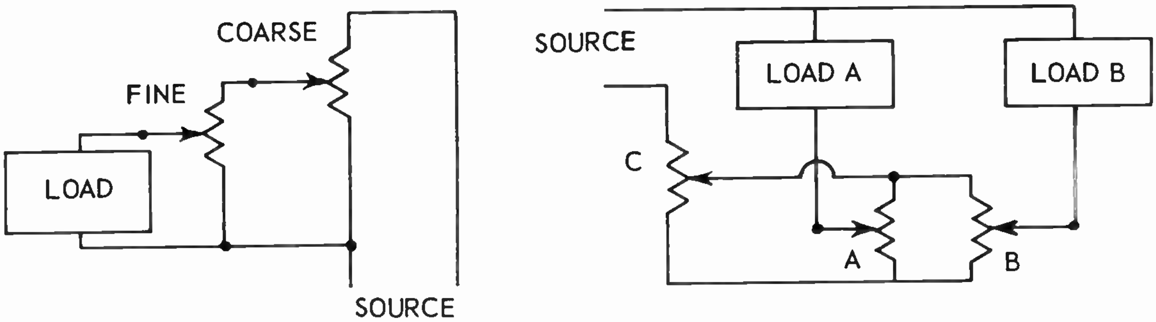


Fig. 24-17. Left: Coarse and fine control. Right: A common control and independent controls for two loads.

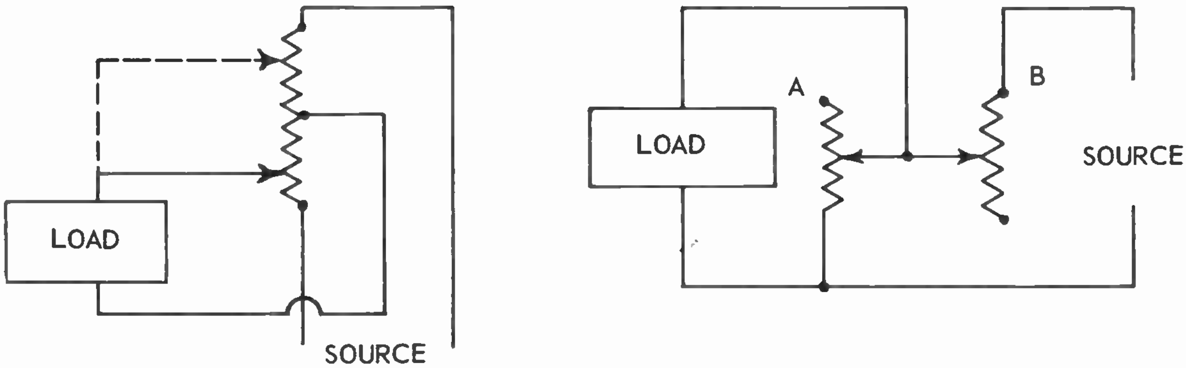


Fig. 24-18. Left: Control using a tapped potentiometer. Right: An L-pad constant inpedance control circuit.

At the left in Fig. 24-16 the load is in parallel with or is shunted across the portion of the control resistance between the rotor and the lower end. Moving the rotor to the upper end of the control resistance applies maximum voltage to the load and sends maximum current through the load. With the rotor all the way to the lower end of the control resistance the load is short circuited, but the entire resistance remains across the source.

In the center diagram of Fig. 24-16 the load is in parallel with the entire control resistance. With the rotor at the upper end of the resistance the source is connected through fixed resistor *Rs* and the control rotor to the top of the load, and is directly connected to the bottom of the load. Current through the load will depend on the relative resistances of the potentiometer element and the load itself, but will be of maximum possible value. With the rotor at the lower end of the control resistance there will be no current in the load because only the resistance at *Rs* will be across the source. Load current is varied by rotor movement.

At the right in Fig. 24-16 there are two loads. With the control rotor at the upper end of the resistance, load *A* is connected directly across the source while in series with load *B* and the source is the entire control resistance. Conditions are reversed between the two loads when the rotor is at the lower end of the resistance. At intermediate positions of the rotor the current in one load will be increased as that in the other load is decreased.

(5) In some control circuits it is necessary to provide a wide range of resistance while allowing very fine or close adjustment anywhere in this range. Any close adjustment requires that a rotor contact move quite a distance while making a rather small change of circuit resistance. Otherwise there is too much change of resistance with the least movement you can give the rotor. The problem may be solved by using two potentiometers as at the left in Fig. 24-17. The coarse adjustment is set for the approximate resistance then the fine adjustment may be changed one way or the other to obtain a precise setting.

At the right in Fig. 24-17 is a method for simultaneously varying the voltage and current to two loads while providing individual adjustments for each load independently of the other. Movement of the rotor in potentiometer *C* affects both loads. Potentiometer *A* allows separate adjustment for load *A*, while potentiometer *B* allows separate adjustment for load *B*. The lower line joining the three potentiometers might be replaced by ground connections, with the circuit completed through chassis metal. Similarly, one side of any of the control circuits being shown might be completed through ground instead of the insulated conductors of the diagrams.

One way of using a tapped potentiometer is shown at the left in Fig. 24-18. The source is connected to the outer ends of the resistance element. The load is connected between the tap and the slider. Positive and negative polarities are marked merely for purposes of explanation, they might be reversed without altering the operating principle. With the rotor between the tap and the negative end of the resistance element, as shown in full lines, the top of the load is connected to a point more negative than that to which the bottom of the load is connected. Were the rotor moved toward the positive end of the resistance element, as shown by broken lines, the top of the load would be made positive with reference to the bottom. Thus the direction of current in the load may be reversed, and the potential difference applied to the load may be varied by movement of the rotor.

a In high quality audio amplifiers it is desirable that resistances across the load and the source remain practically constant while gain or volume is varied. At the right in Fig. 24-18 is shown one way of maintaining nearly constant resistance on the source by employing a dual potentiometer. With the rotors in the position shown the resistance across the source consists of the lower part of element *A* and the upper part of element *B*. If the rotors are up, increasing the input to the load, the resistance across the source includes more of element *A* but less of element *B*, and the total resistance remains unchanged. Moving the rotors down, to reduce input for the load, still leaves the sum of the resistances across the source unchanged. Interchanging the connections for source and load would allow constant resistance across the load. This particular arrangement is called an L-pad attenuator.

TELEVISION

LESSON NO. 25

SWITCHES AND INSULATION

Training Your Mind

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON NO. 25

SWITCHES AND INSULATION

All the service and operating controls in television and radio receivers might be grouped in two general classes. In one class would be everything that alters resistance, capacitance, or inductance of the circuits. Here we would have potentio-meters and other variable resistors, also variable capacitors and variable inductors. In the other class would be all the controls that alter the circuits themselves, not the values of the parts in a circuit, but the actual connections. Here we would have various kinds of switches.

Undoubtedly the type of switch most often used is the rotary selector, of which a fairly typical example is pictured by Fig. 25-1. Units of this general style are used in many ways. In combination receivers they may provide switching between various services, such as television, f-m broadcast, standard broadcast, short wave, and phonograph operation. Rotary switches are used for channel selection in many television receivers, or for changing between low-band and high-band channels.

Service instruments of nearly every kind employ rotary selector switches to provide operation in different bands of frequencies, and to allow measurements in different ranges of voltage, current, resistance, capacitance, or inductance. A rotary switch will open and close almost any number of circuits simultaneously or in any order, will alter the connections between parts, connect and disconnect various parts, shift parts from one circuit to another, and do just about anything desired as you turn the knob or pointer.

The switch of Fig. 25-1 has three sections. The sections sometimes are called gangs, or they may be called decks. Each section is a complete switching mechanism in itself. All are operated together by the single shaft. Depending on what is to be accomplished, there may be anywhere from one to as many as twenty sections.

One section, in a simple switch, may consist of the parts shown by Fig. 25-2. At the left is the contact ring or wafer made of insulating material. Mounted on the ring are six metallic contacts, to the outer ends of which may be soldered the circuit wires. All the contacts extend inwardly beyond the insulation, but the contact at the lower right extends in farther than the others. The drawing at the center shows the rotor, which is a disc of insulation carrying a metallic ring that has an extended tongue. The shaft passes through a slot in the rotor insulation so that this part of the switch may be turned within the stationary contact ring.

With the contact ring and rotor assembled they have the relations shown at the right. The longest contact is on the common terminal to maintain electrical connection with the rotor ring no matter where the rotor is turned. The tongue of the rotor is shown engaging contact number 1, thus making electrical connection from the common terminal through to this number 1 contact and terminal.

If the rotor is moved one-sixth turn clockwise its tongue will engage the contact for terminal 2. For every following one-sixth turn the rotor will engage successive contacts around to number 5. Somewhere in the switch mechanism will be a stop that prevents turning the rotor tongue around to the position of the common terminal.

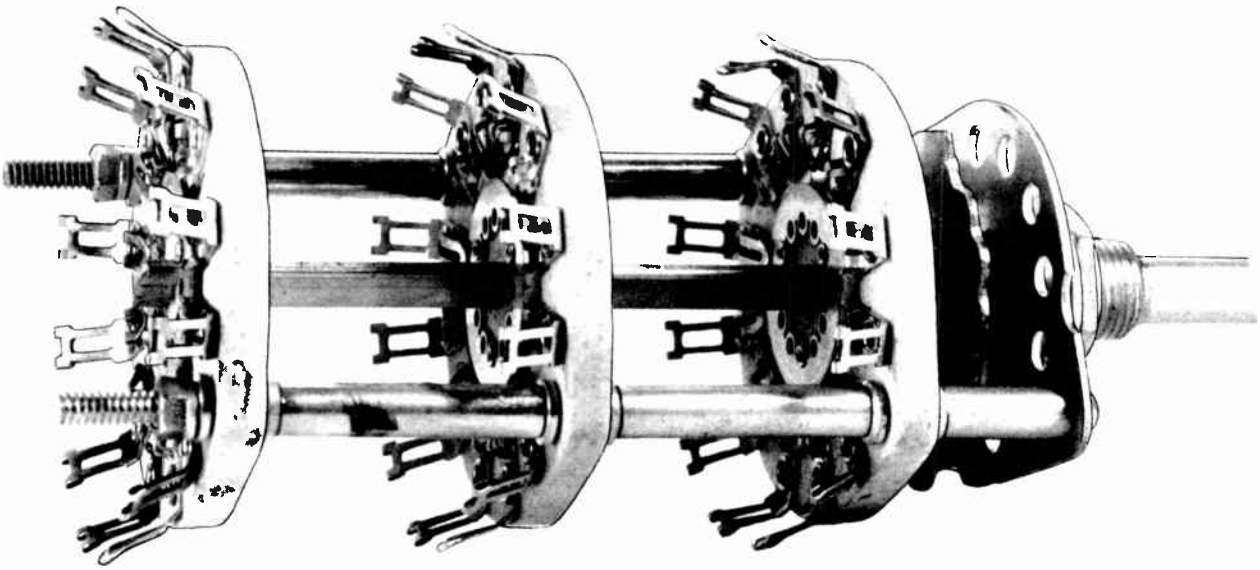


Fig. 25-1. A rotary selector switch in which the insulation is steatite.

The switch section illustrated by Fig. 25-2 has five working positions, or we might say is has five throws. This is one section of what is called a 5-position switch or a 5-throw switch. There might be anywhere from two to more than 20 positions. All other sections of the same switch would have to have the same number of positions. The rotors of all sections would turn together, and in the same direction, because they are operated by the same shaft at the same time.

We have talked about the rotor being turned clockwise from terminal 1 around to terminal 5. Were we looking at the opposite side of the section the rotor would turn counter-clockwise from number 1 to number 5. You might make a small sketch on a piece of paper, draw an arrow showing rotation, then turn the paper over and hold it up to a light to see how direction of rotation depends on how you look at a switch.

The switch of Fig. 25-2, or any other type of 5-position switch, may be shown on diagrams by any of the symbols in Fig. 25-3. The symbol at *A* shows the rotor ring, with contact terminals indicated by arrowheads. At *B* the rotor is indicated by an arrow. At *C* the rotor is shown by an arrow which, quite apparently, may move along the contacts. At *D* the rotor is shown by a straight bar with the tongue an extension at one place on the bar. It would be possible to use any other symbols which make it clear that a single common terminal may be connected to any of several other terminals by operation of the switch.

7 The switch sections of Figs. 25-2 and 25-3 will serve to connect a single conductor to any of several other conductors. The single conductor is connected through to the rotor ring, and all the others go to numbered contacts and terminals. No matter how many of the numbered contacts we place around the ring it still will be possible to switch only the single conductor that goes to the rotor. A design that switches only one

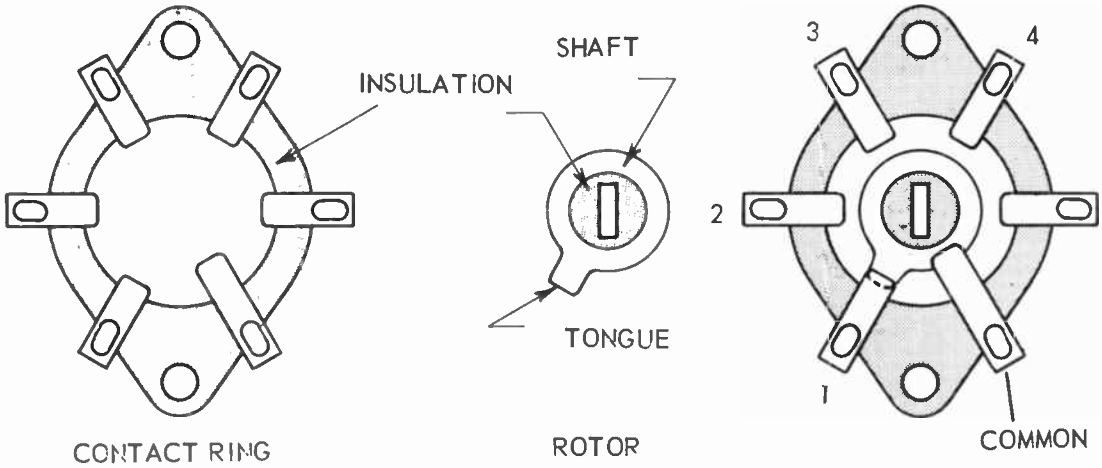


Fig. 25-2. Relations of parts in one section of a selector switch.

conductor is called a single-pole or single-circuit section or switch. It would be possible to switch additional conductors by adding more single-pole sections, using one section for each conductor to be switched.

By using a different rotor construction we may switch two, three, or more conductors with only one section. Fig. 25-4 shows a design for switching three conductors connected to three common terminals marked *A*, *B*, and *C*. On the rotor are three metal segments, each one insulated from the others. The long contact for each common terminal rests on one of the segments, and the extended tongue of each segment reaches one of the shorter contacts.

In the left-hand diagram we have electrical connection from *A* to *A1*, from *B* to *B1*, and from *C* to *C1*. At the right the rotor has been turned 30 degrees clockwise. Now the connections are from *A* to *A2*, from *B* to *B2*. and from *C* to *C2*. Another 30-degree turn would make connections from the common terminals to *A3*, *B3 and C3*. On the switch would be a stop to prevent the rotor tongues from reaching the common terminals.

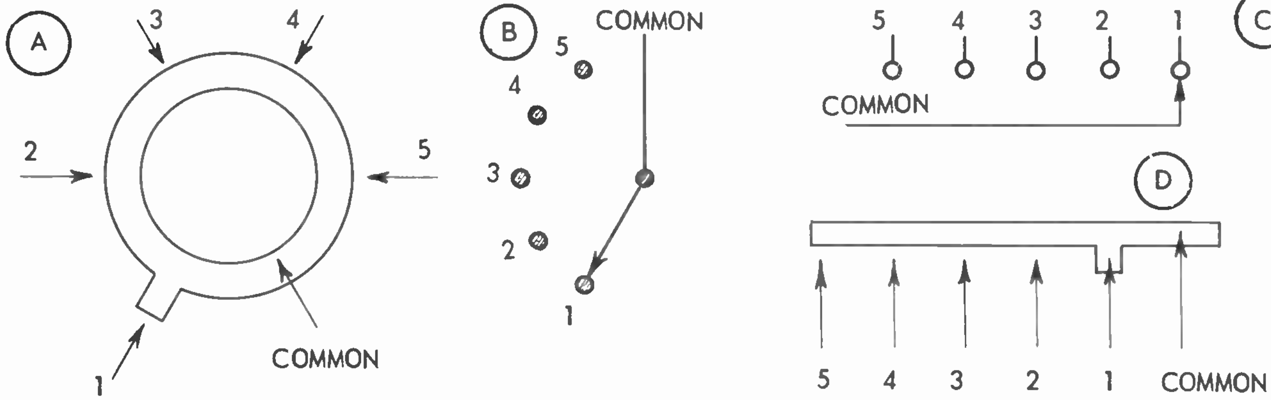


Fig. 25-3. Various symbols used to indicate selector switches on circuit diagrams.

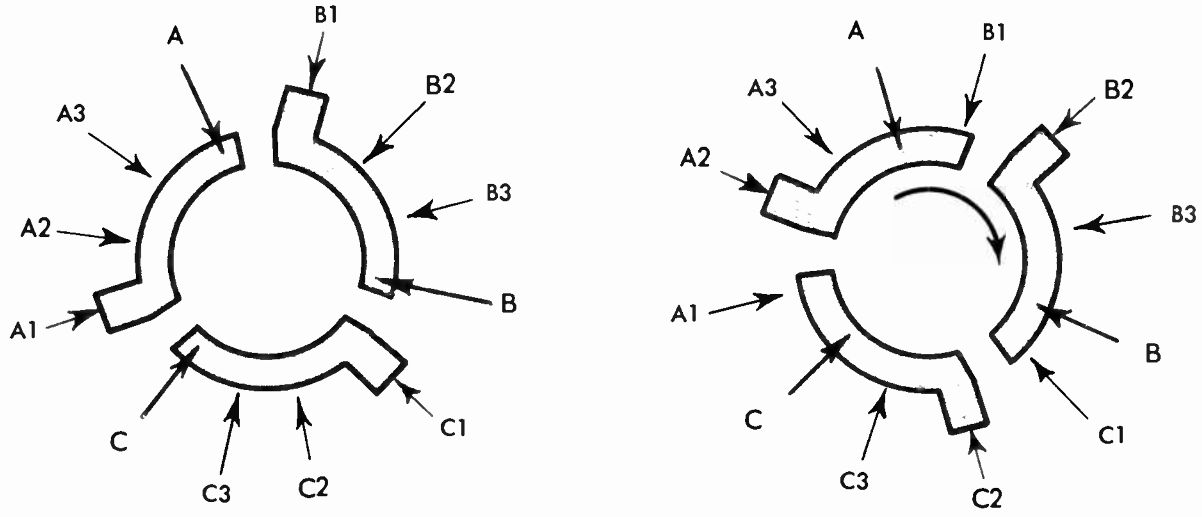


Fig. 25-4. Rotor construction for a 3-pole or 3-circuit switch section.

Here we have a 3-pole switch or a 3-pole section. A complete switch may be built up with sections having the same or different numbers of poles.

Rotary switches are available with either shorting contacts or else with non-shorting contacts. Non-shorting construction is illustrated at the left in Fig. 25-5. The tongue of the rotor is resting on contact 2, and, in broken lines, is shown passing from contact 2 to contact 3. The rotor tongue is narrow enough to break its connection with contact 2 before making connection with contact 3. Contacts 2 and 3 never are electrically connected together through the rotor. This would be true all around the switch, every connection would be broken before the next one is completed.

In the right-hand diagram the rotor tongue is wider. It is shown resting on contact 2, and in broken lines is shown passing between contacts 3 and 4. The tongue retains its connection with contact 3 until it has made connection with contact 4. As this rotor is turned it always will close a new circuit before opening the last one. This design is a shorting construction. Most selector switches in television and radio receivers are of the shorting type, so there never will be complete stoppage of pictures or sound as the switch is operated. Non-shorting switches are common in testing instruments, where sometimes there might be damage were one circuit to be shorted on another during measurements.

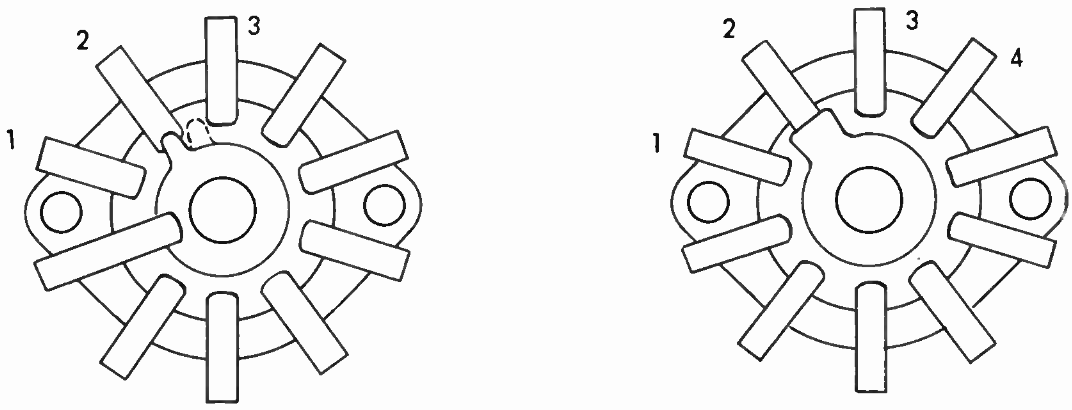


Fig. 25-5. Construction for a non-shorting switch (left) and a shorting type (right).

Fig. 25-6. Rotor design for selecting one contact or terminal while grounding all others.

There are other types of selector switch sections that will complete a circuit to any one terminal while keeping all other terminals shorted together and connected to ground or any other point desired. This is done as shown by Fig. 25-6. There is one metal segment or ring on the front insulation and another on the back. The front and back segments are insulated from each other.

The front segment has an extended tongue for making the line connection to any one of the terminals. The tongue is shown on contact 2. The back segment has an open slot in the same position as the tongue on the front segment, consequently makes no connection to the contact which is at the same time connected to the front segment. The remainder of the back segment makes connection to all the other contacts and terminals. Terminals having like numbers are wired together and to their respective circuits. Fig. 25-7 is a picture of a switch having a single section of this general type.

In some television tuners, and in other parts too, it is required that certain connections of a group be shorted together one after another, and grounded or connected to some one line. This may be accomplished with a switch section such as shown in Fig. 25-8. The rotor here is in position to short together all terminals from 1 through 5, while leaving terminals 6 through 9 separated or alive. A switch of this style may be shown by an extended diagram as at the right.

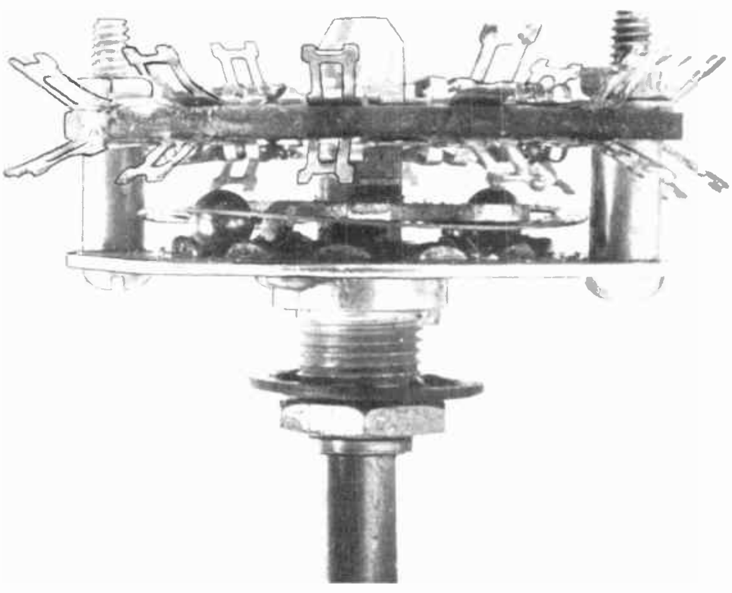


Fig. 25-7. A switch having rotor segments and contacts on both sides of the insulating ring.

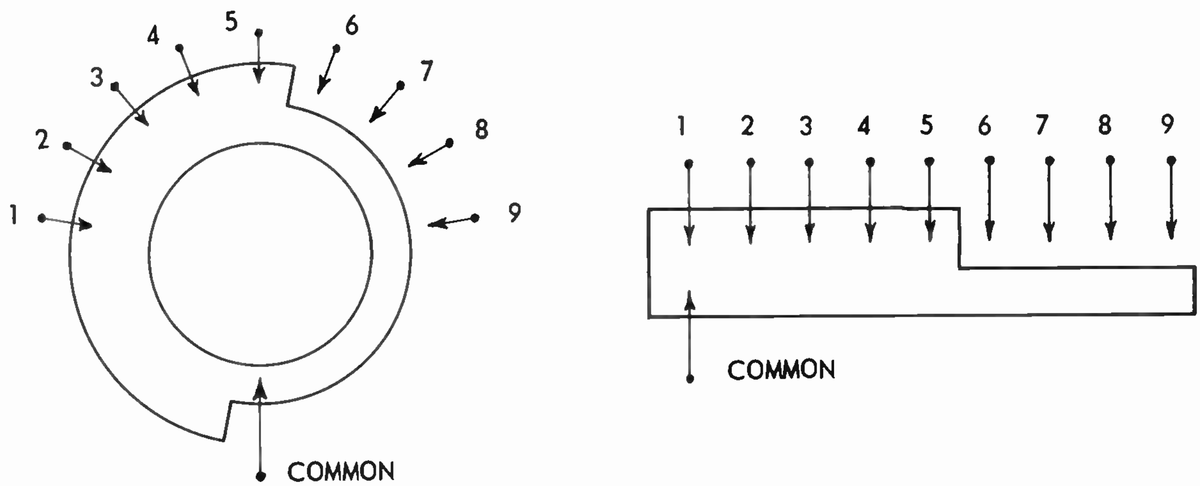


Fig. 25-8. Rotor design for a switch section that shorts its terminals one after another.

Selector switches of styles so far described are of what may be called stock types. They are available in completed form from supply houses, or may be assembled from stock parts. Stock switches or standard switches may be adapted to handle most control circuits, often by using only part of the contacts on one or more sections, by connecting contacts together in various groups, and by using enough sections to meet the requirements.

You will find, however, that controls in most commercial receivers are handled by specially designed selector switches. This allows simplified construction, and the number of sections may be reduced by making rotor segments of such shapes and with such extensions and gaps as will allow one rotor and one section to make many circuit changes.

At the left in Fig. 25-9 is pictured a special rotor having three segments. The rotor is shown in its fully clockwise position, and may be turned from here counter-clockwise through five additional positions. We shall consider the terminals to be numbered counter-clockwise, starting with number 1 at the upper left. There are no terminals or contacts in positions 3 and 12, and there is no extended contact on terminal 7, which is used merely as a convenient junction point for wiring.

Circuit connections for special control switches often are shown as at the right, which is a connection diagram listing all the terminals and all positions. With the switch in its first position, as pictured, terminals 4 and 5 are connected together, also 6 and 8, and 10 and 11. These interconnections are shown on the top line of the diagram. Connections for the remaining positions are shown on the other lines.

SWITCH CONSTRUCTION. Rotary switches usually are designed for mounting in the same manner as potentiometers, through a single panel or chassis opening 3/8 inch in diameter. The method is described in the lesson on potentiometers. Shafts ordinarily are 1/4 inch in diameter, are shaped to take various knobs, and, in replacement units, come long enough for cutting down to the required length. The operating shaft and framework of the switch are insulated from rotor segments and contact rings, as was explained in connection with Fig. 25-2.

5

Although the switch shaft and rotors are easily turned from one position to another by a knob or pointer,

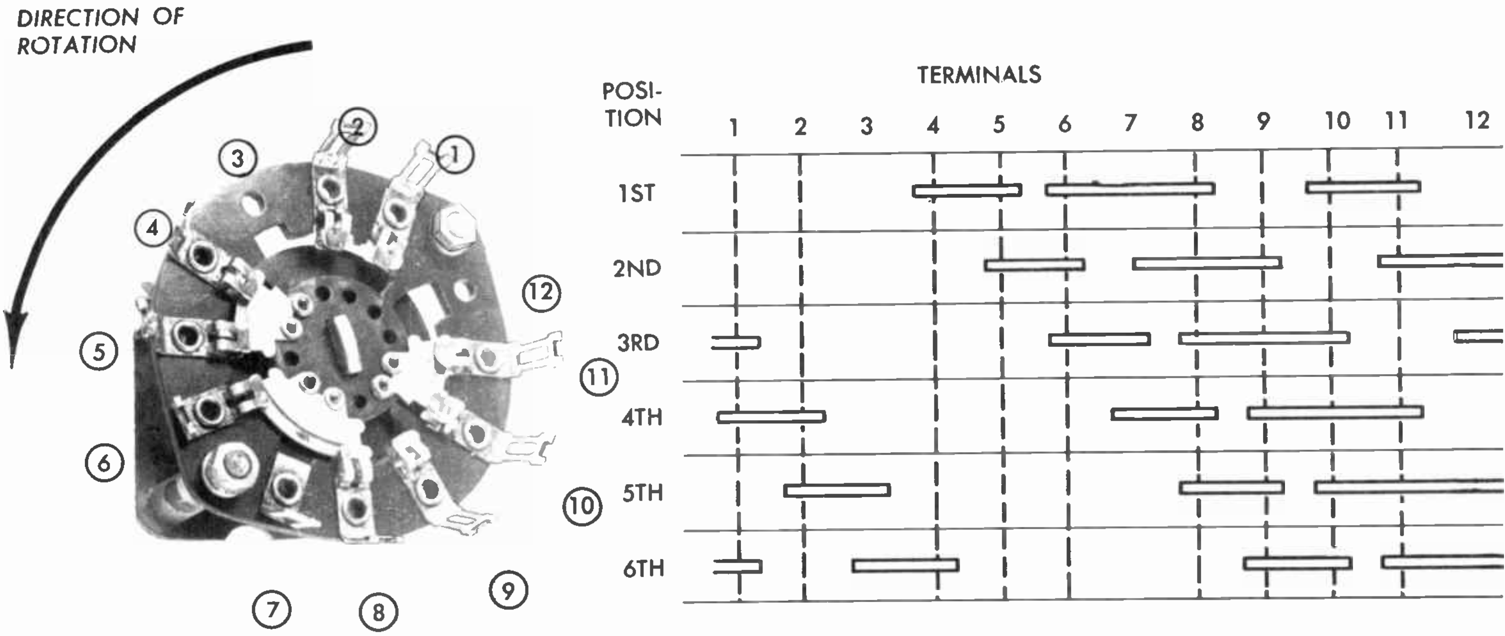


Fig. 25-9. A special type rotary switch and its connection diagram.

the rotors must be held firmly at each of the positions where segments made correct engagement with terminal contacts. This is accomplished by devices called detents. A detent ordinarily consists of one or more hardened steel balls held under spring tension so they snap into depressions or shallow holes. With the construction of Fig. 25-7 the detent balls are carried around through stationary depressions formed on the back plate. One of the balls is plainly visible where it fits into the flexible flat spring that rotates with the shaft. In the switch of Fig. 25-1 the detent ball and spring are stationary, and on the shaft is a disc with edge corrugations in which the ball rests at each switch position.

Fig. 25-10 shows the parts of a disassembled selector switch. The threaded extension for mounting is permanently fastened to the index plate or back plate. The sections are supported on this plate by the long tie bolts, and are held separated by the spacer tubes.

The rotor of some switches may be turned around and around. In others there are one or two stops preventing the rotor from traveling too far in either direction. A stop may consist of a steel pin extending from the shaft so that it strikes against some raised part of the back plate, or it may be an extension on the detent disc, or a screw inserted through the back plate. In Fig. 25-10 is shown an adjustable stop whose extended lip may be put through any of the holes in the index plate, so that this lip will be struck by an extension on the detent disc.

The mounting nut and lock washer usually will hold a switch securely enough in the panel or chassis to prevent turning the whole thing when pressure is applied to the knob or pointer. Additional locking may be provided by a steel pin which extends from the mounting bushing, or there may be an extended lug on the back plate. The pin or lug passes into a small hole in the supporting panel or chassis.

2 *LINE SWITCHES.* The switch for turning a receiver on and off most often is mounted on and operated with the volume control potentiometer. A combined control and switch unit is illustrated at the left in Fig. 25-11. The three terminals of the potentiometer may be seen at the upper right. The two terminals of the line switch are mounted in the insulating back of the unit.

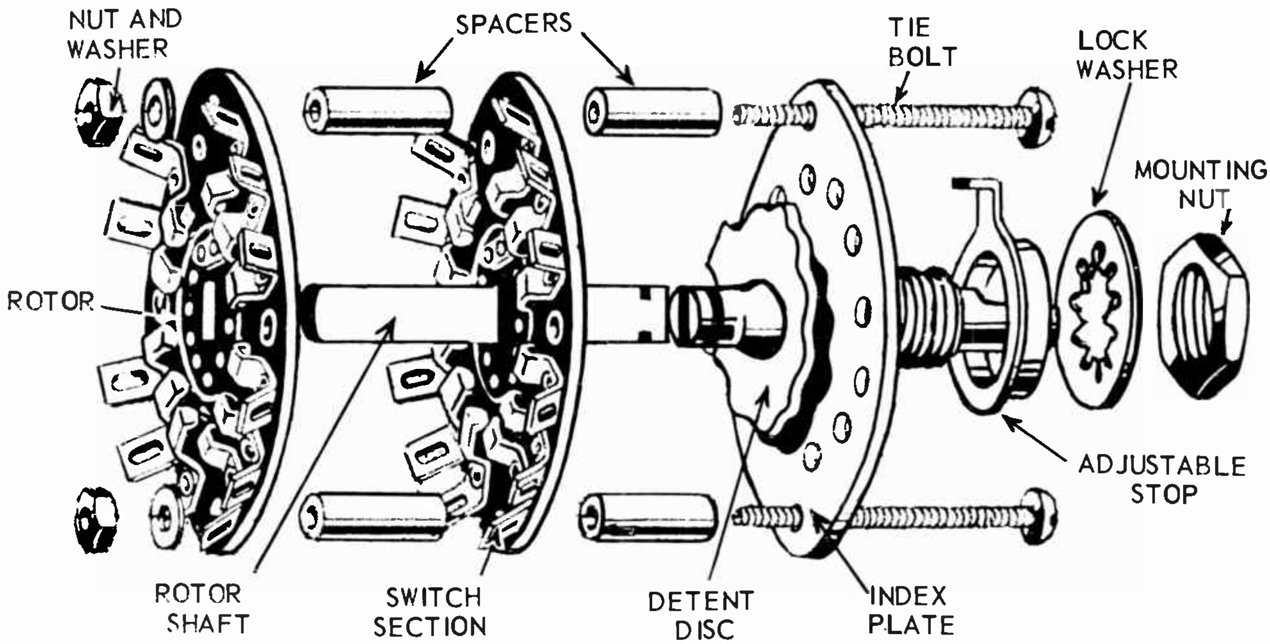


Fig. 25-10. The parts of a Centralab selector switch.

With the volume control knob fully counter-clockwise, from the front of the panel, the line switch is open and volume is minimum or zero. During the first 30 or 40 degrees of clockwise rotation the line switch is closed, and will be again opened when the knob is turned back to its original position. Further clockwise rotation after closing of the switch will increase sound volume.

The line switch is actuated by an extension or finger on the rotor of the potentiometer of Fig. 25-11 the switch finger on the rotor may be seen pointing toward the left. Most replacement potentiometers suitable for volume controls are designed so that a line switch may be attached to the back of any unit having suitable resistance range. The resistance taper of this unit must have practically no increase during that part of the rotation in which the switch is closed and opened. In some controls the resistance element stops short of the point at which switch operation takes place.

4 *SWITCH CARE.* The surfaces of switch contacts, also the insulation of rotors and contact rings, may be cleaned with carbon tetrachloride. This liquid should be applied on flat surfaces with a soft cloth. A small stiff brush may be used for corners, between contacts, and for any spots not easily accessible. This cleaning of contacts surfaces seldom should be needed, inasmuch as most designs provide self-cleaning action as rotor segments rub between the contacts. Rosin which has spread from soldered joints onto metal parts or the insulation may be dissolved and removed by a cloth moistened with denatured alcohol.

To lessen wear and allow smoother action of rubbing contacts they may be lightly lubricated with switch contact oil prepared especially for this purpose. Apply the oil by means of a small swab or any small piece of cloth lightly moistened with the lubricant. Detents and other rubbing or bearing parts of switches may be lubricated with vaseline, or preferably with grease made for the purpose. Either a special grease or else ball bearing grease for bicycles and automobiles will stay in place better than vaseline.

Fig. 25-11. A combination volume control and line switch.

When detent balls become lost and you have no regular replacements, it usually is possible to use bearing balls, which come in several suitable sizes at bicycle repair shops. Cover the new ball with heavy grease, hold the detent spring back with a screw driver blade, and slip the ball into place while grasping it with thin nose pliers or ordinary tweezers.

If switch contacts fail to press tightly enough against rotor segments it often is impracticable to successfully restore the tension. If you decide to try, it is best to first disassemble the switch and remove the rotor from the contact ring. Then the tips of the contacts may be very carefully pushed toward each other from both sides of the ring, making certain that their final position is such that the rotor will engage them correctly.

INSULATION. In this lesson and the one before it we have dealt largely with parts in which insulating housings and supports play an important part. It may be well to get a little better acquainted with the various common insulating materials and some of their properties. The kinds of insulation used in a receiver are, of course, determined by the designer, and there would be nothing you could do about it even were changes desirable—which in all probability they are not. However, when it comes to making repairs and replacements you have a great deal to do with selection of insulation suited to the requirements, and as a good service technician it behooves you to know the essential facts.

3 The prime purpose of insulation is to confine electric currents to certain conductors and prevent potentials from going where they are not wanted, while at the same time providing such support as is necessary for the current-carrying parts. A principal requirement of insulation is high resistance. Yet the matter of resistance turns out to be of little concern, because even such things as paper, cloth, and wood will have more than ample resistance in any reasonable thickness. A piece of the poorest grade paper an inch square and 1/25 inch thick will have resistance between opposite surfaces of something like 500 megohms.

There is possibility of electrical leakage on the surface of insulation. Across one square inch of perfectly clean steatite insulation, such as used for the contact rings of the switch in Fig. 25-1, the resistance is on the order of 250 million megohms. But if that surface gets very dusty or moist or greasy, the surface resistance may drop so low that a few hundred volts will cause appreciable current to flow. Leakage across dirty or wet insulation causes much trouble.

8 More important than its resistance is the dielectric strength or breakdown voltage of insulation. Dielectric

strength is measured as the number of volts potential difference that will put a hole through insulation, whereupon there will be a rush of current through the opening. If you apply voltage to opposite sides of a thin sheet of insulation, and gradually step up the voltage, there will be no measurable flow of current through the insulation until it punctures. When this happens there has been produced such an electrical stress that electrons are literally torn from the molecules of insulating material, and the substance disintegrates. The arc then generates enough heat to char the insulation around the hole, and we call it a burnout. The voltage at which all this happens is just in excess of the dielectric strength of the material.

In tables which list dielectric strengths the values for different substances usually are given in volts per mil, which means volts per thousandth of an inch thickness. You cannot determine actual puncture voltage by multiplying volts per mil by thousandths of an inch thickness, because dielectric strength per unit thickness decreases at varying rates as the material is made thicker.

As an example, hard fibre 0.005 inch thick has dielectric strength of about 200 volts per mil, but when the fibre is 0.250 inch thick its dielectric strength is down below 100 volts per mil. The thicker section will provide more total insulation than the thin one, but the gain will not be proportional to thicknesses. Also, breakdown voltage is less for alternating than direct potentials, is less at high frequencies than at low ones, and decreases with passage of time, with rise of temperature, and with moisture in the material.

We should not forget that air is an insulator. Sparking voltage between two sharp points in air is about 25,000 volts per inch of separation, and this ratio remains unchanged for greater and less spacings between the sharp points. Between two rounded or flat surfaces the dielectric strength of air decreases as separation (thickness of air) becomes greater, just as with solid insulators. Dielectric strength of air is about 10,000 volts for 1/10 inch between rounded surfaces, and becomes about 55,000 volts with one inch of separation. Now you know why we avoid sharp points and edges in all wiring connections for high-voltage television circuits.

It is when we get into the very high frequencies of television and f-m radio that certain properties of insulation take on added importance. It is not so much that there may be more leakage of currents and voltages at these frequencies as because of the many ways in which signal energy may be lost.

Any insulation which is in an electric or electrostatic field is working under conditions similar to those affecting the dielectric in a capacitor. There are electric fields in spaces between all conductors whose potentials are different from one another. Of course, these electric fields are not so concentrated as in a capacitor, but they exist in considerable strength around all wires and coils which are in plate, screen, and grid circuits. Then all insulating materials used near these conductors may cause energy losses like the losses which occur in capacitor dielectrics.

In insulation subjected to alternating electric fields the electrons and the molecules of which they are a part are pulled first in one direction and then in the opposite direction. Electrons are not pulled out of the molecules, nor do the molecules move within the body of insulating material, but the continual stresses and strains cause heating. Energy that produces this heat is lost. The effect is called dielectric loss. A piece of insulation doesn't have to get very warm to be losing or consuming a lot of energy, at least in proportion to the total energy of signal voltages which are doing the work.

Were any substance capable of acting as a perfect dielectric it would behave somewhat like a perfect spring in the field of mechanics. If there were such a thing as a perfect spring, and you compressed it by using some certain amount of energy, every bit of that energy could be recovered as the spring expanded. In any actual spring some of the energy will be wasted by internal friction of the molecules as the spring compresses and expands. This much of the applied energy is lost and cannot be recovered.

Were there such a thing as a perfect dielectric every bit of energy in the electric fields reaching the dielectric would return to the surrounding conductors as the fields collapsed. In a capacitor this would mean that all energy for charging would be returned to the circuit during discharge. No substance is a perfect dielectric, all use up some of the energy entering them from surrounding electric fields. The fraction of the total energy thus lost is called the power factor of the dielectric or insulation. Power factors of high grade insulation range from as little as 3/100 of one per cent up to more than one per cent. You may feel that a loss of one per cent is not serious, but if this loss is repeated at enough places in a receiver it can be very serious.

Still another thing that affects energy loss in insulating material is its dielectric constant. As you will remember, the dielectric constant of a substance is the number of times it will increase the capacitance of a capacitor as compared with air as the dielectric. Dielectric constant might be considered a measure of the relative number of electric field lines drawn into the insulation. Consequently, the greater the dielectric constant the greater will be the energy loss resulting from a large power factor.

Since loss of energy increases with greater power factors and also with greater dielectric constants we may multiply these two together to get an overall measure of loss in the insulating material. The product of dielectric constant and power factor is called the loss factor. As an example, high quality insulating paper has an average dielectric constant of 2.3, and an average power factor of about 0.4. Multiplying these two quantities together gives the average loss factor for paper as approximately 0.92. Some insulating materials used in television and radio receivers have loss factors as low as 0.05, while others have loss factors of 5.0 or more.

INSULATING MATERIALS. Many of our insulating materials are of the class called phenolic resins. A few of the trade names for these substances are Bakelite, Celeron, Formica, and Micarta. The basic ingredients in all of them are phenol and formaldehyde, bonded or "filled" with such things as wood flour, cloth, paper, mica, and various fibres. The phenolic insulators are tough and elastic, with good mechanical strength and freedom from excessive softening or warping at any temperatures likely to be encountered. These materials are easily sawed, drilled, threaded, or filed to shape. The switches of Figs. 25-7 and 25-9 have phenolic insulation in the contact rings and rotors. There are special low-loss grades of phenols having smaller power factors and loss factors than ordinary kinds, suiting these grades for high-frequency applications.

Another class of insulating materials includes those made from steatite. This substance is originally obtained as a very dense soapstone or talc, by quarrying. Some trade names for the finished product are Isolantite and Lava. Steatite insulation often is called ceramic, a name which describes most products made from clays and similar substances. Ceramics would include not only steatite, but also porcelain such as used for some radio insulation.

Steatite insulation is white or light gray. It is mechanically rigid, withstands higher temperatures than most other insulators without weakening, and has relatively little tendency to absorb moisture. Steatite of all common grades is too hard for cutting or threading with shop tools. Power factors and loss factors are small enough to make this material satisfactory for use in very-high frequency apparatus.

Phenolic resins and steatites have been used for insulation since the beginning of commercial radio. Of more recent introduction are the polymerized products called polystyrene and polyethylene. Some trade names for polystyrene are Amphenol 912A, Intelin, Lustron, Styron, and Trolitul. Polystyrene is transparent. It is tough to the extent of bending before breaking, but has less mechanical strength than phenold and steatites. Polystyrene is used for sockets, insulating strips and tubes, bushings, coil forms, cable supports, and other purposes.

Polyethylene is a flexible material used for tubing, bushings, and covering for cables and wire conductors – notably for transmission lines between television antennas and receivers. The most valuable characteristics of polystyrene and polyethylene are their very small power factors and loss factors combined with high dielectric strengths.

Another class of polymerized insulation includes the vinyl resins which most often are manufactured in such a way as to have physical characteristics like rubber; stretching and bending, but returning to their original form and size. The vinyls are used for flexible tubing which may be slipped over conductors and joints, the variety called "spaghetti" in the radio service business. These materials are used also for insulation applied to wires and cables. This kind of insulation may be transparent or of any color. The loss factor is greater than for other materials which have been described.

The accompanying table lists average electrical properties of the more common insulating materials. Although the table provides some basis for comparison between substances, do not forget that values vary widely with working conditions and grade or quality, and that considerations of mechanical strength, workability, and such matters often affect the choice of insulation.

AVERAGE PROPERTIES OF INSULATING MATERIALS

Kind of Insulation	Dielectric Strength, volts per mil	Dielectric Constant	Power Factor, per cent	Loss Factor
Phenolic resins				
Molded, black	350 – 500	5.0	3.6	18.0
Low-loss, mica filled	450 – 650	5.5	0.9	5.0
Low-loss, red, yellow	400 – 500	5.3	0.5	2.7
Steatite				
General purpose	150 – 300	5.7	0.35	2.0
Low-loss	150 – 450	4.4	0.17	0.75
Ceramic, low-loss	100 – 150	6.7	0.08	0.55
Polystyrene	600 – 2200	2.6	0.031	0.08
Polyethylene	1000	2.4	0.035	0.08
Vinyl resins	400 – 500	4.0	1.6	6.4

WIRE AND ITS INSULATION. The diameters of wires used in television and radio are not specified in inches or fractions, but by gage numbers. The biggest wire we ordinarily use would be of number 8 gage, sometimes employed in connections for grounding. This wire is about 1/8 inch in diameter, not counting any insulating covering. As gage numbers increase, the wire diameter becomes smaller and smaller. The smallest common wire, which may be used in some inductors or coils, is number 16. Its diameter is 1½ thousandths of an inch, much less than the thickness of the paper on which this lesson is printed. All wire used in original assembly and in rewiring of receivers is of copper. The copper conductors are tinned for easy soldering.

The most generally used "hookup" wire for running receiver circuits and for service work in general is of number 20 gage, either solid or stranded. Stranded wire, of any gage size, consists of several small conductors whose combined cross sectional area or current-carrying capacity is the same as in a single solid wire of the same gage size. A cable consists of two or more separately insulated conductors, usually stranded. which are enclosed within an outer insulating and protective covering. Shielded wires and cables consist of insulated wires or cables around which is a closely woven flexible covering of stranded or braided copper. There may be an additional protective or insulating covering on the outside of the shield. We shall have more to do with shielded conductors in later lessons.

Hookup wire is available with various kinds of insulation, which may be had in black, white, and a variety of colors and color combinations allowing various circuits or interconnections to be easily distinguished from one another. Insulations include vinyl, previously mentioned, also rubber, and braided fabric impregnated with waxes, with cellulose acetate and other insulating substances, and sometimes lacquered for protection against moisture and fumes.

There are so many gage sizes of wire and so many conductor materials which may be used, especially where high resistance and other special characteristics are desired, that we could fill pages with tables listing the properties. Fortunately, there is a way of quickly determining the resistance of any length of practically any kind of conductor of any gage size. All the necessary information may be put on a single page.

The first thing we need is the accompanying table of wire sizes, showing cross sectional areas in numbers of circular mils corresponding to various gages. A circular mil is the area of a circle whose diameter is 1/1000 inch, it is a unit of area. Next we need the table of conductor resistivities, giving ohms per circular mil-foot for all the metals ordinarily used as conductors. A circular mil-foot is one foot of conductor whose cross sectional area is one circular mil. Finally, we need this simple formula.

$$\text{Ohms} = \frac{\text{wire length, feet} \times \text{ohms per circ. mil-ft}}{\text{cross sectional area, (in circular mils)}}$$

Let's see how it works. What is the resistance of 15 feet of number 24 gage Nichrome II resistance wire? From the table of resistivities we find that Nichrome II has 660 ohms per circular mil-foot. From the table of wire sizes we find the cross sectional area of number 24 gage wire to be 404 circular mils. Placing these values, along with the length, in our formula gives,

$$\text{Ohms} = \frac{15 \times 660}{404} = \frac{9900}{404} = 24.5 \text{ ohms, approximately}$$

Any similar problem relating to conductor resistance may be handled as easily. Figure out the resistance of 100 feet of number 10 U.S. Standard copper wire. It is little more than 1/10 ohm.

WIRE SIZES

Gage No.	Cross Section, circular mils	Diameter, inches	Gage No.	Cross Section, circular mils	Diameter, inches
8	16 510	0.1285	28	159.8	0.01264
9	13 090	.1144	29	126.7	.01126
10	10 380	.1019	30	100.5	.01003
11	8 234	.09074	31	79.70	.00893

12	6 530	0.08081	32	63.21	0.00795
13	5 178	.07196	33	50.13	.00708
14	4 107	.06408	34	39.75	.00631
15	3 257	.05707	35	31.52	.00562
16	2 583	0.05082	36	25.00	0.00500
17	2 048	.04526	37	19.83	.00445
18	1 624	.04030	38	15.72	.00397
19	1 288	.03589	39	12.47	.00353
20	1 022	0.03196	40	9.885	0.00315
21	810.1	.02846	41	7.845	.00280
22	642.4	.02535	42	6.250	.00250
23	509.5	.02257	43	4.850	.00220
24	404.0	0.02010	44	4.000	0.00200
25	320.4	.01790	45	3.063	.00175
26	254.1	.01594	46	2.250	.00150
27	201.5	.01420			

CONDUCTOR RESISTIVITIES

Material Or Trade Name	Ohms circ. mil-ft	Material Or Trade Name	Ohms circ. mil-ft
Advance	294	Manganin	270
Aluminum	17	Nickel	52
Brass, common	49	Nickel silver, 18%	198
Chromel	540	30%	240
Constantan	294	Nichrome II	660
Copper, annealed	10.4	III	540
U.S. Stan'd	10.55	IV	625
hard drawn	10.65	Platinum	72
Ideal	295	Silver	9.75
Iron, pure	60	Steel, galvanized	67
wrought	84	Tungsten	33.2
Magnesium, pure	277		

26 → 30

COMMERCIAL TRADES INSTITUTE

KNOWLEDGE + AMBITION = SUCCESS

LEARN . . .

ACT . . .

Training Your Mind

C.T.I.

Perfecting Your Skill

ENJOY . . .

"The heights by great men reached and kept
Were not attained in sudden flight,
But they, while their companions slept,
Were climbing upward in the night."

COMMERCIAL TRADES INSTITUTE

CHICAGO, ILLINOIS

TELEVISION

LESSON NO. 26

TUNING WITH RESONANT CIRCUITS

Training Your Mind

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind
C.T.I.
Perfecting Your Skill

COMMERCIAL
TRADES
INSTITUTE

Chicago, Ill.

LESSON NO. 26

TUNING WITH RESONANT CIRCUITS

FIRST PRINCIPLES

Were a person who is unacquainted with the history of radio to look inside a modern receiver they might well conclude that the whole thing depends on tubes, for every set appears full of them. Yet we had radio communication over distances of thousands of miles before we had radio tubes.

High-frequency currents and voltages for transmission used to be generated with the help of sparks, arcs, and special types of rotary generators. The resulting high-frequency radio waves were modulated with signals and radiated into space. Receivers of that day were made with coherers and with crystals for detection, and the sound signals were reproduced in headphones.

In text books published as late as 1920 the authors still talked about tubes as something rather new which probably would do a lot for radio, but they gave most attention to the older and better known methods of transmission and reception.

However, no matter how far back into the beginnings of radio we may go, and no matter what kind of apparatus we may examine, we always find one combination of parts without which radio never would have existed, nor would it exist today. This combination is a coil and a capacitor, an inductance and a capacitance. An inductance and capacitance may be made to produce the condition called resonance, and if any one thing could be rated as the prime essential in every branch of radio communication it is the condition of resonance — which we now are ready to investigate.

TUNING WITH RESONANT CIRCUITS

In the United States there are more than 2,000 radio stations operating in the standard broadcast band on carrier frequencies ranging from 540 to 1,600 kilocycles. This band of frequencies is divided into channels 10 kilocycles wide. Each Class 1 station, operating with powers of 10,000 to 50,000 watts, has exclusive use of one clear channel. There are also Class II stations with powers of 250 to 50,000 watts operating in clear channels. Class III stations share a channel with several similar stations, using powers of 500 to 5,000 watts to serve a large center of population and its adjacent rural areas. Class IV stations operate with powers of 100 to 250 watts to give purely local service in channels shared by many similar stations elsewhere.

The radiations from all these stations are "on the air" during most of the hours of every day. Even though your radio receiver is not particularly sensitive it still is within range of at least fifty of them. Then how is it that merely turning the tuning knob picks out one particular program from all those other

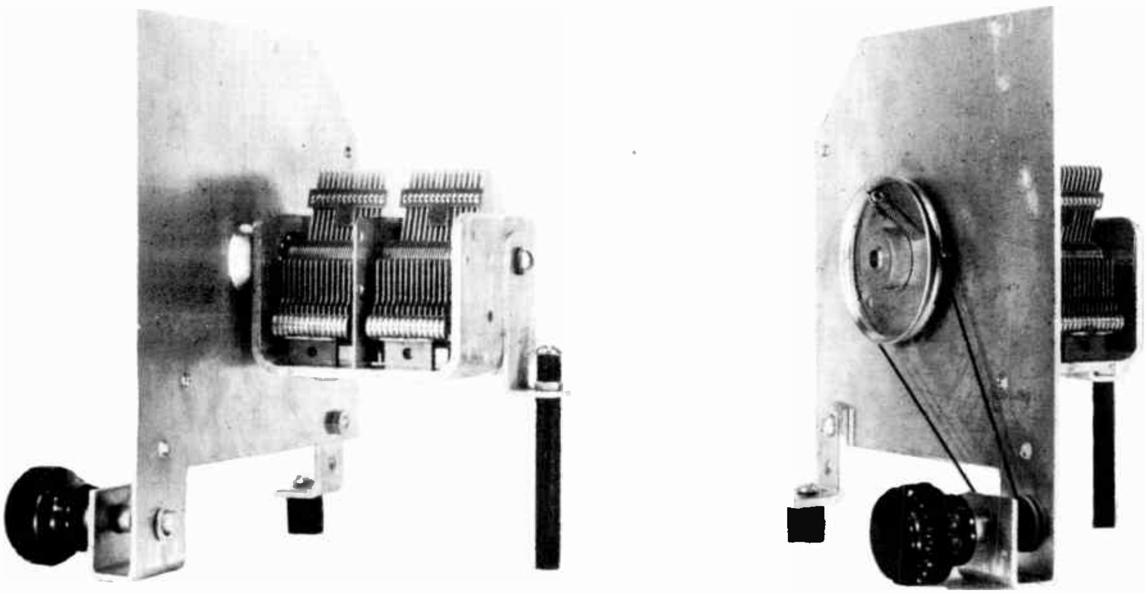


Fig. 26–1. A tuning capacitor with its drive pulleys and belt.

signals flying through space all around you? How is it possible to listen to a station 500 miles away when another is within 10 miles? How is it possible to select one station operating on a channel frequency right between frequencies of two other stations?

The selection is accomplished by using very special combinations of capacitance and inductance in the section of your receiver which is connected to the antenna or to any other conductor that collects the signal energy from space. The process of signal selection is called tuning. The necessary capacitance is furnished by one or more tuning capacitors, and the inductance is furnished by one or more tuning inductors or coils.

A dual or double tuning capacitor used in one style of broadcast receiver is pictured at the left in Fig. 26-1. In each of the two sections of this capacitor is a group of 13 stationary metal plates with air spaces between them. Attached to a shaft which may be rotated are 14 movable plates that are turned farther into or out of air spaces between and on the outsides of the stationary plates as the shaft is turned.

The capacitor shaft that carries the movable plates is turned by the tuning knob visible at the lower left of the supporting panel. The tuning knob, the drive pulleys, and the connecting belt are shown in the picture at the right. The stationary plates of the capacitor are called stator plates, and as a group they are called the stator. The movable plates attached to the shaft are called rotor plates, and their complete assembly is called the rotor.

Even with the rotor plates turned all the way in between the stator plates there remains an air space between each side of every plate and the two adjacent plates. Air in these spaces is the dielectric of the tuning capacitor. The farther the rotors are moved into the spaces between stators the greater is the area of stators and rotors on opposite sides of each layer of air dielectric, and the greater is the effective capacitance.

With the rotors all the way in mesh with the stators the capacitance of one section of this particular capacitor is 445 micro-microfarads. With the rotors turned out as far as they will go the capacitance still is 25 micro-microfarads. At intermediate positions of the rotor plates the effective capacitance is varied between these limits.

When installed in a broadcast receiver the tuning capacitor is connected to a radio-frequency transformer or antenna coupling transformer. One such transformer is shown by Fig. 26-2. At the lower end of the supporting tubing is a rather large coil which will be connected to the tuning capacitor. A little ways above is a slightly smaller coil which will be connected to the antenna. This smaller coil is the primary and the larger one is the secondary winding for the portion of this transformer used in standard broadcast reception.

Signal energy from the antenna enters the primary winding to cause high-frequency voltages and currents. There is inductive coupling between the two coils, so signal energy is transferred from the primary into the secondary winding. Every frequency reaching the antenna thus passes to the secondary winding and the tuning capacitor. At the top of the supporting tubing is another primary and another secondary winding used for short-wave reception where carrier frequencies range between 5 and 20 megacycles. For our first experiments we shall use only the lower secondary winding of this double transformer.

Inductance of the lower secondary winding is 231 microhenrys. This inductance is not changed during the operation of tuning. Inductive reactance of this secondary winding at all frequencies between 540 and 1,600 kilocycles is shown by Fig. 26-3. This reactance increases with frequency at a constant rate, which makes the change of reactance show up as a straight line on the graph.

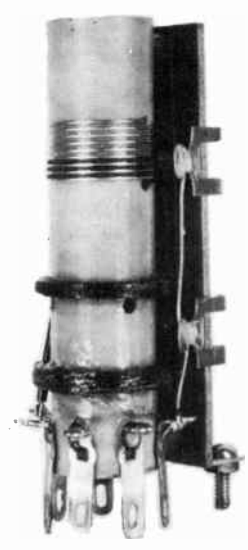


Fig. 26-2. An antenna coupling transformer for standard broadcast and some of the short-wave bands.

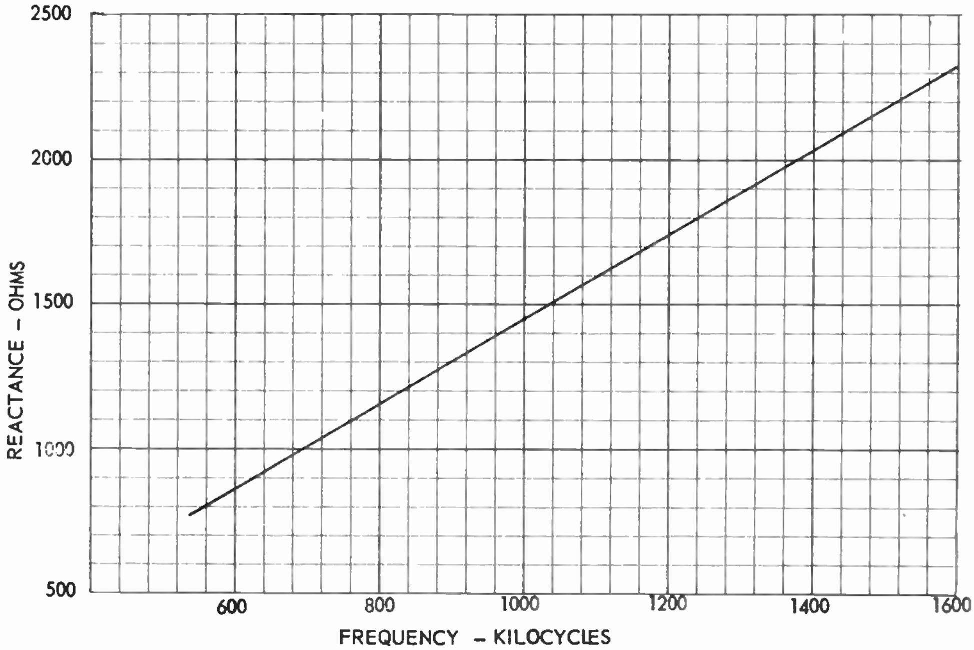


Fig. 26-3. How inductive reactance of the tuning coil changes when there is variation of applied frequency.

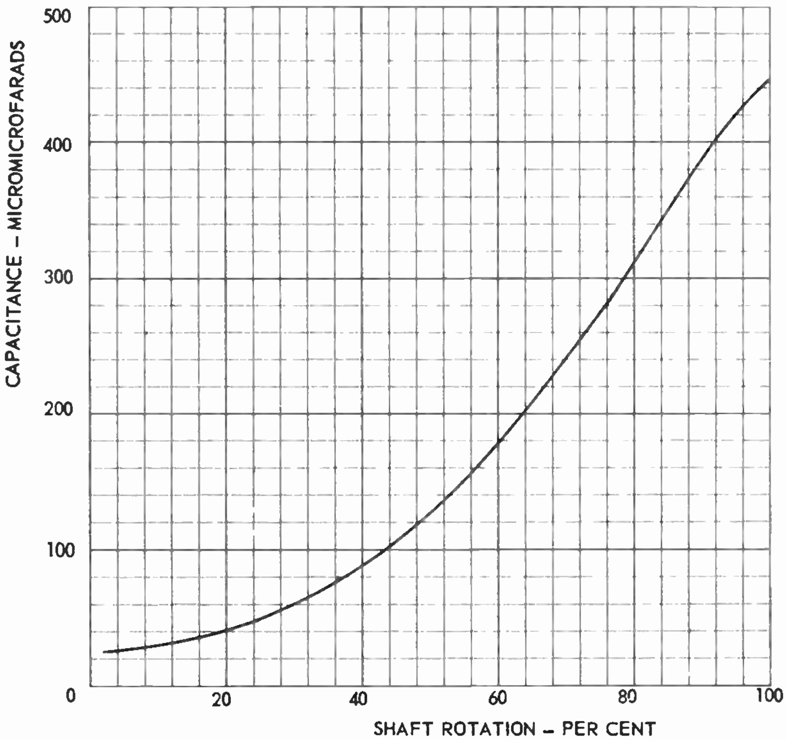


Fig. 26-4. How capacitance of the tuning capacitor varies with turning the rotor shaft.

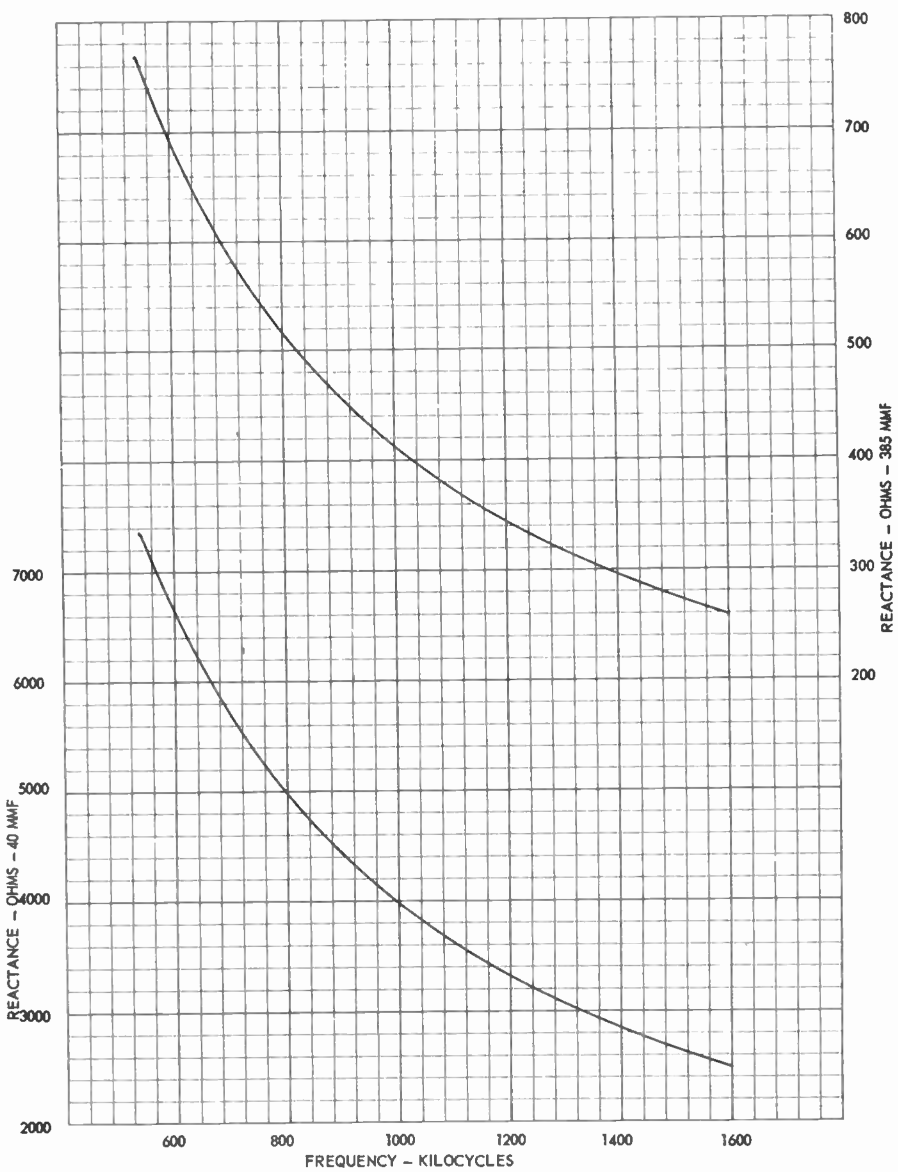


Fig. 26-5. Change of capacitive reactance with variation of frequency for capacitance of 40 micro-microfarads (lower curve) and for capacitance of 385 micro-microfarads (upper curve).

Fig. 26-4 shows how capacitance of the tuning capacitor varies during rotation of its shaft as the rotor plates are moved farther and farther in between the stators. The bottom horizontal scale is for rotation of the shaft in percentages of its full travel. The vertical scale lists corresponding capacitances in micro-microfarads.

With the capacitor adjusted for any capacitance its capacitive reactance becomes less and less as frequency increases. This reactance does not change at a constant rate, but rather as shown by Fig. 26-5.

The lower curve shows how the capacitive reactance changes when the capacitance remains set at 40 mmfd (micro-microfarads) while frequency is varied from 540 to 1,600 kilocycles. The scale for frequencies is along the bottom of this graph. The left-hand vertical scale lists corresponding reactances in ohms.

With the capacitor plates almost all the way in mesh the capacitance becomes 385 mmfd. Reactances of this capacitance at all frequencies between 540 and 1,600 kilocycles are shown by the upper curve of Fig. 26-5. The frequency scale is the same as for the lower curve. The reactance scale for this greater capacitance is written vertically on the right, up toward the top of the graph.

Although the values of reactance in ohms are greatly different for the two very different capacitances, the shapes of the two curves are alike. This shape is typical of curves showing change of capacitive reactance with frequency, no matter what the value of capacitance or the range of frequencies. For any other capacitance between 40 and 385 mmfd we might draw reactance curves, and all would have the same general shape.

Supposing that we wish to receive a signal in the 1,000-kilocycle broadcast channel. At this frequency the inductive reactance of the tuning coil is shown by Fig. 26-3 to be about 1,450 ohms. The capacitive reactance of the capacitor may be made anything between about 410 and 4,000 ohms, depending on how the rotor plates are turned. When you consider that the strength of received signal may be only a few millionths of one volt, either the inductive or capacitive reactance would offer tremendous opposition to signal currents from such a small voltage. The difficulties would be as great for reception at any other channel frequency. It is the reactive oppositions of the tuning transformer coil and the tuning capacitor that hold back the signal currents.

Were signal current to encounter only the resistance of metal parts in the capacitor and of the wire in the coil even these very weak signal voltages from an antenna would cause currents plenty large enough for reception. Resistance in the coil winding is only about 3½ ohms, and in the metal of the capacitor is

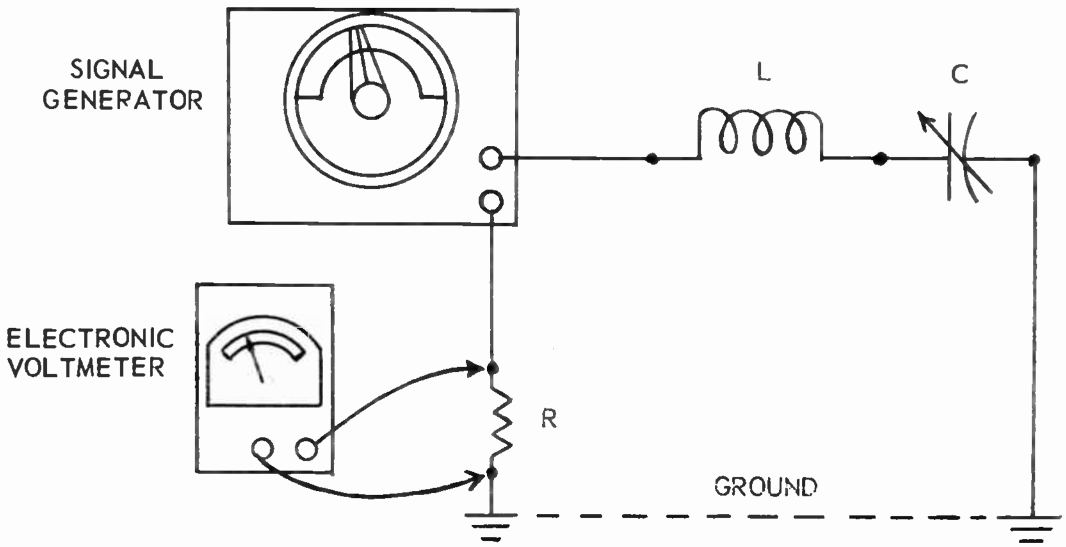


Fig. 26-6. How the instruments are connected for the first tests.

almost too small to be measured. Fortunately, it really is possible to get rid of the reactive oppositions and leave only the circuit resistances if we connect capacitance and inductance together in certain ways and adjust either of them to suitable values.

Strange things happen to high-frequency voltages and currents in circuits which contain both capacitance and inductance. In radio receivers, and television receivers too, everything related to tuning or selection of signals, and almost everything related to amplification of high-frequency signals, depends on the relations between inductance, capacitance, and frequency.

To observe what happens we need only such instruments as are available in nearly every radio service shop. All the tests whose results will be described in following pages were carried out with a service type signal generator furnishing frequencies between 540 and 1,600 kilocycles. These frequencies were applied to one section of the tuning capacitor of Fig. 26-1 and the bottom secondary winding of Fig. 26-2, connected together. Performance was measured with a service type electronic voltmeter.

SERIES RESONANCE. To begin with we shall connect the secondary winding of the radio-frequency transformer in series with the tuning capacitor, as shown by the diagram of Fig. 26-6. The secondary winding is marked *L*, for inductance, and the capacitor is marked *C* for capacitance. The primary winding of the transformer will not be used at present, so it is not connected and is not shown in the diagram.

The object of our first test will be to determine the manner in which changes of tuning capacitance affect opposition to flow of high-frequency current through the coil winding and capacitance in series. The way to do this is to send current from the signal generator through the winding and capacitance, vary the capacitance in small steps, and measure the current at each step. The greater the current the less must be the opposition to current flow, and the less the current the greater must be the opposition.

The diagram shows one terminal of the signal generator connected to one end of the coil. The other end of the coil is connected to one side of the tuning capacitor, and the other side of the capacitor is connected to ground. Ground, in this case, is the metal of the receiver chassis. The other terminal of the signal generator is connected through resistor *R* to ground, the chassis metal. Current from the generator will flow through the coil and capacitor, through the ground connections, and resistor *R*, back to the generator.

The electronic voltmeter is connected across the ends of resistor *R*, to indicate all changes of voltage occuring across this resistor. These changes of voltage will be exactly proportional to changes of current through the resistor. The voltmeter itself has resistance of millions of ohms, consequently takes no appreciable current away from the measured circuit. Then all current flowing in the coil and capacitor will flow also in the resistor, and since readings of the voltmeter will be proportional to current and changes of current through the resistor these readings also will be proportional to changes of current through the coil and capacitor. Thus the voltmeter becomes a current indicator.

The signal generator is adjusted to furnish a frequency of 1,000 kilocycles, and left so. Then the shaft of the tuning capacitor is rotated a little at a time to vary the capacitance in steps of only a few mmfd each time. At every new value of capacitance the voltage from the signal generator must be readjusted to its original value. It is necessary to apply unvarying signal voltage because we wish to observe only the effects of capacitance change. Were the applied voltage to vary it would affect the current, and our measurements would be meaningless.

The only purpose of using the resistor at *R* is to permit the electronic voltmeter to measure current and changes of current through the coil-capacitor circuit. No type of current meter ordinarily available in a service shop is suitable for measuring very small currents at high frequencies. Common types of alternat-

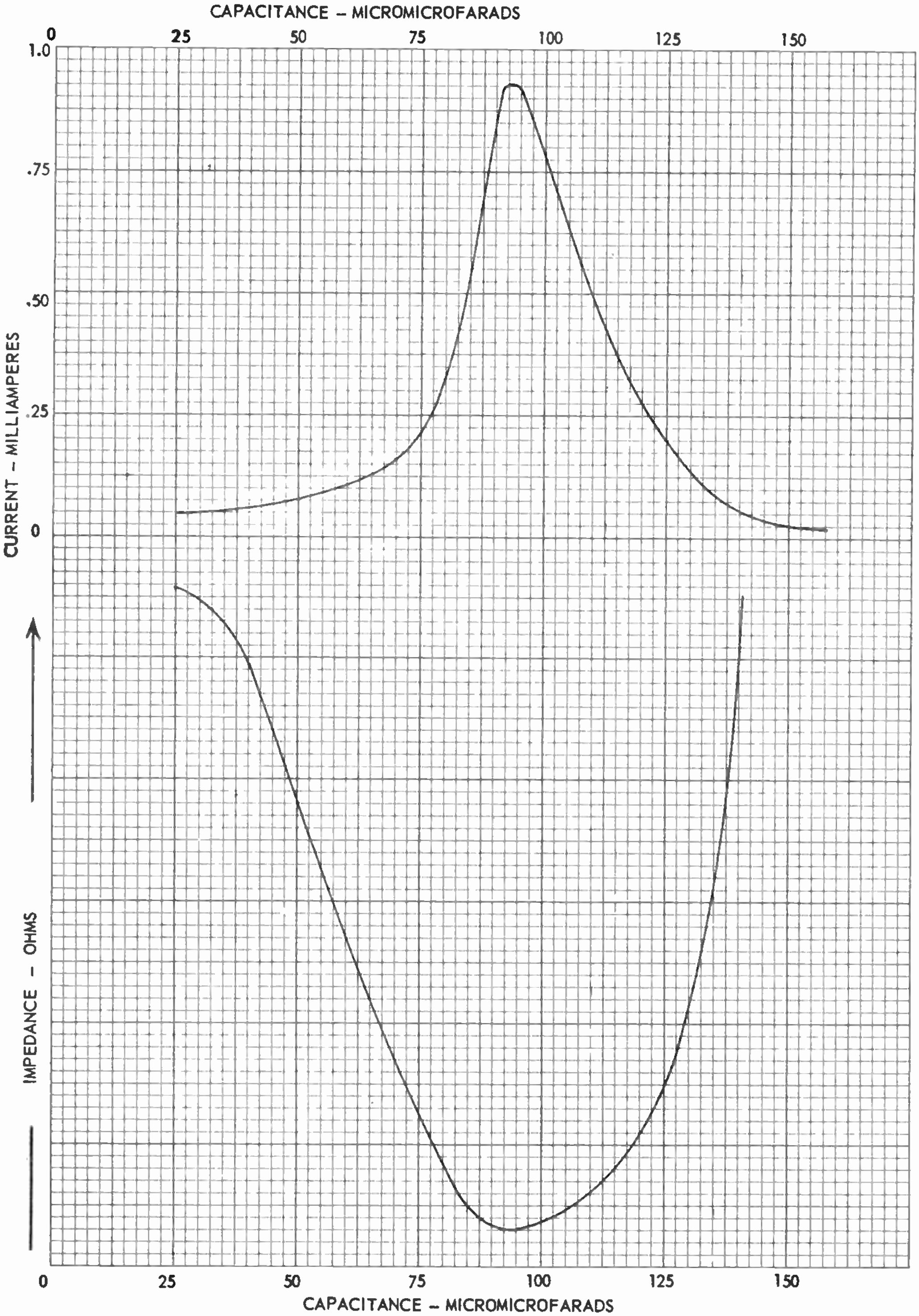


Fig. 26-7. Change of current and of impedance in the coil-capacitor circuit as the capacitance is varied.

ing-current meters either take more current for their operation than would be available in our tests or else their resistance is so great as to practically stop the current in a circuit such as we are using. The electronic voltmeter takes no measurable current for its own operation, and we need add in the measured circuit a resistance of no more than 50 to 100 ohms at most.

7 At the top of Fig. 26-7 is a curve showing how current in the coil-capacitor circuit changes as capacitance is varied. With increasing capacitance the current rises higher and higher until, with capacitance of about 92.7 mmfd, there is maximum flow. With further increase of capacitance the current drops to approximately its original value.

4 This current is flowing through both the coil and the capacitor, because they are connected in series with each other and with the signal generator. Every change of current that occurs in either the coil or the capacitor occurs also, and at the same time, in the other element, No matter how many strange things happen in this circuit during following tests, always remember that current is the same in the coil and in the capacitor so long as these two are in series with each other.

We have seen how variation of capacitance changes the current, but what we really started out to investigate is change of opposition to current flow, or change of inductive and capacitive reactances that oppose flow of current. In the coil there is opposition due to inductive reactance, but there is opposition also due to resistance of the winding wire. In the capacitor there is opposition due to capacitive reactance, also opposition due to resistance of the metallic paths which current must follow in the capacitor.

Of course, it is impossible to separate the oppositions due to resistance from those due to reactance, for we cannot do away with the conductors wherein the resistance exists. Consequently, we shall have to investigate the effect of reactance and resistance working together.

You will recall that opposition to flow of alternating current is called impedance when this opposition is caused by combined reactance and resistance. This makes our present problem one of translating the current changes shown at the top of Fig. 26-7 into corresponding changes of impedance. This is not difficult, because impedance, like resistance alone, is inversely proportional to current when applied voltage remains constant. That is the same as saying that impedance is the reciprocal of current, or is equal to the number 1 divided by current values. At the bottom of Fig. 26-7 is a curve showing changes of impedance with variation of capacitance. This curve was drawn by taking the reciprocals of the values shown by the current curve.

As the rotor plates of the capacitor are turned to insert more and more capacitance in the series circuit the impedance decreases and reaches a minimum at about 92.7 mmfd. With still more capacitance the impedance increases. At the minimum point the impedance is equal only to the circuit resistance. Effects of inductive reactance and capacitive reactance are gone. At this point of minimum impedance the reactance of the coil winding still is 1,451 ohms, and reactance of the capacitor also is 1,451 ohms. Yet these two reactances together amount to zero reactance.

For the time being we shall not concern ourselves with the reason why the two reactances cancel, but instead shall proceed with another test. This time we shall use the same instrument setup as before, but keep the tuning capacitor adjusted for 92.7 mmfd capacitance while changing the signal frequency in small steps from 540 to 1,600 kilocycles. The changes of current, as indicated by the electronic voltmeter across the series resistor, are shown at the top of Fig. 26-8.

This new current curve appears not unlike the one of Fig. 26-7. Current is small at the lower frequencies, rises to maximum at a frequency of 1,000 kilocycles, then drops back at still higher frequencies. Keep in

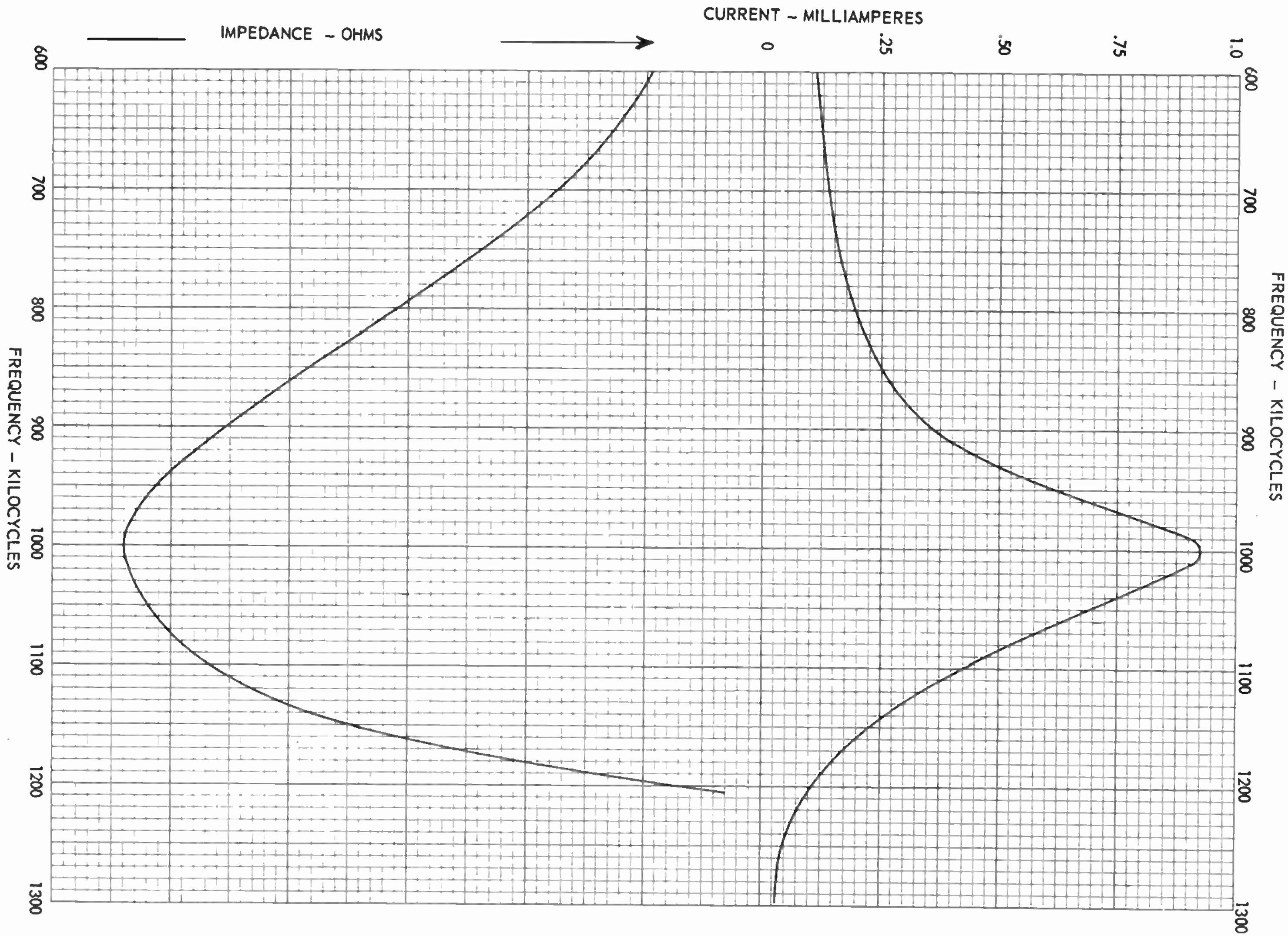


Fig. 26-8. Change of current and of impedance in the coil-capacitor circuit as the applied frequency is varied.

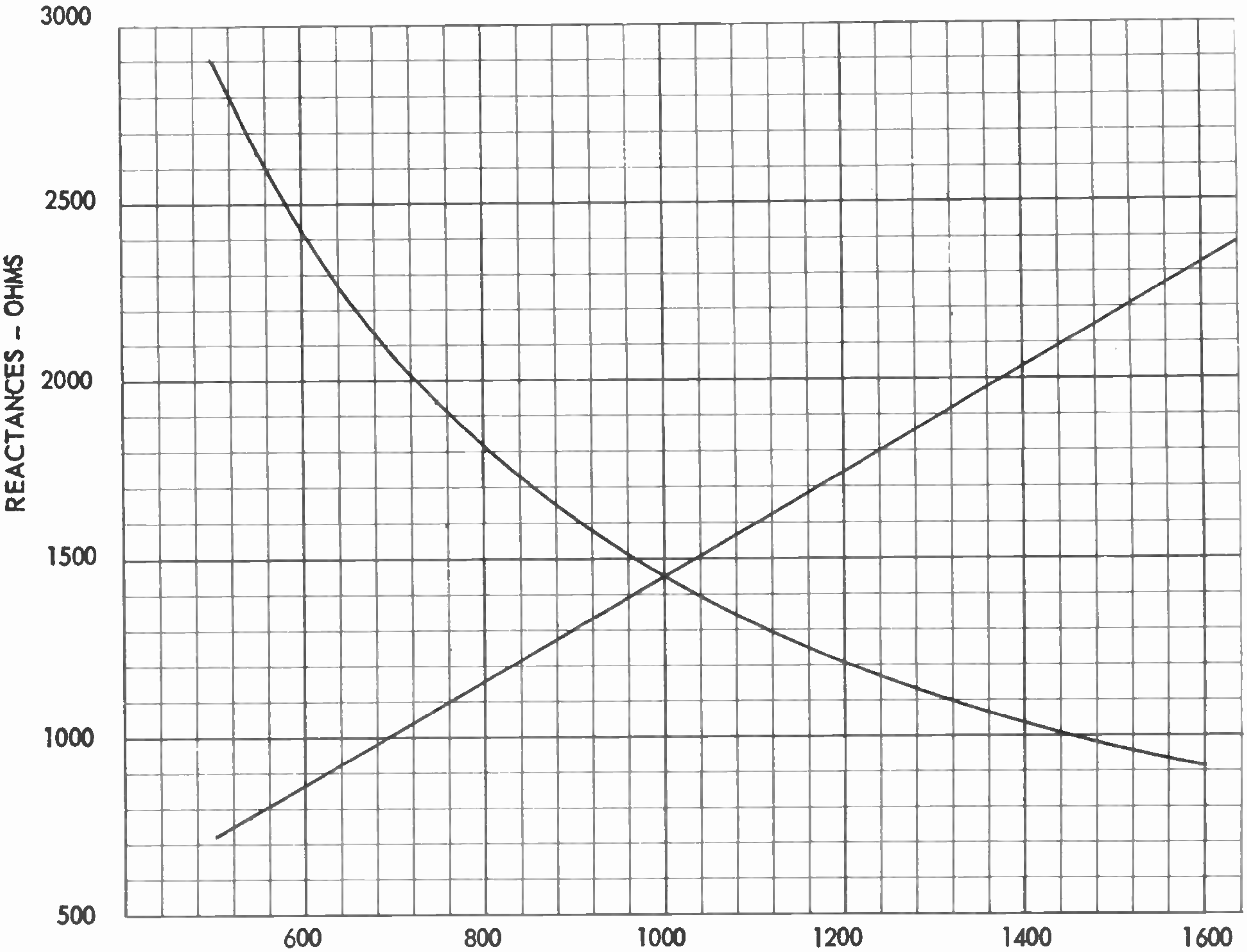


Fig. 26-9. The inductive reactance and the capacitive reactance become equal at a frequency of 1,000 kilocycles.

mind that the capacitance remains fixed at the value giving maximum current at 1,000 kilocycles in our first test.

Down below the current curve of Fig. 26-8 is another curve which shows changes of impedance as the frequency is increased from 600 to 1,200 kilocycles while capacitance remains fixed at 92.7 mmfd. This impedance curve again is the reciprocal of the current curve. Impedance is high at the lower frequencies, drops to a minimum at 1,000 kilocycles, then rises with further increase of frequency. At the frequency of 1,000 kilocycles there is high reactance in the coil and high reactance in the capacitor, yet with the inductance and capacitance connected in series there is no reactance effect, only the opposition due to resistance.

During our first test (Fig. 26-7), the capacitance was varied while keeping the frequency constant. Of course, the inductance also remains constant, because no changes were made in the coil winding. There

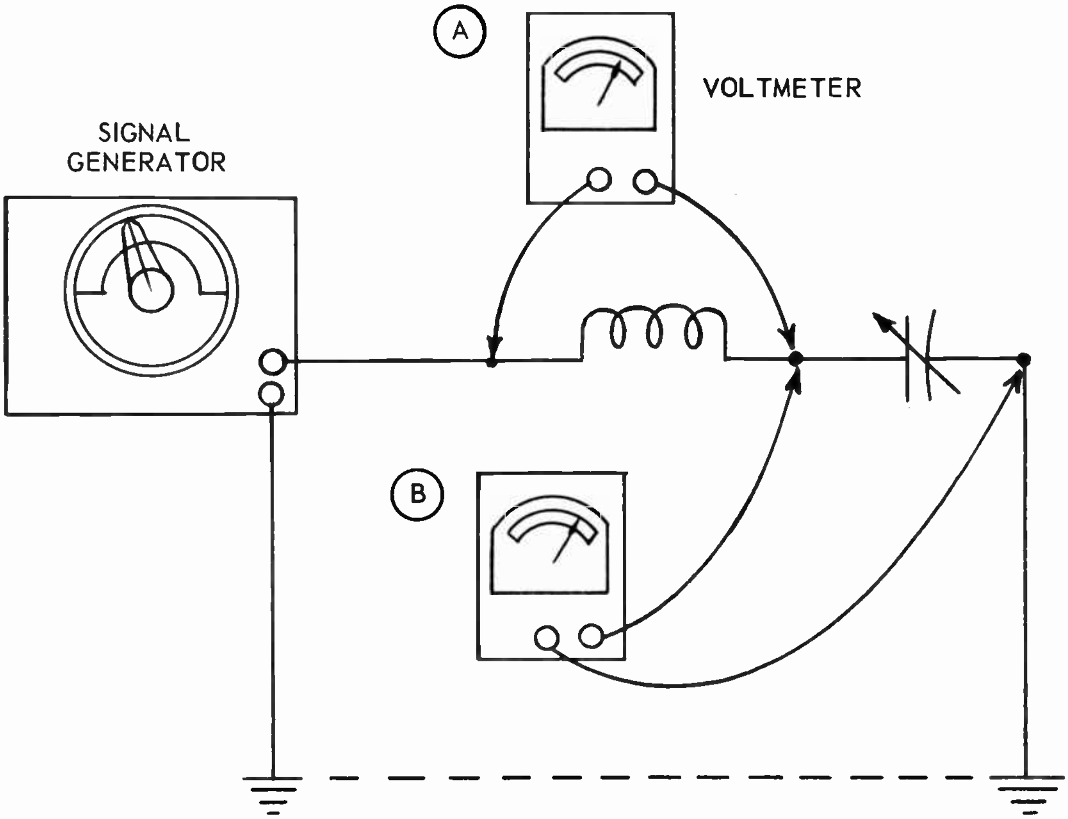


Fig. 26-10. Instrument connections for measurement of reactive voltages across the coil and the capacitor.

was one certain value of capacitance at which the effects of both the inductive and the capacitive reactances disappeared. In the second test (Fig. 26-8) we varied the frequency with a constant value of capacitance (and inductance) and found a certain frequency at which all reactance effects disappeared.

There can be only one explanation for these disappearing reactances. The two kinds of reactance, inductive and capacitive, must have opposite effects when they both are in the same circuit. At one value of tuning capacitance, and one value of frequency, the effect of inductive reactance and the effect of capacitive reactance must exactly balance. Each kind of reactance then is acting oppositely to the other kind, and their effects balance or cancel, to leave only the circuit resistance to oppose flow of high-frequency current.

We may obtain further evidence that the two kinds of reactance oppose and balance each other by making a comparison of how the reactances change with variation of frequency. In Fig. 26-9 the straight line shows the steady rise of inductive reactance in the coil as frequency is increased. The curve shows the drop of capacitive reactance in the capacitor which is in series with the coil. Note carefully the values of the two reactances at the frequency of 1,000 kilocycles. They are equal. If the reactances really do have opposite effects, and are equal to 1,000 kilocycles, it is only natural that they should balance out at this frequency.

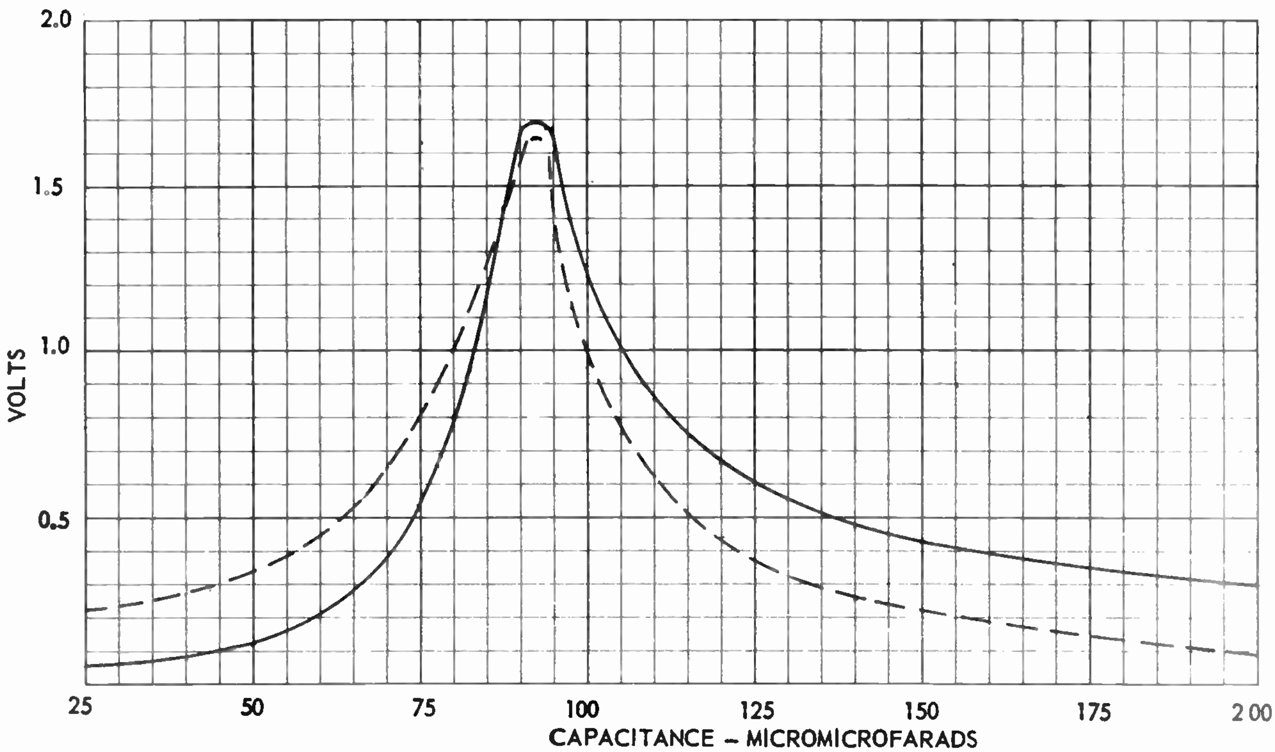


Fig. 26-11. Changes of coil voltage (full-line curve) and of capacitor voltage (broken-line curve) as the capacitance is varied.

5 When capacitive reactance and inductive reactance are equal at a certain frequency the circuit is said to be resonant at that frequency. When frequency is increased in a circuit containing inductance and capacitance, the inductive reactance always rises while the capacitive reactance drops. On the other hand, when frequency is decreased the inductive reactance always drops while capacitive reactance rises. Therefore, for any possible values of inductance and capacitance there must be some frequency at which the reactances become equal and at which the circuit is resonant.

This fact causes plenty of trouble in television receivers, where we have frequencies ranging all the way from 30 cycles per second to 200 million cycles. All kinds of small capacitances and inductances get together to produce resonance effects where such effects are least wanted.

All our examples and tests have been based on having the inductance and capacitance connected in series with each other. This makes a series resonant circuit, and the condition at resonance is called series resonance. Later we shall change the connections to place the inductance and capacitance in parallel with each other, and carry out other tests to determine what happens with parallel resonance. First, however, there are a few more things to be observed in relation to the series resonant circuit.

When running the curves of Figs. 26-7 and 26-8 the electronic voltmeter was used to indicate current, by connecting the meter across a resistor in series with the circuit. Now we shall connect the meter directly across the coil or inductance, and note changes of voltage on the coil. This new test setup is shown by Fig. 26-10, with the voltmeter at *A*.

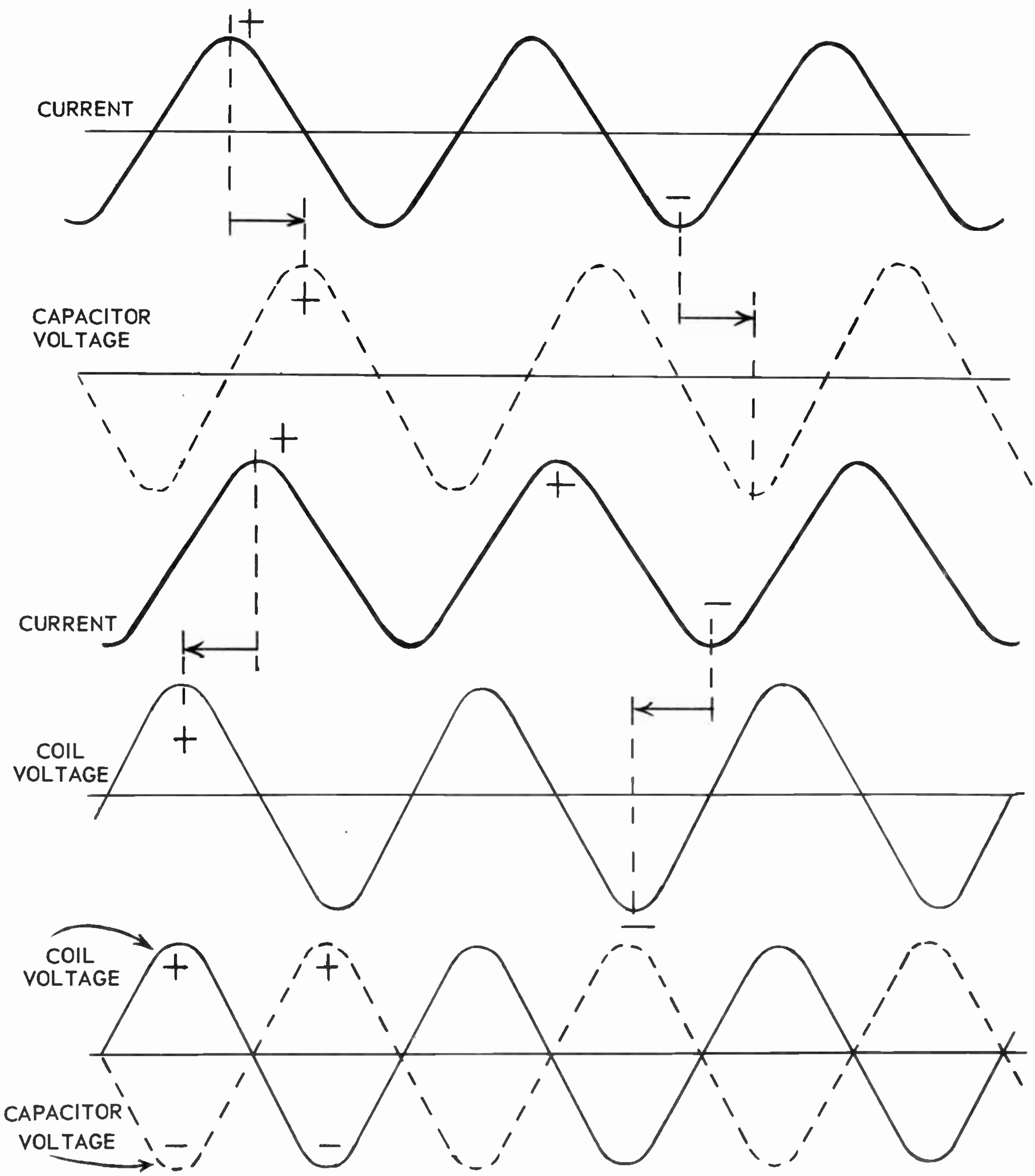


Fig. 26-12. Relations between current and the reactive voltages in capacitor and coil when the circuit is tuned to resonance.

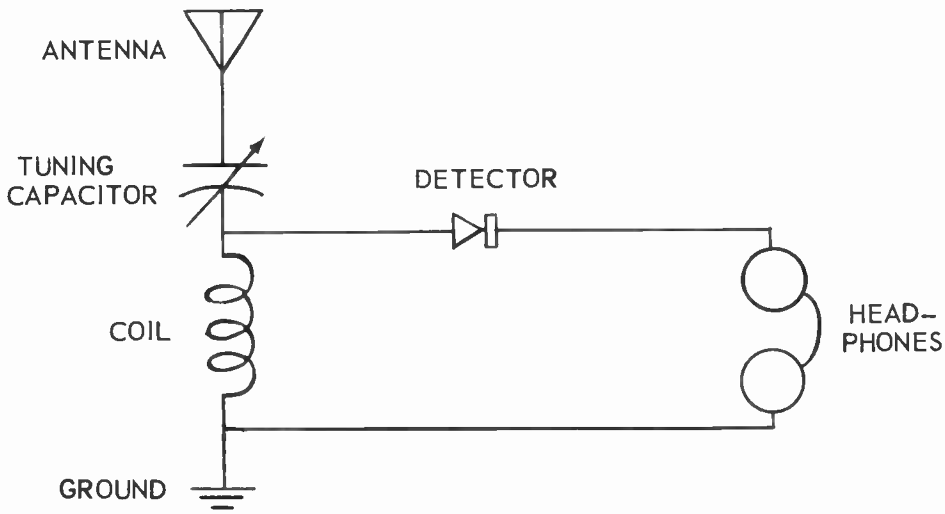


Fig. 26-13. Circuit connections for the crystal receiver.

The full-line curve of Fig. 26-11 shows how coil voltage varies when the tuning capacitance is changed from 25 to 200 mmfd while the applied frequency is held constant at 1,000 kilocycles. This voltage curve for the coil is shaped much like the current curve for the entire circuit, shown in Fig. 26-8.

For still another test we shall connect the electronic voltmeter across the tuning capacitor, as at *B* in Fig. 26-10, and note changes of voltage across the capacitor as its capacitance is varied from 25 to 200 mmfd with the applied frequency still held constant at 1,000 kilocycles. The broken-line curve of Fig. 26-11 shows resulting changes of capacitor voltage.

Voltage across the coil and voltage across the capacitor are maximum with capacitance of 92.7 mmfd. This we might expect, because there is the same current through the coil and capacitor, their two reactances become equal at this value of capacitance, and only the slight difference in resistances remains to affect the currents.

You will remember that we are maintaining a constant voltage from the signal generator throughout each of these tests. This high-frequency signal voltage across the series connected coil and capacitor was maintained at 0.2 volt during tests which yielded the curves of Fig. 26-11. Yet there we show 1.7 volts across the coil and almost as much across the capacitor. Voltage across both the coil and the capacitor is 8.5 times as great as voltage being applied to the entire series resonant circuit. Our circuit has a "voltage gain" of 8.5 when resonant at 1,000 kilocycles.

This means that were this signal voltage brought in from an antenna and applied across the series coil and capacitor the signal strength would be increased 8.5 times with the circuit tuned to resonance for that signal.

Now for a question about the performance which we have observed. If applied signal voltage is held constant across the whole resonant circuit how can the separate voltages across the coil and the capacitor both go so high at the same time? The explanation depends on this fact: The two voltages, at resonance, are away out of time with each other and with the current in the circuit. In the language of radio we would not say the voltages are out of time, we would say they are out of phase.

The electronic voltmeter cannot show differences of phase or time. One high-frequency voltage looks to the voltmeter just like any other high-frequency voltage if both are of the same strength. But if you could look at the manner in which the current is alternating, and at the same time look at alternations of voltage across the capacitor, the two alternating waves would appear as along the two top lines of Fig. 26-12.

The current and the voltage are both alternating at the same rate, or both have the same frequency. Consequently each alternation and cycle of current and of voltage is shown as occupying equal intervals of time – which proceeds from left to right. But every peak of capacitor voltage occurs a quarter of a cycle later than the corresponding peak of current. The alternations or cycles of capacitor voltage are lagging behind the alternations or cycles of current. This happens in any part of a circuit when that part contains capacitive reactance without inductive reactance, as is the case in the portion of our circuit which consists of the capacitor.

Could you look at the current wave and the coil voltage wave at the same time they would appear as at the middle of Fig. 26-12. Current waves are occuring at the same instants of time as up above, because this is the same current that flows in the capacitor. But now the coil voltage waves come earlier than the current waves. Every peak of coil voltage occurs a quarter cycle before the corresponding peak of current. The coil voltage is leading the current. This is true of every part of a circuit wherein there is inductive reactance without capacitive reactance.

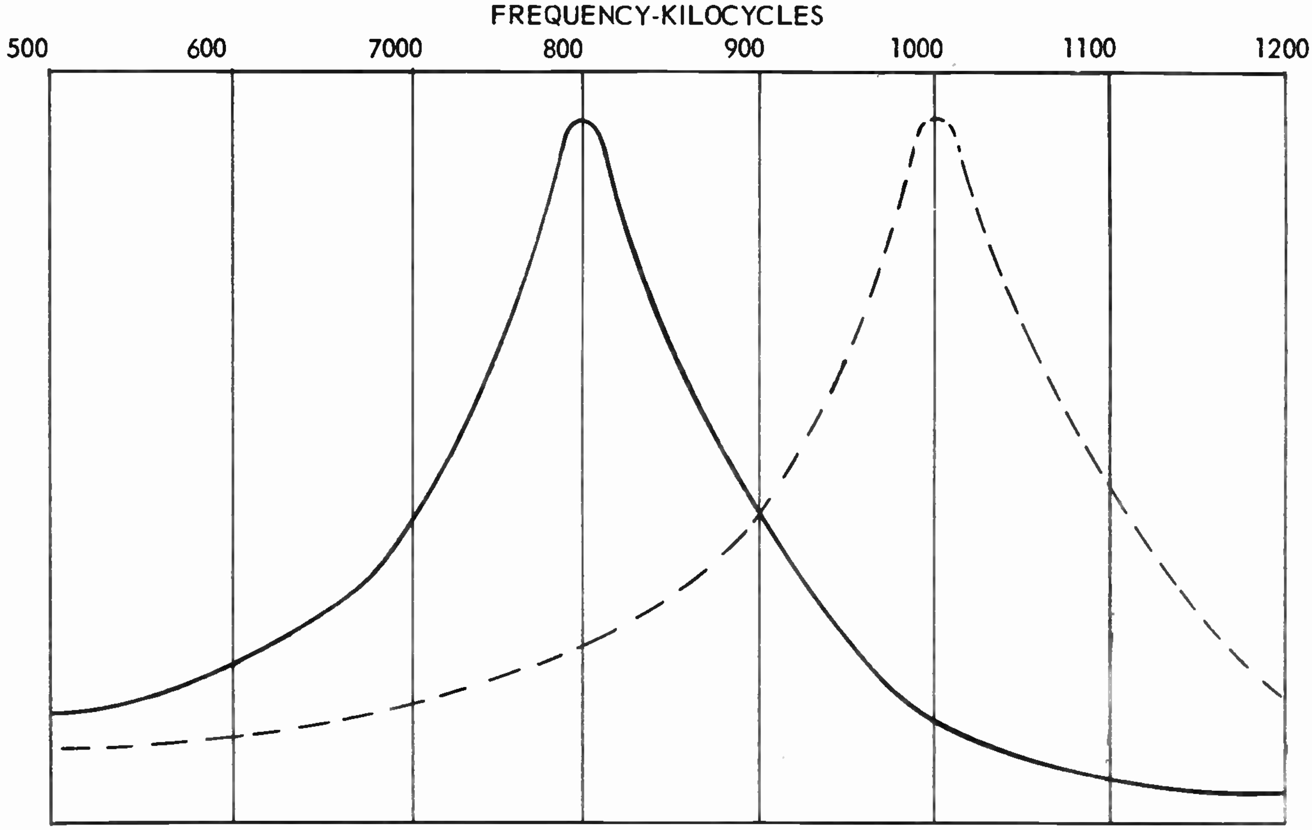


Fig. 26-14. Changing the tuning of a receiver shifts the resonance curve.

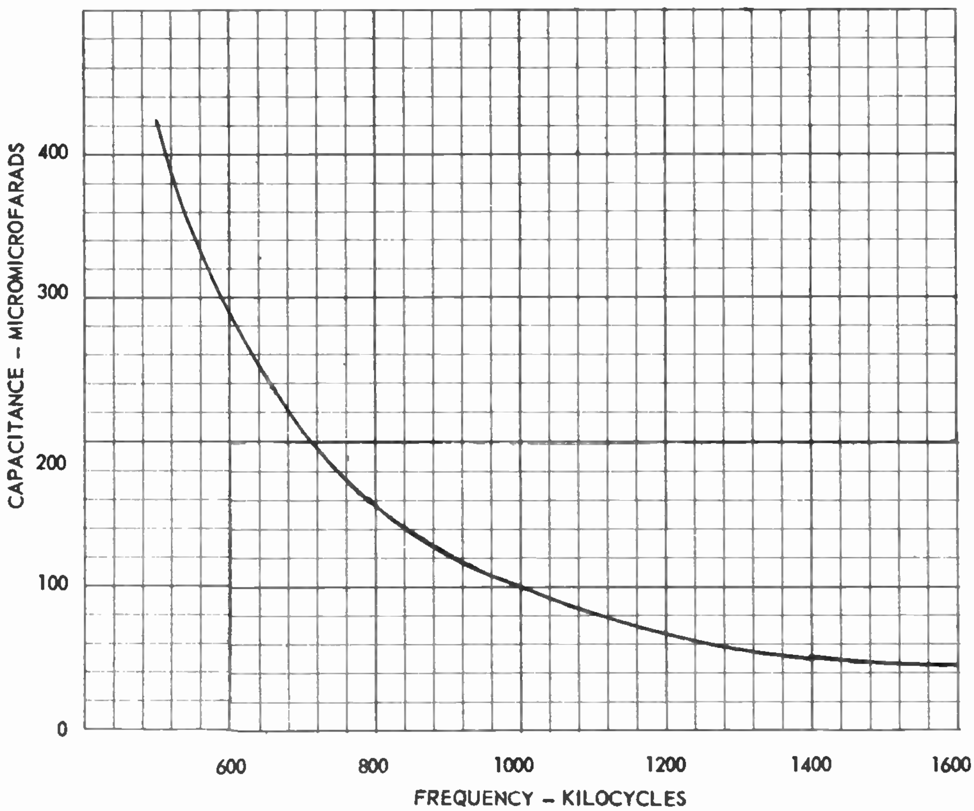


Fig. 26-15. Capacitance required in the tuning capacitor for resonance at frequencies in the standard broadcast band.

At the bottom of Fig. 26-12 the coil voltage and capacitor voltage are shown together, with alternations or waves of each voltage in their correct relative phase or time relation for a circuit in which capacitance and inductance are of values which produce resonance at the applied frequency. Every time the coil voltage becomes positive the capacitor voltage becomes equally negative. Every time the capacitor voltage becomes positive the coil voltage becomes equally negative. These reactive voltages completely cancel each other, at resonance, and their sum is zero.

All that is shown in Fig. 26-12 occurs while the signal generator is furnishing current to the coil and capacitor in series. When the generator looks into the circuit consisting of the coil and capacitor, and sees no voltage or no potential difference due to reactances, the effect, so far as the generator is concerned is as though the circuit possessed no reactance at all. The only thing that remains to oppose flow of current from the generator is resistance in the coil-capacitor circuit.

With reactance out of the way, the current at resonance rises to a value proportional to generator voltage and circuit resistance, just as shown by the curves at the top of Figs. 26-7 and 26-8. Impedance becomes minimum at resonance because all the reactive opposition is balanced out, and the only remaining opposition is that due to circuit resistance. This decrease of impedance at resonance is shown by curves at the bottom of Figs. 26-7 and 26-8.

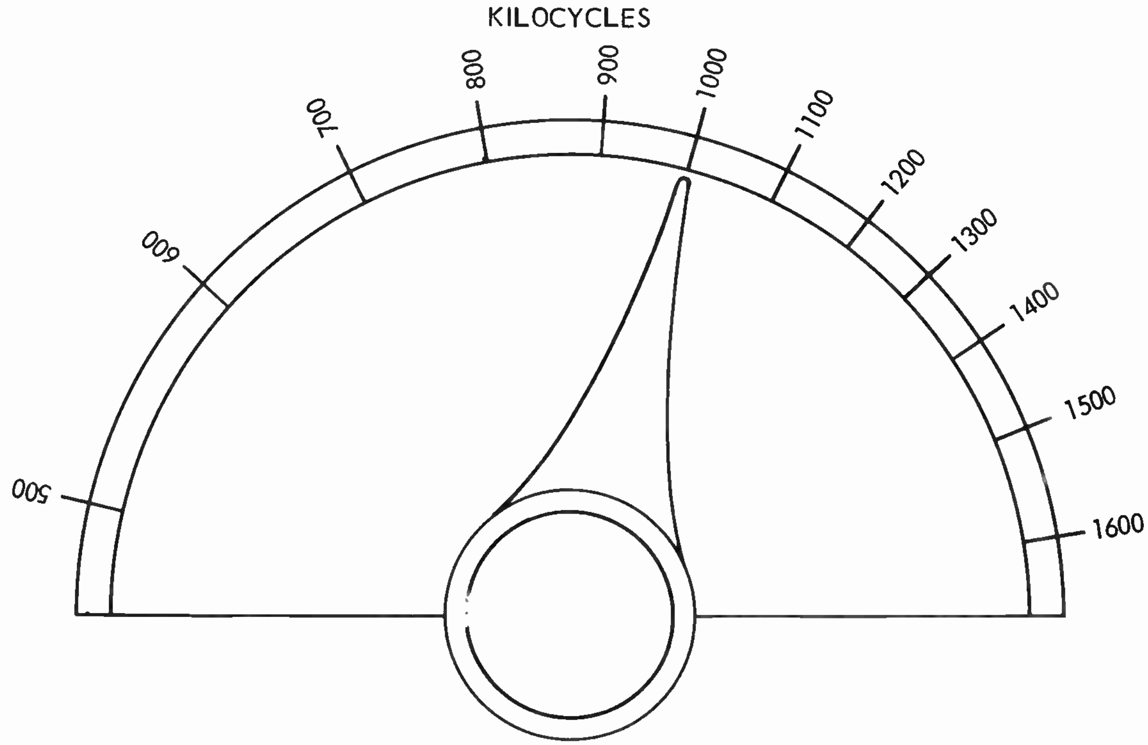


Fig. 26-16. Spacing of the broadcast carrier frequencies across the tuning dial.

TUNING WITH SERIES RESONANCE. By connecting in series the tuning capacitor and coil which we have been testing, and adding a crystal detector and headphones, it is possible to construct a practical radio receiver for the standard broadcast band. The circuit is shown by Fig. 26-13. The high side of the capacitor is connected to an antenna. The low side of the coil is connected to ground. All signal voltages collected by the antenna now are applied to the capacitor and coil,and cause corresponding high-frequency signal currents to flow in these elements.

With the detector and headphones connected across the coil all voltages which appear across the coil are applied to the detector and phones. The detector is merely a rectifier, allowing pulses of high-frequency signal current to flow through it and the headphones in only one direction. As a result of this detector action we have in the phones a one-way current or direct current consisting of pulses at the signal frequency. If the average value of these direct-current pulses varies at an audible rate or at an audio frequency you can hear sounds from the phones.

The direct current pulses do vary at audio frequency because the carrier wave at every transmitter is being modulated or varied in average strength by effects of speech and music which are to be transmitted. Therefore, if signal voltage across the coil and in the headphones becomes strong enough, you can hear the audio frequencies which constitute a radio program.

Now let's see what is happening when you do hear a program. We shall assume that you have adjusted the tuning capacitor for resonance at 1,000 kilocycles. There are many broadcasting stations using this carrier frequency. Any 1,000-kilocycle signal reaching your antenna meets with the very minimum of impedance in the coil-capacitor circuit, as shown by Fig. 26-8. Then this signal produces maximum current in the coil and capacitor. Across the coil there is maximum possible voltage at this 1,000-kilocycle frequency, and this voltage is applied to the headphones through the detector.

Received signals at all frequencies higher and lower than 1,000 kilocycles meet with much greater impedance in the receiver circuit, and these other signals cause relatively weak currents and voltage in the coil. Thus you have strengthened the signal from the transmitter operating with a carrier frequency of 1,000 kilocycles, and at the same time have greatly weakened or wholly excluded the signals from transmitters operating at other frequencies. This is the process of tuning.

Fig. 26-8, of which we have just spoken, shows the current produced in the receiver coil by signals of equal strengths at various frequencies when the coil-capacitor combination is tuned to resonance at 1,000 kilocycles. We say signals of equal strengths because, remember, the output voltage of our signal generator was held constant during every test.

An equally strong signal at 900 kilocycles would produce only about 40 per cent as much current and voltage as the one at 1,000 kilocycles, and a signal at 800 kilocycles would produce only about 20 per cent as much. Going the other direction in frequency, a signal of equal strength at 1,100 kilocycles would cause less than half the current of the one at 1,000 kilocycles, and for a signal at 1,200 kilocycles the response would be little more than 10 per cent of that for the 1,000-kilocycle signal.

Were you to tune your receiver for 800 kilocycles, or make the coil-capacitor circuit resonant at 800 kilocycles, the whole response curve would shift to this new frequency, as shown by Fig. 26-14. Peak response then would be at 800 kilocycles instead of the former 1,000 kilocycles. Loudest sound would come from a station operating on 800 kilocycles, provided its signal were as strong as from the 1,000-kilocycle station, and response would be down at all other frequencies.

The receiving circuit could be made resonant at any frequency within the tuning range of the capacitor, which is the adjustable member of our tuning combination. Fig. 26-15 shows capacitance settings for the tuning capacitor at all signal frequencies between 500 and 1,600 kilocycles. With the capacitor adjusted for any of the capacitances listed on the scale at the left there would be resonance, with the coil inductance of 231 microhenrys, at the frequency shown by the curve.

At the upper end of the curve a large change of capacitance makes a rather small change of tuned frequency, while at the low end a small change of capacitance makes a relatively great change of frequency. If you look back at Fig. 26-4 you will see that, at the high-capacity end of tuning shaft rotation, a fairly small turn makes a large change of capacitance, while at the low capacitance end it takes a very considerable turn to change the capacitance only a little.

The spreading of capacitance change at one end of tuning shaft rotation compensates in large measure for crowding of frequencies at one end of the capacitance scale. By comparing the shaft rotation curve of Fig. 26-4 with the capacitance change curve of Fig. 26-15 we find that station frequencies will be spread quite evenly over the face of a tuning dial, as shown by Fig. 26-16. Tuning capacitor plates are shaped to separate the frequency positions for satisfactory tuning.

TUNING WITH VARIABLE INDUCTANCE. We have been making tests on a circuit with fixed inductance and variable capacitance. Every test could have been carried out just as well by altering the inductance while using a fixed capacitance. Some broadcast sound receivers, a good many f-m receivers, and lots of television receivers are tuned by variable inductors rather than variable capacitors.

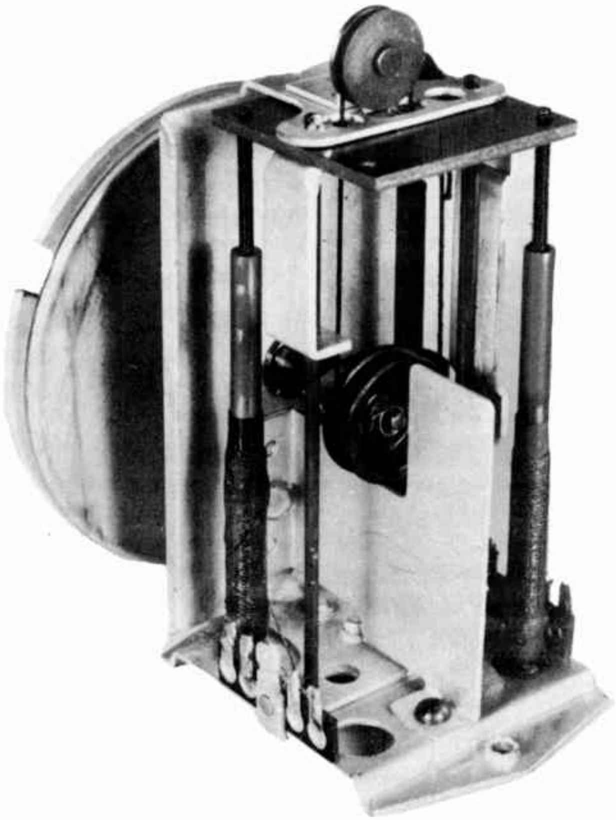


Fig. 26-17. A variable inductance tuner for a radio receiver.

The frequency at which resonance occurs may be changed just as effectively by altering the inductance as by altering the capacitance in tuned circuits. Increasing the inductance, with a fixed capacitance, will lower the resonant frequency in just the same way as increasing the capacitance when using fixed inductance. Conversely, lessening the inductance will raise the resonant frequency in just the same way as lessening the capacitance with fixed inductance.

Inductance and capacitance act alike so far as tuning is concerned. You may use either more inductance or more capacitance to lower the frequency of resonance, or you may use more of both. To raise the resonant frequency you may use less inductance or less capacitance, or you may use less of both.

In our tests with a variable capacitor for tuning we employed a constant inductance of 231 microhenrys. With the tuning capacitor set for about 42 mmfd the circuit was resonant at 1,600 kilocycles, and for resonance at 500 kilocycles we set the capacitor for about 438 mmfd. We might have carried out all the tests with a variable inductor and fixed capacitor. For example, had we employed a fixed capacitor of 100 mmfd it would have been possible to have resonance at 1,600 kilocycles by adjusting the variable inductor to about 97 microhenrys. With the same fixed capacitance there would be resonance at 500 kilocycles by adjusting the inductor to about 1,012 microhenrys.

Fig. 26-17 is a picture of one style of variable inductance unit used for tuning in standard broadcast receivers. The inductance coils stand vertically on the base. Each coil is a radio-frequency transformer consisting of primary and secondary windings suited for one range of frequencies. Sliding inside each transformer coil is a movable core made of finely divided particles of iron cemented into cylindrical shape. These cores are moved up and down by the pulley and belt at the top of the unit and by the large pulley at the back. This large pulley is turned by another belt running to the tuning knob.

The farther the iron cores are moved down inside the transformer windings the greater becomes the inductance, and the farther the cores are raised out of the windings the less is the inductance. When the core of the antenna coupling transformer is moved all the way from top to bottom of its travel the secondary inductance increases from about 86 to 1,000 microhenrys in this particular unit. With a fixed capacitance of 100 mmfd this change of inductance would tune to resonance from a little above 500 kilocycles to about 1,740 kilocycles, completely covering the standard broadcast band with something to spare at each end.

Frequency of resonance depends on the product of inductance and capacitance in the tuned circuit, or on the number of microhenry inductance multiplied by the number of mmfd capacitance. The inductance-capacitance product for resonance at a frequency of 1,000 kilocycles is 25,330. In any possible combination of inductance and capacitance which is resonant at 1,000 kilocycles the product of microhenrys and micromicrofarads must come to 25,330.

10 For resonance at 1,000 kilocycles you could use 100 mmfd capacitance and 253.3 microhenrys inductance, or you could use 100 microhenrys inductance and 253.3 mmfd capacitance, or any other combination whose product is 25,330. For every other frequency of resonance there is an inductance-capacitance product. These products are called oscillation constants.

DISTRIBUTED CAPACITANCE. If you look back at the results of some of the earlier tests with the variable tuning capacitor and fixed inductance you will find that 92.7 mmfd in the capacitor and 231 microhenrys in the coil caused resonance at 1,000 kilocycles. The product of 92.7 and 231 is 21,414 although it just has been stated that the oscillation constant for 1,000 kilocycles is 25,330.

The explanation of this discrepancy is simple, but of great importance in all tuned circuits. There is capacitance in the coil itself. It is called distributed capacitance. The coil used in our tests has distributed capacitance of 17 mmfd along with its 231 microhenrys inductance.

This 17 mmfd of distributed capacitance added itself to the capacitance of the variable capacitor in every test we made. If you add this 17 mmfd to the 92.7 mmfd in the variable capacitor at resonance for 1,000 kilocycles the sum if 109.7 mmfd. If you then multiply 109.7 (mmfd) by 231 (microhenrys) the product is 25,340 approximately. This is within one twenty-fifth of one per cent of the oscillation constant for 1,000 kilocycles, which is as close as you can expect to come in computations such as these.

All through this present lesson we have been careful to talk only about capacitance of the tuning capacitor, not about total circuit capacitance. Every value of capacitance which has been specified with reference to the capacitor must be increased by 17 mmfd of distributed capacitance to arrive at circuit capacitance. Then the oscillation constant will be found to apply in every case.

Now we have piled up a good many things to be investigated at greater length in lessons to follow. We shall have more to do with variable inductance tuning. There is a lot to learn about parallel resonance. Distributed capacitance calls for attention. Oscillation constants can be of help in service operations. In fact, we have just begun the study of resonant circuits.

TELEVISION

LESSON NO. 27

PARALLEL RESONANCE FOR TUNING

Training Your Mind...
C. T. I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind
C. T. I.
Perfecting Your Skill

COMMERCIAL
TRADES
INSTITUTE

Chicago, Ill.

LESSON NO. 27

PARALLEL RESONANCE FOR TUNING

In Fig. 27-1 we have an inductance coil and a small variable tuning capacitor connected in series. This coil-capacitor circuit may be considered as starting at the free end of the wire attached to the back of the coil, going through the coil, thence to the stator of the capacitor, through the capacitor, and from the rotor through the forward extending wire attached to the capacitor frame. The connections are shown by symbols at the right. This you will recognize as a series resonant circuit.

In Fig. 27-2 the same coil and capacitor are connected in parallel with each other. The stator of the capacitor remains connected through a light colored wire to the near end of the coil. The rotor of the capacitor now is connected to the far end of the coil, through a second light colored wire. The two black wires lead to any external circuit. Again the connections are shown by symbols at the right. This is a parallel resonant circuit.

Any inductance and capacitance may be connected for either series resonance or for parallel resonance. The tuning capacitor and r-f transformer winding which were connected in series for our tests in the preceding lesson later were reconnected in parallel with each other, to form a parallel resonant circuit. The signal generator was adjusted to furnish a frequency of 1,000 kilocycles. The electronic voltmeter was used to measure high-frequency voltage across the coil and capacitor. With the tuning capacitor varied between 25 and 180 mmfd capacitance the voltage across this parallel resonant circuit changed as shown by Fig. 27-3.

The condition of parallel resonance is indicated by peak voltage. This peak occurs when the tuning capacitance is 92.7 mmfd, exactly the same value as caused resonance at 1,000 kilocycles with the coil and capacitor connected for series resonance.

For another test the tuning capacitor was left adjusted for capacitance of 92.7 mmfd and the frequency from the signal generator was varied from 650 to 1,350 kilocycles. The electronic voltmeter across the coil and capacitor in parallel showed the voltage changes of Fig. 27-4. There is peak voltage, indicating resonance, at 1,000 kilocycles — just as with the same coil and capacitor connected for series resonance.

Further tests with any values of inductance and capacitance would show that resonant frequency is the same with the elements connected in parallel as with them connected in series. Parallel resonant and series resonant circuits behave similarly so far as relations between values of inductance, capacitance, and frequency are concerned. This, however, is about the only way in which the two types of resonant circuits do behave similarly.

You will find that wherever a certain change of frequency causes current to rise in a series resonant circuit the current will drop with parallel resonance. Whatever causes less voltage across a series resonant

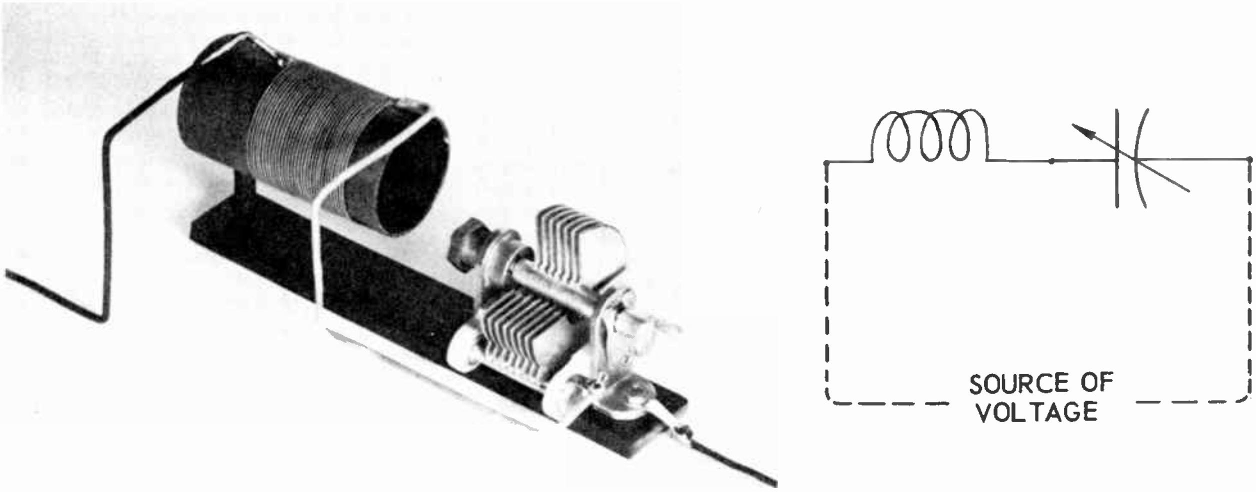


Fig. 27-1. A series resonant cirouit.

circuit will cause more voltage across the parallel resonant circuit. For the frequency at which impedance is minimum in the series circuit it becomes maximum with the parallel connection.

We could continue to investigate the performance of parallel resonant circuits by using our original tuning capacitor and transformer winding, whose values were suited for standard broadcast frequencies. Instead we shall use capacitance and inductance of values such as employed in television and f-m sound receivers. Instead of working with standard broadcast frequencies of a few hundred thousand cycles we shall use frequencies of tens of millions of cycles per second.

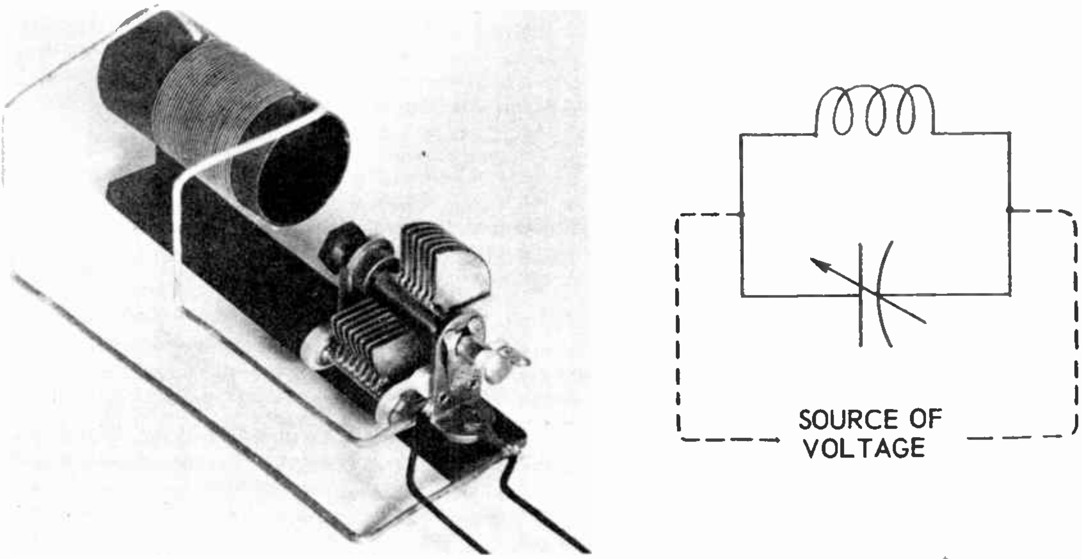


Fig. 27-2. A parallel resonant circuit.

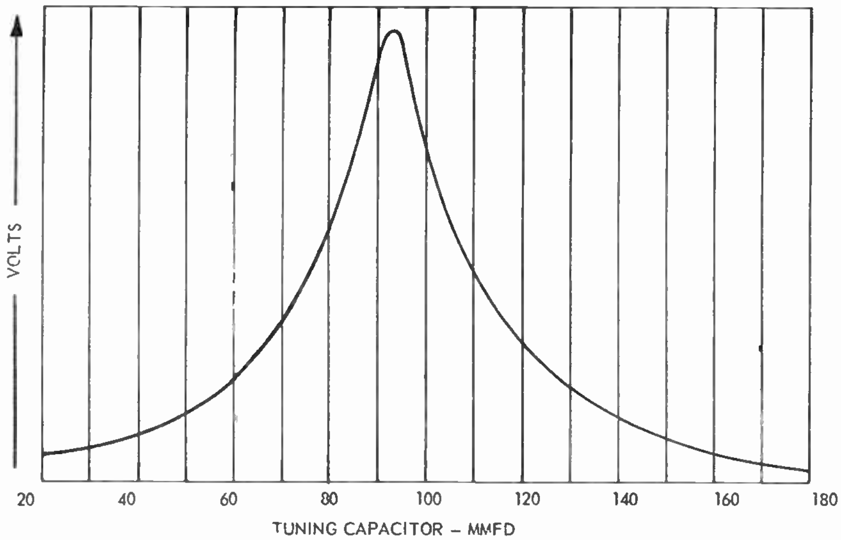


Fig. 27-3. The change of voltage across the coil and capacitor of a parallel resonant circuit when the capacitance is varied.

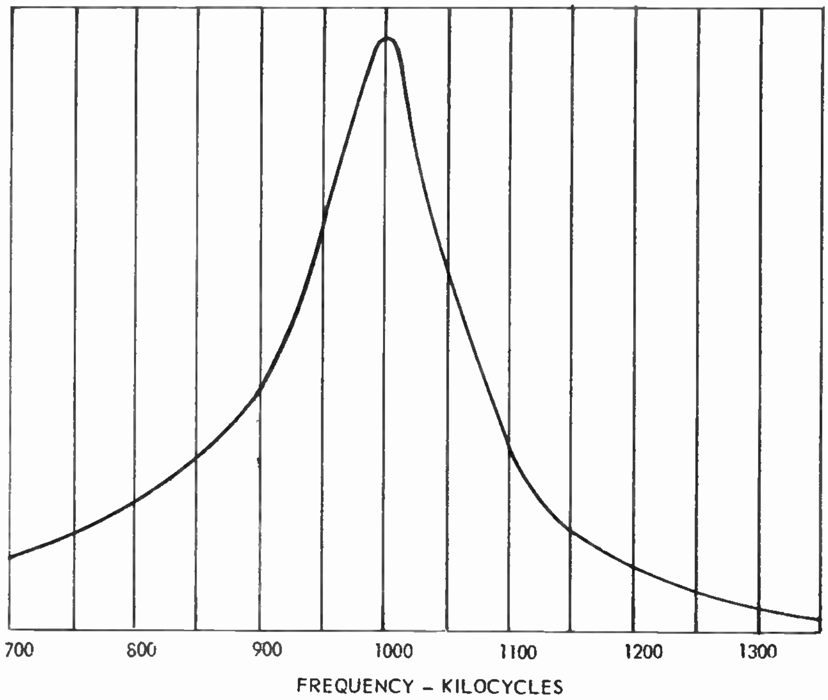


Fig. 27-4. Change of voltage across the parallel resonant circuit when the applied frequency is varied.

A

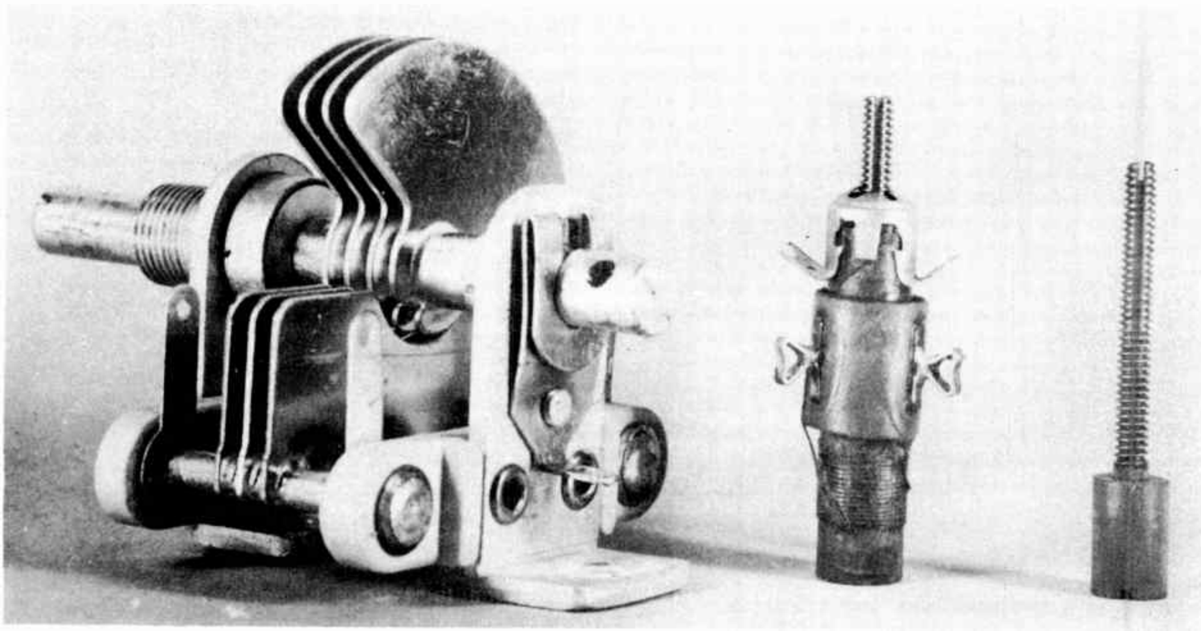


Fig. 27-5. The variable capacitor and the coil whose inductance may be varied by a movable core.

The new experimental coil and capacitor are illustrated in Fig. 27-5. The coil winding consists of 17 turns extending over a length of about 0.30 inch with a diameter of 0.28 inch. This coil is designed for variable inductance tuning by means of a movable core made of powdered iron cemented into shape. A core is shown standing alongside the coil form in the picture.

When the core is inserted into the coil form the long screw on the core extends through the metal ferrule at one end of the form. A screw driver slot on the end of the screw allows turning the core to various positions in relation to the winding. This unit is a coupling coil used between intermediate-frequency amplifier tubes in a television receiver.

With the movable core adjusted to a position where it extends all the way through the winding which is on the outside of the form, the inductance is maximum at about 6 microhenrys. With the core adjusted so that no part extends into the winding the inductance drops to around 2 microhenrys.

The variable tuning capacitor of Fig. 27-5 really is not a type ordinarily used in television receivers, but it will provide a range of capacitances matching anything found in actual circuits, and will allow making tests. With the rotor plates of the capacitor turned all the way into mesh with the stator plates the capacitance is about 47 mmfd, and with the rotor all the way out the remaining capacitance is 6.5 mmfd.

2 Possibly you wonder why we don't use a variable tuning capacitor such as found in television sets for investigating the performance of television circuits. The answer is, in many of the tuned circuits in television we don't use any tuning capacitor at all, yet we tune with capacitance and inductance. Before proceeding with tests on parallel resonance in television circuits it will be well to determine where the tuning capacitance does come from.

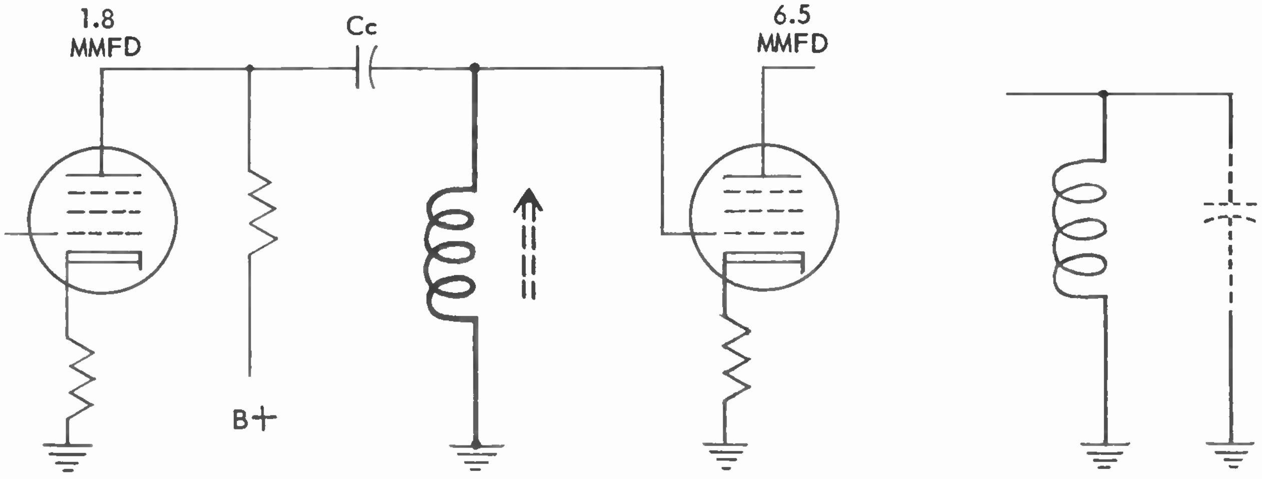


Fig. 27-6. The inductance coil used for coupling between two video i-f amplifier tubes.

DISTRIBUTED AND STRAY CAPACITANCES. When we get into the high frequencies used for television and f-m radio we tune with inductances and capacitances which are only small fractions of those used at standard broadcast frequencies. For television intermediate frequencies we may use three or four microhenrys inductance and maybe ten to fifteen mmfd capacitance.

This makes it necessary to consider carefully the distributed capacitances in coils, capacitances in tubes, and various "stray capacitances" in other parts. Although these capacitances are small in themselves, only a few mmfd, they are large in comparison with the capacitances required for tuning to resonance. An example will illustrate what we may expect in television circuits and also in f-m sound receivers.

Let's say we have a television set employing a video intermediate frequency of 25 megacycles, and wish to use in one of the resonant circuits a coil having inductance of 4 microhenrys. To have resonance at 25 mc (megacycles) with this much inductance requires about 10.1 mmfd total capacitance.

Distributed capacitance in our test coil and its terminal connections on the supporting form will be on the order of 1.6 mmfd. This, seemingly, will leave the difference between 10.1 and 1.6 mmfd, or 8.5 mmfd, required for adjustable tuning. But before connecting a variable tuning capacitor into the coil circuit we must look at several other things.

Quite probably our coil will be connected between amplifier tubes as shown by Fig. 27-6. We shall assume that both these tubes are 6AG5's, a type quite generally used in video i-f amplifiers. The top of the coil is directly connected to the grid of the second tube, and the bottom of the coil is connected to ground (chassis metal).

The cathode of that second tube is connected to ground through a biasing resistor. The grid of the tube is metal, its cathode is metal, and between them is a vacuum. As a result, the grid and cathode act like the two plates of a capacitor, with the vacuum for dielectric, and within the tube we have capacitance between grid and cathode. Actual capacitance on the input or grid side of a 6AG5 tube will average 6.5 mmfd. This input capacitance, shown by a symbol in the small diagram at the right, is in parallel with the coil and with distributed capacitance in the coil. The tube and coil capacitances combine with the coil inductance as part of a parallel resonant circuit.

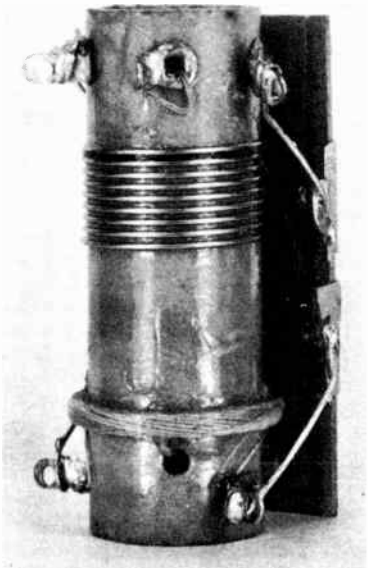


Fig. 27-7. Turns in the upper winding are spaced apart to lessen distributed capacitance.

We started out by needing 8.5 mmfd, in addition to distributed capacitance, for tuning to resonance at 25 mc. The input capacitance of the second amplifier tube turns out to be 6.5 mmfd, and is connected across the coil. So now we need only an additional 2.0 mmfd capacitance for tuning.

The top of our tuning coil is connected through coupling capacitor *Cc* to the plate of the first amplifier tube. In this tube there are internal capacitances between the plate, the cathode, and other metallic elements. Their total, forming the output capacitance of this 6AG5 tube, is 1.8 mmfd. This capacitance also is across the coil. Coupling capacitor *Cc*, in series between plate and coil, causes reduction of about 0.01 mmfd in output capacitance across the coil, but this is such a slight change that we may figure on the entire output capacitance.

A few moments ago we needed an additional 2.0 mmfd for tuning to resonance at 25 mc. Now we have found 1.8 mmfd in the first amplifier tube. This leaves only an extra 0.2 mmfd needed for tuning the coil to resonance at 25 mc when coil inductance is 4 microhenrys.

We need not look far for this small additional capacitance, because there is lots of it in the wiring connections. If you take one inch of ordinary hookup wire and place it tightly against chassis metal there is capacitance of about 5 mmfd between that one inch of wire and the chassis. Should you raise the inch of wire a quarter inch off the chassis metal there still would be capacitance of at least 0.3 mmfd between them.

A look at the wiring in even the best of television receivers makes it apparent that in the connections for any coupling coil there must be much capacitance between the wires themselves and between wires and chassis metal. The fact is, outside a laboratory setup, you couldn't make the necessary connections to the coil and introduce only 0.2 mmfd of stray capacitance.

So far we haven't used any tuning capacitor, yet have found distributed capacitance, tube capacitance, and stray capacitance totaling probably 14 mmfd as a minimum. With our assumed 4 microhenrys in the coil this much capacitance without any tuning capacitor at all would be resonant at 21.28 mc instead of the desired 25 mc. To raise the resonant frequency to 25 mc it would be necessary to reduce the capacitance. The only capacitance that can be reduced is that in the wiring, and getting rid of all of it wouldn't be enough.

The solution of our difficulty is to reduce the inductance. Here we have one reason why so many high-frequency circuits are tuned with adjustable inductance. With so much distributed, stray, and tube capacitance which cannot be varied, we have to have adjustable inductance. By screwing the core farther out of the coil winding we can drop the inductance to 2.9 microhenrys. Then there will be resonance at 25 mc with the circuit and tube capacitance of 14 mmfd.

3

We have been talking about distributed, tube, and stray capacitances as they affect television circuits, because at high frequencies these capacitances often form the entire tuning capacitance. At standard broadcast frequencies we use relatively large adjustable capacitance in the tuning capacitors, or may use large adjustable inductance. Then circuit capacitances are not so important, although they can make trouble. The chief difficulty occurs at the higher frequency end of the broadcast tuning range, where the variable capacitor or inductor is adjusted too near its minimum value.

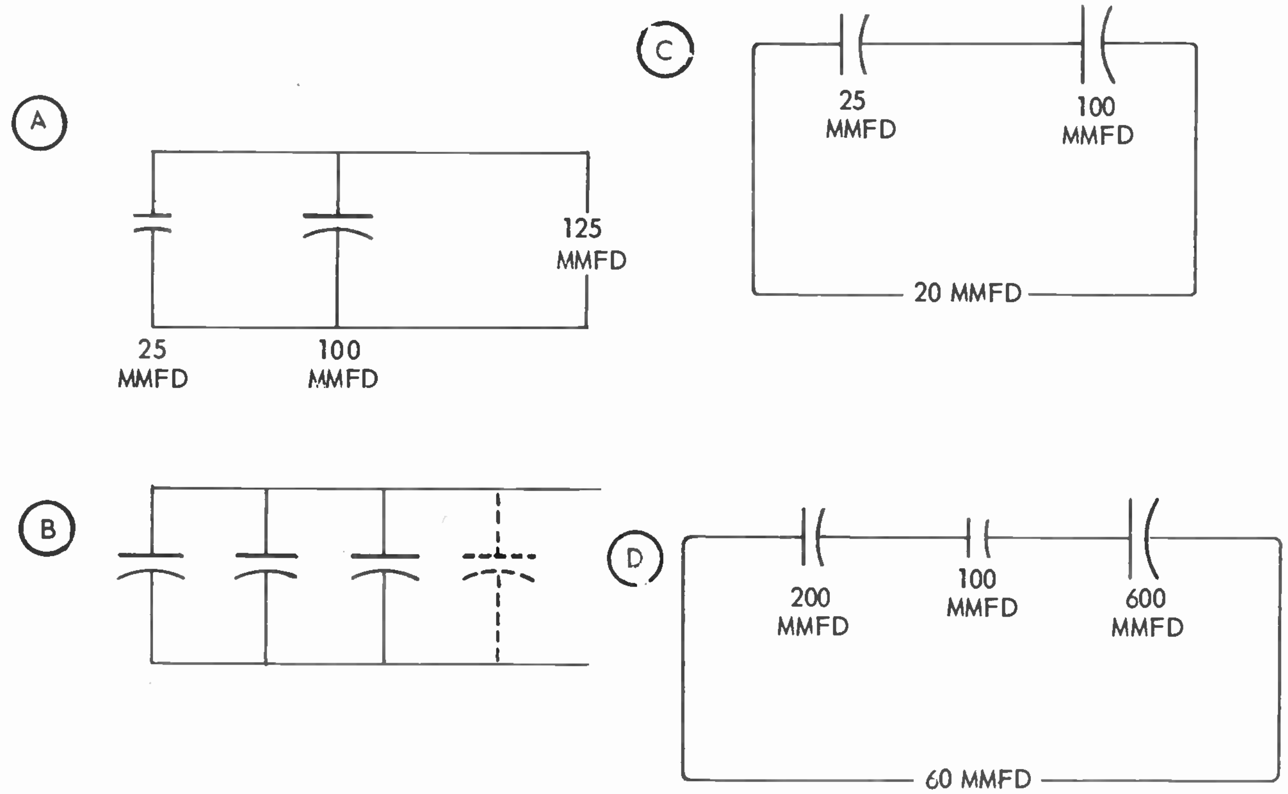


Fig. 27-8. Effective capacitances of capacitors connected together in parallel and in series.

During some tests in the preceding lesson we used 231 microhenrys inductance, a minimum of 25 mmfd in the variable capacitor, and figured on 17 mmfd distributed and circuit capacitance. This combination tuned to about 1,617 kilocycles. Had there been only an extra 10 mmfd of stray capacitance, making a total of 52 mmfd working with the inductance, the highest tunable frequency would have dropped to 1,455 kilocycles. Fifteen of the 10-kilocycle channels at the top of the band would have been completely cut off.

Distributed capacitance in a coil exists between turns of the winding. Although any one winding is a continuous conductor there is a difference of potential between any one turn and the turns on either side when current flows in the winding. When there is a difference of potential between conductors separated by insulation or dielectric, there is an electric field between the conductors. Therefore, there are electric fields between turns of the coil, and this is equivalent to the effect of electric fields between any other conductors – it is capacitance.

There are various structural features of a coil which will either lessen or increase the distributed capacitance, depending on how they are carried out. If the turns of a coil are spaced apart, as in the upper winding of Fig. 27-7, the distributed capacitance is reduced. Extra spacing is equivalent to greater separation between plates of a regular capacitor. In a coil of many turns a spaced winding might make the overall length too great. A relatively short coil may be made by criss-crossing the turns back and forth over one another instead of laying them side by side. The criss-cross method makes a doulateral or a honeycomb coil, in which distributed capacitance is relatively small.

In any winding the distributed capacitance increases more rapidly than the number of turns. That is, 20 turns of given diameter will have more than twice as much capacitance as 10 turns of the same diameter, and 40 turns will have more than twice as much as 20 turns. To reduce this effect some coils are wound in several sections or "pies", with only part of the total turns in each pie. Then all the pies are mounted on one form and connected together.

Using wire in which the metallic conductor is of small diameter reduces distributed capacitance, just as reducing the size or area of plates in an ordinary capacitor lessens the capacitance. Also, the less the dielectric constant of the wire insulation and of the supporting material the less is the capacitance. A self-supporting coil of bare wire has less capacitance than any construction using insulation but otherwise similar. If connections are made to several places along a winding, and some of the turns are not included in the active circuit, the "dead end" turns increase the distributed capacitance.

Stray capacitance exists between every wire and every other nearby wire, also between fixed resistors,

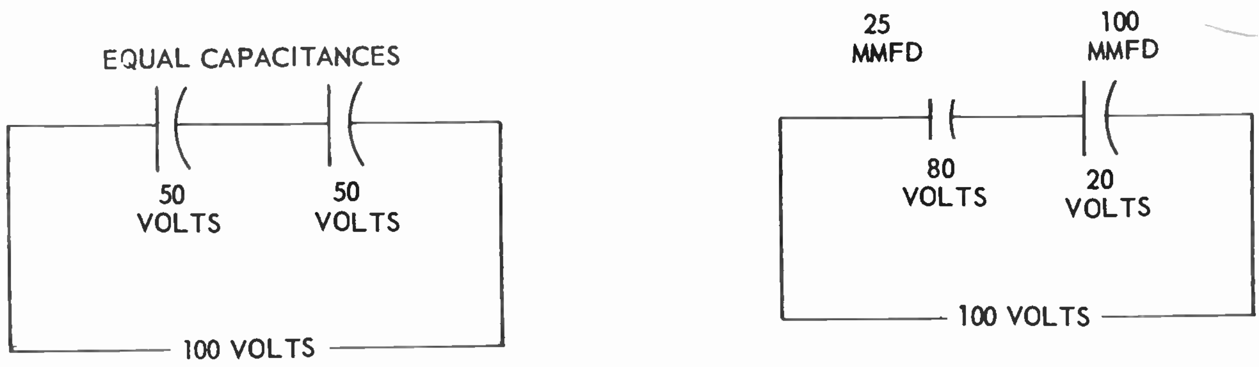


Fig. 27-9. Voltage division between capacitors connected together in series.

fixed capacitors, and other small parts which are near one another. There is capacitance between tube base pins, between lugs on tube sockets, and between insulated terminals of all other parts. There is capacitance also between chassis metal and every insulated part of the receiver.

Stray capacitances may be lessened by keeping everything which contains metal as far as possible from everything else which contains metal. This is especially important with conductors which carry signal currents, as in all grid circuits and all plate circuits.

5 Parts which are of small physical size or of small dimensions have less capacitance to each other and to chassis metal than have larger parts. This is one reason for favoring small sized fixed resistors, fixed capacitors, and other circuit units in television and f-m receivers. Don't conclude that small parts were used to cut costs, they probably indicate careful engineering. Stray capacitance is reduced also by using materials of low dielectric constant for all insulating supports.

CAPACITANCES IN PARALLEL AND SERIES. When discussing the input and output capacitances of the tubes between which our tuning coil is connected in Fig. 27-6 it was stated that these capacitances are in parallel with the distributed capacitance of coil, and that such capacitances add together. It was stated also that the coupling capacitor reduces the effect of tube output capacitance. Unless we get clearly in mind the effects of connecting capacitances in parallel and in series with one another we shall encounter many puzzling situations while working with resonant circuits, and, for that matter, while working with other radio circuits. Therefore, before going further, we shall take a little time to look into the elementary principles of capacitor connections.

If you connect two capacitors in parallel with each other, as at *A* in Fig. 27-8, they act like a single capacitor whose capacitance is the sum of the two separate capacitances. You may connect any number of capacitances in parallel, as at *B*, and the total capacitance always will be the sum of the separate capacitances. Now we know why the input capacitance of the second amplifier tube in Fig. 27-6 adds itself to the distributed capacitance of the coil – the two capacitances are in parallel.

The fact that capacitors in parallel add their capacitances often helps out when making service replacements. If you lack a capacitor of a required value it may be possible to connect in parallel two smaller ones, the sum of whose capacitances makes up the necessary total. Of course, three or even more capacitors may be paralleled if there is space available.

7 The case of capacitors in series with one another makes computations somewhat more difficult. If the series capacitances at *C* and *D* of Fig. 27-8 are all of the same value their effective capacitance will be equal to one capacitance divided by the number of capacitances. For example, the effective capacitance of two 100-mmfd capacitors in series is found by dividing 100 by 2, and is 50 mmfd. Were there four of the 100-mmfd units in series their effective capacitance would be 100 divided by 4, or 25 mmfd.

6 To determine the effective capacitance of two unequal capacitances connected in series, as at *C*, we use the same method of computation employed for resistors in parallel. That is, the effective capacitance is found by dividing the product of the two capacitances by their sum. This is the way it works out for capacitances of 25 and 100 mmfd.

$$\text{Effective capacitance} = \frac{25 \times 100}{25 + 100} = \frac{2500}{125} = 20 \text{ mmfd}$$

If these are more than two unequal capacitances in series with one another, first determine the effective capacitance of any two. Then use this effective capacitance together with the capacitance of a third unit in the regular formula, and keep on including one more capacitance each time until all have been considered.

Supposing we have capacitances of 200, 100, and 600 mmfd in series. Here are the computations.

First: $\frac{200 \times 100}{200 + 100} = \frac{20\,000}{300} = 66.7$ mmfd, approximately

Second: $\frac{66.7 \times 600}{66.7 + 600} = \frac{40\,000}{666.7} = 60$ mmfd

It is important to remember this: The effective capacitance of any number of capacitances in series always must be less than the least of the separate capacitances. Now we know why the coupling capacitor *Cc* in Fig. 27-6 reduces the effective output capacitance of the first tube so far as the coil circuit is concerned. The tube capacitance and coupling capacitance are in series with each other.

All the rules for paralleled and series connected capacitances apply whether the units are microfarads or micro-microfarads. In the first case the answer will be in microfarads, in the second case the answer will be in micro-microfarads. Don't mix the two units in the same formula at the same time.

The rules just mentioned apply to capacitances, which we measure in microfarads or in micro-microfarads. Capacitive reactance, which we measure in ohms, is treated just like resistance, which also is measured in ohms. Resistances which are in series add together, and so do capacitive reactances when they are in series.

The effective or combined reactance of any number of equal capacitive reactances connected together in parallel is equal to the value of one reactance divided by the number of reactances, exactly the same method used for resistances in parallel. If the capacitive reactances in parallel are not equal, we determine the effective or combined reactance from dividing the product by the sum, just as though the values of reactance were values of resistance.

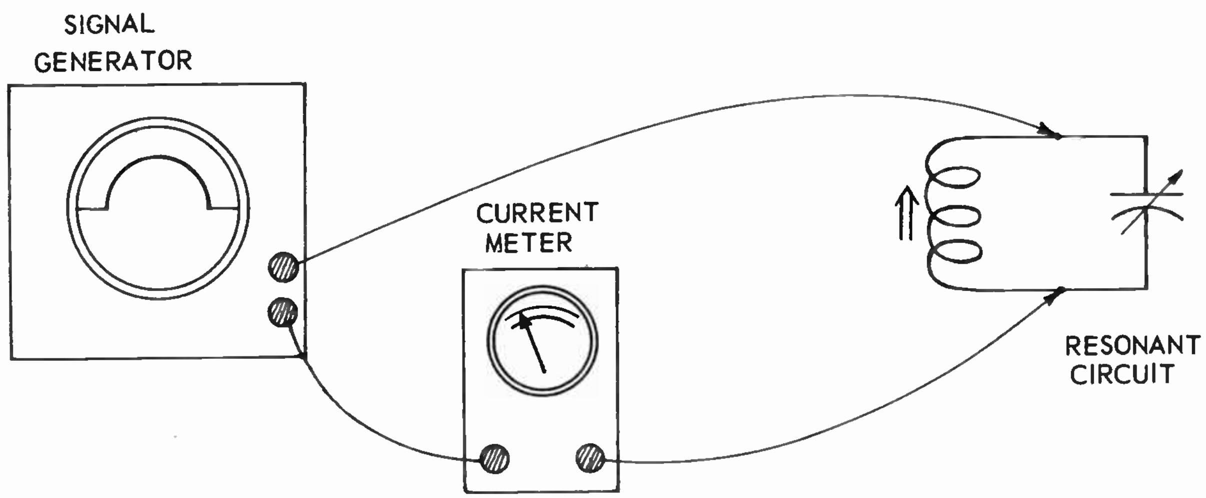


Fig. 27-10. The test setup for measuring current in the parallel resonant circuit.

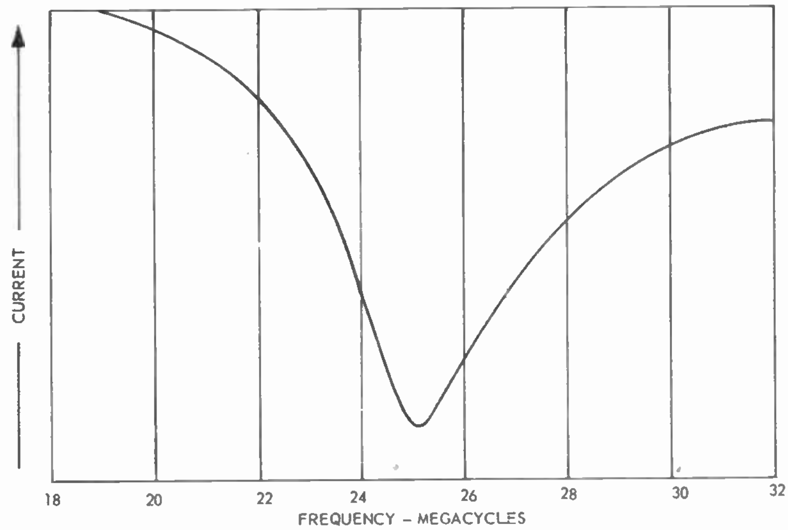


Fig. 27-11. How current changes in the parallel resonant circuit when applied frequency is varied.

Of course, to have capacitive reactances in series we must have capacitances in series, for we don't have this kind of reactance unless there is capacitance. Also, to have capacitive reactances in parallel we must have capacitances in parallel. The differences between methods of computation depend entirely on whether you are figuring with capacitance or with capacitive reactance. Which method to use is easy enough to remember. If what you are working with is measured in ohms (measuring reactance), make computations just as though working with resistances. If it is not measured in ohms (meaning capacitance) use the reverse methods.

The last thing we need consider about capacitors in parallel and in series relates to voltages acting on each separate capacitor. If you look back at diagrams *A* and *B* of Fig. 27-8 it is easy to see that the voltage must be the same across any number of capacitors connected directly together in parallel. The values, of capacitances in the separate units have no effect on the voltage. With all terminals on each side connected together there can be only one potential on each side, and there must be the same difference of potential across the two sides and across all the capacitors.

When capacitors are connected in series across a source of voltage we have to consider the values of capacitance when determining the voltage across each unit. At *A* in Fig. 27-9 are two equal capacitances in series on a 100-volt source. The total voltage always divides equally between equal capacitances. With any number of equal capacitances connected together in series the overall voltage will divide into equal parts across each unit. For instance, with 4 equal capacitances in series each one would be subjected to 1/4 the overall voltage.

At *B* we have unequal capacitances in series. The overall voltage divides inversely as the capacitances. The 25-mmfd unit has 1/4 as much capacitance as the 100-mmfd unit, so the smaller one takes 4 times as much voltage as the larger one. We might say this in reverse. The 100-mmfd capacitor has 4 times as much capacitance as the 25-mmfd capacitor, so the larger unit takes 1/4 as much voltage as the smaller one. Remember this: The smallest capacitance in series always gets the most voltage, and the largest series capacitance always gets the least voltage.

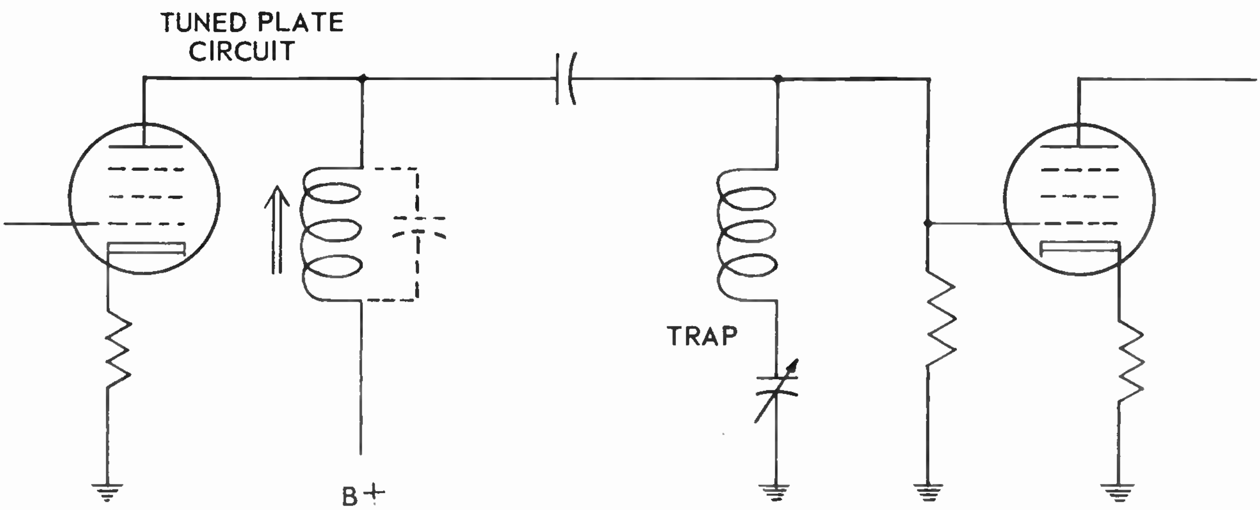


Fig. 27-12. A series resonant coil and capacitor used as a trap in a television amplifier circuit.

The relations between the capacitance and the voltage across a capacitor are just as true with direct voltages as with alternating or signal voltages. Signal voltages in radio and television receivers usually are too small to cause danger of capacitor breakdown. We may, however, get into trouble when using small series connected capacitors in plate and screen circuits where direct voltages may be quite high.

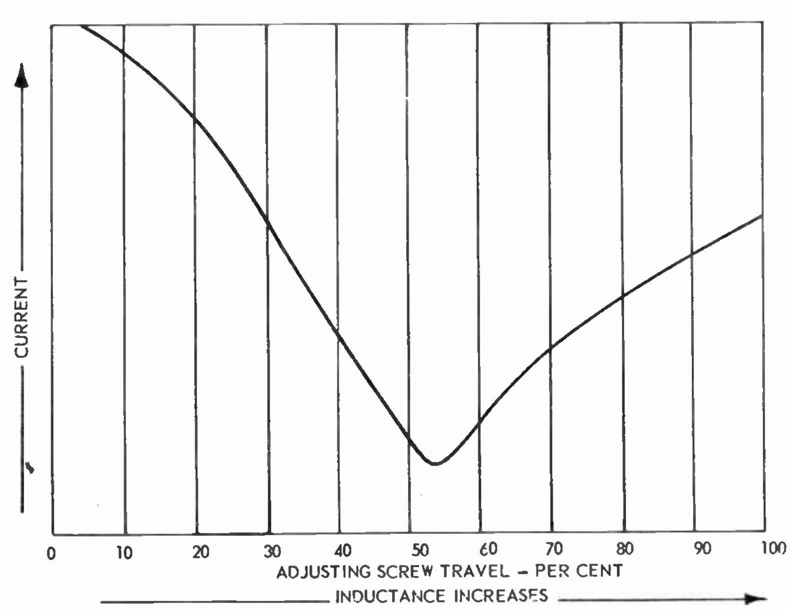


Fig. 27-13. Change of current in the parallel resonant circuit when its inductance is varied.

The safe way is to use no capacitors whose voltage ratings are not amply high for the highest voltage in the circuit, regardless of whether or not any of the capacitors are in series with others. This method also protects remaining capacitors in the circuit in case one of them breaks down. If one of two series connected capacitors should short circuit, the full voltage of the source will come on the capacitor that did not break down. If the voltage rating of the remaining good capacitor is not high enough to withstand full circuit voltage, that good unit will immediately become a bad one.

PARALLEL RESONANT CIRCUITS. Having gotten some of the preliminaries out of the way we now may proceed with our line of investigation on performance of parallel resonant circuits. For the first test the movable core of the coil is adjusted for 2.9 microhenrys inductance and the tuning capacitor is set for 14 mmfd capacitance, the values for resonance at a frequency of 25 mc. Then the paralleled coil and capacitor are connected across the output of the signal generator as in Fig. 27-10. In series with the resonant circuit and the generator is a meter for measuring high-frequency current.

As frequency is gradually increased the current decreases as shown in Fig. 27-11. At 25 mc the current becomes minimum, with only a small fraction of its starting value. With further increase of frequency the current rises, and finally becomes almost as great as at the lowest frequency. A decrease of current must mean that there is increase of impedance in the resonant circuit. When current is minimum the impedance must be maximum, at 25 mc.

Keeping in mind what happened to current at resonance in the series resonant circuit examined earlier, we now may state an important difference between series resonant and parallel resonant circuits. At the frequency of resonance there is maximum current in the series circuit, but there is minimum current in the parallel circuit. Also, at the resonant frequency there is minimum impedance in the series circuit, but there is maximum impedance in the parallel circuit.

Series resonant circuits are used wherever it is desired to provide a path of very small impedance for currents at some certain frequency while preventing or greatly opposing the flow of currents at all other frequencies.

A typical use of series resonant circuits is as traps in television receivers and sometimes in radio receivers. A television trap is shown by Fig. 27-12. The purpose is to get rid of undesired signals that interfere with normal reception. With a series resonant circuit connected from a signal-carrying circuit to ground, and tuned to the frequency of an interfering signal, currents at the interference frequency will flow away to ground with little impedance. Desired signal currents at other frequencies meet relatively high impedance in the trap and are prevented from escaping.

Parallel resonant circuits are used where it is desired to have maximum impedance at a certain frequency. A common use for parallel resonant circuits is in the plate circuit of an amplifier tube, as shown for the left-hand tube in Fig. 27-12. The capacitances which tune the coil to resonance are indicated by the broken-line symbol.

When studying about amplifier tubes in an earlier lesson we learned that the greater the resistance, or impedance, in the plate circuit the greater is the amplification. By having parallel resonant inductance and capacitance in the plate circuit, and tuning to resonance at the frequency to be amplified, that one frequency is greatly strengthened while other frequencies are amplified only a little.

Now for a second test on the parallel resonant circuit. We shall keep the signal generator tuned to a frequency of 25 mc and keep the tuning capacitor set for 14 mmfd capacitance while moving the core inside the coil form. The setup is the same as shown in Fig. 27-10. Commencing with the movable core as far out

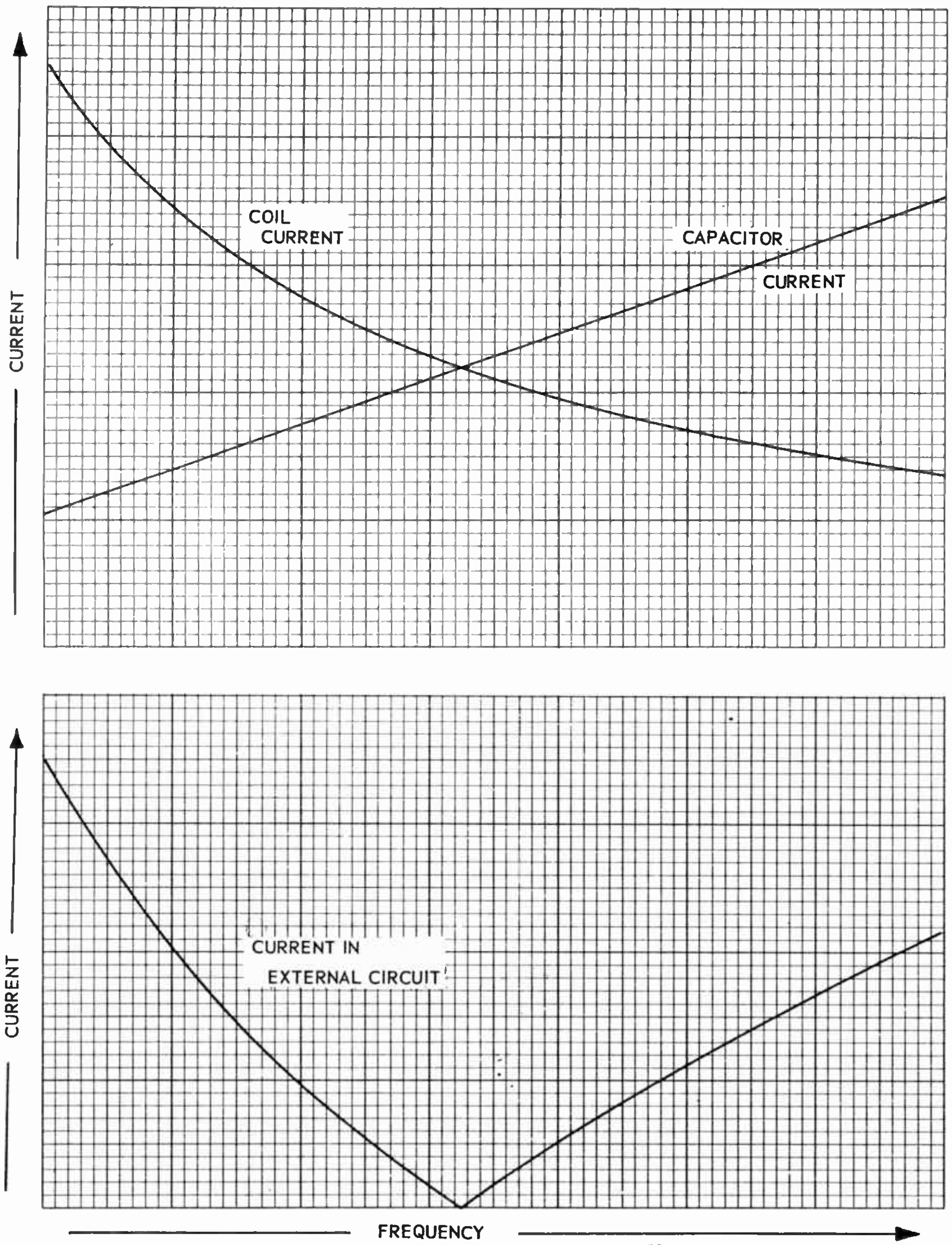


Fig. 27-14. Behavior of a parallel resonant circuit as the applied frequency is varied. At the top are shown changes of current in the coil and capacitor, at the bottom are changes of current in the external circuit.

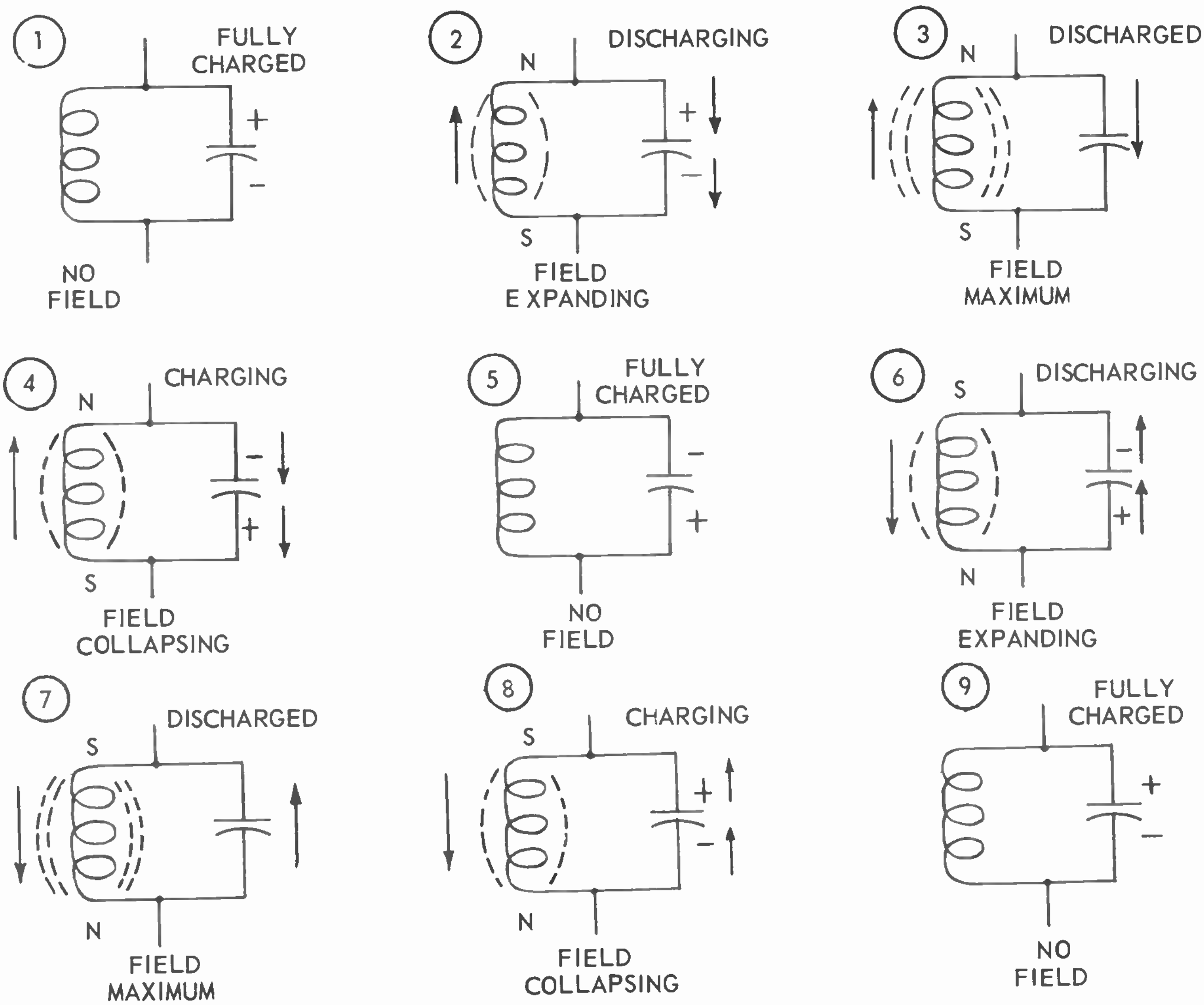


Fig. 27-15. Currents during one cycle in the coil and capacitor of a parallel resonant circuit when tuned to resonance.

of the winding as it will go, we turn the adjusting screw until the core is moved all the way inside the winding. There is change of current as shown by Fig. 27-13.

At first there is large current. As inductance is increased by moving the core farther and farther into the winding the current becomes smaller and smaller. At one certain position of the core there is minimum current. Here the circuit must be tuned to resonance at the applied frequency of 25 mc. With further increase of inductance the current becomes greater, but not so large as in the beginning.

At the point of minimum current the core of the tuning coil is in the position which causes inductance of 2.9 microhenrys. This inductance, with 14 mmfd capacitance, brings about the condition of resonance at a frequency of 25 mc.

CURRENTS IN A RESONANT CIRCUIT. The reasons for changes of current which have been observed in a circuit containing paralleled inductance and capacitance are interesting, and the explanation is quite simple. First we must realize, and keep in mind, that voltage across the coil always must be the same as voltage across the capacitor of a parallel resonant circuit. This is true because the two ends of the coil are connected to the two sides of the capacitor. At the end of the coil which is connected directly to one side of the capacitor there can be only one potential. At the other end of the coil and other side of the capacitor, likewise connected together, there can be only one potential. Then the difference between these two potentials is across both the coil and the capacitor.

High-frequency voltage applied from a source, such as our signal generator, causes high-frequency currents to flow in both the coil and the capacitor. Naturally, the current in the coil must depend on coil impedance and source voltage. Current in the capacitor must depend on capacitor impedance and source voltage. Resistances in coil and capacitor are small in comparison with inductive and capacitive reactances, except at resonance, so we may talk about changes of reactance as controlling the currents.

As we have learned before, inductive reactance of the coil rises with increase of frequency. Then coil current must drop as frequency increases. The drop of current in the coil is shown by the curve at the top of Fig. 27-14.

Capacitive reactance of the capacitor becomes less and less as frequency increases. Then current in the capacitor will become greater with increasing frequency. This rise of current is shown by the straight line at the top of Fig. 27-14.

At any one instant the current in the coil is flowing in a direction opposite to current in the capacitor, if we take these currents with reference to the external circuit going to the source. That is, when current is flowing toward the high side of the source in the coil, the current in the capacitor is flowing away from the high side of the source.

Then, so far as the external circuit is concerned, the coil current and capacitor current are flowing in opposite directions. But you cannot have current flowing both ways in the same conductor at the same time. Actual current in the external circuit, in the leads to the signal generator, is the difference between the coil current and the capacitor current.

The difference between the two currents is shown by the curve at the bottom of Fig. 27-14. Compare this curve, arrived at by computation, with the curves of Figs. 27-11 and 27-13, which show actual measured currents. The only difference is that circuit resistances are affecting the measured currents, while the curve at the bottom of Fig. 27-14 shows only the effects of reactance, without resistance. At the frequency of resonance the reactive currents in coil and capacitor are equal. Their difference is zero, and there is no reactive current in the external circuit. At resonance the only current in the external circuit is that corresponding to resistance.

Fig. 27-15 shows what is happening at resonance during one complete cycle of high-frequency current in the parallel resonant circuit. There is no current, or the least possible current in the external circuit, but there are large currents in both coil and capacitor. The instantaneous directions of these internal currents are shown by arrows alongside the symbols for coil and capacitor. Here is an explanation of each of the numbered diagrams.

1. We come in at the instant in which the capacitor is fully charged in the polarity marked, and when there is no current flowing in either direction.

2. The capacitor begins to discharge through the coil. The resulting current produces a magnetic field around the coil. Energy is going from the capacitor charge (or electric field) into the magnetic field around the coil.

3. There is maximum current and maximum magnetic field around the coil at the instant in which the capacitor is fully discharged. Remember, no charge means only that electron quantities have at this instant become equal in the two sides of the capacitor, it does not mean that there are no electrons. Note that the top of the coil is shown as its magnetic north pole, and the bottom as the south pole.

4. Current continues in the same direction, as energy coming out of the collapsing field of the coil goes into a new charge of the capacitor.

5. The capacitor becomes fully charged just as the field disappears from around the coil. Since the capacitor can take no more charge at the existing voltage, current stops during this instant. This is the reason why there is no magnetic field around the coil, there can be no such field unless current is flowing in the coil. Note the electric polarity of the capacitor charge and compare it with the polarity of diagram 1.

6. The capacitor commences to discharge, sending current through the coil to produce a new magnetic field.

7. There is maximum current through both capacitor and coil, and maximum magnetic field around the coil at the instant in which the capacitor loses all its charge or the instant in which electron quantities become equal on both sides of the capacitor. Note the magnetic polarity of the coil and compare it with the polarity in diagram 3.

8. Now the magnetic field is collapsing and energy is passing into a new electric field and new charge being formed in the capacitor.

9. The capacitor becomes fully charged just as the magnetic field disappears. Current must stop as the capacitor reaches full charge. This is where we came in, as you will see by comparing this diagram with number 1.

Currents in the coil and capacitor just surge back and forth between these two elements. The strength of this circulating current is limited only by resistance in the coil and capacitor. Were there no resistance at all, even the smallest voltage would cause current infinitely great, for at resonance the reactances neutralize each other and furnish no opposition to flow of current.

Back in Fig. 27-14 we saw that coil current is greater than capacitor current at frequencies below resonance, while at frequencies above resonance the capacitor current is greater than coil current.

These facts are shown in another way by Fig. 27-16. At frequencies below resonance there is more coil current than capacitor current. Part of this coil current is that which circulates back and forth between coil and capacitor. The remainder of the larger coil current must flow back and forth in the line or in the external circuit. Since all the line current is flowing in the inductance, or the coil, the parallel resonant circuit acts at frequencies below resonance like an indnctance so far as the external circuit or the line is concerned.

At resonance there is minimum line current, while circulating currents in coil and capacitor are equal or nearly so.

At frequencies above resonance there is more capacitor current than coil current. Part of this greater capacitor current is that which circulates back and forth through the coil and capacitor. The remainder of the larger capacitor current flows also in the line. Now all the line current is flowing in the capacitor. As a result the parallel resonant circuit at frequencies above resonance acts like a capacitance so far as the external circuit is concerned.

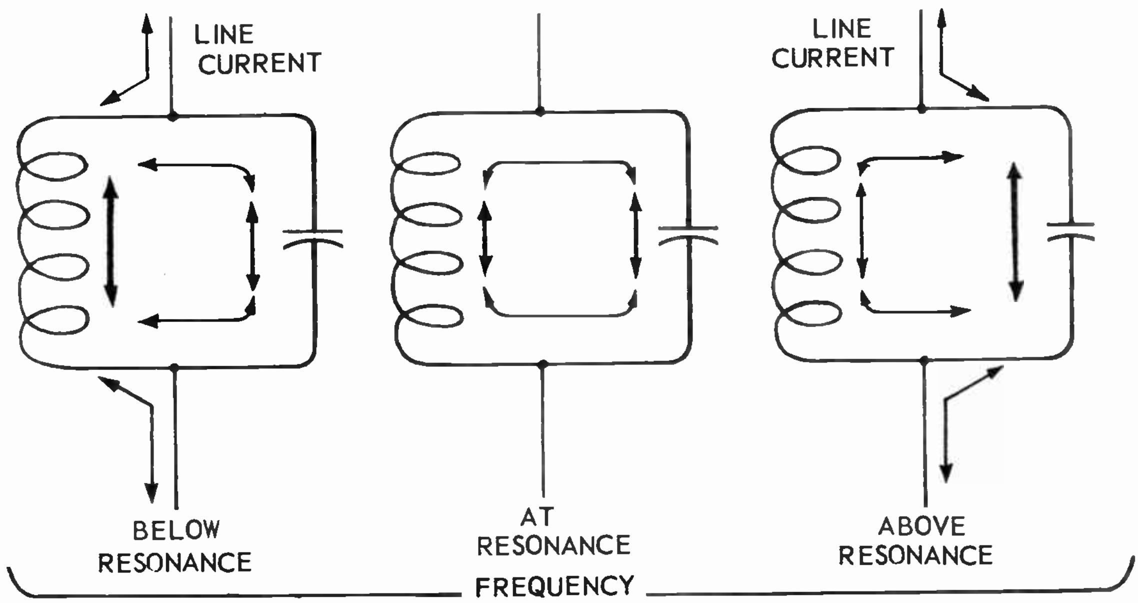


Fig. 27-16. Line current and currents in the paralleled coil and capacitor at frequencies below and above resonance.

USEFUL FORMULAS. In this lesson have been mentioned many combinations of capacitance, inductance, and frequencies of resonance. You might like to compute some of these relations for yourself. Someday you may have to do so. Here are the formulas

$$\text{Kilocycles} = \frac{159155}{\sqrt{\text{mmfd} \times \text{microhenrys}}} \qquad \text{Megacycles} = \frac{159.155}{\sqrt{\text{mmfd} \times \text{microhenrys}}}$$

When you wish to determine the capacitance required for resonance at a certain frequency when using a given inductance it may be done as follows.

$$\text{Mmfd} = \frac{25\,330\,000\,000}{\text{kc}^2 \times \text{microhenrys}} \qquad \text{Mmfd} = \frac{25330}{\text{mc}^2 \times \text{microhenrys}}$$

In other cases you may wish to learn the inductance needed for resonance at a certain frequency when using a given capacitance. It is done with the following formulas.

$$\text{Microhenrys} = \frac{25\,330\,000\,000}{\text{kc}^2 \times \text{mmfd}} \qquad \text{Microhenrys} = \frac{25330}{\text{mc}^2 \times \text{mmfd}}$$

In the preceding lesson we used the oscillation constant for 1,000 kilocycles, which is 25,330. Doubtless you remember that an oscillation constant is a number which, divided by microhenrys inductance, gives the number of mmfd for resonance, and when divided by mmfd capacitance gives the number of microhenrys for resonance. In other words, the oscillation constant is the product of inductance and capacitance for the specified frequency. You may determine the oscillation constant for any frequency with these two very simple formulas, one for frequency in kilocycles and the other in megacycles.

$$\text{Constant} = \frac{25\,330\,000\,000}{\text{kc}^2} \qquad \text{Constant} = \frac{25330}{\text{mc}^2}$$

TELEVISION

LESSON NO. 28

PHASE RELATIONS IN A-C CIRCUITS

Training Your Mind

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

COMMERCIAL TRADES INSTITUTE

Training Your Mind
C. T. I.
Perfecting Your Skill

Chicago, Ill.

LESSON NO. 28

PHASE RELATIONS IN A-C CIRCUITS

When you talk with television technicians they speak of phase shifts, phase angles, phase detectors, phase controls, phase differences, phase displacements, and talk about "phase" as familiarly as about voltages and currents. If we learn to talk and think the same way the remainder of our tests on resonant circuits and almost everything else we do with alternating voltages and currents will have much more real meaning.

1 Phase really is a simple thing. It refers to the point in a cycle at which an alternating voltage or current has arrived at a certain instant of time. If we are considering two voltages, two currents, or a voltage and a current which are alternating, one of them may have progressed farther or not so far as the other one at at certain instant of time. Their relative positions in their cycles are described as the phase of one in relation to the other. In most discussions you could substitute the word "time" for the word "phase" with no great change in the meaning.

Were you going to do nothing but service sound radios there would be little need for learning about phase. Signal voltages at the various frequencies for sound may get out of time (phase) to almost any degree and the human ear never will notice any resulting distortion of speech or music. But if there is any considerable phase shift (time shift) between television signal voltages of different frequencies there is plenty of trouble. Our eyes are highly sensitive to picture distortion resulting from phase shifts.

Although changes or differences of phase (time) will cause trouble in many television circuits, such changes are put to very good use at other places. As just one example, the majority of automatic controls for horizontal sweep of the electron beam across the picture tube screen depend on phase (time) differences between two voltages. All in all, the time we spend in learning about phase will be regained many times over in easier and quicker understanding of much of our future work.

PHASE IN A SINE WAVE. At the top of Fig. 28-1 is a magnet between whose poles is a uniform magnetic field, a field in which the magnetic force is uniformly distributed. Rotating within the magnetic field is a conductor. The conductor cuts through the magnetic lines of force, and an emf is thus induced in the conductor. Were the conductor part of a closed electric circuit there would be electron flow or current flow in the circuit.

The right hand drawing of Fig. 28-1 shows an end view of the magnetic field and the rotating conductor. The circle around which the conductor travels may be considered as divided into 360° (360 degrees), just as every other circle may be so divided. Let's assume that the conductor begins one rotation at the point in the circle marked 0°, and also 360° because this is both the beginning and the end of the circle. For just an instant the conductor is moving parallel with the magnetic lines rather than cutting through any of the lines. Therefore, at this instant, there is no induction of emf.

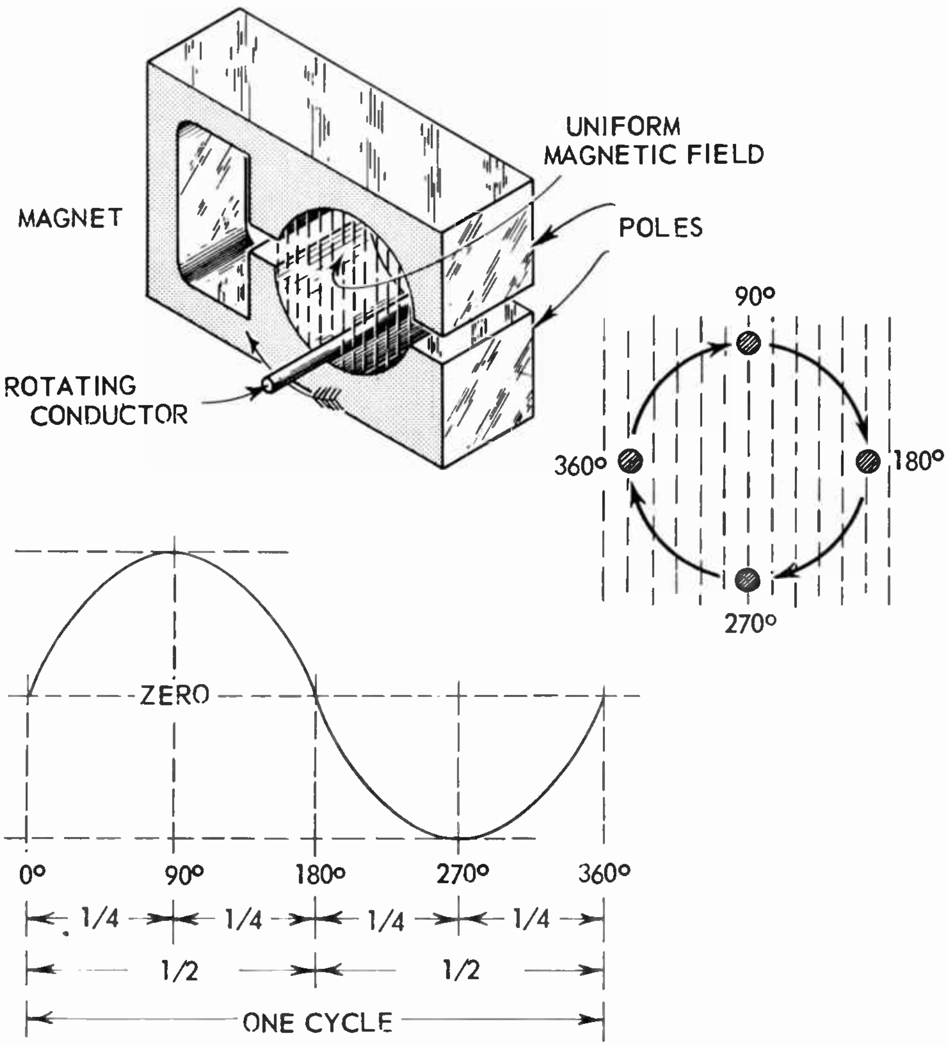


Fig. 28-1. How a sine wave of voltage or current is developed.

As the conductor continues to move at constant speed it cuts through more and more magnetic lines in any given interval of time. Since induced emf is proportional to the rate of cutting, the emf becomes greater and greater until the conductor comes to the position marked 90°. Here the rate of cutting is maximum, because the conductor is moving at right angles to the magnetic lines, and there is maximum induced emf.

With continued rotation of the conductor the rate of cutting decreases, and becomes zero at the point marked 180°, where travel again is parallel with the magnetic lines. Accordingly, the induced emf drops to zero at this point. Then there is another increase in rate of cutting and proportional increase of induced emf while the conductor moves from 180° to 270° around the circle. This is followed by a decrease in the cutting rate and in the induced emf as the conductor moves on from 270° to 360°, the end of this rotation. All this repeats over and over again as the conductor continues its rotation.

The lower drawing of Fig. 28-1 shows how the induced emf increases, decreases, increases, and again decreases while the conductor completes one circle. This curve, like the circle traveled by the conductor, is marked off in degrees. From 0° to 180° the emf is shown as acting in one direction or one polarity, by drawing the curve above the horizontal zero line. From 180° to 360° the emf is shown as of opposite direction or polarity, by drawing the curve below zero. This reversal of polarity occurs because, during the first half of the circle, the conductor moves through the magnetic lines from left to right, and during the second half it moves from right to left. Polarity depends on the direction of cutting in relation to the direction of lines of force.

We have shown one complete cycle of alternating emf. Were the emf acting between points in an electric circuit this would be one cycle of alternating voltage. Were the emf to be causing electron flow or current, our curve could represent one cycle of alternating current. The waveform shown by this curve is called a sine wave.

Any point in the cycle which has been reached by emf, voltage, or current may be specified by stating the number of degrees from the beginning of the cycle to that point. We may speak of the first 180° as a half-cycle, and of the second 180° as another half-cycle. Or we may call each 90° portion of the cycle a quarter-cycle.

A cycle and its divisions in degrees or fractions are not related to frequency. Whether the frequency is 60 cycles or 60 megacycles, or anything else, we still have 360 electrical degrees, four quarter-cycles, and two-half cycles. Neither is the cycle or its divisions related to amplitude or strength of emf, voltage, or current. We might show one alternating voltage by the full-line curve of Fig. 28-2, and some different voltage by the broken-line curve, with different amplitudes indicated by different heights of the curves. Two currents might be similarly indicated.

PHASE DIFFERENCES. A phase difference is the number of electrical degrees separating similar peak values of two alternating electrical quantities existing together. By similar peaks we mean two adjacent positive peaks or two adjacent negative peaks. By alternating electrical quantities we mean two alternating voltages, two alternating currents, or an alternating voltage and an alternating current.

In Fig. 28-2 the two similar peaks always occur at the same instant of time. There is no phase difference. We say that the two quantities (voltages or currents or both) are "in phase". We might say also that there is zero phase angle between the two quantities.

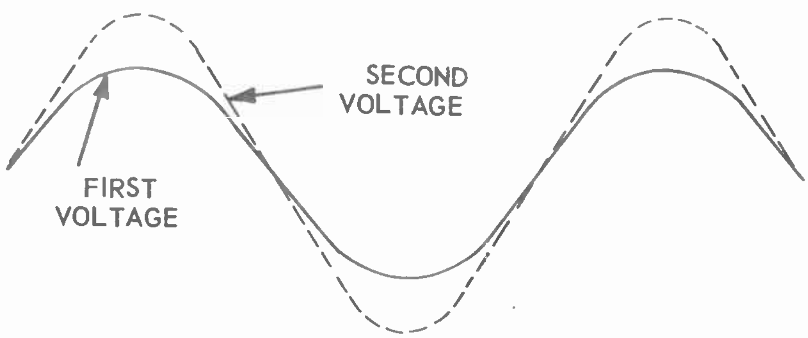


Fig. 28-2. Two voltages which are in phase.

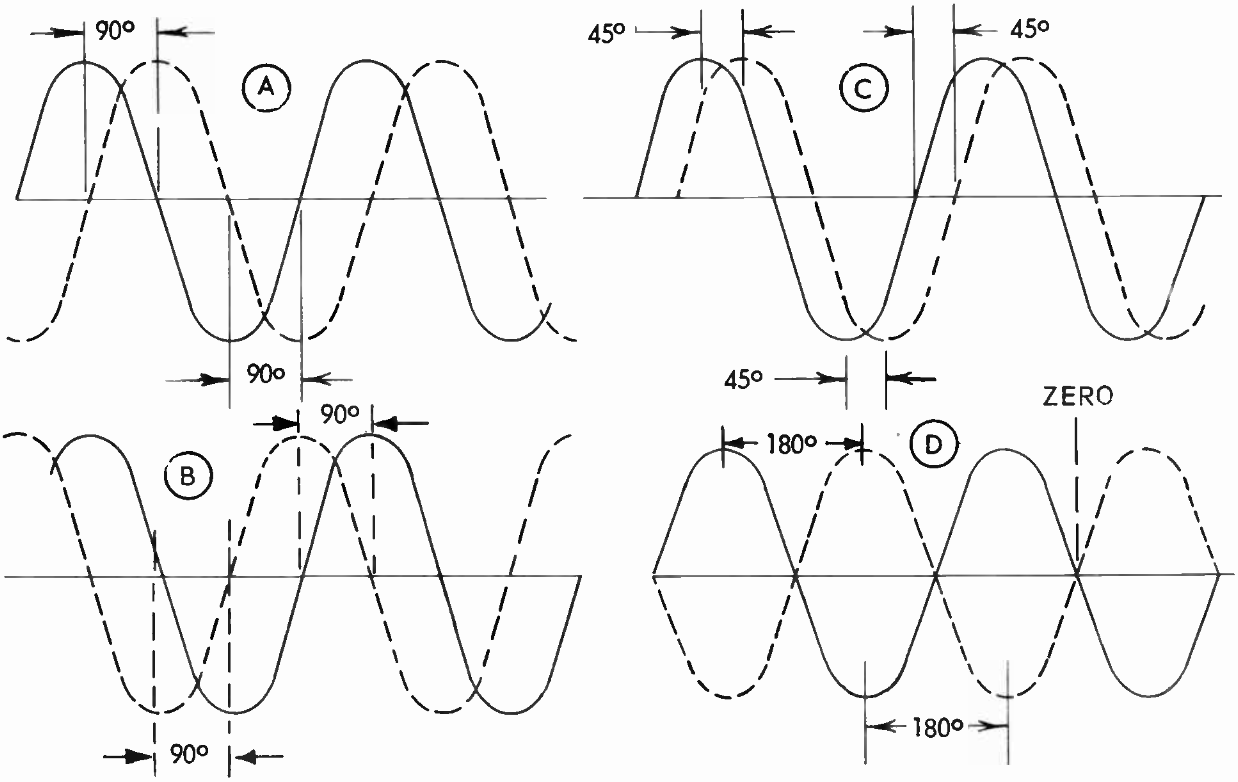


Fig. 28-3. Phase differences between two voltages, two currents, or a voltage and a current.

At *A* in Fig. 28-3 there is a phase difference of 90°. There is separation, in time, of 90° between each pair of positive peaks and between each pair of negative peaks. Since 90° equals a quarter-cycle there is a phase difference of a quarter-cycle.

Curves such as we are looking at always refer to time. Time is assumed to increase from left to right along the curves. Anything shown toward the left occurs before anything shown toward its right, or any value shown toward the left on a curve is occuring earlier than whatever is toward the right. The electrical quantity (voltage or current) represented by the full-line curve at *A* in Fig. 28-3 is going through all its changes before the similar changes of the quantity represented by the broken-line curve. Then we say that the quantity shown by the full-line curve is "leading" the one shown by the broken-line curve. We might say just as correctly that the quantity shown by the broken line curve is "lagging" the other quantity, because all its changes occur later than similar changes of the quantity shown by the full-line curve.

In diagram *B* we again have a phase difference of 90°. Here the quantity shown by the broken-line curve leads the other quantity by 90°, and the one shown by the full-line curve lags by 90°. Assume that the full-line curves represent an alternating voltage while the broken-line curves represent the current in the same circuit. Then in diagram *A* the voltage leads the current and the current lags the voltage. In diagram *B* the current leads the voltage and the voltage lags the current.

Diagram *C* shows a phase difference of 45° between the two quantities. Note that in diagrams *A*, *B*, and

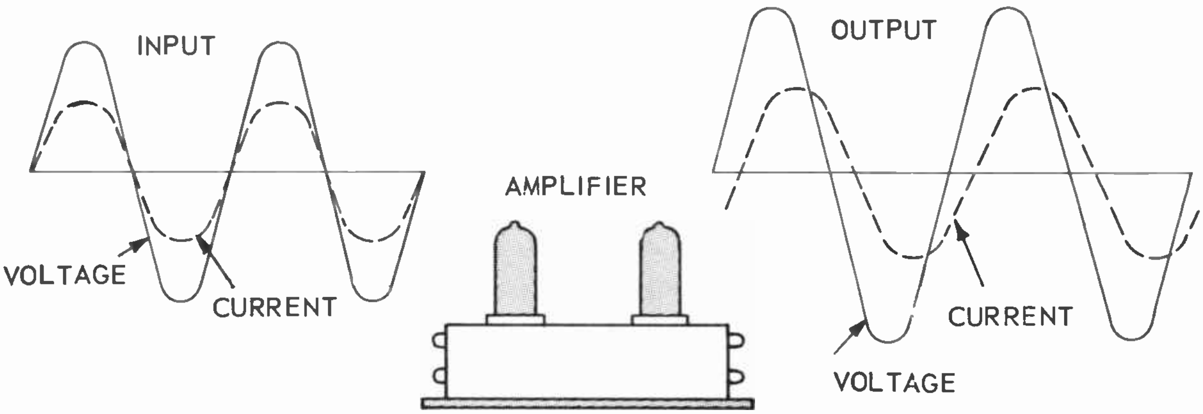


Fig. 28-4. Phase shift occurring between input and output of an amplifier.

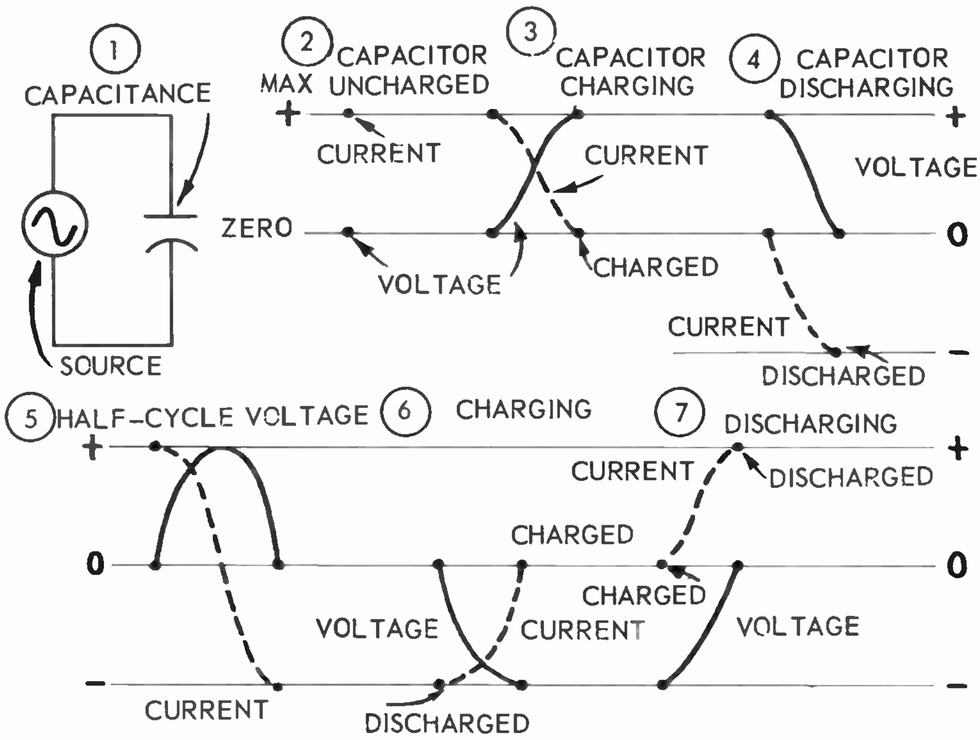


Fig. 28-5. Alternating voltage and current in a capacitance.

C there is the same phase difference between instants at which the two quantities go through zero values as there is between the instants at which they go through peak values.

At *D* we have a phase difference of 180°. Since this is a half-cycle we might speak of a phase difference of a half-cycle. Positive peaks of one quantity always occur at the same instant of time as negative peaks of the other quantity. Consequently, the two quantities are said to be in "opposite phase".

In every case which has been shown by Fig. 28-3 the two electrical quantities are "out of phase". Whenever there is a phase difference between two quantities, no matter what its value, the two quantities are out of phase. The only instances in which the quantities are not out of phase is when they are "in phase", as shown by Fig. 28-2.

When we talk about a phase difference we ordinarily refer to what is happening to two electrical quantities existing together at the same point in a circuit. We might be referring to voltage and accompanying current at the input to an amplifier, or at the output. But sometimes the phase of two quantities at the output of an amplifier or other circuit is not the same as at the input to the same amplifier or circuit.

As an example, Fig. 28-4 shows voltage and current in phase at the input to an amplifier. In the output from this amplifier the current lags the voltage, or the voltage leads the current. This is one of the difficulties which designers, and service technicians, have to contend with in television video amplifiers and in other parts. This kind of change in phase is called a phase shift rather than a phase difference. The phase of the current has shifted with reference to phase of voltage, or voltage has shifted with reference to current, as the signal goes through the amplifier.

5 *ALTERNATING VOLTAGE AND CURRENT IN A CAPACITOR.* Now we are going to go through a complete explanation of something which is decidedly interesting, but of which you need remember only the final result. The explanation is necessary because, without it, you never would believe the final result to be possible. What we are about to prove is this: In a capacitance with which is associated neither resistance nor inductance the current always leads the voltage by 90°. Think of it; changes of current which result from an alternating voltage occur a quarter-cycle before the changes of voltage. It doesn't sound reasonable, yet it is true.

Here we go for the explanation. The following numbered paragraphs refer to similarly numbered diagrams of Fig. 28-5. The voltage being shown is that being applied from the source. Current is that flowing out from and back toward the source, it is the charging and discharging current of the capacitor.

1. A capacitance or capacitor, shown by its symbol, is connected to a source of alternating voltage and current. The source is represented by a circle enclosing a sine wave. This is a standard symbol for any kind of a-c source.

2. We enter the circuit at an instant in which the capacitor is uncharged, and in which the alternating applied voltage is commencing to increase from zero in a positive direction. The capacitor now has no voltage of its own with which to oppose the applied voltage, and is assumed to have no resistance. Consequently, the least bit of voltage from the source causes a very great rate of electron flow in the capacitor plates, it causes maximum current. Then for a starting point we show the voltage at zero and the current at maximum.

3. During the first quarter-cycle the voltage rises to its maximum while the capacitor is charging. While the capacitor is charging there is an increasing potential difference between its plates. This potential difference between its plates. This potential difference or voltage of the capacitor opposes the applied

voltage. It is only the difference between applied voltage and capacitor voltage that is acting to carry on the charging. As the difference becomes less, with increasing charge, there must be a decreasing rate of charge or decreasing charging current. So we show current decreasing as applied voltage rises. When applied voltage reaches its peak the capacitor voltage immediately comes up to an equal value. Then there is no difference of voltage to continue the charge, charging current stops, and we have zero current at the time of maximum voltage.

4. These two curves begin where the preceding ones left off, with maximum voltage and zero current. During this second quarter-cycle the applied voltage drops from maximum to zero. As applied voltage drops, it leaves more voltage on the capacitor plates than from the source, and the capacitor commences to discharge. The rate of discharge, which is the current, becomes greater and greater, and reaches its maximum at the instant in which applied voltage goes to zero. As a result of all this we must show current increasing from zero to maximum while applied voltage goes from maximum to zero.

5. **Putting together the changes of applied voltage and of current shown by diagrams *3* and *4* completes our curves for the first half-cycle, and here they are. Applied voltage has changed from zero to maximum to zero. The capacitor has charged and discharged.**

6. Here we begin the second half-cycle, with zero voltage, maximum current, and the capacitor discharged – as at the end of diagram *4*. What happens now is really exactly the same as described in connection with diagram *3*, except that the polarities of voltage and current are reversed. We are working now with the voltage going from zero to maximum negative while the capacitor is charged. Of course, the charge on the capacitor also is reversed, the plate formerly made positive now is made negative, and the one formerly made negative now becomes positive.

7. We continue the action where it ended in the preceding diagram. Except for reversals of polarity we have here the same performance witnessed in diagram *4*. Applied voltage goes from maximum negative back to zero, and current goes from zero to maximum as the capacitor discharges. At the end of the action of this diagram we are right back where we began in diagrams *2* and *3*, with zero applied voltage, a discharged capacitor, and maximum current. We have completed one cycle.

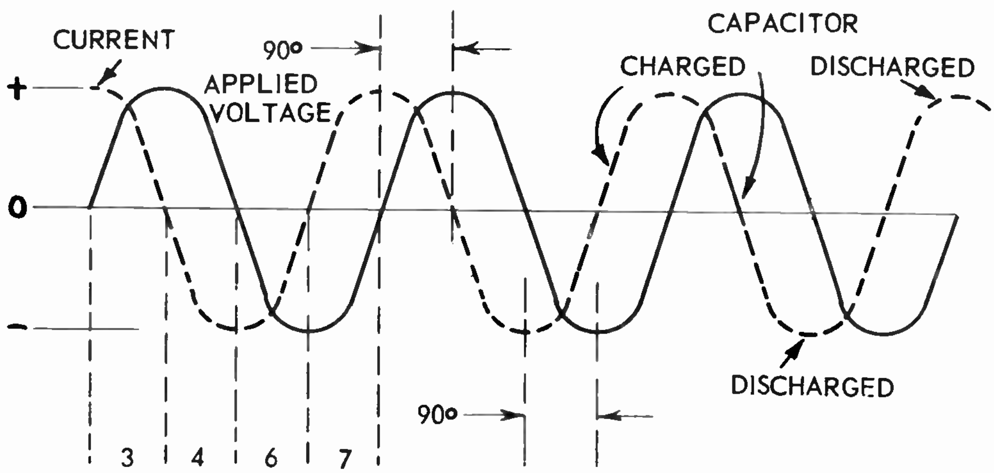


Fig. 28-6. In a capacitance the current leads the voltage.

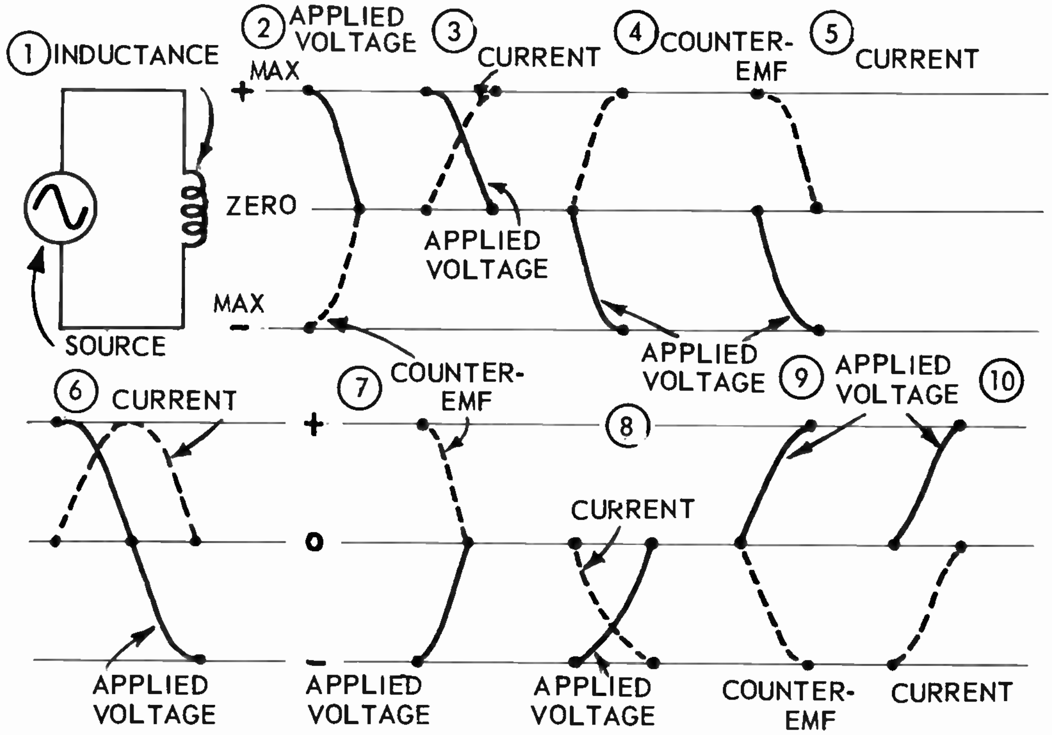


Fig. 28-7. Alternating voltage and current in an inductance.

The cycle is shown all put together at the beginning of Fig. 28-6, where the first four quarter-cycles are the same as explained in connection with diagrams *3, 4, 6,* and *7* of the preceding figure. This cycle is followed by any number of others just like it. Always the current is ahead of the voltage by 90°, current always leads the voltage by 90° or a quarter-cycle in a circuit containing only capacitance.

ALTERNATING VOLTAGE AND CURRENT IN AN INDUCTOR. Supposing someone were to ask you to state the phase relation between voltage and current in an inductance or inductor assumed to have no resistance. The chances are that you would give the correct answer. You would recall that inductance and capacitance always act oppositely, you now know the phase relation of voltage and current in a capacitance, and you would say that current lags the voltage by 90° in a resistance-less inductor. This, of course, is correct.

Doubtless you would like to know exactly why the current lags the voltage in an inductance. The explanation will help toward a better understanding of alternating voltages and currents in general, although there is no real need to remember all the steps we shall go through in arriving at the conclusion. You must, however, never forget the final conclusion, that current really does lag the voltage by a quarter-cycle in an ideal inductance having no resistance.

The following numbered paragraphs refer to the numbered diagrams of Fig. 28-7. In these examinations we shall be talking about the counter-emf which results from electromagnetic induction, as well as about the voltage applied from the source. This is because what happens to counter-emf is easily determined, and that tells us what the current must be doing. A review of the lessons on magnets and on induction will refresh your memory on the subject of counter-emf.

1. The a-c source and the inductance, shown by symbols, are connected in series.

2. We come into the circuit at an instant in which applied voltage is of maximum positive value, and follow the first quarter-cycle during which this applied voltage drops to zero. This voltage causes a current of changing value to flow in the inductor. The changing current produces a changing magnetic field around the inductor, and the changing magnetic field induces a counter-emf in the inductor. A counter-emf always is nearly equal in strength to the applied voltage, but is of opposite polarity – it opposes the applied voltage. Consequently, during this quarter-cycle we must show the counter-emf as changing from its maximum negative value to zero.

3. Now to determine what the current is doing in this first quarter-cycle. We know that the strength of the counter-emf must be proportional to the rate of change of current and of magnetic field, so we know that current must be changing at a rate which will induce the kind of counter-emf existing in the circuit. The question becomes, when or where will there be the maximum rate of current change, and where will the rate of change be least?

The rate of change must be maximum just after current commences to increase from zero. Here is the reason: As current changes from 1 per cent to 10 per cent of its full value during a certain period the rate of change is 10 times. With equal increase from 10 to 20 per cent during a like period the rate of change is only 2 to 1, the current has merely doubled. In the next like period an equal change from 20 to 30 per cent is a rate of change of only 1½ to 1, and so it goes with a decreasing rate of change. Accordingly, the rate of change must be maximum when current commences at zero, which is the instant in which applied voltage and counter-emf are maximum. Then current during the first quarter-cycle must be shown increasing from zero to maximum positive, as in the diagram.

4. 5. Now for the second quarter-cycle, in which applied voltage goes from zero to maximum negative. Counter-emf must go at the same time from zero to maximum positive. Again we have the question, what kind of current change will induce this particular counter-emf? According to the previous explanation of where the rate of current change is greatest and least we have maximum rate of change when the

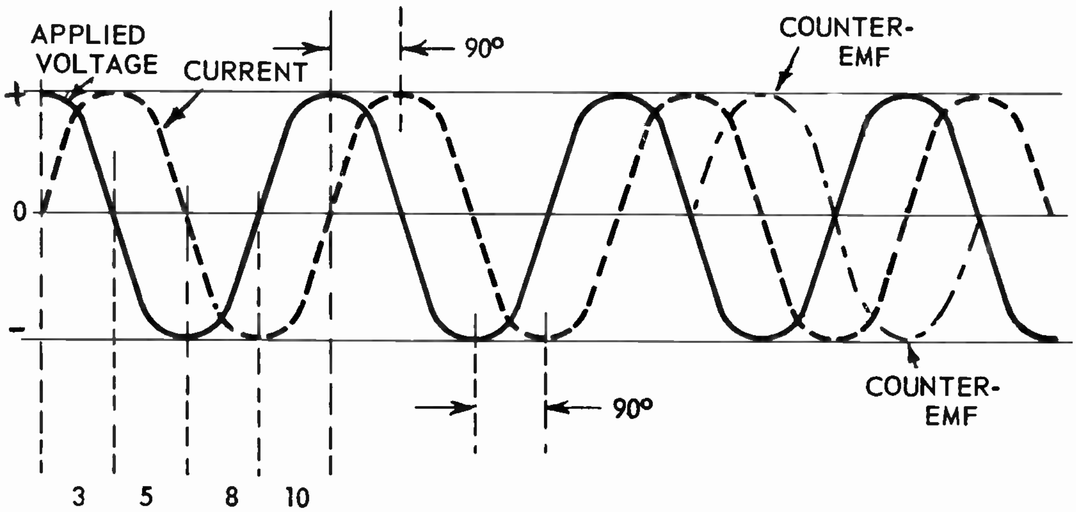


Fig. 28-8. Current lags the voltage in an inductance.

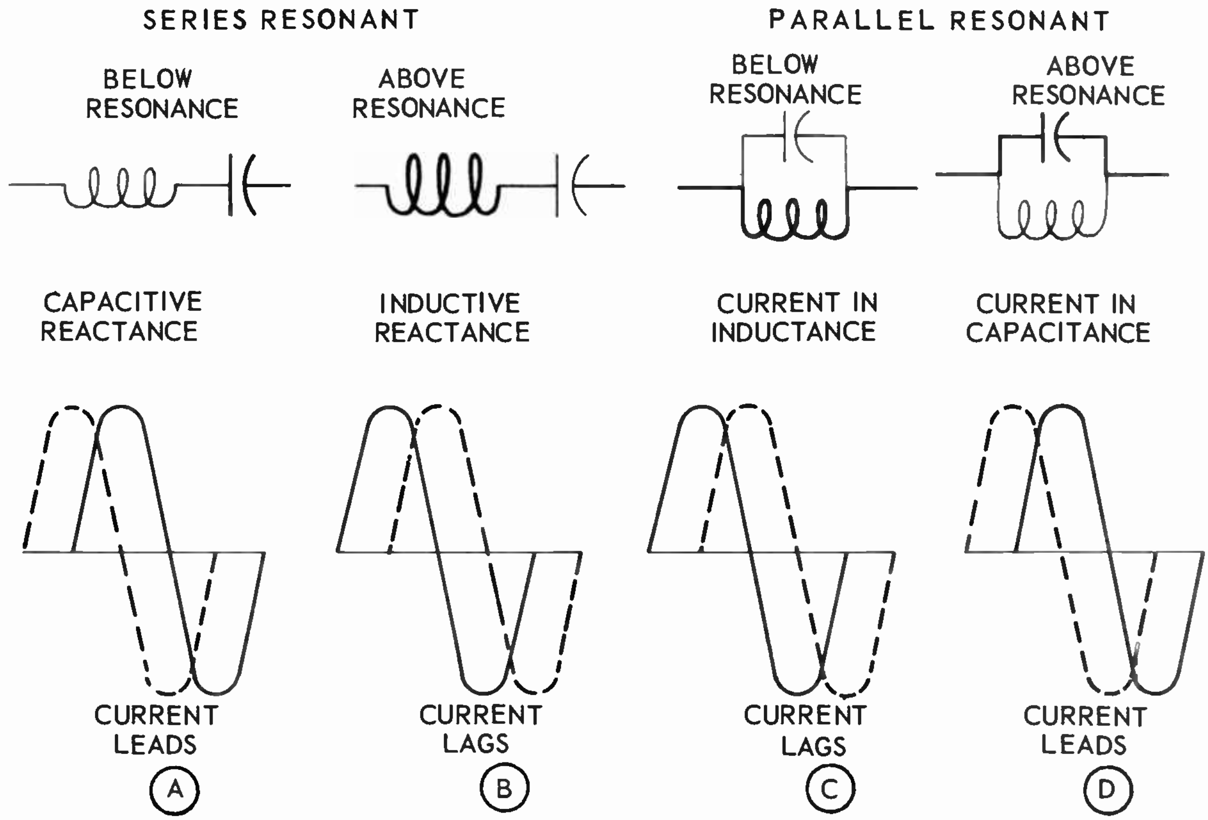


Fig. 28-9. Phase relations at frequencies below and above resonance.

current is zero or approaching zero, and the minimum rate of change when current is maximum and has ceased to increase. Then the current curve for this second quarter-cycle must be shown as going from maximum positive to zero.

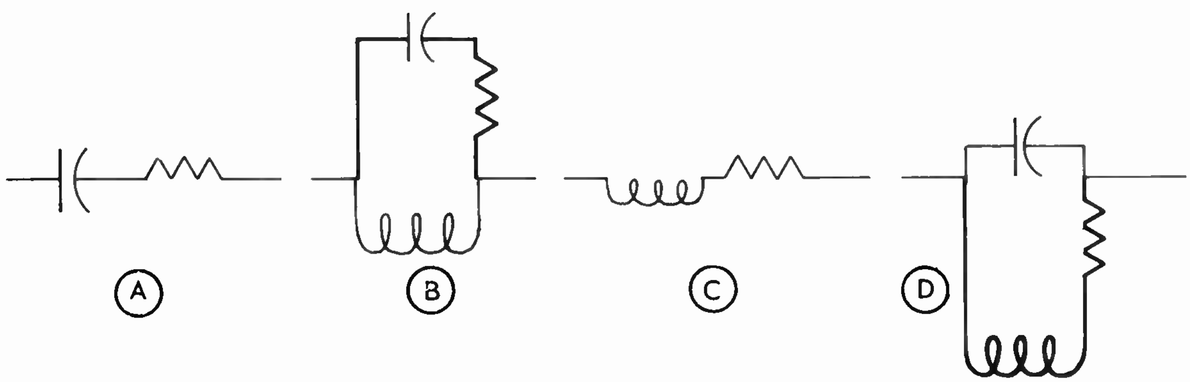


Fig. 28-10. Resistance alters the angle of lag or lead.

6. Putting together the curves for applied voltage and current in the first two quarter-cycles gives these curves for the first half-cycle. We may neglect the counter-emf. Applied voltage has changed from maximum positive through zero to maximum negative, while current has changed from zero to maximum positive and back to zero.

7. Now we begin the second half-cycle with applied voltage and current where we left them at the end of the action in diagrams *5* and *6*. The action is just the same as in diagram *2* except that polarities of applied voltage and counter-emf are reversed. Applied voltage goes from maximum negative to zero, and counter-emf from maximum positive to zero.

8. Here the action is the same as explained for diagram *3* except for reversed polarities. While applied voltage goes from maximum negative to zero, current goes from zero to maximum negative.

9. The same as diagram *4*, except for reversed polarities.

10. The same as diagram *5*, except for reversed polarities. At the end of the actions shown here we are back at the start of the whole thing in diagram *2*. We have completed one cycle of applied voltage and current for the inductance.

At the left in Fig. 28-8 are traced the applied voltage and current during the cycle whose quarters we examined in diagrams *3, 5, 8,* and *10* of the preceding figure. Toward the right are the changes of applied voltage and accompanying current occuring in following cycles. Current in the resistance-less inductance always lags the applied voltage by 90°. Over at the right is traced one cycle of counter-emf. This emf always is in opposite phase to the applied voltage. Incidentally, another way of saying "in opposite phase" is to say "in phase opposition", which means the same thing.

ALTERNATING CURRENT AND VOLTAGE IN RESISTANCE. Supposing we have a circuit in which there is only resistance, with neither capacitance or inductance in any appreciable amount – then what is the phase relation between alternating voltage and accompanying current?

In a pure resistance there is none of the opposing voltage that builds up in a capacitance. Neither is there any of the counter-emf that opposes applied voltage in an inductance. As a result, the current and voltage always remain in phase for any circuit in which there is nothing but resistance or in which capacitance and inductance, or both, are negligible. There is nothing more to the story of the effect of resistance on phase relation of alternating voltage and current, they simply remain in phase.

ALTERNATING CURRENT AND VOLTAGE AT RESONANCE. If a circuit contains such values of capacitance and inductance as are resonant at the applied frequency what is the phase relation of the alternating voltage and current? You can hardly help but know the correct answer to this one. Current and voltage are in phase.

At resonance, as you well know, the reactive effect of the capacitance is exactly balanced by the reactive effect of the inductance. The capacitance tries to make current lead the voltage by 90°. The inductance tries to make current lag the voltage by 90°. The net result is neither lead nor lag, the current stays right in phase with the applied alternating voltage – at resonance.

Here is a question that may not be quite so easy, although previously you have been shown every separate fact that contributes to a correct answer. What is the correct phase relation of current and voltage in a series resonant circuit at frequencies below resonance? The answer is shown at *A* in Fig. 28-9. Here are the facts which combine to give the answer. At frequencies below resonance the capacitive reactance exceeds and cancels the lower inductive reactance, in ohms. Therefore, the source sees the remaining capacitive reactance and the circuit acts as though composed chiefly of capacitance. In a capacitance the current leads the voltage, so that is the answer.

What is the phase relation in a series resonant circuit at frequencies above resonance? Of course, you surmise that the answer is opposite to that for frequencies below resonance, which is correct, as shown at *B*. Inductive reactance exceeds capacitive reactance, the circuit acts as though composed mostly of inductance, and in an inductance the current lags the voltage.

What about the phase relation in a parallel resonant circuit at frequencies below or above resonance? Below resonance, as at *C* in Fig. 28-9, most of the current from the source flows through the inductance

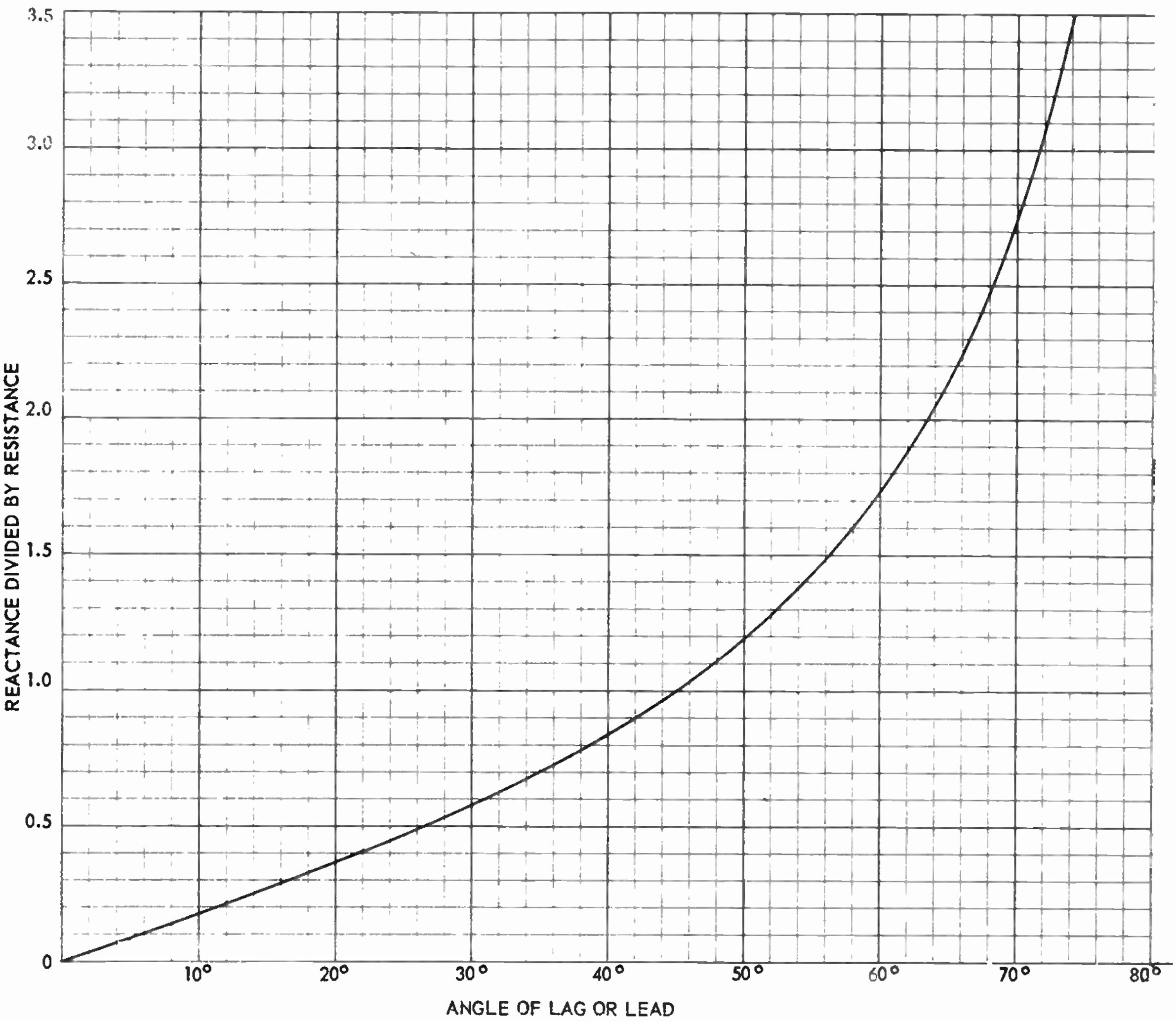


Fig. 28-11. Relations between circuit resistance and the angle of lag or lead.

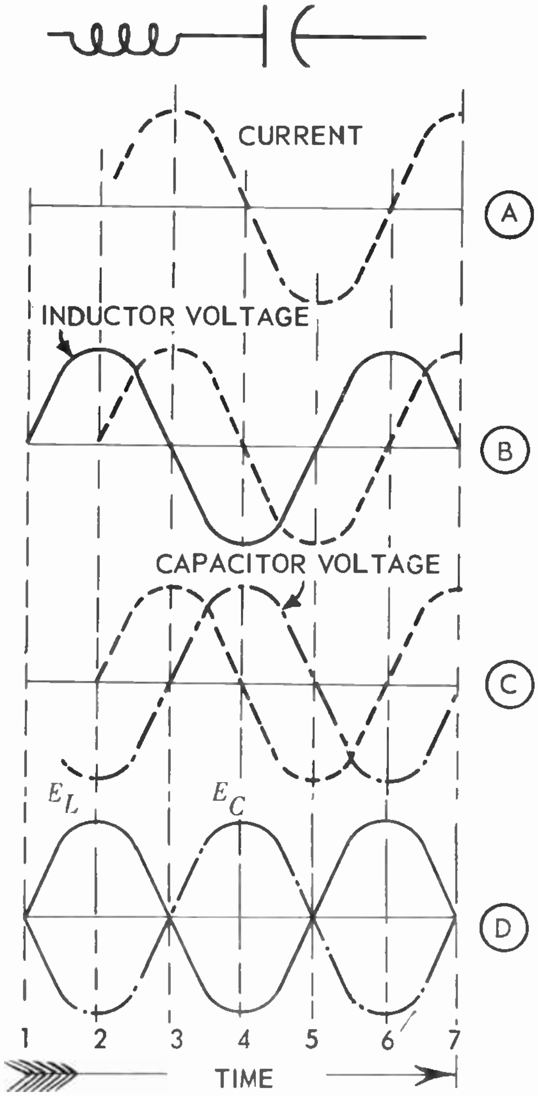


Fig. 28-12. The reasons why voltages cancel at resonance in a series resonant circuit.

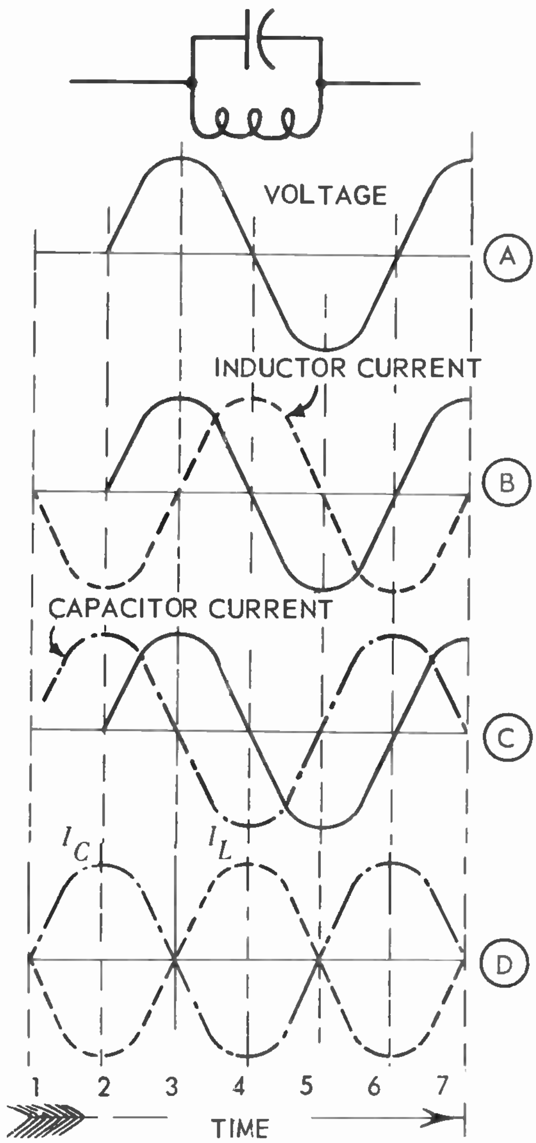


Fig. 28-13. Reasons why currents cancel at resonance in a parallel resonant circuit.

because inductive reactance is less than capacitive reactance at such frequencies. Since this current is supplied by the source, the capacitive effect is negligible and the circuit acts inductive. Current lags voltage in an inductance, and this is the answer for frequencies below resonance. Above resonance most of the current flows in the capacitance, because at the higher frequencies the capacitive reactance is less than the inductive reactance. Then the circuit acts like a capacitance, in which current leads the voltage.

We are explaining these matters of phase relations in resonant circuits so that you can have a good idea of the important part they play in our work with television receivers later on. Remember all you can about these phase relations, and remember too where to find this information when you wish to come back for it.

RESISTANCE EFFECT ON LAG AND LEAD. We have learned that in a pure capacitance the current will lead the voltage by 90°, and that in a pure resistance there is no lead, the current is in phase with the voltage. What will happen if we have both capacitance and resistance in the same circuit, as at *A* or *B* of Fig. 28-10? It is rather obvious that the current will lead the voltage, but by something between zero and 90°.

The actual lead of the current, in electrical degrees, will depend on the relation between capacitance and resistance. If we wish to determine the actual number of degrees of lead it is necessary to figure out the capacitive reactance in ohms, then divide by the resistance in ohms.

Fig. 28-11 allows determining the number of degrees of lead for any ratio of reactance to resistance between zero and 3.5. Supposing the capacitive reactance is 500 ohms and the resistance is 1,000 ohms. Dividing 500 by 1,000 gives 0.5. Following along the line for 0.5 to the curve of Fig. 28-11, thence downward to the scale for angles, shows that the current will lead the voltage by a little more than 26°. Again, supposing capacitive reactance is 5,200 ohms and resistance is 2,000 ohms, which is a ratio of 2.6. The angle of lead for this ratio is shown by the graph to be about 69°. We know the phase difference is a lead of current, not a lag, because we have capacitance, not inductance.

The same method and same graph are used for determining the angle of lag when we have inductance and resistance in the same circuit, as at *C* and *D* of Fig. 28-10. We then divide the inductive reactance by the resistance, both in ohms, and use the ratio to find the angle of lag. There will be lag rather than lead of current because we are dealing with inductance.

The graph of Fig. 28-11 illustrates two things of importance. First, the greater the resistance in comparison with capacitive or inductive reactance the more nearly the current and voltage will be in phase, and the greater the reactance compared with resistance the greater will be the phase difference. The second matter of importance is that the phase difference varies with frequency. It varies with frequency because we use the ratio of reactance to resistance, and reactance varies with frequency.

If you have capacitance and resistance in the same circuit, the capacitive reactance will decrease as frequency increases. Then the ratio of reactance to resistance becomes less, and there is a smaller angle of current lead – as you can see from the graph. Increase of frequency in a circuit having capacitance and resistance will bring the current more nearly into phase with the voltage.

With inductance and resistance in the same circuit, the inductive reactance will increase as frequency rises. Then the ratio of reactance to resistance becomes greater, and there is a larger angle of current lag. When frequency increases in a circuit having inductance and resistance the current will get farther out of phase with the voltage. It seems never to fail that capacitance and inductance have opposite effects.

PHASE EFFECTS IN RESONANT CIRCUITS. The results of almost all tests which we previously made on resonant circuits are explained by the facts that inductance makes current lag the voltage, while capacitance makes the current lead. There is no need to go back over every one of those tests and show how phase relations account for the results, but one or two examples will be instructive.

Being able to follow an explanation based on phase relations will be of help when you work from television service instructions issued by many manufacturers of receivers. In those instructions it is taken for granted that you are familiar with phase relations between voltages and currents as these relations are affected by inductance, capacitance, and resistance. The instructions will tell you why some certain circuit works as it does, and why adjustments are made in certain ways. But you are presumed to understand the ordinary language of television and radio, in which, phase is one of the most familiar words.

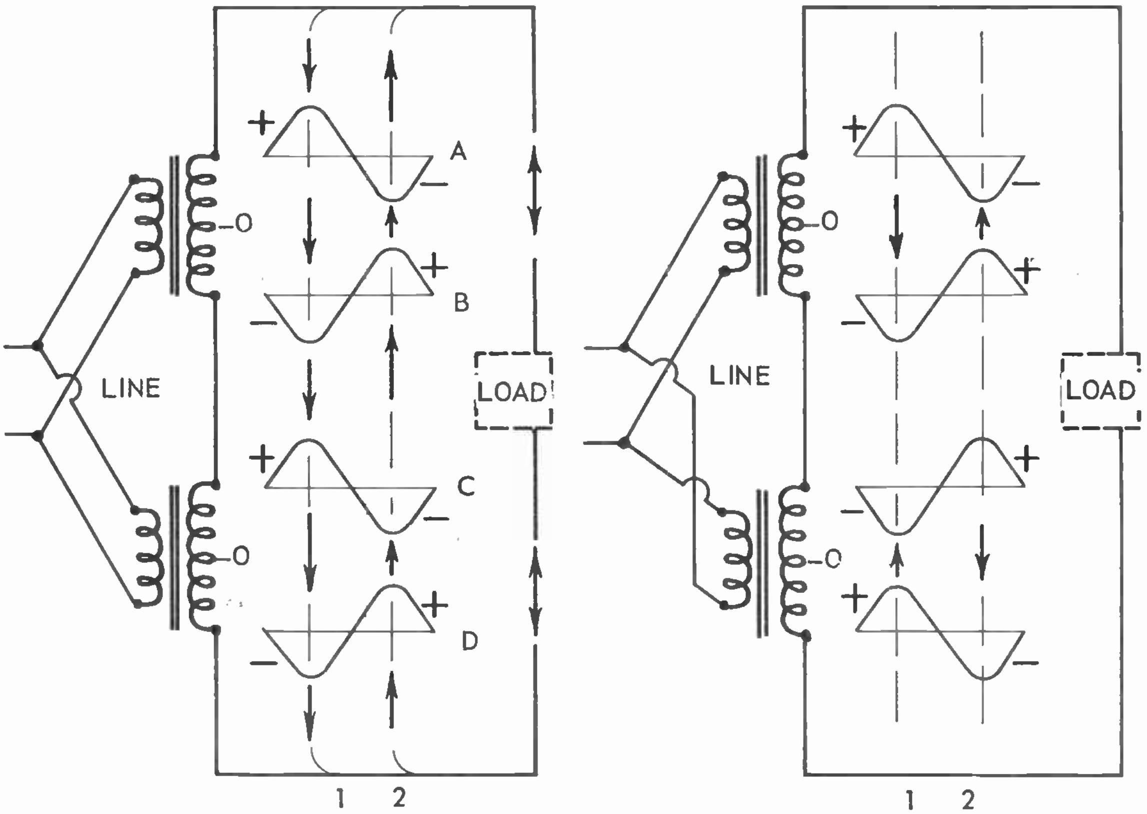


Fig. 28-14. Emf's and electron flows with series transformers connected correctly (left) and incorrectly (right).

For one of our examples we shall determine why voltage is minimum and current maximum in a series resonant circuit at resonance. The explanation depends on the fact that current is the same in both the inductance and the capacitance of any series resonant circuit – it is, in fact, the same current that flows in both parts. Only the voltages across the inductance and capacitance may differ. Therefore, for the series resonant circuit, we may show only variations of voltage phase with reference to a current of fixed phase.

At *A* in Fig. 28-12 is a curve representing the single current that flows in both the inductor and the capacitor. At *B* we have added the curve representing voltage across the inductor. Current lags the voltage, or voltage leads the current by 90°. At *C* is the curve representing voltage across the capacitor, along with the original curve for current. Current leads the capacitor voltage, or capacitor voltage lags the current, by 90°.

At *D* we have the curves for inductor voltage, marked E_L, and for capacitor voltage, marked E_C, by themselves. The phase relation between these two voltages is the same as in the separate preceding diagrams. Vertical lines numbered from *1* to *7* each denote an instant of time which is the same through all the curves. The two voltages are in opposite phase, or 180° out of phase. They also are of equal amplitude. Consequently, the two voltages cancel out to leave zero voltage drop across the series resonant circuit at resonance. If there is no drop of voltage there must be no reactance to oppose the flow of current, and current becomes maximum. This, we know, is just what would happen at resonance if it were possible to construct a circuit where there is no resistance.

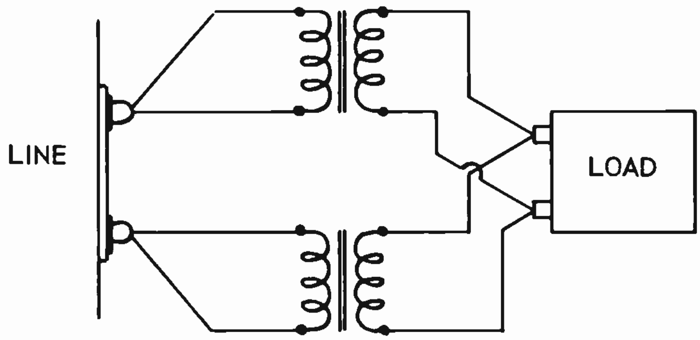


Fig. 28-15. Transformer secondaries connected in parallel on a load.

In any practical series resonant circuit there must be more or less resistance in addition to the react - ances in the inductance and capacitance. When there is resistance, the phase difference between current and voltages cannot be 90° in either one, it must be something less. Then the two voltages do not completely cancel each other, there is remaining net voltage whose value depends on the amount of resistance. Then current must be something less than maximum, for flow of current is being opposed by the remaining resistance. By using such helps as the graph of Fig. 28-11 we could figure out the current and the two voltages for any combination of inductance, capacitance, resistance, and frequency

Now let's see how phase relations affect performance of a parallel resonant circuit. In such a circuit, at resonance, the voltage across the circuit is maximum while current from the source through the resonant circuit is minimum. We must remember, as a basis for the following explanation, that voltages across the inductance and across the capacitance are equal in a parallel resonant circuit. Really there is only one voltage. It exists between the points where the parallel connection is made.

At *A* in Fig. 28-13 is a curve representing the single voltage across the parallel resonant circuit. Accompanying this single voltage we may have, and will have, different currents in the inductance and capacitance. The currents may change in a parallel resonant circuit, but not the voltage.

At *B* is shown the curve for inductor current, along with the original curve for voltage. Inductor current lags the voltage by 90°, or the voltage leads the current by 90°. At C is a curve for capacitor current, in addition to the curve for voltage. Capacitor current leads the voltage, or voltage lags this current, by 90°. At *D* we have the two current curves by themselves. Time relations or phase relations have been preserved, as you will see along the time lines numbered from *1* to *7*.

Capacitor current, identified as I_C, and inductor current, I_L, are equal and opposite. They are in opposite phase, or 180° out of phase. Then the currents cancel each other, and there is no current at all from the source through the parallel resonant circuit at resonance. If there can be no current from the source there must be infinitely great impedance in this circuit. Across an infinite impedance there would be exerted the full voltage of the source. Therefore, we have maximum voltage and minimum current.

Again we have been assuming a circuit in which there is no resistance, only inductance and capacitance. Since there must be some resistance in the inductance and capacitance in any practical circuit, the currents will not be a full 90° out of phase with the voltage in each and they will not be in phase opposition with each other. Some current still will flow from the source. With a current flow there cannot be infinite impedance. So, in a practical parallel resonant circuit, there will be very great impedance, but not infinite, and there will be very little current, but not zero current.

TRANSFORMERS IN SERIES AND IN PARALLEL. Sometimes the secondary windings of two or more transformers are connected in series with one another in order that the sum of their voltages may be applied to a load. Such a connection is shown made correctly at the left in Fig. 28-14, and incorrectly at the right. First we shall look at the correct connections.

The two primary windings are connected to the power line. We may assume that alternating potential differences from the two secondary terminals of the upper transformer are as shown at *A* and *B*, with reference to center tap, and from the secondary terminals of the lower transformer are as at *C* and *D*, with reference to its center tap. Peak voltages during the first half-cycle lie on the vertical line marked *1*, which represents a certain instant of time. Peaks for the second half-cycle occur at the instant indicated by time line number *2*. Directions of electron flows during these two instants are shown by arrows on the respective time lines.

Electron flows within the secondary windings are from positive to negative, which is the direction in which the emf acts inside any kind of source. Between the windings the electron flows are from negative to positive, which is the direction of electron flow in any circuit external to a source. All the directions are downward during instant *1*, which means there will be electron flow through the source in the corresponding direction. Note that emf's in the two windings are in the same direction, so they add their two forces so far as the load circuit is concerned.

During instant *2* there is reversal of the emf's being induced in the secondary windings, and reversal of electron flow in the entire circuit, including the load. Again the two emf's add their forces.

In the right-hand diagram the connections between the lower primary winding and the line have been reversed. This reverses the emf's in the lower secondary winding during both the time instants. These emf's now oppose those induced in the upper winding, where the primary has not been changed on the line. If the emf's or voltages of the two secondaries are equal there will be a balance of forces, and no electron flow in the load circuit nor anywhere else.

How can you tell when the two secondary windings are correctly "phased"? Simply connect an a-c voltmeter of suitable range across the load. Try reversing one primary, but not both. The connection giving higher voltage on the load is correct. Exactly the same things would result from reversing the connections on the two terminals of either secondary. If you reverse the connections to both secondaries, or to both primaries, there will be no change in load voltage because phases have been reversed in both transformers.

Two transformer secondaries sometimes are connected together in parallel, to furnish more load current than either winding could deliver by itself. Such a connection is shown by Fig. 28-15. Phasing is checked by testing the load voltage with an a-c voltmeter and reversing connections to one primary or one secondary winding until the higher of two load voltages is shown by the meter.

In case you should make a parallel connection of two transformers to a single load there are certain precautions which must be observed if the transformers are not to be burned out. Voltage ratings of the two transformers must be alike. Otherwise one of them will pump electrons through the other as well as through the load. The two transformers must have the same turns ratios and same ratio of secondary to primary impedance. They should have the same or nearly the same voltage regulation. It adds up to this: paralleled transformers should be identically alike.

When connecting two transformer secondaries in series it is necessary to make sure the current output won't exceed the current rating of either unit, and to be certain that the higher voltage won't exceed the ability of insulation to withstand it in either transformer.

Now we shall leave the subject of phase for a while. But don't forget, whenever you work with two or more alternating voltages or currents the results of your work will depend to a great extent on the phase relations which may be involved.

TELEVISION

LESSON NO. 29

CAPACITORS AND INDUCTORS FOR TUNING

Training Your Mind...

C.T.I.

Perfecting Your Skill

DO NOT TEAR — CUT ALONG THIS LINE AND SEND IN FOR GRADING.

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

COMMERCIAL
TRADES
INSTITUTE

Training Your Mind
C. T. I.
Perfecting Your Skill

HIGH SCHOOL DIVISION Chicago, Ill.

LESSON NO. 29

CAPACITORS AND INDUCTORS FOR TUNING

In receivers for television and also in those for sound alone we find these major divisions or sections. In a television receiver there are all ten sections. In a sound receiver there are five of them.

Name of Section	Kind of Receiver In Which Used	
Radio-frequency, or tuner	* Television	* Sound
Video i-f amplifier	* Television	
Sound i-f amplifier	* Television	* Sound
Video detector	* Television	
Sound detector	* Television	* Sound
Video amplifier	Television	
Audio-frequency amplifier	Television	Sound
Synchronizing section	Television	
Sweep section	* Television	
Power supply	Television	Sound

Six of the television receiver sections are marked with stars, and three of the sound receiver sections are similarly marked. In each of these starred sections there are from one to five adjustable capacitors or adjustable inductors which tune the circuits to resonance. Some of these adjustments are used only during servicing. Others are continually used by the operator. There are adjustable or variable tuning adjustments of one kind or another in sixty per cent of the major divisions of every receiver. Without any doubt, tuning is of real importance.

In every tuned circuit there are three electrical elements; capacitance, inductance, and resistance. Capacitance acts electrically in almost exactly the same way as the spring in Fig. 29-1 acts mechanically. The spring is supporting a weight, which acts mechanically in the same way that inductance acts electrically. Resistance to flexing of molecules in the spring, and to movement of the weight through air, acts like electrical resistance in a circuit.

Over toward the right are a tuning capacitor, a tuning inductor or coil, and a resistor. The whole picture shows a mechanically resonant circuit. It would be interesting for you to fix up an equivalent mechanically resonant circuit by hanging any kind of weight on the end of a rubber band.

Pull down on the weight, then let it go. The weight bobs up and down as the spring, or rubber band, contracts and expands. Energy is passing back and forth between the weight and the spring. When the spring is stretched, all the energy of the system is momentarily stored in the spring, and this energy will be used to

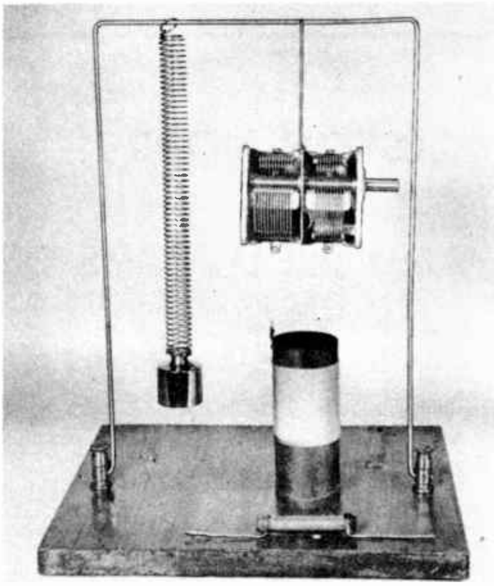


Fig. 29-1. A mechanically resonant system and the parts of an electrically resonant system.

raise the weight. When the weight is at its highest point all the energy is in the weight, because of its elevated position. This energy will be used to stretch the spring when the weight falls downward. The weight and the spring will come to rest when all the energy has been used up on overcoming mechanical resistance.

When the capacitor is charged, all the electrical energy is in the electric field between plates of the capacitor. As the capacitor discharges through the coil, the energy is carried by moving electrons from the capacitor over to the coil, and there the energy reappears as a magnetic field around the coil. Then the magnetic field collapses, induces emf and current in the coil, and energy goes back to the capacitor where it builds up another electric field. Energy finally will stop oscillating back and forth between the two fields when all of it has been used up in overcoming circuit resistance and other losses.

If you use a longer spring, or more capacitance, the movements recur less rapidly. You have lowered the resonant frequency. Using a heavier weight, or more inductance, will have the same effect of lowering the resonant frequency. If you use a shorter spring, or a shorter rubber band, the up and down swings will occur more rapidly; you have increased the frequency. The same thing happens with less capacitance.

Now set the weight to bobbing up and down. Every time the weight is all the way down, and starting back upward, give it an upward tap with the end of a pencil. Get your pencil taps in time or in phase with natural movement of the weight. It takes only very light taps to keep the oscillations going indefinitely, and with slightly stronger taps you can greatly increase the amplitude of the oscillations.

You and the pencil are the applied frequency, or represent signal energy at the applied frequency. If the frequency of your tapping is exactly the same as the natural resonant frequency of weight and spring, the oscillations are maintained or increased in amplitude. But change the rate of tapping to some different frequency. Now you are out of tune, out of phase, or out of resonance with the spring and weight. Oscillations

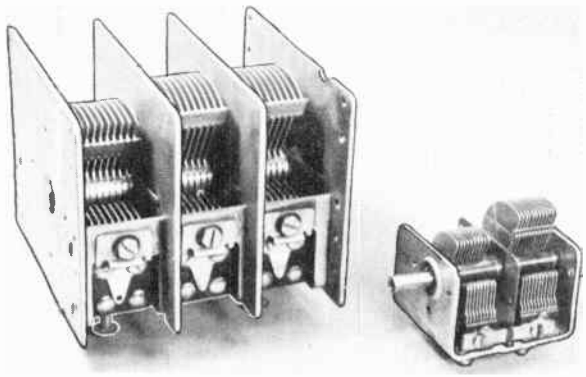


Fig. 29-2. A three-gang tuning capacitor and a two-gang unit.

are reduced or stopped. If your tapping is energetic enough you can force the mechanical system to work at a frequency different from that at which it is naturally resonant.

Exactly similar things happen, electrically, when an applied voltage is at a frequency which is the same as or different from the frequency at which a capacitor and inductor are resonant.

You may experiment with resistance by fixing a strip of cardboard so that the weight rubs as it moves up and down. Friction between weight and cardboard is like electrical resistance, which is due to friction between electrons and atoms. The greater the friction (resistance) the less becomes the amplitude of motion (or of voltage and current).

Our mechanical model shows the effects of capacitance (resilience), of inductance (weight, mass, or inertia), of resistance (friction), and of frequency. And this is the story of resonance.

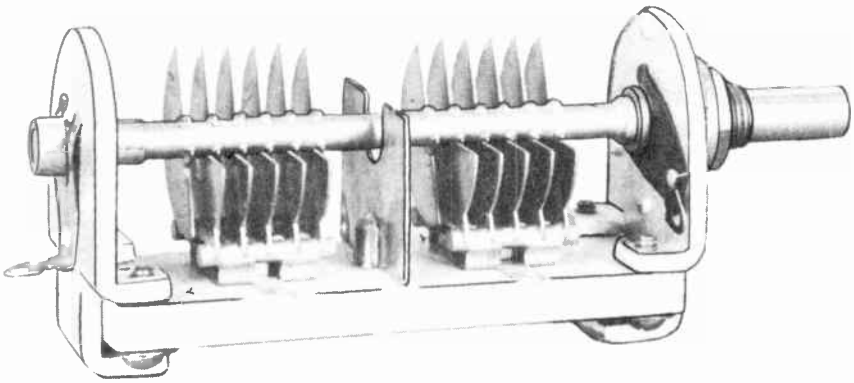


Fig. 29-3. Dual type variable capacitor. Capacitance in each section is maximum at 15 mmfd.

VARIABLE CAPACITORS WITH AIR DIELECTRIC. In earlier lessons we have looked at a number of variable tuning capacitors having air for the dielectric between plates. This general type of construction is illustrated by the two units of Fig. 29-2. The movable plates, rotor plates, are mounted on a shaft which is turned by the tuning dial or knob to bring greater or smaller areas of the rotor plates and stationary (stator) plates opposite one another with the air dielectric between them. Thus the effective capacitance is varied.

The unit at the left in the picture is a 3-gang capacitor with three sets of rotors and three sets of stators which allow simultaneous tuning of three different circuits as the three rotor groups are moved together on the single shaft. At the right is a 2-gang capacitor with two sets of each kind of plates, for simultaneous tuning of two different circuits.

Both kinds of plates are made of sheet brass, aluminum, or rust-proofed steel. The rotor plates are welded, brazes, soldered, or securely clamped on the shaft. The stators are similarly fastened to metallic end supports which electrically join all the plates into a single unit. The stator unit or units are supported on the capacitor frame by one or more pieces of solid insulating material. Fig. 29-3 is a picture of a small-capacitance 2-gang or dual variable capacitor in which the two stator groups are carried by a long piece of steatite insulation extending between the two end frames.

In a well designed capacitor the insulating supports are not too close to the edges of the plates nor to the outer ends of the groups of plates. At these points the electric field extends for some distance beyond the plates, and if the field passes through much insulation we have the effect of a partial solid dielectric in which energy losses are greater than in air. There should be a reasonably long distance along the insulation surface between nearest points of stator and rotor elements, in order to prevent electrical leakage. Insulation preferably is of a material having low dielectric constant.

2

The stators of any tuning capacitor should be connected to the grid side or plate side of the tuned circuits. This usually is called the high side of the circuit. The rotor side of the capacitor is connected to chassis ground, to B-minus, or any point not leading directly to grids or plates. Then the rotor shaft, which may be closely approached by the operator's hand, is on the relatively insensitive side of the tuned circuit, and tuning will be little affected.

9

In every capacitor there is at least some small loss or waste of signal energy. For example, at very high frequencies most of the current flowing into and out of the capacitor tends to travel only on the surfaces rather than being distributed uniformly through the metal. This is called skin effect, because current travels in the "skin" of the metal. The result is increased resistance, because all the current flows in only a limited cross sectional area of conductor. Also, if magnetic fields cut through the capacitor plates there is induction of "eddy currents" which circulate in the metal and cause energy loss due to resistance of the metal opposing flow of these currents.

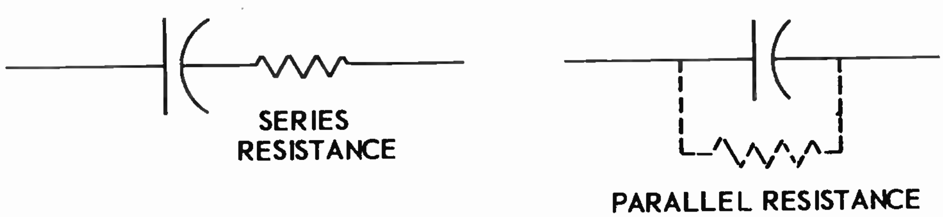


Fig. 29-4. Energy losses in capacitors may be represented as equivalent resistances.

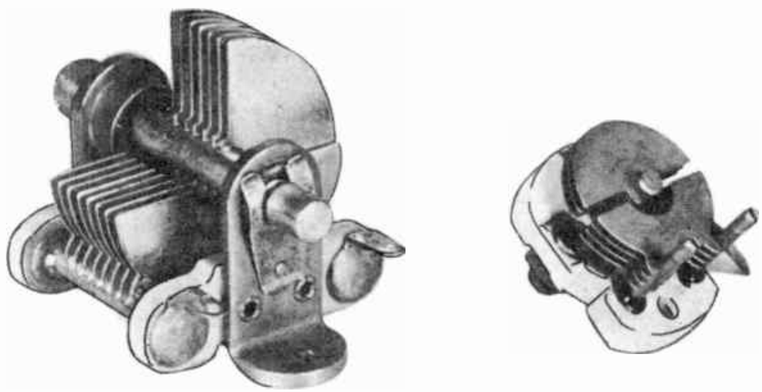


Fig. 29-5. Small variable capacitors having semi-circular plates.

The two losses mentioned, and any others which would or could cause heating, have the same effect as resistance in series with the capacitor, as represented at the left in Fig. 29-4. Their effect is called equivalent series resistance, of a number of ohms which would cause the same loss were the capacitor itself electrically perfect.

Leakage across the insulation, and other losses which allow current to flow right through the capacitor instead of charging it, act like a resistance in parallel with the capacitor, as at the right in Fig. 29-4. The effect is called equivalent parallel resistance, and it too may be measured in ohms.

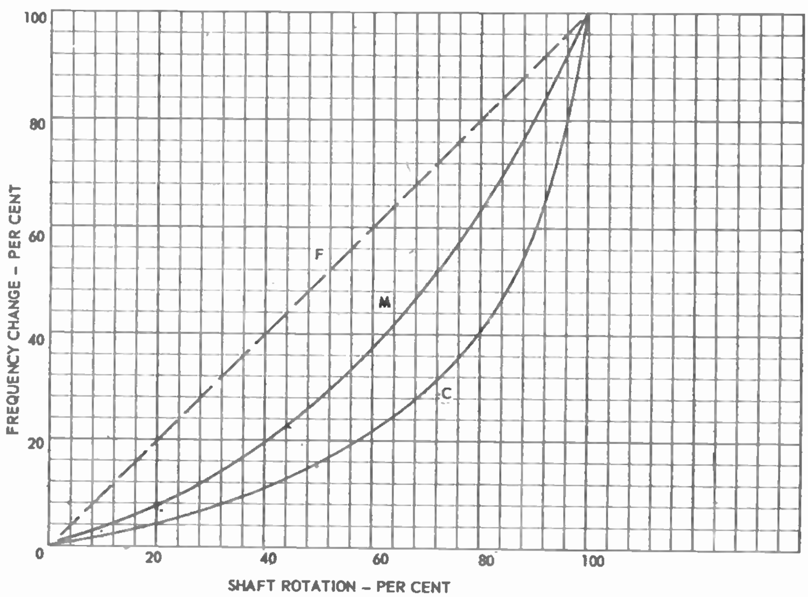


Fig. 29-6. Relations between shaft rotation and change of tuned frequency with capacitor plates of three shapes.

The variation between minimum and maximum capacitance of a tuning capacitor has to be much greater than the desired change between highest and lowest tuned frequencies. This is explained by the formula for tuning capacitance, of which one form is as follows.

$$\text{Micro-microfarads} = \frac{25\,330}{\text{megacycles}^2 \times \text{microhenrys inductance}}$$

If you assume any two frequencies which are to be tuned by use of the same inductance, and figure out the required capacitances, you will find that the change of capacitance must be equal to the square of the change of frequency. For instance, a 3-to-1 change of frequency calls for a 9-to-1 change of capacitance. The square of 3 is 9.

Actually the variation of tuning capacitance must be even greater than the square of the change in frequency. This comes about because there is a very considerable fixed capacitance, consisting of distributed capacitance in inductors, stray capacitance of wiring, and internal capacitances of tubes.

Supposing you have a tuning capacitor variable from 30 to 300 mmfd. This is a change of 10 to 1. But the total fixed capacitance maybe 30 mmfd. Then the actual minimum capacitance is the sum of the fixed capacitance and the capacitor minimum, which comes to 60 mmfd. Maximum capacitance is the sum of fixed capacitance and maximum capacitance of the capacitor, a total of 330 mmfd. Now the total variation, 60 to 330 mmfd, is only 5½ to 1. Without any fixed capacitance the variable capacitor would have tuned a frequency range of 3.16 to 1 in megacycles or kilocycles. Adding the fixed capacitance brings the tuning range down to 2.35 to 1. Watch this when you buy tuning capacitors for replacement or when designing a circuit, get as much capacitance variation as is possible.

The less the maximum capacitance of a tuning capacitor the greater is its minimum capacitance as a fraction of the maximum value. Around 300 mmfd maximum you can get minimums as low as 13 to 15 mmfd, a ratio of about 20 to 1. With a maximum of 30 mmfd the minimum hardly ever will be less than 6 mmfd, a ratio of 5 to 1.

When tuning with a variable capacitor, the manner in which the resonant frequency is changed in relation to rotation of the capacitor shaft depends on the shapes of stator and rotor plates. In the two capacitors of Fig. 29-5 the stators and rotors are semi-circular, they are the shape of half-circles. Capacitors with such plates may be called straight line capacitance types, because their capacitance changes uniformly with rotation of the shaft.

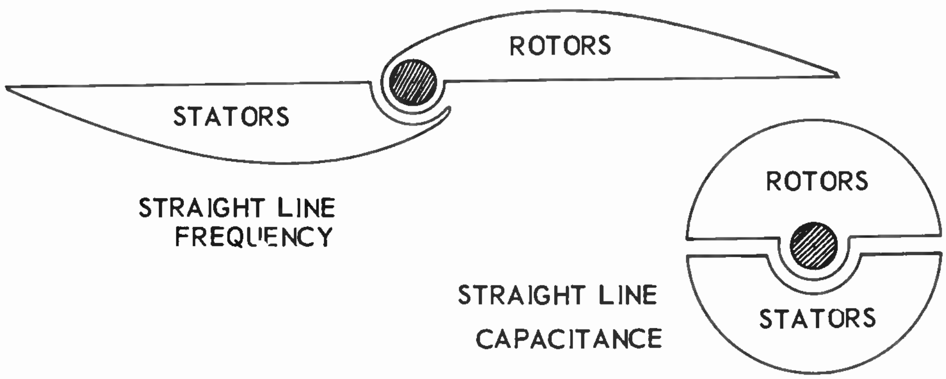


Fig. 29-7. Straight line frequency plates would occupy much space.

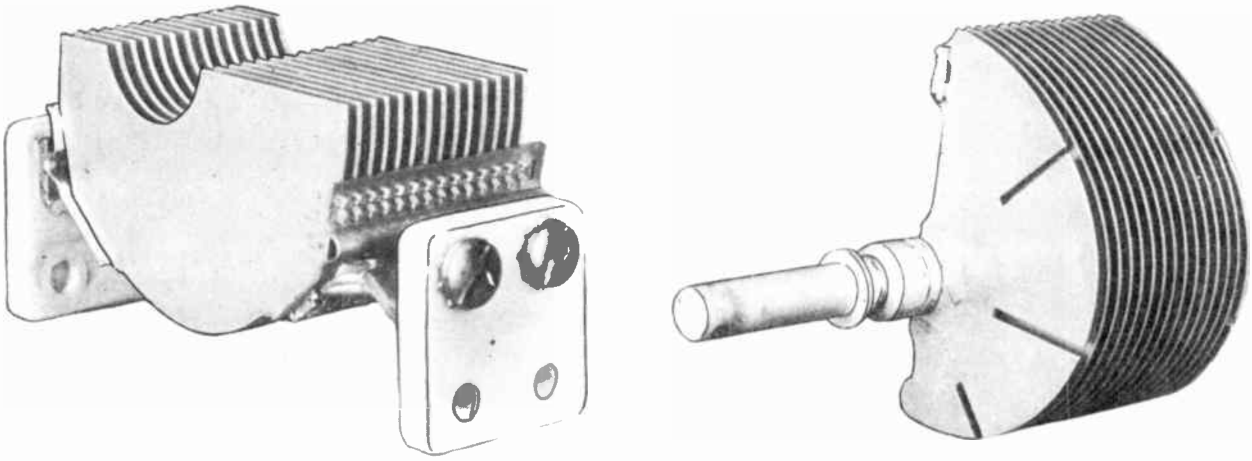


Fig. 29-8. Stator plates and rotor plates of a modified straight line frequency tuning capacitor.

When a capacitor of the straight line capacitance type is used for tuning, the relation between shaft rotation and change of frequency is as shown by the curve marked *C* in Fig. 29-6. With the rotor shaft turned through the first 50 per cent of its travel the frequency has changed by only 16 per cent. With the remaining 50 per cent of shaft rotation the frequency changes by 84 per cent of its total change. Broadcast frequencies would be far apart on the tuning dial during the first part of the rotation, and would be jammed closely together during the final part. Capacitors having plates of this type are used in testing and measuring instruments, and for auxiliary tuning in some receivers, but not in ordinary broadcast receivers.

The ideal broadcast receiver capacitor would cause spacing of frequencies uniformly all across the tuning dial. The relation between shaft rotation and frequency would be as shown by the straight line marked *F* in Fig. 29-6. A capacitor of this kind would be called a straight line frequency type. Its plates would have to be of the general proportions shown at the top in Fig. 29-7, as compared with the straight line capacitance plates shown down below. The straight line frequency capacitor would take up too much space to be practical in modern receivers.

4 Most of the broadcast tuning capacitors are of a modified straight line frequency type which provide a relation between shaft rotation and frequency change about as shown by the curve marked *M* in Fig. 29-6. Frequencies are somewhat spread out at the lower end of the tuning range, are quite uniformly spaced through most of the rotation, and are slightly crowded at the high-frequency end of the range.

At the left in Fig. 29-8 is a group of stator plates from a modified straight line frequency capacitor. The end brackets which fasten the plates together are supported by insulating blocks of steatite which mount on the frame. At the right are shown the rotor plates from the same capacitor. You will note that the end plates of the rotor are slotted. This allows bending of these end plates toward or away from the stators to slightly increase or decrease the capacitance at some points in the rotation of the shaft. This adjustment is used by service technicians for aligning or "tracking" of two or more circuits tuned simultaneously by the capacitor. We shall get acquainted with such adjustments a little later.

1 *TRIMMER CAPACITORS.* A trimmer capacitor is a variable capacitor, usually of rather small capacitance, used with a variable tuning capacitor or with a tuning inductor for making service adjustments of total circuit capacitance. Mounted on the supports for the stator plates of the capacitors in Fig. 29-2 you can see

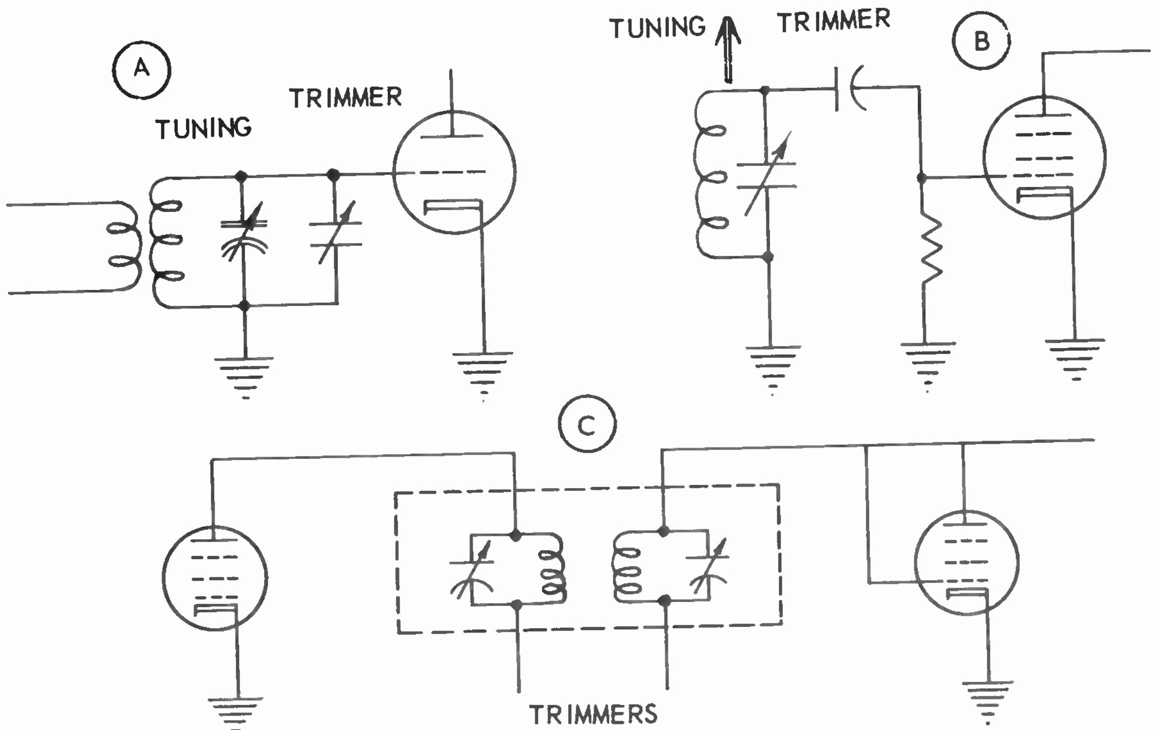


Fig. 29-9. How trimmer capacitors may be shown in circuit diagrams.

trimmer capacitors with their screw adjustments. The trimmer for each gang or section is permanently connected in parallel with the main tuning unit. One side of the trimmer is connected to the stator plates. The other side is connected to the frame of the capacitor, and through the frame and shaft to the rotor plates.

A trimmer for one tuning capacitor or one section of a tuning capacitor may be shown in a circuit diagram as at *A* in Fig. 29-9. Unfortunately there is no special symbol for indicating a trimmer capacitor as distin-

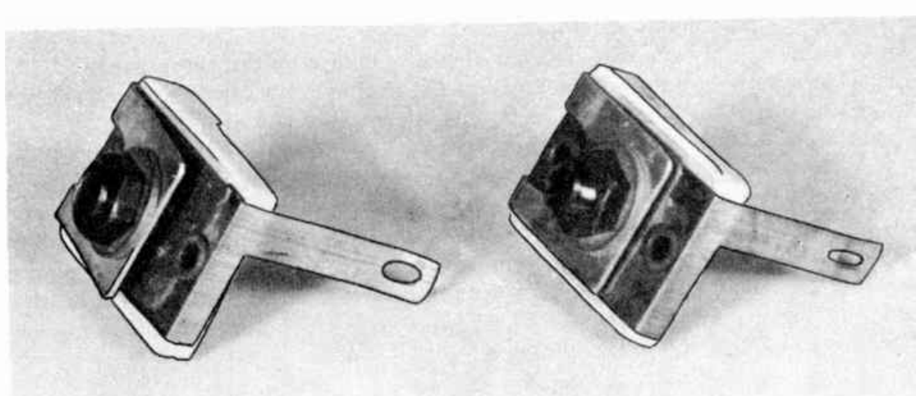


Fig. 29-10. Mica-dielectric trimmer capacitors of the compression type.

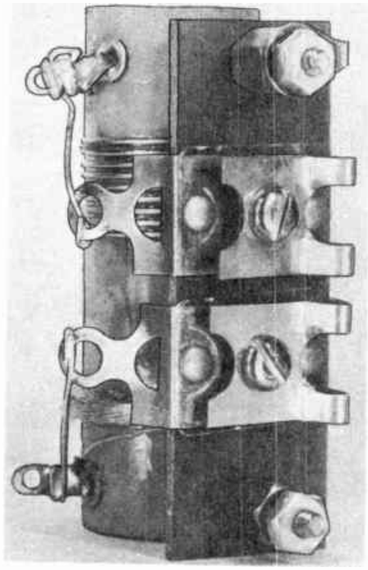


Fig. 29-11. A mica-dielectric trimmer capacitor is connected to each coil winding.

guished from a main tuning capacitor. At *B* is shown a trimmer capacitor in parallel with an inductor or coil that is tuned by means of a movable core. At *C* there are trimmers in parallel with each of the windings of a coupling transformer.

Trimmers that are built onto variable tuning capacitors usually have a thin sheet of mica for the dielectric. The movable plate of the trimmer is moved toward or away from its stationary plate by means of a screw. Mica-dielectric trimmer capacitors which may be mounted by themselves are pictured in Fig. 29-10. At the left the plates are widely separated, for minimum capacitance, and at the right the plates have been forced close together for maximum capacitance.

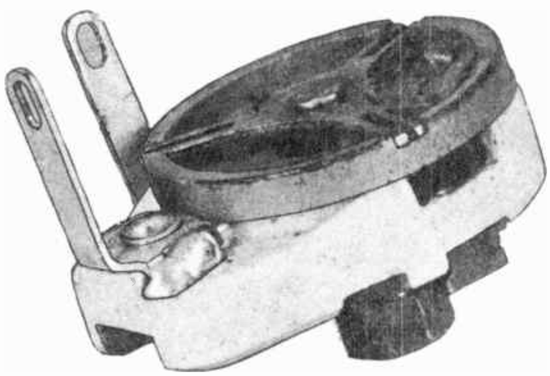


Fig. 29-12. Ceramic trimmer capacitor. The rotating member is on top, the adjusting screw is below.

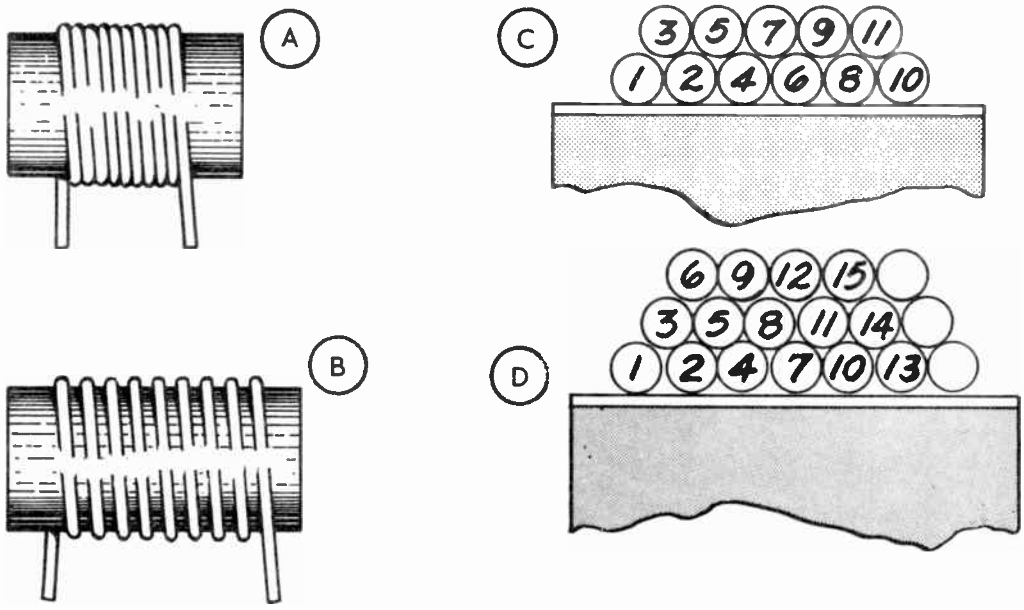


Fig. 29-13. Single-layer and multi-layer coil windings.

These compression type mica-dielectric trimmers are made with anywhere between two and eight or more plates, half for one side, the half for the other side of the capacitor with sheets of mica between every plate and those on either side. The adjustment range of capacitance may be something like 2 to 30 mmfd, or it may be as great as from 100 to 500 mmfd.

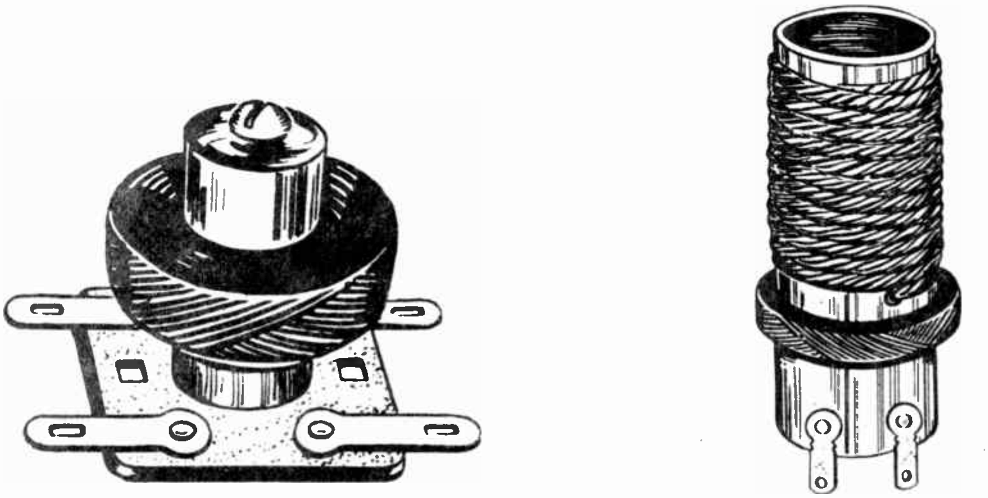


Fig. 29-14. A duolateral winding (left) and an r-f transformer having duolateral primary and bank-wound secondary (right).

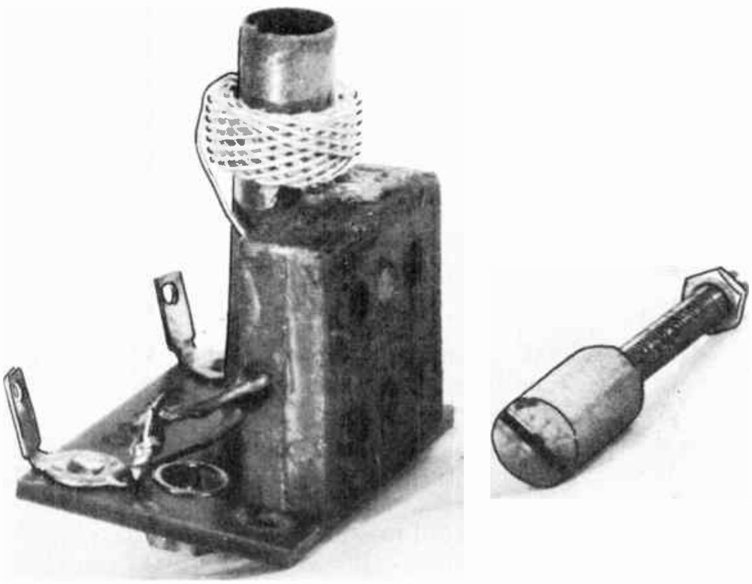


Fig. 29-15. This inductance-tuned trap has a honeycomb coil. The coil core has been removed and is lying at the right.

Trimmers sometimes are mounted on the coil forms or supports instead of on the main tuning capacitors, as shown by Fig. 29-11. Here there is one trimmer for each of the two coils that are wound on the one form. This coil and trimmer unit might be used with the two-gang variable tuning capacitor at the right in Fig. 29-2, thus providing two resonant circuits simultaneously tuned, but with separately adjustable trimmers.

A trimmer capacitor more commonly used in high-frequency circuits is the ceramic type, of which one example is illustrated by Fig. 29-12. The name ceramic refers to various porcelain-like materials. Steatite often is used. With one construction the ceramic base is perfectly flat with a thin coating of silver acting as one of the plates. On the other side of the flat base is a flat piece of metal which is rotated by the adjustment screw or nut. This is the second plate. Both plates are of semi-circular shape so that rotation through 180° changes the capacitance from minimum to maximum. Ceramic trimmers may have capacitance ranges as small as from 2 to 6 mmfd or as great as from 20 to 125 mmfd or even more.

Still another kind of trimmer is called a tubular type. This style of trimmer looks much like the small inductor with a movable core that we used for making tests of resonance. The dielectric is the tubular or cylindrical form. One plate is a coating of metal on the outside of the form. The other plate is a solid or tubular piece of metal that fits inside the form and is moved lengthwise by a screw adjustment. Capacitance is increased by turning the inner plate farther into the form, so greater areas of the two plates are opposite each other with the dielectric between them. Capacitance ranges are small, usually from one mmfd or less up to about 6 or 8 mmfd between minimum and maximum.

Small variable capacitors of the air-dielectric type sometimes are used as trimmers, especially for work at very high frequencies and in some testing instruments. These trimmers look like the units pictured in Fig. 29-5. The shafts usually have screw driver slots for adjustment, and are held in position by lock nuts.

INDUCTORS OR COILS FOR TUNING. An inductor or coil which is satisfactory for use in high-frequency circuits will meet the following requirements. It will be of the smallest physical size that allows the necessary inductance. There will be only a small magnetic field extending around the outside of the coil. Distributed capacitance and other causes of high-frequency energy loss will be small. The construction will be durable enough to withstand ordinary handling.

Simplest of all coil windings is the single-layer close wound type shown at *A* in Fig. 29-13. The adjacent or successive turns are laid as close together as permitted by the size of the wire. There is considerable distributed capacitance, and the coil becomes rather long or of large diameter when any great amount of inductance is provided.

At *B* is a single-layer spaced winding. Adjacent turns are separated from each other by a space of about half the wire diameter or somewhat more. There is a very worth while reduction of distributed capacitance for a given inductance. However, to provide the same indnctance as in a close wound coil the spaced type must have more turns and will be much longer or of greater diameter.

The greater the number of coil turns in any given length and diameter the greater is the inductance of any type of coil. Inductance increases as the square of the number of turns. Therefore, if you put twice the number of turns in the same cross sectional area you will have four times as much inductance. Of course, the distributed capacitance will be increased by having more turns close together.

One way of providing many turns in small space with least distributed capacitance is to use a two-layer bank winding made up as shown at *C* in Fig. 29-13. The turns are wound on in the order shown by the numbers. After winding turns number 1 and 2 side by side, turn 3 is laid on top of them. Then turn 4 is laid alongside number 2, and turn 5 goes on top. Thus the winding is carried on. For still more turns in a given length we may use a three-layer bank winding, as shown at *D*. The order in which the turns are laid on is indicated by the numbers. A multi-layer coil would by completing one layer before starting the next layer on the outside would have so much distributed capacitance as to be of little use at radio frequencies.

Another multi-layer winding that has reasonably small distributed capacitance is the duolateral type shown at the left in Fig. 29-14. The winding is laid on with groups of side-by-side turns successive groups cross one another at an angle. At the right is a coupling transformer for standard broadcast frequencies in which the lower coil (the primary winding) is a duolateral type and in which the upper coil (the secondary) is bank wound.

5 In Fig. 29-15, at the top of the tubular form, is a coil made with a honeycomb winding. Each layer consists of spaced turns. The turns of each layer cross all the turns of layers inside and outside at an angle. For a given inductance a honeycomb winding will take up more space, (diameter, length, or both) than will either a bank winding or a duolateral winding. But the honeycomb type has less distributed capacitance than any other multi-layer type with the same inductance. The entire unit pictured by Fig. 29-15 forms a trap, a parallel resonant circuit that is resonant at some certain undesired frequency and which decreases the gain or amplification at this one frequency.

WIRE FOR COIL WINDING. Wire for close wound coils having the small inductances required in high-frequency circuits usually is insulated with enamel or with some type of plastic coating, and sometimes with silk or rayon. Wire of any diameter that provides sufficient mechanical strength is amply large for the small currents in high-frequency circuits of receivers. The smaller the wire, or the smaller the conductor inside the insulation, the less will be the distributed capacitance.

Space wound high-frequency coils usually are made with uninsulated wire, but the copper wire then is coated with silver or cadmium, or is tinned, to prevent oxidation of the copper. Solid silver wire sometimes is used, but it has little advantage over silver coated copper wire.

One of the causes for waste of signal energy in coils is skin effect, which was mentioned in connection with capacitors. Skin effect, in a coil, forces high-frequency currents to travel almost entirely at and near the outer surface of the wire. The inside might as well not be there. In some high-frequency transmitter circuits the inductors are made of copper tubing, thus providing what amounts to only the outside of the conductor. Silver coated wire reduces skin effect losses because the highly conductive silver is out where there is maximum current density.

7 The wire for windings used at standard broadcast carrier frequencies often is of a type called Litz wire. This name is an abbreviation of the word Litzendraht. Litz wire consists of from 5 to 15 strands, each separately coated or insulated with enamel, and all braided together so that every strand comes to the surface over the same length of wire as every other strand. The use of Litz gives a decided reduction of skin effect at frequencies between about 300 kilocycles and 3 megacycles. At frequencies much higher the Litz is poorer than solid wire. At lower frequencies the skin effect is slight, so the use of the more costly Litz is not justified.

The accompanying table of *Wires Sizes and Turns* gives the diameter in fractions of an inch of various sizes of solid copper wire commonly used for coil windings. Included also are numbers of turns per running inch of close wound coils made with several types of wire. The abbreviation "SSC" means single silk covered, and "SSE" means single silk enameled, for a wire which is enameled and then covered with a single layer of silk. These latter two columns would apply generally when rayon is used instead of silk. The number of turns per inch won't always be exactly as listed, because different manufacturers may use coverings of slightly different thicknesses.

COIL FORMS AND SUPPORTS. The tubular or cylindrical forms on which coils are wound may be of hard fibre, of wax impregnated paper or cardboard, of wax impregnated wood, or of phenolic compounds such as various grades of Bakelite. All these are used for coils operating at frequencies through the standard broadcast carriers and sometimes for higher frequencies as well.

Gage No.	Diam, bare	Turns per Inch Bare	Enamel	SSC	SSE
14	0.0641	15.5	15.1		
16	.0508	19.7	18.9	18.6	18.0
18	.0403	24.8	23.7	23.2	22.3
20	0.0320	31.2	29.4	29.0	27.6
22	.0253	39.5	37.0	36.0	34.1
24	.0201	49.7	46.5	44.5	42.0
26	0.0159	62.9	58.2	55.0	51.5
28	.0126	79.4	73.1	67.2	62.5
30	.0100	100	91.5	82.2	76.0
32	0.0080	125	115	98.4	91.0
34	.0063	159	143	118	109
36	.0050	200	178	167	129

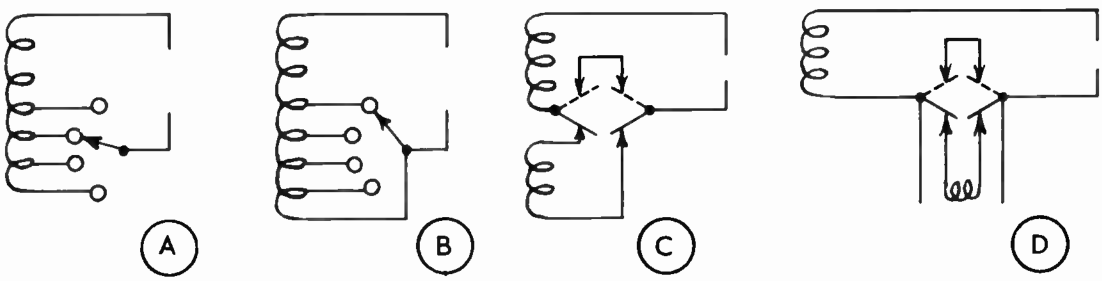


Fig. 29-16. Methods of coil switching used for varying the total or effective inductance.

Special grades of wax impregnated paper forms are in general use for the very high frequencies. In the very-high frequency bands it is also common practice to use forms and supports of the low-loss dielectrics such as polystyrene and steatite. The least loss of waste or signal energy is attained by using self-supporting coil windings made of wire that is rigid enough to need no form. Self-supporting coils are mounted by soldering their ends to any terminals through which circuits are completed. A self-supporting coil of bare wire or of wire without insulating covering has the lowest high-frequency losses of any type which may be used.

When a coil is of such length, diameter, or wire size that it cannot be made self-supporting, the losses may be made almost as low by stiffening the winding with long, narrow strips of polystyrene to which the turns are cemented. Such a coil may be supported by the strips or at the ends of the wire. The winding may be space wound, since the turns are held securely in position by the cement. Another low-loss type of support consists of a cylindrical or tubular form on the outside of which are lengthwise ridges. The winding rests only on the outermost edges of the ridges.

The smaller the quantity of insulation which is on or close to the wire conductor of a coil the lower will be the energy losses due to the action called dielectric absorption, and to the action whereby atoms in the insulating dielectric are strained one way and the other by the alternating fields. Although in any ordinary commercial design there will be quite a bit of metal close to the inductors, it still is a fact that large energy losses occur with such construction. When any metal is located in the field of a coil the magnetic lines cut through the metal and induce eddy currents. These are small electric currents that whirl around and around within the body of metal where the magnetic lines are passing through. It requires energy for these eddy currents to overcome the resistance of the metal, and this energy must be taken from signal energy in the coils.

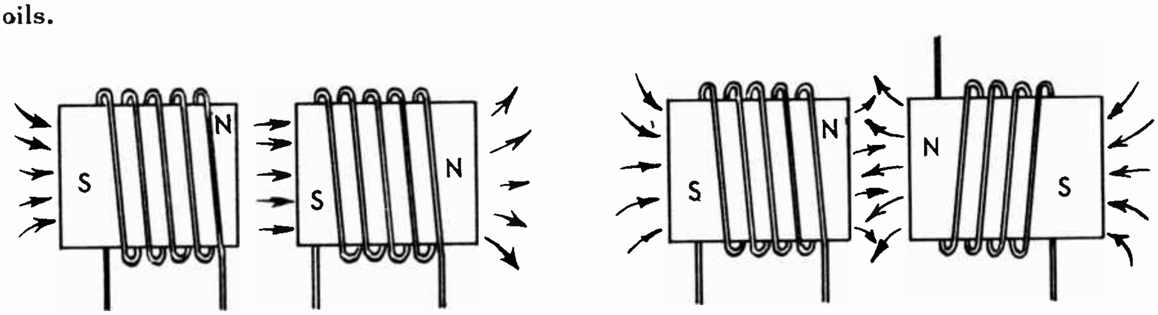


Fig. 29-17. Mutual inductance is positive at the left, is negative at the right.

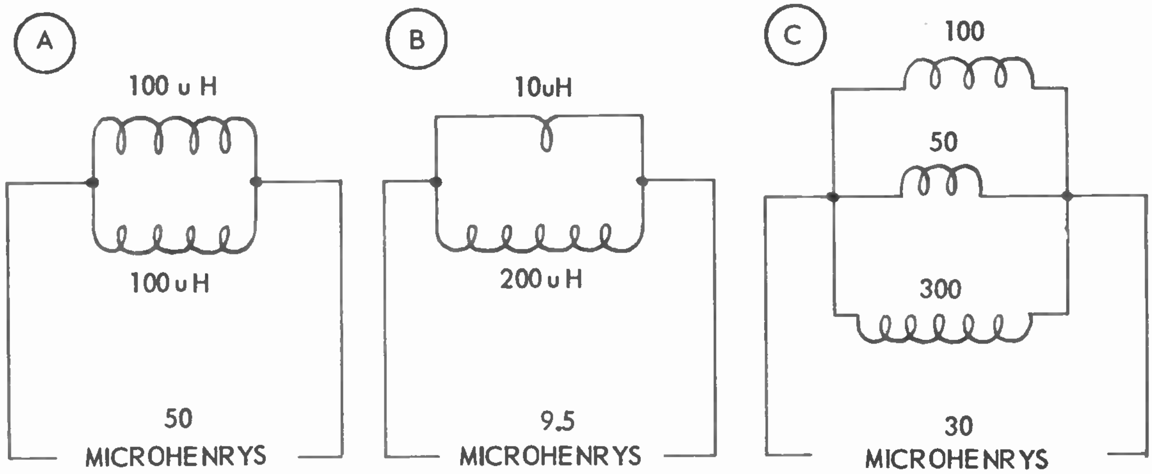


Fig. 29-18. Combined self-inductance of inductances in parallel is determined similarly to combined resistance of resistors in parallel.

Coil turns may be held in place and various parts of the windings may be given added protection by a thin coating of coil cement. Various types of cements are available, but for inductors operating at radio frequencies it is advisable to use the kinds made from polystyrene because of its small dielectric constant and low factor of energy loss.

Coils often are made moisture proof by heavy coatings of wax. Paraffin wax and ceresix wax have small dielectric constants, with maximum of about 2.5, and have loss factors comparable with that of steatite. Where there is likelihood of moistnre collecting on the coils these wax coatings are helpful, but they should not be used otherwise, for the same reasons that all insulation and dielectric material should be avoided so far as is possible.

COIL SWITCHING. The inductance in a circuit sometimes is altered by connecting or disconnecting part of the turns of an inductor. Several methods are illustrated by Fig. 29-16. At *A* there is a tap switch for connecting the lower line to any of several points along the coil, increasing the connected turns and the inductance as the switch is moved downward. The unused turns, below the switch, are called dead end turns. The field of the coil induces emf's in the dead end turns. Energy is expended in these emf's and in connection with distributed capacitance of the unused turns there will be small currents in this part of the winding. All this means a waste of signal energy.

The arrangement at *B* sometimes is an improvement and again it is not. Here the dead end turns, below the switch contact, are short circuited through an additional connection to the switch arm. Currents induced in the unused portion of the coil will circulate through this connection.

With the switching arrangement shown at *C*, or anything equivalent, the unused turns of the coil are completely disconnected from the main portion. With the switch arms downward, as in full lines, the lower section of the inductor is in series with the upper main section to provide maximum inductance. With the switch arms upward, as in broken liues, both ends of the unused turns are open circuited. The unused turns still are inductively coupled to the main part of the winding, and in the disconnected turns there will be induced emf's and currents which waste energy.

If the second section of coil circuit is not inductively coupled to the first section, as indicated at *D*, there will be no induced emf's and currents in this second section when it is open circuited and disconnected.

(6) *INDUCTORS IN PARALLEL AND IN SERIES.* Whenever inductances are connected together in series, as *C and D* of Fig. 29-16, the inductances add together. If one coil or section of a winding has, for example, inductance of 200 microhenrys and the other coil or section has inductance of 100 microhenrys, the total inductance will be 300 microhenrys with them in series. This statement of series inductances adding is strictly true only when there is no coupling between the two inductors.

If there is inductive coupling between the two inductors we have not only the self-inductance of each one but also the mutual inductance due to the coupling. Doubtless you recall that mutual inductance is the property of coupled circuits by which each produces in the other an emf whenever there is a change of current and of magnetic field in either circuit. Mutual inductance exists in addition to the self-inductances.

Something to be remembered about mutual inductance is that it may either add to or subtract from the sum of the self-inductances which are in series. At the left in Fig. 29-17 are two inductors or coils in which currents at the same instant are in such directions that the magnetic fields act in the same direction. Then magnetic lines through one of the inductors pass also through the other. The mutual inductance adds itself to the two self-inductances, and the total inductance is greater than the sum of the two self-inductances.

At the right in Fig. 29-17 one of the coils has been turned end for end, or its connections have been reversed. Now the two magnetic fields are of such polarities at any one instant as to oppose. The magnetic lines through one coil do not pass through the other one. We still have mutual inductance, for the coils are inductively coupled and the field of each induces emf's in the other, but the emf's of mutual induction oppose those of self-induction. Now the mutual inductance is subtracted from the sum of the two self-inductances, and the total inductance is less than the sum of the self-inductances.

With conditions as in the left-hand diagram we say that the mutual inductance is positive. At the right the mutual inductance is negative. Going back to Fig. 29-16, at *C* there is coupling and there is mutual inductance between the two parts of the inductor. Depending on relative directions of current in the two sections the mutual may be positive or negative, and may add to or subtract from the separate self-inductances. At *D* there is no coupling, and there is no mutual inductance. The total inductance will be the sum of the two self-inductances under all conditions.

The lower coil at *D* in Fig. 29-16 might be called a loading coil. A loading coil adds to the self-inductance in a circuit without acting in the transfer of energy to or from that circuit.

When two inductors or coils are connected in parallel with each other the total inductance is not only less than the sum of the self-inductances, it is less than either of the separate inductances. At *A* in Fig. 29-18 are represented two inductances of 100 microhenrys each, connected in parallel. The total inductance is 50 microhenrys, which is equal to one of the separate equal inductances divided by the number of inductances, 2. Maybe this reminds you of something we have heard about several times in the past.

At *B* in Fig. 29-18 there are inductances of 10 microhenrys and of 200 microhenrys connected in parallel. The total inductance is 9.5 microhenrys, approximately. If you divide the product by the sum of these two inductances the answer will be 9.5. At *C* there are three inductances in parallel; 100, 50, and 300 microhenrys. The total inductance is 30 microhenrys. In this case you can use the rule of dividing the product by the sum of 100 and 50. The answer will be 33.3. Then you can use the same rule with the numbers 33.3 and 300, which gives the total inductance, 30 microhenrys.

The total or effective values of self-inductances in parallel are computed in exactly the same way as are the effective resistances of separate resistances connected in parallel. With equal paralleled inductances the total is equal to one of them divided by their number. With other combinations we generally use the rule of dividing the product by the sum.

A type of continuously variable inductor used for laboratory measurements makes use of the change of inductance with two coils connected in series and in parallel. One coil is stationary and the other may be rotated so that their axes vary from in line to a right angle rotation. This changes the coupling from maximum to near zero, and varies the total inductance accordingly. For further reduction of inductance the two coils are reconnected in parallel, and again rotated from maximum to minimum coupling. The total variation of inductance is in the ratio of about 25 to one. Inductors utilizing these principles were used for tuning in some of the earliest radio receivers, then being called variometers.

With reference to all our discussion of inductances in series and in parallel you must bear in mind that we are talking about inductances and not about inductive reactances. Inductive reactances are measured in ohms, like all other oppositions to flow of current. Inductances are measured in microhenrys, millihenrys, and henrys. Reactances in series add together, just like resistances in series, and so do inductances. Reactances in parallel are treated like resistances in parallel, and inductances in parallel are treated likewise. But do not confuse quantities of inductance with quantities of inductive reactance.

Do Not Tear — Cut Along This Line And Send In For Grading.

TELEVISION

LESSON NO. 30

DETECTORS FOR AMPLITUDE MODULATION

"Training Your Mind"

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

COMMERCIAL
TRADES
INSTITUTE

Chicago, Ill.

LESSON NO. 30

DETECTORS FOR AMPLITUDE MODULATION

You are well acquainted with the construction and action of all kinds of capacitors, inductors, and resistors, both fixed and variable. You understand the operation of diodes, triodes, pentodes, and beam power tubes. You are familiar also with power supply systems which furnish direct and alternating voltages and currents for plates, screens, grids, and heaters of tubes. These things are the basic elements from which all radio and television receiving circuits are built.

It won't be long until you can perform all the ordinary service operations on types of broadcast receivers such as the one whose circuit diagram is shown in Fig. 30-1. What's more, when you do a service job you will be able to do it right and in the shortest possible time. This is because you will know why things work as they are supposed to work, which makes it easy to determine what is wrong when things don't do what they are supposed to do.

About all we need do now is put capacitors, inductors, resistors, and tubes into various kinds of receivers to get acquainted with circuit connections that produce sound and pictures. It will be easiest to understand what all the various circuits are good for it we begin with the simplest possible arrangement that is capable of changing a radio signal into sound.

The simplest type of radio receiver in common use is the five-tube superheterodyne, of which Fig. 30-1 is a good example. The first tube at the left is a pentagrid type used as a frequency converter. Next comes a pentode used as an intermediate-frequency amplifier. Then there is a duodiode triode whose single envelope contains two diode plates that work with the cathode as a detector, and contains also a grid and a plate which work with the same cathode as a triode for audio-frequency amplification. Feeding the loud speaker is a beam power tube, the output amplifier. Down below is the power supply, consisting of a half-wave rectifier, a filter, and the tube heaters connected in series on the power line.

Although this receiver is just about the ultimate in simplicity according to present-day standards, you will be astonished at the number of parts which may be dispensed with while still being able to hear radio programs. It would be possible to take away everything except the parts shown at *A* in Fig. 30-2, and still have a practical radio receiver of the kind which furnished entertainment is about 1920. All that remains is the antenna coupling transformer and the diode portion of the duodiode triode tube, this portion of the tube acting as a detector. A detector is indispensable for reception. It separates the audio-frequency signal voltages from the radio-frequency signals arriving by way of the antenna.

The antenna transformer includes a primary winding connected between antenna and ground, and a secondary winding across which are a variable tuning capacitor and a trimmer capacitor. The secondary winding and the capacitors form a series resonant circuit that may be tuned to resonance at the frequency of whatever signal is to be selected.

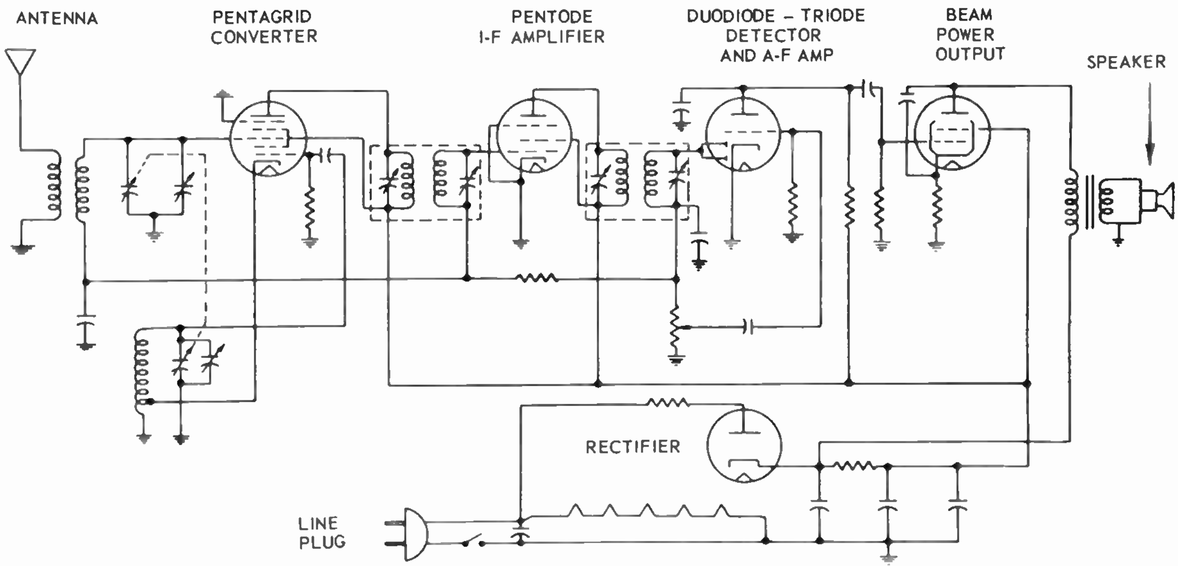


Fig. 30-1. The complete circuit diagram for a five-tube superheterodyne receiver.

The portions remaining of the duodiode triode tube are two diode plates and the cathode. The two plates are connected together to act like one. There are two plates merely because the original tube is made that way in order to handle some types of circuits that require two diodes, as might be found in other receivers.

Taken by themselves the remaining parts of our broadcast receiver would be connected as at *B* in Fig. 30-2. Here the detector is shown as an ordinary two-element diode, which is all we need. The connection from the stator of the tuning capacitor, which formerly went to a grid in the pentagrid converter tube, now goes to the plate of the diode detector. Because everything beyond the detector has been discarded, including the loud speaker, we must add a headphone to produce audible sounds – just as nearly everyone did in 1920. The line which formerly went from the detector output through the volume control resistor to the grid of the audio amplifier now goes to one side of the headphone.

All the ground connections except the one at the bottom of the antenna circuit in diagrams *A* and *B* are to chassis metal, so all of them go to this continuous metal conductor and all are effectively connected together. Probably it will be easier for you to follow the paths of signal currents when the circuits are redrawn as at *C*, with the ground connections represented by big dots joined together by a line.

In diagram *C* the series resonant circuit that includes the transformer secondary and the tuning capacitor shows up quite clearly as being completed through fixed capacitor *Ca*. This capacitor has small reactance to radio-frequency currents in the secondary winding and tuning capacitor, but has high reactance to the low-frequency audio currents. Consequently, audio frequencies are forced to go through the detector and volume control resistor while radio-frequency currents flow freely in the tuned resonant circuit.

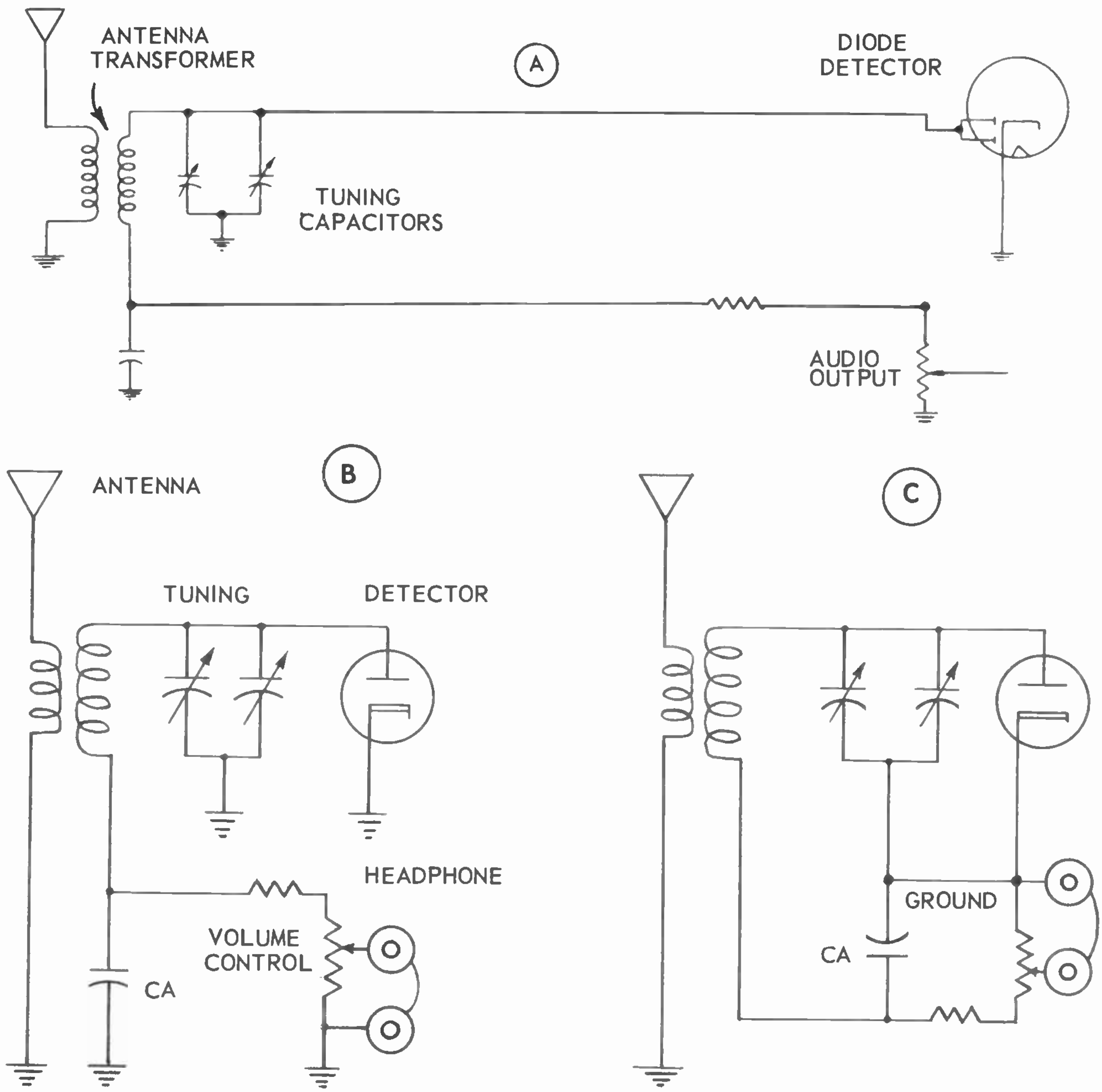


Fig. 30-2. The minimum number of parts which will provide broadcast reception.

The parts which have been removed as nonessential have taken with them all the superiorities of today's receivers over those of 1920. We have lost a great part of the original selectivity, which allows separating the signals of one station from those of other stations. Now, when we listen to one program, we shall hear in the background the programs of other stations whose carrier frequencies are close to the one selected.

We have lost also most of the sensitivity of our original five-tube superheterodyne. Sensitivity is the property of a circuit or an entire receiver that allows listening to stations far away. Still, we have a real radio receiver. Later we shall put back the other parts, one by one, and discover why they give the selectivity and sensitivity that nowadays are expected as normal performance.

In checking over the action of the greatly simplified receiver we need spend no time on the tuned transformer secondary, for we already know about resonance and what it accomplishes. But we must find out how the diode detector works. Diode detectors were used in the very earliest radio receivers. Then came more sensitive detectors, but they did not deliver tones as pure even though they did make the sounds louder. Now, in all receivers, we have gone back to diode detectors, both in sound radios and in the video circuits of all television receivers. This is because we now have other and better means than the detector for providing lots of sensitivity and because we must have faithful reproduction of signals, which the diode provides.

AMPLITUDE MODULATION. To find out how any detector works we must know something about the kind of signal on which it works. Sound radio signals for standard broadcasting are transmitted by a method called amplitude modulation. Amplitude modulation is employed also for transmitting all picture signals and all synchronizing signals for television.

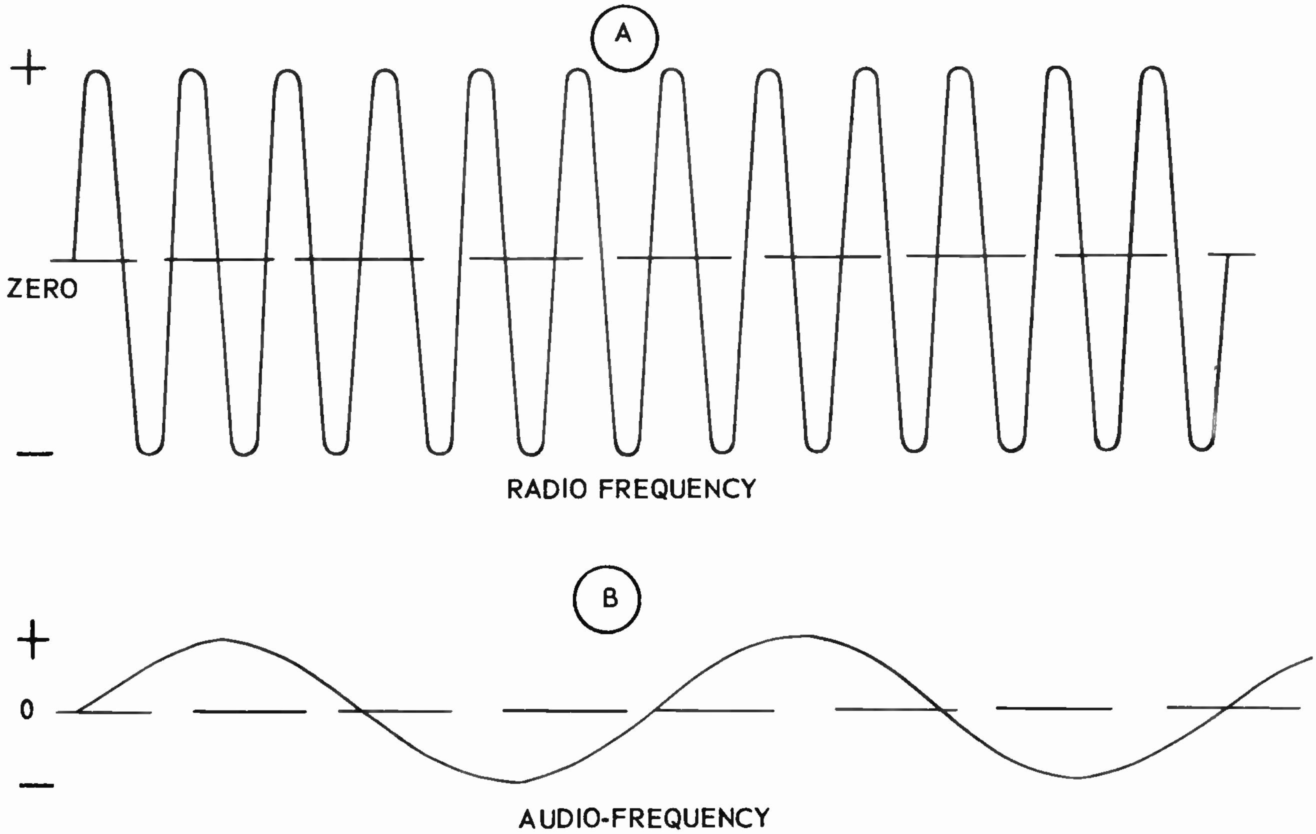


Fig. 30-3. A radio-frequency carrier voltage and an audio-frequency voltage used for modulation.

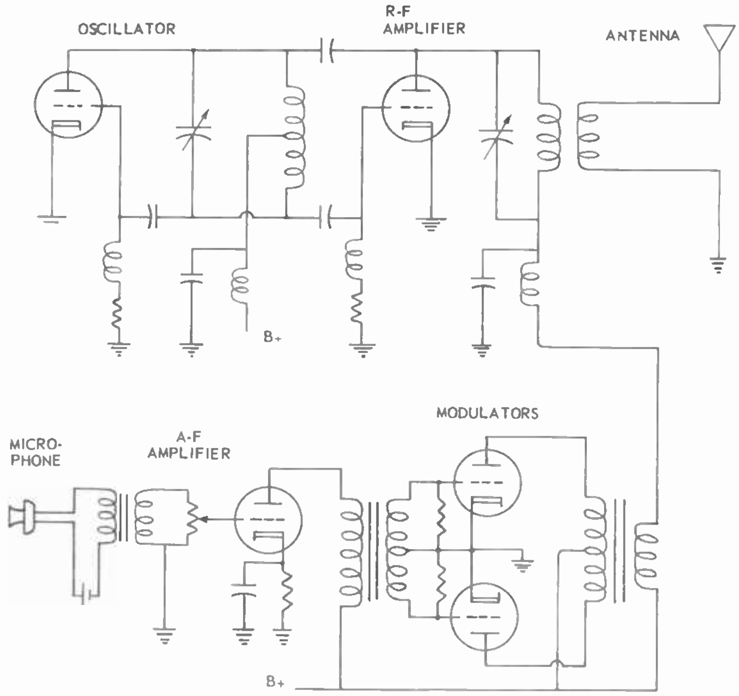


Fig. 30-4. Circuits in an amplitude-modulation radio transmitter.

A radio transmitter produces and radiates from its antenna a carrier wave, even when no sound signals are being transmitted. That is, a station may be "on the air" during intervals when there are no sounds. A carrier wave flies through space and produces in the receiving antenna a voltage alternating at radio frequency. A few cycles of r-f voltage, carrying no sound signal, may be represented as at A in Fig. 30-3. This voltage from the receiving antenna will appear in the primary of the antenna coupling transformer, and by inductive coupling will pass into the tuned secondary.

At *B* is represented a sound signal, at audio frequency, which is to be transmitted. For the sake of simplicity this sound voltage is shown as a sine wave, which would give a single constant tone. Actually the audio voltage will vary in frequency in order to deliver various tones, or sounds of different pitch. The actual audio voltage will vary also in amplitude, to deliver sounds that are louder or less loud.

Fig. 30-4 shows the basic circuits for a complete radio transmitter of small size and power. Radio-frequency voltages and currents which produce the radiated carrier wave originate in the oscillator, are strengthened in the r-f amplifier, and go to the antenna of the transmitter. Audio-frequency voltages to be transmitted originate in the microphone, are strengthened by the a-f amplifier, and go to the modulator tubes.

If you commence tracing at the plate of the r-f amplifier in the transmitter, and follow through to the B+

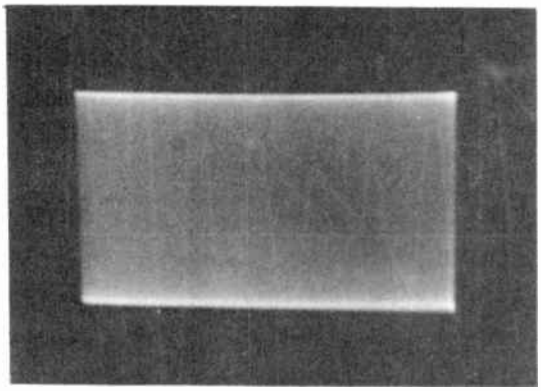


Fig. 30-5. The unmodulated r-f voltage.

connection at the bottom of the diagram, you will find that this plate circuit of the r-f amplifier goes through the secondary winding of a transformer whose primary is connected to the plates of the modulators. Thus the audio voltages from the modulators get into the output of the r-f amplifier, and r-f voltages and currents going to the antenna are made to vary in accordance with the audio signal voltages. This is modulation of the r-f carrier.

In your own service shop or laboratory you can reproduce this whole performance, and experiment to your heart's content with all manner of combinations of r-f voltages modulated by a-f voltages. This is because the signal generator which you will use for servicing contains an oscillator for production of r-f voltages, another oscillator for production of a-f voltages, and means for combining the two voltages in a modulated output. If you connect the output of the signal generator to the input of your service oscilloscope, on the screen of the oscilloscope will appear pictures of what happens. This was done in making some pictures which follow.

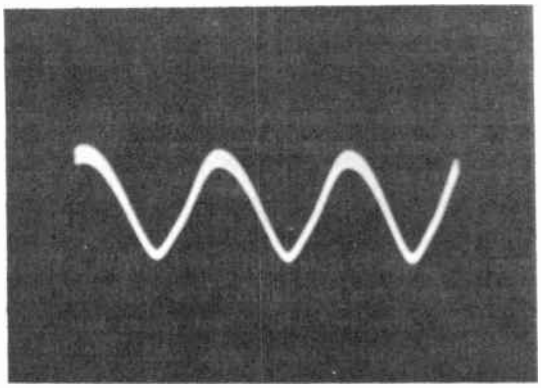


Fig. 30-6. Audio voltage used for modulation.

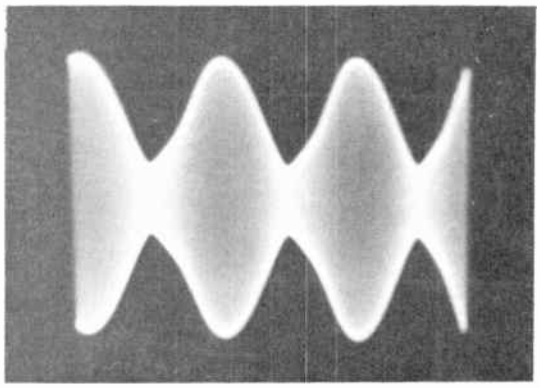


Fig. 30-7. The r-f voltage, strongly modulated.

First you would turn off the audio modulating voltage in the signal generator and look at the unmodulated r-f voltage. Fig. 30-5 is a photograph of an oscilloscope screen on which is appearing the illuminated trace produced by an r-f voltage at a frequency of 500 kilocycles per second. This picture shows the r-f voltage during a time period of about 7/1000 second. Within this brief time there occur 3,500 complete r-f cycles. Successive cycles come so close together that they appear as only an extended blur, but here we have in fact what is shown in principle for only a few cycles at *A* in Fig. 30-3.

Your signal generator probably will be of a type which will furnish the a-f voltage by itself, as well as when modulating the r-f voltage. If you apply only the a-f voltage to the oscilloscope this voltage will show up as in Fig. 30-6. The audio frequency used in making this picture was 400 cycles per second. There are about 2.8 complete cycles, which occur during a time period of 7/1000 second when the frequency is 400 cycles. Here we have the actual equivalent of the audio voltage shown at *B* in Fig. 30-3.

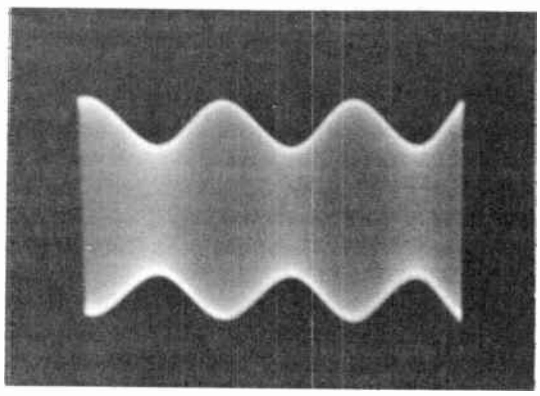


Fig. 30-8. Effect of reducing the modulating voltage.

The next stop would be to modulate the 500-kilocycle r-f output of the signal generator with the 400-cycle audio voltage, just as might be done at the radio transmitter. The result is shown by Fig. 30-7. Here the amplitude or the up and down change of the r-f voltage is varying in exact accordance with the changes of audio modulating voltage.

A horizontal line drawn through the middle of the modulated voltage wave would represent zero voltage, just as at *A* in Fig. 30-3. The r-f voltage changes back and forth between maximum positive at the top through zero to maximum negative at the bottom. The effect of modulating the r-f voltage is to vary the maximum voltages at successive instants on both the positive side and the negative side of the zero line. Either the top or the bottom of the modulated voltage looks like the a-f voltage that is doing the modulating. If you examine the picture quite carefully it will be plain that the wavy outline along the bottom is the same as the wavy outline across the top except for being turned upside down.

Note, in Fig. 30-7, that the rise and fall or the vertical changes along the wavy line of modulation are of exactly the same extent as changes of amplitude in the modulating voltage shown by Fig. 30-6. The

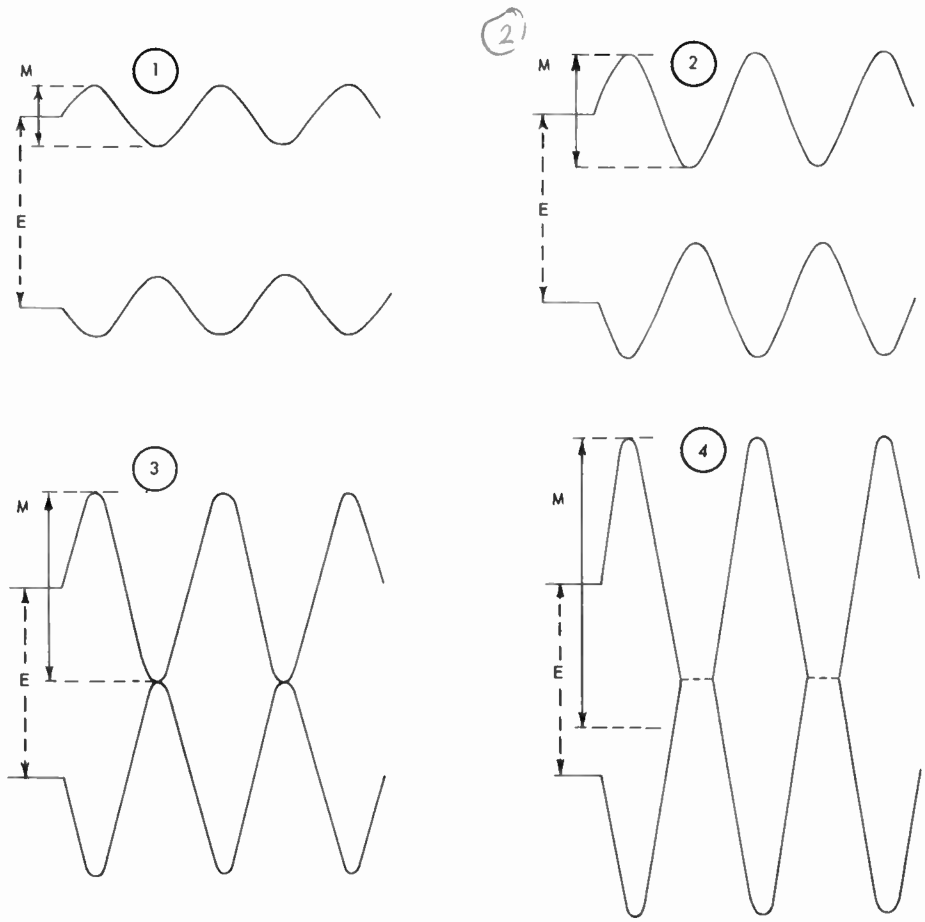


Fig. 30-9. Percentages of modulation. 1, thirty per cent. 2, sixty per cent. 3, one hundred per cent. 4, over-modulation.

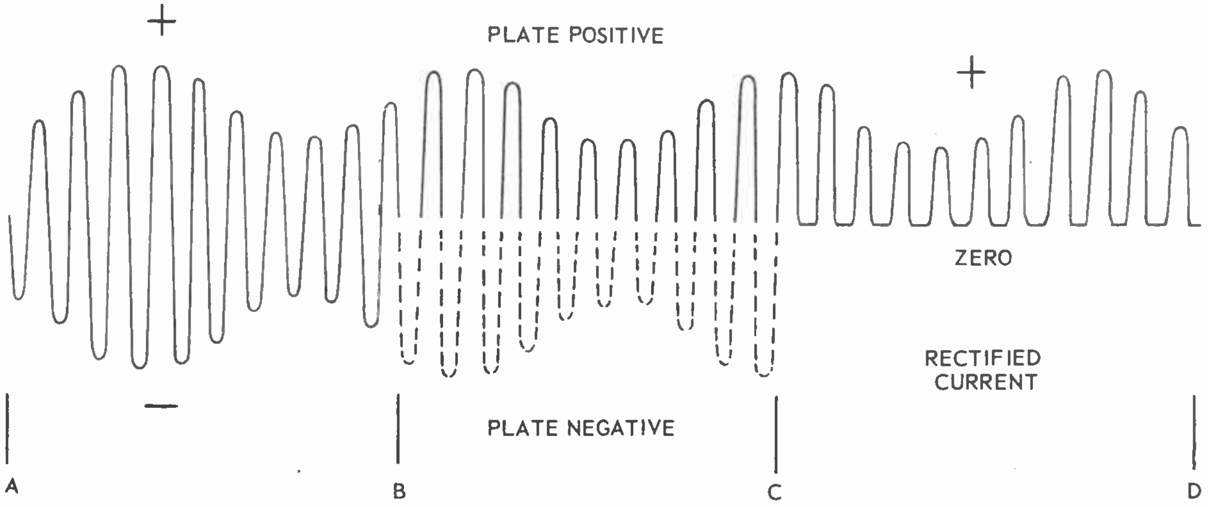


Fig. 30-10. What the diode detector does with the modulated r-f voltage.

wavy lines along top and bottom of the modulated r-f voltage are called the modulation envelope. The shape of the modulation envelope is the same as changes of amplitude in the modulating voltage.

If now you were to reduce the strength or amplitude of the modulating audio voltage the result would be as in Fig. 30-8. Modulation of the r-f voltage has been decreased proportionately to decrease of amplitude in the modulating voltage. The modulation envelope has changed with change of modulating voltage.

The number of waves per second in the modulating or the frequency of the modulation is just the same as the frequency of the modulating voltage. The photographs of the two voltages cover equal periods of time. There are the same numbers and fraction of cycles in the modulation as in the audio voltage. Any and every change of frequency in the modulating voltage will be reproduced in the modulation.

In order that you may make service tests and adjustments under actual operating conditions, signal generators furnish r-f and a-f voltages like those produced in a receiving antenna by carrier waves coming from a radio transmitter through space to a receiving antenna. With the help of a signal generator and service oscilloscope we have seen amplitude modulation in action. Very soon we shall be ready to use various types of detectors for recovering the audio signal or audio modulation from the r-f voltage, but first there are a few additional matters to be cleared up with reference to amplitude modulation.

Here is one of the things which should be understood. In a modulated signal there always remains the radio frequency of the carrier regardless of what frequency may be used for modulation. As an example, in Figs. 30-5, 30-7, and 30-8 we always have the frequency of 500 kilocycles. Only the amplitude of the r-f voltage is undergoing change.

The modulating frequency itself does not exist in the r-f wave. That is, in the waves just mentioned there is no 400-cycle frequency. This frequency will reappear only after the detector does its work. There is, however, a frequency equal to the sum of the radio and audio frequencies which, in our examples, would be 500,400 cycles per second, or 500 kilocycles plus 400 cycles. And there is another frequency equal to the

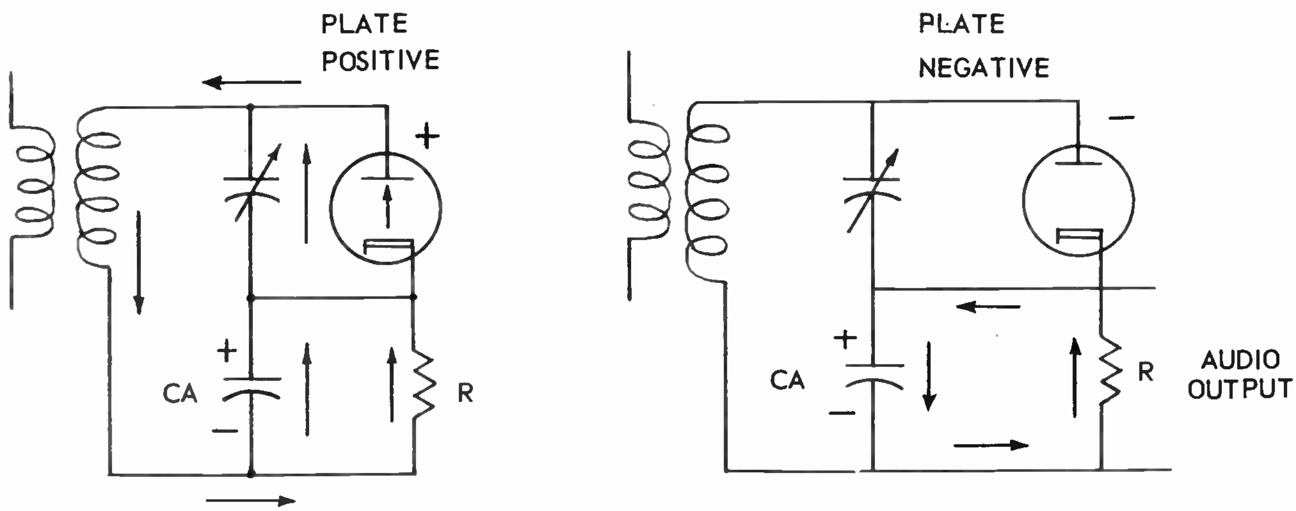


Fig. 30-11. Electron flows in the diode detector circuit during positive and negative r-f alternations.

difference between radio and audio frequencies. This difference frequency would be 499,600 cycles per second in the waves at which we have looked. Frequencies which are the sum of carrier and modulating frequencies, and are higher than the carrier, are called upper sideband frequencies. All which are equal to the difference, and are lower than the carrier, are lower sideband frequencies.

Normally the sideband frequencies would be varying, because in speech and music there are frequencies all the way from about 50 to 15,000 cycles or more. A radio-frequency amplifier must handle, and should amplify, all frequencies that may be produced in upper and lower sidebands.

Another thing which we should understand is the matter of modulation percentage or the percentage of modulation. This refers to the extent of the modulation as compared with the amplitude of the unmodulated r-f voltage or unmodulated carrier wave. The percentage of modulation is the percentage of the unmodulated carrier amplitude represented by the amplitude of modulation. For example, were you to have an unmodulated carrier with peak to peak amplitude of 100 volts, and modulate it with an audio voltage having peak amplitude of 30 volts, the carrier would be modulated 30/100 or 30 per cent.

Fig. 30-9 shows how several percentages of modulation would appear on the oscilloscope. Peak-to-peak amplitude of the unmodulated r-f voltage is marked *E*. It is the same in all four diagrams. Peak-to-peak amplitude of the modulating voltage is marked *M* on each diagram. In diagram 1 we have 30 per cent modulation. Amplitude of the modulating voltage is 30 per cent of the amplitude of the unmodulated r-f voltage or the carrier wave. In diagram 2 the modulation has been increased to 60 per cent. Now look back at Fig. 30-7. There the modulation is between 58 and 59 per cent. In Fig. 30-8 the modulation is somewhat less than 25 per cent.

Back to Fig. 30-9, in diagram 3 there is 100 per cent modulation. The modulating amplitude equals the amplitude of the unmodulated r-f voltage. This is as far as we can go while preserving the true shape of the modulating voltage. In diagram 4 the modulation is more than 100 per cent. This is called over-modulation. Parts of the modulating voltage wave are cut off and there are breaks in the r-f voltage or the carrier wave. When a detector works on an over-modulated signal the resulting sound will be greatly distorted.

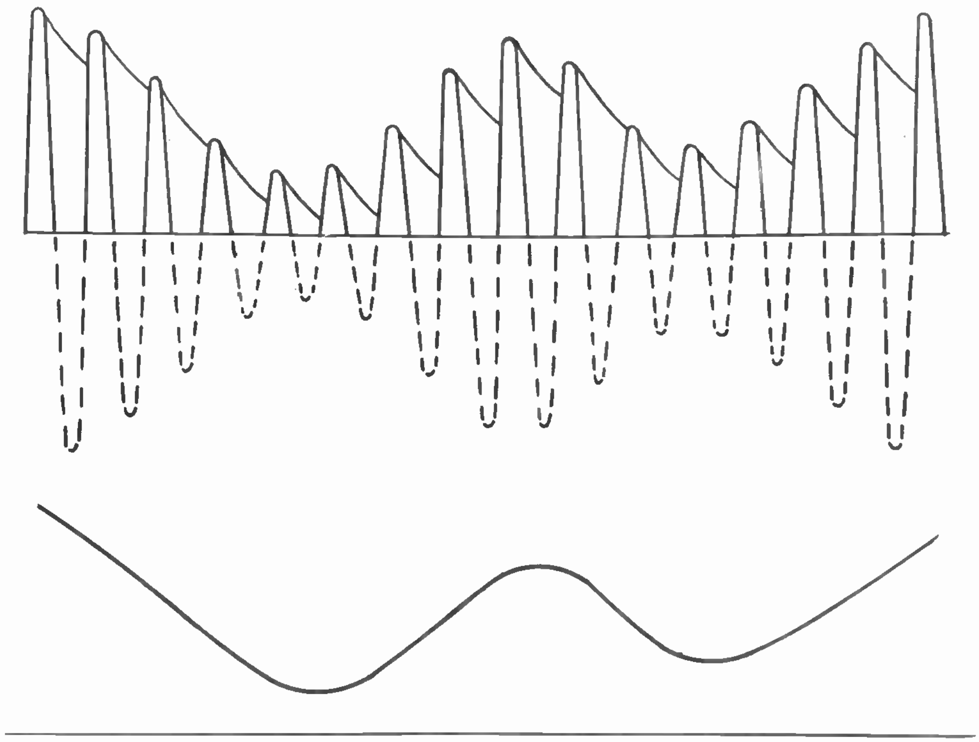


Fig. 30-12. Charge and discharge of the capacitor in the diode detector circuit.

DETECTORS. In the process of amplitude modulation we commence with an r-f voltage of constant amplitude, combine with it an a-f voltage of varying amplitude and varying frequency, and end with the modulated r-f voltage. The process of detection is the reverse of modulation. We commence with a modulated r-f voltage, separate from it the a-f signal voltage, and discard the remaining r-f voltage. The process of detection often is called demodulation, and the detector may be called a demodulator.

In following portions of this lesson we shall examine the action of detectors used for amplitude modulation. Detectors for frequency modulation, as used in f-m receivers, operate on entirely different principles. F-m detectors or demodulators will be examined when we come to the subject of frequency modulation in general.

6 *DIODE DETECTOR.* The diode detector acts primarily as a rectifier. When the modulated r-f voltage is applied across the plate and cathode of the diode this alternating voltage makes the plate first positive and then negative with reference to the cathode. While the plate is positive there is conduction through the diode, and while the plate is negative there is no conduction.

This rectifying action is illustrated by Fig. 30-10. Between *a* and *b* is represented the modulated r-f voltage, with positive alternations above and negative alternations below. The positive alternations of voltage, shown by full line curves between *b* and *c*, cause electrons to flow from cathode to plate in the

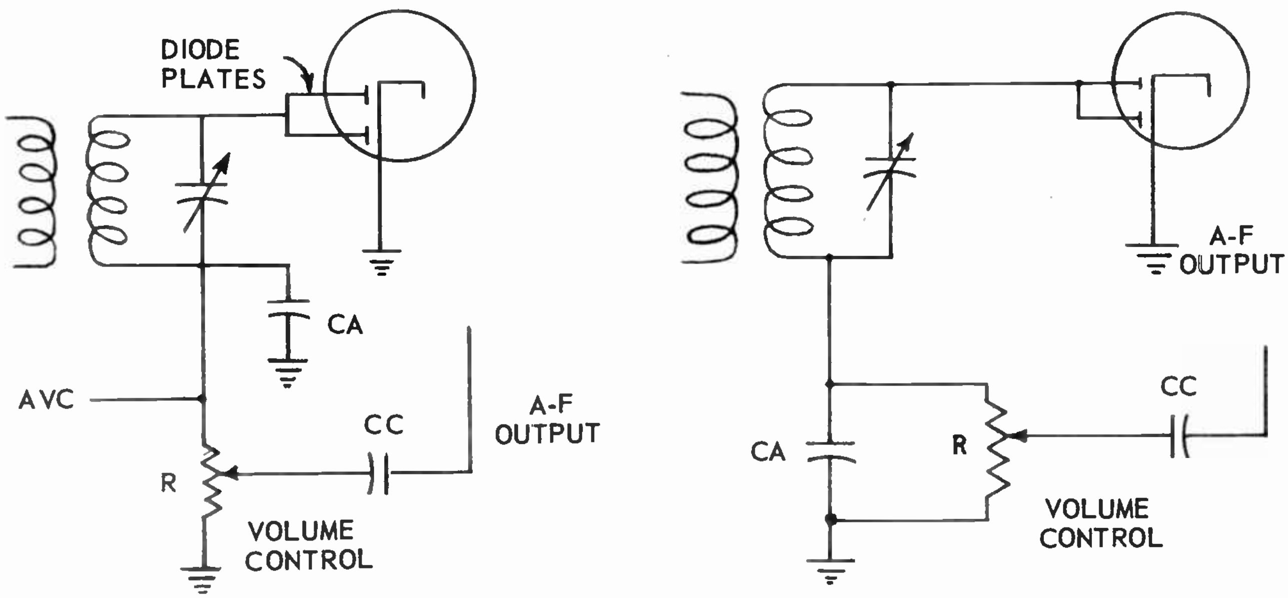


Fig. 30-13. Diode detector circuits used in standard broadcast receivers.

diode and the connected circuit. The negative voltage alternations, in broken-line curves, have no effect. The result, shown between *c* and *d*, is a rectified current that flows in pulses as it varies between zero and positive peaks whose values depend on modulation of the original input voltage.

Electron flows during positive alternations and negative alternations of r-f voltage are shown by Fig. 30-11, in which are simplified circuit diagrams of our diode detector. At the left are shown conditions during the voltage alternation that makes the diode plate positive. Voltage from the secondary winding of the r-f transformer is being applied between the top of the tuning capacitor and the bottom of capacitor *Ca*, also,

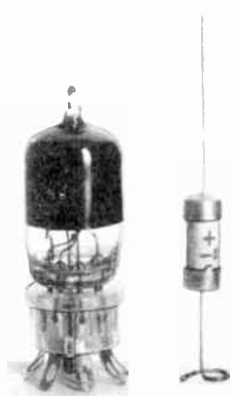


Fig. 30-14. A crystal diode and a miniature radio tube.

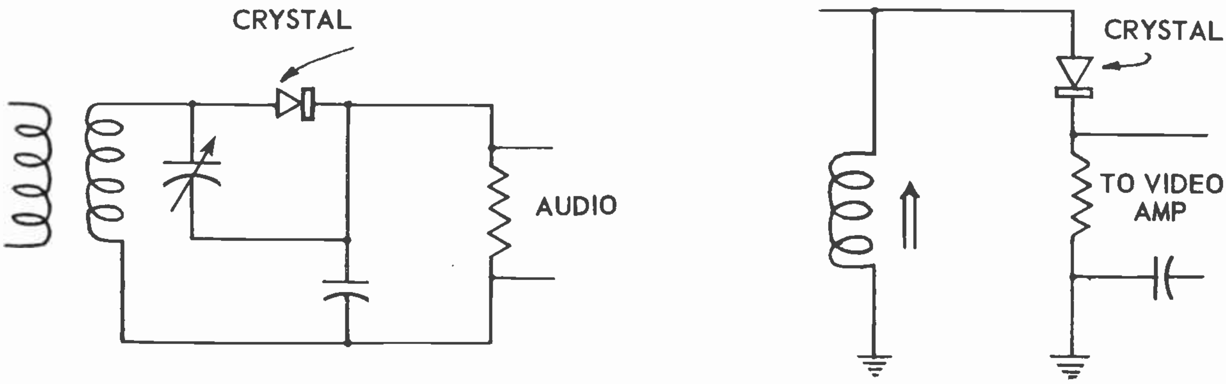


Fig. 30-15. Circuits for crystal diode detectors.

between the plate of the diode and the bottom of resistor R. Resulting electron flows are shown by arrows. There is electron flow upward through capacitor *Ca* and the tuning capacitor. There is also an upward flow through resistor *R* and from cathode to plate in the diode. Capacitor *Ca* is being charged, with its upper plate positive.

Conditions during the r-f voltage alternation that makes the diode plate negative are shown by the right-hand diagram. Because the diode plate is negative with reference to the cathode there can be no electron flow in the diode. But capacitor *Ca* now is discharging through resistor *R*, with electron flow in the direction of the arrows. Note that this electron flow through *R* is in the same direction as during the opposite voltage alternation.

Capacitor *Ca* charges during voltage alternations of one polarity, and discharges during alternations of the opposite polarity. To what extent will the capacitor discharge between pulses of charging current, or how much of the capacitor voltage will be lost before it is again built up by the next charging pulse? It is entirely a matter of the time constant of capacitor *Ca* and resistor *R*, and of the frequency of the r-f voltage.

Discharges of the capacitor between successive peaks of charging current are shown by the lines between the peaks in Fig. 30-12. The effect is somewhat exaggerated because the peaks have to be drawn far enough apart to illustrate the action. The variation of charge between successive pulses would be very small, and there would be a practically smooth variation of capacitor charge and voltage as shown by the wavy line at the bottom of the diagram. It is quite apparent that this wavy line of average capacitor voltage is the voltage of audio modulation used on the r-f carrier coming to the receiver.

The time constant of the capacitor and resistor in the diode circuit must be long enough that there won't be much discharge between r-f pulses, yet short enough that discharge (and recharge) can follow the audio frequency of modulation.

Capacitor *Ca* and resistor *R* are in parallel with each other. Therefore the voltage always will be the same across both units. This average voltage is the audio modulation or the audio signal being transmitted. So we may take the audio output of our diode detector from across the capacitor or the resistor and apply this voltage to a following a-f amplifier or, in our experiments, to the headphone.

Resistor *R* is called the detector load resistor. For the majority of service operations classed as alignment adjustments in television receivers we connect either an electronic voltmeter or an oscilloscope across the load resistor of the video detector.

An increase of resistance in the detector load resistor gives some increase of audio output voltage. In tests on a broadcast receiver, doubling the detector load resistance increased the audio output voltage by 10 to 25 per cent, with the greater increases secured on relatively weak r-f inputs.

7 A correctly designed diode detector circuit is capable of handling both weak and strong signals with little distortion. This means that the audio output voltage will follow the modulation quite faithfully at all signal levels. A disadvantage of the diode detector is that it forms, in effect, a resistance in parallel with the tuned input circuit. The detector is equivalent to a paralleled resistance because it allows additional rectified current to flow in the tuned circuit, as is evident in Fig. 30-11. This effect is called loading of the input. Loading, with current or resistance, acts to broaden the band of frequencies in which the input circuit delivers relatively high voltage, which means a reduction of selectivity. The whole subject of selectivity will be investigated very shortly.

In connection with diode rectifiers you are going to hear the word "perveance". Perveance describes the ability of a diode detector to rectify high frequency voltages with little loss or drop of voltage in the rectifier itself. A diode having high perveance will not only have fairly low internal resistance to the range of current it is supposed to handle, but will have also reasonably low internal capacitance between plate and cathode. This internal capacitance is in parallel with the tuning capacitor.

It should not be forgotten that connections for the detector in the receiver from which we started have been changed around to simplify the explanations. We replaced all the ground connections with the conductor shown as a line in Fig. 30-11, the line from between the two capacitors over to the diode cathode.

The original connections, including the grounds, are as at the left in Fig. 30-13. Capacitor *Ca* and resistor *R* are the same as in our explanatory diagrams. Capacitor *Cc* is between resistor *R*, the volume control, and the line to the grid of the a-f amplifier. This capacitor allows only the alternating audio voltages to go to the a-f amplifier, and holds back the rectified pulsating direct current which is in the volume control resistor. Voltage from the volume control is used as grid bias voltage for other tubes and provides automatic volume control in a way we shall learn about in due time. This bias voltage goes through the connection marked "AVC", an abbreviation for automatic volume control. Were you to connect together all the grounds in Fig. 30-13 you would have the same electrical system as shown in Fig. 30-11.

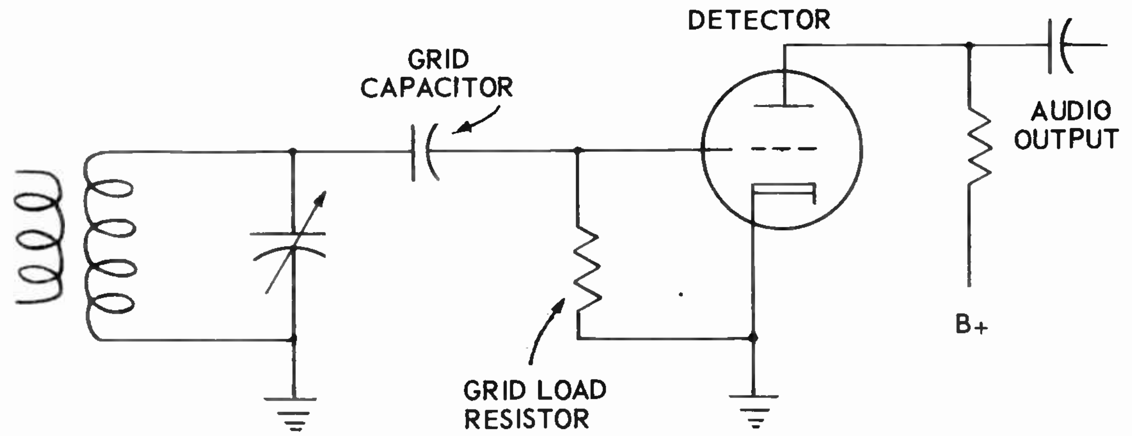


Fig. 30-16. A circuit for a grid-leak detector.

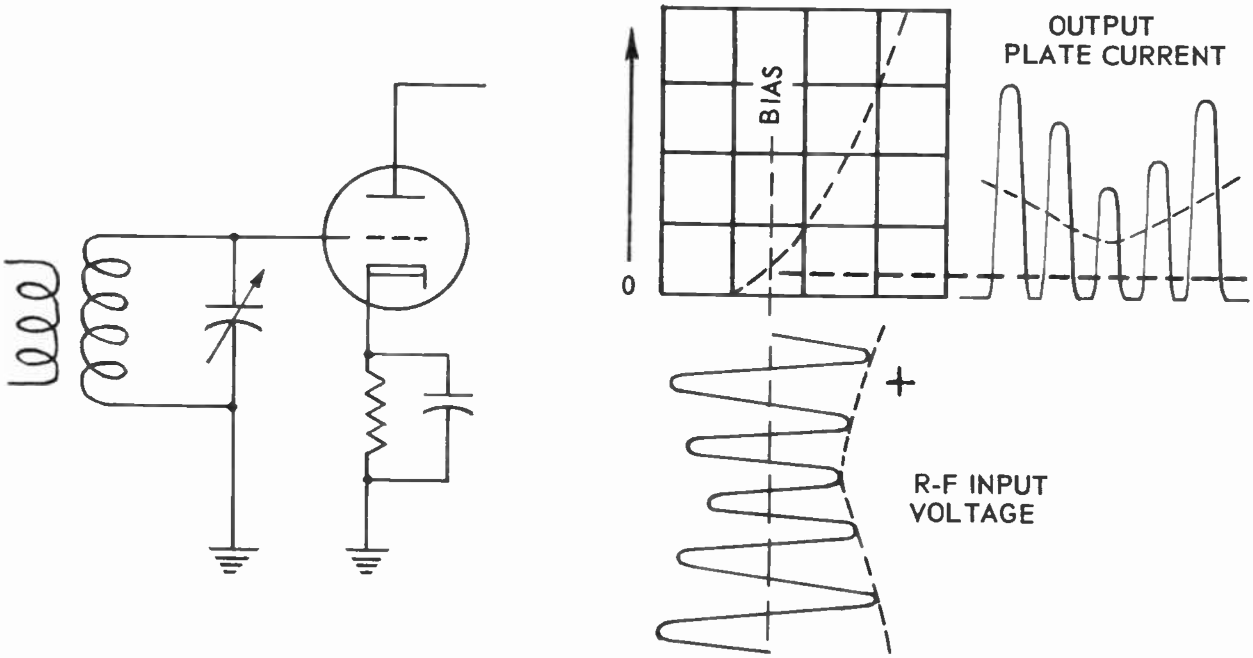


Fig. 30-17. A plate detector circuit, and the action of such a detector.

At the right in Fig. 30-13 is another way of showing the same electrical circuit for the diode detector. Again *Ca* and *R* are the same as in preceding diagrams, and capacitor *Cc* is in the a-f output line. Connecting all the grounds together would give our regular circuit for the diode detector. Neither of the detectors in Fig. 30-13 is across an antenna coupling transformer, both are across the secondary winding of intermediate-frequency coupling transformers. Whether the high frequency voltage for the detector comes from the antenna circuit or from some other circuit makes no difference in detector action.

Diode detectors in sound radio receivers practically always are sections of multiple tubes in which there are also a grid and a plate which, with the common cathode, form a triode, or in which there are the additional elements which form a pentode. The triode ordinarily is used as an audio-frequency amplifier. The pentode may be used as either an audio-frequency amplifier following the detector or as a radio-frequency amplifier preceding the detector.

In all the circuits which have been examined, the diode detector acts as a half-wave rectifier. Although half-wave rectification or detection is almost the universal practice it is possible to use the two diode plates in a full-wave detector circuit whose connections to a center-tapped secondary winding of the input transformer are like those of a full-wave power rectifier to its center-tapped power transformer. The half-wave diode detector delivers greater output voltage than a full-wave type with equal r-f input voltages, but delivers somewhat less power. Since audio output voltage usually is more to be desired than added power, there is no particular advantage in the full-wave detector.

CRYSTAL DIODE DETECTORS. The detectors in most receivers are electronic tubes of the hot cathode type, with either a heater-cathode or a filament-cathode. There are, however, some sound radio receivers and a considerable number of television receivers employing crystal diode detectors. Fig. 30-14 is a picture of one style of crystal diode alongside one of the smallest of the miniature hot-cathode electronic tubes with its socket. The crystal diode in the picture is 11/16 inch long and has a maximum diameter of 1/4 inch. There are other types even smaller, about 3/8 inch long and somewhat less than 1/4 inch in diameter.

The crystal illustrated is enclosed within a ceramic tube and has exposed metallic caps at both ends. Other insulations may be used, including glass for some types, and there may be no exposed metal. All these crystal diodes are moisture-proofed and most of them are hermetically sealed.

The crystal element is a piece of the metal called germanium in detectors used for all ordinary kinds of receivers, or for operation at frequencies up to about 100 megacycles. Other kinds of elements, including silicon, will operate satisfactorily at frequencies up to 30,000 megacycles and more. In contact with the surface of the germanium crystal element is the end of a small-diameter wire of tungsten or of platinum. In some crystal diodes the wire, called a cat whisker, makes pressure contact on the germanium and in other types it is welded in place.

The cathode end or cathode terminal of a crystal diode may be marked "Cath" or may be identified by a minus sign (–). The anode end, equivalent to the plate in a tube, may be identified with a plus sign (+). Sometimes both ends are marked, and again only one is marked. Obviously, if you can identify either end there can be no difficulty in identifying the other end, since there can be only a cathode and an anode.

A crystal diode may be used in any kind of detector circuit where a tube diode would work. At the left in Fig. 30-15 is a detector circuit containing a crystal diode for demodulation of audio frequencies. At the right is one style of crystal detector circuit used in television receivers. Distributed and stray capacitances are sufficient for most of the tuning in television circuits. The symbol for a crystal diode is like that for the selenium rectifier used in power supplies, it is the symbol that represents any kind of "contact rectifier" in which there is oneway conductivity at the contact between two elements.

A crystal diode is mounted by soldering its pigtail leads to terminal lugs or other convenient parts in the circuit. The crystal requires no socket, and since it has no heater or filament there is no wiring other than to the cathode and anode. The resistance of the crystal to electron flow from cathode to anode is less than the internal resistance of a diode tube to similar flow, but resistance of the crystal to reverse electron flow is no where near as high as in a diode tube, Therefore, the crystal diode permits a very small electron flow in the reverse direction, from anode to cathode, during voltage alternations which make the anode negative with reference to the cathode.

GRID-LEAK DETECTOR. Now we are going to spend a few minutes in examination of a detector that once was in nearly universal use for radio broadcast reception, and now is practically obsolete. We cannot very well skip this detector, because it is familiar to so many experienced radio technicians and they are certain to mention grid leak detection once in a while. You cannot afford to appear entirely ignorant on such a subject.

This obsolete detector is the grid-leak type. You will recall that in an earlier lesson we studied grid-leak biasing, which is far from obsolete and, in fact, is in common use for many tubes in television receivers. With that method of biasing we employ a grid capacitor and grid resistor or grid leak connected as in Fig. 30-16. This circuit could be labeled either grid leak biasing or grid leak detection.

If the grid-leak circuit is used to produce a negative bias proportional to the strength of any incoming alternating voltage it is a grid-leak biasing circuit. If it is used to demodulate an r-f voltage and recover the audio signal it is a grid-leak detector circuit. The difference is not in the circuit, it is in the purpose for which the circuit is used.

Unless you recall the operating principles of grid-leak biasing it may be well to go back and read about them in the earlier lesson. There you will find that every increase in amplitude of the incoming alternating voltage makes the grid more negative, while every decrease of amplitude lets the grid become less negative. Grid voltage follows the changes of amplitude in the applied signal, and since changes of amplitude in an r-f signal voltage constitute the modulation of the sigual, the grid voltage of a grid-leak detector follows the modulation.

The grid being used for detection is the grid of a triode tube. Voltage on the plate circuit of the triode is of a value suitable for moderate amplification. Then the triode amplifies the changes of voltage which are appearing on its grid, just like any triode amplifier. These changes of grid voltage are following the modulation of the incoming signal, so the output in the plate circuit of the grid-leak detector is the amplified audio signal.

Another way of looking at the action of the grid-leak detector is to consider the grid and cathode as forming a diode detector. The grid is acting like the diode plate. The grid, the plate, and the cathode are acting together as a triode amplifier. First there is detection in the grid-cathode circuit acting as a diode detector. Then there is amplification in the triode circuit, which includes the plate.

The grid-leak detector is more sensitive than any other type on weak signals. But there is distortion in the plate circuit output because grid voltage and plate voltage cannot be held at such relative values as will cause uniform amplification of all signal strengths while maintaining detector action. This distortion becomes quite objectionable on all except weak incoming signals. Nowadays we don't need the sensitivity of the grid-leak detector on weak signals, because our modern amplifier tubes have high enough transconductance to furnish all the amplification or sensitivity that can be used. Consequently, the grid-leak detector, with its distortion, has passed out.

PLATE DETECTOR. There is one more detector once commonly used in sound radio receivers, and now nearly obsolete. However, as a competent television-radio technician you are presumed to know at least the principles of all these things, even though not now in general use. Incidentally, after the disappearance of grid-leak detectors, no one used grid-leak-capacitor circuits for much of anything during many years. Then television made it necessary for all the technicians to learn all about grid-leak biasing. You never can tell what is going to happen.

This other nearly obsolete detector is variously called a plate detector, grid bias detector, plate rectification detector, or infinite impedance detector. The English call it an anode bend detector. A circuit for plate detection is shown at the left in Fig. 30-17. Here we have a triode tube with cathode bias. Instead of cathode bias we might use fixed bias from the d-c power supply. Instead of a triode we might use a pentode. In any case the circuit would look just like an amplifier circuit. It becomes a detector circuit by choice of grid bias and plate voltage that allow detection instead of amplification as the principal result.

Action in the plate detector is illustrated at the right in Fig. 30-17. Bias is made so highly negative that operation is at or near plate current cutoff. Then only the positive alternations of r-f input voltage cause any great flow of plate current, with the negative alternations driving the grid beyond the cutoff voltage. Now we have rectified the r-f voltage, very much as was done with the diode back in Fig. 30-10. The average value of plate current follows the changes of modulation, as may be seen on the right-hand side of the graph.

The plate detector has a real advantage in that the grid remains so highly negative that no current can flow in the grid circuit. This means that detector input resistance is effectively very high, there is negligible loading of the tuned input circuit, and selectivity is good. Sensitivity to weak signals is greater than with a diode detector, less than with a grid-leak detector. There is some distortion, due to working the tube on the sharpest part of the bend in the mutual characteristic. This distortion is lessened by using cathode bias instead of fixed bias.

Now we understand detectors and how they are used. A detector plus a tunable antenna circuit plus a headphone equals a radio receiver. The receiver will have tone quality as near perfection as can be had from whatever signal comes from the antenna. But we lack selectivity, we lack sensitivity, and we lack enough audio output strength to operate a loud speaker. Our receiver wouldn't be very salable in today's market, so in the next lesson we shall commence improving the performance.

31 → 35

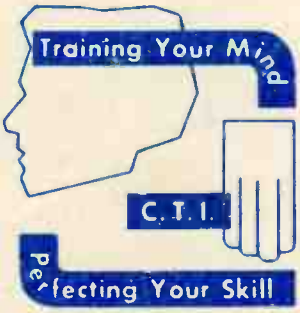
COMMERCIAL TRADES INSTITUTE

KNOWLEDGE + AMBITION = SUCCESS

LEARN . . .

ACT . . .

ENJOY . . .



"The heights by great men reached and kept
Were not attained in sudden flight,
But they, while their companions slept,
Were climbing upward in the night."

TELEVISION

LESSON NO. 31

SELECTIVITY AND Q-FACTOR

"Training Your Mind"
C.T.I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON NO. 31

SELECTIVITY AND Q-FACTOR

If we were to go about improving our radio receiver in the way the early experimenters did it we would try first for more audio output. Most people wanted to get rid of the headphone and acquire a "loud speaker set". Then they could tell their friends about some program coming in so loud that everyone in the room could hear it. The first attempts to get more volume consisted of adding, after the detector, a stage of audio-frequency amplification, then two stages, and even three. Tone quality usually became progressively worse after the first added audio stage.

Soon it became apparent that audio amplification after the detector didn't really increase distance over which reception could be had. If you could hear a program from the headphone you could hear it louder from the speaker, but if the signal never reached the detector it was like multiplying zero by a thousand, the answer still was zero.

The folks then went in for radio-frequency amplifications adding one or more r-f amplifier tubes between the antenna circuit and the detector. If everyone kept quiet and you put your ear far enough into the horn of the loud speaker, you almost always could hear KDKA from Pittsburgh even though you lived away out in Omaha. Station KDKA, as you may know, was the pioneer broadcaster.

For our own experiments we shall go first to radio-frequency amplification. This will be a step in the right direction, because in a great many radio broadcast receivers we shall be working with a stage of r-f amplification immediately following the antenna, and there are one or more such stages following the antenna in nearly every television set.

Adding a stage of r-f amplification means adding between antenna circuit and detector input the parts enclosed by broken lines in Fig. 31-2. Originally we had the tuned secondary winding of the antenna coupling transformer connected to the detector plate and to the audio output circuit of the detector, from which in the latter circuit there are connections through ground to the detector cathode. With the added r-f stage we connect the tuned secondary of the antenna coupling transformer to the grid of the r-f amplifier and to ground. From ground there is a connection through the cathode bias resistor and bypass capacitor to the cathode of the r-f amplifier.

Modulated r-f voltage from the antenna transformer formerly went into the plate-cathode circuit of the diode detector. Now it goes to the grid-cathode circuit of the r-f amplifier. The modulated r-f voltage is accompanied in the plate circuit of the r-f amplifier tube by a corresponding r-f current. This r-f current, with the audio modulation, is in the primary of the tuned r-f transformer. It induces a corresponding emf, at the modulated radio frequency, in the secondary of this transformer. The secondary emf or voltage is applied to the detector circuit just as was r-f voltage from the antenna transformer in the earlier hookup.

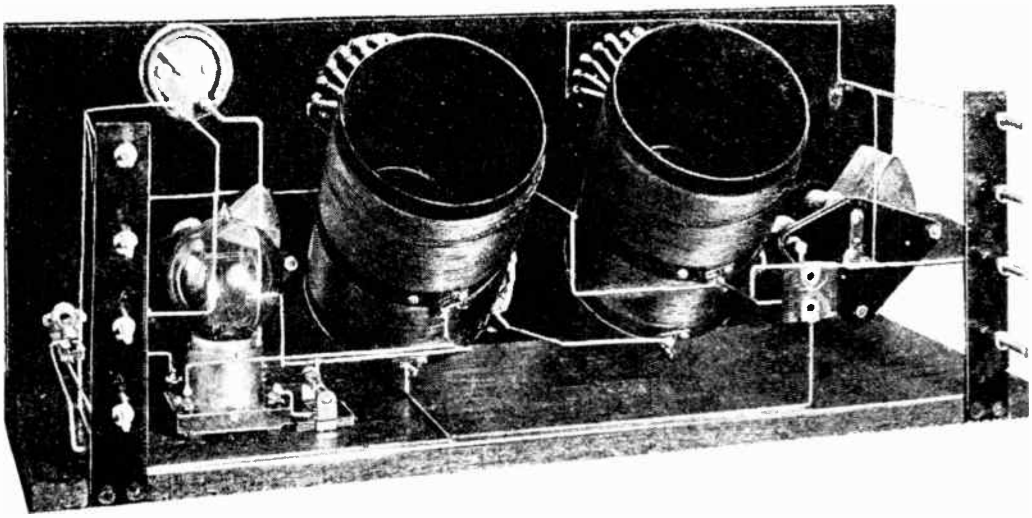


Fig. 31-1. A one-tube receiver of 1920. The coils and variable capacitors for tuning the antenna circuit are connected to the grid-leak detector at the left end of the baseboard.

The amount of amplification in the r-f tube depends on the type of tube and on the voltages at which it is operated. There should be very considerable amplification or increase of modulated r-f voltage between input and output of the tube. There may be some additional voltage gain in the secondary of the r-f transformer.

If there is close coupling between primary and secondary of the r-f transformer, and low-loss construc-

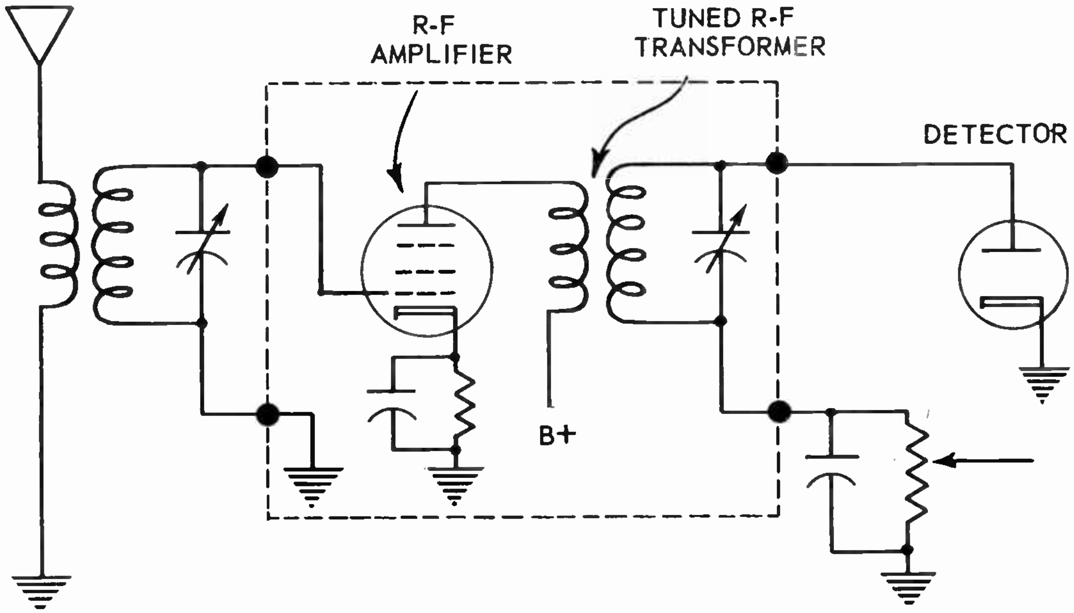


Fig. 31-2. The parts added for r-f amplification are within the broken lines.

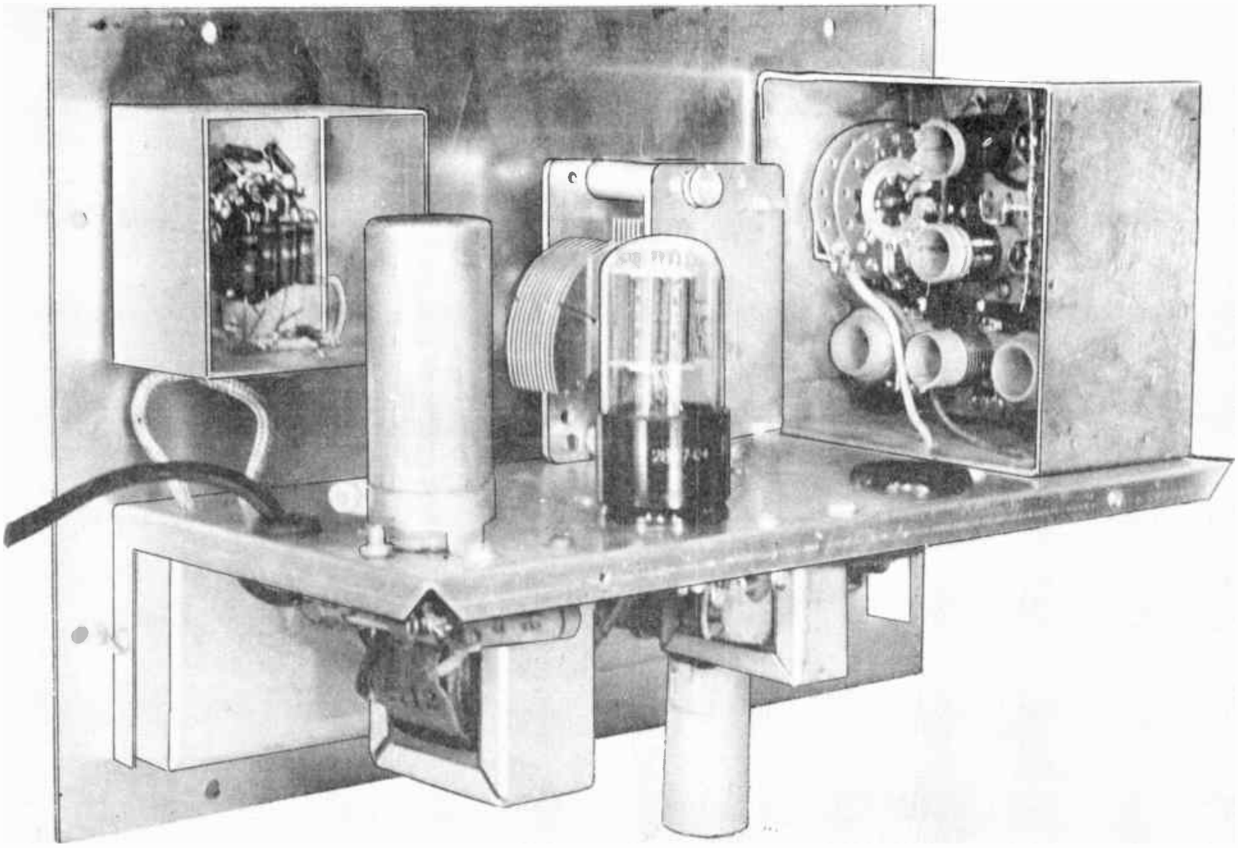


Fig. 31-3. The tuning coils which are exposed at the right in this small signal generator are of high-Q construction.

tion in the whole transformer, there is likely to be worthwhile gain. With loose coupling, poor construction, or both, there may be no gain or there may be an actual loss of voltage in the transformer. But in any case, due to amplification in the tube, we should come to the input of the detector with a much higher r-f voltage than before the r-f stage was added.

In a receiver of ordinary commercial design and construction there might be a voltage amplification of 6 times between grid and plate of the r-f tube. There might be gain of 1½ times in the r-f transformer. Then total voltage gain of the r-f stage would be 6 times 1½, or 9 times in all. Even were there no gain in the transformer we still should have the tube amplification. Were there a slight loss of voltage in the transformer we would come to the detector with less gain than that of the tube alone, but still with more signal strength than from the antenna transformer secondary before the r-f stage was added.

Whatever voltage gain may be had in the r-f stage will add to the sensitivity of the receiver. This will permit good reception from stations much farther away and there will be more volume from stations which formerly were weak. At the same time there will be a decided improvement in selectivity. Instead of hearing two or three programs at once you will be able to cut out all but the one you wish to hear, unless

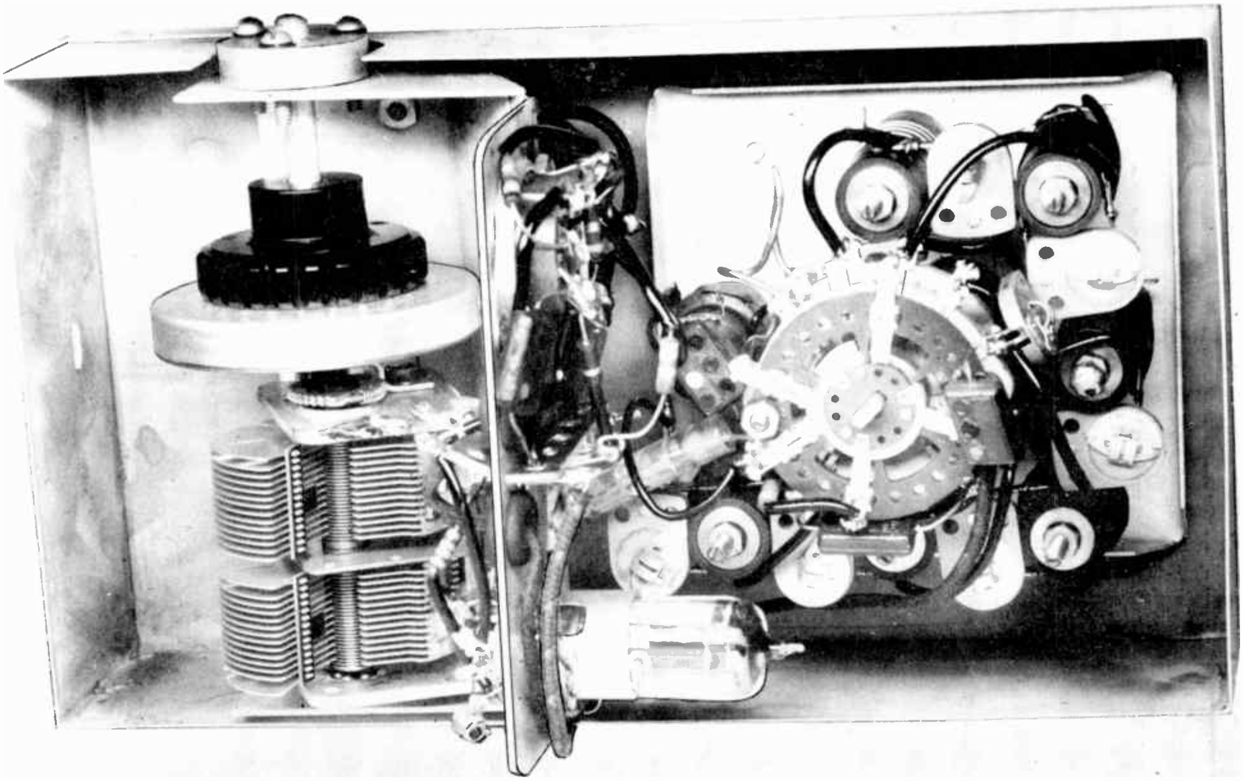


Fig. 31-4. The tuning section of a signal generator capable of producing frequencies in excess of 100 megacycles. High-Q construction is essential at such frequencies.

station frequencies are very close together. How much or how little improvement is realized will depend to a great extent on how well the r-f transformer is designed and constructed. This is a matter which we must investigate.

What we do to the r-f transformer to increase gain and selectivity can be done also to the antenna coupling transformer, with equally beneficial results. The same things can be done to any other high-frequency transformer which may be added in later steps. We shall try to increase the "Q" of the circuits and parts.

Q-FACTOR. Radio and television technicians often speak of high-Q and low-Q radio-frequency circuits and parts of such circuits. They talk about obtaining higher "Q" with certain changes, or about making the "Q" lower with other changes. The "Q" or the Q-factor of a part of a circuit is the ratio of its reactance to its high-frequency resistance. Using the ordinary symbols for reactance and resistance, the Q-factor is equal to X/R, or to reactance in ohms divided by resistance in ohms. We shall find that high-Q is all-important in selectivity and in voltage gain of resonant circuits.

(6) With no further explanation than the mere statement of the ratio, X/R, anyone might assume that a coil with inductive reactance of 1,000 ohms and wire resistance of 2 ohms would have a "Q" of 500, which is the reactance divided by the resistance. More probably the "Q" would be around 100, maybe less. The explanation is that Q-factor is defined as the ratio of reactance to high-frequency resistance. High-frequency resistance is quite different from ordinary ohmic resistance of the wire with which the coil is wound.

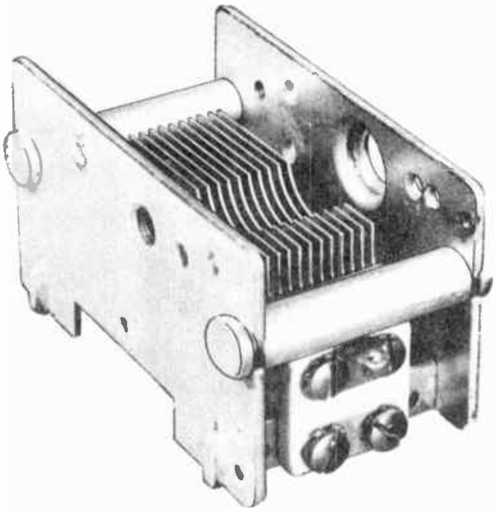


Fig. 31-5. Energy losses in tuning capacitors depend largely on quality, position, and spacing of insulation that supports the stator plates.

High-frequency resistance refers to every waste of signal energy and takes into consideration every energy loss which may occur. The sum of all these energy losses is spoken of as resistance, because the same waste of signal energy could be had by connecting a resistance in series with or in parallel with a circuit which, in itself, had no loss of energy. The number of ohms in this equivalent real resistance would be the number of ohms of high-frequency resistance in the actual circuit.

Let's get down to the real significance of this ratio of reactance to high-frequency resistance, the ratio that we call "Q". First consider the high-frequency resistance, which is a resistance that would cause energy loss equal to all kinds of energy loss put together. It goes without saying that nearly always we shall wish to conserve signal energy, not waste it. The less the waste of signal energy the more acceptable will be the performance of an inductor, a capacitor, or an entire circuit.

A small loss will mean a small value of the term R in the ratio X/R. The smaller the value of R the larger becomes the ratio. That is, the ratio of 1 to 2 is larger than the ratio of 1 to 4, or 1/2 is larger than 1/4. Then a small value of R gives a high Q, because Q equals X/R. Thus we find that high-Q means the same as low-loss.

Now about the X part or reactance part of the ratio. Consider first a series resonant circuit from which we expect to obtain, across either the inductor or the capacitor, a signal voltage greater than the voltage applied across the whole circuit. The larger the reactance of the inductor or capacitor the greater will be the voltage across it for any given current through it. So we should like to have the highest possible reactance.

Next consider a parallel resonant circuit whose prime purpose is to provide high impedance at the frequency of the applied voltage. This impedance is increased by greater inductive reactance or greater capacitive reactance. So again we wish to have the highest possible reactance.

The greater the reactance of the inductor, the capacitor, or both, the higher is the value of "Q", because the X that means reactance is the top term of the ratio or fraction. The larger the top term the larger is the

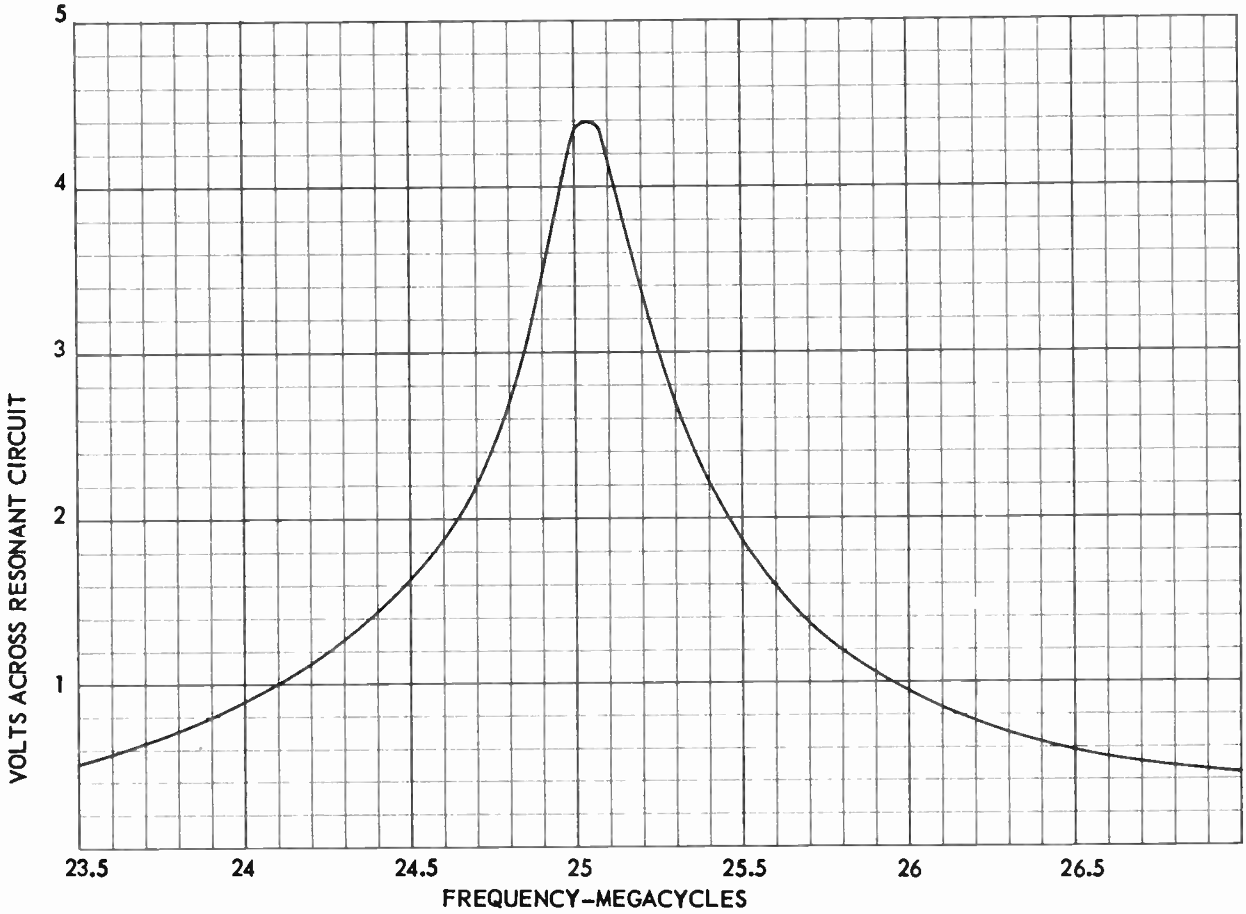


Fig. 31-6. Voltage resonance curve of a tuning coil used in television receivers.

Q-factor. For example, 3/4 is larger than 1/4. High-Q means that we shall obtain high voltage or high impedance with relatively small loss of signal energy. Low-Q means lower voltage or less impedance in comparison with whatever losses of energy may exist.

ENERGY LOSSES. When discussing the construction of inductors and capacitors in an earlier lesson there were mentioned some of the things which allow waste of signal energy. These included distributed capacitance, skin effect, eddy currents, and others.

High-frequency resistance includes everything that allows signal energy or power to change into heat instead of progressing on through the circuits. Heating effect in receiver circuits won't be enough to notice as a rise of temperature – at least not the heating due to waste of signal energy. But remember, we start from the antenna with only a few millionths of a watt of power. All of it could change to heat without causing an apparent rise of temperature, though the power and energy would be wasted.

High-frequency resistance includes the ordinary ohmic resistance of conductors, for this kind of resist-

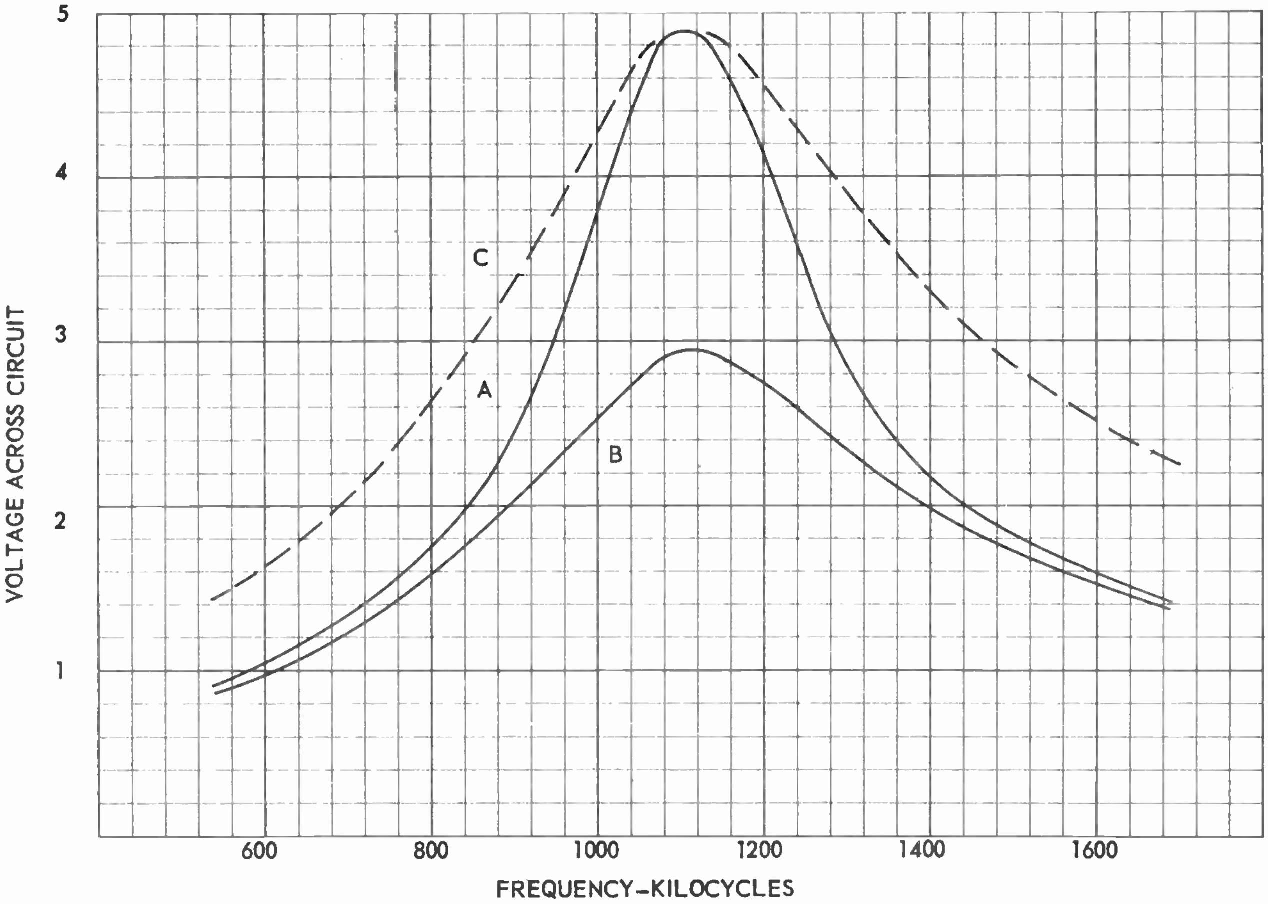


Fig. 31-7. Effect of resistance in parallel with a tuning coil and capacitor. Curve B shows the reduction of gain. Curve C shows the loss of selectivity.

ance changes electrical energy into heat. This ohmic resistance is, however, a small part of the total high-frequency resistance, which usually is 10 to 100 times the ohmic resistance.

Distributed capacitance causes energy loss because high-frequency currents flow through this capacitance, as through any other kind, and in conductors of the circuit these currents produce heat. Furthermore, the distributed capacitance and the coil inductance may be resonant at some high frequency other than the tuned frequency. Then there will be large currents at this self-resonant frequency, and these currents will cause much heating and energy loss.

Now about skin effect. The magnetic field of a conductor carrying current exists inside as well as outside the conductor. In fact, this field is most concentrated at the center of the conductor. High-frequency magnetic fields induce strong counter-emf's inside the conductor. The counter-emf's oppose the applied voltage, with the result that most of the current can flow only near the outside or only in the "skin" of the conductor where applied voltage meets less opposition. So far as flow of current is concerned the size of the conductor is reduced, and the greater part of the current in the reduced cross sectional area produces

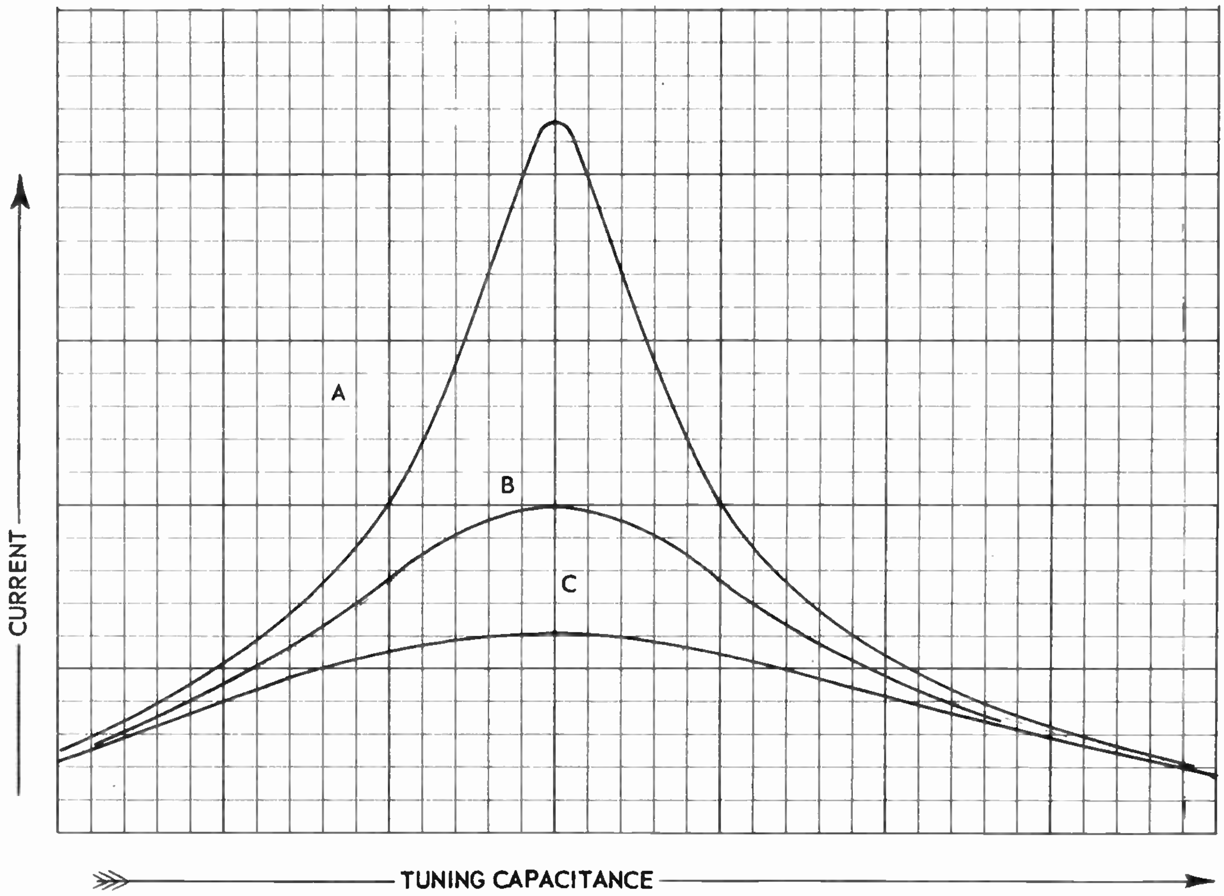


Fig. 31-8. How current is affected by added resistance in a series resonant circuit.

more heat. This skin effect loss increases with rise of frequency, but, fortunately, only in proportion to the square root of frequency, not directly with frequency.

The smaller the diameter of a wire the greater is its surface in proportion to total volume of metal in the wire. As a consequence, skin effect is less in small wires than in larger ones. At 1,500 kilocycles the skin effect in number 32 copper wire is only 1/10 as bad as in number 14 wire. Skin effect is worse in close wound coils than in inductors of more open construction. Skin effect in capacitors is negligible, because the conductors have such large surfaces in comparison with their volumes.

There is what we call proximity effect in conductors that are close together. Magnetic fields from one conductor induce emf's in the other. If currents in the two conductors are in opposite directions at the same instant, both currents are forced into portions of the conductors that are nearest each other. Then there is increased high-frequency resistance in both conductors. In a bundle of wires or in stranded wire most of the current is forced to flow in the outer strands.

Eddy currents are induced in and circulate within any conductor which is being cut by magnetic field lines. Energy to keep these currents flowing must come from the coil or circuit in which the magnetic field originates. Eddy current loss is greatest in iron or steel, because of the rather high ohmic resistance of these metals.

There are energy losses also in all insulating and dielectric materials reached by alternating electric fields. As the force of the fields pulls electrons one way and the other in the atoms there is heating of the material. This action is employed to good advantage in industrial electronics for heating the inner parts of non-conductors, as when setting the plastic cement between layers of plywood used in airplane construction. But the effect isn't any advantage in radio and television.

Were it possible for a capacitor to have a perfect dielectric the current during charge and discharge would lead the voltage by 90°. But when there is energy loss, which has an effect similar to resistance, the lead is less than 90°. Then some of the energy used during charging is not returned to the circuit during discharge, but acts instead to produce heat in the dielectric.

The percentage of total power thus wasted in heating is called the power factor of the capacitor or of the dielectric material. The power factor or percentage of loss is the same for a given material whether it is being used as the dielectric in a capacitor or as an insulating support. Dielectric power factors in mater-

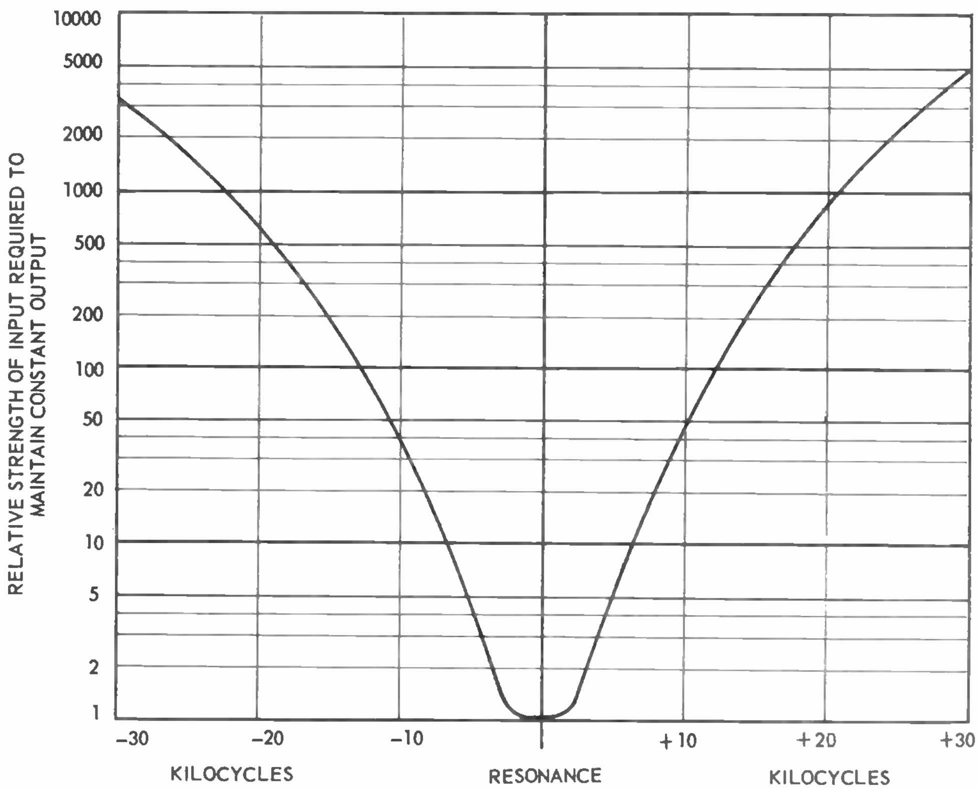


Fig. 31-9 A selectivity curve showing signal strengths required to maintain a constant output as tuning of an amplifier or receiver is varied.

ials used around high-frequency circuits range from about 0.03% up to as much as 3%. Polystyrene is the best performer so far as dielectric power factor is concerned.

Quite often you will hear references to dielectric loss factor. This factor is equal to the product of dielectric power factor and the dielectric constant of the material. For example, steatite with a power factor of 0.2% and dielectric constant of 5.0 would have a dielectric loss factor of 1.0%.

Q-FACTOR AND FREQUENCY. There are more kinds of energy losses and larger losses associated with coils or inductors than with capacitors. Associated with inductors and their magnetic fields we have distributed capacitance, skin effect, eddy currents, and proximity effects. Skin effect is of no importance in capacitors. Eddy currents may exist, but they derive their energy from nearby inductors. With well designed and well constructed capacitors the dielectric losses and the leakage are small.

Because most of the energy losses and most of the high-frequency resistance are in the inductors, the "Q" of a resonant circuit depends chiefly on the "Q" of its inductor. Greatest improvement in circuit "Q" results from improvement in the inductor. The Q-factors of inductors in general use range between 100 and 800.

4 All kinds of high-frequency resistance increase with increase of operating frequency, so the R term of the ratio becomes larger at high frequencies. But inductive reactance increases directly with frequency, which makes the X term of the ratio become larger at high frequencies. With both terms increasing or decreasing together, the value of the ratio, or the "Q" of the inductor, may remain about the same throughout a considerable variation of frequency.

9 The Q-factor of a capacitor must always decrease as frequency rises. As frequency goes up, capacitive reactance becomes less and the X term of the ratio gets smaller. At the same time there is increase of high-frequency resistance with rise of frequency, and the R term of the ratio gets larger. Then the value of the ratio, or of the Q-factor, becomes progressively less.

SELECTIVITY. Selectivity, as mentioned before, is the ability of a circuit, an amplifying stage, or an entire receiver, to select a single frequency and its modulation while more or less completely excluding other frequencies. When studying resonance in earlier lessons we examined curves showing voltages developed in resonant circuits at various frequencies. We always found much higher voltage or greater response at the resonant frequency than at frequencies below and above resonance.

7 A curve showing relations between applied frequency and the signal voltage or current developed in a circuit is called a resonance curve for that circuit. The sharper the peak and the steeper the sides of a resonance curve the greater is the selectivity, because this means greater signal voltage at the selected resonant frequency and less voltage at all other frequencies.

Fig. 31-6 is a voltage resonance curve showing performance of a small coil such as used in television intermediate-frequency amplifiers. The coil was tuned to resonance at a frequency slightly above 25 mc, using an air-dielectric tuning capacitor to represent circuit and tube capacitances which would do the tuning in an actual television receiver. The voltage rises much higher at resonance than at frequencies either side of resonance. A signal at the resonant frequency would be strengthened much more than signals at other frequencies. This indicates good selectivity.

Fig. 31-7 shows the effect of resistance in a resonant circuit on both gain and selectivity. The tests whose results are shown here were made with an antenna coupling transformer of a type often used in small broadcast receivers for sound radio. An adjustable capacitor was used to tune the transformer secondary winding to resonance at about 1,100 kilocycles. Curve *A* shows voltage developed across the secondary at

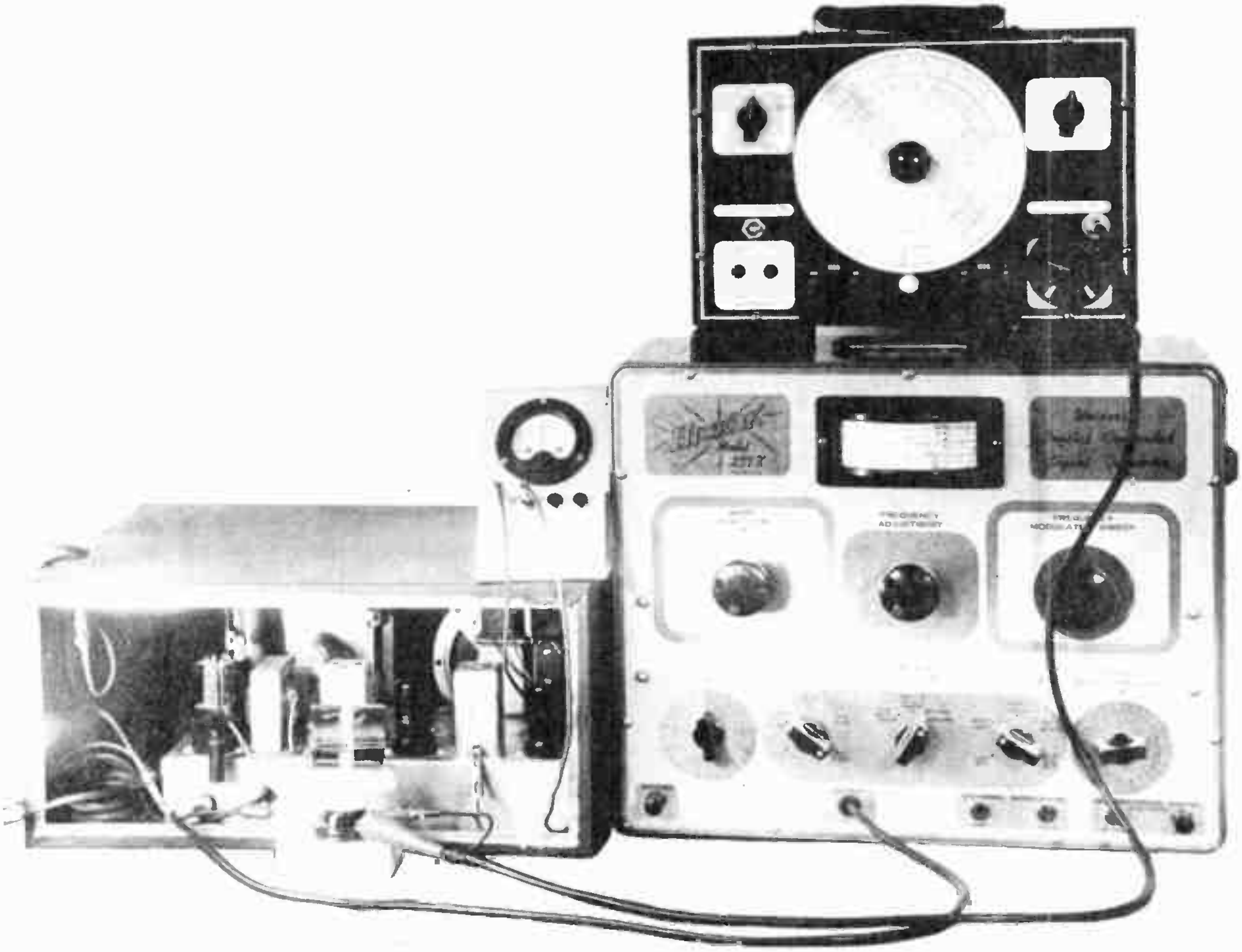


Fig. 31-10. Measuring selectivity by means of two signal generators, one of which furnishes an interference frequency.

frequencies between 600 and 1,000 kilocycles with constant voltage input to the primary winding from a signal generator. Then a resistor of 15,000 ohms was connected in parallel with the secondary winding and the tuning capacitor. The result on developed voltage is shown by curve *B*. Maximum voltage is much lower, which indicates a considerable reduction of gain.

To compare the selectivity with and without the paralleled resistor it is necessary to multiply all the voltage values on curve *B* by a factor which makes the peak voltage equal to the peak of curve *A*. This is equivalent to turning up the volume control of a receiver. The result is shown by broken-line curve *C*. This curve is much less sharply peaked than curve *A*, showing that selectivity has been reduced by the resistance.

A similar reduction of selectivity and gain would result from resistance connected in series with the transformer winding or with the tuning capacitor. The fixed resistor used for testing is the equivalent of high-frequency resistance in the resonant circuit, since there is dissipation or loss of energy in the resistor just as there would be in any kind of high-frequency resistance.

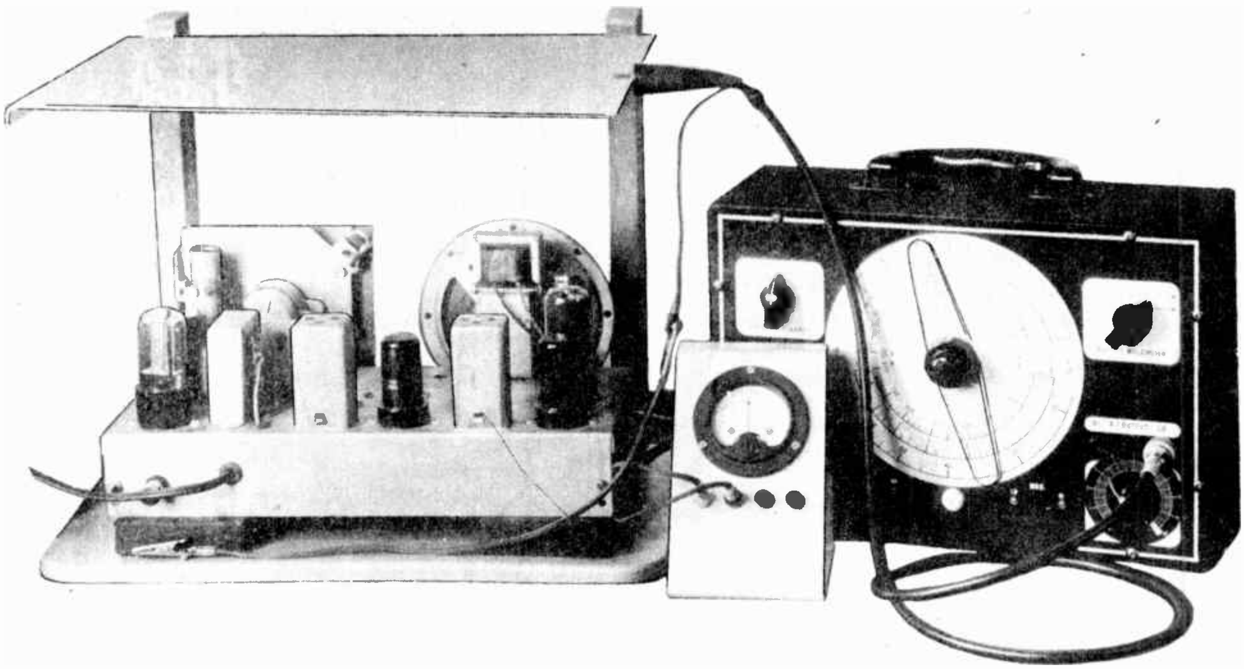


Fig. 31-11. Selectivity is affected by signal pickup on wiring and parts of a receiver. This is a test of such pickup in an r-f field.

Fig. 31-8 illustrates the effect of resistance in a series resonant circuit. The curves show changes of current in the resonant circuit as the tuning capacitance is varied. Curve *A* was made with the coil and capacitor alone, which had a measured resistance of approximately 5 ohms. Curve *B* was made with a 5-ohm fixed resistor added in series with the resonant circuit, to double the resistance. Curve *C* was made with a 10-ohm added series resistor. The added resistors, which are equivalent to added high-frequency resistance, cause serious reductions of gain and selectivity.

So far, we have discussed selectivity with reference to resonance curves. Another method of showing selectivity is illustrated by Fig. 31-9. Here we have what is called a selectivity curve, not a resonance curve. The selectivity curve shows relative strengths of input signal, at frequencies below and above resonance, required in order to maintain the same output as at resonance.

In making a selectivity curve the tested circuit or amplifier is tuned to resonance at any frequency, usually at the center frequency of some broadcast channel. A signal generator used for supplying the input voltage is tuned to the same frequency. Output of the circuit or amplifier is measured by means of any suitable voltmeter. To begin with, the signal generator voltage is made as low as will allow obtaining a distinct indication of resonance on the voltmeter.

Then the circuit or amplifier is detuned to some frequency slightly below resonance, without changing the signal generator frequency, and the generator voltage is increased to bring the output back to its original value or to bring the reading of the voltmeter to its original value. This process is continued, with the circuit or amplifier tuned further and further from resonance while noting the number of times the generator voltage must be increased at each frequency to keep the output constant.

These multiples of the original generator voltage at each frequency are shown by the lower-frequency side of the selectivity curve. Next, the circuit or amplifier is tuned through many frequencies above resonance while generator voltage again is increased as required to maintain the original output. These multiples of the original generator voltage are shown by the higher-frequency side of the curve.

Fig. 31-9 shows selectivity as measured on a small receiver for standard radio broadcast. In the standard broadcast band there are adjacent channels 10 kilocycles below and 10 kilocycles above that of the center frequency at which the circuit or amplifier was first tuned to resonance. At 10 kilocycles below resonance the curve shows that about 40 times as much input is required, and at 10 kilocycles above resonance nearly 50 times as much, to produce the same output as at resonance. These relative strengths of input signal are a measure of "adjacent channel sensitivity."

The second channels are 20 kilocycles below and above the first channel. Our selectivity curve shows at 20 kilocycles below resonance an input signal about 600 times, and at 20 kilocycles above resonance an input almost 900 times as great as at resonance for equal outputs. These are measures of "second channel selectivity." Selectivity curves have to be drawn with a logarithmic vertical scale in order to cover the great change of input voltage while clearly showing conditions near resonance on a graph of any reasonable size.

A selectivity test as just described does not duplicate the conditions of actual operation, because there is only a single input signal at any one time. In practice, the selectivity or lack of it shows up when a second interfering signal is trying to come in along with the signal you desire.

Actual conditions are more nearly duplicated by using two signal generators, as in Fig. 31-10. One gen-

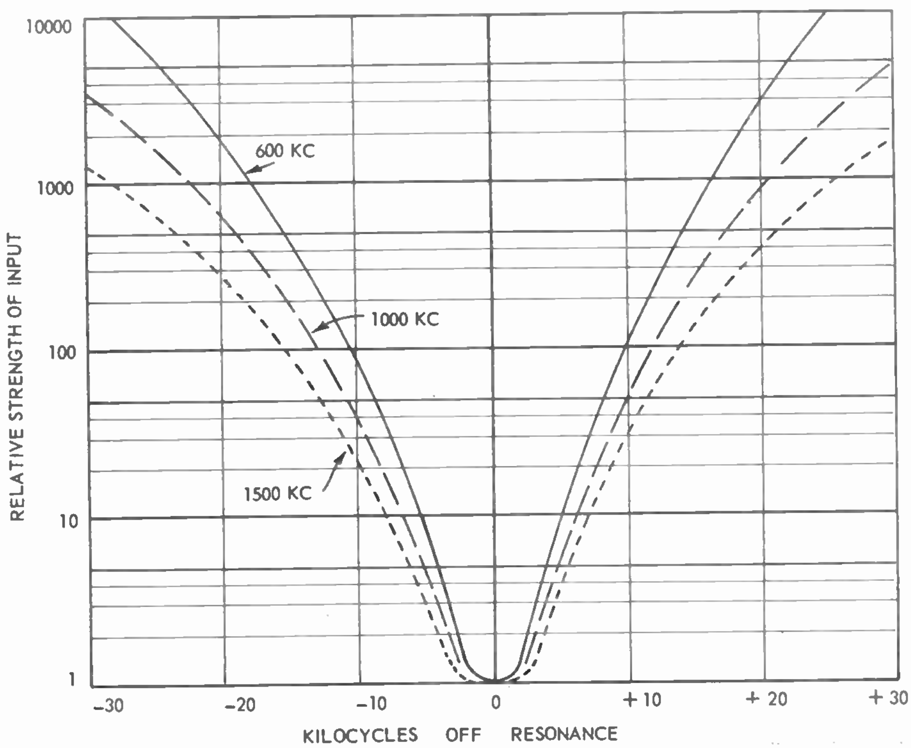


Fig. 31-12. Selectivity is better at low signal frequencies than at high ones. These curves show selectivity of a certain broadcast receiver at three center frequencies.

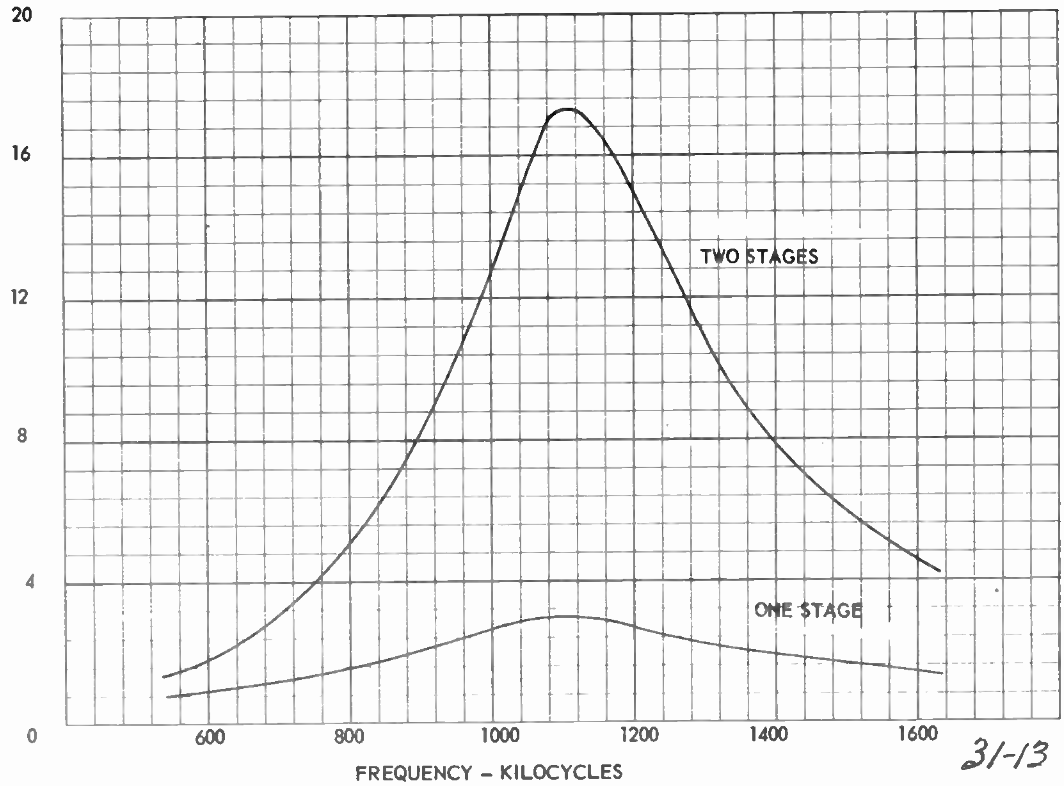


Fig. 31-13. How gain and selectivity would be improved by using two similar r-f amplifying stages instead of one stage.

erator supplies a signal at the frequency to which the receiver is tuned. The second generator supplies a signal at any other "interfering" frequency. Signals from both generators are fed to the antenna input terminals of the receiver being tested. The receiver in the picture is enclosed within a copper box which forms a shield against any signals not intentionally introduced from the generators. Interfering radio frequencies are removed from the power line connection by a filter at the left of the receiver. Receiver output is measured by a type of output meter which responds to the audio signal or to the modulation used on the r-f voltages from the generators.

After both signal generators are adjusted to give equal readings of the output meter at the tuned frequency, modulation is turned off in the generator giving the desired frequency but its r-f signal is allowed to remain. Modulation remains turned on in the generator furnishing the interfering signal, to cause readings on the output meter.

Now the interference signal generator is tuned to various frequencies below and above the tuned resonant frequency of the receiver. At each frequency the interfering voltage is increased to bring the output meter reading back to its original value. The number of times the interference voltage must be increased at each frequency is used in drawing a selectivity curve like that of Fig. 31-9.

10 Still another kind of selectivity test is pictured by Fig. 31-11. This is a check on the degree of interference pickup by wiring and other parts of the receiver. The receiver is mounted between but insulated from two metal plates to which are connected the leads from the signal generator. In the space between the plates, and around the receiver, will exist a radio-frequency electromagnetic field similar to that from inter-

fering radio stations. The field frequency and strength are controlled by adjustment of the signal generator. Receiver response may be measured by any suitable type of output meter.

With the signal generator tuned to various frequencies its output voltage is made such as will maintain the same receiver output as at the tuned frequency for desired reception. These voltages are plotted on a selectivity curve.

You will find that the selectivity of any receiver is better when tuned to low frequencies than to high ones. Fig. 31-12 shows three selectivity curves made from the same broadcast receiver. The full-line curve was made by tuning 30 kc below and above 600 kc, with the signal generator remaining at 600 kc. The curve with long dashes was made with a center frequency of 1,000 kc. The short dashed curve was made with a center frequency of 1,500 kc. These are the three frequencies usually tested in the standard broadcast band.

At 10 kc below resonance an interfering signal has to be about 90 times as strong as at resonance on the 600 kc curve, but on the 1,500 kc curve the interfering signal has to be only a little more than 20 times as strong as at resonance to produce equal outputs. At 20 kc below resonance the interfering signal on the 600 kc curve has to be nearly 2,000 times as strong, but on the 1,500 kc curve only 290 times as strong as at resonance. All this comes about because high-frequency resistance or energy loss is much greater at 1,500 kc than at 600 kc.

Selectivity is increased also by passing a signal through additional resonant circuits tuned to the same frequency. To illustrate what would happen with more tuned circuits we shall first go back to curve *B of* Fig. 31-7, the curve which shows the poor selectivity and reduced gain resulting from resistance in a tuned circuit. Assume that curve *B* shows voltage output from a stage of r-f amplification when the input signal strength is constant at 0.5 volt for all frequencies. At resonance, around 1,100 kc the output is nearly 3.0 volts, so at resonance there is gain of nearly 6 times. The gain is less at other frequencies.

The output voltage shown by curve *B* might be fed into another amplifying stage of exactly the same characteristics, becoming the input for this second stage. Assuming that this second stage has no more gain and no more selectivity than the first one, the signal at each frequency would be amplified to the same degree as in the first stage.

The result of adding the second stage of tuned r-f amplification is shown by Fig. 31-13. The curve at the bottom represents the same voltage at each frequency as was shown by curve *B* in Fig. 31-7. The reason the curve appears even flatter than before is because we are using a different voltage scale, a scale that extends from zero to 20 volts rather than to only 5 volts as in the earlier figure. With this lower curve representing the input for the second stage of amplification, the output voltage from that second stage would be as shown by the upper curve.

The added stage of r-f amplification has, by itself, just as little gain and just as poor selectivity as the first stage. Yet the overall result shows much better selectivity and, naturally, shows much more total gain in voltage. Selectivity still isn't at all good, but it is a great improvement over that of a single amplifying stage. Were both stages to be of high-Q construction, with good selectivity in both, the overall selectivity would be very good.

Now that we have become acquainted with the relations between selectivity and any resistances or energy losses in resonant circuits it will be easier to understand the effects on selectivity of using different types of detectors. Connecting the output side of a tuned circuit to a diode detector has the same effect on selectivity as connecting a resistor across the tuned circuit. This is true because there is flow of current in the diode detector just as there is flow of current in the resistor, and the effects are alike. There would be similar loss of selectivity in a resonant circuit connected to a grid-leak detector, because a detector of

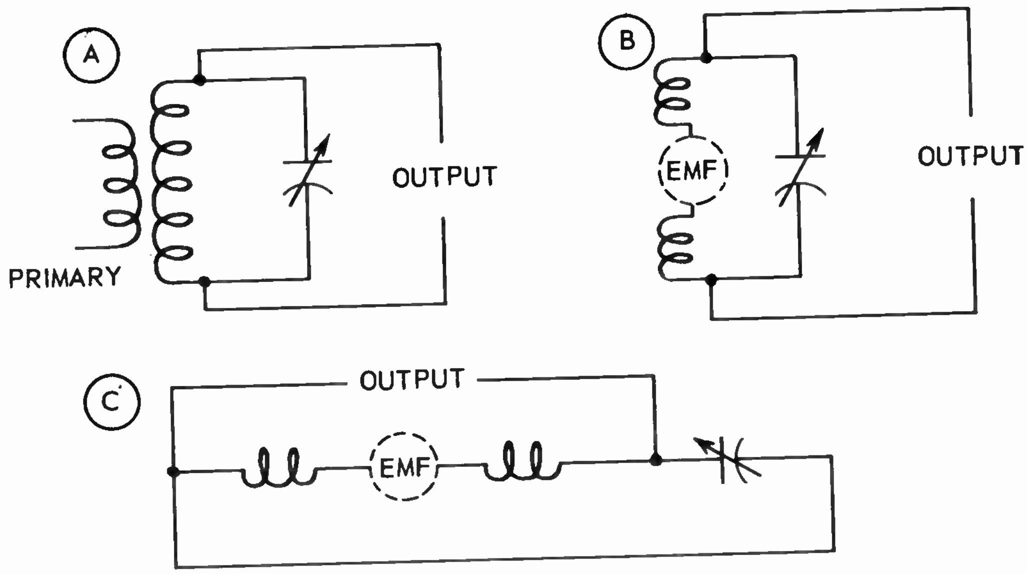


Fig. 31-14. The secondary circuit of a tuned r-f transformer is really a series resonant circuit.

this type allows current to flow in its grid circuit. But a plate detector, which allows no current in its grid circuit, would not lessen the selectivity.

Especially when working with television receivers you will be struck by the great difference between selectivity of transformers connected to the video detector, a diode type, and the selectivity of similar transformers in stages ahead of the detector. There is the same effect in broadcast sound receivers which employ diode detectors.

You will discover too that a transformer whose primary winding is in the antenna circuit usually is less selective than a similar transformer whose primary is in the plate circuit of an amplifier tube. This is because antennas and their connections have rather large energy losses, and the effect of the high-frequency resistance in the primary circuit is carried over into the secondary by the mutual inductive coupling between primary and secondary.

TUNED R-F TRANSFORMERS. When studying power transformers we learned about turns ratio and voltage ratio, and learned that there may be a step-up of voltage from primary to secondary. This relation between turns ratio and voltage ratio is due to the very close coupling between primary and secondary in such transformers. There is close coupling because the iron core provides a closed magnetic circuit which confines the magnetic field lines so that almost all the lines from the primary cut through the secondary turns.

In air-core r-f transformers the coupling is quite loose. The magnetic field of the primary winding spreads out and many of its lines fail to cut through the secondary turns. In r-f transformers having powdered iron cores the coupling is closer than with air cores, but because the powdered iron core does not form a closed magnetic circuit there is spreading of the field and many of the lines from the primary fail to cut the secondary.

In r-f transformers there is no direct relation between primary to secondary turns ratio and the voltages in primary and secondary. Were the voltage ratio to be the same or nearly the same as the turns ratio, using

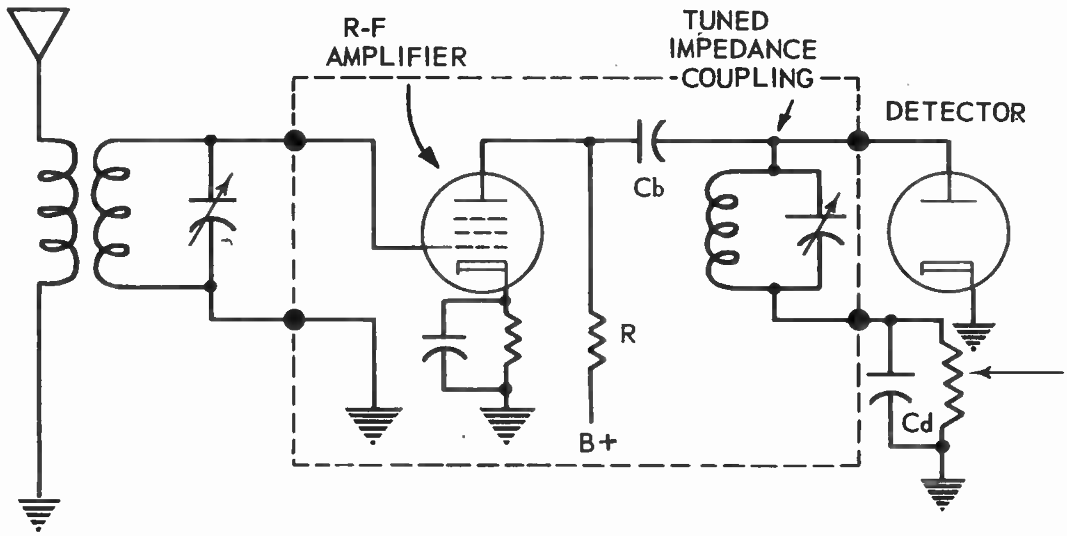


Fig. 31-15. A tuned impedance coupling between the tuned antenna circuit and the diode detector.

fewer turns in the primary would mean higher secondary voltage in relation to primary voltage. But with fewer primary turns you will have less secondary voltage, because you are reducing the coupling and the transfer of energy.

More primary turns with the same number of secondary turns will increase the secondary voltage in relation to primary voltage. This is because the bigger primary allows closer coupling with the secondary, and greater transfer of energy to the secondary. Also, more primary turns increase the primary inductance, and more inductance in either winding or in both will increase the coupling and the transfer of energy.

When a primary winding is coupled to a tuned secondary winding in an r-f transformer there is an increase of apparent inductance in the secondary. The resonant frequency of the secondary is somewhat lower than it would be without the primary present. The added effective inductance is coupled into the secondary from the primary. The effect of added inductance increases with greater inductance in the primary and with closer coupling between the two windings. With usual constructions of r-f transformers the secondary frequency is dropped by five to ten per cent under what it would be without the primary. Consequently, a secondary winding is designed with somewhat less inductance then required for tuning with whatever capacitance is present. The difference is made up by coupling to the primary.

When you look at a diagram showing a tuned r-f transformer, as in Fig. 31-2, the secondary winding and tuning capacitor look like a parallel resonant circuit. But the tuned secondary acts like a series resonant circuit, and is in fact a series resonant circuit.

Fig. 31-14 will show why the tuned secondary is a series resonant circuit. At *A* we have the usual form of diagram, with a primary coupled to the secondary. Changes of current and magnetic field in the primary winding induce an alternating emf in the secondary. This induced emf is the only source of voltage in the secondary circuit. Therefore, as in diagram *B*, we may take away the primary and show the induced emf in series with the coil and capacitor of the secondary circuit. Output voltage is taken from across the coil or from across the capacitor.

Now we have in series with one another the inductance of the coil, the capacitance of the capacitor, and the source of energy, just as in any series resonant circuit. The true state of affairs becomes more evident when we redraw the circuit as at C. Although connections are exactly the same as before, this diagram looks more like a series resonant circuit. Output voltage is being taken from across the inductance, as is usual practice with series resonant circuits.

3 We have learned that voltage across either the inductance or the capacitance of a series resonant circuit is much greater at resonance than is the applied voltage. In the present case the applied voltage is the induced emf. Voltage across the ends of the inductance, and across the capacitance too, should be much greater than the value of the induced emf, and it will be much greater if the Q-factor of the circuit is reasonably high.

The higher the "Q" of the secondary circuit the greater will be the gain with reference to the induced emf. The induced emf is increased by closer coupling between primary and secondary, for then there is greater transfer of energy. As a consequence of all this there is effective gain of secondary output over primary input in a well designed r-f transformer. The voltage gain, or step-up, is not the result of high turns ratio but is the result of close coupling and of resonant gain in the secondary, which is acting like a series resonant circuit.

TUNED IMPEDANCE COUPLING. We commenced this lesson, back in Fig. 31-2, by adding to our original one-tube receiver a stage of r-f amplification in which output of the r-f amplifier is coupled to the detector input by means of a tuned r-f transformer. In Fig. 31-15 we have taken out the r-f transformer and substituted a tuned impedance coupling. Here we have a parallel resonant circuit that not only looks like a parallel resonant circuit but acts like one.

Direct plate current and voltage come to the plate of the r-f amplifier tube through resistor *R* from the B+ line of the power supply. Capacitor *Cb* is a blocking capacitor that allows modulated radio-frequency current from the plate circuit to pass through the tuned impedance coupling, but keeps direct voltage and current out of the coupling. Modulated radio-frequency voltage from across the parallel resonant circuit of the coupler is applied to the detector.

There are really two plate circuits, one for direct voltage and current, the other for r-f voltage and current. The d-c plate circuit is completed through resistor *R*. Here we have high resistance to r-f current as well as to direct current, and most of the r-f current has to pass through capacitor *Cb*. The r-f plate circuit is completed through capacitor *Cb*, the tuned impedance coupling, capacitor *Cd*, the ground connections, and the capacitor across the biasing resistor of the r-f amplifier back to the cathode of this tube.

Impedance to r-f current in the plate circuit is maximum at the frequency to which the parallel resonant coupler is tuned. Since amplification of a tube depends on impedance in its plate circuit there will be maximum amplification at the tuned frequency and relatively little amplification at frequencies below and above resonance, where the impedance is less.

Increase of signal voltage in the r-f stage of Fig. 31-15 is due entirely to amplification of the tube. Nothing is contributed by any resonant gain in the tuned circuit, because there is no such gain in a parallel resonant circuit used to provide maximum impedance at a tuned frequency. That is, there can be no gain in the sense of having greater voltage across either the inductance or the capacitance than the voltage applied to the circuit, as there is with series resonance. The whole object of the parallel resonant circuit is to provide greatest possible impedance in the plate circuit at the frequency to be amplified, and thus to raise the tube amplification as high as possible at this frequency.

Tuned impedance r-f coupling seldom is used in sound radio receivers. In the video i-f amplifiers of television receivers this type of coupling is used more than any other. We shall have much to do with tuned impedance coupling as we get into the details of television receivers.

TELEVISION

LESSON NO. 32

THE SUPERHETERODYNE CIRCUIT

Training Your Mind

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

COMMERCIAL
TRADES
INSTITUTE

Chicago, Ill.

LESSON NO. 32

THE SUPERHETERODYNE CIRCUIT

Before anyone added radio-frequency amplification to a receiver the experimenters tried all kinds of tricks to make the detector more sensitive to weak signals. We had Armstrong regenerators, Bishop ultra-regenerators, Cockaday tuners, Flewelling super-regenerators, Reinartz tuners, and a host of others. Some of these old time hookups are shown by Fig. 32-1. Most of them depended on tuning both the antenna and the detector circuits to get utmost resonant gain, and then they fed some of the energy from the plate circuit back into the grid circuit to give the grid a still greater push on every signal alternation.

Adjustments were too critical for any except the most determined operators to master. Just when the signal came in loudest for ordinary folks it would change into the wail of a banshee in distress. That was because the detector changed itself into an oscillator and began producing signals on its own.

Adding one stage of r-f amplification helped a lot, but not enough. Two r-f stages really would bring in the stations so long as you stayed with the low frequencies, near the bottom of the broadcast band. The minute you tuned to higher frequencies one or both the r-f amplifiers would change themselves into oscillators and produce loud shrieks instead of sweet music from the speaker. The worst of it was that when an amplifier became an oscillator the receiver became a transmitter and every listener in the neighborhood heard your wailing from their own set. The receivers putting on this performance were called "bloopers".

The bloopers blooped because of energy fed back from plate to grid through capacitance within the tubes. In those days there were no pentodes, not even screen grid tetrodes, only triodes. Then came the era of playing tricks on this feedback, of trying to balance the internal feedback with an equal and opposite external feedback. We had the isofarad balancing circuit, the Rice balancing circuit, the Roberts balancing circuit, and, probably best known of all, the neutrodyne sets. For a long time the neutrodyne was just about the last word in radio reception.

The big trouble with all these "tuned radio frequency" receivers was oscillation at the higher broadcast frequencies, with amplifiers changing into oscillators and producing unearthly sounds. The alternative was to hold amplification at a low level. Only slightly less troublesome was the difficulty of tuning. Popular sets, like the one pictured in Fig. 32-2, had three separate tuning dials on the front panel. To tune the first r-f amplifier, the second r-f amplifier, and the detector input all to the same frequency at the same time was a genuine accomplishment, if you could do it. Adding a third r-f amplifier to bring up the output made oscillation practically certain and made tuning next to impossible.

During those days of fighting unwanted oscillation and the intricacies of three tuning dials there was one receiver that didn't become a blooper at even the highest broadcast frequencies, that had a world of amplification, and that could be tuned to any frequency by manipulating only two dials. This was the super-

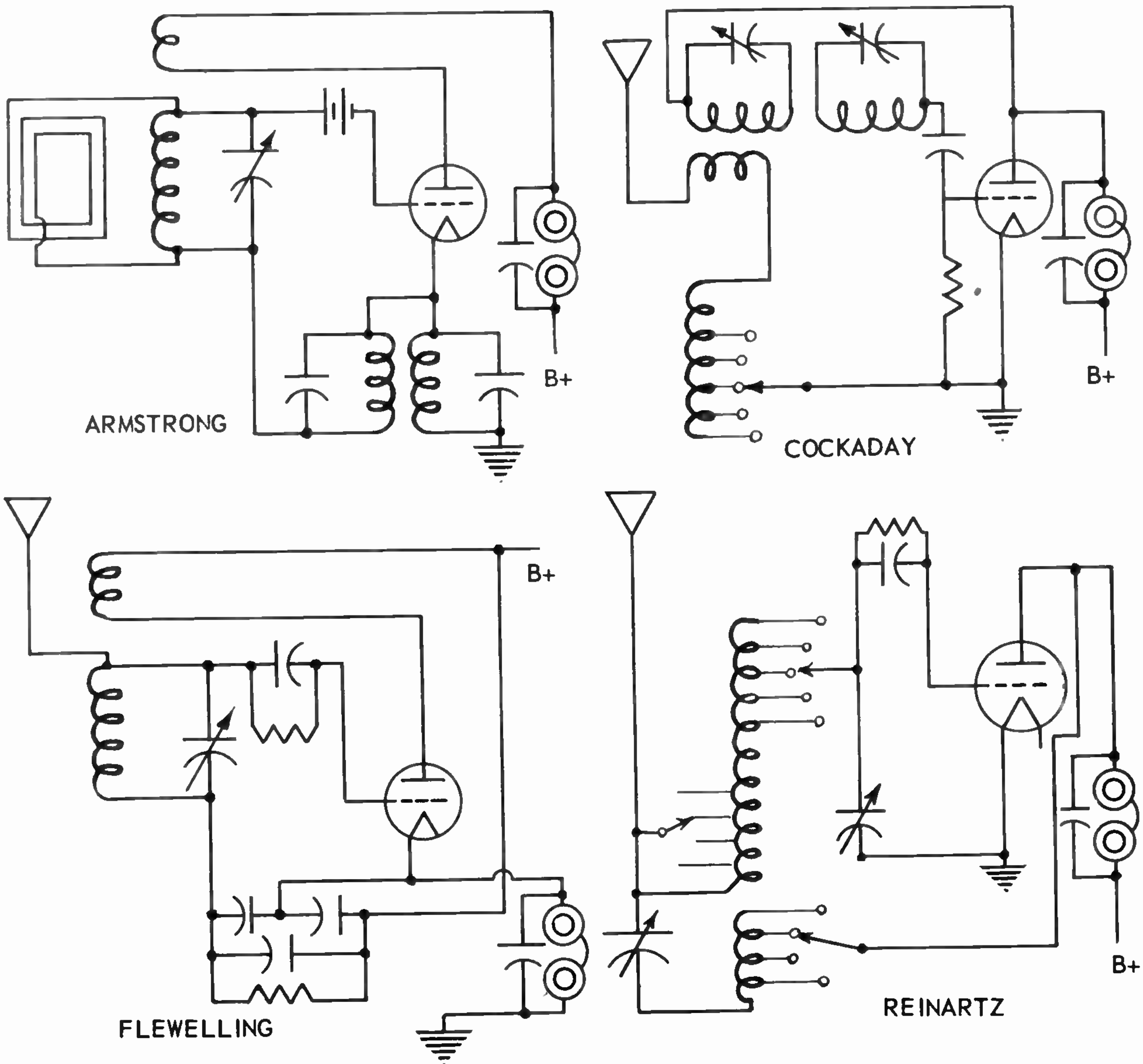


Fig. 32-1. Some of the old-time radio tuning circuits.

heterodyne. Probably everyone would have used the superheterodyne circuit except for the fact that the principle was thoroughly covered by patents.

Nowadays the superheterodyne and its associated improvements, including only a single dial for tuning, may be used under a license agreement with the patent owners. And what has been the result? Every television receiver has a superheterodyne tuner. The front end of every f-m broadcast receiver is a superheterodyne tuner. And except for a few of the smallest every standard broadcast receiver is a superheterodyne.

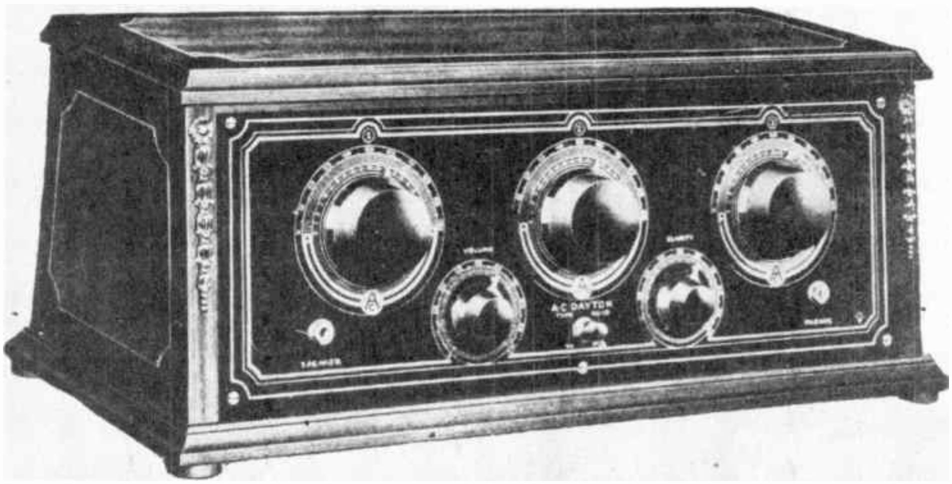


Fig. 32-2. A three-dial tuned radio frequency receiver of 1925.

Obviously we must learn how the superheterodyne, by using an oscillator that oscillates intentionally, does away with oscillation in amplifier and detector circuits and reduces tuning to the utmost simplicity.

The superheterodyne avoids oscillation in amplifying stages by carrying out almost all the amplification at a frequency so low that there is no real tendency to oscillate. This requires changing all the different received frequencies into one lower frequency which is called the intermediate frequency. Because this intermediate frequency remains the same no matter what may be the received carrier frequency all the intermediate-frequency amplifying stages are tuned once, and require no changes when going from station to station or from channel to channel.

3 All the received frequencies are changed to the single intermediate-frequency, still carrying the original modulation, by a process called beating. The incoming signal frequency is made to beat with or mix with another frequency produced by an oscillator which is built into the receiver. The received frequency and the local oscillator frequency beat together in a tube called the mixer. Out of the mixer comes the lower intermediate frequency. By tuning the local oscillator to a frequency suitable for each different received carrier frequency the intermediate frequency is made to remain the same no matter what carrier is being received. Then it is simply a case of amplifying this relatively low intermediate frequency with as many stages as seem desirable.

10 One of the two tuning dials of the early superheterodynes adjusted the frequency of the local oscillator to suit each received frequency. The other dial tuned the antenna circuit or tuned the circuit of the r-f amplifier tube. Operating only the two dials would change any frequency coming from the antenna to the one intermediate frequency, which then could be tremendously amplified by following stages of i-f (intermediate-frequency) amplification. In all later superheterodynes a single tuning dial operates the two capacitors or two inductors for both the local oscillator and the carrier frequency.

Fig. 32-3 is a picture of a television tuner. On top may be seen three miniature tubes. The one at the left is the r-f amplifier. In the center is the mixer. At the right is the oscillator. Immediately below the tube shelf appear some of the coils, resistors, and small fixed capacitors which are parts of the tuned circuits.

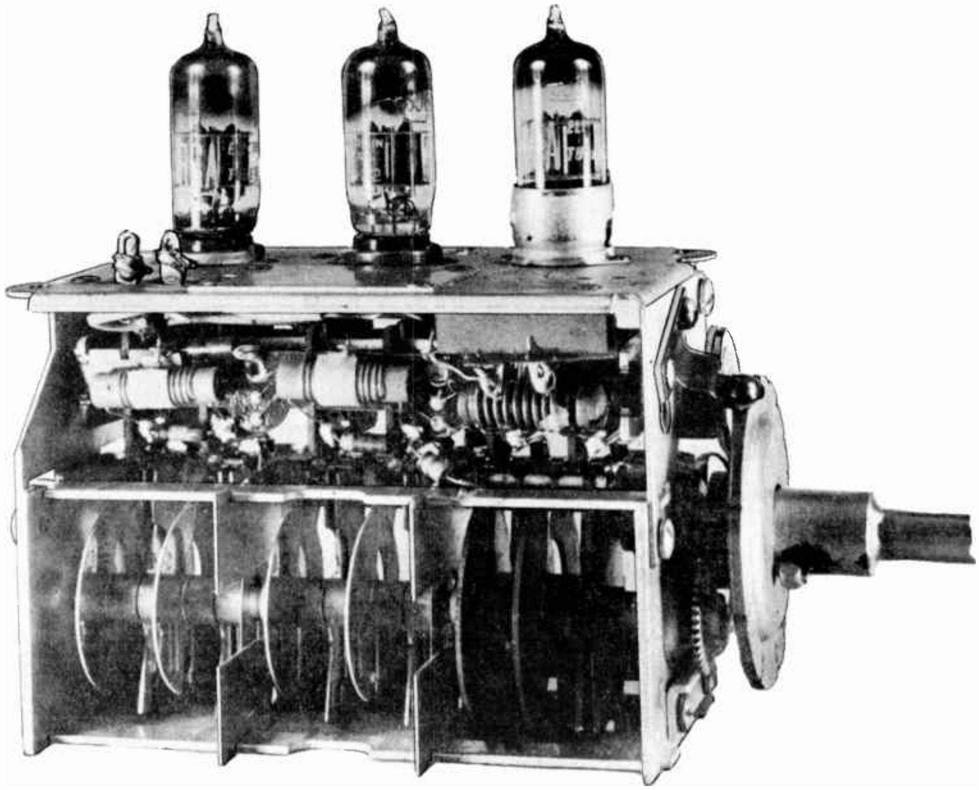


Fig. 32-3. Capacitance tuned television tuner with separate tubes for r-f amplifier, oscillator, and mixer.

The circuits in this particular tuner are made resonant at frequencies for the various channels by means of the variable capacitors whose movable plates are visible in the lower part of the unit. This is only one of the many styles of tuners with which we shall become acquainted.

Before we can discuss the details of tuners and how they utilize the principles of the superheterodyne circuit it will be necessary to understand the fundamental principle of producing the intermediate frequency by beating, and to understand how the local oscillator produces the frequency which beats with the carrier. The method of producing a beat frequency is quite simple, so we shall give it our first attention.

BEAT FREQUENCIES. We shall assume, to begin with, that a received carrier frequency is unmodulated, that it has passed through a stage of r-f amplification and is brought into the grid circuit of our mixer tube. This travel of the carrier frequency voltage is shown by Fig. 32-4. At the same time the oscillator is producing another voltage at somewhat higher frequency, and this oscillator voltage is put into the grid circuit of the mixer. Now we have at the mixer grid or in the grid-cathode circuit of the mixer two different voltages at two different frequencies. One is the carrier signal voltage, the other is the oscillator voltage.

What happens to the two voltages is illustrated by Fig. 32-5. At the top are represented 14 cycles of signal voltage. For purposes of explanation we shall assume that the time is one second, or that the frequency of the signal voltage is 14 cycles per second. Below the signal voltage are represented 16 cycles of oscillator voltage. Since the time, from left to right, is the same for both voltages the frequency of the oscillator voltage must be 16 cycles per second.

How the amplitudes of the two voltages combine is shown at the bottom of the figure. At instant 1 the signal voltage and oscillator voltage are in phase. They add together to produce a voltage or amplitude equal to their sum. So at this instant we have a high peak in the combination voltage down below. Then, due to the difference between frequencies and time periods per cycle, the two voltages get farther and farther out of phase until, at instant 2, they are of opposite phase. Now the combined voltage is only the difference between oscillator and signal voltages, and there is a low peak or small amplitude in the combined voltage.

At instant 3 the signal and oscillator voltages have come back into phase, and there is a second high peak in the combined voltage that is going to the mixer. At instant 4 the original voltages again are of opposite phase, and produce a low amplitude, and at instant 5 they are back in phase to produce a high amplitude. Thus the action continues.

It is perfectly plain that we have at the bottom of Fig. 32-5 a modulated high-frequency voltage. The modulation has resulted not from applying a low modulating frequency to a high frequency, but from beating of two high-frequency voltages. What is the modulation frequency, in cycles per second? The variations of high-frequency voltage commence with a maximum amplitude and go through minimum amplitude back to maximum twice. Then we have two cycles of modulation. The time is one second, so the modulation frequency is two cycles per second. But the difference between 16 cycles per second of oscillator voltage and 14 cycles per second of signal voltage is two cycles per second. It always is true that mixing or beating of two different frequencies produces a new frequency equal to their difference. This difference frequency is called the beat frequency.

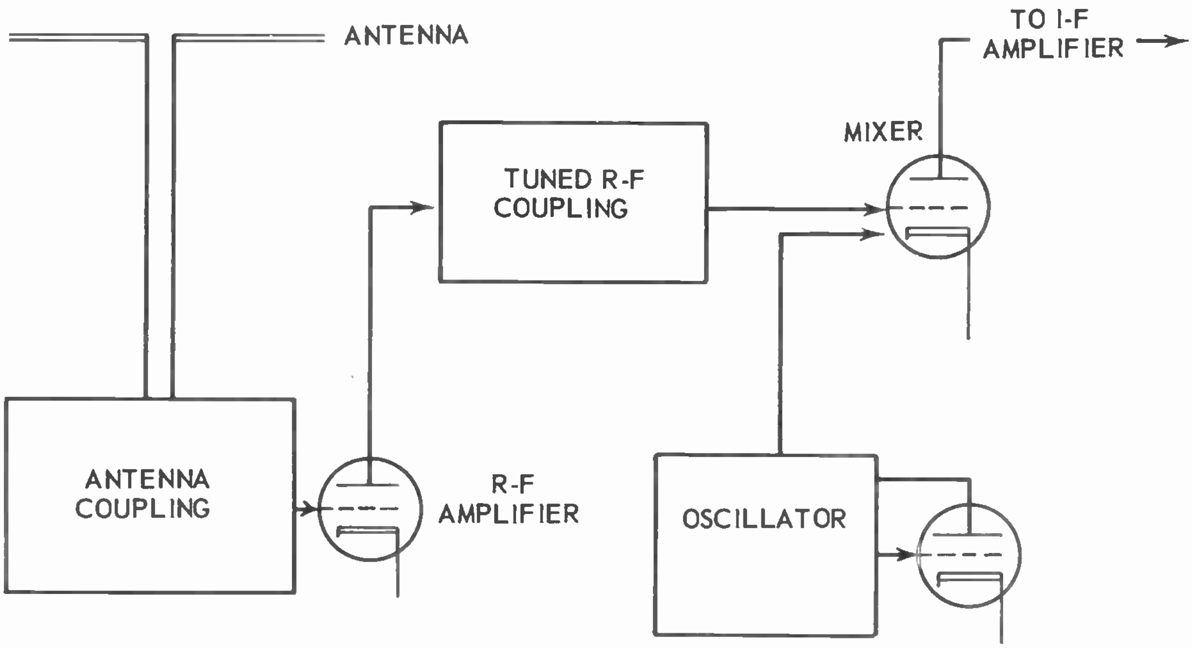


Fig. 32-4. The received signal frequency and the oscillator frequency are fed to the mixer.

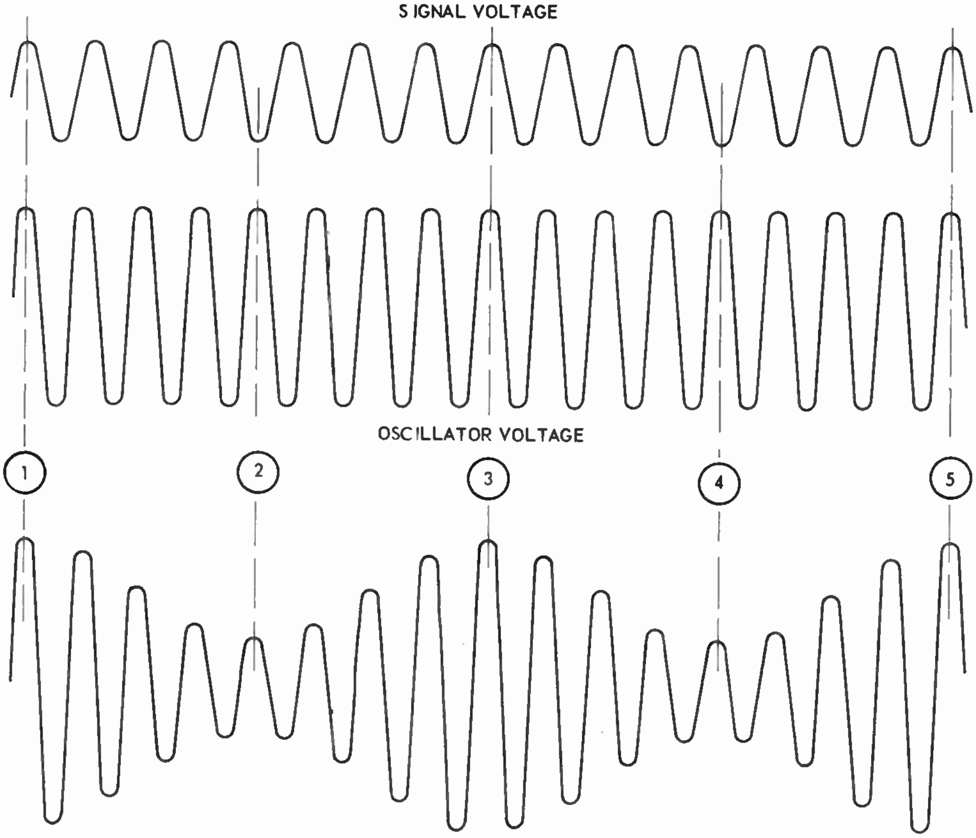


Fig. 32-5. The radio frequency and oscillator frequency combine to produce modulation at the beat frequency.

THE MIXER. In the voltage wave shown at the bottom of Fig. 32-5 we do not have the beat frequency, we have only a high-frequency voltage that is modulated at the beat frequency. To obtain the beat frequency itself we must do with this modulated voltage just what we would do with any other modulated voltage in order to obtain the modulation frequency, which here we are calling the beat frequency. To obtain the beat frequency we put the modulated high-frequency voltage into a detector and take the beat frequency from the output of the detector.

The detector used for obtaining the beat frequency ordinarily is called the mixer. Some manufacturers call it the first detector, because it really is a detector and because farther along in the receiver we shall use a second detector for demodulating the modulated intermediate-frequency signal voltage. Nearly all mixers are triodes or pentodes operated either as grid-leak detectors or else cathode-biased detectors.

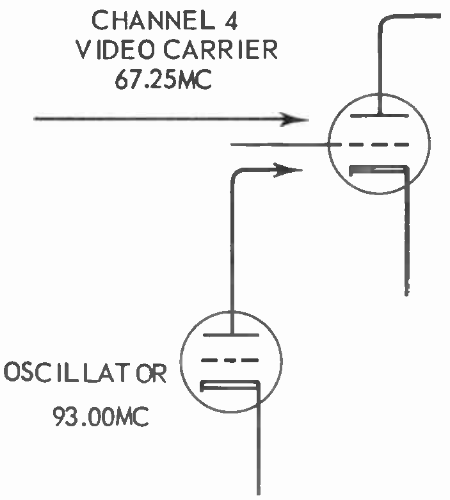
With either method of operation the mixer grid is biased so far negative that only the more positive alternations of the applied high-frequency signal voltage cause pulses of plate current to flow. Then the average value of plate current varies according to the modulation or according to the beat frequency. Current and voltage in the plate circuit load of the mixer tube vary at the beat frequency.

It is apparent in Fig. 32-5 that the amplitude of the modulation is exactly the same as that of the signal voltage shown at the top of the figure. That is, the difference between the highest and lowest peaks of the combined voltage is the same as between positive and negative peaks of the incoming r-f signal. Every change in r-f signal voltage will cause a proportional change in the beat frequency voltage. With actual operating conditions there will be changes in the incoming r-f voltage, because that voltage will be modulated with the television signal or a sound signal for standard broadcast. Since every change in the r-f voltage will appear in the beat-frequency voltage, the television signal or sound signal modulation will reappear in the beat-frequency voltage. This beat voltage is the intermediate frequency to be further amplified, and so we find the original television or sound signal modulation coming through as modulation of the intermediate frequency voltage.

The mixer not only changes the original signal-modulated carrier frequency to a signal-modulated intermediate frequency, but it also provides a certain amount of amplification or voltage gain. This we should expect, because in both the grid-leak detector and the cathode-biased detector the changes of voltage produced on the grid are amplified by triode or pentode action in the plate circuit, and the mixer is one or the other of these detectors.

4 You will recall that the increase of signal strength in amplifier tubes is proportional to the characteristic which we call transconductance. Transconductance is the ratio of signal current in the plate circuit to signal voltage on the grid, and is equal to the quotient of dividing the microamperes of plate signal current by the volts of grid signal. Such a division gives transconductance in micromhos. The transconductance of a mixer tube is arrived at in just the same way, and it is equal to a number of micromhos found from dividing microamperes of plate signal current by volts of grid signal. This characteristic of a mixer tube is called its conversion transconductance. All this is mentioned because you will hear a good deal about conversion transconductance, and you should understand its meaning.

INTERMEDIATE FREQUENCIES. In the demodulated output from the mixer tube we have the beat frequency or difference frequency which is to be the intermediate frequency for our receiver. But, while the average value of plate current is varying at the beat frequency, we still have pulsations at all the original frequencies.



OUTPUT FREQUENCIES

CARRIER	OSCILLATOR	DIFFERENCE	SUM
		93.00	93.00
		–67.25	+67.25
67.25	93.00	25.75	160.25

INTERMEDIATE FREQUENCY

Fig. 32-6. The four principal frequencies in the output of a mixer.

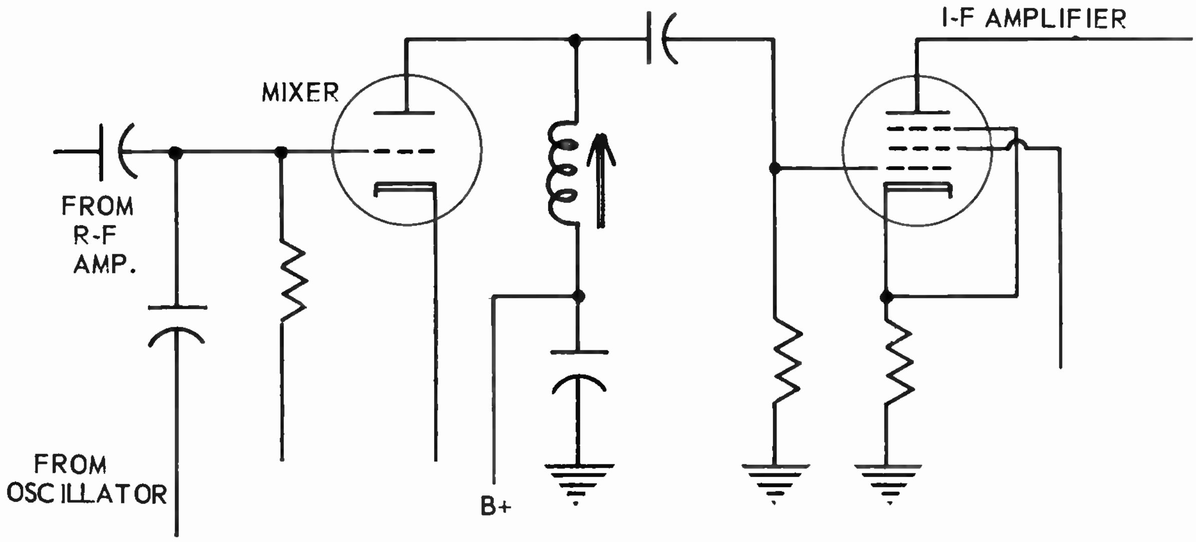


Fig. 32-7. A coupling between mixer and first i-f amplifier is tuned to the beat frequency or intermediate frequency.

In order to have some actual frequencies for illustrating what happens we shall assume, as in Fig. 32-6, that we are tuned to television channel 4 where the video carrier frequency is 67.25 mc (megacycles), and that the local oscillator is tuned for 93.00 mc. These are the two frequencies applied to the mixer grid. In the mixer output or in the mixer plate circuit we have four frequencies. We still have the original carrier frequency of 67.25 mc and also the original oscillator frequency of 93.00 mc. We have the beat frequency which is the difference between oscillator and carrier, or is 25.75 mc. This is the intermediate frequency which we intend to use. The fourth output frequency is the sum of the oscillator and carrier f equencies, or 160.25 mc.

In addition to the two original frequencies and their difference and their sum there will be multiples of the original frequencies. These multiples are called harmonic frequencies. Later on we shall learn more about harmonics. Just now we need know only that any tube which is biased to act like a detector will produce multiples or harmonics of all the frequencies applied to its grid circuit. We here would have the second harmonic of the carrier frequency, which would be 134.50 mc, and a second harmonic of the oscillator frequency, which would be 186.00 mc, and still higher harmonics as well.

All that we need do in order to get rid of everything except the difference frequency or the desired intermediate frequency is couple the output of the mixer to the grid of the following intermediate-frequency amplifier tube through a circuit tuned to resonance at the intermediate frequency. Such a coupling, of the tuned impedance type, is shown by Fig. 32-7. Whatever the type of coupling it is tuned to the intermediate frequency which, in our example, would be 25.74 mc. This frequency receives maximum amplification, while all the other unwanted frequencies receive little or no amplification.

Supposing now we wish to change from television channel 4 to channel 5 for reception. In channel 5 the video carrier frequency is 77.25 mc. We do not want to alter the intermediate frequency. In fact, we could not alter the intermediate because our whole i-f amplifying system is permanently tuned to this one frequency. This fixed intermediate frequency is one of the big advantages of the superheterodyne.

Since we cannot change the intermediate frequency we must change the oscillator frequency so that it will beat with the carrier of channel 5 to produce our fixed intermediate frequency of 25.75 mc. The new oscillator frequency must be 103.00 mc because the difference between 103.00 mc and the channel-5 carrier frequency of 77.25 mc is our intermediate frequency of 25.75 mc. The way we actually determine the required oscillator frequency is by adding the carrier frequency and the intermediate frequency; 77.25 mc plus 25.75 mc equals 103.00 mc.

Every time we wish to change channels or change the received signal it is necessary only to re-tune the r-f amplifier and also re-tune the local oscillator so that the difference between oscillator and carrier frequencies always remains equal to the intermediate frequency for which the receiver is designed and adjusted. Following is a list of video carrier frequencies for the various television channels, together with the oscillator frequency required for each channel when the video intermediate frequency is 25.75 mc.

Channel	Carrier	Oscillator	Channel	Carrier	Oscillator
2	55.25	81.00	7	175.25	201.00
3	61.25	87.00	8	181.25	207.00
4	67.25	93.00	9	187.25	213.00
5	77.25	103.00	10	193.25	219.00
6	83.25	109.00	11	199.25	225.00
			12	205.25	231.00
			13	211.25	237.00

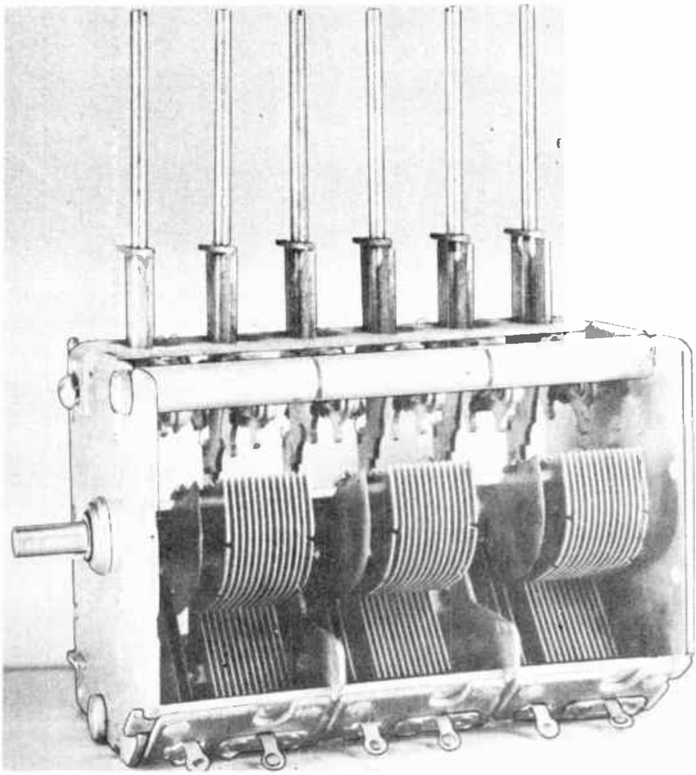


Fig. 32-8. Ganged capacitor for tuning r-f, oscillator, and mixer circuits in a combination a-m and f-m receiver. The upward extending rods are for push buttons that will tune five signal frequencies.

Were the intermediate frequency to be something different from 25.75 mc all the oscillator frequencies would have to be changed accordingly. Most intermediate frequencies of television receivers are between 25.75 and 26.75 mc, although some are in the range around 36 to 37 mc. As an example, were the video intermediate frequency of a receiver to be 26.60 mc the oscillator frequency for channel 4 would have to be the sum of the channel-4 video carrier frequency, 67.25 mc, and the intermediate frequency of 26.60 mc, which comes to 93.85 mc for the oscillator. The oscillator frequencies for other channels would undergo similar changes.

Now let's see how signal modulation of the carrier reappears as the same signal modulation of the intermediate frequency. We shall assume that the television receiver is tuned to channel 4, where the video carrier frequency is 67.25 mc, and assume also that there is signal modulation of 3.25 mc at some one instant of time. We take these figures merely to provide a simple explanation; and other channel frequency and any other modulation frequency would work out in similar fashion. The accompanying tabulation lists all the frequencies existing at the selected instant. The oscillator frequency remains constant so long as we are tuned to the same channel.

		No Modulation	Modulated (upper sideband)	Modulated (lower sideband)
Oscillator frequency	=	93.00	93.00	93.00
Carrier frequency	=	67.25	70.50	64.00
Intermediate frequency	=	25.75	22.50	29.00

The intermediate frequencies are the difference frequencies between oscillator and carrier. With no modulation the intermediate center frequency is 25.75 mc, for which we assume the receiver is designed. The intermediate frequency resulting from the upper sideband in the carrier is 22.50 mc. This differs from the unmodulated intermediate frequency by 3.25 mc. The intermediate frequency resulting from the lower sideband is 29.00 mc, which again differs from the unmodulated intermediate by 3.25 mc. Thus we find exactly the same signal modulation in the intermediate frequency as in the carrier and its modulation.

You will note that there is reversal of upper and lower sideband frequencies between carrier and intermediate. The upper sideband frequency on the carrier is higher than the frequency of the unmodulated carrier, but the resulting modulation of the intermediate is lower than the unmodulated intermediate frequency. There is similar reversal of the other sideband frequency. This makes no real difference in reception, because the voltages are not yet detected or demodulated and we may use either the positive or the negative side of the voltage wave.

OSCILLATORS. While examining the principles of superheterodyne reception we have mentioned the oscillator as producing a high-frequency voltage, but have said nothing about how the oscillator does its work. The oscillator which is part of the superheterodyne tuner used d-c power from the B-supply of the receiver to maintain alternating currents and voltages at the resonant frequencies of inductance and capacitance in the oscillator circuit. From now on we shall call this part of the tuner the r-f oscillator rather than the local oscillator to distinguish it from other oscillators used in the same receiver for sweeping the electron beam in the picture tube.

More than a dozen different types of oscillator circuits are in use or have been used in television and radio apparatus to produce high-frequency voltages of approximate sine-wave form such as required in the superheterodyne tuner. We shall examine two of these types in detail, and later may look briefly at some of the others. One of the two types is called a Colpitts oscillator, used in the tuners of the great majority of television receivers. The other is the Hartley oscillator, used in a great many sound radio receivers. Both types are used in signal generators and other testing instruments.

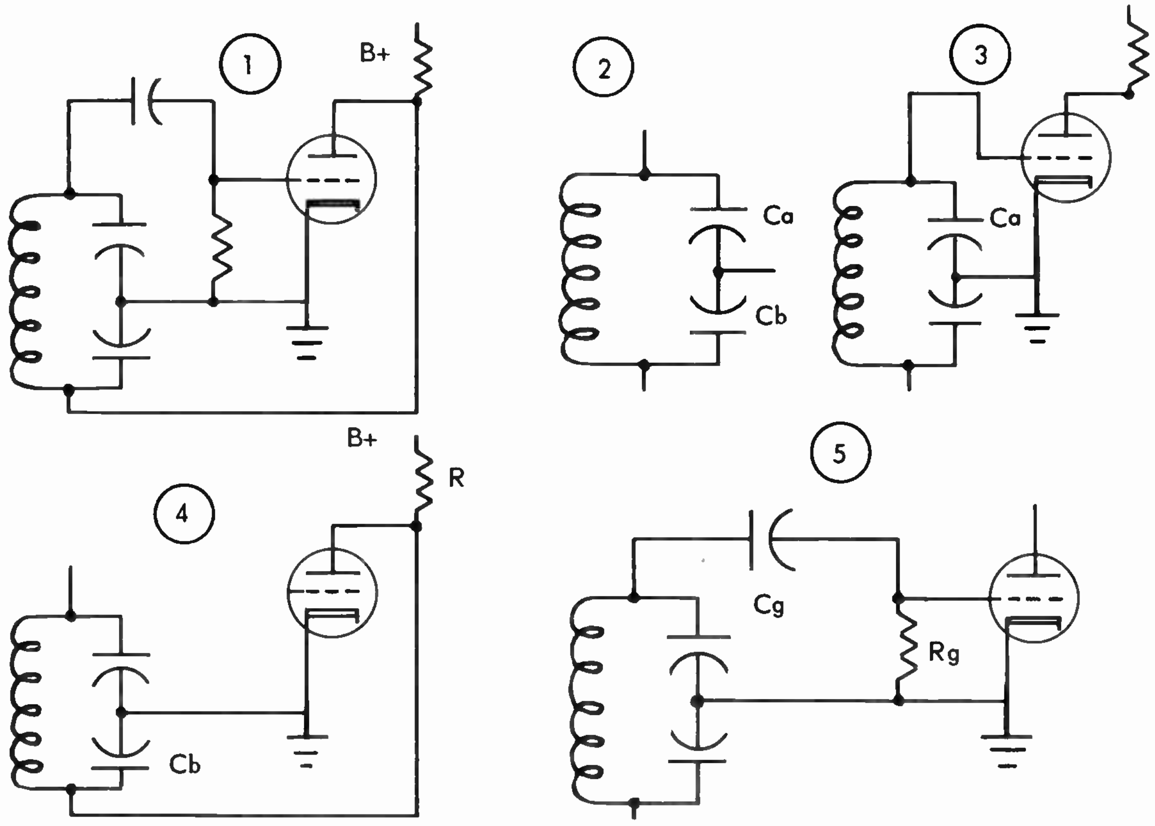


Fig. 32-9. The parts of a Colpitts oscillator circuit.

Colpitts and Hartley oscillators, as well as most other well known types, belong to the class called feedback oscillators in which the action is sustained by feeding energy from the plate circuit back into the grid circuit of a tube. The tube acts much like an amplifier in which alternating voltage applied to the grid circuit is secured from the plate circuit of the same tube instead of from an external signal source.

The elementary circuit of a Colpitts oscillator is shown at 1 in Fig. 32-9. Without the tube and its connections we have the parallel resonant circuit of diagram 2. The only difference between this and any ordinary parallel resonant circuit is that there are two capacitors in series instead of the usual single capacitor. So far as resonant behavior of the circuit is concerned the two series capacitors act just like a single unit.

When the capacitors are charged they contain a store of energy in the electric fields within the dielectric. As the capacitors discharge they will force an electron flow through the coil or inductor, and a magnetic field will be built up around the coil. Then the energy is in the magnetic field. The magnetic field immediately collapses, induces emf in the coil, and the resulting electron flow is in a reversed direction recharges the capacitors. Then the energy is back in the electric fields. All this, as you must have recognized, is simply the action that takes place in any parallel resonant circuit.

The energy swings back and forth or oscillates between the capacitors and the coil at a frequency determined by values of capacitance and inductance in the circuit. The energy is transferred back and forth by electron flow, or alternating current, at the frequency for which the capacitance and inductance are resonant. Once the oscillations are started they will continue until the energy is used up in overcoming high-frequency resistance or losses in the circuits elements.

If the loss of energy can be continually replaced with more energy the oscillations will continue indefinitely. This is where the tube comes in, it is a means for replacing high-frequency losses in the tuned circuit by energy taken from the B-power supply of the receiver through the plate circuit of the tube.

As a first step in replacing the lost energy we shall put back the grid circuit connections, as at 3 in Fig. 32-9. Now, with both capacitors charging and discharging at the resonant frequency, the alternating voltage across capacitor *Ca* is applied between grid and cathode of the tube. This will cause an alternating plate current, at the resonant frequency.

In diagram 4 we have the plate circuit by itself. The necessary direct plate voltage from the B-power supply is applied through resistor *R*. The high resistance of *R* opposes flow of the alternating plate current, so most of this high-frequency current takes the easier path through capacitor *Cb* back to the cathode of the tube. This current continually adds to the charges and voltage of capacitor *Cb* and replaces the energy being lost in the resonant circuit, of which *Cb* is a part. The frequency of the plate current is the same as the tuned resonant frequency of the whole system, because plate current is controlled by resonant voltage from capacitor *Ca* applied to the grid. Pulses or alternations of plate current remain in phase with or in time with the current that is charging and discharging the capacitors.

With this oscillator circuit it is easy to put energy into the resonant circuit at any rate required to replace the high-frequency losses. We have feedback of energy from the plate circuit, and oscillation will be maintained.

If you were to put diagrams 3 and 4 together there would be a conductive connection from the highly positive direct voltage at the tube plate through the coil of the resonant circuit to the grid. This would be bad, for the grid must not be positive, it must be negatively biased. To block the direct voltage away from the grid we insert capacitor *Cg* of diagram 5. To obtain negative grid bias we use this capacitor in connection with grid resistor *Rg* as a grid-leak biasing system. Now, by putting diagrams 4 and 5 together you have the complete Colpitts oscillator, as in diagram 1.

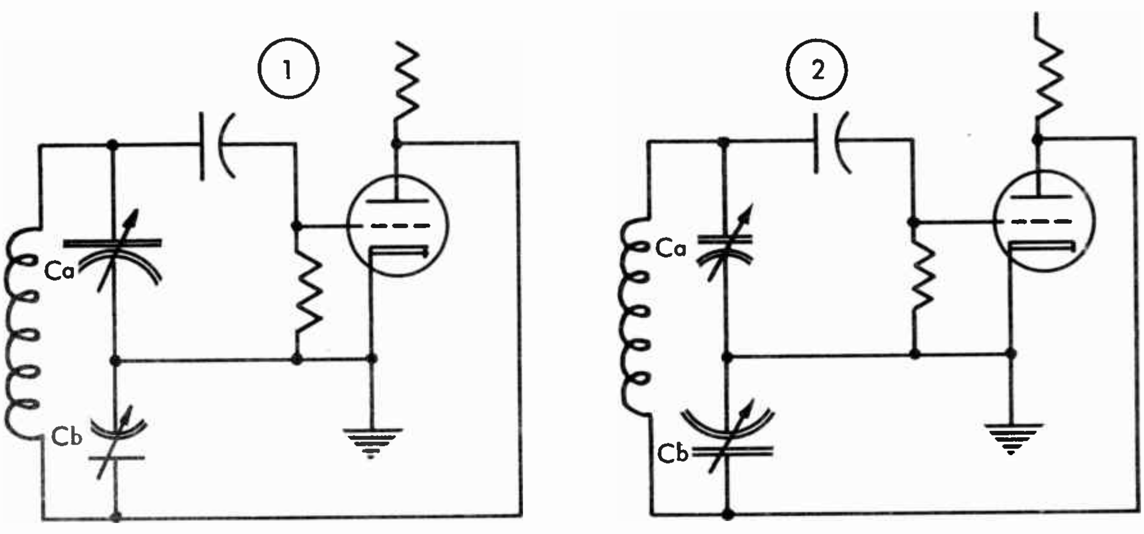


Fig. 32-10. Capacitance changes which alter the feedback in a Colpitts oscillator.

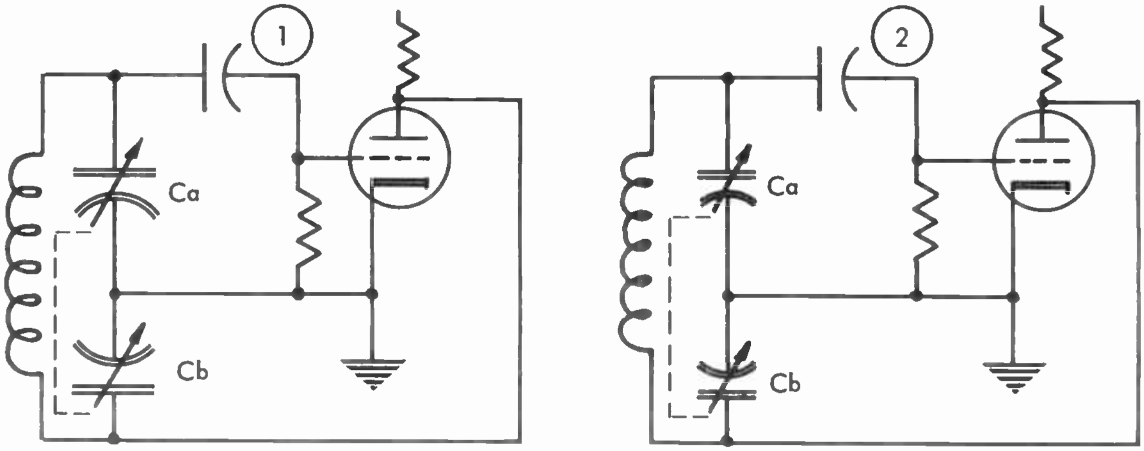


Fig. 32-11. Capacitance changes which vary the frequency of a Colpitts circuit.

COLPITTS OSCILLATOR ADJUSTMENTS. To understand the effects of altering one or both capacitances in the Colpitts oscillator circuit we must keep these facts in mind. Applications of the facts are shown by Figs. 32-10 and 32-11.

More capacitance in a tuned circuit means a lower resonant frequency. More capacitance also means less capacitive reactance and a lower voltage across the capacitance at any given frequency.

Less capacitance means a higher resonant frequency. Less capacitance means also more capacitive reactance and high voltage across the capacitance at any given frequency.

Feedback is increased by more voltage across capacitor *Cb* in the plate circuit, and is decreased by less voltage across *Cb*.

The effect of feedback in the grid circuit of the tube is increased by more voltage across capacitor *Ca*, in the grid circuit, and is decreased by less voltage across *Ca*.

The effect of feedback really depends on the ratio of voltages across *Cb* and *Ca* rather than on either voltage considered by itself. That is, increasing the voltage across capacitor *Cb* has the same effect as decreasing the voltage across *Ca*, and vice versa.

In Fig. 32-10 diagram 1 represents large capacitance at *Ca* and small capacitance at *Cb*, or low voltage across *Ca* and high voltage across *Cb*. The effect of feedback has thus been increased, because of high reactance in feedback capacitor *Cb*.

In diagram 2 of Fig. 32-10 there is small capacitance and higher voltage across *Ca*, in the grid circuit, and large capacitance and lower voltage across *Cb*, in the plate circuit. The effect of feedback is thus decreased, due to less reactance and less amplification for feedback voltage in the plate circuit.

Now what about the effect on resonant frequency? If capacitance at *Ca* in diagram 1 is equal to capacitance at *Cb* in diagram 2, and if *Cb* of diagram 1 is equal to *Ca* of diagram 2, the combined capacitance is the same in both diagrams. This combined capacitance is that in the resonant circuit, and since capacitance has not been changed there is the same resonant frequency in both cases. Feedback is changed without changing the resonant frequency.

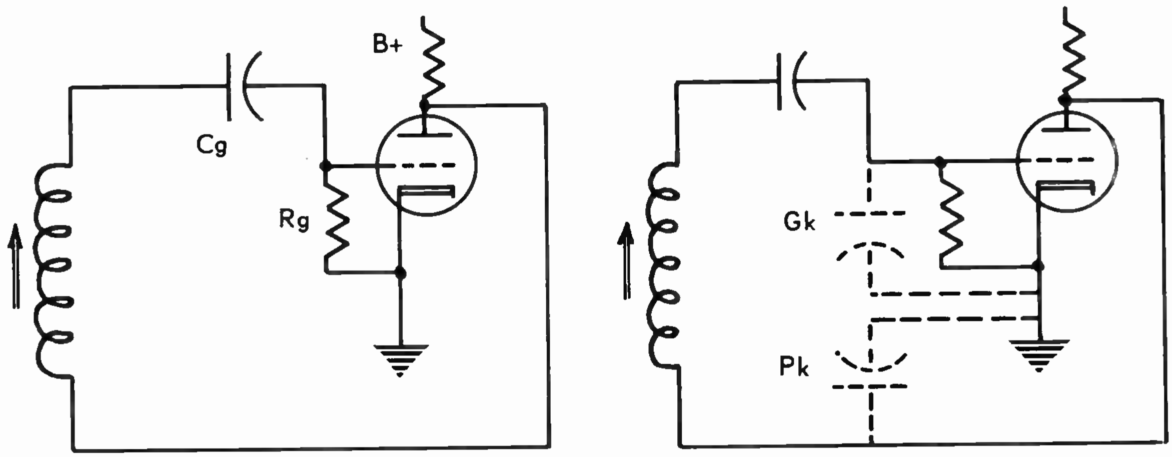


Fig. 32-12. Colpitts circuit employing tube capacitances for feedback. Tuning is by means of a movable core in the coil.

In Fig. 32-11 diagram 1 represents large capacitance in both *Ca* and *Cb*. This greater capacitance lowers the resonant frequency. In diagram 2 there is small capacitance in both capacitors, and this smaller capacitance raises the resonant frequency.

If capacitances at *Ca* and *Cb* of the two diagrams have been altered together in such manner as not to change the ratio of one to another there has been no change of effective feedback. That is, if both capacitances were halved, or changed in any other similar relation, the reactances and voltages would likewise change together and would remain of the same ratio as earlier voltages. In Fig. 32-11 we have changed the frequency without changing the feedback.

5 When two variable capacitors are used in a Colpitts oscillator circuit the two most often are ganged, on a single shaft, so that resonant frequency may be altered without changing the feedback to any material degree. Ganging of the two capacitors is indicated by the broken line connections in Fig. 32-11.

MODIFIED COLPITTS OSCILLATORS. In the service manuals for many television receivers you will find the statement that the r-f oscillator is a modified Colpitts type. This means that the oscillator operates with the Colpitts principle, but has some arrangement of inductance or capacitance that differs from the basic circuit that we have examined.

One modification is shown at the left in Fig. 32-12. Both the capacitors which do the tuning and provide feedback are absent, but the two capacitances still exist. One of the capacitances is that between the grid and cathode of the oscillator tube, as indicated by broken lines at *Gk* in the right-hand diagram. The other capacitance is that between the plate and cathode of the tube, as indicated at *Pk*. These tube capacitances, in connection with capacitances of the tube socket, wiring connections, and other parts, will provide the tuning and feedback necessary for oscillator operation.

Other modified Colpitts oscillator circuits are shown by Fig. 32-13. Feedback in both these circuits is through the internal capacitances of the tube, as explained in connection with Fig. 32-12. Tuning is varied by adjustable capacitors *Ct* connected across the coils. In the left-hand diagram there is cathode bias by means of resistor *Rk*, with resistor *Rg* used only to provide a conductive grid return to the cathode. Blocking capacitor *Cb* keeps direct current and voltage of the B-power supply out of the tuning coil and grid circuit while providing a path for high-frequency currents between the tube plate and the bottom of the oscillator coil.

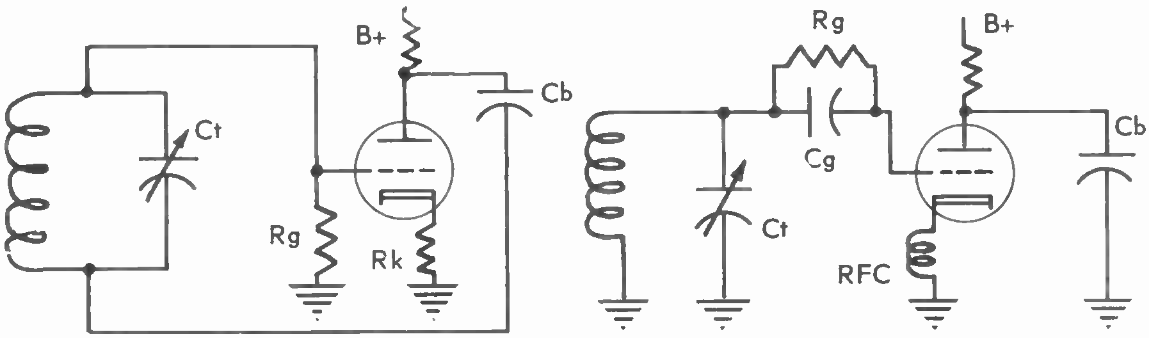


Fig. 32-13. Modified Colpitts oscillators.

In the right-hand diagram we have grid-leak bias by means of resistor *Rg* and capacitor *Cg*. The ground connections effectively join together one side of the coil, one side of the tuning capacitor *Ct*, one side of the blocking and r-f plate current capacitor *Cb*, and the lower end of the cathode circuit. The radio-frequency choke *RFC* between cathode and ground has very great reactance or impedance to the high-frequency currents in plate and grid circuits, but only negligible resistance to direct current in the cathode circuit. Consequently, this choke isolates the cathode from the high-frequency circuits while allowing free flow through it of direct current for the plate circuit or the B+ circuit.

There are a great many other variations or modifications of the fundamental Colpitts circuit, as we shall discover when examining television oscillators. Quite often the tuning is varied by means of a movable core in the coil rather than with an adjustable capacitor, or both the capacitor and the movable core may be used for service adjustments.

HARTLEY OSCILLATOR. The principles of the Hartley oscillator are shown by diagram 1 in Fig. 32-14. The parallel resonant circuit consists of the tuning capacitor *Ct* and the tapped coil whose parts are marked *Lg* and *Lp*. Capacitor *Cg* and resistor *Rg* provide grid-leak bias for the tube. Blocking capacitor *Cb* keeps direct plate voltage and current of the B-supply out of the tuned circuit while allowing high-frequency plate current to reach the tuned circuit.

The portion of the coil marked *Lg* is in the grid circuit of the tube, because it is connected to the grid through capacitor *Cg* and to the cathode through the tap connection on the coil. The portion of the coil marked *Lp* is in the plate circuit, because it is connected to the plate through capacitor *Cb*, and to the cathode through the tap connection. The oscillating current that swings back and forth between the coil

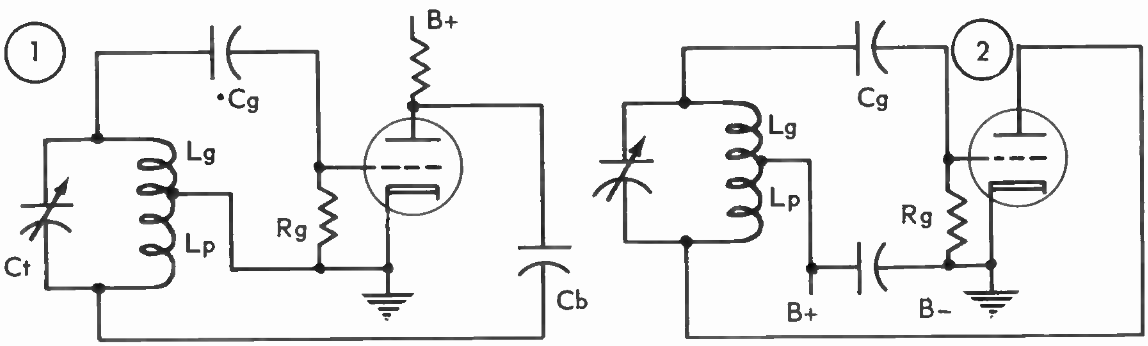


Fig. 32-14. The Hartley oscillator circuit.

and capacitor *Ct* must flow in both parts of the coil, because the parts are in series with each other so far as the resonant circuit is concerned. Since the same oscillating current flows in both the plate circuit and the grid circuit there is feedback of energy from plate to grid. Plate current replaces in the oscillating circuit all the energy that is lost due to high-frequency resistance.

The inductance coil of the Hartley circuit may be a continuous winding, tapped at or near its center, or it may be two separate windings connected together and to the cathode of the tube. The two parts of the inductance may be in such positions that there is inductive coupling and mutual induction between them, or they may be separated and so placed as to have little or no inductive coupling. Inductive coupling is not necessary, because the same oscillating current flows in both inductances no matter how they are positioned.

Diagram 2 of Fig. 32-14 shows a Hartley oscillator circuit which does not require the blocking capacitor in the plate circuit. The B+ connection is made to the tap on the inductance coil. Direct plate current flows in the lower portion of the coil, *Lp*. There is connection from the tap to the tube cathode from B+ to B– and ground, through the power supply.

In the oscillator circuits of Fig. 32-14 one side of the tuning capacitor is connected or coupled to the grid circuit and the other side is connected or coupled to the plate circuit. We say that both sides of the tuning capacitor are at "r-f potential", which really means that neither side of the capacitor is connected to ground or to B-minus. Were you to touch the rotor shaft or the adjusting screw of a tuning capacitor which is at r-f potential there would be a decided change of operating frequency, or the oscillation might stop altogether. This would be due to what is called body capacitance.

Body capacitance effect may be avoided by rearranging the Hartley oscillator circuit as in diagram 1 of Fig. 32-15. Here the rotor side of the tuning capacitor is directly connected to chassis ground. Now the rotor shaft or adjusting screw is at ground potential, not at r-f potential, and you can come close to or touch the tuning adjustment without greatly affecting the tuned frequency. When one side of the tuning capacitor is grounded the tube cathode cannot be grounded. Then the cathode is at r-f potential. When the tuning capacitor is at r-f potential it is possible to ground the tube cathode, as in Fig. 32-14.

Whether to ground the tuning capacitor or the tube cathode of any of the oscillators depends on how the circuit is to be operated and on the preferences of the designer. For example, were the capacitor to be adjusted by the operator of a receiver it would be better to ground the capacitor, but if this is only a service adjustment it could be handled with an insulating screw driver or wrench, and the cathode could be grounded to simplify the wiring. Note that in the Colpitts oscillator at the left in Fig. 32-13 the tuning capacitor is at r-f potential, while with the arrangement at the right the capacitor is grounded.

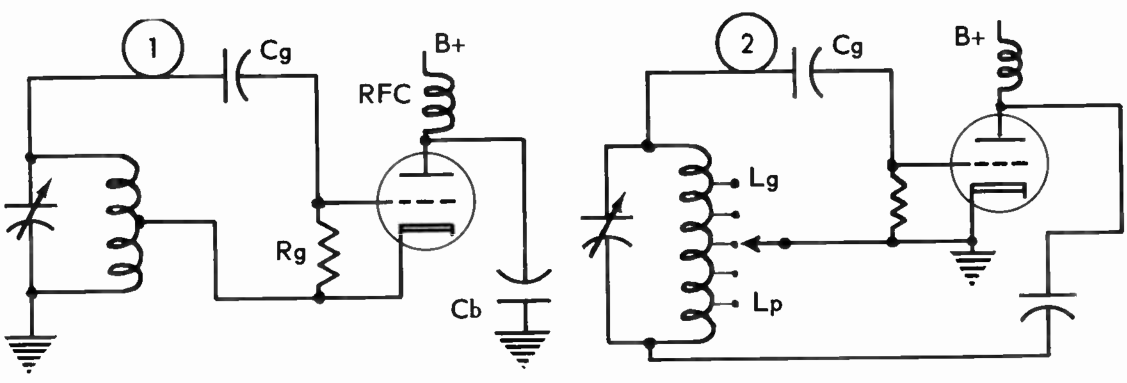


Fig. 32-15. Modifications of the Hartley oscillator.

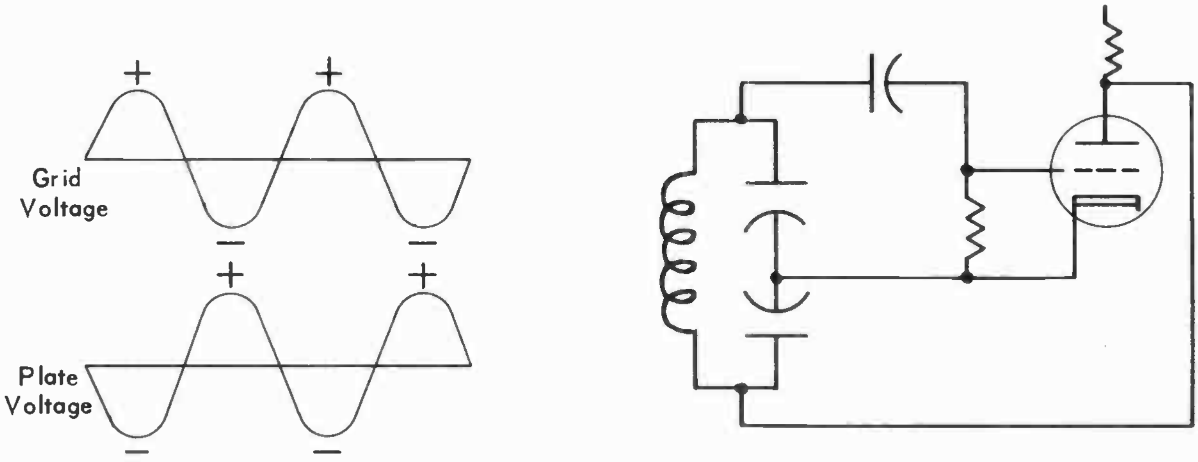


Fig. 32-16. Phase relation of grid and plate voltages in an oscillator.

7 The strength of feedback in the Hartley oscillator depends on the ratio of plate circuit inductance to grid circuit inductance. The greater the plate circuit inductance in comparison with the grid circuit inductance the stronger is the feedback. Diagram 2 of Fig. 32-15 shows a method of adjusting the feedback by making connection to any of a number of tap points on the coil. This adjustment makes no material change of oscillating frequency, since the total inductance in the resonant circuit is not altered. Usually it turns out that a tap connection near the coil center allows strongest oscillation or maximum oscillating current and voltage. Getting the tap too near the grid end makes too great a reduction of inductance and reactance in the grid circuit. With the tap too near the plate end there is too little inductance and reactance or impedance in the plate circuit to let the tube act as an efficient amplifier for grid voltage.

In the B+ lines of Fig. 32-15 there are radio-frequency choke coils, *RFC*, instead of resistors. A choke

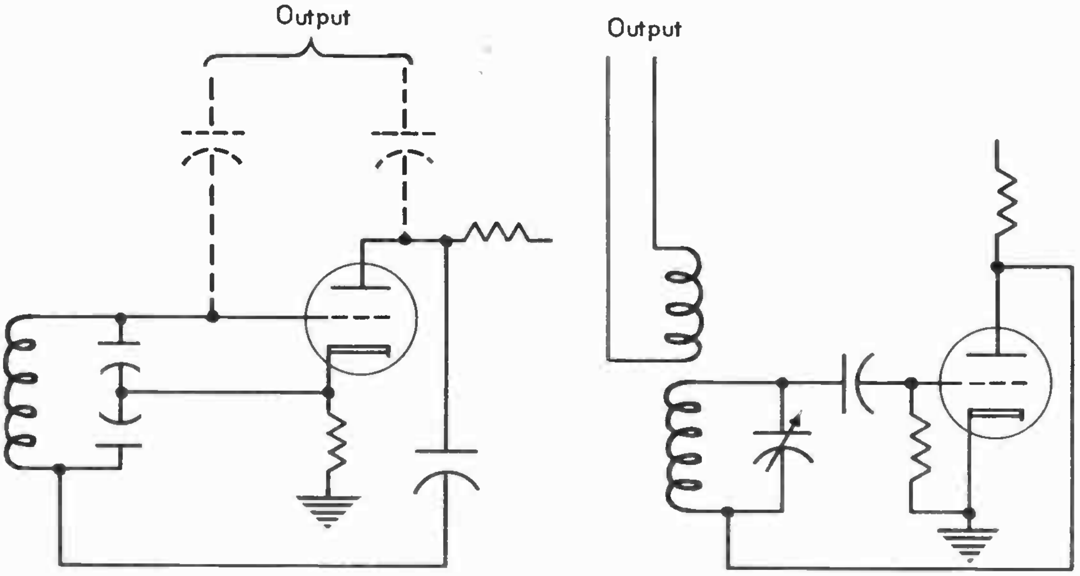


Fig. 32-17. Methods of taking voltage outputs from oscillators.

provides greater impedance than a resistor against escape of high-frequency currents and has lower resistance to direct plate current. Resistors are more commonly used because they may be needed for voltage dropping, they give generally satisfactory operation, and cost less than chokes.

2 *OSCILLATOR CIRCUITS.* An oscillator tube acts essentially as an amplifier for voltages applied to the grid. Earlier we learned that in any amplifier tube the plate voltage becomes less positive when grid voltage is becoming less negative or is going positive. That is to say, plate and grid voltages are of opposite phase. But when plate voltage is fed back to the grid, to maintain oscillation, the feedback voltage at the grid must be in phase with the alternating voltage already existing at the grid. Consequently, the plate voltage must be inverted before it reaches the grid.

The opposite-phase relation of grid and plate voltages is represented at the left in Fig. 32-16. At the right is a Colpitts oscillator circuit. In this circuit the coil or inductor extends between the grid and plate of the tube. We know that in a coil carrying alternating current at resonance, or carrying the current which flows back and forth between coil and capacitor, the potential is positive at one end of the coil while it is negative at the other end. Therefore, when positive voltage from the plate is applied to one end of the coil there is negative voltage at the other end, and vice versa. Thus we have feedback voltage in correct phase relation to maintain oscillation. The same thing happens in a Hartley oscillator, which also has a single inductor extending between plate and grid. An equivalent phase reversal must occur in every other oscillator which depends on feedback for its action.

Any oscillator begins to operate when voltage is first applied to the tube, because any change of plate voltage, either great or small, results in a feedback voltage to the grid, Current commences to flow in the resonant circuit, the capacitor or capacitors are charged in one polarity or the other, and oscillating current commences to flow back and forth between capacitors and coils.

Oscillator tubes have either grid-leak bias or cathode bias. With either method there is zero bias at the instant in which voltage first is applied to the tube. This allows the momentary flow of grid current which is necessary to start electron flow in the resonant circuit. Then a negative bias is built up by continued pulses of grid current with grid-leak bias or with continued plate current for cathode bias.

High-frequency voltage may be taken from any type of oscillator circuit by either of the methods illustrated in Fig. 32-17. In the left-hand diagram a small capacitor is connected from either the grid circuit or the plate circuit of the oscillator to the grid circuit of the mixer tube. Thus the oscillator voltage is put into the mixer along with the amplified carrier signal voltage. The coupling capacitance in television tuners usually is something between 1 and 4 mmfd.

In the right-hand diagram the oscillator output is taken through a coil inductively coupled to the coil in the resonant circuit of the oscillator. This output coil might be the tuned secondary winding of the transformer which is between the r-f amplifier and the mixer, with both coils wound on the same form or otherwise supported close enough to each other for coupling to take place.

In all r-f oscillators it is usual practice to have what is called shunt feed for the direct voltage and current in the plate circuit, which means that this B-power voltage and current are fed to the tube plate without going through the tuned resonant circuit. This method of connection is shown in all diagrams of this lesson except diagram 2 of Fig. 32-14.

Incidentally, the resonant circuit or tuned plate-grid circuit of an oscillator quite often is called the "tank circuit". In describing a shunt feed connection we might say that there is no direct voltage or current in the oscillator tank circuit.

TELEVISION

AND SEND

LESSON NO. 33

SUPERHETERODYNE TUNING METHODS

Training Your Mind

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind
C.T.I.
Perfecting Your Skill

COMMERCIAL

TRADES

INSTITUTE

Chicago, Ill.

LESSON NO. 33
SUPERHETERODYNE TUNING METHODS

In this lesson we shall learn how r-f oscillators, mixers, and r-f amplifiers are connected together in television receivers, f-m (frequency-modulation) sound receivers, and a-m (amplitude modulation) sound receivers. Incidentally we shall practice reading circuit diagrams of the kinds ordinarily used in service instructions.

Up to this point in our work each given type of circuit has been drawn in the same general way for every diagram. This has made it easier to follow electron paths and to figure out what is happening, because you have seen certain arrangements over and over until they have become familiar. To continue in this way might cause you trouble later on, for you would come to feel that particular circuit details should be shown in only one way, and you could be confused by other types of diagrams.

There is no standard way of drawing the circuit diagram for any part of any television or radio receiver. It is possible to take something so simple as a Colpitts oscillator and draw its circuit in so many different ways that the originator, E.H. Colpitts, would have to look hard at some of the layouts to recognize his own brain child.

Of course, there are standard symbols which everyone uses. But the symbols for the parts which go into some tuner, amplifier, or other section of apparatus may be placed on a diagram in any relative positions which appeal to the person making the layout. There may be a close relation between the positions of symbols on the diagram and positions of parts in the actual apparatus, or there may be no relation at all. Regardless of the positions of the symbols, when lines are drawn between them to show correct electrical paths we have a diagram that is electrically correct.

The same symbols might be arranged on other diagrams in any of a dozen or more different positions, and joined with lines showing correct electrical connections. Then there would be a dozen or more different diagrams, all with the same symbols, all alike electrically, but looking as though they represented a dozen different kinds of apparatus. You should know your circuits, at least the common kinds, and be able to trace them out no matter how the symbols and connections are placed in a diagram, just as we shall do in several of the figures to follow.

TUNER CIRCUITS. First we shall look at the connections in one style of television tuner, as shown by Fig. 33-2. The r-f amplifier is a pentode, as you can tell from the symbol. The grid is connected through a capacitor to the antenna coupling transformer, also through a resistor to a source of biasing voltage not shown in the diagram.

The plate of the r-f amplifier is coupled to the grid of the mixer through an r-f transformer having adjustable tuning capacitors across both the primary and the secondary windings. There is shunt feed from the r-f

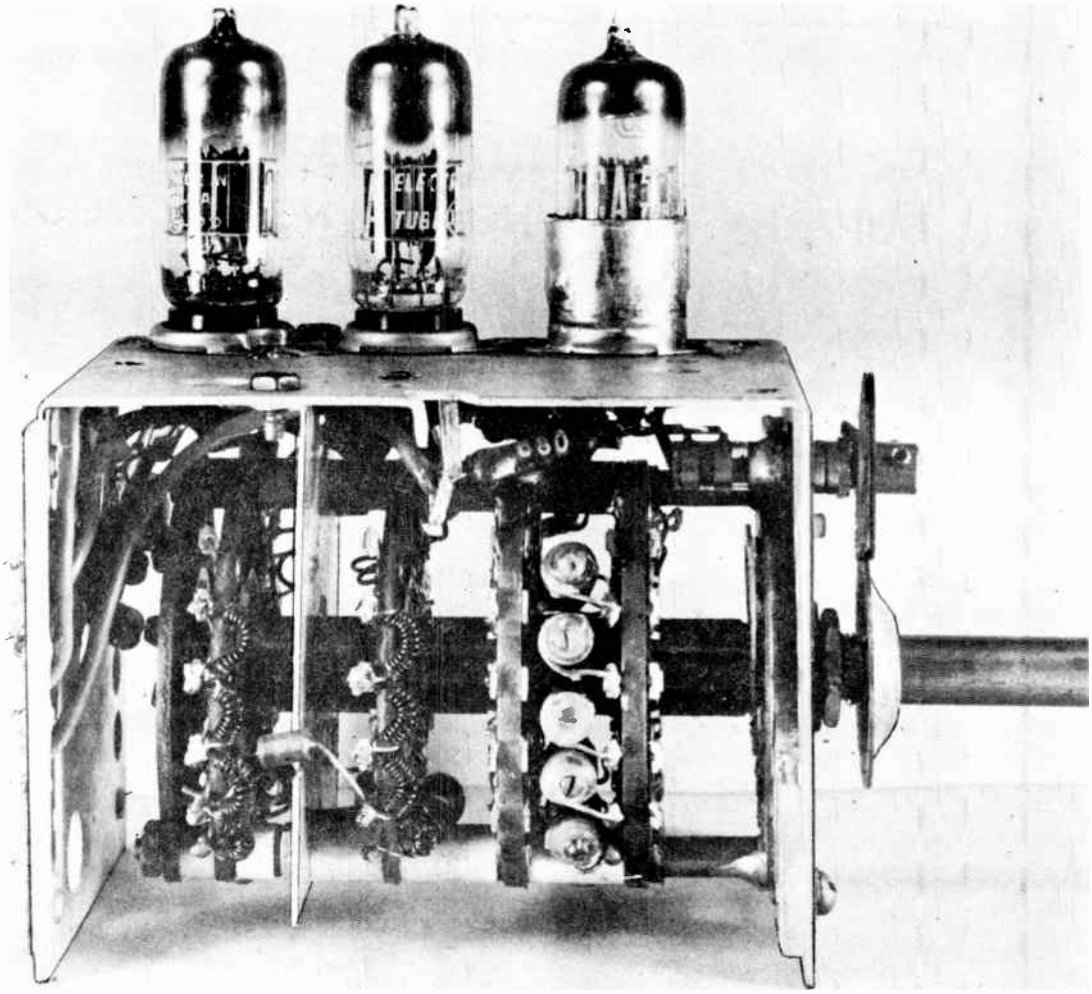


Fig. 33-1. A television tuner in which r-f and oscillator frequencies are obtained by switching suitable inductors into the circuits.

plate to the transformer, with blocking capacitor *Ca* allowing r-f currents to flow from plate to transformer, but confining direct plate current and voltage to the path which includes a resistor going to B+. The mixer tube is a triode, with grid-leak bias by means of capacitor *Cg* and resistor *Rg*.

Doubtless you recognize the r-f oscillator as a Colpitts type. There is grid-leak bias. The plate circuit is completed through capacitor *Cb* and ground to the lower end of the tuning coil. Across this coil is a tuning capacitor *Ct*. The tube cathode is isolated from r-f currents in the plate-grid circuit by means of the r-f choke *RFC*. High-frequency oscillator voltage is taken from the upper end of the tuned coil through coupling capacitor *Cc* to the mixer grid circuit.

In the large diagram of Fig. 33-2 the mixer and oscillator are shown as two separate triode tubes. Instead

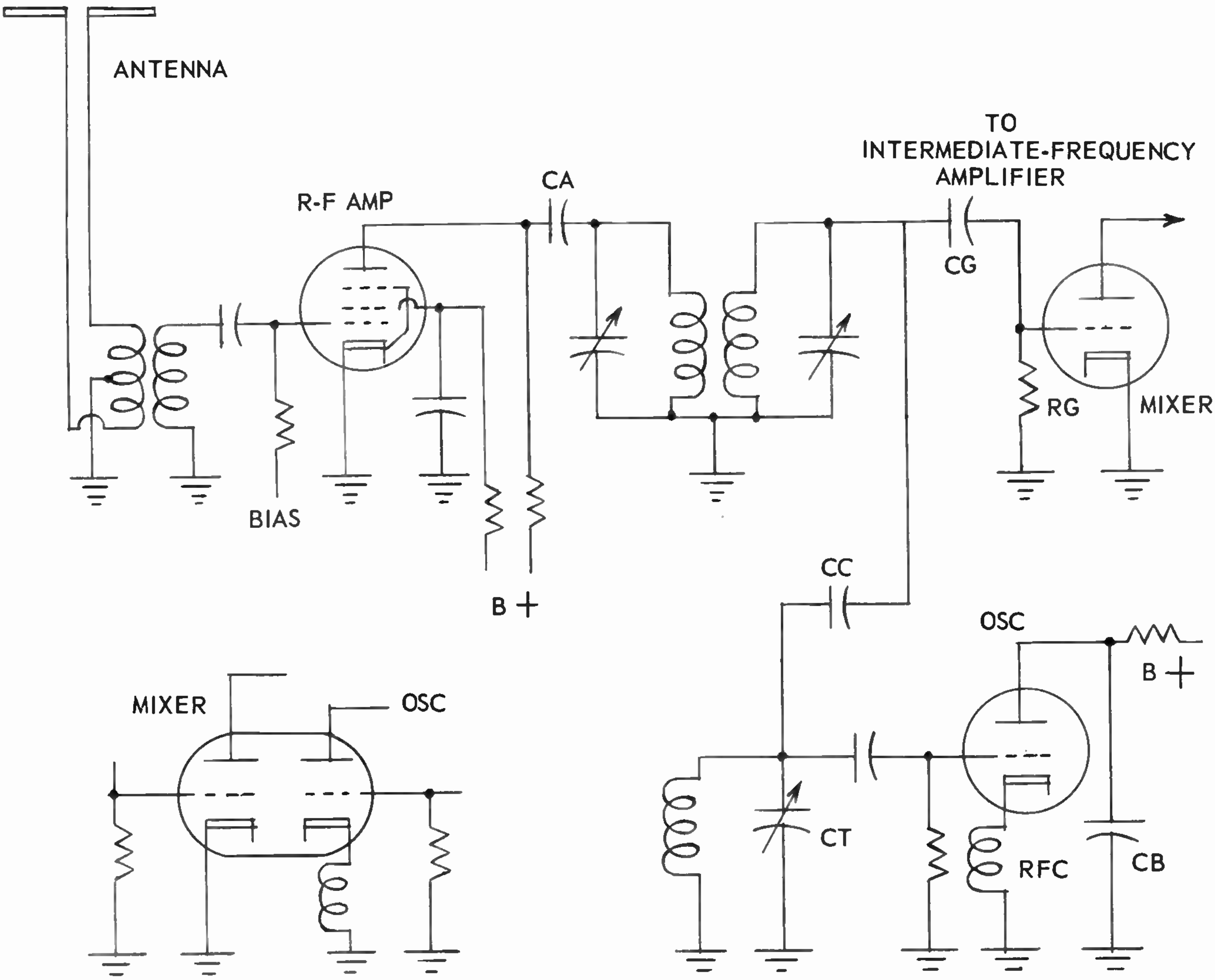


Fig. 33-2. Connections such as may be used in a television tuner.

of using separate tubes it would be more common practice to use the two sections of a twin triode tube as shown by the small sketch at the lower left. Connections to plates, grids, and cathodes of the twin triode would be the same as shown for the separate triodes in the large diagram. For connections such as shown here many television sets use a type 12AT7 tube, and many f-m sound receivers use a type 7F8.

(4) In Fig. 33-3 we have the tuning section for one style of f-m sound receiver. There is no r-f amplifier. The secondary winding of the antenna coupling transformer is in the grid circuit of the mixer, with this winding tuned by means of a variable capacitor and a paralleled trimmer capacitor. There are a large number of the smaller receivers for f-m sound and also those for a-m sound in which the antenna is coupled to the mixer grid without any intervening r-f stage. The mixer is one section of a twin triode tube, with the other half used as the r-f oscillator. Note that this mixer has cathode-bias for its grid voltage.

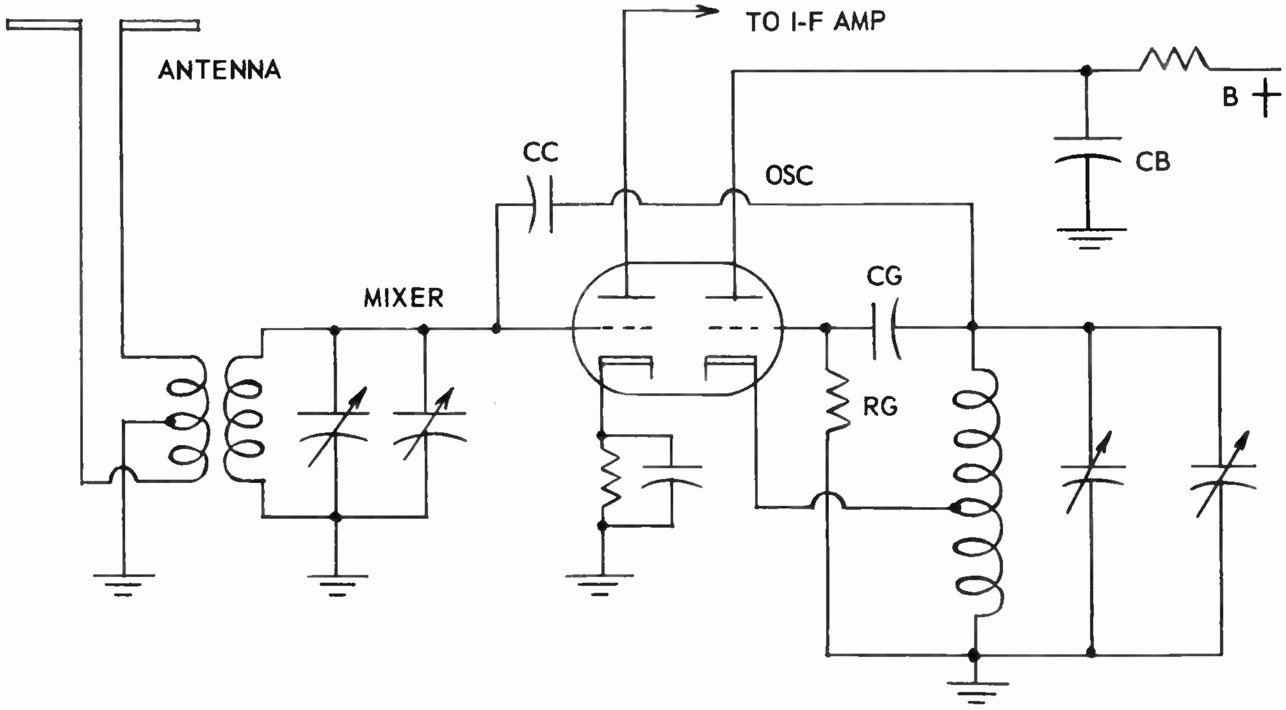


Fig. 33-3. Tuning circuits for an f-m sound receiver.

The r-f oscillator is a Hartley type, as is evident from the fact that the tube cathode is connected to a tap on the tuned coil. The coil in the grid-plate circuit is tuned by means of a variable capacitor and a paralleled trimmer capacitor. Bias for the oscillator is of the grid-leak type, with capacitor *Cg* and resistor *Rg*. The plate circuit is completed through capacitor *Cb* and ground to the lower end of the tuned coil. High frequency oscillator voltage is taken from the upper end of the tuned coil, or from the "tank circuit", through capacitor *Cc* to the mixer grid circuit. The capacitance at *Cc*, and also at the similarly marked capacitor in Fig. 33-2, is 1.5 mmfd.

TWIN TRIODE WITH SINGLE CATHODE. In Fig. 33-4 we again have the mixer and oscillator as the two sections of a twin triode tube, but here there is only a single cathode for both sections instead of the two separate cathodes of earlier diagrams. The circuits of this figure are part of those for a popular type of television tuner. The tube is a type 6J6. When there is only one cathode element serving both mixer and oscillator the cathode circuit must be the same for both functions. Usually the cathode is grounded, although in a few cases there is cathode bias when the same biasing voltage is suitable for both mixer and oscillator operation.

The plate circuit of the r-f amplifier, not shown in the diagram, is coupled to the grid circuit of the mixer through a transformer whose primary (plate) winding is marked *Lr* and whose secondary (grid) winding is marked *Lm*. The grid winding is tuned by an adjustable capacitor and by the input capacitance of the tube and wiring. Grid-leak bias for the mixer is provided by the capacitor in series with the grid circuit and the resistor which is between grid and cathode.

The oscillator of Fig. 33-4 is a Colpitts type. The oscillator coil, *Lo*, is tuned by the tube capacitances in the usual way and also by fixed capacitor *Ca* and adjustable capacitor *Cb*. Note that *Ca* is connected from the grid side of the circuit to ground, that *Cb* is connected from the plate side to ground, and both these capacitors are connected through ground to the cathode of the tube. Consequently, the connections between the capacitors and the cathode are electrically the same as in any Colpitts oscillator having capacitors for tuning and feedback.

Capacitor *Cg* and resistor *Rg* provide grid-leak bias for the oscillator. Capacitor *Cg* acts also as a blocking capacitor for the plate circuit, since it prevents direct conductive connection from the plate to ground or to the grids of the mixer and oscillator. High-frequency oscillator voltage is carried into the mixer grid circuit through inductive coupling between oscillator coil *Lo* and mixer grid coil *Lm*. These two coils are supported on the same form, with spacing of about 5/16 inch between their ends.

TUNED GRID OSCILLATOR WITH TICKLER FEEDBACK. In Fig. 33-5 there is still another mixer-oscillator combination with a single-cathode twin-triode tube. This particular circuit is used for a-m tuning in a combination a-m f-m sound receiver. You may not recognize the oscillator circuit, for it is of a type not previously considered.

The oscillator grid circuit, at the lower right, consists of coil *Lg* which is tuned to resonance by a variable capacitor and a trimmer capacitor. Inductively coupled to the grid coil is coil *Lp* which is in series with the plate of the oscillator. Energy is fed back from plate to grid through this coupling. The coil in the

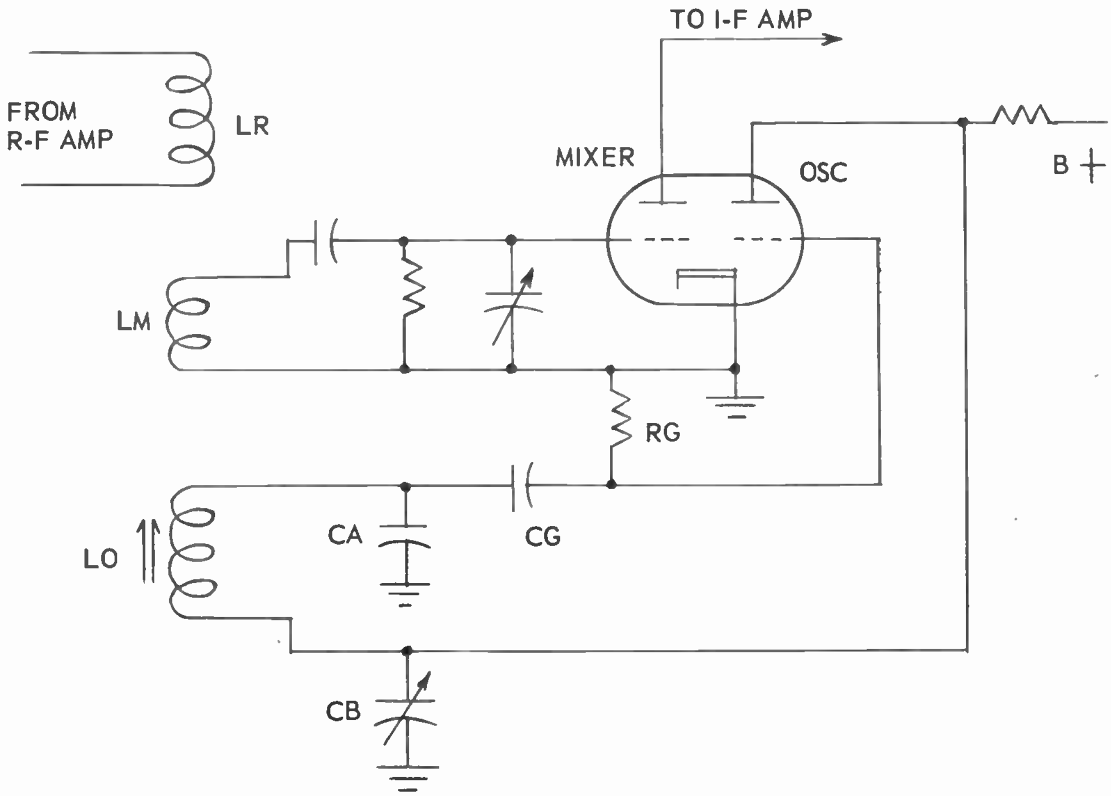


Fig. 33-4. Twin-triode mixer-oscillator tube having a single cathode.

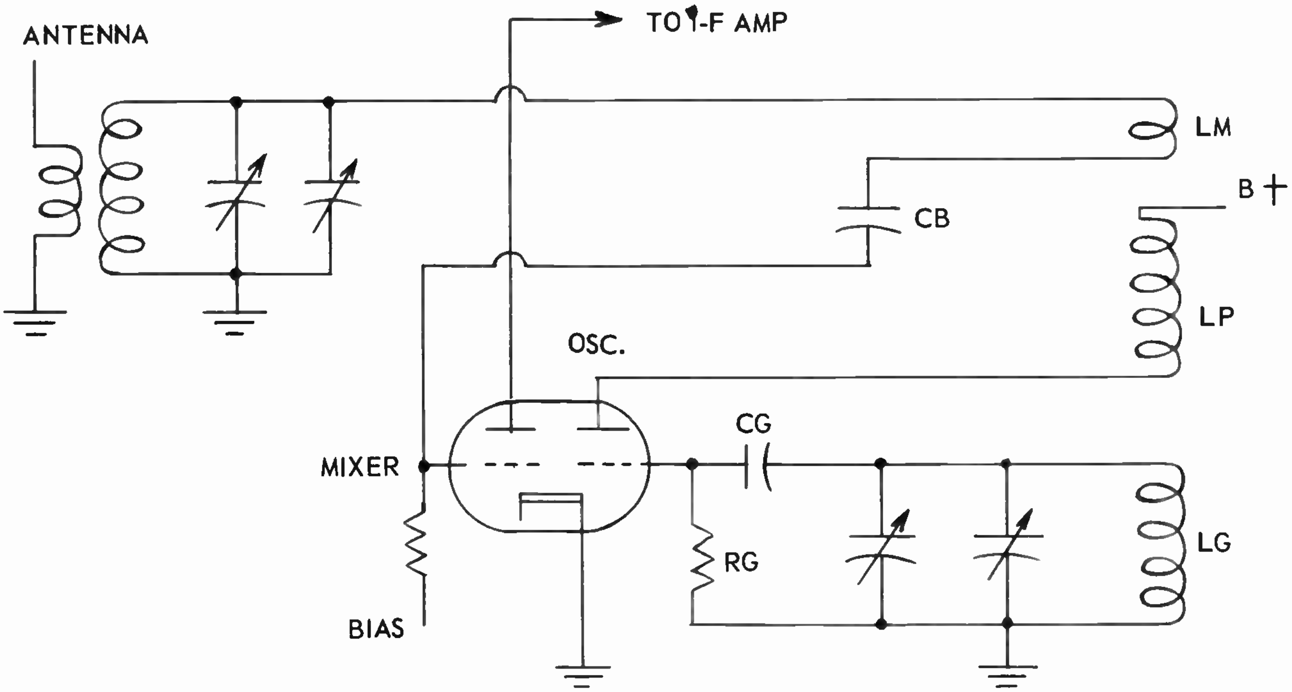


Fig. 33-5. Twin-triode mixer-oscillator with tuned-grid tickler feedback in the oscillator circuit.

plate circuit commonly is called a tickler coil or tickler winding. The oscillator is grid-leak biased with capacitor *Cg* and resistor *Rg*.

High-frequency oscillator voltage is inductively coupled into winding *Lm* from the tickler winding,and thus is put into the grid circuit of the mixer, with which *Lm* is in series. The mixer grid is negatively biased from a source not shown in the diagram. The direct bias voltage is blocked from a ground connection

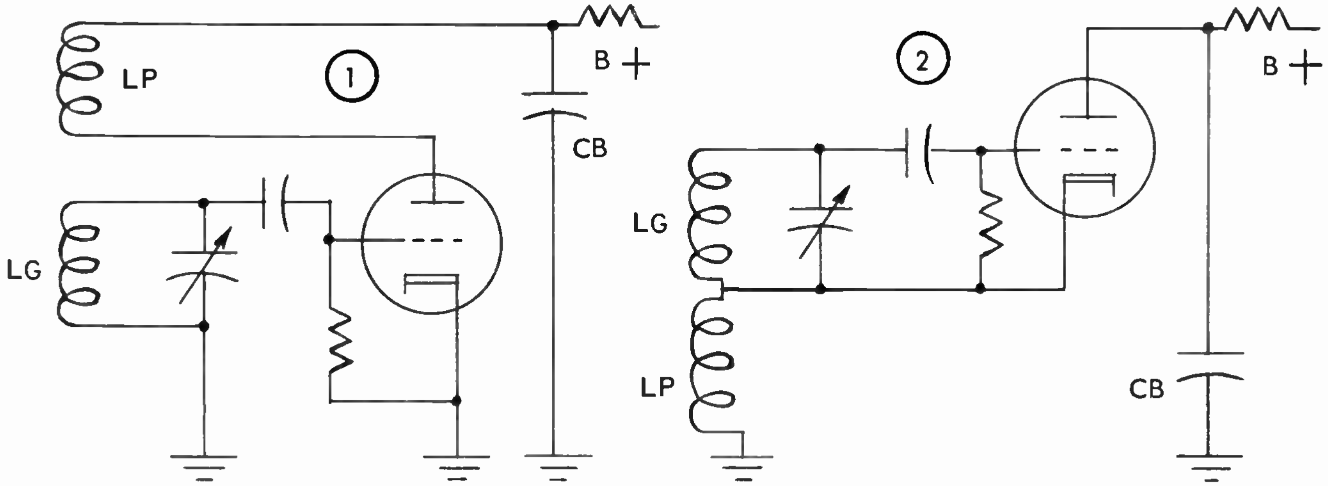


Fig. 33-6. Typical circuits for tuned-grid tickler feedback oscillators.

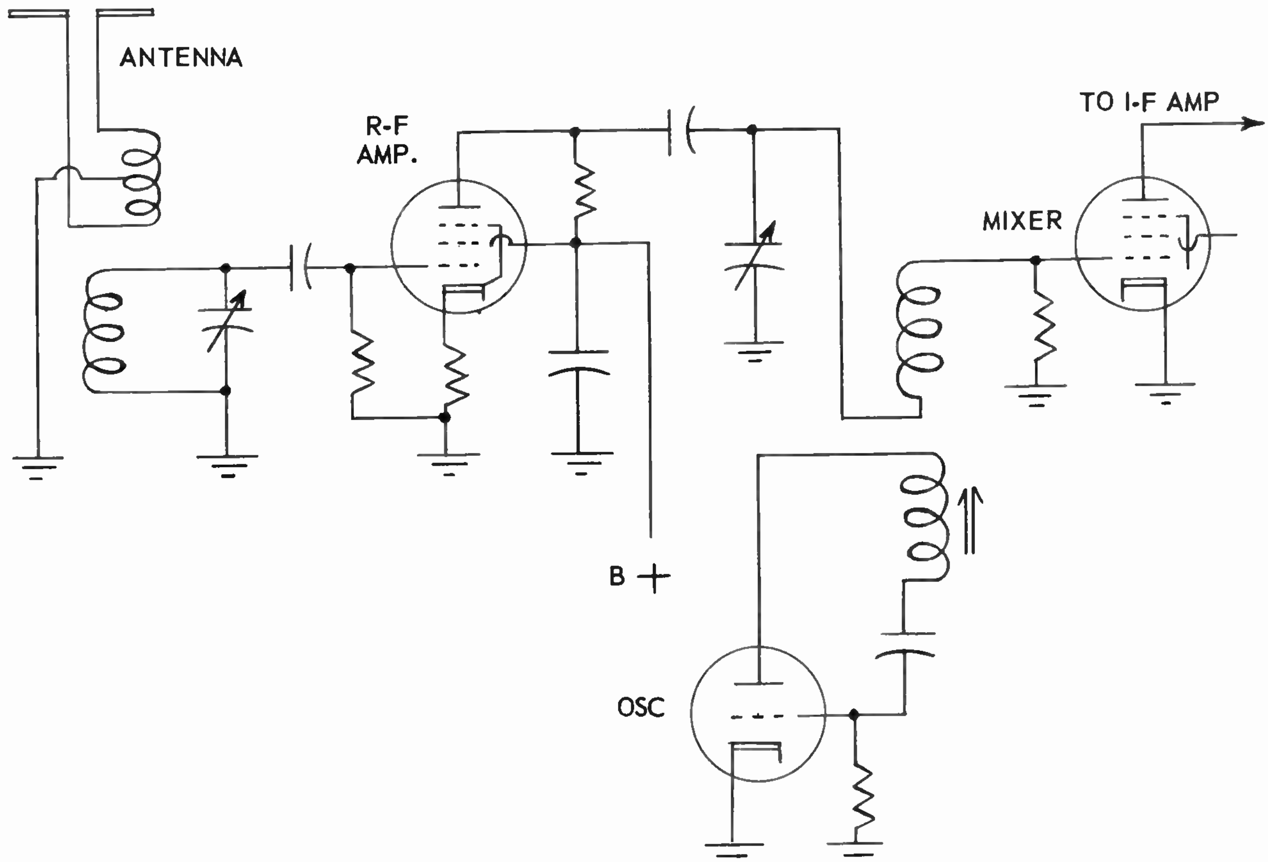


Fig. 33-7. Pentode mixer in a television tuner circuit.

through the antenna transformer secondary by means of capacitor *Cb*. Without this capacitor the d-c bias could not be maintained at the mixer grid. In this tuning arrangement there is no r-f amplifier, the mixer grid being coupled to the antenna through the antenna transformer.

5 Tuned-grid tickler feedback oscillators often are shown by diagrams generally similar to that at *1* in Fig. 33-6. The grid winding *Lg* is tuned by a variable capacitor. The feedback winding or tickler winding is marked *Lp*. The strength of feedback energy is increased by closer coupling between the two windings, and is decreased by looser coupling. Any change of coupling alters the mutual inductance in both windings and varies the resonant frequency of the grid circuit because of increased or decreased total inductive effect in the grid circuit. The r-f feedback circuit is completed from the plate through blocking capacitor *Cb* and through ground to the lower end of the winding in the grid circuit.

Diagram 2 of Fig. 33-6 illustrates another tuned-grid tickler feedback oscillator circuit. The grid winding *Lg*, tuned by a variable capacitor, is connected between the grid and cathode of the tube. The feedback winding or tickler winding *Lp* is inductively coupled and also is conductively connected to the grid winding. R-f plate current for feedback of energy to the grid passes through blocking capacitor *Cb* and ground to the lower end of the feedback winding, and through this winding to the tube cathode. There is grid-leak bias for the oscillators in both diagrams of Fig. 33-6.

COMMERCIAL TRADES INSTITUTE

The circuit of diagram 2 bears considerable resemblance to that for a Hartley oscillator, inasmuch as a tap on the winding is connected to the tube cathode, and both sides of the tuning capacitor are at r-f potential. In this circuit, however, only the grid winding is tuned by the variable capacitor, while in a Hartley circuit both windings are tuned by the one capacitor.

10 Our apparent emphasis on the different basic kinds of oscillator circuits may have led you to believe that it is essential to identify the type. Identification is not absolutely necessary, but often it is convenient in order that you may know what to expect when servicing various tuners. As an example, you will learn that Hartley oscillators and tuned-grid tickler feedback types are capable of delivering strong voltage outputs, but are not so well adapted as Colpitts types to operation at the very-high frequencies of television tuners. The chief reason is that Colpitts oscillators have less tendency to generate strong harmonic frequencies. The characteristics of the several oscillator circuits will become more apparent as we get further along in our work with television and testing instruments.

PENTODE MIXERS. Fig. 33-7 shows the circuit connections for a television tuner in which there is a pentode r-f amplifier, a pentode mixer, and a triode oscillator. There are three separate tubes, one for each function. The oscillator and mixer are not combined as a dual tube in a single envelope because, at present the available triode-pentode combinations are not well suited to very-high frequency operation. Therefore, tuners with pentode mixers and r-f stage are three-tube types.

The grid of the r-f amplifier is coupled to the antenna through a transformer whose secondary is tuned by an adjustable capacitor. There is cathode bias for the grid of this tube. Signals from the r-f plate go to the mixer grid through a modification of tuned impedance coupling.

The oscillator of this tuner is a Colpitts type using tube and wiring capacitances for feedback. Biasing is by the grid-leak method. High-frequency voltage from the oscillator is put into the mixer grid circuit by inductive coupling between the coils or windings in the two circuits. Note that the symbol for the oscillator coil indicates that there is a movable core for varying the inductance and the resonant frequency of the

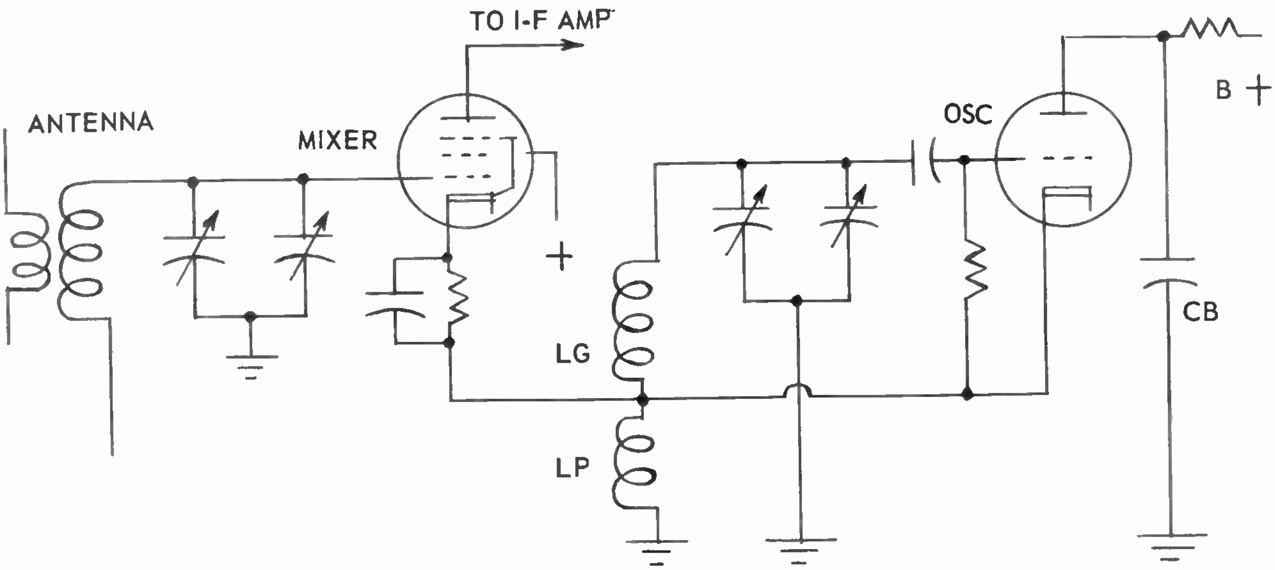


Fig. 33-8. Oscillator coupled to mixer through cathode line of mixer.

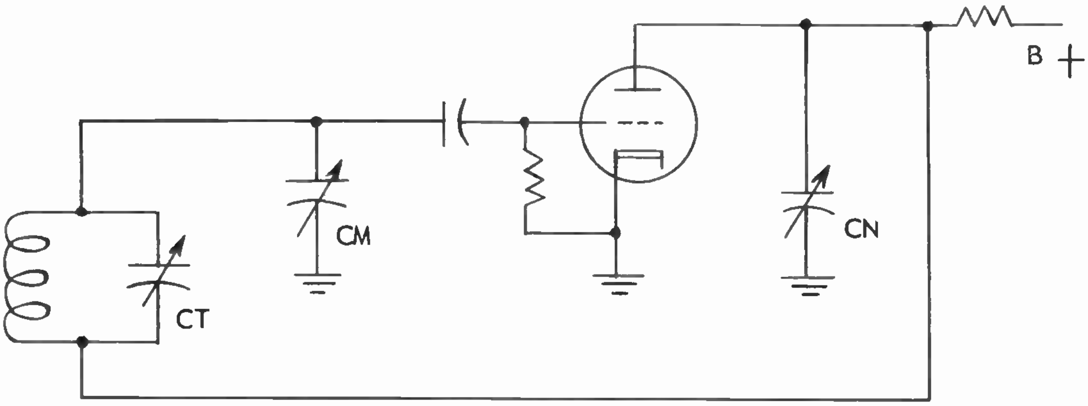


Fig. 33-9. An ultraudion oscillator circuit.

oscillator circuit. Separate triode oscillator tubes for television and f-m tuning usually are type 6C4, type 6AB4, or may be one plate and one grid of a 6J6 twin triode with the other plate and grid grounded.

Fig. 33-8 shows connections in the tuner section of an a-m sound radio. Here again we have a pentode mixer and a triode oscillator. There is no r-f amplifier. The tuned grid circuit of the mixer includes the secondary winding of the antenna coupling transformer and a variable tuning capacitor with paralleled trimmer capacitor. There is cathode-bias for the grid. Note especially that the cathode line from the mixer goes first through the biasing resistor and its bypass capacitor, then through winding *Lp* of the oscillator circuit and to ground.

The oscillator is a Hartley type with the tap of the tuned coil connected to the tube cathode. The variable tuning capacitor and paralleled trimmer are connected across the entire coil from the upper end or grid end to the lower end through ground. The plate circuit is completed for r-f currents and voltages through blocking capacitor *Cb* and ground to the lower end of the oscillator coil. Biasing is by means of the familiar grid-leak method.

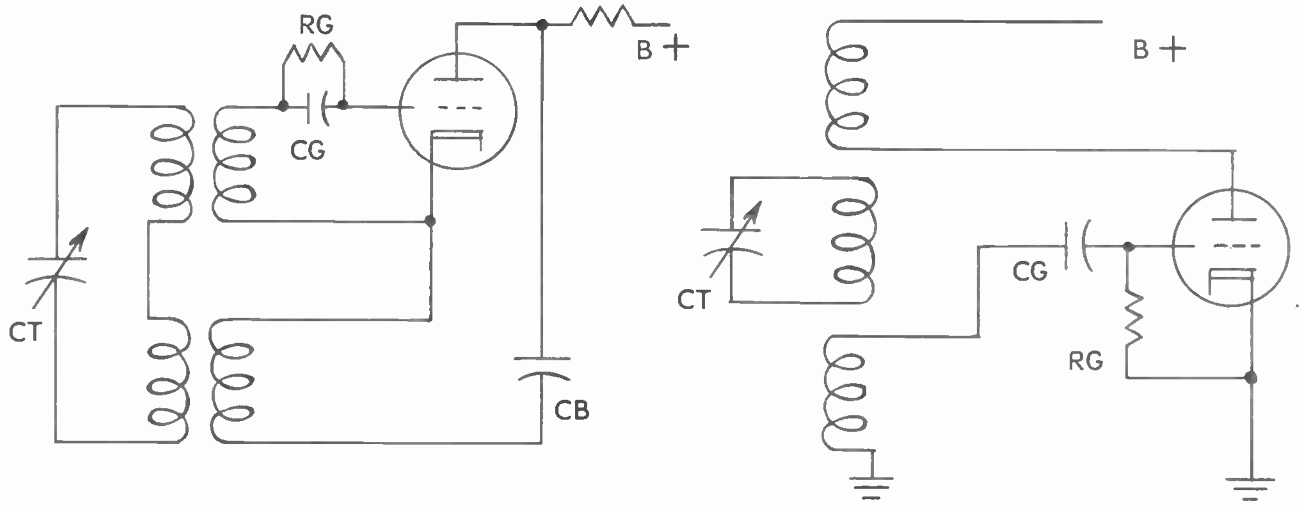


Fig. 33-10. Two variations of the Meissner oscillator circuit.

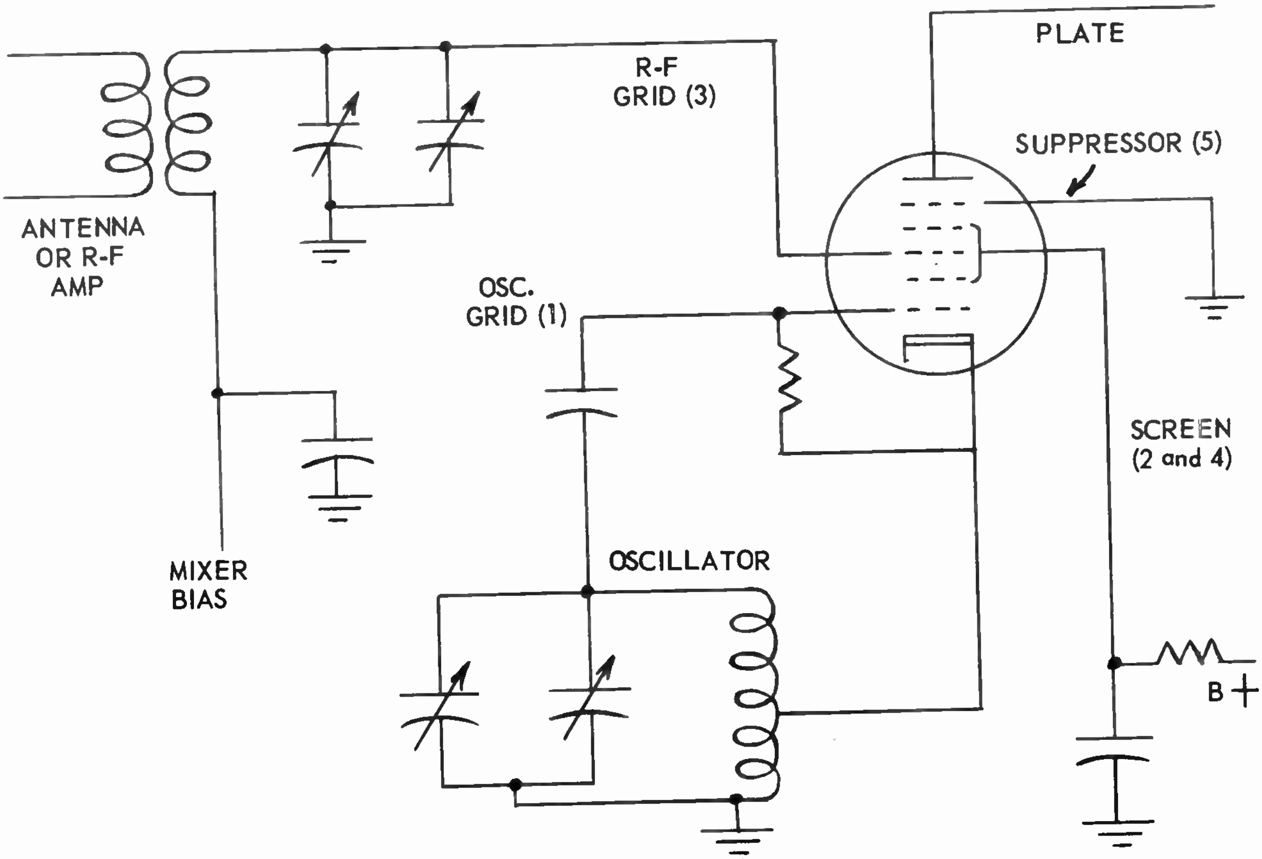


Fig. 33-11. A pentagrid converter performing the functions of oscillator and mixer.

Here we have a new method of getting the high-frequency voltage from the oscillator circuit into the mixer grid circuit. As noted before, the mixer cathode line goes through section *Lp* of the oscillator coil. In this section are oscillator current and voltage, consequently the cathode current of the mixer is varied or modulated at oscillator frequency. The mixer cathode is in both the plate circuit and the grid circuit of the mixer, and so we have the oscillator frequency in the mixer grid circuit. This is known as cathode coupling for the oscillator and mixer. It is a method often employed in standard broadcast radio receivers.

OTHER OSCILLATORS. For a little additional practice in following the action of feedback oscillators we shall look at two more types. One of the oldest oscillator circuits is the ultraudion, for which connections are shown by Fig. 33-9. It is equivalent to a Colpitts oscillator to which has been added the variable tuning capacitor *Ct* and in which there are either one or both the feedback control capacitors *Cm* and *Cn*.

Capacitor *Cm* is between the grid connection and the cathode, just as is the grid capacitor in the Colpitts circuit. Capacitor *Cn* of the ultraudion oscillator is between the plate and the cathode of the tube, as is the Colpitts plate capacitor. Capacitor *Ct* of the ultraudion circuit is in the same position as the added tuning capacitor in one modification of the Colpitts oscillator. Often it is difficult to know whether you are looking at a Colpitts or an ultraudion, and some television manufacturers have stated in their instruction literature that they use an ultraudion type, although it might also be considered a Colpitts.

An oscillator having separate plate and grid inductance is the Meissner type, of which two variations are shown by Fig. 33-10. The resonant circuit which contains tuning capacitor *Ct* may have two windings, as at the left, or only a single coil, as at the right. The coil in the grid circuit and also the one in the plate circuit are coupled to the tuned resonant circuit. Feedback energy from the plate coil passes through the tuned circuit to the grid coil and grid circuit. The Meissner circuit once was popular in broadcast radio receivers, but now is seldom used except in certain industrial electronic applications.

CONVERTERS. In our work with oscillators and mixers we have seen the r-f signal voltage and the oscillator voltage applied together to the grid of the mixer. We also have seen the oscillator voltage introduced through the mixer cathode. Now we come to a class of combined oscillator and mixer tubes in which there are seven elements serving both functions, but with only a single electron stream passing from cathode to plate through all the other elements. These tubes are called frequency converters, or more usually are called simply converters.

In most circuits previously studied the oscillator voltage has been carried into the mixer circuit through a small capacitor or else by means of mutual induction between two coils or windings. In converter tubes there is no need for either capacitive or inductive couplings between oscillator and mixer, for the r-f voltage and oscillator voltage both act on the single electron stream. The electron flow is modulated by both voltages. This is electron coupling.

Fig. 33-11 shows one kind of tuning circuit in which there is a converter instead of separate oscillator and mixer tubes, or instead of a twin triode with two electron streams. The converter has five grids between its cathode and plate, so usually is called a pentagrid converter, from the Greek word pronounced penta and meaning five. If we count also the cathode and plate the converter has seven active elements in all, and so may be called a heptode converter, from the Greek word hepta which means seven.

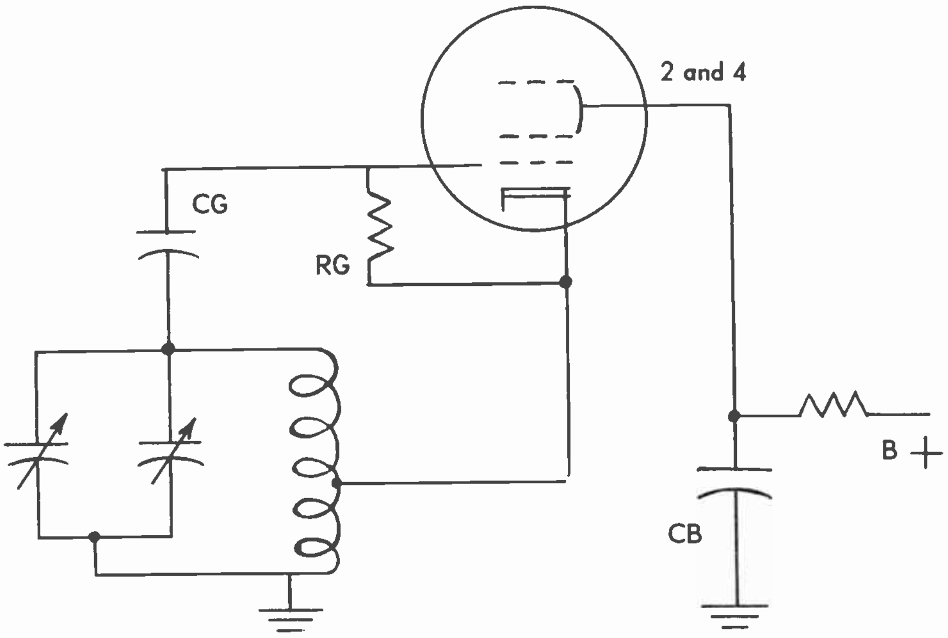


Fig. 33-12. The oscillator circuit of the converter.

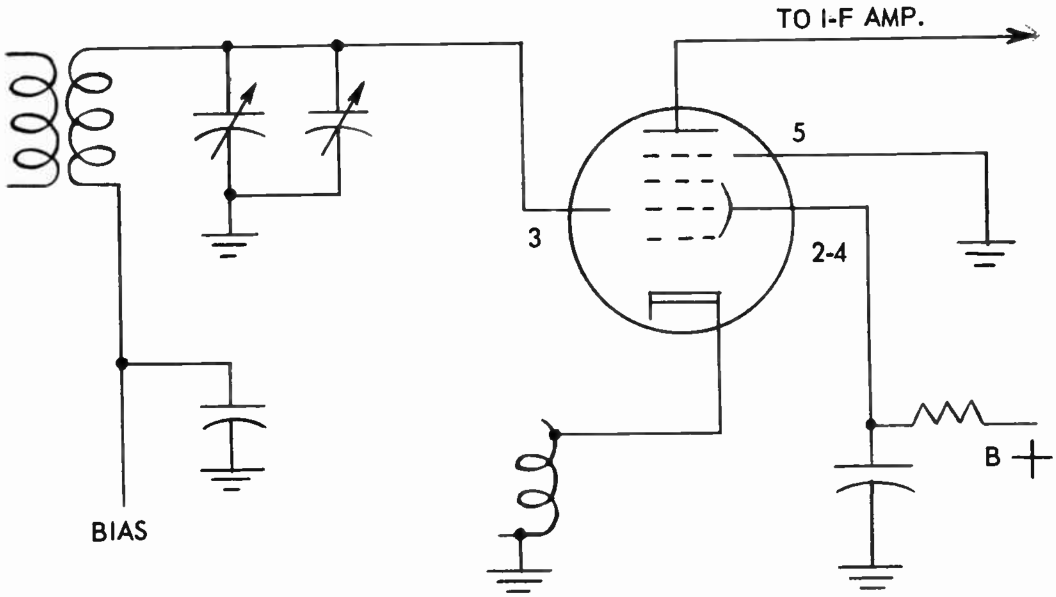


Fig. 33-13. The mixer circuit of the converter.

The names and numbers of the several grids are marked on the conductor lines leading to them in the diagram. Grid number 1, nearest the cathode, is the oscillator grid. Grids number 2 and number 4 are connected together within the tube, and act as a screen for the r-f grid, number 3' The screen reduces the internal capacitance of the tube and also accelerates electrons between cathode and plate, just as does the screen in any other tube. Grid number 5 is a suppressor, serving the same purpose as the suppressor in a pentode.

The oscillator circuit, a Hartley type, is shown by itself in Fig. 33-12. The tap on the coil is connected to the tube cathode. The upper end of the coil is connected through capacitor *Cg* to grid number 1, which is the oscillator grid. Grids number 2 and 4 act as the plate or anode for the oscillator. These grids are supplied with B+ voltage, and are connected through capacitor *Cb* and ground to the lower end of the oscillator coil, thus completing the feedback circuit. The oscillator is biased by the grid-leak method, with capacitor *Cg* and resistor *Rg*. Thus we have the Hartley oscillator circuit operating in connection with a cathode, a grid, and what amounts to a plate or anode, just as though these parts of the converter were a triode tube.

The mixer circuit of the pentagrid converter is shown by itself in Fig. 33-13. For all practical purposes we have here a pentode mixer. R-f signal voltage is applied to grid number 3. Grids 2 and 4 form the screen, with suppressor grid number 5 between screen and plate, just as in a regular pentode. Grid number 3 is negatively biased to make this section of the tube act like a detector and deliver the beat frequency in the same manner as with any other mixer.

8 The electron stream flowing from cathode to plate is first acted upon by the high-frequency voltage from the oscillator, applied to grid number 1. Then the same electron stream, after passing through screen grid number 2, is acted on by the r-f signal voltage applied to grid number 3. Then the doubly modulated electron stream proceeds to the plate.

Fig. 33-14 is another pentagrid converter circuit. The functions of each of the grids are the same as previously explained. For the oscillator we now have a tuned grid tickler feedback arrangement. The tuned grid coil is marked *Lg*. The tickler or feedback coil, marked *Lp*, actually is connected between the tube cathode and ground, but is effectively in the plate circuit. This is because by far the stronger r-f current in the cathode line is plate current, and because the r-f circuit for the plate is completed through capacitor *Cb* and ground back to the cathode through coil *Lp*.

In the converter of Fig. 33-14 the suppressor is internally connected to the cathode. This is true of type 6SA7-GT, an octal based glass tube, of type 6BE6, a 9-pin miniature tube, and also of the lock-in type 7Q7. In the converter of Fig. 33-13 the suppressor is connected to a separate base pin. This is the construction used in the type 6SA7 and type 6SB7-Y metal tubes and also in the type 6BA7 9-pin miniature tube. The five types just mentioned include nearly all the heater-cathode converters now in general use. The types 6SA7, 6SA7-GT and 7Q7 are used in a-m broadcast receivers and in some combination a-m standard broadcast and short-wave receivers. All the other types are used in f-m sound receivers as well as in standard broadcast and short-wave sets.

Pentagrid converters 6SA7, 6SA7-GT, and 7Q7 operate well at radio frequencies up to about 12 or 15 megacycles. At higher frequencies it becomes difficult to maintain uniform oscillator output voltage, the signal sensitivity drops off quite rapidly, and there is a marked tendency for tuning of the r-f input circuit to alter the oscillator frequency. In the other converters which have been mentioned these troubles have been reduced to an extent that allows satisfactory operation through the f-m broadcast band, or up to 108 megacycles. None of the converters presently available are used for television receivers where, on the high-band from channel 7 through channel 13, signal frequencies range from 174 to 216 megacycles.

In standard broadcast a-m receivers and also in combination standard broadcast and international short wave sets we find converters. In combination f-m and standard broadcast receivers there may be a converter

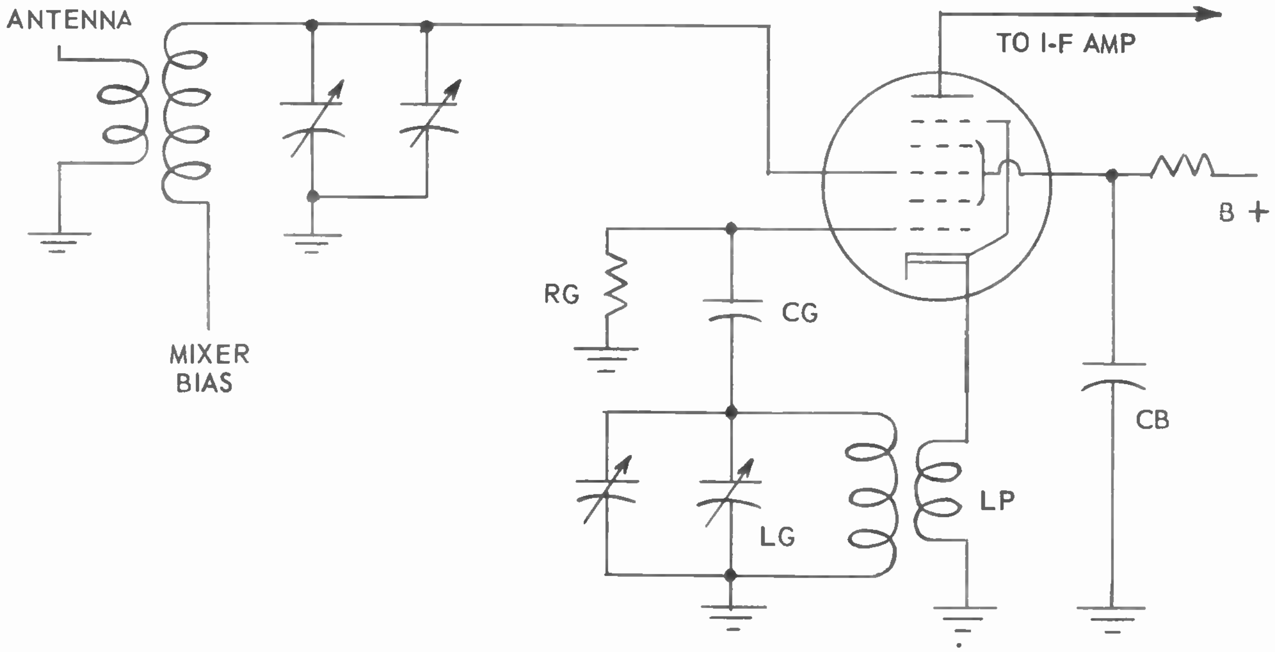


Fig. 33-14. Pentagrid converter with oscillator feedback from winding in cathode line.

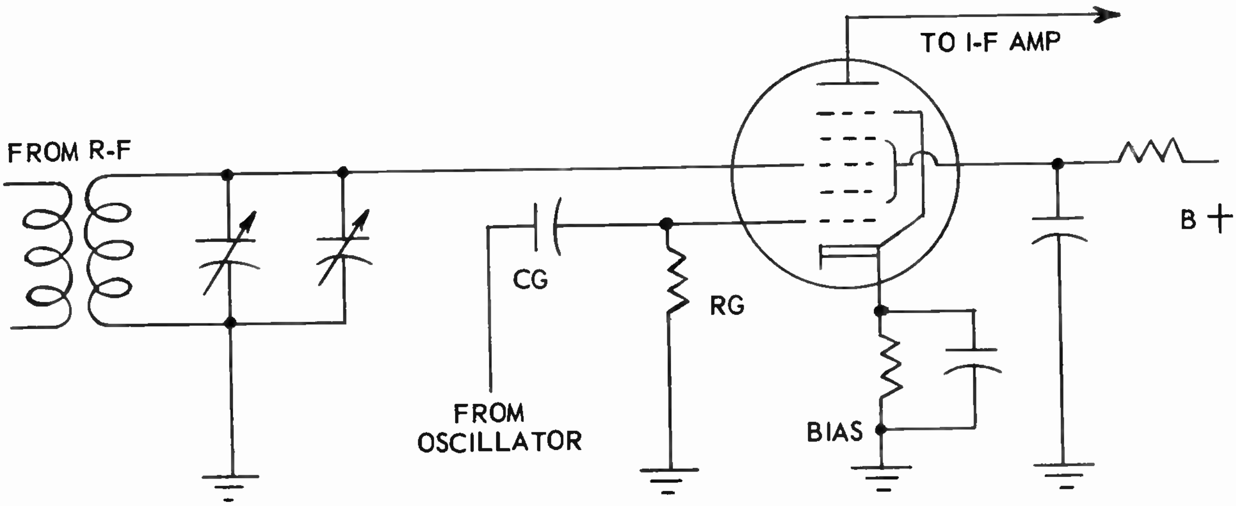


Fig. 33-15. Pentagrid converter tube used as mixer, with separate oscillator.

suitable for f-m service, or there may be a low-frequency converter for the standard broadcast band and also separate mixer and oscillator tubes for the f-m band. The converter and also the separate mixer and oscillator tubes are used in many combination receivers for television, f-m broadcast, and standard a-m broadcast service. In straight television receivers, straight f-m broadcast receivers, and in combination television and f-m sets there will be separate mixer and oscillator tubes,or a twin triode serving the two functions.

In some receivers other than television types you will find a pentagrid converter used as a mixer tube, with a separate triode oscillator. This is done to reduce or prevent changes of oscillator frequency when the r-f signal circuit is tuned. Such an effect on oscillator frequency usually is called "pulling". Connections for a pentagrid converter used as a mixer are shown by Fig. 33-15. Grid number 1, the oscillator grid, is connected through a small capacitor or possibly by inductive coupling to any type of separate oscillator. Bias for the mixer is shown as a cathode resistor type. With the r-f grid shielded from the oscillator grid by the screen grids 2 and 4 there is only electron coupling within the tube. Any of the pentagrid converters may be used in this manner. One of the older types of tubes, the 6L7 pentagrid mixer, was especially designed for this kind of circuit, but its frequency characteristics are not satisfactory for other than standard a-m broadcast and the lower short-wave frequencies. The 6L7 is now considered obsolete.

OSCILLATOR VOLTAGE. The greatest conversion gain or greatest signal voltage output is realized from a mixer or converter when oscillator r-f voltage is kept as high as possible. The limiting voltage is that which, when beating with the r-f signal, gives peaks which equal the mixer grid bias voltage. Anything more than this causes flow of grid current in the mixer or mixer section, and results in distortion of the signal.

2 Earlier we talked about how oscillator frequency is determined by the resonant frequency of inductance and capacitance in the oscillator tuned circuit, but said nothing about how far the oscillating voltage may swing in the positive and negative directions. For any given type of oscillator tube this oscillation amplitude is determined chiefly by the applied plate voltage and the impedance or resistance of the plate circuit load through which flows the plate current. The action is as follows.

We may assume, to begin with, that feedback is such that the grid voltage is becoming less and less neg-

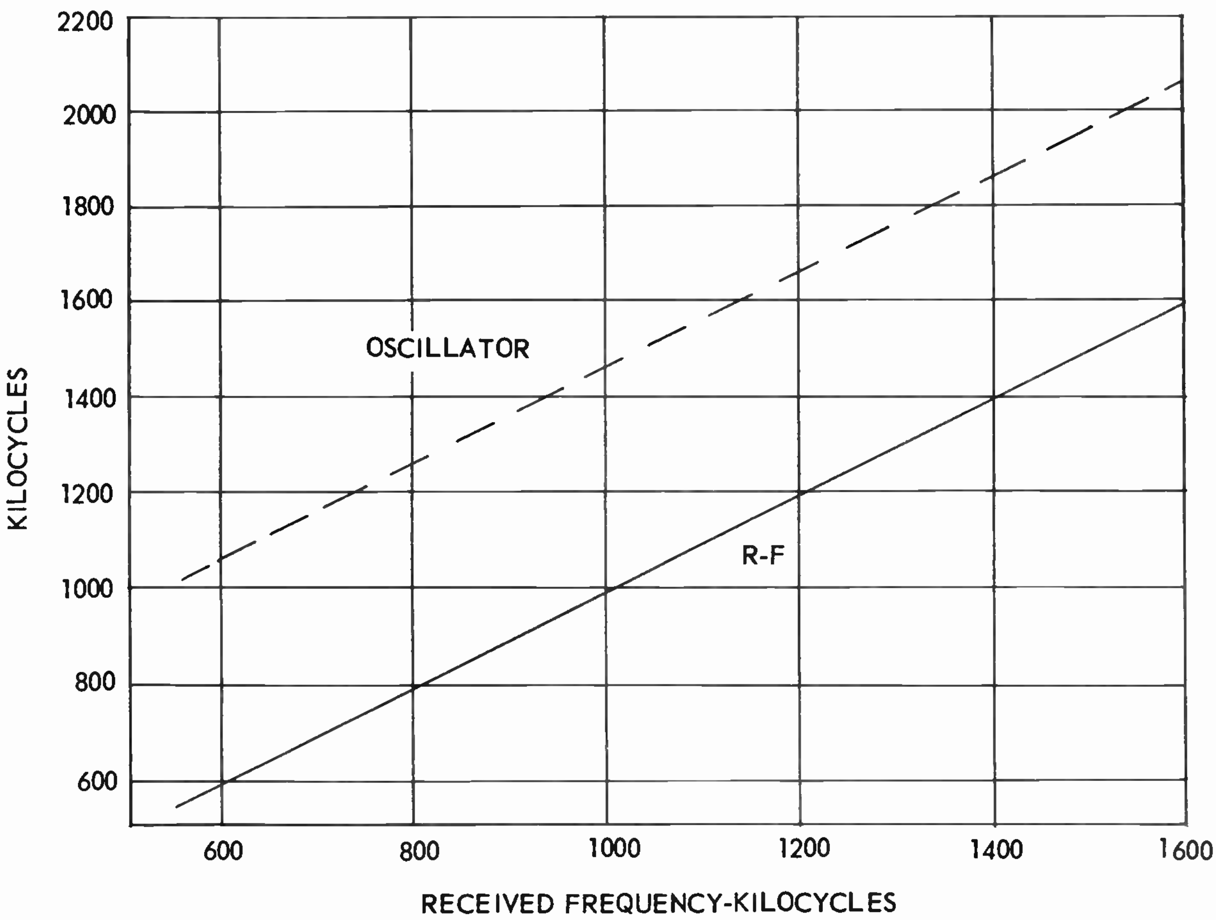


Fig. 33-16. Oscillator frequency remains 455 kc higher than radio frequency of the received signal.

ative or is changing in the positive direction. The result is an increase of plate current and a continued feedback that drives the grid even more positive. At the same time the increase of plate current through the load causes a greater drop of voltage in the load, and less and less positive voltage remains at the plate of the tube. This decrease of plate voltage tends to prevent the rise of plate current. Eventually the plate voltage drops so low that plate current can become no greater, even though the grid still is going positive. It is the *change* of plate current that has been causing the feedback, and when there is no further change the feedback ceases.

With no feedback there is nothing to drive the grid more positive, and grid voltage commences to change from maximum positive back toward zero. This really is a change of grid voltage in the negative direction, and a negative-going grid always causes plate current to decrease. The decreasing plate current now produces a feedback in a polarity which is the reverse of that with which we commenced, and the grid is driven more and more negative. The lessened plate current means less voltage drop in the plate load impedance or resistance, and voltage increases at the tube plate. But in spite of this rising plate voltage the grid continues going more and more negative and eventually it reaches the value for plate current cutoff.

When plate current is cut off, or when this current reaches zero value, there can be no further *change* of current. With no further change there is no feedback. With no feedback the grid can be driven no further negative, and grid voltage begins to return toward zero. This is a change of grid voltage in the positive direction. We commenced this explanation by assuming a grid voltage changing in the positive direction, and now we are back at that same point in the cycle. The process repeats over and over again, once for each cycle of oscillation.

Everything described in the three preceding paragraphs could happen in the most minute fraction of a second so far as the tube itself is concerned. How fast it does happen, or can happen, depends on the resonant frequency of the tuned grid-plate circuit. The maximum plate current reached before the increase is stopped by lack of plate voltage, and also the value of negative grid voltage at which there is plate current cutoff, both these depend on applied plate voltage and on the average grid voltage or the grid bias voltage at which the tube is operating. Amplitude is increased by higher applied voltage on the plate circuit. Also, since the oscillating voltage goes alternately positive and negative from the average voltage or bias voltage, the amplitude of oscillation is increased by making the bias more negative.

With biasing by the grid-leak method, as employed for nearly all oscillators, the value of negative bias depends on the resistance of the grid leak resistor and on the amount of rectified current flowing in this resistor. Bias voltage is equal to the product of ohms and milliamperes divided by 1,000 which is our regular rule for determining voltage when knowing resistance and current. The bias voltage may be measured directly without any very great error by connecting a sensitive electronic voltmeter across the grid leak resistor.

In no practical r-f oscillator circuit is there the same grid current or same grid leak current at all frequencies within the normal operating range. Consequently there will be changes of bias and of oscillating voltage amplitude with changes of tuned frequency. Total change throughout the tuning range ordinarily is no more than 20 per cent. For example, the lowest bias voltage might be 10 volts and the highest 12 volts.

If the receiver or other apparatus containing an r-f oscillator tunes in more than one frequency band, as in both standard broadcast and short-wave bands, the oscillator voltage may be quite different in the various bands. As a rule the voltage will be relatively high on low-frequency bands and will be lower on the higher-frequency bands. This is far from being a universal rule, as may be seen from the following oscillator grid voltage measurements made on a combination receiver having a standard a-m broadcast band, an a-m short-wave band, and an f-m band. All measurements were made with an electronic voltmeter across the grid resistor.

	Frequency	Volts	Frequency	Volts
Standard broadcast	550 kc	– 2.7	1,600 kc	– 5.2
Short-wave	9.5 mc	– 10.5	16.0 mc	– 15.5
F-m broadcast	88 mc	– 9.0	108 mc	– 6.5

In this particular receiver the oscillator grid voltage is lowest in the lowest frequency (standard broadcast) band, is highest in the medium frequency (short-wave) band, and is in between on the highest frequency (f-m) band. In the standard broadcast and short-wave bands the oscillator grid voltage increases negatively as the tuned frequency is increased, while in the f-m band the grid voltage decreases as the tuned frequency is increased.

7 If oscillator grid voltage drops to zero at some frequency, or shows very wide fluctuations with a small change of frequency, there is trouble in the oscillator circuit. As an example, the tuning capacitor might have a short circuit between rotor and stator plates at some points in tuning.

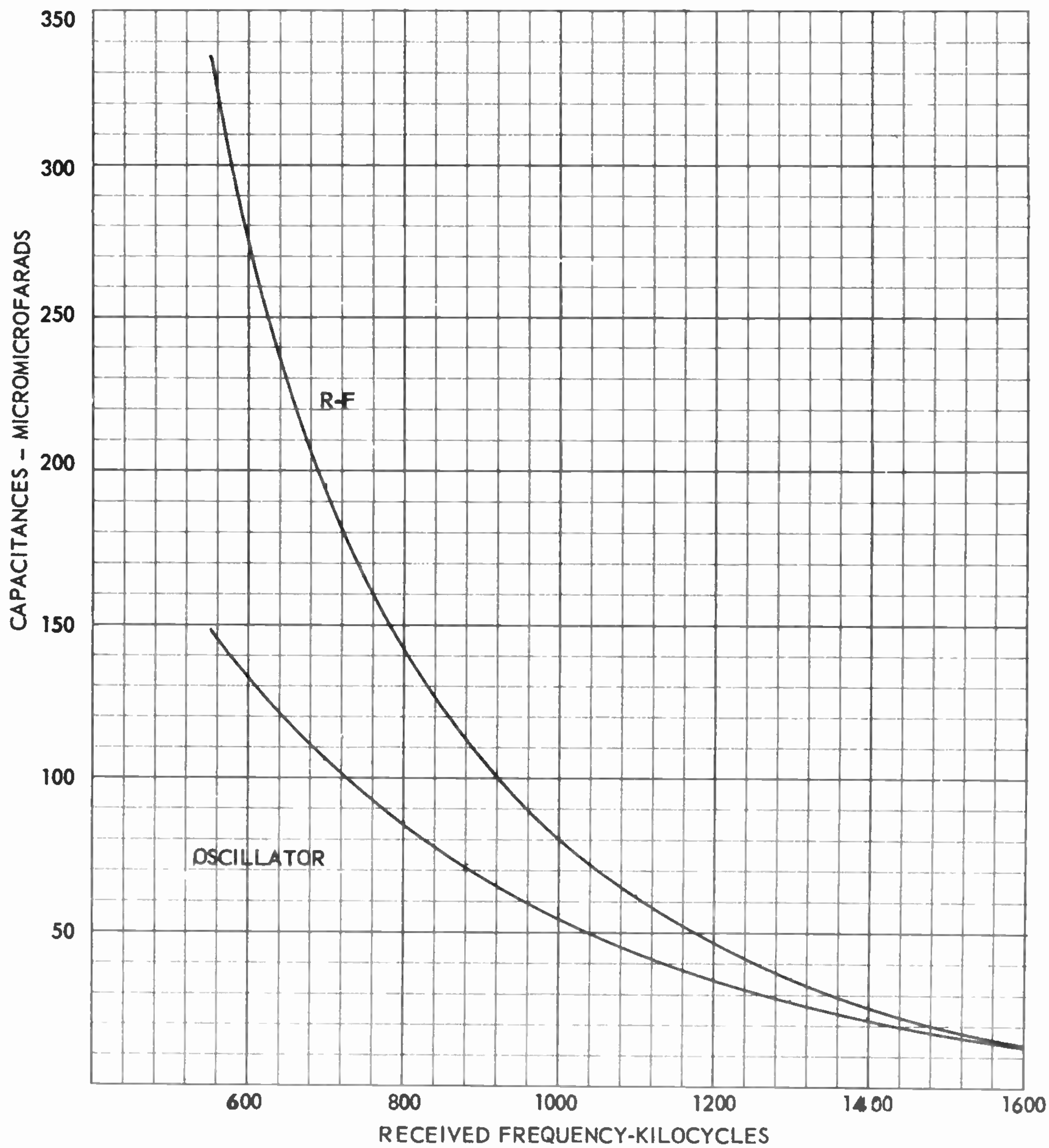


Fig. 33-17. How radio frequency and oscillator frequency must change together throughout the range of received signal frequencies.

TRACKING OF OSCILLATOR AND R-F TUNING. In nearly all receivers designed for standard broadcast and international short wave reception the intermediate frequency is either 455 or 456 kilocycles. In f-m receivers the intermediate frequency is quite well standardized at 10.7 megacycles. The intermediate frequency for any superheterodyne receiver must remain constant for all received carrier frequencies. Since the intermediate frequency is the difference between oscillator frequency and received radio frequency it is necessary to maintain a constant difference between these two frequencies.

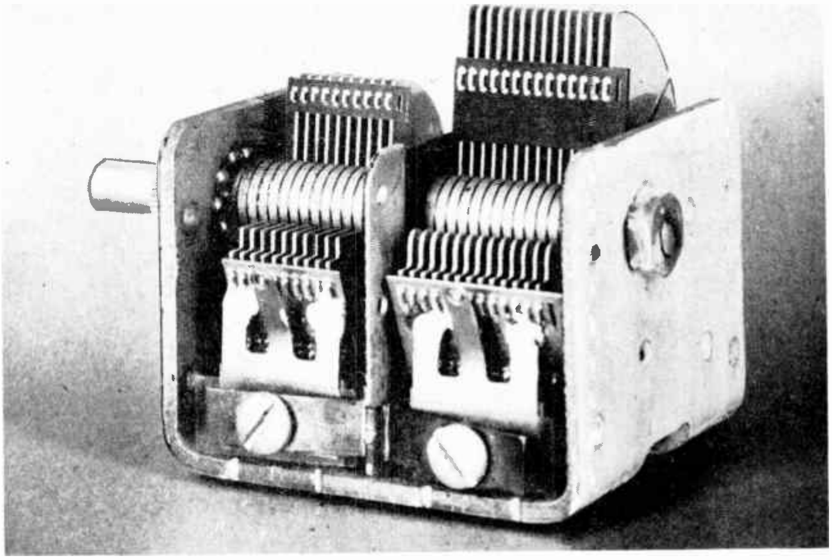


Fig. 33-18. A cut plate variable tuning capacitor.

The oscillator frequency almost always is higher than the received radio frequency. This means that in standard a-m broadcast and most short-wave receivers the oscillator frequency must remain 455 or 456 kc higher than the radio frequency, and in f-m receivers the oscillator frequency must remain 10.7 mc higher than the received frequency throughout the tuning range. Oscillator and radio frequencies for a standard a-m broadcast receiver may be represented as in Fig. 33-16. Throughout the reception range the oscillator frequency is 455 kc higher than the radio frequency which is applied to the mixer.

If tuning is by means of fixed inductances and variable capacitors the tuning capacitances must change together in such manner as to preserve the correct difference between oscillator and radio frequencies. If tuning is with variable inductances and fixed capacitances there must be similarly suitable changes of inductances. Circuits with which the frequency difference remains constant or practically so are said to be tracked.

As an example of what is necessary in order to accomplish satisfactory tracking we shall consider a case of standard broadcast tuning with variable capacitors covering the range from 550 to 1,600 kc with a 455-kc intermediate frequency. Since oscillator frequency always remains higher than radio frequency we may use less inductance and less capacitance for oscillator tuning than for r-f tuning. For our example we shall assume in the r-f coil an inductance of 230 microhenrys, and 140 microhenrys, in the oscillator coil.

By using formulas given in an earlier lesson you can figure out the changes of r-f and oscillator tuning capacitances required with these inductances. To tune the r-f circuit from 550 to 1,600 kc you will find that the capacitance must change from approximately 365 mmfd to 43 mmfd. To tune the oscillator circuit from 1,005 to 2,055 kc (455 kc above r-f tuning) the capacitance must be varied from about 179.2 mmfd to 42.9 mmfd.

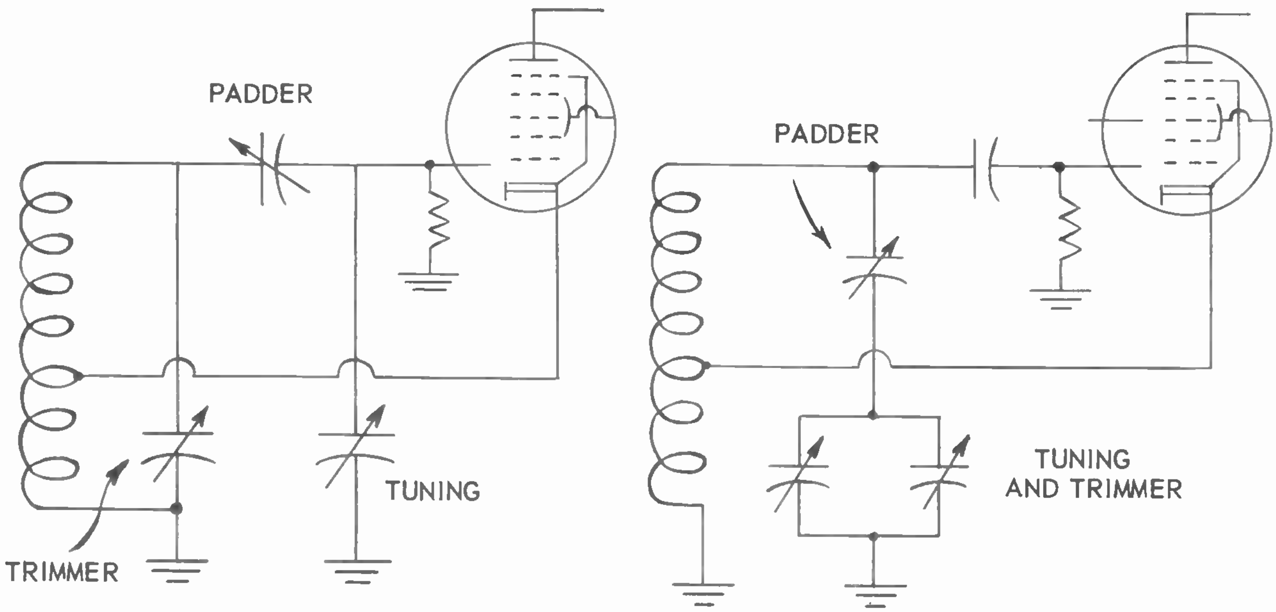


Fig. 33-19. Connections for padder capacitors which maintain tracking.

Now we shall make the further assumption that fixed capacitances in each circuit amount to 30 mmfd. This will include distributed capacitances of the coils, input capacitances of the tube or tubes, and stray capacitances in wiring and other parts. Subtracting these 30-mmfd fixed capacitances from the required variations of tuning capacitance shows that the r-f tuning capacitor itself must provide a change from 335 to 13 mmfd, and that the oscillator tuning capacitor must provide 149.2 to 12.9 mmfd. By computing the tuning capacitances for many received frequencies in between 550 and 1,600 kc we would obtain values for drawing the two curves of Fig. 33-17. These curves show capacitance changes required to maintain a 455-kc intermediate frequency with our assumed values of inductance.

One way of providing the required simultaneous changes of r-f and oscillator capacitances is to use a "cut plate" tandem tuning capacitor such as illustrated by Fig. 33-18. The section of smaller plates and smaller capacitance, at the left, is for the oscillator circuit. The section of larger capacitance, at the right, is for r-f tuning.

A cut plate tandem capacitor may be designed to cover any one frequency band. But if we wish to use a tandem tuning capacitor for one or more additional frequency bands the same cut plate design will not work in the added bands. If you care to figure out the changes of capacitance for any short-wave band, as from 8 to 25 mc, and compare the required changes with those of Fig. 33-17 it will be apparent that relations between simultaneous oscillator and r-f capacitances are quite different.

For tuning in more than one band with the same set of capacitors we may employ identical capacitor sections for both the oscillator and the r-f circuits, but in series with the oscillator tuning capacitor we must connect an adjustable capacitor called a "padder". Two methods of connecting padders are shown by Fig. 33-19. In either case the padder is effectively in series with the oscillator tuning capacitor, so far as the resonant circuit is concerned. The combined capacitance of series capacitors is less than that of either unit alone, so the padder reduces the capacitance of the oscillator circuit throughout the tuning range. The combined capacitance for any tuned frequency is equal to the product divided by the sum of the two capacitances.

If the oscillator coil is made with inductance suitable for use with a padder system it becomes possible to have precise tracking at three places in the tuning range, at high, low, and middle frequencies, and to have only small errors at other points. Obtaining correct inductance is a matter of original design.

At high frequencies, which require very small capacitance in the tuning capacitor, even large changes of padder capacitance have little effect on combined series capacitance. For example, with the tuning capacitor adjusted to 13 mmfd, changing a padding capacitance all the way from 300 to 200 mmfd varies the combined capacitance only from 12.46 to 12.20 mmfd.

6 At low frequencies the tuning and padding capacitances are more nearly equal, and every change of padder capacitance causes a large variation of combined capacitance or effective tuning capacitance. Because of this fact the padder is adjusted to obtain correct tracking or correct tuning at or near the low-frequency end of the tuning range.

3 But at the high-frequency end of the tuning range, with the tuning capacitor adjusted for least capacitance, the trimmer capacitor in parallel with the variable tuning capacitor has maximum effect. As an example, with a tuning capacitor set for 13 mmfd, a change of trimmer capacitance from 5 to 15 mmfd alters the combined parallel capacitance from 18 to 28 mmfd. For this reason the trimmer is adjusted for correct tuning at or near the high-frequency end of the frequency range. The trimmer and padder are service adjustments.

In a multi-band receiver using the same tuning capacitors for all frequency bands it is necessary to provide separate inductance coils and separate adjustable padders for each band. The higher the frequency band the smaller must be the tuning inductance and the greater must be the padder capacitance.

1 Tracking is a problem in standard broadcast receivers because of the high ratio of maximum to minimum frequencies. The ratio of 1,600 kc to 550 kc is about 2.9 to 1. Similarly high ratios are found in many shortwave bands. But in the f-m broadcast band the maximum received frequency is 108 mc and the minimum is 88 mc, for a ratio of only about 1.23 to 1. This makes tracking a minor problem. The frequency ratio of television channels 2 through 6 is 1.49 to 1, and of channels 7 through 13 it is about 1.2 to 1. Most television receivers have tuning inductances or capacitances, or both , which are separately adjustable or adjustable in groups for the channels or bands, which obviates any overall tracking problem. Television receivers having continuously adjustable inductances for the total variation of frequency have suitable means for adjusting the tracking. These matters will be investigated when we come to service alignment of television tuners.

TELEVISION

LESSON NO. 34
INTERMEDIATE-FREQUENCY AMPLIFIERS

Training Your Mind
C. T. I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind
C.T.I.
Perfecting Your Skill

COMMERCIAL
TRADES
INSTITUTE

Chicago, Ill.

LESSON NO. 34
INTERMEDIATE-FREQUENCY AMPLIFIERS

We have succeeded in obtaining from the mixer or converter a signal whose center frequency remains the same for every station or channel received, yet which carries the modulation for pictures or sound of each different program. The constant center frequency of this signal is the intermediate frequency coming from the mixer or converter. This modulated signal voltage will be put through the intermediate-frequency amplifier system of the receiver and then will go to the detector.

Since the intermediate frequency remains the same for all received signals, the transformers or other couplings in the i-f (intermediate-frequency) amplifier system may be tuned for this frequency and left that way. Tuning of these transformers or couplers is a service adjustment, it is not under the control of the operator. The only tuning to be done by the operator is that for the r-f oscillator, for the mixer, and for the antenna if the antenna coupling is a tuned type. Here we have advantage number one of the superheterodyne; simplicity of tuning.

The intermediate frequency always is lower than any received frequency. In an earlier lesson we learned that the lower the resonant frequency at which tuned circuits are operated the greater is their selectivity. So the second advantage of the superheterodyne is improved selectivity.

When talking about some of the old-time receivers we learned that undesired oscillation, with its resulting whistling and howling, occurs at the high-frequency end of the tuning range – unless amplification is kept at a very low value. Since the i-f amplifier of the superheterodyne operates at a frequency even lower than the low end of the tuning range, we may use very great gain or amplification in this amplifier without running into difficulty with oscillation. This possibility of high i-f gain is advantage number three of the superheterodyne.

Although the choice of intermediate frequency is made by the designer of the receiver, and cannot be changed to any great extent by service adjustments, we should understand a few of the facts which influence the choice.

In early superheterodynes for broadcast reception the intermediate frequencies were as low as 50 to 175 kilocycles. These low intermediates were chosen to avoid feedback through the plate-to-grid capacitance inside the amplifier tubes, for at that time all amplifiers were triodes, screen grid tubes and pentodes had not yet appeared.

The old triode amplifiers had transconductances of only 400 to 800 micromhos, and to obtain sufficient total amplification it was necessary to use three or more i-f stages with low-loss or high-Q couplings in every stage. Using many stages of amplification with low-loss construction increases the tendency to oscillate. To prevent such trouble it was necessary to use the low intermediate frequencies.

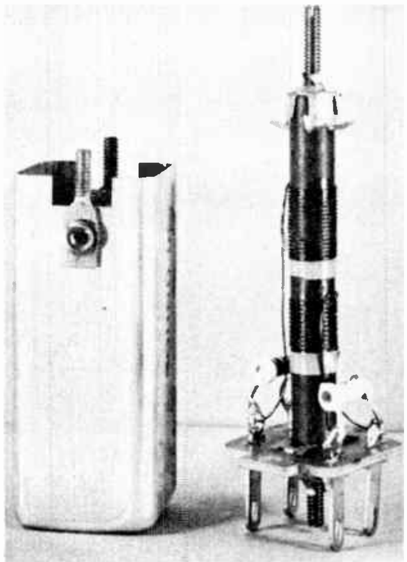


Fig. 34-1. An i-f transformer for television sound. The shield has been removed and set to one side.

Nowadays practically all i-f amplifier tubes are pentodes, in which there is very small plate-to-grid capacitance and through which there cannot be enough feedback to cause oscillation at any intermediate frequency which may be used. Furthermore, present i-f amplifier pentodes have transconductances of 2,000 to 6,000 micromhos, which makes it unnecessary to use many stages of i-f amplification to obtain all the gain that is desirable.

One of the principal advantages of high intermediate frequencies is a reduced tendency to pulling between mixer and oscillator tuned frequencies. Naturally, if the oscillator frequency differs from the incoming radio frequency by something like 455 kc the two circuits are much less likely to pull toward the same frequency than with a difference of only 175 kc or less. Another important advantage of higher intermediates is greater separation of "image frequencies" from receiver frequencies. This matter of image frequencies requires a little explanation.

IMAGE FREQUENCIES. Assume that your superheterodyne broadcast receiver operates with an intermediate frequency of 455-kc, and that you have it tuned to receive an r-f signal at 600 kc. Since the oscillator frequency will be equal to the radio frequency plus the intermediate frequency, the oscillator frequency will be 1,055 kc in this particular case. Supposing now that another broadcast signal at 1,510 kc can reach the mixer or converter of your receiver. This other signal frequency will beat with the oscillator frequency to produce a beat frequency equal to their difference. The difference between the 1,510-kc signal frequency and the 1,055-kc oscillator frequency is 455 kc, which is the intermediate frequency at which your receiver operates.

If both the desired r-f signal at 600 kc and the undesired r-f signal at 1,510 kc reach the mixer grid, both will produce a 455-kc intermediate frequency, and this intermediate frequency will carry the modulations or programs of both stations. The i-f amplifier of your receiver will be tuned for the intermediate frequency of 455 kc, so both programs will be amplified, detected, and reproduced from the speaker at the same time.

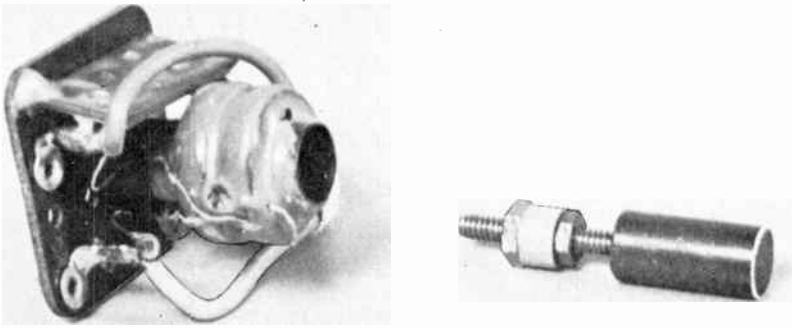


Fig. 34-2. A "slug tuned" i-f transformer for standard broadcast.

The 1,510-kc frequency is called an image frequency. It is 455 kc above the oscillator frequency, while the desired 600-kc signal is 455 kc below the oscillator frequency. An image frequency is one that is just as far above the oscillator frequency as the desired signal frequency is below the oscillator frequency. Another way of saying the same thing is to state that an image frequency differs from the desired signal frequency by twice the intermediate frequency. Twice our intermediate frequency of 455 kc is 910 kc. The image frequency of 1,510 kc is just 910 kc higher than the desired frequency of 600 kc.

To keep the image frequency away from the mixer grid or the r-f grid of a converter, we must have one or more tuned circuits between the grid and the antenna, into which, both the desired frequency and the image frequency are induced. If the circuit or circuits between antenna and mixer are tuned to the desired signal frequency, and if these circuits are reasonably selective, they will exclude the image frequency while accepting the desired signal.

The higher the intermediate frequency the farther the image frequencies will be placed above the desired signal frequencies. Supposing we were to use an intermediate frequency of 535 kc. When tuned to the lowest

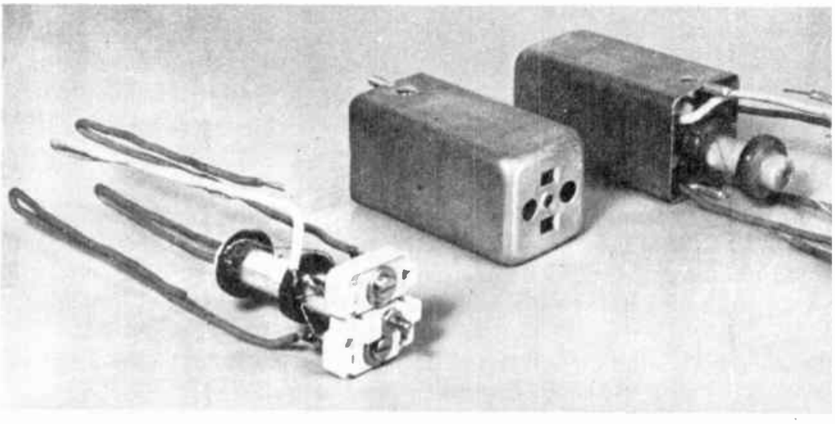


Fig. 34-3. I-f transformers with widely separated windings providing loose coupling.

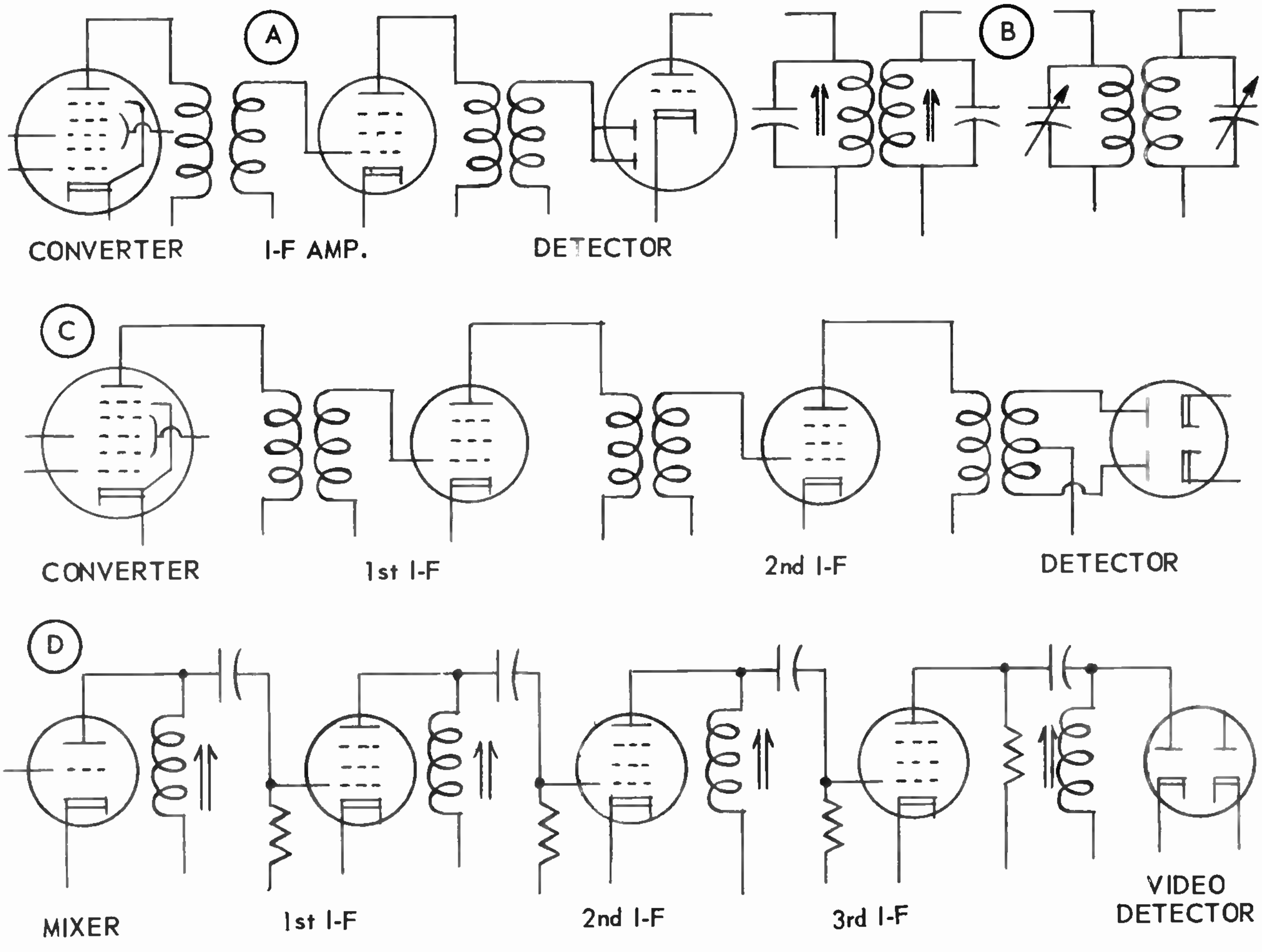


Fig. 34-4. Basic circuits for i-f amplifier sections of standard broadcast, f-m broadcast, and television receivers.

American broadcast carrier, 550 kc, the oscillator frequency would be 535 plus 550, kc, or would be 1,085 kc. The image would be at 1,085 kc (oscillator) plus 535 kc (intermediate), or would be 1,620 kc. This would be above the highest broadcast carrier frequency.

But with the i-f amplifier tuned to 535 kc it would require a high degree of selectivity to prevent the lower broadcast carriers, around 550 kc, from being picked up directly by wiring and parts in the i-f amplifier system, without having to come through the antenna and mixer circuits. To avoid such pickup it is necessary to use an intermediate frequency quite a bit lower than the lowest r-f signal frequency to be received.

If you assume an intermediate frequency of 455 kc and figure out the images for received frequencies between 550 and 690 kc you will find that all these images are within the standard broadcast band, between 1460kc and 1600 kc. If you compute the images for received frequencies between 840 and 1,600 kc you will find they fall between 1,750 and 2,510 kc, a range within which other radio services are operating.

INTERMEDIATE-FREQUENCY AMPLIFIERS. Fig. 34-4 represents, in greatly simplified form, the most common practices in i-f amplifiers for standard a-m broadcast, f-m broadcast, and television receivers. Diagram A shows standard broadcast practice. A pentagrid converter is coupled to the i-f amplifier pentode by means of an i-f transformer. The output of the i-f amplifier tube is coupled to the diode detector through a second i-f transformer. Both transformers are tuned to the intermediate frequency used in the receiver. The diode detector is one section of a duodiode-triode tube whose other section, the triode, acts as an audio-frequency amplifier.

Both the primary and the secondary windings of the i-f transformers may be tuned to resonance at the intermediate frequency in either of the ways shown by diagram *B*. With one method there are adjustable cores in each of the windings, to vary their inductances, and fixed capacitors are connected across each winding. With the other method there are adjustable capacitors across each of the windings.

Diagram *C* represents common practice for i-f amplifiers in f-m broadcast receivers. There is a pentagrid converter of a type suited for operation at the f-m carrier frequencies. There are two i-f amplifier pentodes with three i-f transformers for coupling between converter and first i-f amplifier, between the two i-f amplifiers, and from the second amplifier to the detector. The detector is a twin-diode connected into a special circuit suited for demodulation of f-m signals. The matter of f-m demodulation or detection will be investigated a little later. The coupling transformers may be tuned either by adjustable cores or else by adjustable capacitors, as shown by diagram *B*.

Diagram *D* of Fig. 34-4 shows what probably is most common practice in television i-f amplifiers which handle the picture or video signals. Following the mixer are three pentode i-f amplifiers. A good many sets use four such amplifiers. All the couplings between tubes consist of a single coil or inductor tuned to resonance by an adjustable core, for varying the inductance. Capacitance for the resonant circuit is furnished by internal capacitances of the tubes, by distributed capacitance of the coil, and by stray capacitances in wiring and other parts. The diode type video detector usually is one section of a twin-diode, with the second section employed for some other purpose.

More i-f amplifying stages are needed for f-m reception than for standard broadcast because it is not possible to obtain as much gain per stage at the higher intermediate frequency used in f-m receivers. We need still more i-f stages for television picture or video signals than for f-m broadcast reception because of the still higher intermediate frequencies employed with television, and the resulting still lower gains per stage.

There is less gain per stage at higher frequencies partly because high-frequency resistance goes up and the "Q" of tuned circuits goes down as we go into the ranges of intermediate frequencies used for f-m broadcast and television reception. However, the big reason for drop of stage gain at high frequencies is the decrease of reactance in tube capacitances, distributed capacitances, and stray capacitances. These decreased capacitive reactances act like bypass capacitors across all the tuned circuits, and it is impossible to get enough impedance in the plate circuits to obtain much amplification per tube.

The elementary circuits of Fig. 34-4 are only the most common kinds. There are many exceptions. In f-m receivers there may be more than two i-f stages, or there may be one or more single-coil couplers instead of all transformers. In television i-f amplifiers there may be two-winding transformers instead of single-coil couplers, or there may be some of each kind in the same amplifying system.

The principal object of any coupling device used in an i-f amplifier is to cause maximum amplification at the intermediate center frequency and the modulation frequencies on both sides of the center, while allowing minimum amplification at all other frequencies.

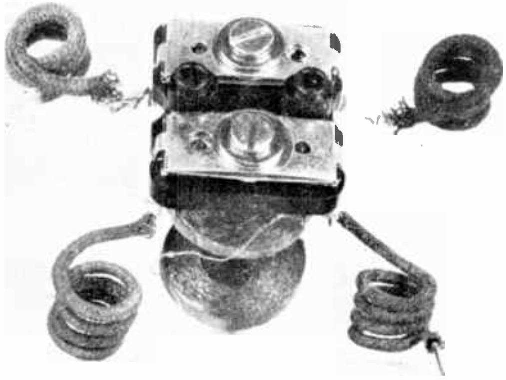


Fig. 34-5. An i-f transformer having adjustable mica capacitors for tuning the primary and secondary.

In the standard broadcast band each channel is 10 kc wide. This means that audio-frequency modulation may extend to 5 kc below the intermediate center frequency and to 5 kc above this center frequency. Any audio frequencies higher than these limits would mean that sideband frequencies are extending over into lower or higher channels. It follows that the ideal coupling device for standard broadcast reception would have maximum gain throughout a range of 10 kc, and zero gain everywhere else, as represented at *A* in Fig. 34-6.

For f-m sound broadcasting the maximum variations of frequency go to 75 kc below and 75 kc above the center frequency. Then the ideal i-f coupling device for this class of service would have frequency response as at *B*, giving maximum amplification throughout a range 150 kc wide, and zero everywhere else.

Transmission and reproduction of sound for television programs is handled by the frequency-modulation method, but maximum frequency variation extends to only 25 kc below and 25 kc above the center frequency, rather than to 75 kc as for f-m sound broadcasting. Consequently we would need maximum frequency response over a range of only 50 kc for f-m sound in television receivers.

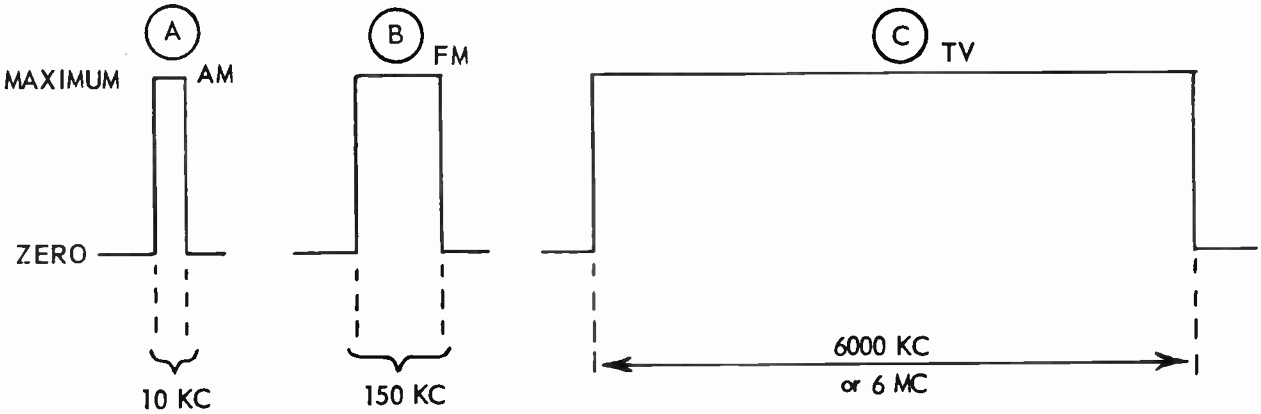


Fig. 34-6. Passbands are relatively narrow for a-m and f-m reception, but very wide for television.

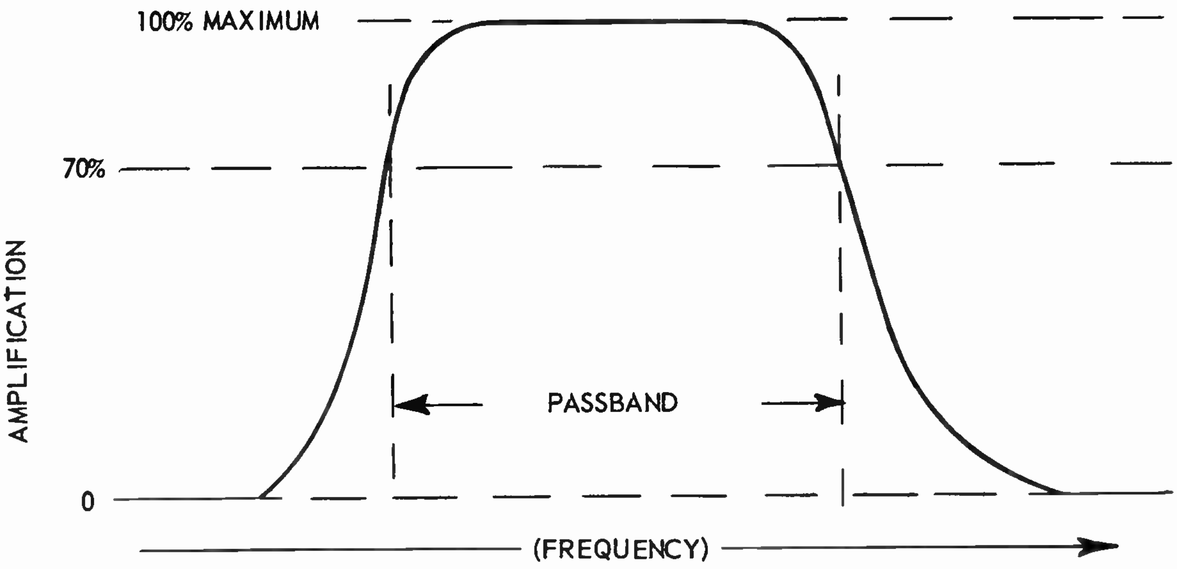


Fig. 34-7. Response curve showing coverage of a wide range of frequencies.

Signals for television pictures, for blanking, and for synchronization, are handled by amplitude modulation. These "video" signals extend over a maximum frequency range of about 5½ megacycles in the television i-f amplifier. Both these a-m video signals and the f-m sound signals for a television program are within a channel extending over a total frequency range of 6 megacycles. Consequently, the ideal frequency response for the television i-f amplifier would be 6 megacycles, as at *C* in Fig. 34-6.

It is impossible to construct i-f amplifiers with cutoffs of amplification so sharp at certain frequencies as indicated in Fig. 34-6, but the closer we come to the ideals the more satisfactory will be the performance. The actual range of frequencies throughout which there is voltage amplification of no less than 0.707 or about 70% of the maximum amplification usually is considered as the effective band width of an amplifier or of any single stage of amplification. Amplification or voltage gain at various frequencies may be shown by a curve like that of Fig. 34-7. Then the range of frequencies within the points at which amplification falls off to about 70 per cent may be called the passband of the amplifier or the stage, or it may be called the bandpass.

I-F TRANSFORMERS. It is impractical to construct either a single-coil coupler or a two-winding transformer to provide satisfactorily high gain, good selectivity, and at the same time allow a passband so wide as required for television. A single television channel covers a range of frequencies about six times that of the entire standard broadcast band, in which there are more than 100 channels. To adequately amplify all frequencies in a television channel and provide sharp enough cutoff at both ends it is necessary to use several i-f amplifier couplings tuned to different frequencies, as shown by Fig. 34-8. The overall frequency response or overall amplification will be about as shown at the bottom of the figure. By suitable choice of the separate frequencies the overall response may be made as wide as required. This is called staggered tuning.

The passband for f-m sound broadcast reception is only 2½ per cent of that for television. A two-winding transformer may be constructed and adjusted to cover this narrower range without difficulty. For a-m stan-

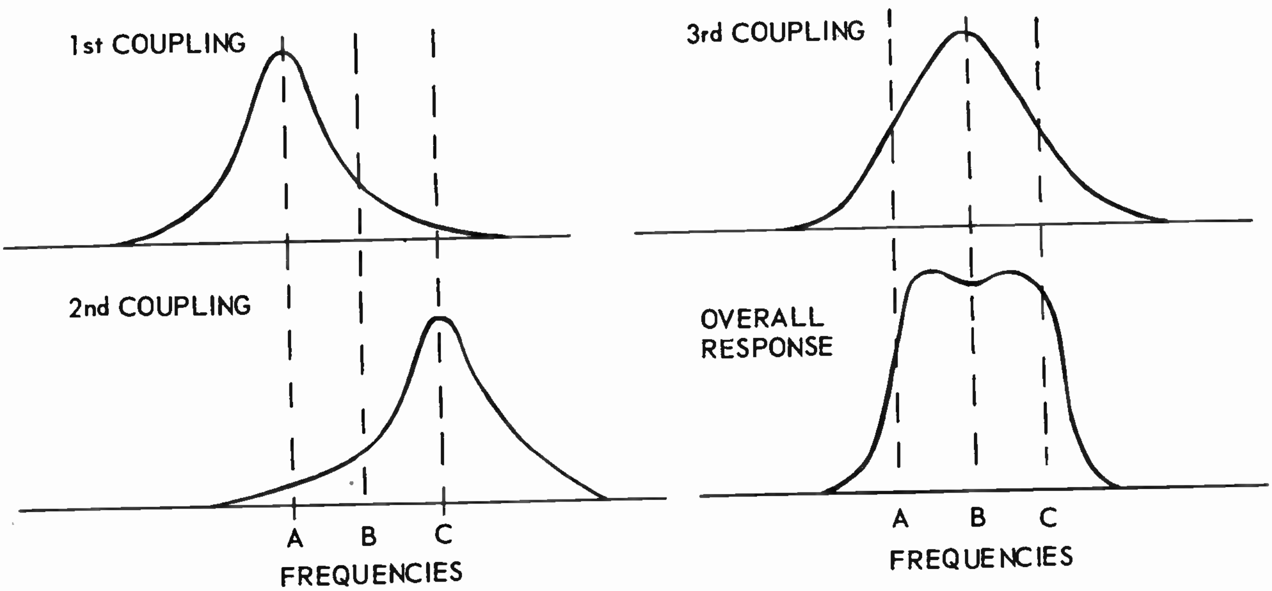


Fig. 34-8. Stagger tuned stages provide a wide overall frequency response with sharp cutoffs.

dard broadcast the passband need be only 1/6 of one per cent of that for television, so it is a simple matter to cover this very narrow range with a two-winding transformer.

The basic design of i-f transformers for f-m receivers is like that for a-m standard broadcast receivers, except for changes required in handling the higher intermediate frequency. The same basic design is used for sound i-f transformers of television receivers. All this makes it logical to examine the i-f transformers for standard broadcast, F-M broadcast, and television sound all at the same time. Video i-f amplifying systems for television will be taken up a little later.

2 I-f transformers having primary and secondary windings separately tuned by adjustable capacitors or by adjustable iron cores are pictured in Figs. 34-1, 34-3, and 34-5, and are shown by symbols at *B* in Fig. 34-4. If the two windings of a transformer were of identical construction and Q-factor, had very loose coupling, and were tuned to the same frequency, the response curve would have a single rather sharp peak. Because of the loose coupling the voltage and current induced in the secondary would not be very strong; there would not be an efficient transfer of power through the transformer.

As the coupling between windings is made closer and closer there will be an increase of power transfer, while the frequency response will continue to show a single peak. Finally you would reach a degree of coupling at which there is maximum power transfer, and maximum induced voltage and current in the secondary. With either greater or less coupling the secondary voltage and current would drop lower. This is the condition called critical coupling.

With increase of coupling beyond the critical value the single peak of response will separate into two peaks, one at a frequency lower than the original resonant frequency and the other above the original frequency. Changes of frequency response will be somewhat as shown by Fig. 34-9. At 1 we have less than critical coupling. At 2 there is critical coupling. At 3 we have "overcoupling" to a degree slightly greater than the critical value. At 4 the coupling has been greatly increased, causing two distinct peaks with relatively weak response at the original resonant frequency, which now lies between the two peak frequencies.

The two peaks appear because the mutual inductance alternately adds to and subtracts from the inductances of the two windings, producing the effect of two different inductances and the corresponding two peaks of response.

We have assumed identical primary and secondary circuits with equal Q-factors. If the two circuits are not alike, but have different Q-factors, and if there is overcoupling. the voltages of the separate peaks will be less than the voltage for a single peak with critical coupling. The greater the Q-factor of either circuit the sharper will be its peak. High Q-factors in both circuits make both peaks sharper and increase the depth of the "valley" in between them.

There is a great deal of interesting theory dealing with the effects of over-coupling, but to follow the theory we would have to know the coupling coefficients and how they vary, and would have to know the Q-factors of the two circuits. These things would not be known during ordinary service work, and itwould not be worth our while to go into them because few i-f transformers are constructed with adjustable couplings.

Nearly all i-f transformers are designed with fixed coupling. There are adjustments that allow you to alter the frequency at which the peak or peaks appear, but during service operations you cannot materially change the shape of the response curve other than to distort it beyond all usefulness. The shape of the frequency response, when there are correct adjustments, depends on the original design.

Fig. 34-10 shows some of the changes in frequency response which might be seen by a service technician during adjustment of a double-tuned overcoupled i-f transformer. These curves were taken on an i-f transformer in an f-m receiver employing an intermediate frequency of 10.7 mc, the intermediate frequency used in all modern f-m sets.

In diagram 1 both windings are far out of adjustment. There is a low peak at 10.5 mc, a higher peak at 10.83 mc, and between them is a fairly flat response. Adjustment of one or the other of the windings will bring the peaks closer together, as at 2. Further adjustment brings the peaks still closer, as at 3, and there is an increase of height on the lower peak because of greater power transfer. Now it becomes possible, by adjusting either winding, to produce the effect shown at 4. Here there is a high, sharp peak at the left and a lower one at the right. The high peak might be at the right instead of at the left.

Final adjustment will produce the response shown at 5 in Fig. 34-10. The center is at 10.7 mc, with

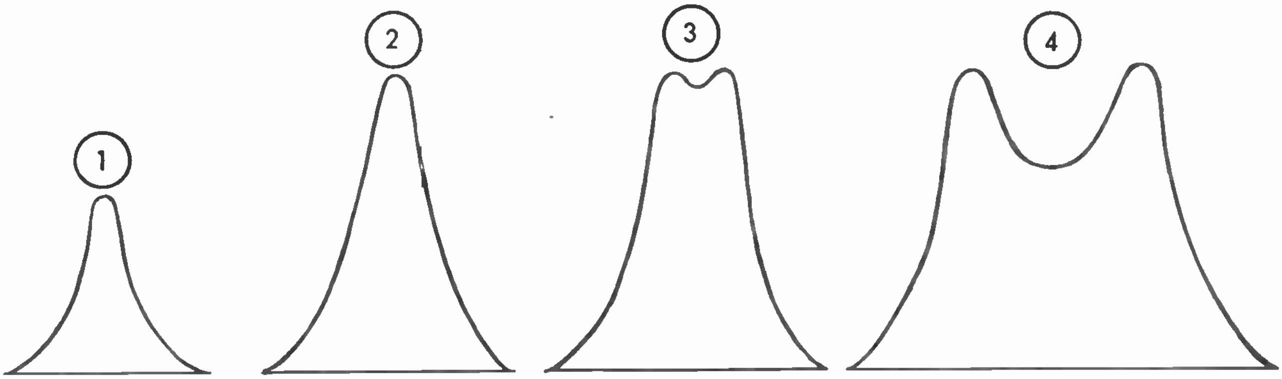


Fig. 34-9. How frequency response is altered by increasing the coupling.

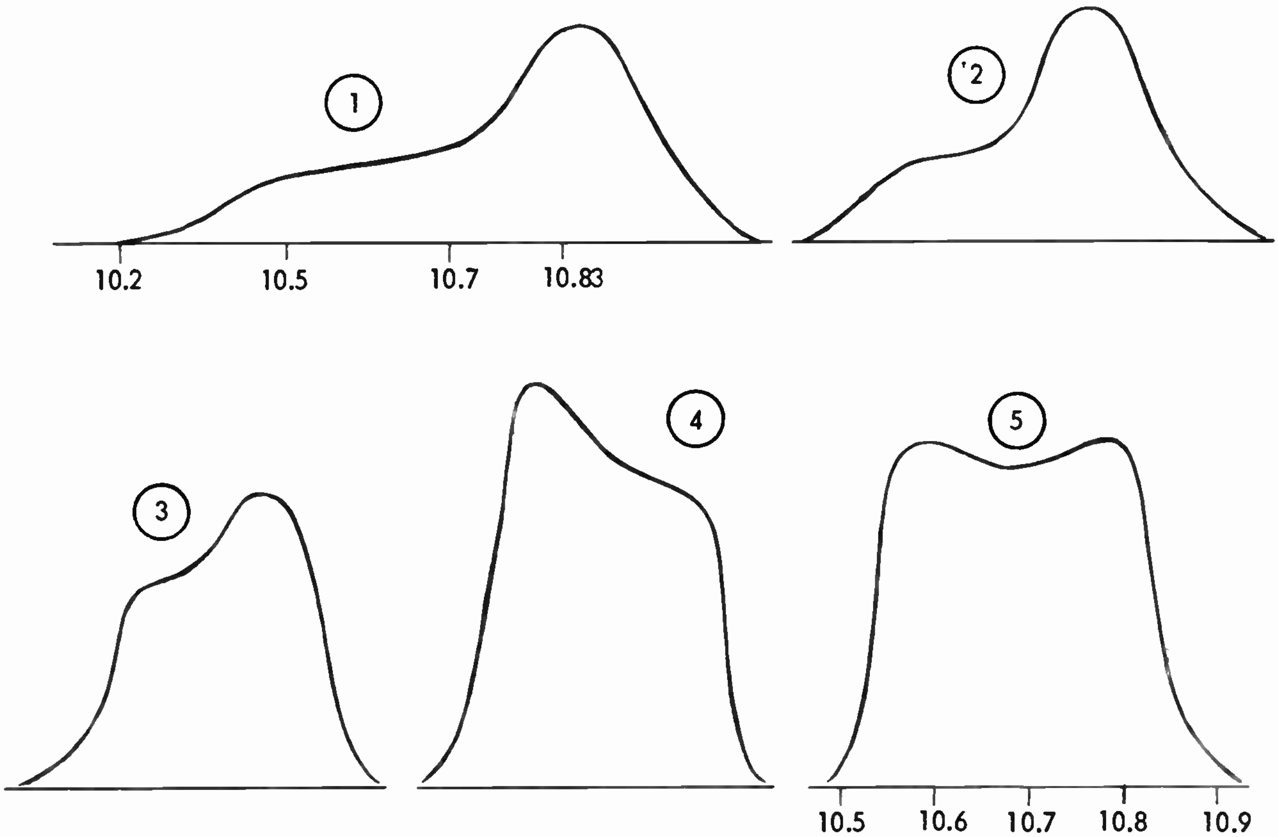


Fig. 34-10. Frequency responses obtained during tuning or "alignment" of an i-f transformer in an f-m receiver.

peaks 0.1 mc lower and 0.1 mc higher. There is almost complete cutoff at frequencies 0.2 mc lower and 0.2 mc higher than the center frequency. The effective passband is about 0.3 or 0.3 mc. The valley is very little lower than the peaks.

In examining this i-f response you should keep in mind that maximum frequency modulation or frequency change (for an f-m sound receiver) will extend to 75 kc or to 0.075 mc above and below the center frequency. That is, the total change of frequency will be twice 0.075 mc or will be 0.15 mc. Since our passband is 0.30 mc there is sure to be practically uniform amplification of all modulation frequencies within the 0.15 mc range of reception.

With an i-f transformer designed for less than critical coupling, or with one in which overcoupling cannot be had, it still is possible to have two peaks or a single fairly broad peak by tuning the primary and secondary to slightly different frequencies. Such i-f transformers are used in many f-m and a-m sound receivers, and in some television video i-f amplifiers. This method of coupling may be shown as at the left in Fig. 34-11. Primary and secondary are tuned by adjustable cores. There is neither a fixed nor an adjustable capacitor across either winding. But across the primary is the output capacitance of the first tube, and across the secondary is the input capacitance of the second tube. These tube capacitances are indicated by broken-line symbols.

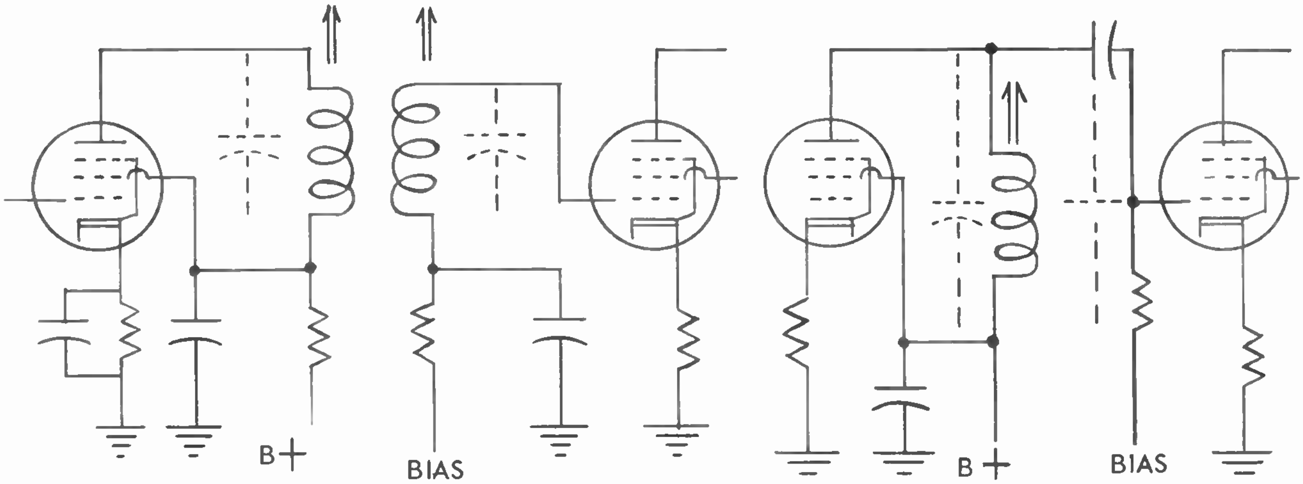


Fig. 34-11. Interstage coupling with a double-tuned transformer (left) and with a single tuned coil (right)

The diagram at the right in Fig. 34-11 shows the coupler between two amplifying tubes as a single tuned coil rather than a two-winding transformer. Now both tube capacitances are across the single coil. Since the capacitances are in parallel with each other there is more capacitance across the coil than across either of the windings in the transformer at the left.

One of the advantages of a double-tuned transformer as compared with a single coupling coil is that each tube capacitance and also the circuit capacitance act with only their own winding, and the tuning of each winding may be exactly suited to the capacitances with which it must operate. A double-tuned transformer is capable of producing a frequency response with a flatter top and steeper sides than obtainable with a single coil.

The frequency response of an i-f transformer does not have to have such steep sides or such sharp cut-offs as indicated in Fig. 34-10 by diagram 5. The double-peaked response at the left in Fig. 34-12 would be entirely satisfactory, as would also the broad single peak at the right – especially when there is more than one amplifying stage. When one stage feeds into another the effect is to produce an overall response with sides much steeper than in the response of any one of the stages by itself. The responses of Fig. 34-12 are photographs of oscilloscope traces.

Some i-f transformers have their primary and secondary windings not as separate coils close together but with the turns of each winding wound right along with those of the other winding. Such a transformer is made up by laying the wires for the two windings parallel and right against each other, then winding both together around the form which is to support them. Such windings sometimes are called bifilar, a name originally applied to two threads running side by side on the body of a machine screw or bolt.

These bifilar windings may be indicated by symbols such as illustrated in Fig. 34-13. Any symbol showing two coils with only one adjustable core for both indicates a winding of this general type. Moving the core farther into the windings increases the inductance and lowers the tuned frequency, while turning the core out of the windings lessens the inductance and raises the frequency of resonance. Although changing the position of the core makes large variations of resonant frequency, it has very little effect on the coupling.

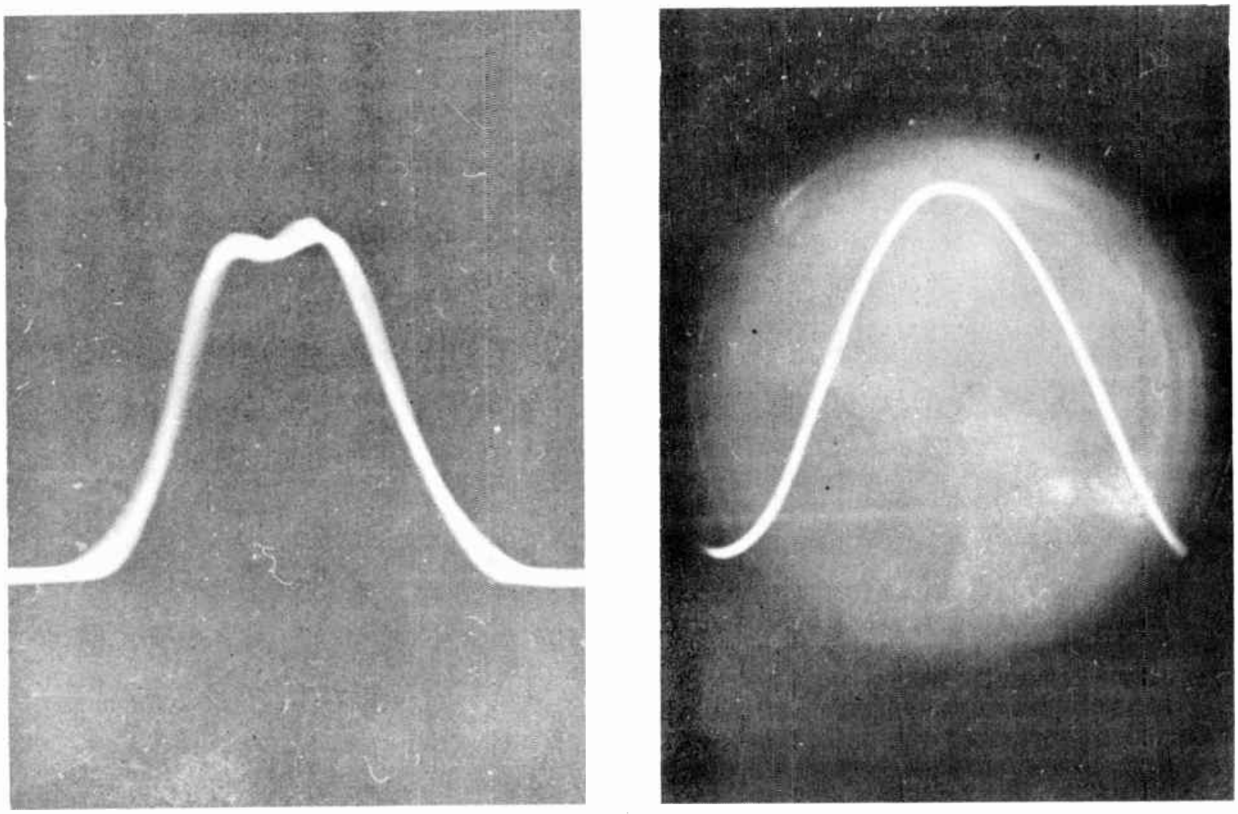


Fig. 34-12. Frequency responses with double and single peaks as observed on an oscilloscope.

This is because the core does not form a complete magnetic circuit and changing its position cannot alter the proportions of iron and air in the magnetic circuit nor make any material change of permeability.

When there is only a single adjustable core, as at the right in Fig. 34-11 and in Fig. 34-13, the frequency at which there is peak response is varied by moving this core farther into or out of the winding or windings. With two adjustable cores, as at the left in Fig. 34-11, both cores must be moved to change the peak frequency while maintaining maximum gain or maximum height of the response curve. If the frequency is to be lowered, both cores must be turned farther into their windings to increase the inductance. To raise the frequency both cores must be turned farther out of their windings to lessen the inductance. If there are two adjustable capacitors, as at *B* in Fig. 34-4, both capacitances must be increased to lower the peak frequency, or both must be decreased to raise the peak frequency.

When adjustment is by means of a movable core or slug, there will be maximum inductance and lowest tuned frequency with the core centered along the length of its winding or windings. Movement either way from this center position decreases the inductance and raises the peak frequency. As a consequence, you sometimes may screw a core one direction to make the peak frequency lower only to find that continued movement in the same direction finally brings the frequency up again. This means that you have turned the adjustment past the center point.

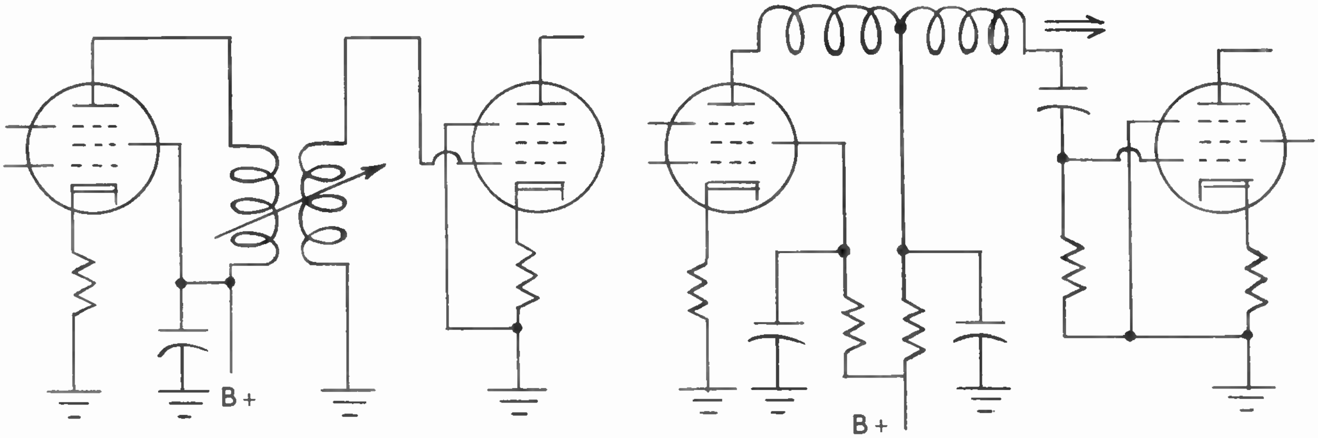


Fig. 34-13. Transformers with bifilar windings may be shown by symbols of these types.

INDUCTANCE TUNING. You have noticed that some circuits are tuned to resonance by adjustable inductance, with fixed capacitance, while others are tuned by adjustable capacitance with fixed inductance. Inductance is adjusted by means of a movable iron core inside a coil winding. If this inductance is to be varied only during service operations, as when tuning the i-f transformers, the adjustment ordinarily is altered by turning a threaded stud or screw to which the core is attached. If the inductance is to be varied by the set operator, for tuning to various stations or channels, the change is made by means of a knob or dial with some form of pulley and cord arrangement for moving the core.

When variable capacitance is to be used by the set operator for tuning to stations or channels, the capacitor is an air dielectric type with movable rotor plates and fixed stators. Small air dielectric capacitors may also be used for service adjustments, with the end of their rotor shaft slotted to take a screw driver, or of square or hexagon shape to take a socket wrench. A capacitor more commonly used for service adjustments is the mica compression type having a screw for bringing the plates closer together or letting them spring apart. We have looked at a number of these mica compression capacitors used as trimmers. Two such trimmers are on each of the i-f transformers pictured in Figs. 34-3 and 34-5.

Adjustable capacitors used as trimmers or for any service adjustments sometimes are of the rotary ceramic type, with thin metallic plates formed on the surfaces of ceramic discs. There also are tubular ceramic trimmer capacitors in which the dielectric is a thin ceramic cylinder. The fixed plate is a metallic coating or a metal tube or a winding of wire around the outside of the dielectric cylinder. The other plate is either a solid cylinder or else a tube of metal inside the dielectric cylinder, arranged to be moved lengthwise by a threaded stud or screw. Capacitance is increased by turning the inside adjustable plate farther into the outer fixed plate.

6 By using a movable iron core it is possible to obtain a necessary inductance with a much smaller coil winding than when using air-core construction, this is because the iron greatly increases the permeability of the magnetic circuit and the inductance of the winding. The smaller winding will have less distributed capacitance and less high-frequency resistance, thus tending to increase the Q-factor. But there are losses due to eddy currents and hysteresis in the iron core. These losses would not be present with air-core construction. The net result is that there is little difference between Q-factors and resonant gains of circuits tuned by adjustable inductance and by adjustable capacitance. The chief advantage of iron-core construct-

COMMERCIAL TRADES INSTITUTE

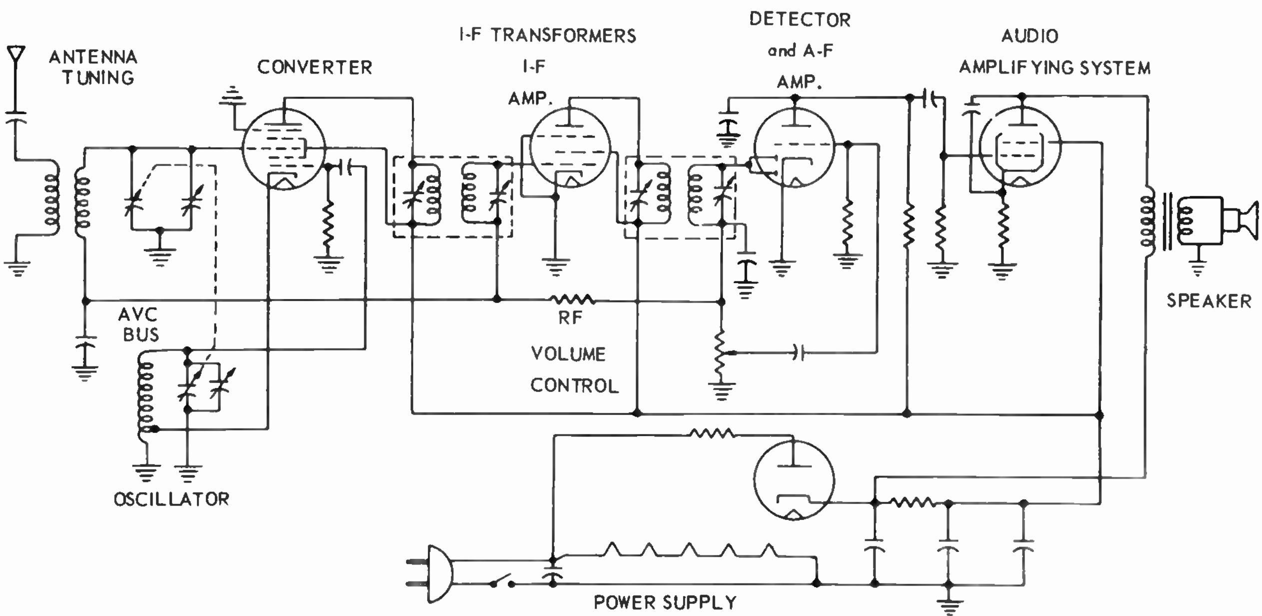


Fig. 34-14. The superheterodyne circuit with which we started our investigations of this type of receiver.

ion is reduction of the space needed, because the coil winding may be smaller and because the movable core is inside the coil and adds nothing to the overall size.

COMPLETING THE SUPERHETERODYNE. When first commencing the study of superheterodyne receivers we looked at the circuit diagram for a five-tube set and saw many parts which then were wholly unfamiliar. That diagram is reproduced in Fig. 34-14. Now only a few parts are unfamiliar, and these can be disposed of in short order because you are well versed in principles that explain their operation.

Up to this point we have had to deal chiefly with fundamentals, to make sure of having a firm foundation for future practical applications. Now we are almost ready to undertake some of the most important service operations. Of course, there are many principles still to be learned, but once we get started in actual servicing procedures it will be easy to grasp all the remaining basic facts. You will learn these factors just as any beginning service technician learns new things every day he works. He learns by running into problems that demand additional knowledge, by acquiring that knowledge, and immediately putting it to use in solving the problems.

To finish what little preliminary work remains we shall examine our receiver diagram to discover which parts or circuits require additional explanation. Over at the left-hand side of the diagram we have a converter tube, also a tuned circuit between antenna and r-f grid, and we have a Hartley oscillator circuit connected to the oscillator grid.

Next comes an i-f transformer, then the i-f amplifier tube, and another i-f transformer. These parts have been discussed in the present lesson. From the second i-f transformer the signals go to the diode detector which is part of the duodiode-triode tube. We know how the diode detector performs. Down below is the B-power supply and the series heater connections. We are well acquainted with power supplies.

There remain only two parts to be examined; the audio amplifying system between detector and speaker, and the connection from the grid returns of the converter and i-f amplifier to the top of the volume control resistor. This latter connection provides automatic volume control which compensates for variations in strength of received signals.

The circuits for automatic and manual volume control are shown by themselves in Fig. 34-15. As we learned when studying diode detectors, the demodulated signal voltage in the detector circuit appears across the volume control resistor. Part of this signal voltage is taken through the slider contact and capacitor *Cg* to the grid of the triode audio-amplifying section of the combined a-f amplifier and detector. The farther the slider is moved away from the grounded lower end of the volume control resistor the stronger is the signal voltage taken to the grid of the a-f amplifier, and the greater will be the sound volume from the speaker. Moving the slider toward the grounded end of the control resistor reduces the volume. This is the action of the manual or hand-operated volume control.

Now to explain the action of the automatic volume control, a name usually abbreviated to "avc" or "AVC". First look at the grid return circuits shown by figure 34-14 of the converter and i-f amplifier. From the converter grid the return circuit goes through the transformer winding to the line marked AVC Bus, thence through resistor *Rf* and the volume control resistor to ground. This grid circuit is completed through ground to the lower end of the oscillator coil and through part of this coil back to the converter cathode.

From the grid of the i-f amplifier the return circuit goes through the transformer winding to the avc bus, through resistor *Rf*, the volume control resistor, and through ground back to the cathode of the i-f amplifier.

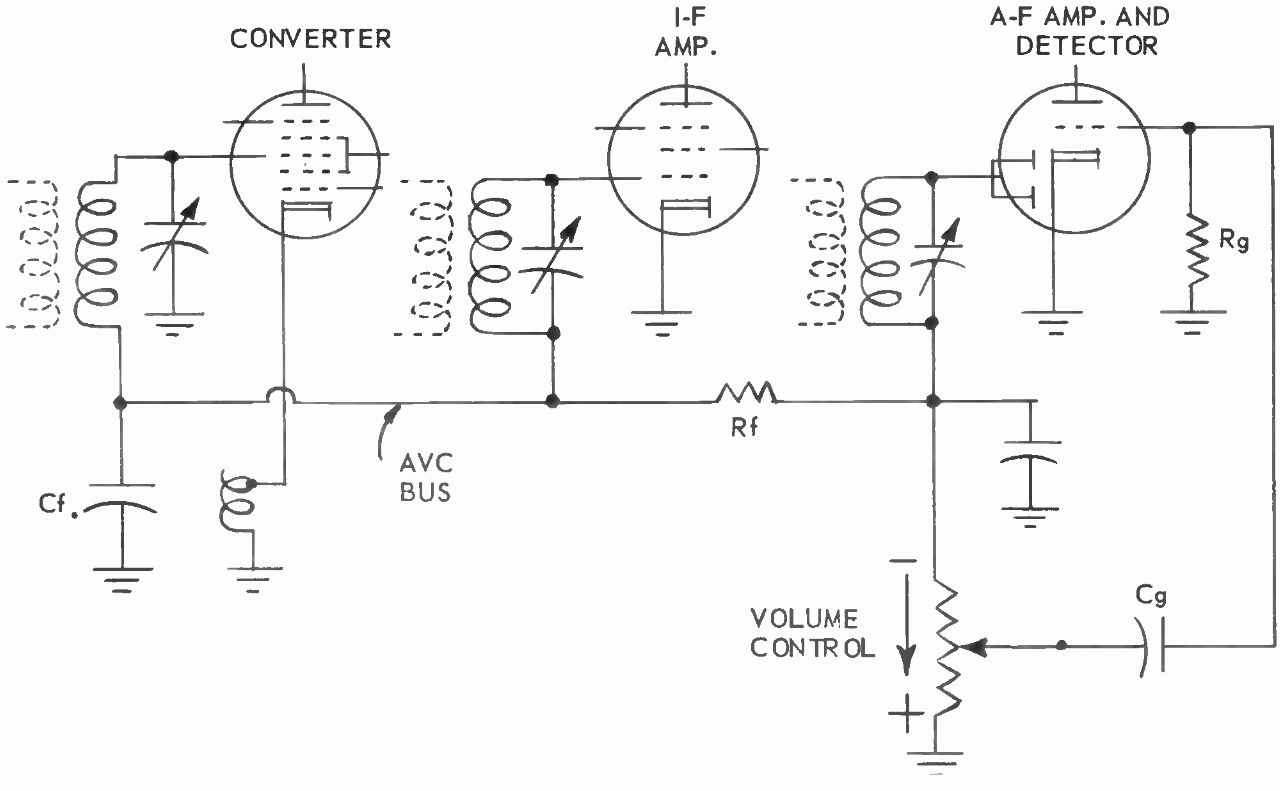


Fig. 34-15. Manual and automatic volume control circuits of the superheterodyne.

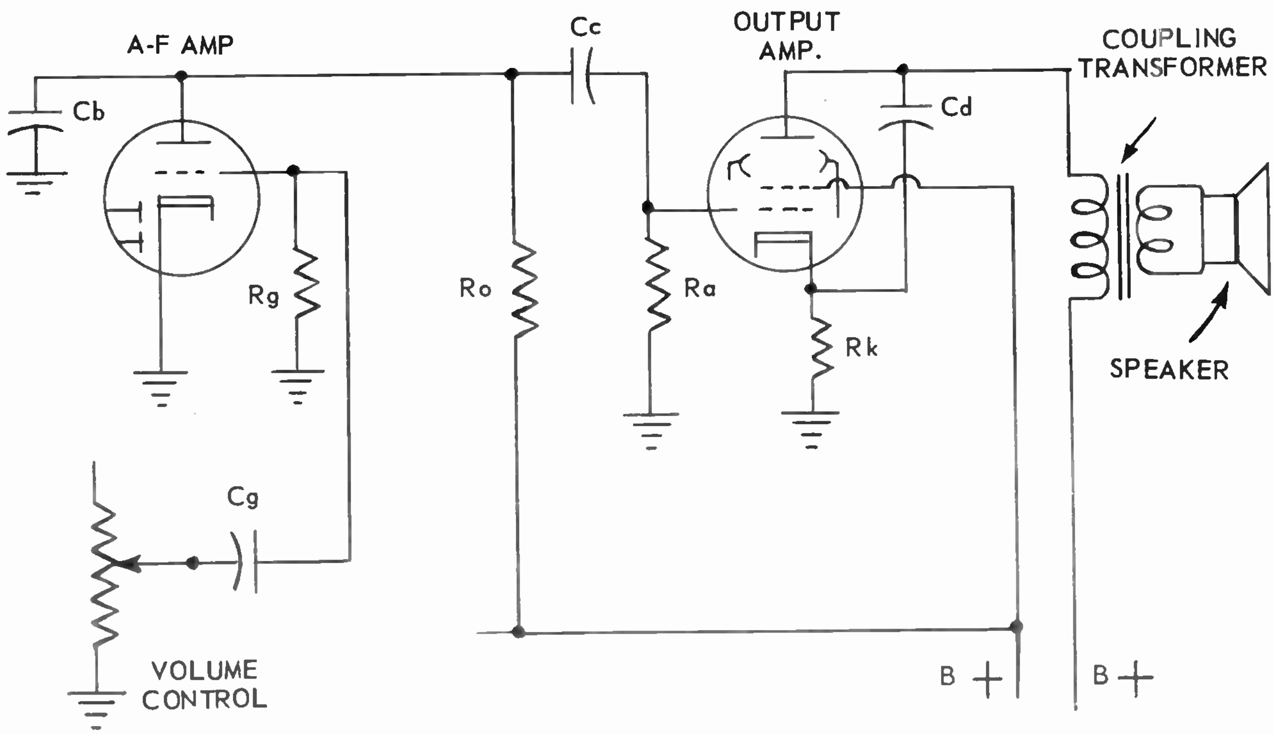


Fig. 34-16. The audio amplifying system of the superheterodyne receiver.

The grids of both tubes are connected to the top of the volume control resistor. The cathodes of both tubes are connected through ground to the bottom of this resistor. Therefore, the voltage or potential difference across the volume control resistor must be the grid bias voltage for the two tubes, because this voltage is the difference between potentials of the grids and the cathodes.

Electron flow in the volume control resistor comes from the diode detector plates through the transformer winding and passes from top to bottom of the resistor. The electron flow continues on through ground and back to the cathode of the detector tube. This direction of electron flow means that the top of the volume control resistor is negative with reference to the bottom. Consequently, the grids of the converter and i-f amplifier (connected to the top of the resistor) are negative with reference to their cathodes (connected to the bottom of the resistor) and there is negative grid bias on both tubes.

1 The grid bias voltage is varied by changes of strength in the antenna signals which are amplified by the converter and i-f amplifier. A stronger signal delivered from the converter and i-f amplifier to the detector will increase the average electron flow through the volume control resistor. The greater electron flow means increased voltage across this resistor. The increased voltage makes the bias for the tubes more negative. As the bias becomes more negative it reduces the amplification, thus counteracting the stronger received signal. If the incoming signal gets weaker there is less voltage across the volume control resistor, and bias on the converter and i-f amplifier tubes becomes less negative. The less negative bias allows amplification to increase – again counteracting the effect of the weaker signal.

6 We have looked at the simplest of automatic volume controls. A stronger incoming signal acts to decrease

the amplification, while a weaker incoming signal acts to increase the amplification. Variations of strength in received signals are largely compensated for, and sound volume from the speaker depends chiefly on the setting of the manual volume control.

Next we shall look at the audio amplifying section, shown by itself in Fig. 34-16. The a-f amplifier is the triode portion of the combined detector-amplifier tube. The grid is biased by the grid-leak method, with grid resistor *Rg* and capacitor *Cg*. Capacitor *Cb* bypasses to ground any intermediate frequencies which might get into the audio amplifier system. Resistor *Ro* is the load resistor in the plate circuit of the a-f amplifier. Amplified audio signal voltages developed across *Ro* go through coupling capacitor *Cc* to the grid of the output amplifier. Long ago we learned how amplified signal voltages are produced across a load resistor in a plate circuit.

The output amplifier is a beam power tube. It is operated with cathode bias by means of bias resistor *Rk*. The grid return is through resistor *Ra* to ground and the cathode. Amplified signal output from this tube is coupled to the speaker through a transformer. The transformer primary provides the load impedance in the plate circuit of the tube. A small amount of signal voltage is fed from the plate to the cathode of the output amplifier through capacitor *Cd*. This is a degenerative feedback for lessening distortion of reproduced sound, something to be examined at length at a later date.

Finally we have arrived at the speaker of a complete superheterodyne receiver after following the signals all the way from the antenna. Principles which have been studied in many lessons here have come together and found their applications in the superheterodyne system which is the basis of all a-m receivers, all f-m receivers, most of the short-wave receivers, and all television receivers.

To make this a-m receiver over into an f-m type would require only two major changes; the i-f amplifier section would be designed to operate at higher frequencies, and we would use a different kind of detector. Then, to adapt the f-m receiver for use as the sound section in a television set we could take off the f-m tuner, connect the i-f amplifier to the regular tuner of the television receiver, and design the i-f amplifier for still higher frequencies.

To use our a-m superheterodyne for the r-f and video sections of a television receiver we would replace the converter with separate mixer and oscillator, add more i-f stages, and connect the output of the diode detector through a video amplifier to the picture tube instead of through the audio amplifier to the speaker. Always we would retain the basic superheterodyne principle.

TELEVISION

LESSON NO. 35

ALIGNING THE SUPERHETERODYNE

Training Your Mind...

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

DO NOT TEAR — CUT ALONG THIS LINE AND SEND IN FOR GRADING.

COMMERCIAL TRADES INSTITUTE

Training Your Mind... C.T.I. Perfecting Your Skill

Chicago, Ill.

LESSON NO. 35

ALIGNING THE SUPERHETERODYNE

Many a service man has begun his professional career with less knowledge of radio than you already possess. Let's imagine that you are on your own, and have the job of aligning the six tube-two-band superheterodyne of Fig. 35-1. To align a receiver means to make all necessary tuning adjustments in circuits for antenna and r-f amplifier, the r-f oscillator, and the intermediate-frequency amplifier.

The minimum equipment for alignment of a standard broadcast set is a narrow bladed screw driver with an insulated shaft, and a thorough knowledge of what you are trying to accomplish and how it should be done. A trained technician can do more of a job with a screw driver and his ears, to hear with, than can be done by someone with less mental equipment and a bench full of signal generators, electronic voltmeters, output meters, oscilloscopes, and everything else that he doesn't know how to use. Don't conclude, however, that good equipment is not worth while. With suitable instruments and tools the trained man can do even a better job and do it in far less time.

Quite likely you will have available the manufacturer's service information, which will be of the general order shown under the heading of "Service Information" in this lesson. The instructions tell how the manufacturer believes the job should be done to insure best results with the testing equipment mentioned. As a general rule it is assumed that you have a signal generator and some kind of meter for measuring signal output from the set. After a few pages of preliminaries we shall discuss each of the alignment steps.

Accompanying the service information will be a circuit diagram. Fig. 35-2 is such a diagram for the particular receiver we intend to align. This diagram gives the values of all fixed capacitors and fixed resistors. In accordance with usual practice the capacitances are in micro-microfarads for all values less than 1,000 micro-microfarads, and are in decimal fractions of a microfarad for all greater values. Resistor values followed by the letter M are in megohms, those followed by the letter K are in thousands of ohms, and with no letter the values are in ohms.

A competent service technician would recognize all the parts and their purposes from the symbols, the relative positions, and circuit connections. For convenience in explanation there has been added to the circuit diagram a series of reference letters from A to N. These letters would not appear on a regular service diagram. Information to be gained from each reference point is as follows.

A. This is part of the band switch for selecting either standard broadcast or short-wave reception. The switch rotors are shown in the broadcast positions.

B. The antenna tuning transformer. Short-wave windings are above and the standard broadcast windings below. Trimmer capacitors are across each secondary. In this multi-band set the trimmers must be across the coils, not on the tuning capacitor sections, because the same tuning capacitors are used for both bands and each band requires separate alignment. A 40K resistor is across the broadcast primary to broaden the tuning by increasing high-frequency resistance. The lower ends of the primaries connect to chassis ground through an.01 mf capacitor, to insulate the antenna from the "hot" chassis.

SERVICE INFORMATION

Tuning Ranges: Broadcast, 540 to 1,600 kc. Short-wave, 6 to 16 mc.
Intermediate Freq: 455 kc.
Power Supply: D-c or 50-60 cycle a-c. 115 to 120 volts.

ALIGNMENT

Connect receiver to 110V line through isolating transformer if possible, otherwise connect 0.1 mf between low side of generator and chassis.

Output indicator: High-resistance voltmeter in series with 300K resistor, or electronic voltmeter, between high side of volume control and chassis. Meter negative to volume control, positive to chassis (volume control not marked on schematic)

SIGNAL GENERATOR INPUT	SETTING	RECEIVER DIAL	ADJUSTMENT
	I-f Transformers		
1. Converter pin 8, r-f grid, through 1000 mmf. Note "a"	455 kc	For no signal	Secondary and primary of 2nd i-f transformer for maximum
2. Same as step 1.	455 kc	For no signal	Secondary and primary of 1st i-f transformer for maximum.
	Wave Trap		
3. Antenna terminal through 1000 mmf.	455 kc	600 kc	Wave trap for minimum. Increase generator output.
	Broadcast R-f (Antenna)		
4. Antenna terminal through 200 mmf. Note "b".	1400 kc	1400 kc	Broadcast oscillator trimmer for maximum.
5. Same as step 4.	1400 kc	1400 kc	Broadcast antenna trimmer for maximum.
6. Same as step 4.	600 kc	600 kc	Padder capacitor for maximum. Note "c".
	Short-wave R-f		
7. Antenna terminal through 400-ohm carbon resistor	16 mc	16 mc	Short-wave oscillator trimmer for maximum. Tighten, then loosen to second peak.
8. Same as step 7.	16 mc	16 mc	Short-wave antenna trimmer for maximum.

Note a: Use signal generator unmodulated for all steps in alignment.

Note b: Band switch at broadcast position for steps 4 through 6, at short-wave position for steps 7 and 8.

Note c: Rock receiver tuning dial back and forth around 600 kc marking to find position allowing maximum meter reading, which may not be at exactly 600 on the dial.

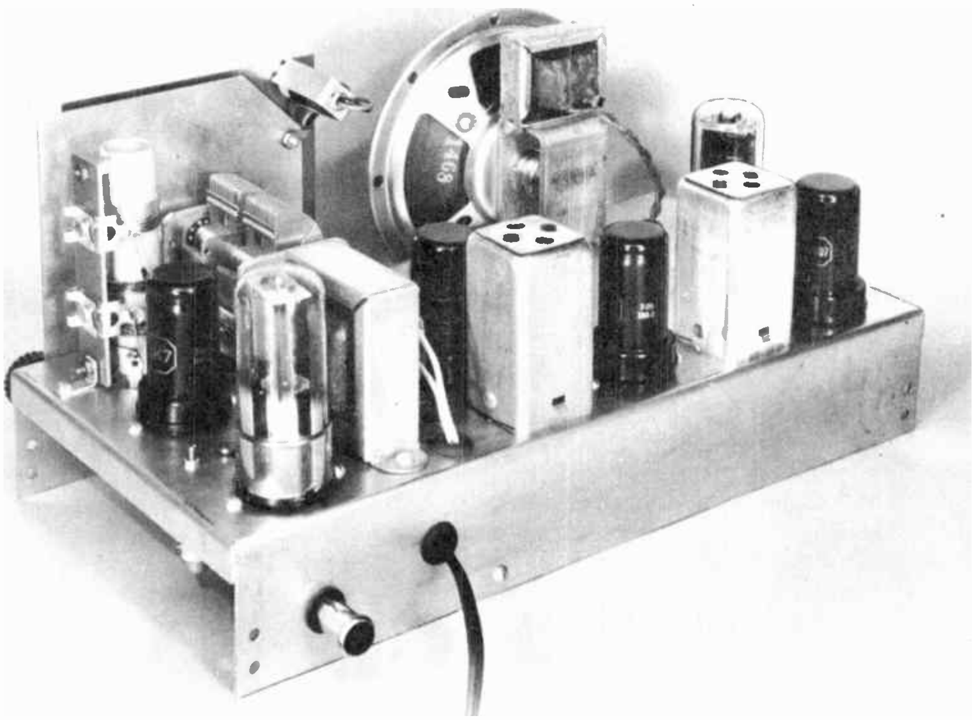


Fig. 35-1. The six-tube two-band superheterodyne receiver. The antenna tuning transformers and i-f transformers are visible.

C. This coil with adjustable core and the paralleled .005 mmf capacitor form a wave trap in series with the cathode of the r-f amplifier. The trap will be tuned to 455 kc, the intermediate frequency. Plate or cathode currents due to signals at this frequency which might come to the r-f tube meet the high impedance of the parallel resonant trap circuit and are greatly weakened instead of passing through the converter to the i-f amplifier. There are other radio services, including some code stations, operating at and near this frequency.

D. Cathode-bias resistor for the r-f and i-f amplifier tubes. Note that the cathodes of both these tubes connect through this 100-ohm biasing resistor to ground.

E. and F. The first and second i-f transformers. There are capacitor trimmers on each winding. In other sets there might be adjustable cores with screws reached from the top or bottom of the can, or with one winding adjusted from above and the other from below. Alignment procedure is exactly the same with inductance trimmers as with capacitance trimmers.

G. The diode detector is part of the duodiode-triode tube whose triode section is the a-f amplifier.

H. This is the avc bus connected to the top of the 500K volume control resistor. The grid return of the r-f amplifier connects to this avc bus through either of the secondaries in the antenna coupling transformer.

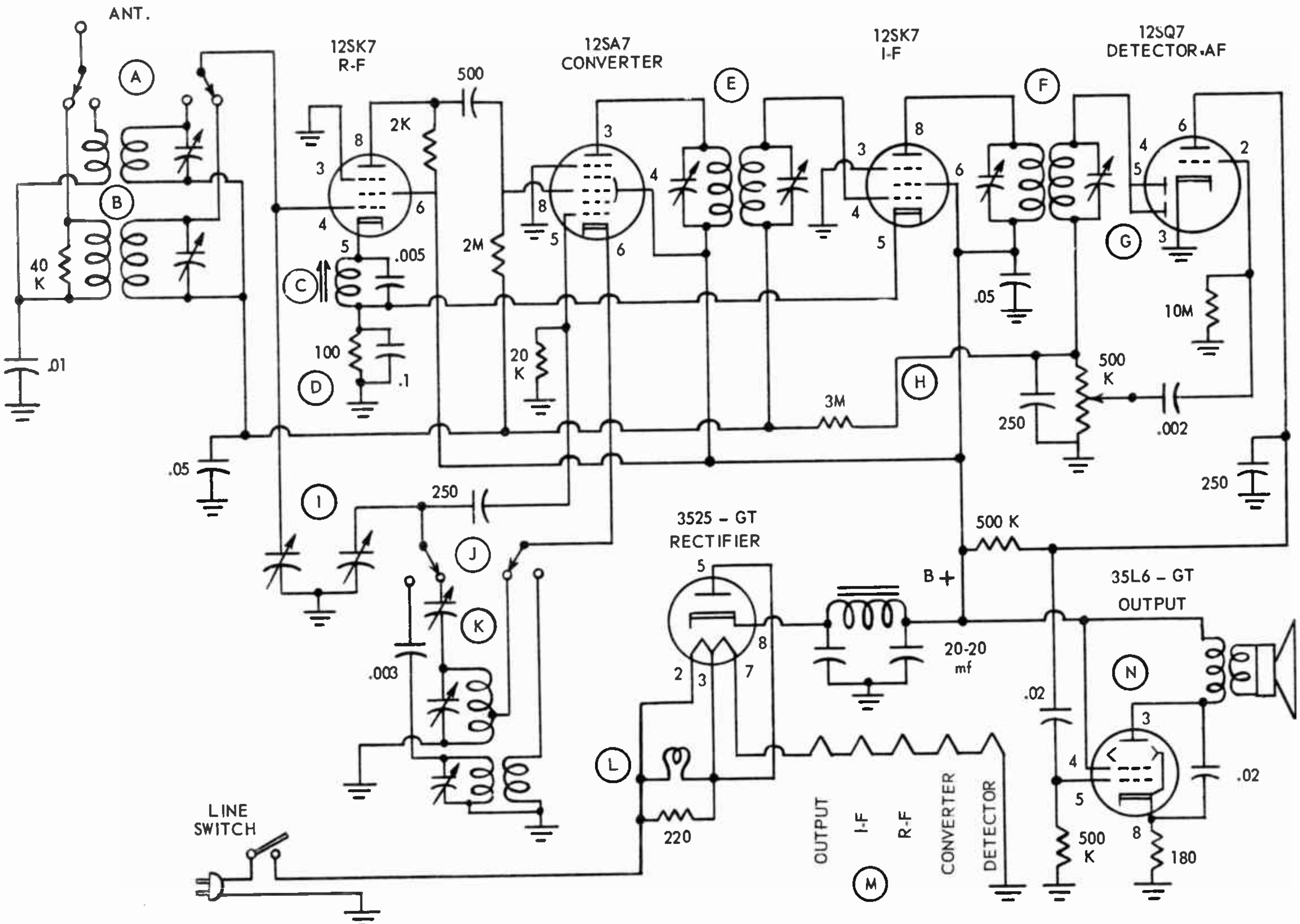


Fig. 35-2. The complete service diagram for the 6-tube superheterodyne.

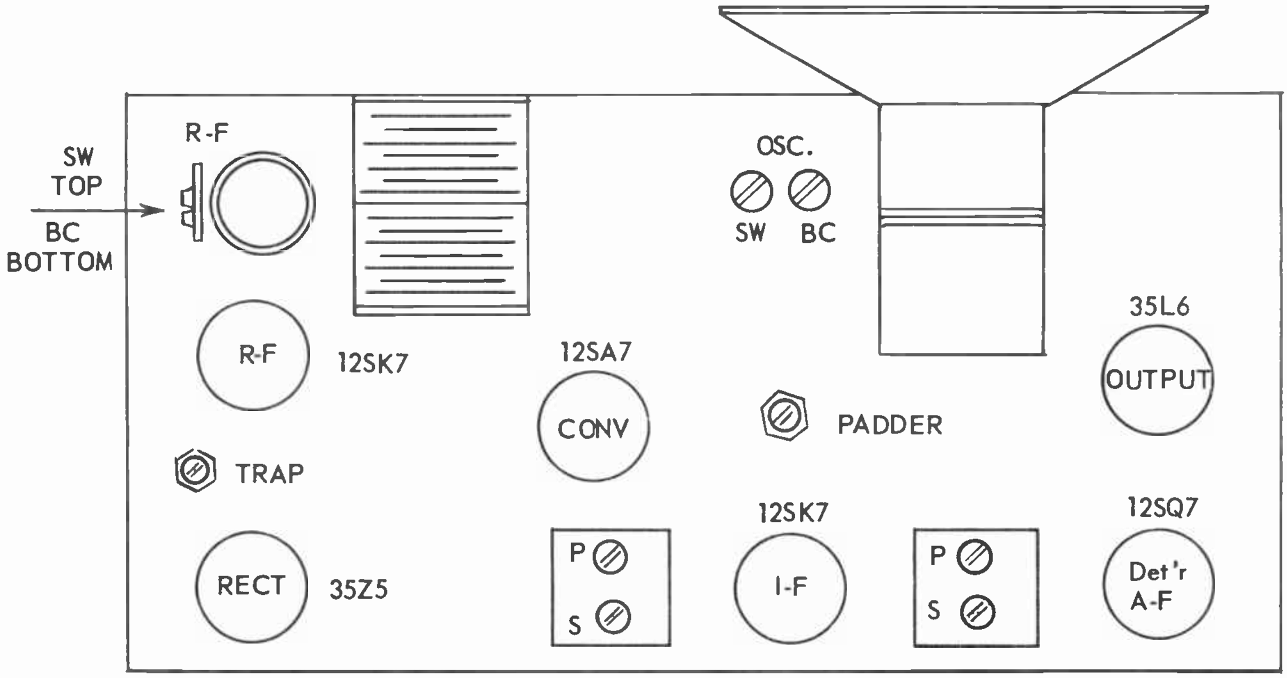


Fig. 35-3. Alignment adjustments reached from the top of the chassis.

The r-f or mixer grid of the converter connects to the avc bus through a 2M resistor. Thus the r-f amplifier and the mixer section of the converter have automatic volume control. We noted previously that the r-f and i-f amplifiers have cathode bias.

With this rather unusual avc and biasing system there is cathode bias and avc action on the i-f and r-f amplifiers and there is avc bias on the converter mixer section. A change of strength in the incoming signal will alter the avc voltage and cause a change of cathode current in the r-f amplifiers and in the biasing resistor at D. This resistor also biases the i-f amplifier.

When a change of avc voltage acts to decrease cathode current in the r-f amplifier this lessened current in the biasing resistor at D lessens the voltage across the resistor and allows the r-f amplifier grid to become even more negative with respect to cathode than it would be made by avc action alone. A given change of signal strength causes maximum avc action on the r-f amplifier, it has less effect on the i-f amplifier, and still less effect on the converter. In this way the r-f amplifier is given a great control over strong signals, which prevents overloading of the following converter and i-f amplifier.

You will come across many peculiar avc systems, especially in the r-f and video sections of television receivers – where they are called automatic gain control systems. Fortunately, these controls do not affect ordinary alignment procedures. At least, they do not alter the general methods by which alignment is carried out. Only when there are troubles with overloading and distortion will we have to consider these volume and gain controls.

I. This is the tandem variable tuning capacitor. The section at the left, in the circuit diagram, tunes the antenna transformer, and the right-hand section tunes the r-f oscillator.

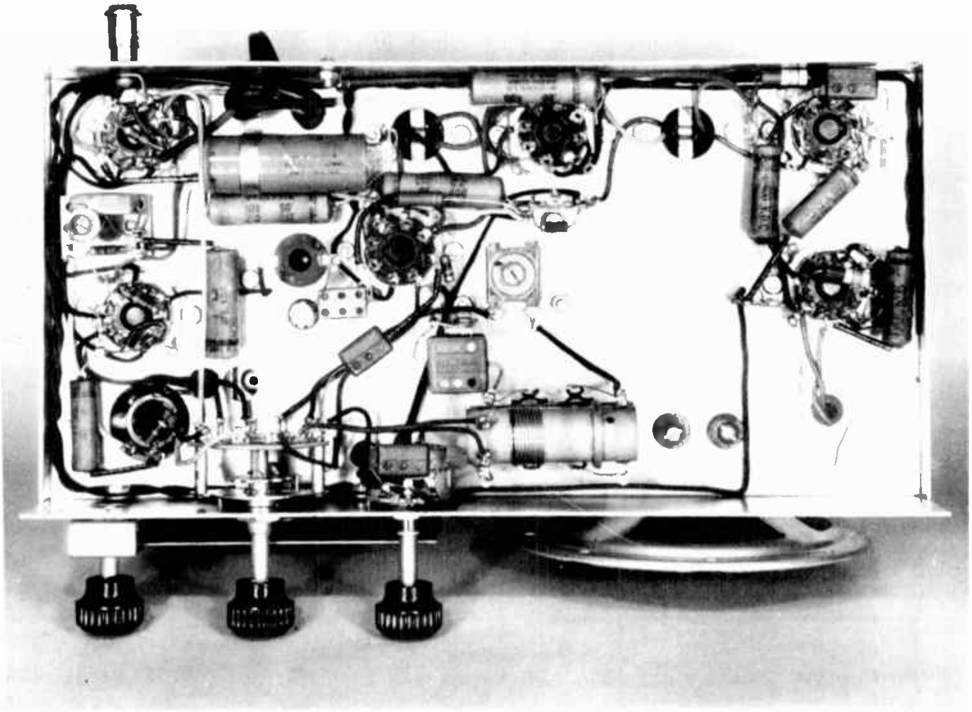


Fig. 35-4. The selector switch, tube sockets, most of the small parts, and wiring, are reached from underneath the chassis.

J. Another part of the band selector switch. This part connects the oscillator grid and cathode of the converter tube to the upper oscillator coil for broadcast reception or to the lower oscillator coils for short-wave reception.

K. This is the padder capacitor for alignment at the low-frequency end of the standard broadcast band. There is no adjustable padder for the short-wave band, but there is a fixed .003 mf capacitor in series with the short-wave oscillator coil. Note that there are trimmer capacitors across the oscillator coils for both broadcast and short-wave.

L. The pilot lamp, in parallel with a 220-ohm resistor and part of the heater in the rectifier tube.

M. The series heating string. Heaters for detector and converter, the tubes most sensitive to 60-cycle line-frequency hum, are connected at the grounded end of the string.

N. Note especially the connections on the output beam power tube and the a-f amplifier. Signals from the a-f plate come to the grid of the output tube through the .02 mf coupling capacitor. The a-f plate is connected to the B+ line through a 500K resistor. From the a-f plate to ground there is a 250-mmf capacitor for bypassing any high frequencies which would add to the load on the output tube even though these frequencies are above audibility. The .02 mf capacitor between plate and cathode of the output tube provides degenerative feedback for improvement of sound quality.

In addition to the circuit diagram there might be provided a diagram like that of Fig. 35-3 which shows the location of service adjustments as seen when looking down on top of the chassis. This type of diagram is a convenience but is not essential because you could locate all the adjustments by looking at the parts on top of the chassis and following connections underneath.

Underneath the chassis are the fixed resistors and capacitors, the band switch, volume control, oscillator coils, tube socket lugs, and most of the wiring, all as pictured by Fig. 35-4.

The actual layout of wiring and parts bears little resemblance to the connections as shown on the circuit diagram, but there are certain "landmarks" which make it easy to check the actual connections against those on the diagram. These landmarks are the tube socket pins and the terminals of some of the more easily recognized parts. By looking at the circuit diagram you can tell where the leads must go from each of the numbered pins on the tubes and the similarly numbered lugs on the sockets. Small capacitors and resistors of certain values may be identified by their color coding, while the larger paper and electrolytic capacitors are plainly marked with words and numerals. Terminals on parts such as padders and traps are easily identified. By watching the band switch while turning the rotor shaft you can tell which terminals are used for each frequency band.

ALIGNMENT METHODS. Having completed our examination of the receiver we may get ready to follow the instructions for alignment. Naturally, the exact methods for alignment will vary somewhat with different receivers. If detailed instructions are at hand they should be followed in order to save time and insure satisfactory results without loss of time. However, afterdoing a few actual jobs, you will be well able to make suitable adjustments without any printed instructions, and you even can get along without a service diagram by taking enough time to trace connections in the actual wiring.

The first step is to check indications of the tuning dial against positions of tuning capacitor rotors or tuning inductor cores. Some dials have special markings at which rotors or cores should reach the limits of

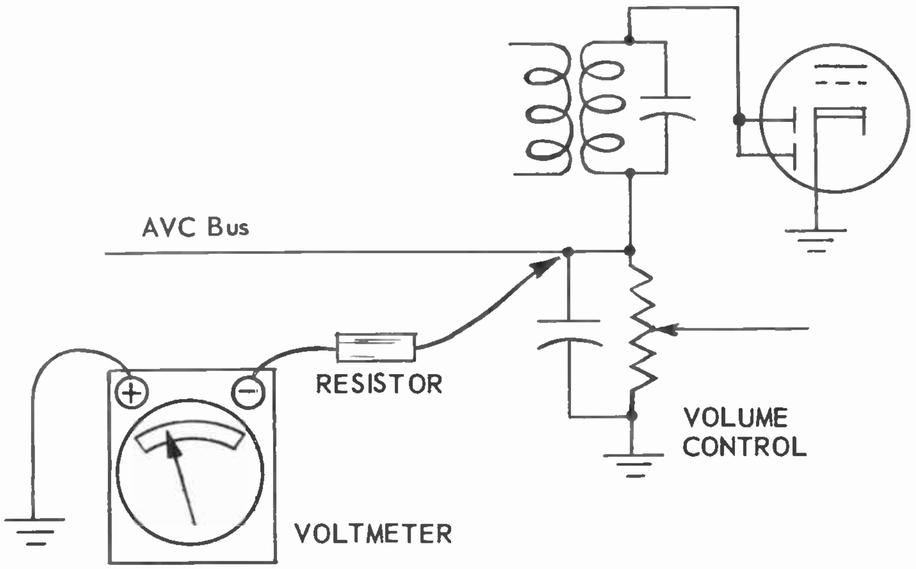


Fig. 35-5. Connecting the d-c voltmeter or electronic voltmeter for alignmeut measurements.

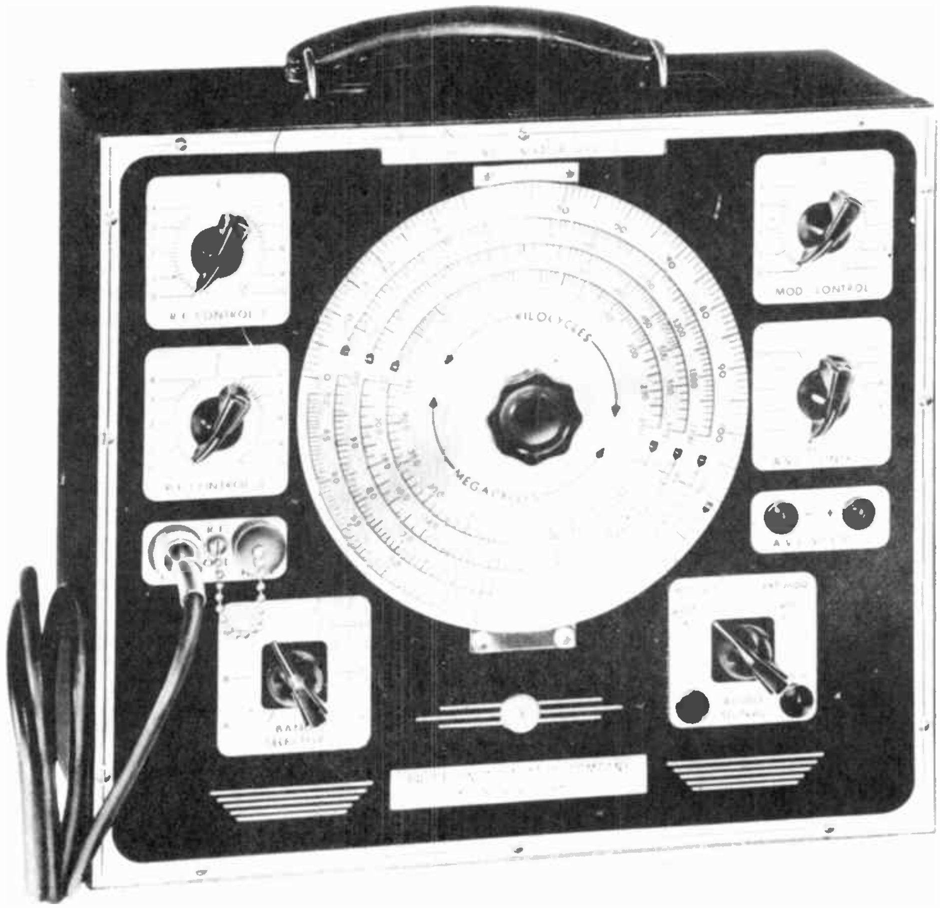


Fig. 35-6. A signal generator such as used for alignment of standard broadcast and short-wave bands.

their travel. At any rate there should be maximum capacitance or inductance somewhat below the lowest frequency marking, and minimum capacitance or inductance somewhat above the highest frequency marking. Dial pointers usually are adjustable by mechanical means which should be evident upon inspection.

a Under the heading of *Alignment* in our service instructions it is recommended that an isolating transformer be used between receiver and line. An isolating transformer is a power type with primary and secondary insulated from each other, and usually with a voltage or turns ratio of one-to-one. The purpose is to insulate a "hot" chassis from both sides of the power line while applying line voltage to the receiver. This avoids danger of severe electrical shock when touching the chassis, and prevents possible damage or burnout of testing instruments which may be connected to the chassis.

3 If no isolating transformer is available the signal generator is insulated from the receiver with a large capacitor (0.1 mf is recommended) in series with the normally grounded lead from the generator. This capacitor must have a working voltage rating of at least 150 volts.

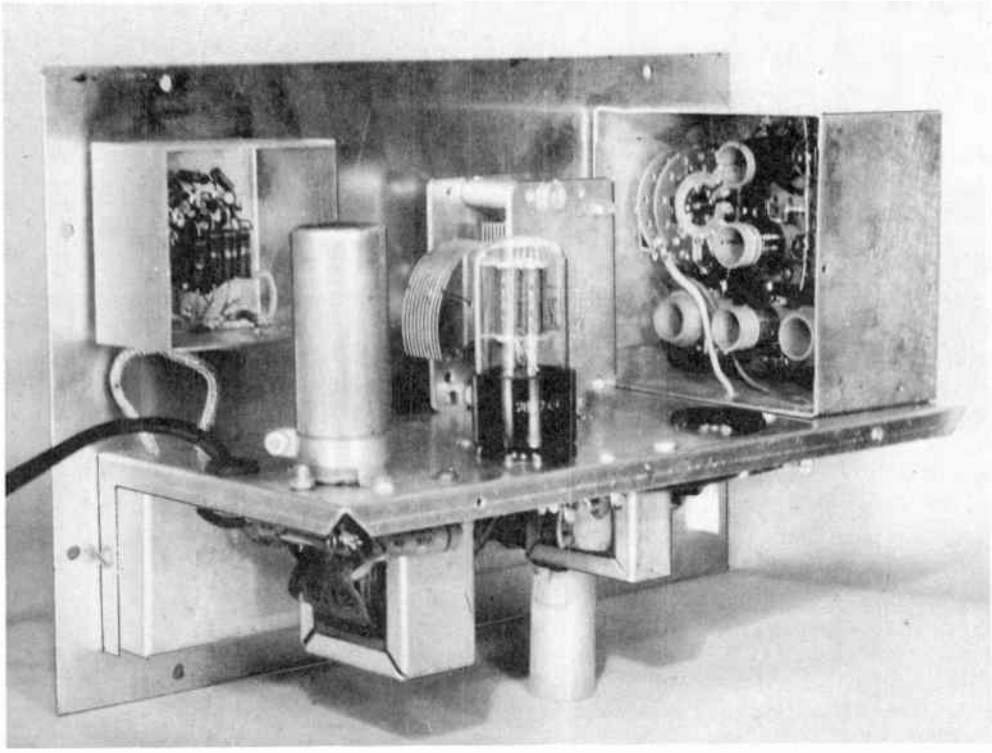


Fig. 35-7. The inside of a small service-type signal generator. Shielding covers and r-f oscillator tube have been removed to show attenuator and band tuning coils.

The next instruction is with reference to the output indicator, which is to be a high-resistance voltmeter. This means a d-c voltmeter with sensitivity or internal resistance of 20,000 ohms per volt or more. Connecting the meter between the top of the volume control and ground, as in Fig. 35-5, allows measuring changes of voltage across the control resistor. As we have learned before, the stronger the signal coming from the converter and i-f amplifier to the detector, the greater is the current in the volume control and the voltage across the control. Consequently, our voltmeter will read proportionately to the incoming signal and its amplification in converter and i-f amplifier. When alignment adjustments are correct there will be maximum amplification and maximum reading of the voltmeter.

If we use the 10-volt scale of a 20,000 ohms per volt meter the meter resistance will be only 200,000 ohms. The 300K series resistor is used to prevent too much of the volume control current from flowing through the resistance of the meter, with too little remaining in the regular volume control resistor. If an electronic voltmeter is available it may be used without a series resistor, because the resistance of an electronic voltmeter is several megohms on any of its scales. Either type of meter may be connected to any point along the avc bus in case the volume control terminal is difficult to reach. The volume control slider or manually operated knob may be at any position, it will have no effect on the alignment readings.

When we come to the alignment of television receivers you will find that the great majority of measurements are made with a d-c meter or electronic voltmeter connected to the detector load resistor, which is equivalent to the volume control resistor in the present type of receiver.

SIGNAL GENERATOR. The usnal source of signal voltages for i-f, and r-f or antenna alignment is the type of instrument commonly called a signal generator. Fig. 35-6 is a picture of a fairly typical generator suitable for a-m broadcast receivers. Inside the generator is a radio-frequency oscillator, usually of either the Hartley or the Colpitts type, with a number of coils or inductors which allow tuning through a number of frequency ranges. A range suitable for the job in hand is selected by a rotary switch whose knob is on the front panel. There is an output voltage control or attenuator for regulating the strength of signal delivered to the output cable. The attenuator control ordinarily consists of two parts, one being a coarse control, and the other a fine control or vernier control for close adjustment of the output voltage.

Practically all service generators contain also an audio-frequency oscillator whose principal purpose is to modulate the r-f output voltage. The audio frequency usually is fixed at about 400 cycles per second. On the front panel of the signal generator will be a switch that allows applying the modulation to the r-f output or cuts it off to leave a "pure" r-f output or an unmodulated output. The audio voltage is available by itself from a separate connection or jack on the panel so that it may be used for some tests on a-f amplifier and speaker systems.

The generator illustrated provides for adjustment of the percentage of modulation anywhere between zero and 100 per cent. In instruments having no such adjustment the modulation commonly is fixed at about 30 per cent, or there may be two degrees of modulation selected by a switch to provide something like 30 per cent and 80 per cent.

The remaining part of the signal generator is its power supply system. There will be a half-wave or full-wave rectifier, a filter which usually consists of a choke coil and large filter capacitors, and the voltage dropping resistors with additional bypass or filter capacitors in the lines to oscillator plates and screens.

The inside of a small signal generator is shown by Fig. 35-7. At the upper left is the coarse attenuator from which the shield cover has been removed. Underneath the tube shelf is the fine attenuator. Near the center is a twin-triode tube, one section of which is used as the a-f oscillator while the other section, with plate and grid connected together, is used as the half-wave rectifier. Behind this tube is the tuning capacitor.

The r-f oscillator tube has been removed to expose the tuning coils in the right-hand compartment, from which the cover has been removed. There is a separate set of windings for each frequency range. Parts of the power supply, except for the rectifier tube and filter capacitor can, are underneath the tube shelf.

5 The r-f voltage output from any signal generator is delivered to the receiver through a shielded cable. This is a cable with a central flexible copper conductor surrounded by insulation, with a braided flexible copper shield over the insulation. The outside of the braided shield is protected by an insulating cover. The shield is connected to chassis metal or ground inside the generator, and the other end has a clip for connecting it to chassis metal or B-minus of the receiver. This outer grounded shield is referred to as the low-side connection. The central conductor, carrying output voltage from the oscillator plate and attenuator, is called the high side. The shielded cable is necessary in order that r-f fields may not escape, to be picked up on various parts of the receiver, and so that the desired signal voltage may be delivered only where it is wanted.

The first two columns of our alignment instructions tell where the high side of the generator cable is to be connected in the receiver, how the connection is to be made, and the frequency at which the generator is to be set for the various tests.

ALIGNMENT CONNECTIONS AND ADJUSTMENTS. Before any alignment adjustments are undertaken the signal generator and receiver should be connected to the power line, turned on, and allowed to warm up

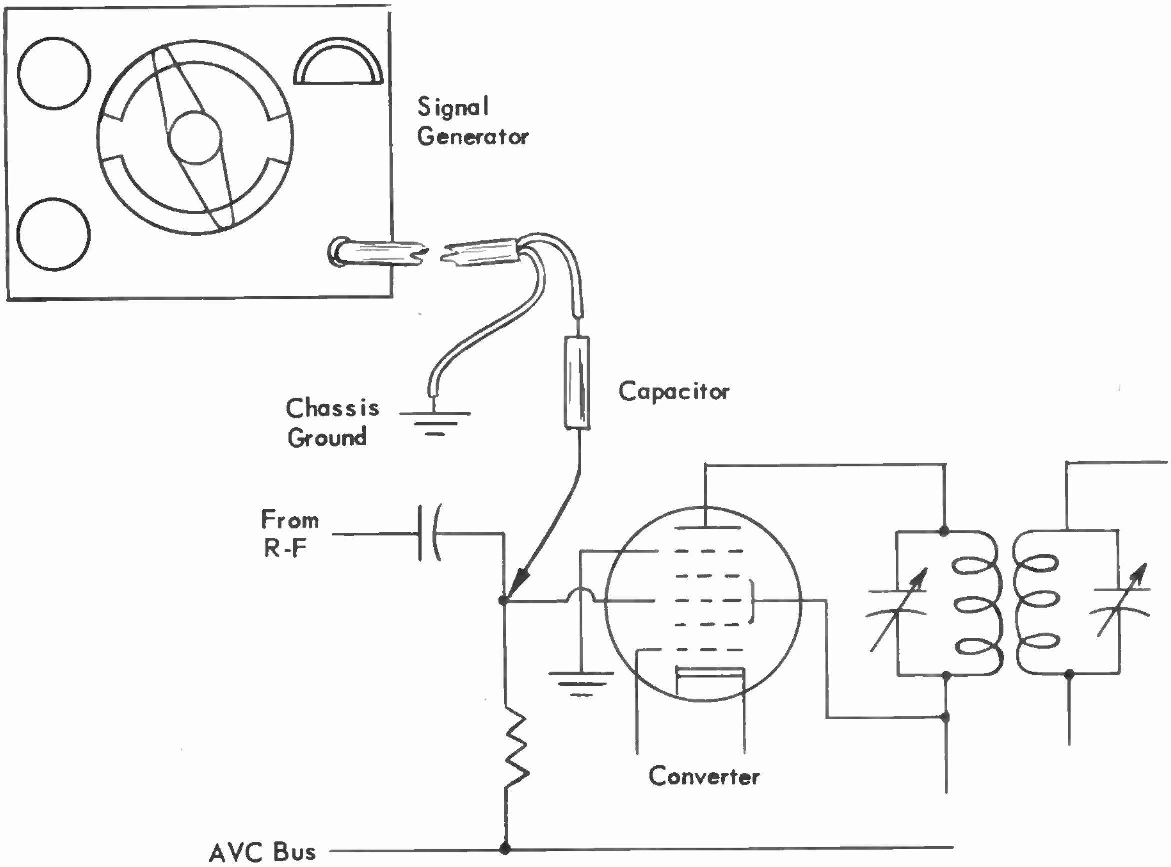


Fig. 35-8. Connection of signal generator to mixer grid of converter tube.

to normal operating temperature in all their parts. During this warming up period there will be changes in values of resistors, capacitors, and inductors. Were adjustments made during this period they would not be satisfactory during regular operation of the receiver. The warm up should be allowed to continue for ten minutes at the very least, and twenty minutes to a half hour will be much better.

No metallic shielding covers should be removed from tubes or any other parts while alignment adjustments are being made. Such shielding covers change the capacitance of protected parts and also alter the effective inductance of coils. Adjustments made without the shields would be incorrect when these parts are replaced.

Most alignment adjustments are provided with slotted head screws. Some will have hexagon shaped or possibly square headed screws. It is best to use screw drivers or wrenches made especially for alignment. These alignment tools have handles and shafts of plastic or fibre and have a minimum of metal at their points. For adjustments of very high frequency parts the entire tool must be of insulating material, with no metal at all.

When using the voltmeter on the volume control or avc bus, the volume control knob of the set may be in

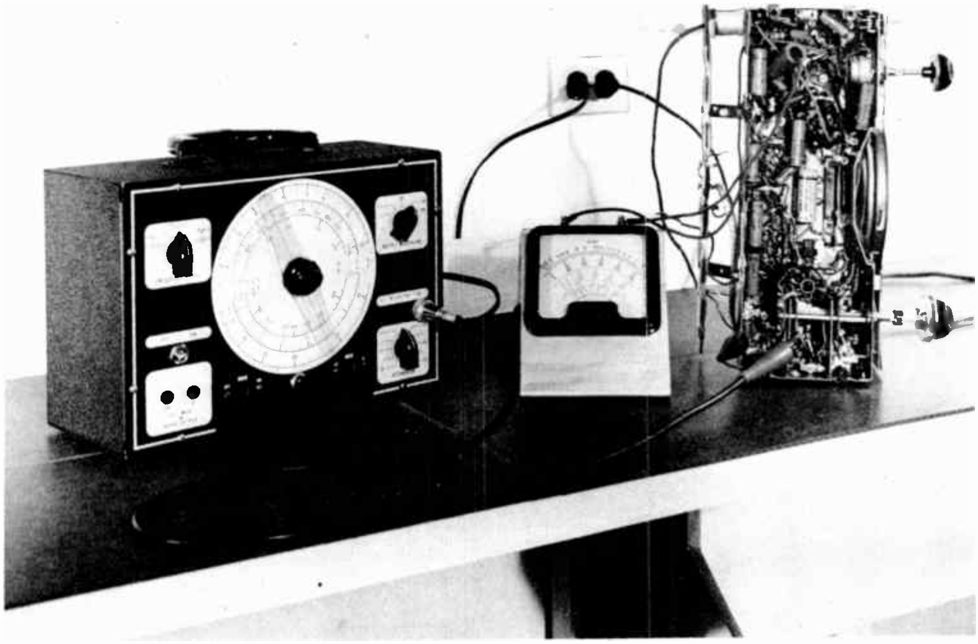


Fig. 35-9. Test setup of signal generator, high-resistance voltmeter, and a receiver to be aligned.

any position, as mentioned before. Usually the volume will be turned all the way down. With other methods of alignment a special type of output meter (described later) is connected to the speaker. Then the volume control must be kept at its maximum setting. If the receiver has push-button tuning or any kind of automatic tuning, this control should be turned off and ordinary manual or hand tuning used throughout the alignment process. Some tone controls affect the passband or selectivity of the receiver. Any tone control should be placed in the position for speech reception, or a so-called normal position, or at the middle one of several possible positions.

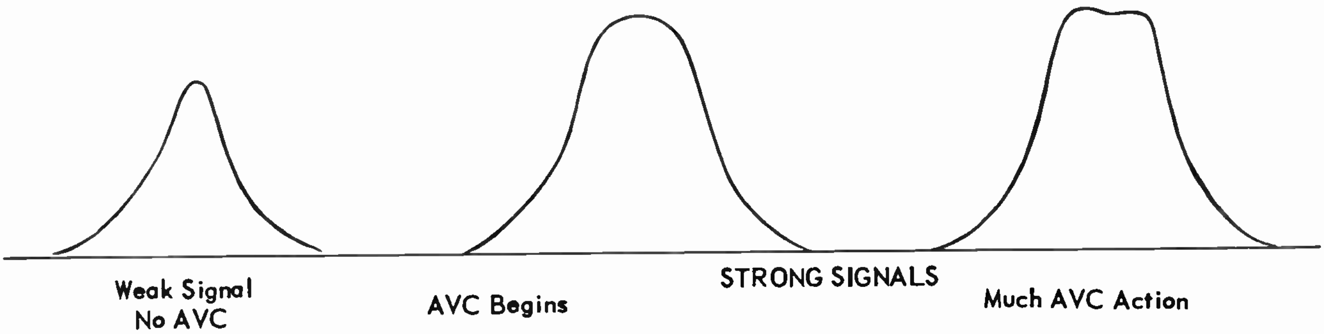


Fig. 35-10. Voltage peaks are flattened if there is avc action during alignment.

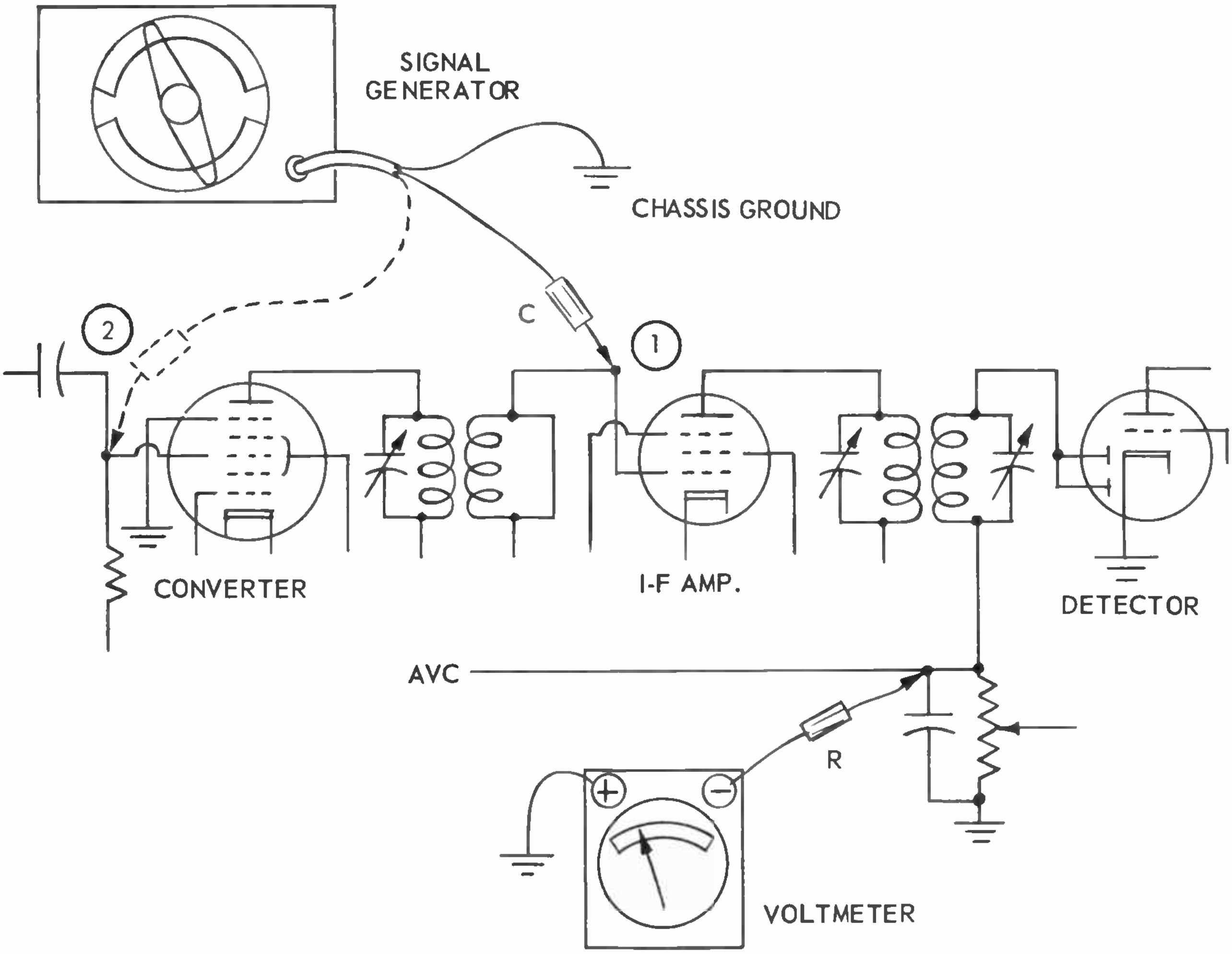


Fig. 35-11. Signal generator may be connected first to the grid of the i-f amplifier for aligning the second i-f transformer.

Step number 1 of our alignment instructions says to connect the signal generator (high side) to pin 8 of the converter tube through a fixed capacitor of 1,000 mmf capacitance. This connection is to the r-f grid or mixer grid of the converter, as in Fig. 35-8. The capacitor must be used between the generator cable and grid connection to avoid upsetting the bias voltage. The avc bias for the converter, like any other bias voltage, is applied through a conductive path or a d-c path extending all the way from the grid around to the cathode. In the output attenuator of the signal generator there is a conductive path to ground through the control resistors. Were the output lead of the generator attached directly to the grid of the tube, the conductive resistor path inside the generator would be in parallel with the grid return circuit of the tube, and bias voltage would be decreased. This is prevented by the insulating dielectric of the capacitor. R-f and i-f currents go through this capacitor, but it stops or blocks the d-c biasing voltage for the converter.

In step 1 of the instructions there is a note (a) saying to use the signal generator unmodulated. The unmodulated signal fed to the mixer grid of the converter is amplified in the converter, passes through the first i-f transformer, is further amplified by the i-f amplifier tube, and passes through the second i-f transformer to the detector. The detector demodulates the amplified high frequency voltage and delivers to the volume control resistor a direct current whose accompanying voltage is measured by the d-c voltmeter or electronic voltmeter connected across the volume control resistor.

The signal generator is tuned to deliver signal voltage at 455 kc, the intermediate frequency at which the two i-f transformers are to have peak response. The instructions say to tune the receiver dial for no signal. This means to set the dial at any point where no regular broadcast signal would be received, at any point where no nearby station is operating, or at some point between stations. This will prevent a broadcast signal picked up by the receiver from affecting or interfering with the signal from the testing generator.

We now have a setup as illustrated by Fig. 35-9, with the signal generator connected to the converter grid and the voltmeter on the volume control or avc bus. This picture shows a different receiver than the one for which we have detailed instructions, but the connections and general method of alignment would be the same. The receiver with which we are working is merely a typical example. Any other a-m broadcast receiver would be handled in practically the same way.

7 To complete step 1 of the instructions we now use our alignment tool for adjusting the secondary and primary trimmers of the second i-f transformer to produce maximum reading of the voltmeter being used as an output indicator. It is absolutely necessary to use the weakest voltage output from the signal generator that will give distinct indications on the voltmeter. Unless we use the weakest possible signal there will be avc action in the receiver. With such action the amplification or the response will be limited, and instead of having a sharp peak of output voltage as at the left in Fig. 35-10 we shall have peaks flattened by avc action as at the right. A sharp peak is easily identified by maximum voltmeter reading with only one exact adjustment of the trimmers. With flattened voltage peaks it is possible to shift the adjustments quite a ways in one direction or another with little change of output voltage, and the correct adjustment point is difficult to recognize.

To insure working with the weakest possible signal, always use the lowest range on the output voltmeter. Depending on the make and model of meter this may be 3 volts, 5 volts, or 10 volts. Generator output should be such that the meter reading varies only about 1 volt throughout the entire range of trimmer adjustments, from zero to peak. As you bring the adjustments closer and closer to their correct positions the meter reading will rise. Then you must use the attenuator of the signal generator to reduce its output, and thus prevent avc action from setting in. This matter of using the weakest possible input signal cannot be overemphasized. It is impossible to make correct trimmer adjustments with a strong signal.

For step number 2 of the alignment instructions the connections and settings are the same as for step 1. Now we adjust the trimmers for the secondary and primary of the first i-f transformer, the one following the converter, in exactly the same way as was done for the second i-f transformer. The output as indicated by the voltmeter will increase as this first transformer is brought into adjustment, and it will be necessary to keep reducing the signal voltage from the generator.

There is a certain rather definite order in which alignment adjustments are made, whether you are working with a-m, f-m, or television receivers. First comes the i-f amplifier. In aligning the i-f amplifier we commence with the transformer or coupler nearest the output indicator and farthest from the signal generator, then work back toward the generator connection. After aligning the i-f amplifier we go to the r-f oscillator. The work then is completed by adjustment of whatever r-f and antenna transformers or couplers may be used.

Wave traps are used in relatively few a-m and f-m receivers, but they are found in most television receivers. Wave traps, when used, are aligned immediately following the i-f amplifier. If there are short-wave bands, as in our present receiver, their alignment is carried out after finishing work on the standard broadcast band.

When aligning the i-f transformer nearest the detector while the signal generator is connected to the converter or mixer, it sometimes happens that adjustments are so far out of line that no distinct peak reading can be obtained on the voltmeter. Then you should connect the signal generator, through the usual blocking capacitor, to the grid of the tube preceding the last i-f transformer, as shown at 1 in Fig. 35-11. The voltmeter or other output indicator remains in its usual position.

Now it will be possible to align the secondary and primary of this last transformer for peak reading on the voltmeter. Then the generator connection is moved back to the mixer grid of the converter, as at 2, or to the grid of a separate mixer tube, and the first transformer is aligned for peak reading of the voltmeter. After completing alignment of the first transformer you should re-check the adjustments of the second transformer to get them exactly on the peak. This method of moving the generator connection progressively back from the grid of the last i-f amplifier until reaching the mixer grid often is required when working on television i-f amplifiers in which there are three or more i-f tubes and four or more couplers.

For step number 3 of the alignment instructions we adjust the wave trap. The purpose of the trap is to guard against undesired signals coming from the antenna. Therefore, we connect the signal generator to the antenna terminal of the receiver, and use the 1,000-mmf capacitor on the high-side lead. With the generator tuned to 455 kc the movable core of the wave trap coil is adjusted for *minimum* reading on the voltmeter, which still is connected across the volume control. The voltage output from the signal generator should be increased as this trap adjustment proceeds, since the object is to attenuate or reduce the effect of a strong signal on the antenna.

In steps numbers 4, 5, and 6 we align the tuner, the r-f section, or the antenna transformer for standard broadcast reception. The high-side of the signal generator lead is connected to the antenna terminal of the receiver through a 200-mmf capacitor. This capacitor often is called a dummy antenna, because it is presumed to make the capacitance load of the generator about the same as the capacitance of the usual antenna, or to provide about the same capacitance as would exist in an ordinary antenna.

There is no hard and fast rule about the capacitance to be used as dummy antenna. For standard a-m broadcast alignment, service technicians use almost anything between 75 and 500 mmf. The purpose is not to upset the effects of alignment when the signal generator is disconnected and a regular antenna connected in its place. Alignment instructions sometimes specify the use of a standard IRE (Institute of Radio Engineers) dummy antenna, which is made up as in Fig. 35-12. The fixed capacitors should be mica or ceramic types. The resistor should be a carbon type.

8 The dummy antenna for short-wave broadcast bands is a fixed carbon resistor of 400 ohms, connected be-

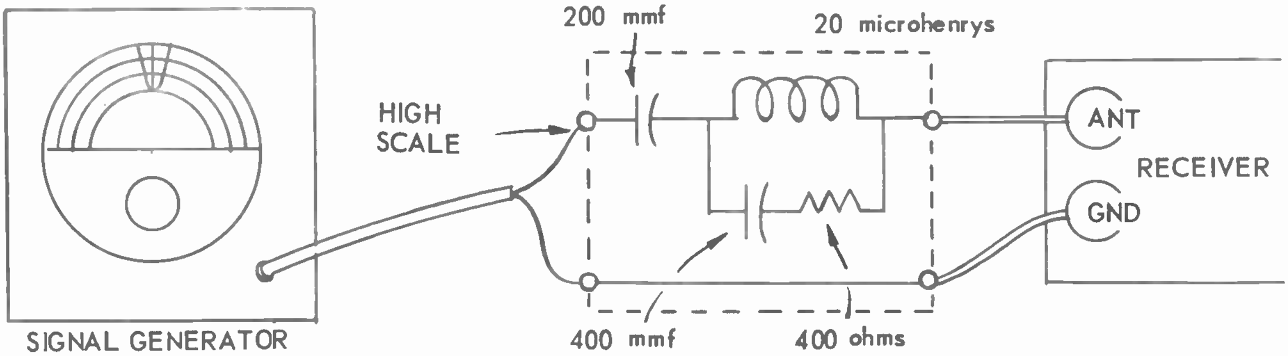


Fig. 35-12. The standard IRE dummy antenna for alignment in the broadcast band.

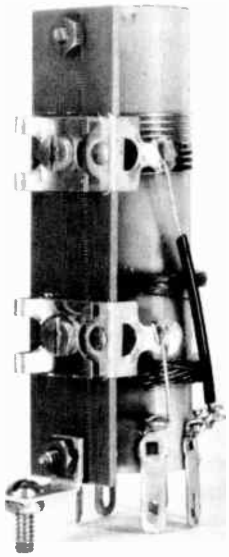


Fig. 35-13. Trimmer capacitors for broadcast and short-wave bands are mounted on the coil form that carries the antenna transformers.

tween the high-side lead of the signal generator and the antenna terminal of the receiver just as the fixed capacitor or IRE dummy antenna is used for the standard broadcast band.

In step 4 of the alignment instructions it is specified that the generator and receiver both be tuned to 1,400 kc. The two settings always must be alike, but they may be almost anywhere between 1,400 and 1,600 kc. This frequency must be one at which there are no sounds or other evidences of a broadcast signal being received and amplified. The generator must furnish the only signal voltage. Now we adjust the broadcast oscillator trimmer for maximum reading on the voltmeter. Next, in step 5, we adjust the broadcast antenna trimmer for maximum voltmeter reading.

If the trimmers are far out of adjustment to begin with you may have difficulty in obtaining a peak reading on the voltmeter. Then the thing to do is leave the signal generator at its specified frequency and operate the receiver tuning dial until you do get a peak reading. If the receiver dial is now at a frequency higher than that of the generator the trimmers are set for too much capacitance or inductance. You have had to reduce it by turning the dial to a higher frequency point. The thing to do is adjust the trimmers for less capacitance or inductance until it becomes possible to pick up the peak reading of the voltmeter with generator and receiver dials at the same frequency settings. If the receiver dial is at a frequency too low, lower than the generator frequency, the thing to do is adjust the trimmers for more capacitance or inductance. Then make a final precise setting.

If the receiver tuning dial is varied throughout its range you are practically certain to get voltage peaks of various values at several frequencies. These peaks are due to broadcast signals being picked up by the receiver in spite of having the antenna disconnected. To avoid working on a broadcast signal peak rather than on one from the signal generator, shift the generator tuning dial a little ways one direction or the other. If this changes the meter reading very decidedly you have the true signal generator peak. If changing the generator tuning has little or no effect on the meter reading you have a broadcast signal.

Now we proceed to step 6 of the alignment instructions. The signal generator remains connected as before.

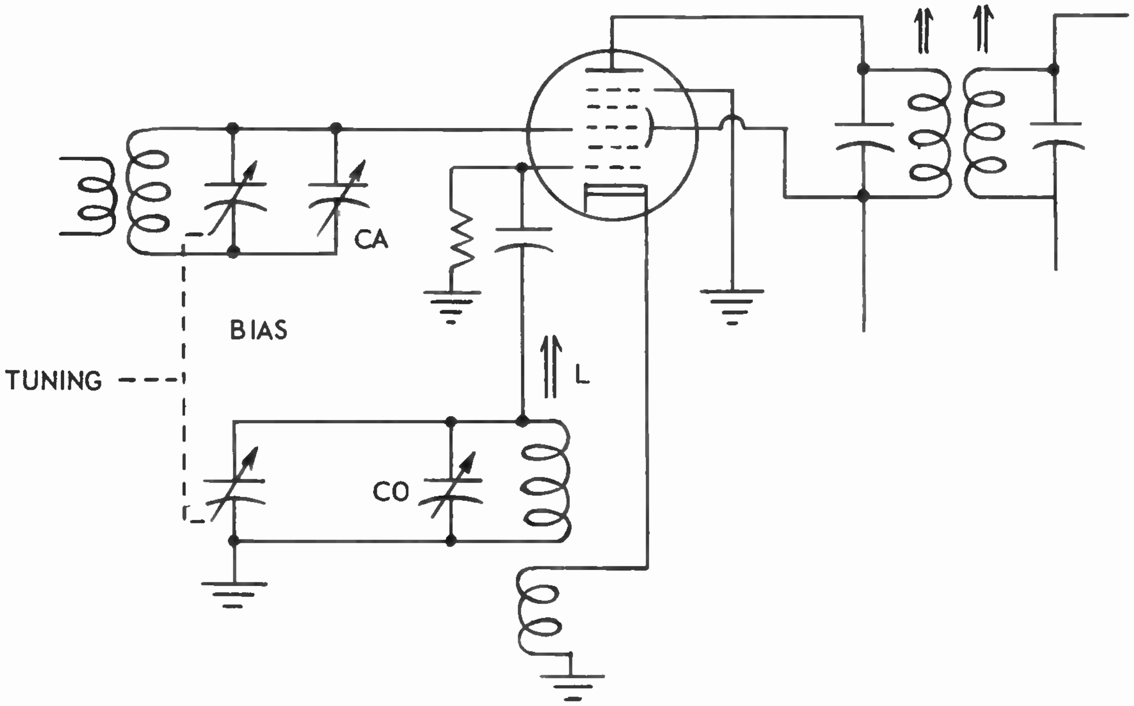


Fig. 35-14. Inductance adjustment with movable core in oscillator coil, for aligning the low-frequency end of the tuned band.

Frequencies of both the generator and receiver are readjusted for 600 kc, at which frequency the padder capacitor is adjusted for maximum voltmeter reading. The footnote on the instructions (c) says to rock the tuning dial of the receiver while making this peak voltage adjustment. The padder is to be adjusted for peak voltage even though the receiver dial is slightly off its 600 kc marking.

In receivers having no padder, which means the majority of them, the broadcast band alignment is completed when you adjust the oscillator and antenna transformer trimmers, or when you adjust the oscillator trimmer and a trimmer for a tuned r-f stage if the set has such a stage. In the receiver with which we are working the coupling between r-f amplifier plate and converter mixer grid is untuned, hence requires no alignment adjustment. When there is no padder or any equivalent arrangement for adjusting the tuning at the low-frequency end of the broadcast band the only alignment adjustments will be trimmers for the oscillator and the antenna or r-f couplings. These adjustments are made at some frequency between 1,300 and 1,500 kc. The low end of the frequency band is assumed to tune satisfactorily.

In a number of receivers there are capacitor trimmers for adjustment of antenna or r-f transformers and also for the r-f oscillator at the high-frequency end of the standard broadcast band, and there is an adjustable core in the oscillator coil for alignment at the low-frequency end of the band. One such system is shown by Fig. 35-14. Trimmer capacitors *Ca* and *Co* are adjusted in the usual way at some frequency between 1,400 and 1,500 kc. Then the adjustable core *L* in the oscillator coil is adjusted at a frequency of 600 kc, just as a padder capacitor would be adjusted. The adjustable core is the equivalent of a padder capacitor in its effect. It should be noted that either capacitor or inductor padders may be used in sets tuning only in the standard broadcast band as well as in multi-band types.

After making alignment adjustments for the oscillator at 1,400 to 1,500 kc and also at 600 kc you always should re-check the high-frequency adjustment by returning the signal generator to the high frequency and tuning the receiver dial to get a peak reading on the voltmeter. Unless the adjustments were very nearly correct before you touched them, the receiver dial reading no longer will be at 1,400 kc or at whatever frequency you originally used. Adjustment of any padder always changes the resonant frequency controlled by the trimmer. The procedure is to readjust the oscillator trimmer at the original high frequency. Then re-check the receiver dial setting with the signal generator at 600 kc. It will have changed because every adjustment of the trimmer affects the resonant frequency as adjusted by the padder. So you readjust the padder at 600 kc. This will throw out the high-frequency position of the receiver dial, and you give the trimmer a slight readjustment.

It is a case of working back and forth between the two frequencies until the receiver dial reading checks very closely with the generator frequency at both of the test frequencies. Finally, make a check and any necessary readjustment of the antenna or r-f trimmer at the high-frequency setting of the signal generator.

At last we arrive at steps numbers 7 and 8 of the alignment instructions. The high-side of the signal generator is connected to the receiver antenna terminal through a 400-ohm carbon resistor, the correct antenna for short-wave broadcast bands. Generator and receiver are both adjusted for 16 mc, or they could be set together for any other frequency near the high end of the short-wave range.

On the short-wave bands the oscillator trimmer usually will be capable of altering the frequency over such a wide range that oscillator frequency may be made either higher or lower than the received frequency. In both cases there will be a beat which is the intermediate frequency of the receiver.

6 We wish to tune the oscillator at a frequency higher than the received signal frequency. The adjustment instruction for step 7 tells how to do this. By first turning the trimmer all the way in, for maximum capacitance, you will tune the oscillator to a frequency far below that of the generator signal. As you loosen the trimmer to reduce its capacitance the oscillator frequency will increase until there is a peak reading of the voltmeter with the oscillator far enough below the generator frequency to produce a beat of 455 kc. This causes a peak reading of the voltmeter. Then you continue reducing the trimmer capacitance, and raising the oscillator frequency, until there is a second peak reading of the voltmeter. This indicates that the oscillator frequency is far enough above the generator frequency to produce the intermediate frequency of 455 kc.

The same result might be accomplished by first turning the trimmer all the way out, for least capacitance and highest oscillator frequency, then turning it in until the first voltmeter peak reading appears. This method is not so good, because you have identified only one peak. It could be the wrong one. With the method first described you are certain of the results, because you check both peaks. In a few cases it is possible to adjust the oscillator either above or below the generator frequency in the broadcast band. The same checks may be used to select the higher frequency, the one above the received signal frequency.

There is a final step not mentioned in the instructions. It is this. Start all over again at step number 1 and check every adjustment with the signal generator correctly connected and tuned to the specified frequencies. This is the only way of making certain that the job is right in every particular.

Here is a question. How should you align a superheterodyne when you have nothing but a screw driver and your ears to hear with? First you tune the set to the weakest broadcast station you can hear with the volume control all the way up. Then you go through all the steps of the alignment instructions, so far as adjustments are concerned, and make the settings for loudest reception. As the signal becomes louder, tune to weaker and weaker stations or disconnect the antenna. Now go through all the adjustments once more. The alignment may be a little lopsided as compared with what you can do with regular test equipment, but after you gain some experience in this work it will be good enough to satisfy most people.

Alignment by ear can produce passable results. But alignment with suitable test instruments correctly used will increase the sensitivity, sharpen the selectivity, and make a great improvement in tone quality. These are the things that give you the kind of reputation that builds profitable business.

36→40

COMMERCIAL TRADES INSTITUTE

KNOWLEDGE + AMBITION = SUCCESS

LEARN . . .

ACT . . .

"The heights by great men reached and kept
Were not attained in sudden flight,
But they, while their companions slept,
Were climbing upward in the night."

ENJOY . . .

COMMERCIAL TRADES INSTITUTE
CHICAGO, ILLINOIS

TELEVISION

LESSON NO. 36

FREQUENCY MODULATION

Training Your Mind

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind
C.T.I.
Perfecting Your Skill

COMMERCIAL

TRADES

INSTITUTE

Chicago, Ill.

LESSON NO. 36
FREQUENCY MODULATION

Our next big project in the way of service operations will be alignment of a frequency-modulation receiver. From there we shall proceed to combination f-m and a-m receivers, a class which includes a large percentage of all presently popular sound receivers. In becoming acquainted with these f-m receivers we shall cover also most of the work to be done in the sound sections of television receivers, except for a few minor modifications.

To understand the methods of aligning f-m sound receivers we first must learn something about frequency modulation and how it works. Fig. 36-1 represents an audio-frequency modulating voltage and the carrier that is frequency modulated by means of the audio signal. The process of modulation occurs at the transmitter.

a While there is no modulating voltage, as at the left, the carrier is of constant frequency and of constant amplitude, just like any unmodulated carrier wave. This unmodulated frequency of the carrier is called its *center frequency*. When modulating voltage is applied, and increases in a positive direction, it causes the carrier frequency to increase. There is no change of amplitude. Maximum positive audio modulating voltage causes maximum frequency of the carrier.

As the modulating audio voltage falls back toward its own zero value the carrier goes back toward its center frequency. When modulating voltage then increases in the negative direction it causes a decrease of carrier frequency, which becomes lower than the center frequency. With maximum negative modulating voltage there is minimum carrier frequency.

What happens during each complete cycle of modulating voltage is shown from *a* to *b* in Fig. 36-1. The carrier is first at its center frequency, then goes to maximum frequency, back through the center frequency to minimum frequency, and returns to the center frequency.

The greater the amplitude or strength of the audio signal the farther the carrier will be driven away from its center frequency. That is, the greater the sound "volume" the higher and lower the carrier frequency will be driven, and the greater will be the total variation of carrier frequency. The change of frequency away from the center value is called *frequency deviation*, or simply deviation. Deviation is the change with respect to the center frequency. For example, if the frequency goes alternately to a maximum of 75 kc above the center frequency and then to a minimum of 75 kc below the center frequency the deviation is 75 kc. In this case the total variation of frequency would be 150 kc, from 75 kc above to 75 kc below the center value. The total change is equal to twice the deviation.

The greater the number of cycles per second of modulating audio voltage the more often the carrier will deviate above and below the center frequency during each second.

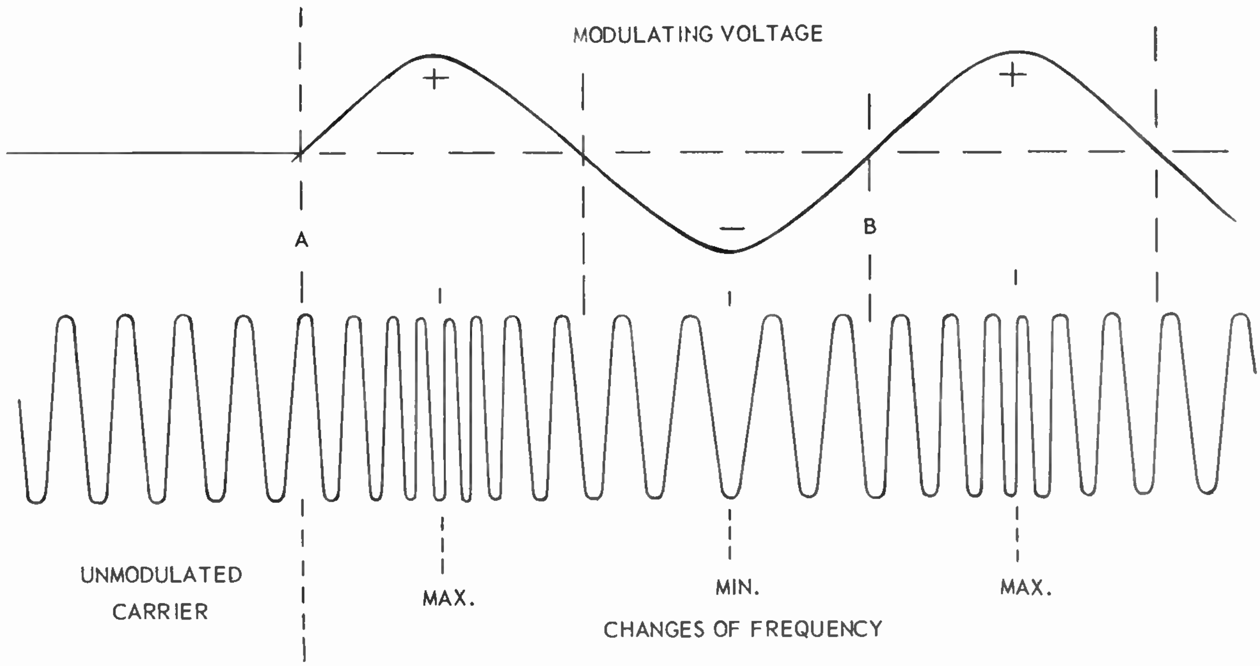


Fig. 36-1. Frequency modulation corresponding to an audio modulating voltage.

When studying amplitude modulation it was easy to get a clear mental picture of how the modulating audio voltage affected the carrier wave, for the envelope of the carrier looked just like the modulating voltage wave. But with frequency modulation the carrier amplitude is not changed at the transmitter, there is no change of shape in the carrier envelope, and it becomes somewhat more difficult to visualize the relations between the audio signal and the carrier frequency.

Some of the relations are shown by Fig. 36-2. At 1 the carrier is transmitting a loud sound and at 2 it is transmitting a relatively weak or soft sound. The difference is that there is a greater deviation of frequency, or a greater change of frequency with the loud sound than with the weaker one. There is greater difference between maximum and minimum frequencies for loud sounds than for weak ones.

In diagram 3 the carrier is transmitting a sound of high pitch or high audio frequency, and at 4 is transmitting a sound of relatively low pitch or lower audio frequency. Here the difference is that we have within a given period of time, more changes of carrier frequency for sounds of high audio frequency than for sounds of lower audio frequency.

Carrier amplitude remains the same for loud sounds and weak ones, also for high pitched and low pitched sounds. The differences are in the extent of frequency deviation, which corresponds to sound volume, and in the number of frequency changes or deviations per second, which corresponds to sound pitch or frequency.

The frequency-modulated carrier leaving a transmitter remains at constant amplitude so long as the radiated power is not changed. If, however, the power were to be increased there would be an increase of carrier amplitude, and with less power there would be a decrease of amplitude. If the received carrier is of relatively great amplitude it constitutes a strong incoming signal, and if of lesser amplitude there is a

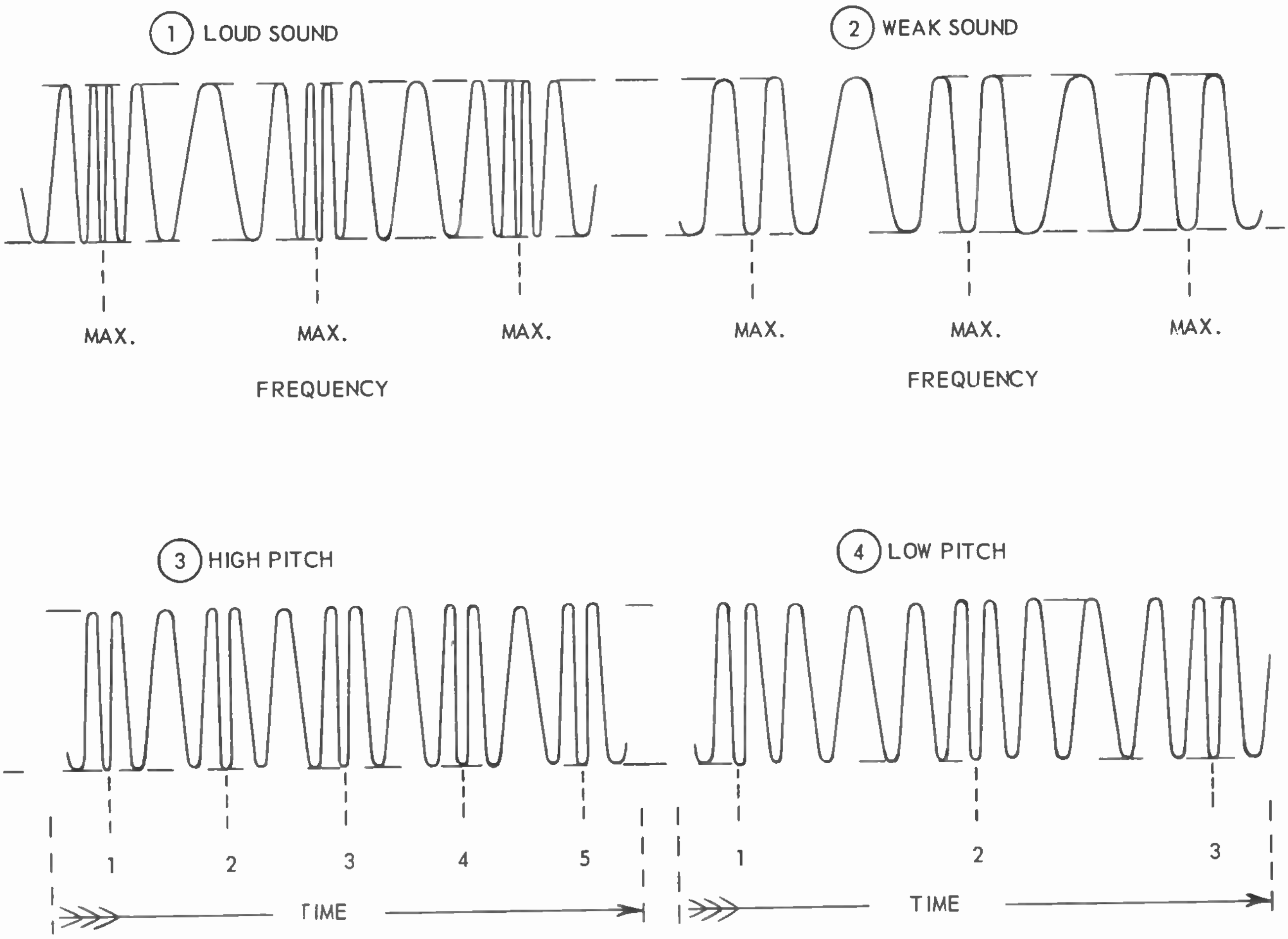


Fig. 36-2. Frequency variations resulting from different audio signals.

weaker received signal. Carriers of any strength (amplitude) may carry the same audio signals, for the volume and pitch of audio signals depend on changes of carrier frequency, not of carrier amplitude.

A frequency-modulated signal is amplified in the receiver just as an amplitude-modulated signal is amplified. Both kinds of signals are strengthened by increasing their amplitude. The amplification occurs in the r-f amplifier, in the mixer or converter, and in the i-f amplifier section. Amplification does not alter the carrier frequency nor the intermediate frequency, nor does it alter the deviations of frequency.

To illustrate, we might take the frequency-modulated voltage at the left in Fig. 36-3 and amplify it as at the right. Frequencies and deviations of frequency are the same before and after amplification. Only the amplitude is altered. We must not alter the frequency nor the changes of frequency, for that would alter the sounds that are being carried as frequency modulation. We may alter the amplitude to any extent, for amplitude represents only signal strength and has nothing to do with the modulation signal being carried.

For f-m sound broadcast and reception the standard maximum deviation is 75 kc. That is, the loudest sound which is to be transmitted is allowed to produce a deviation of 75 kc above and 75 kc below the center frequency. This amount of deviation represents maximum amplitude of the audio voltage that can be handled. Any greater audio voltage would cause still greater deviation, and over-modulation. Then there would be distortion of reproduced sound, because all the f-m receiver apparatus is designed to handle maximum deviation of only 75 kc either side of center frequency. For television sound, which is transmitted by an f-m carrier, the standard maximum deviation is 25 kc either side of center frequency.

Just because there can be greater maximum deviation for f-m sound broadcast than for television sound does not mean that louder sounds can be transmitted. It means only that apparatus used for f-m sound broadcasting is arranged to handle up to 75 kc deviation, while that for television sound is designed for maximum deviation of only 25 kc. The loudest sound transmitted by an f-m broadcast carrier would cause deviation of 75 kc. The same sound or the same loudness for television would cause deviation of 25 kc. It is merely a matter of how the transmitting and receiving apparatus for the two services is designed, adjusted, and operated.

Here is a comparison of frequency modulation with amplitude modulation.

	CHARACTERISTICS OF SIGNALS	FREQUENCY MODULATION	AMPLITUDE MODULATION
Audio Signal	Louder	Greater deviation of frequency.	Greater variations of amplitude.
	Softer	Less deviation of frequency.	Less variation of amplitude.
	Higher pitch	More changes of frequency per second.	More changes of amplitude per second.
	Lower pitch	Fewer changes of frequency per second.	Fewer changes of amplitude per second.
Carrier wave or voltage	Stronger	Greater amplitude	Greater average amplitude.
	Weaker	Less amplitude.	Less average amplitude.

The chief reason for employing frequency modulation rather than amplitude modulation for any kind of sound transmission is relative freedom from the kind of interference we call static. Static interference is caused by electric cars and trains, by ignition systems of automobiles and airplanes, by flashing electric signs, by electric switches and other electrical devices which have sparking contacts, by lightning and other atmospheric electrical disturbances, and all similar things. Static impulses vary the amplitude modulation of any carrier wave in an irregular manner. This irregular amplitude modulation acts in the diode detector of an a-m receiver just like normal sound modulation, and is reproduced from the speaker as noise.

The amplitude of frequency-modulated carrier waves in space is varied by static impulses, but this does not alter the variations of frequency that represent the desired sound signals. In the detector system of the f-m receiver there are means for greatly reducing or completely eliminating any variations of amplitude in voltages resulting from the received carrier. This removal of unwanted amplitude modulation leaves the frequency modulation unaffected. The remaining frequency modulation is detected to yield the transmitted audio signal. Were the same means for removing amplitude modulation to be employed in an a-m receiver the desired sound signal would be removed at the same time as the static, for both are carried as amplitude modulation.

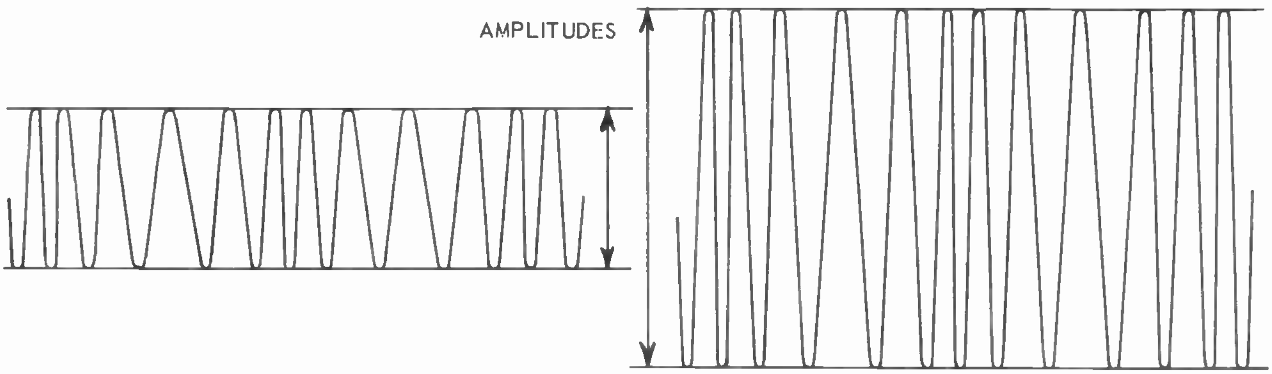


Fig. 36-3. An f-m signal may be amplified without altering the extent or rate of deviation.

F-M *CHANNELS.* With maximum deviation of 75 kc above and below the center frequency for f-m broadcasting the total change of frequency is 150 kc or 0.150 mc. To accomodate this range, while preventing possible interference between transmission, each f-m channel extends over a frequency range of 200 kc or 0.200 mc. The entire f-m broadcast band extends from 88 to 108 mc. This f-m broadcast band is just above the low band of television broadcast frequencies, as shown by Fig. 36-4. Within the total of 200 mc included within the f-m broadcast band there is room for 100 channels, each 0.200 mc or 200 kc wide.

The lowest-frequency f-m channel extends from 88.0 to 88.2 mc, with its center frequency at 88.1 mc. The next higher channel extends from 88.2 to 88.4 mc, with a center frequency of 88.3 mc. Then come other center frequencies at 88.5, 88.7, 88.9, and so on as far as the last center frequency at 107.9 mc, which is in the channel extending from 107.8 to 108.0 mc. Channels usually are specified according to their center frequencies. All the center frequencies are in decimal fractions of odd tenths; 0.1, 0.3, 0.5, 0.7, or 0.9.

At the very-high frequencies used for f-m broadcast transmissions the carrier waves behave in space like those used for television broadcasting. That is, both f-m and television carrier waves follow a straight line, much as does the beam from a searchlight. Consequently, the maximum distance of reception from a given transmitter is practically the line-of-sight distance at which you could see a powerful searchlight, or is normally no more than 15 to 30 miles. It follows that f-m transmitters separated by 75 to 100 miles may operate in the same channel without interfering with each other.

Best results are secured with an f-m receiver when signals are collected by some form of dipole antenna of the same general type used for television receivers. This type of antenna may be elevated out of doors,

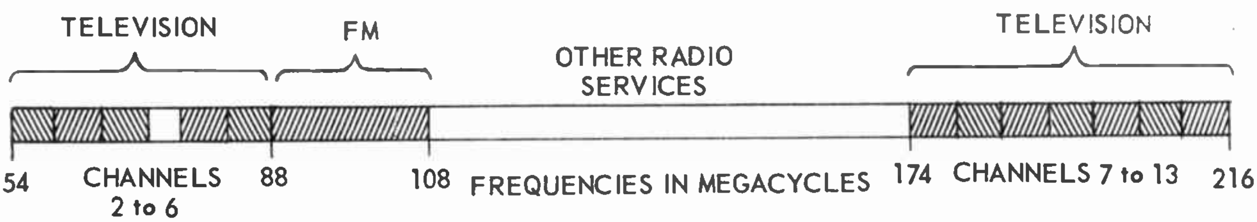


Fig. 36-4. Position of the f-m broadcast band in relation to television broadcast bands.

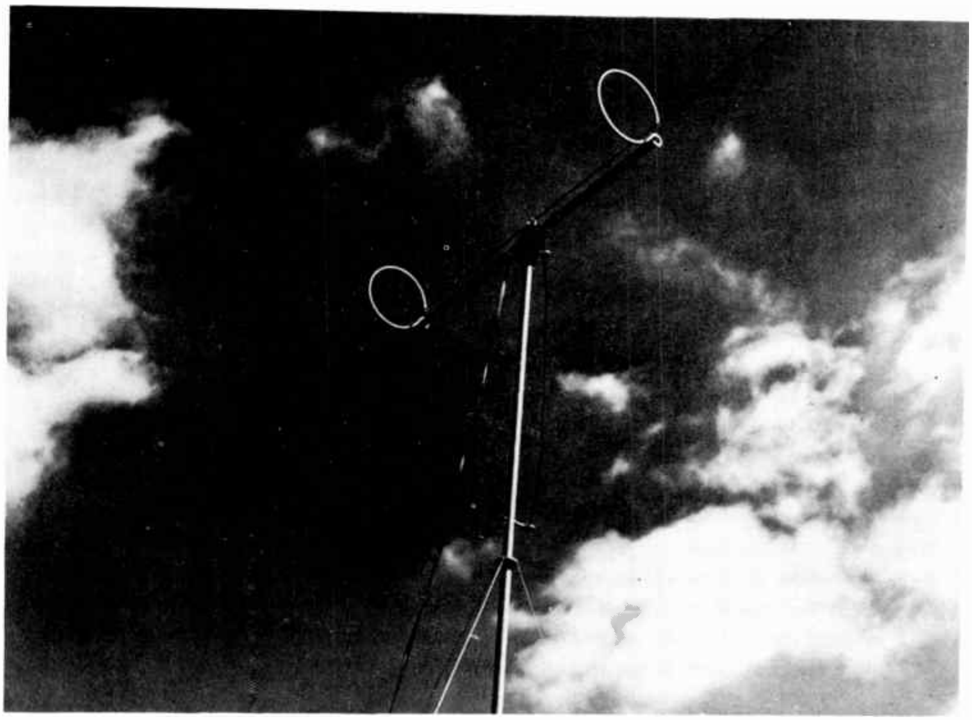


Fig. 36-5. A Tricraft antenna suitable for f-m and television reception.

or may be indoors, or may be built into the receiver cabinet. Because of the similarities of wave propagation and antenna problems for f-m broadcast and television broadcast we shall consider this whole subject at one time when we come to study television antennas.

F-M RECEIVERS. There are but few makes or models of "straight" f-m receivers, which tune only in the f-m broadcast band. Nearly all receivers which tune in the f-m band have provision for tuning also in the standard a-m broadcast band.

If we consider only the f-m portions of these combination sets a design that represents a considerable percentage of the models is shown by the simplified block diagram of Fig. 36-6. Coupled to the antenna is a converter tube performing the functions of mixer and r-f oscillator. Next there are two i-f amplifier tubes. Then, instead of the diode detector used for demodulation of a-m signals, we have what is called a ratio detector which demodulates f-m signals. The remainder of the receiver consists of the usual audio amplifier tubes, the speaker, and the power supply.

Instead of having the antenna feed through a tuned coupling to the mixer section of the converter you are just as likely to find the antenna feeding an r-f amplifier, as in Fig. 36-7. The output from the r-f stage goes to a separate mixer tube or to a twin triode in which one section is used as the mixer. There is either a separate r-f oscillator or else the second section of the twin triode is used as the oscillator.

Following the mixer in this design are two i-f amplifiers. The output of the second i-f amplifier most often

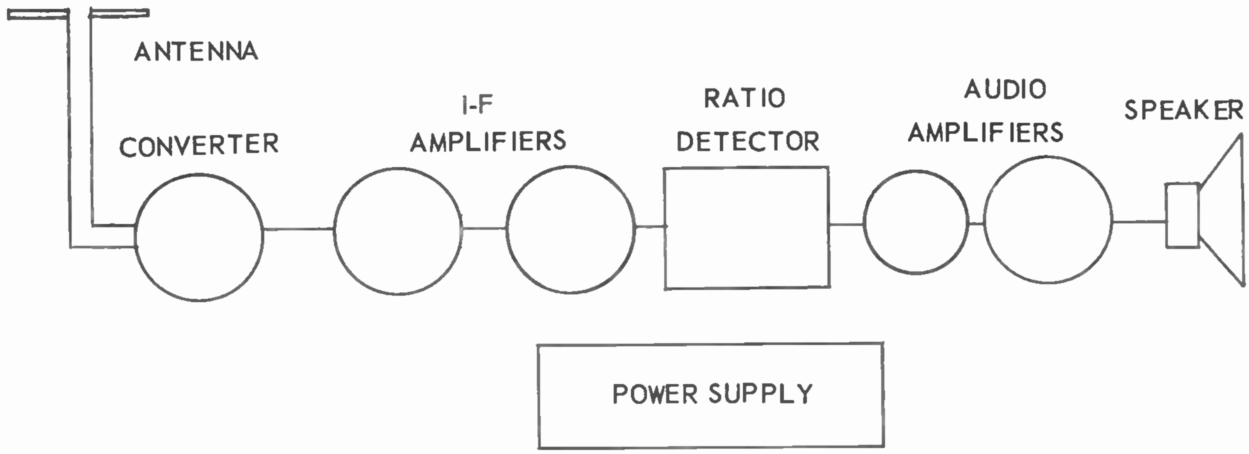


Fig. 36-6. Relations of parts in one of the simplest f-m receivers.

goes to a ratio detector for demodulation, but in some sets the demodulation is effected by a limiter stage and a discriminator. The limiter and discriminator perform the same work as the ratio detector, but in a somewhat different way. Then we have the audio amplifiers, speaker, and power supply to complete the receiver.

Other than being designed for operation at much higher frequencies there is no fundamental difference between the tuning sections of f-m sets and a-m sets. The same principles and almost the same constructions are found in both kinds of receivers.

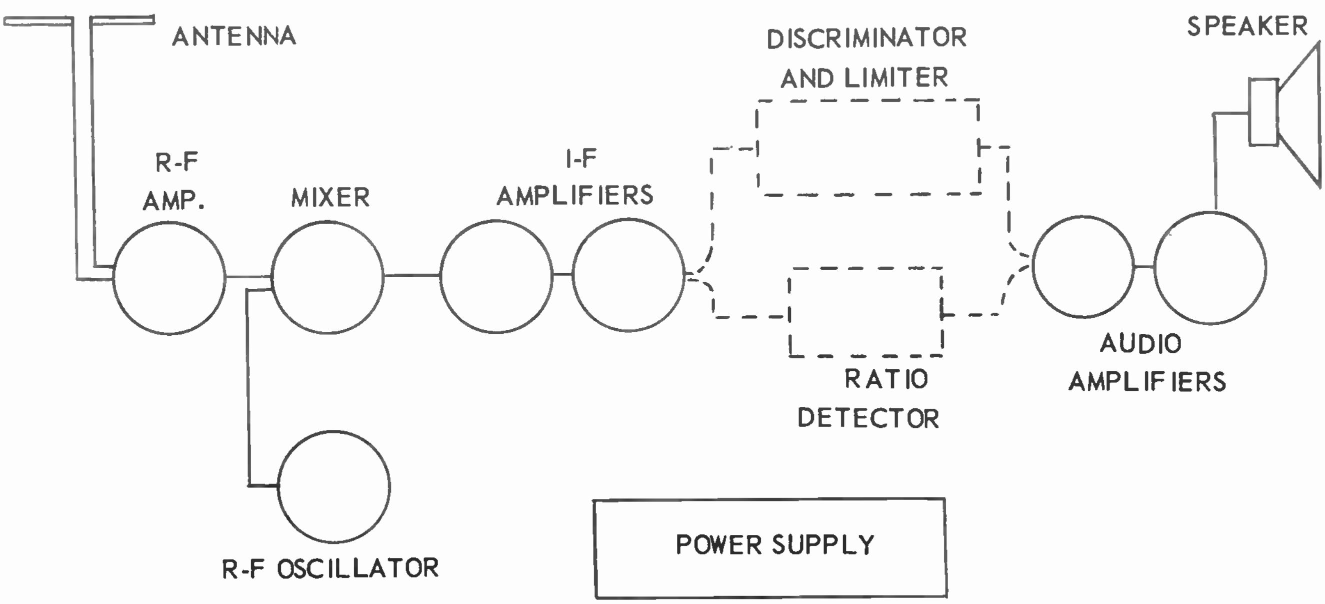


Fig. 36-7. Relations of parts in an f-m receiver having separate mixer and r-f oscillator.

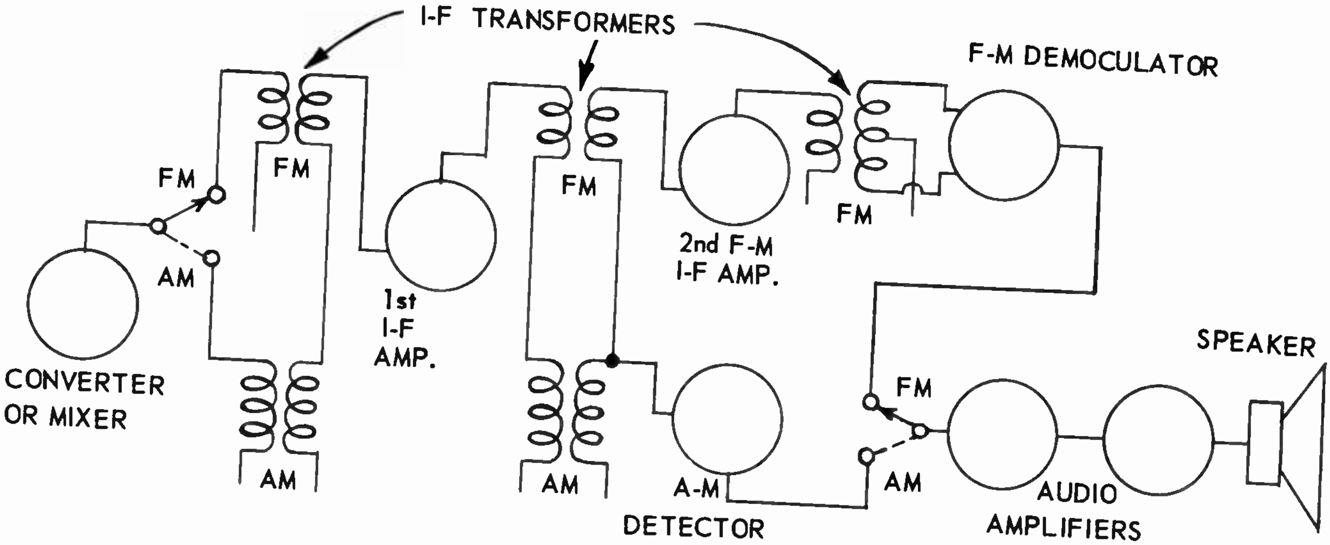


Fig. 36-8. Amplification systems for f-m and a-m signals in a combination receiver.

If a straight f-m receiver the i-f transformers would be similar to those in a-m receivers, except for changes which adapt them to high intermediate frequencies. In a combination fm-am receiver there will be two sets of i-f transformers, one for f-m and the other for a-m reception. The transformers for both bands often are enclosed within a single shield or are in the same "can".

Fig. 36-8 shows a common arrangement for combination fm-am receivers. The output of the converter or mixer is switched to either an f-m transformer or an a-m transformer, depending on which band is to be received. The secondaries of both transformers feed into the first i-f amplifier tube. The output of this tube goes to two more i-f transformers, one for f-m and the other for a-m reception. The widely different intermediate frequencies affect only the transformer tuned to one or the other, and will pass through the second unit without causing any difficulties. We shall examine the entire i-f amplifier system later on.

The secondary of the second f-m i-f transformer feeds the input of the second i-f amplifier tube, which is used only for f-m amplification. The output of this f-m i-f amplifier goes through still another f-m transformer to the f-m demodulator, either a ratio detector or a discriminator systsm, and from here the signal goes to the audio amplifiers and speaker.

Were the f-m system of our receiver to have a discriminator for demodulator there would be one additional tube, the limiter, and one more coupling transformer between limiter and discriminator tubes. The purpose of the limiter stage is to remove any amplitude modulation which may have added itself to the frequency-modulated carrier, so that a signal of constant amplitude is fed to the discriminator stage. The ratio detector type of f-m demodulator is capable of ridding itself of amplitude modulation effects, and a limiter is not essential.

Going back to the second a-m i-f transformer of Fig. 36-8, we find its secondary connected to the a-m detector, which will be a diode type. The output of this a-m detector goes to the same audio amplifiers and speaker that are used for f-m reception, the changeover being made by a second part of the band switch. For f-m reception we now have an extra i-f amplifier tube, possibly an added limiter, and necessary extra i-f transformers – all for feeding an f-m demodulator. For a-m reception we have one i-f amplifier tube and two i-f transformers for feeding a diode detector, just as in a separate a-m receiver.

RATIO DETECTOR. Our examination of the block diagrams showing typical f-m receiver designs brings out the fact that there is only one part that differs fundamentally from anything in an a-m receiver. This wholly different part is the demodulator. We shall commence our detailed examination of f-m sets with the demodulator, which nearly always is either a ratio detector or a discriminator and limiter, although a few other methods are used. Action of these demodulators in f-m receivers is identically the same as in television sets, where again we nearly always find either a ratio detector or a discriminator.

The ratio detector is used more often than the discriminator, chiefly because the ratio detector requires no extra limiter tube for removal of amplitude modulation, also because there may be somewhat more effective reduction of noise interference on very weak received signals. The ratio detector is a comparatively recent development. All early f-m and combination fm-am sets used discriminators, and the discriminator system still is used in many high-quality f-m and television receivers. Sometimes a ratio detector is marked on a service diagram as being a discriminator. The difference will become apparent as we proceed.

For f-m demodulation by any method there are three requirements. First, when the i-f signal is at its center frequency (unmodulated) there must be zero audio signal output. Second, when the i-f signal deviates to a higher frequency there must be audio output voltage in one polarity, of strength or amplitude proportional to the amount of frequency deviation. Third, with deviation to a lower frequency the audio output must go

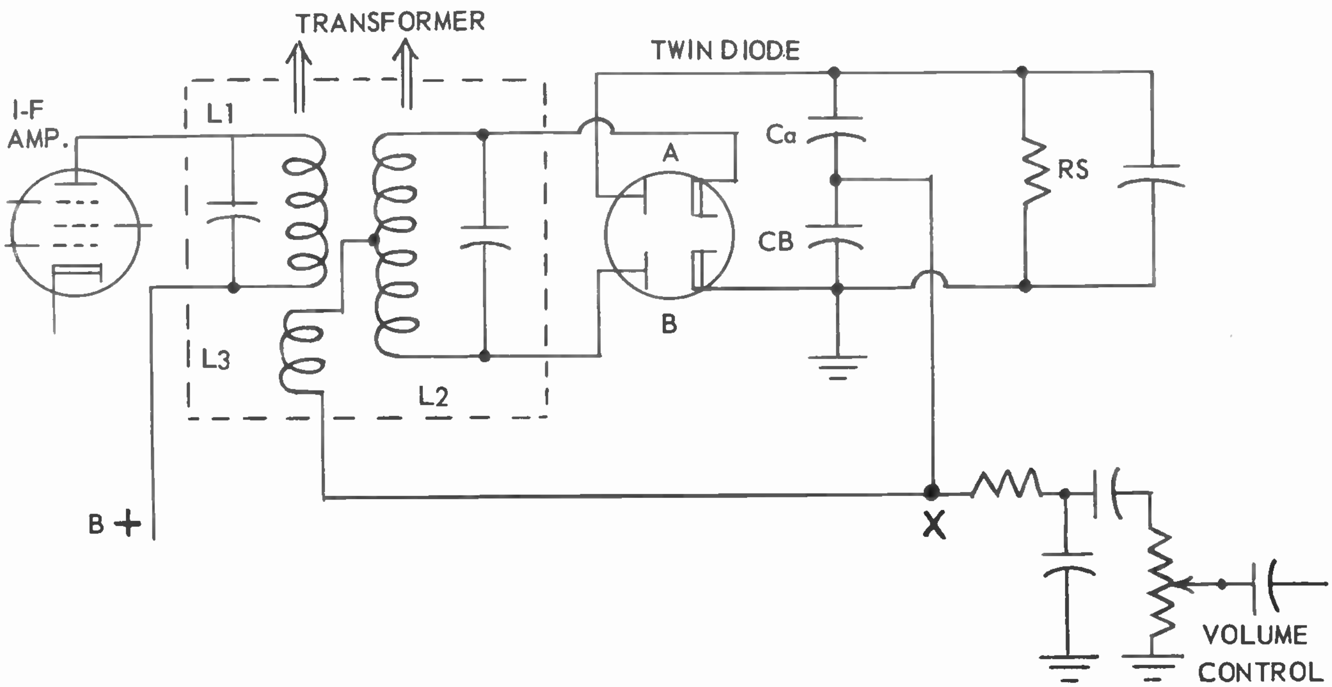


Fig. 36-9. Connections for a popular type of ratio detector.

to opposite polarity, in strength proportional to deviation. To commence the explanation of ratio detectors we shall assume an i-f signal centered at 10.7 mc and having equal deviations above and below this frequency.

The circuit connections for ratio detectors may be modified in a great variety of ways without affecting the basic operating principle. One of the circuits in general use is shown by Fig. 36-9. Here we have the two diodes required for any f-m demodulator. The two units are in a single twin-diode tube. The diodes are coupled to the last i-f amplifier tube through the detector transformer.

The primary winding of the detector transformer, in the plate circuit of the i-f amplifier, is marked *L1*. The center-tapped secondary, whose outer ends are connected to the diodes, is marked *L2*. A third winding, *L3*, is inductively coupled to the primary. The upper end of *L3* is connected to the center tap of the secondary and the lower end is connected to the output side of the diodes. Both the primary and the secondary are tuned to resonance at the intermediate center frequency, 10.7 mc.

Electron flow paths in the two diodes and other parts of the circuit are shown by Fig. 36-10. These flows occur only during half-cycles of intermediate frequency which make the upper end of the secondary negative and the lower end positive with reference to each other, for only then are the diode cathodes negative and their plates positive with reference to each other.

Electron flow for diode *A*, shown by broken-line arrows, is as follows. Diode plate. Capacitor *Ca*. Point *X*. Winding *L3*. Upper half of secondary *L2*. Back to the diode cathode.

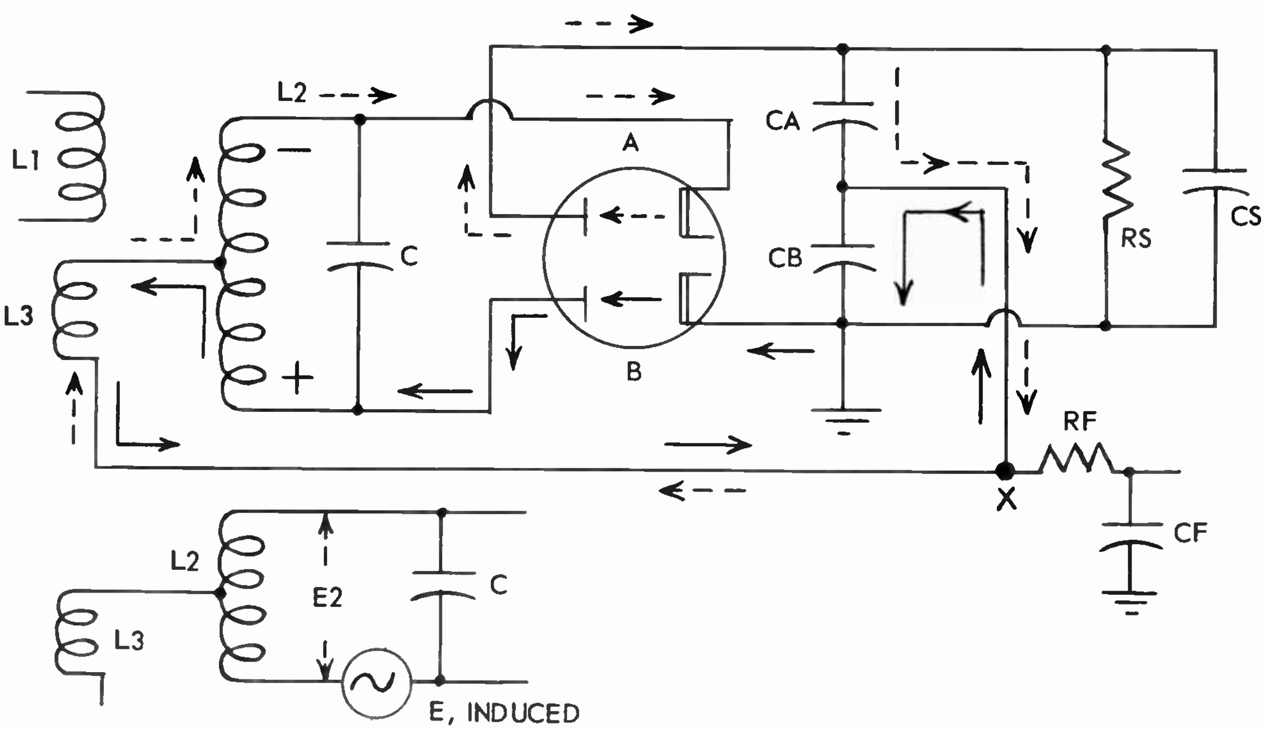


Fig. 36-10. Electron flows in the diode circuits of the ratio detector having two capacitors in the audio output section.

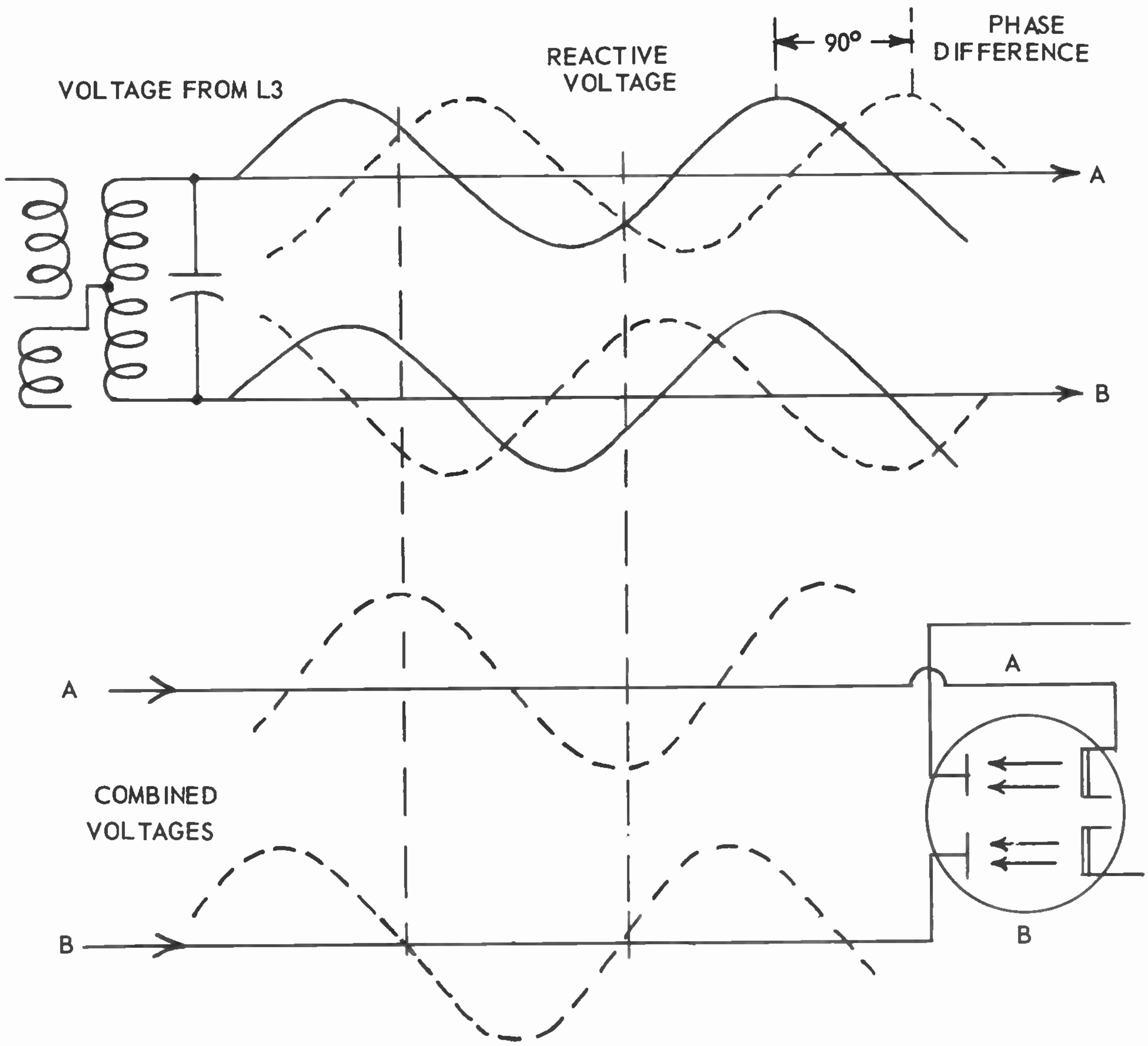


Fig. 36-11. Voltages in ratio detector secondary and diodes when there is no deviation.

Flow for diode *B*, shown by full-line arrows, takes this path. Diode plate. Lower half of secondary *L2*. Winding *L3*. Point *X*. Capacitor *Cb*. Back to diode cathode.

Deviations of frequency in the primary of the detector transformer will cause greater conduction and electron flow first in diode *A* and then in diode *B*. These electron flows will become alternately greater and less at the same rate as the rate of deviation to higher and lower frequencies. Since the changes of frequency deviation follow the audio frequency of the transmitted signal we shall have at point *X* an electron flow alternating at this audio frequency. Point *X* is the audio-frequency output point of the ratio detector. Now we shall learn what happens in the secondary of the transformer to cause variations of the diode currents.

7 Two voltages are produced simultaneously in the secondary winding. One is the voltage fed in at the center tap from winding *L3*. Because this voltage is introduced at the center of the secondary winding it appears without change of polarity or phase at both top and bottom of the secondary. That is, when voltage from *L3* is positive it remains positive at both ends of the secondary, and when it is negative it remains negative at both ends of the secondary.

The voltage brought in from winding *L3* is 180° out of phase with the primary voltage across *L1*, just as the induced voltage across the secondary winding *L2* is 180° out of phase with the primary voltage. The voltage induced across the secondary winding *L2* acts as a generator in series with *L2* and the capacitor connected across it, as shown by the small diagram at the lower left in Fig. 36-10.

At the resonant i-f frequency no deviation will occur. Therefore, the current *I2* due to the induced voltage is in phase with the induced voltage, because a series resonant circuit acts resistive at resonance. Since current *I2* is allowed to flow, a reactive voltage *E2* developed across the secondary is 90° out of phase with the voltage that caused the flow. This phase relationship of current and voltage is true of any inductance. Then the voltage *E2* developed across the secondary winding is 90° ahead of the current *I2*.

5 At any one instant the reactive voltage *E2* will be of opposite polarity at opposite ends of the secondary. When the reactive voltage is positive at the top it will be negative at the bottom, and vice versa. The reactive secondary voltage *E2* will be 90° out of phase with the primary voltage. This comes about because both primary and secondary are tuned to resonance at the center frequency. In any double-tuned transformer there is a 90° phase difference between primary voltage and reactive secondary voltage *E2* when the resonant frequency is applied to the primary.

Fig. 36-11 shows the two voltages in the secondary winding when we apply to the primary the center intermediate frequency, at which both primary and secondary are resonant. The reactive voltage is shown by broken-line curves. Voltage from *L3* is shown by full-line curves. Relations of the two voltages shown on the line extending from the top of the secondary are as they appear at this end of the winding. Along the line extending from the bottom of the secondary are shown the two voltages as they appear at this end of the winding.

All conditions previously mentioned are fulfilled in Fig. 36-11. Voltage from *L3* is of the same polarity and phase at both ends of the secondary. The reactive voltage *E2* is 90° out of phase with voltage from *L3*, which is 180° out of phase with primary voltage. Reactive voltage at the top and bottom of the secondary is of opposite polarity, there is 180° phase difference.

4 The two voltages at each end of the secondary combine their strengths or amplitudes as shown on the two lines down below in Fig. 36-11. These curves, showing combined voltages at top and bottom of the secondary, are arrived at by adding together the original voltages when both are positive or both are negative, and by subtracting when one is positive and the other negative. The principal thing to be noted is that amplitudes or strengths of combined voltages from top and bottom of the secondary are equal. The voltage from the top of the secondary acts on diode *A*, and voltage from the bottom acts on diode *B*. Since both voltages are of equal strength the two diode currents are equal. Remember, this is the condition existing at the center frequency, when there is no deviation.

The equal diode currents will flow in the paths we traced in Fig. 36-10. Current coming toward point X from either direction will be balanced by current flowing away from this point in the same direction. Of course, you don't actually have two separate currents in the same conductor at the same time, you have a single current or electron flow which is the sum or difference. In the present case the sum and the difference are zero, so there is no electron flow to or from point X, which is the audio output point.

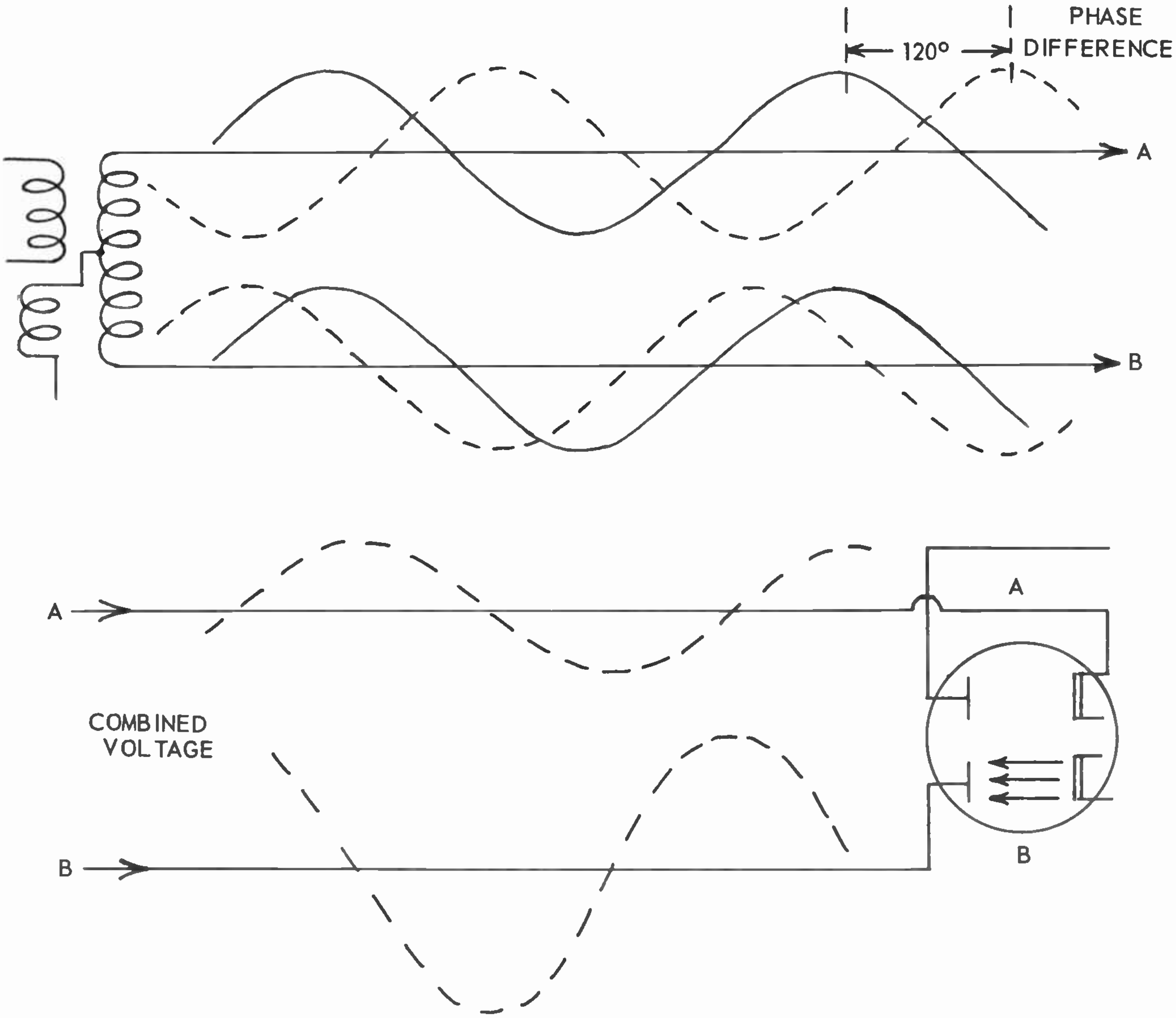


Fig. 36-12. Ratio detector voltages when there is deviation to a frequency lower than the center frequency.

Now we have satisfied the first of our three original requirements; when the i-f signal is at its center frequency (unmodulated or with no deviation) there is zero audio signal output.

Next we shall turn to Fig. 36-12, where are shown the two voltages in the secondary winding as they exist when there is deviation to a frequency lower than the center frequency. Voltage from L3, shown by full-line curves, is the same as before.

Even though a change of applied frequency were to change the timing of the voltage of *L3*, induced in that winding by primary current variations, it would make no difference for it is merely our reference voltage with which the reactive secondary voltage $E2$ will combine. There has been a change of phase difference between primary voltage and reactive secondary voltage. The difference was 90° in Fig. 36-11, now it is 120°. This change of phase difference is explained by facts with which we have been acquainted for a long time.

First, when an emf or voltage is induced in a tuned secondary circuit that circuit behaves like a series resonant circuit. At frequencies lower than resonance in a series resonant circuit there is an excess of capacitive reactance. Second, when there is an excess of capacitive reactance the current leads the applied voltage or emf. This is the same as saying that the voltage lags the current, or we may say that increase of capacitive reactance makes the voltage peaks occur later in time or in phase. These two facts explain the whole performance when deviation is to a lower frequency and the secondary acts capacitive.

Here is the explanation: The voltage induced in $L2$ and $L3$ is 180° out of phase with the primary voltage, as before. Voltage $L3$ is the reference voltage, as explained. The induced voltage causes a current 12 to flow in the circuit. Because the circuit is operating below the resonant frequency the series inductance and capacitance behaves like a capacitor, and $I2$ leads the induced secondary voltage by a certain amount, in degrees, depending on the amount of deviation below the resonant frequency.

A reactive voltage $E2$ is developed across $L2$, because of the current $I2$, which is 90° ahead of $I2$ due to the coil inductance. The voltage at the upper end of $L2$ is 180° out of phase with the voltage at the lower end. The voltage at this end then is 90° behind the current $I2$. In a capacitive series circuit the voltage lags, its peaks occur later in time. In Fig. 36-12 the voltage has taken the lag. Its positive peaks now occur 120° after the reference peaks of $L3$, whereas in Fig. 36-11 the reactive voltage peaks occur only 90° after the reference peaks. We are using this particular change of phase only to illustrate the action. Were the deviation something less than has been assumed there would be less lag, and were the deviation still greater there would be more lag of reactive voltage.

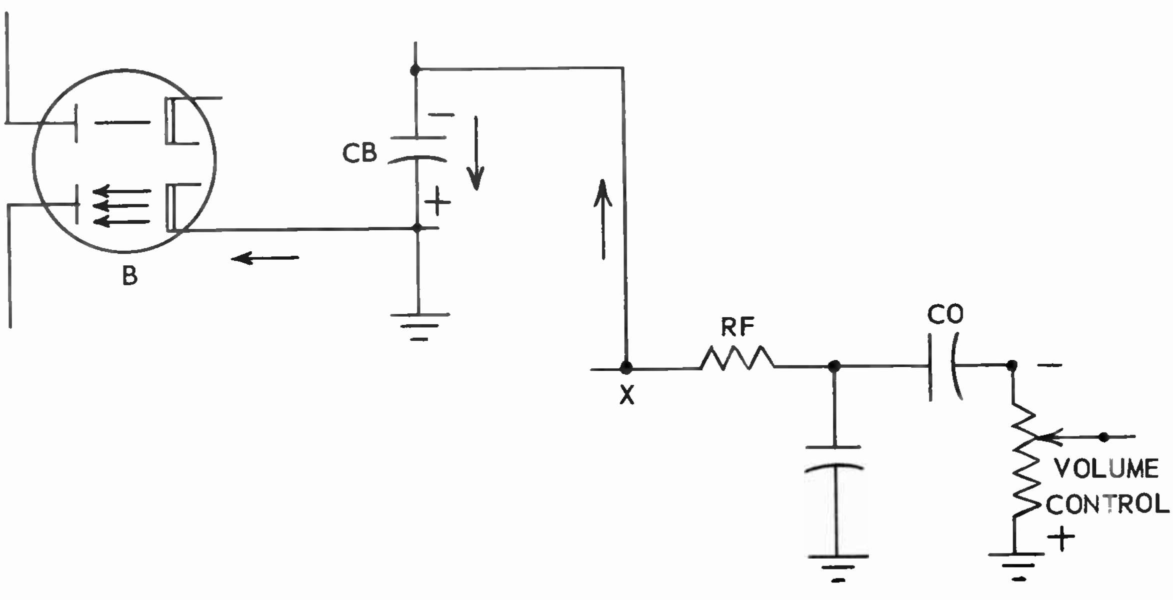


Fig. 36-13. Path of increased current when deviation is to a lower frequency.

Down below in Fig. 36-12 are shown the results of the change in phase that occurs when there is deviation to a lower frequency. Again the curves from combined voltages have been computed by adding and subtracting the separate voltages up above. There is a considerable decrease of voltage applied to diode A, and a large increase of voltage applied to diode B. Now diode A carries less current, while diode B carries more current than before. Were there greater deviation there would be still greater change in diode currents. With less deviation there would be less change in diode currents.

Next we may go back to Fig. 36-10 and note how the changed diode currents act in the connected circuits. To more easily see what happens we may look at only the portion of the circuit associated with diode B and capacitor Cb, as this portion is reproduced in Fig. 36-13.

There is more current in diode B, which means that more current is flowing into and out of capacitor Cb. The result is an increase of charge on Cb and an increase of potential difference across this capacitor. According to the direction of electron flow we can see that the top of Cb is becoming more negative and the bottom more positive. The top of this capacitor is connected through point X, resistor Rf, and capacitor Cc to the top of the volume control resistor. The bottom of Cb is connected through ground to the bottom of the volume control. Consequently, potential difference across the volume control resistor must be undergoing the same change as potential difference across capacitor Cb.

The top of the volume control, and also its slider that connects to the following audio amplifier, are becoming more negative with reference to ground. This is the result of deviation to a lower frequency. Thus we have satisfied the second of our three original requirements; when the i-f signal deviates to a lower frequency there is audio output voltage in negative polarity, of amplitude proportional to the amount of deviation.

Now in Fig. 36-14, we may see what happens when there is deviation to a higher frequency. Along the upper two lines, connected to top and bottom of the transformer secondary, are $L3$ and reactive voltages as they appear at the respective ends of the secondary. The phase difference has decreased to only 60°, it is less than with no deviation. Again the voltage induced in the secondary is 180° out of phase with primary voltage. Above resonance the series inductance-capacitance circuit acts inductive. Then the resulting $I2$ lags the induced voltage by a certain amount, in degrees, depending on the deviation. $I2$ develops a voltage $E2$ across the secondary winding 90° ahead of $I2$, and since the voltage at the bottom of the secondary is 180° out of phase with the top, then the voltage at this end is 90° behind the current $I2$.

7 The same two facts that explained the change of phase due to lower deviation also explain this new and opposite shift of phase due to higher deviation. At any frequency higher than resonance the reactance of a series resonant circuit becomes more inductive than capacitive. When there is an excess of inductive reactance the current lags the voltage or the voltage leads the current. This means that voltage peaks occur earlier in time or phase. The earlier voltage has reduced the phase difference to 60°.

The combined voltages are shown in the lower part of Fig. 36-14. Deviation to a higher frequency has brought about an increase of voltage on diode A and a decrease on diode B, just the opposite of what happened with deviation to a lower frequency. There is a resulting increase of current in diode A and a decrease of current in diode B.

During the interval of time between the former deviation to a lower frequency and the present deviation to a higher frequency the charges on capacitors Ca and Cb have been equalizing themselves by electron flow occuring between these capacitors and resistor Rs and capacitor Cs which are shown in Figs. 36-9 and 36-10. The charges and potential differences on Ca and Cb will have dropped back to low values or to zero due to electron flow through the paralleled resistor Rs.

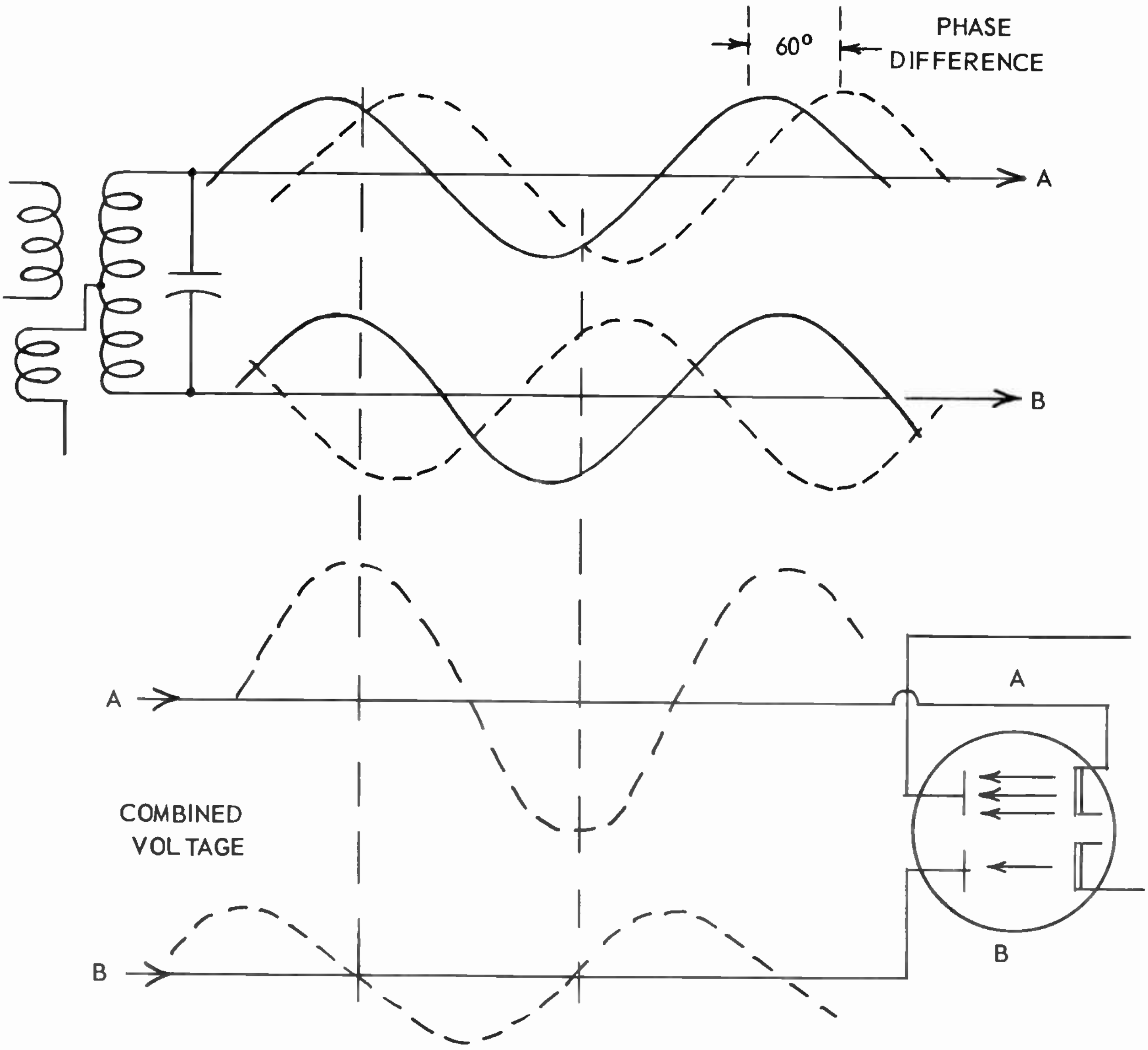


Fig. 36-14. Ratio detector voltages when there is deviation to a frequency higher than the center frequency.

With the new relatively large charging current through capacitor *Ca* and the small current through capacitor *Cb* the charge and potential difference of *Cb* will be much smaller than before. This same smaller potential difference will exist across the volume control resistor, which is connected to top and bottom of capacitor *Cb* as previously explained. The top of *Cb* still is negative and the bottom still is positive, but the difference of potential with the present deviation to higher frequency is not so great as with the earlier deviation to lower frequency.

Across capacitor *Cb* we have a direct voltage which is becoming alternately more negative and less negative with deviations to lower and higher frequencies. This changing direct voltage is applied to the volume control resistor through capacitor *Co*. But capacitor *Co*, like any other capacitor, does not conduct for direct voltage – it passes only the alternating component of the direct voltage. Then in the volume control we have alternating current and voltage which becomes negative when there is deviation to lower frequencies and becomes positive when there is deviation to higher frequencies.

We have satisfied the third and last of our original requirements; with deviation to a higher frequency the audio output voltage swings positive, in strength proportional to the amount of deviation. The ratio detector has changed deviations of frequency into an alternating audio voltage that follows the modulation put on the carrier at the transmitter.

Before proceeding to learn how the ratio detector gets rid of amplitude modulation in the carrier and i-f signal voltages it will be well to look at another typical circuit as shown by Fig. 36-15. If you compare this detector with the one of Fig. 36-9 it is apparent that the transformer and its connections to the i-f amplifier and to the twin diode have not been altered. There are, however, changes on the output side or load side of the twin diode. Instead of the two capacitors *Ca* and *Cb* of the earlier circuit there are now two resistors, *Ra* and *Rb*. We have retained capacitor *Cs* but have omitted the single paralleled resistor *Rs* which appears in Fig. 36-9.

The effect of frequency deviations on diode voltages and currents is exactly the same as previously described. Electron flow paths for the two diode circuits are as shown by Fig. 36-16, where broken line arrows indicate flow for diode *A*, and full-line arrows for diode *B*.

The path for diode *A* is as follows: Diode plate. Resistor *Ra*. Ground, to lower end of capacitor *Cf*. Through *Cf* to point *X*. Resistor *Rc*. Winding *L3*. Upper half of secondary *L2*. Back to the diode cathode.

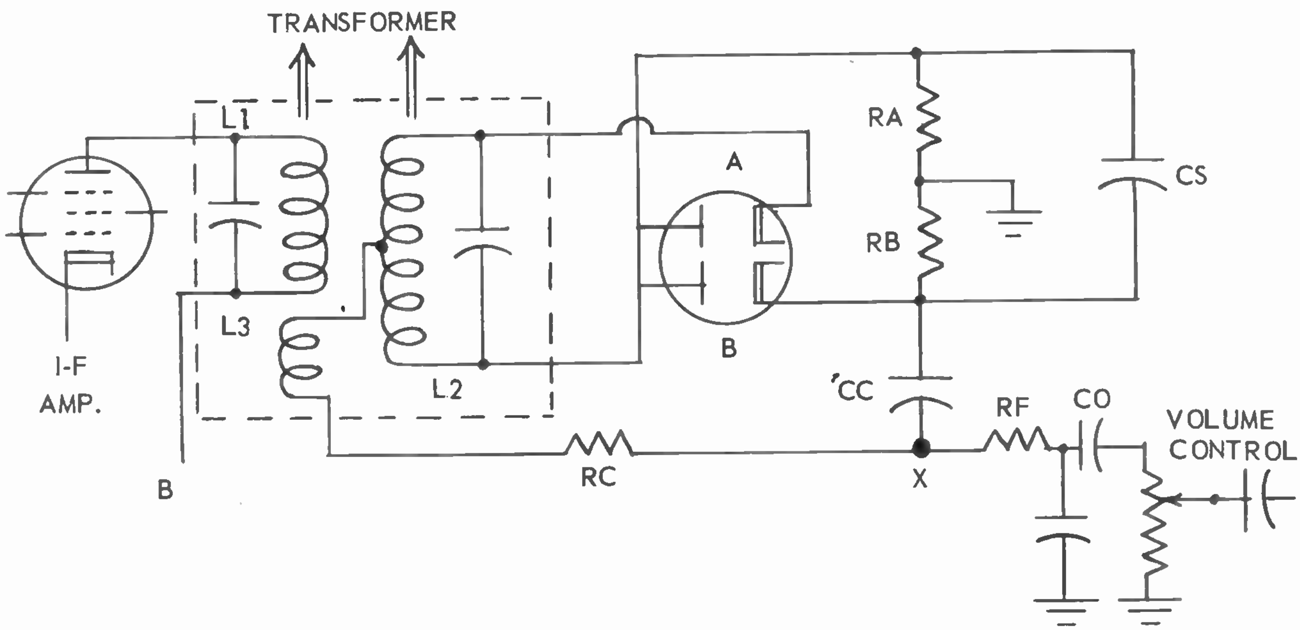


Fig. 36-15. Ratio detector having two resistors instead of two capacitors in the audio output circuit.

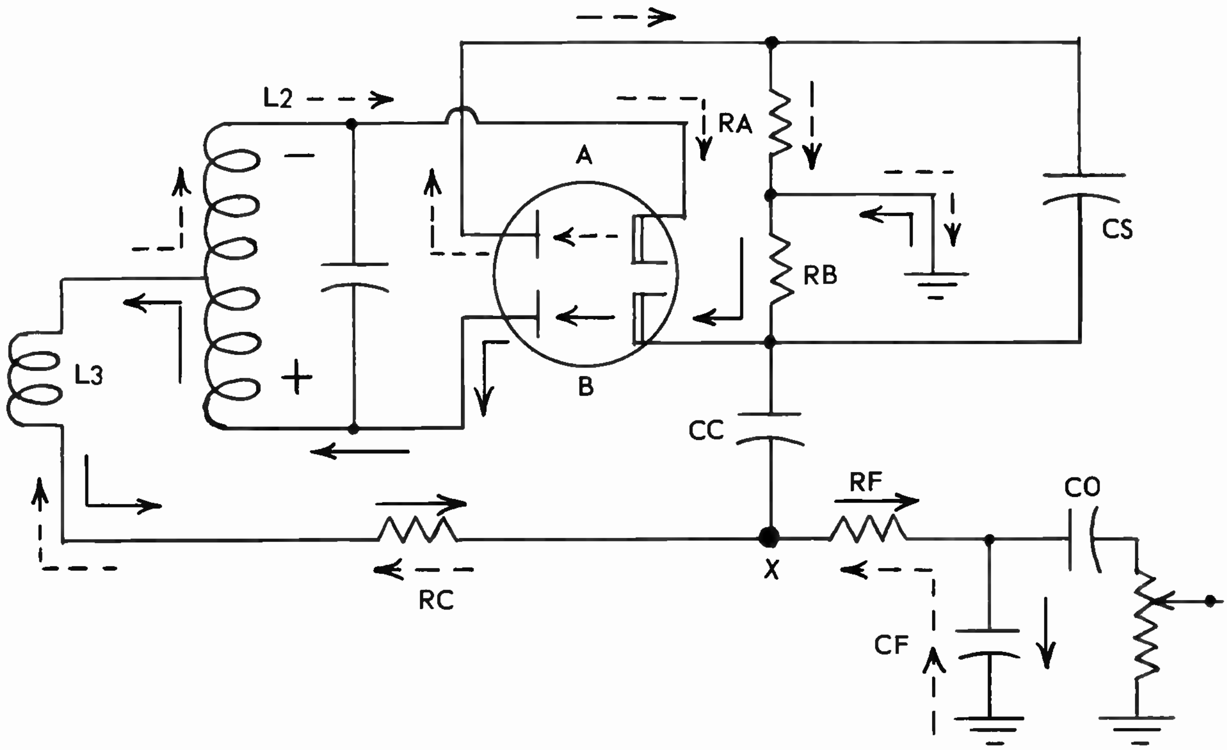


Fig. 36-16. Electron flows in the ratio detector having two resistors in the audio output circuit.

For diode *B* the flow is as follows: Diode plate. Lower half of secondary. *L2*. Winding *L3*. Resistor *Rc*. Point *X*. Resistor *Rf*. Capacitor *Cf* to ground. Through ground to *Rb*. Back to diode cathode.

When there is deviation to a lower frequency the current in diode *B* increases. Then, as shown by full-line arrows in Fig. 36-16, there is increase of electron flow through capacitor *Cf* to ground and the top of *Cf* accordingly becomes more negative, the top of *Cf* is connected through capacitor *Co* to the top of the volume control, so the top of the volume control resistor and its slider become more negative with reference to ground. Thus we have an audio signal voltage going negative when deviation is to a lower frequency.

Deviation to a higher frequency causes an increase of current in diode *A*. Along the path of the broken-line arrows in Fig. 36-16 there is increased electron flow. This increased flow passes from ground upward through capacitor *Cf*, making the top of this capacitor more positive. This increasingly positive voltage appears at the top of the volume control resistor, making the top of this resistor and its slider more positive with reference to ground. Here we have the audio signal going positive when there is deviation to a higher frequency. The audio signal goes alternately positive and negative as the deviation swings above and below the center frequency.

There are numerous other variations of the ratio detector circuit. Differences between them are principally in the means by which the alternating audio voltage is taken from the output side or load side of the diodes. Fortunately, these differences make little or no alteration in the methods of aligning a ratio detector. The

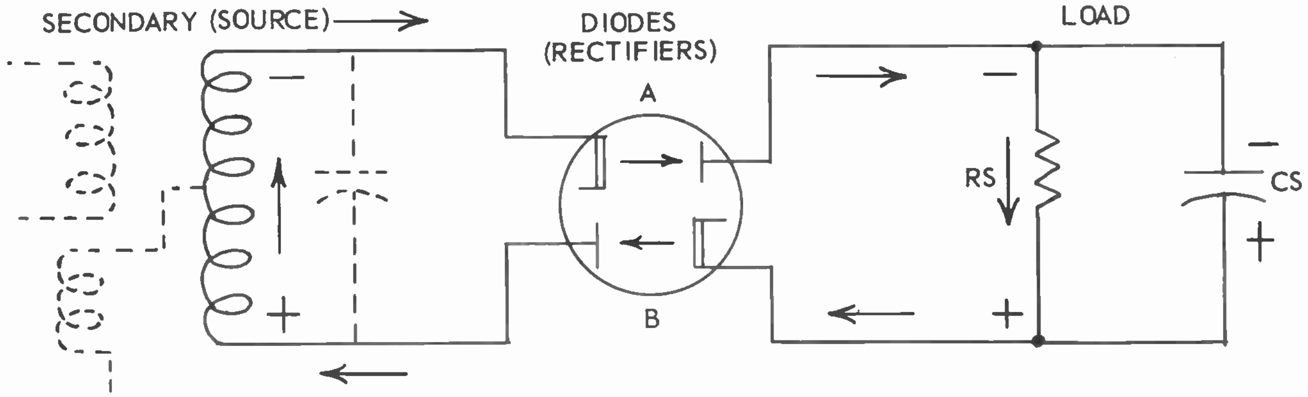


Fig. 36-17. Charging circuit for the large capacitor of the ratio detector.

point which we have identified as X is the one at which we make most of our alignment measurements for detector adjustments, and this point is quite easily located through its connection to the volume control no matter how the remainder of the detector circuit may be arranged.

REMOVAL OF AMPLITUDE MODULATION. On the output side or load side of ratio detector circuits are one or more resistors totaling 12,000 to 40,000 ohms, connected from the plate of one diode to the cathode of the other diode. In parallel with the resistor or resistors is an electrolytic capacitor or sometimes two such capacitors having total capacitance of one to twenty or more microfarads. The single resistor is marked *Rs* in Figs. 36-9 and 36-10. Its place is taken by the two resistors *Ra* and *Rb* in Figs. 36-15 and 36-16. The paralleled capacitor is marked *Cs* in all the diagrams.

As shown by Fig. 36-17, resistor *Rs* is electrically in series with the two diodes and the secondary of the transformer. The source of voltage for this series circuit is the emf induced in the secondary. The voltage is rectified by the two diodes, with the result that there is one-way electron flow in the path shown by arrows. This flow is from the negative upper end of resistor *Rs* to the positive lower end.

The average rate of electron flow and the average potential difference across resistor *Rs* will depend on the amplitude of the i-f voltage, or on strength and amplitude of the received signal. A stronger received signal will increase the potential difference across *Rs*, and weaker signal will decrease this potential difference.

Because capacitor *Cs* is in parallel with resistor *Rs*, or resistors *Ra* and *Rb*, the capacitor voltage must be the same as the average voltage across the resistance. Since the capacitance is very large the time constant of capacitance and resistance is proportionately long. The time constant is longer than the period of the lowest audio frequency and far longer than the period of the intermediate frequency. Consequently, the charge of the capacitor and the voltage across capacitance and resistance cannot vary to any extent either at intermediate frequency or at audio frequency. This voltage can vary only when there is a change of average signal strength, or when there is change of carrier amplitude and i-f amplitude continuing long enough to alter the charge of capacitor *Cs*.

Static interference causes voltage pulses at audio frequencies. The resulting momentary changes of current in the detector output are absorbed by the large capacitance at *Cs* with no appreciable change of capacitor charge or voltage. Any other amplitude modulation at audio frequencies is similarly absorbed. This is why the audio output of the ratio detector does not vary with nor respond to amplitude modulation.

Frequency deviation drops the voltage on one of the small capacitors (Ca or Cb) in Fig. 36-10 while raising it on the other small capacitor. Deviation similarly drops the voltage across one of the resistors *(Ra* or *Rb)* in Fig. 36-16 while raising it on the other resistor. But the two small capacitors or the two resistors are in parallel with the large capacitor *Cs*, whose voltage remains constant unless there is change of signal strength. Therefore, the sum of the voltages across the two small capacitors or across the two resistors cannot change at audio frequency.

The separate voltages across the two small capacitors or across the two resistors may vary at audio frequency to any extent so long as their sum never is more or less than the constant voltage across capacitor *Cs*. It is variation between the voltages or the difference between voltages across the small capacitors or the two resistors that is responsible for the output voltage that varies at audio frequency.

If there is increase of strength in the received signal the sum of the voltages across the audio output capacitors or resistors will increase, along with increase of voltage across capacitor *Cs*. This permits greater voltages across each of the audio output capacitors or resistors, and there is a stronger alternating audio voltage at the volume control and at point *X* in the diagrams. A weaker received signal allows capacitor *Cs* to lose some of its charge. This reduces the sum of the audio output voltages and decreases the maximum voltage on each of the small capacitors or the resistors. Then there is weaker alternating audio voltage at point *X* and at the volume control.

So long as the received signal is of constant strength the only change at an audio frequency rate in the output or load circuit of the detector is in the ratio of one voltage to the other voltage on the two small capacitors or the two resistors. The sum voltage cannot vary at audio frequency, but the ratio of separate voltages can vary. This is the reason for calling this type of demodulator a ratio detector.

* * * * *

We have spent a great deal of time examining the action of a ratio detector, for the following very good reasons.

The ratio detector is used for sound reproduction in most f-m receivers and in most of the recently designed television receivers.

The basic principles of the ratio detector are used also in the discriminator circuit, which is employed for sound reproduction on practically all other f-m and television receivers.

The same basic principle of combining out-of-phase voltages is used in both discriminators and phase detectors which are employed for automatic control of sweep frequency in television receivers.

TELEVISION

LESSON NO. 37
DISCRIMINATORS AND LIMITERS

Training Your Mind...
C. T. I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind
C. T. I.
Perfecting Your Skill

COMMERCIAL
TRADES
INSTITUTE

Chicago, Ill.

LESSON NO. 37

DISCRIMINATORS AND LIMITERS

The discriminator type of demodulator for f-m reception is subject to numerous modifications, as is the ratio detector type. But again none of the modifications in general use make any change in fundamental operating principles, and the procedure for alignment is practically the same for all arrangements or circuits. 3 Fig. 37-1 shows a discriminator circuit widely used in f-m receivers and in the sound sections of many television sets.

5 The most noticeable difference between the discriminator and the ratio detector is the connection of the transformer secondary to both diode plates in the discriminator, rather than to the plate of one diode and the cathode of the other as in the ratio detector.

2 Another difference plainly apparent upon examination of service diagrams is in the value of the capacitor 4 connected across the output side or load side of the demodulator. This capacitor is marked *Cd* in Fig. 37-1. It was marked *Cs* in our ratio detector circuits. For the discriminator this capacitance is only 50 to 500 micro-microfarads, while in the ratio detector the value is one microfarad or more. Each of the load resistors *Ra* and *Rb* for the discriminator is of 100K to 500K resistance, whereas in the ratio detector using two of these resistors the value is only 6K to 20K each.

In the discriminator transformer the primary and secondary are both tuned to the intermediate frequency, just as in the ratio detector transformer. Again the secondary is center-tapped with a connection from this tap to the primary circuit. In Fig. 37-1 the primary voltage is brought to the secondary center tap through capacitor *Cc*. In some discriminator circuits the primary voltage is introduced from a third winding, as in the ratio detector transformers which we examined. There are ratio detector transformers in which coupling from primary to secondary center tap is through capacitance. With any type of coupling, the primary i-f voltage is brought to the secondary center tap.

In addition to the primary voltage introduced at the center tap of the discriminator transformer there is another secondary voltage induced by means of inductive coupling between the two windings. These two secondary voltages act in the discriminator transformer just as they act in the ratio detector transformer. When there is no deviation, and the applied frequency is the center frequency, the combined voltages applied to the two diodes are equal and the diode currents are equal. When there is deviation to a higher frequency there is increase of voltage and electron flow in diode *A*, and decrease in diode *B*. With deviation to lower frequency there is increase of voltage and current in diode *B*, and decrease in diode *A*. We learned all about this action when studying ratio detectors.

Electron flows in the two diodes and their connections are shown by Fig. 37-2. The upper diagram shows the flow in diode *A*, from the diode plate through the upper half of the transformer secondary, from the center tap through the common lead to the bottom of resistor *Ra*, upward through this resistor, and back to the

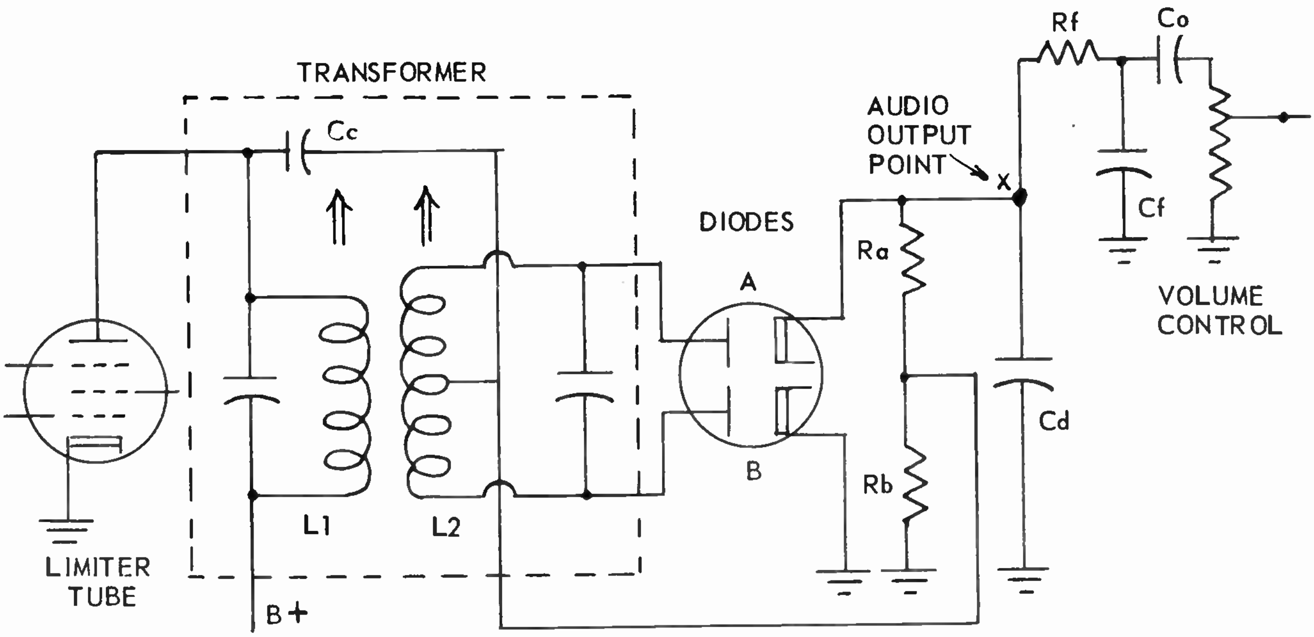


Fig. 37-1. Typical discriminator circuit for demodulation of f-m signals.

cathode of diode *A*. With this direction of electron flow the top of resistor *Ra* is positive with reference to the bottom. This flow occurs during half-cycles of i-f voltage that make the top of the secondary positive with reference to the bottom and the center tap, for only during these half-cycles is the plate of diode *A* positive with reference to its cathode. There is no flow in diode *B* because during these half-cycles its plate is made negative with reference to its cathode.

The lower diagram of Fig. 37-2 shows electron flow in diode *B* and its connections. This flow occurs only during i-f voltage half-cycles that make the bottom of the transformer positive with reference to its top and the center tap, since only then will the plate of diode *B* become positive with reference to its cathode. From the diode plate the electron flow is upward through the bottom half of the secondary, thence from the center tap through the common lead to the top of resistor *Rb*, downward through this resistor, and through ground back to the cathode of diode *B*. The top of resistor *Rb* is negative with reference to its bottom. Note that polarities of the two load resistors are opposite or are opposed to each other.

Now we shall assume that there is no deviation, that frequency applied to the transformer primary is the center frequency. With no deviation there will be equal currents in the two diodes and in the two load resistors, as represented by diagram 1 of Fig. 37-3. We shall assume further that the equal electron flows are accompanied by 5-volt potential differences across each of the load resistors. The equal voltages cancel each other, because of their opposite polarities, and the net voltage from the top of *Ra* to the bottom of *Rb* is zero. This is the voltage between the audio output point, *X*, and ground. Therefore, with no deviation, the audio output voltage is zero.

Diagram 2 of Fig. 37-3 shows what happens when deviation is to a higher frequency. Now there is more current in diode *A* and less in diode *B*, and more current in resistor *Ra* while there is less current in resistor *Rb*. We shall assume that these changes of resistor currents are accompanied by a potential difference

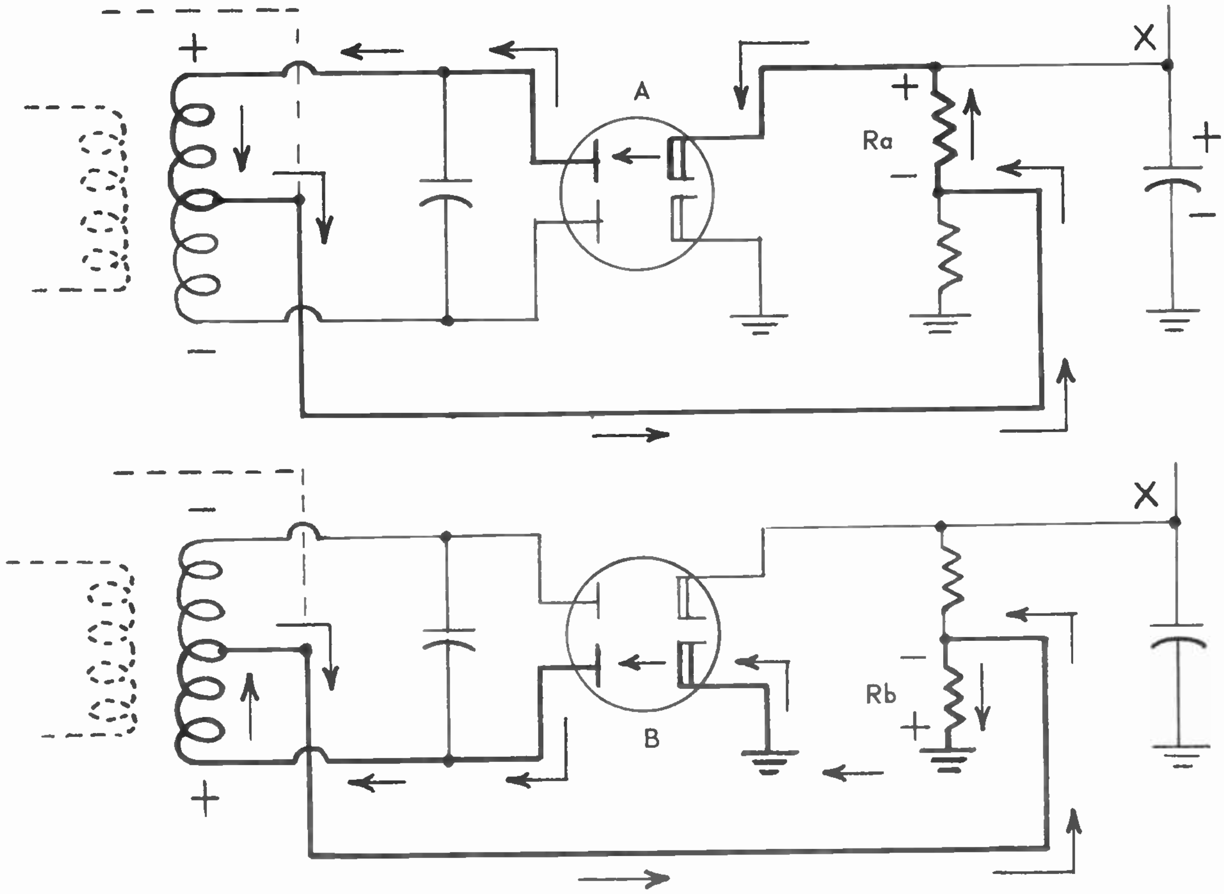


Fig. 37-2. Electron flows in the discriminator circuit during deviations to higher and lower frequencies.

increasing to 7 volts in *Ra* and by one decreasing to 3 volts in *Rb*. Now the greater voltage across *Ra* overcomes the smaller voltage across *Rb*, and leaves the difference of 4 volts as the net potential difference from top to bottom and between audio output *X* and ground. Thus the deviation to higher frequency has caused the audio output voltage to go positive.

Still greater deviation would cause still greater difference between currents in the diodes and the load resistors, and there would be still greater net positive voltage at the audio output point *X*. Any lesser deviation would cause less difference between currents in the diodes and the load resistors, and there would be a smaller net positive audio output voltage.

Diagram 3 of Fig. 37-3 shows what happens when deviation is to a frequency lower than the center frequency. Now there is more current in diode *B* and resistor *Rb* while there is less current in diode *A* and resistor *Ra*. If this deviation is equal in frequency change to the former high-frequency deviation the voltages across the two load resistors will be reversed. Again the difference is 4 volts, but now this net voltage is negative because the greater voltage is across resistor *Rb* whose upper end, toward the audio output point, is negative with reference to ground. With this lower deviation the audio output voltage swings to 4 volts negative.

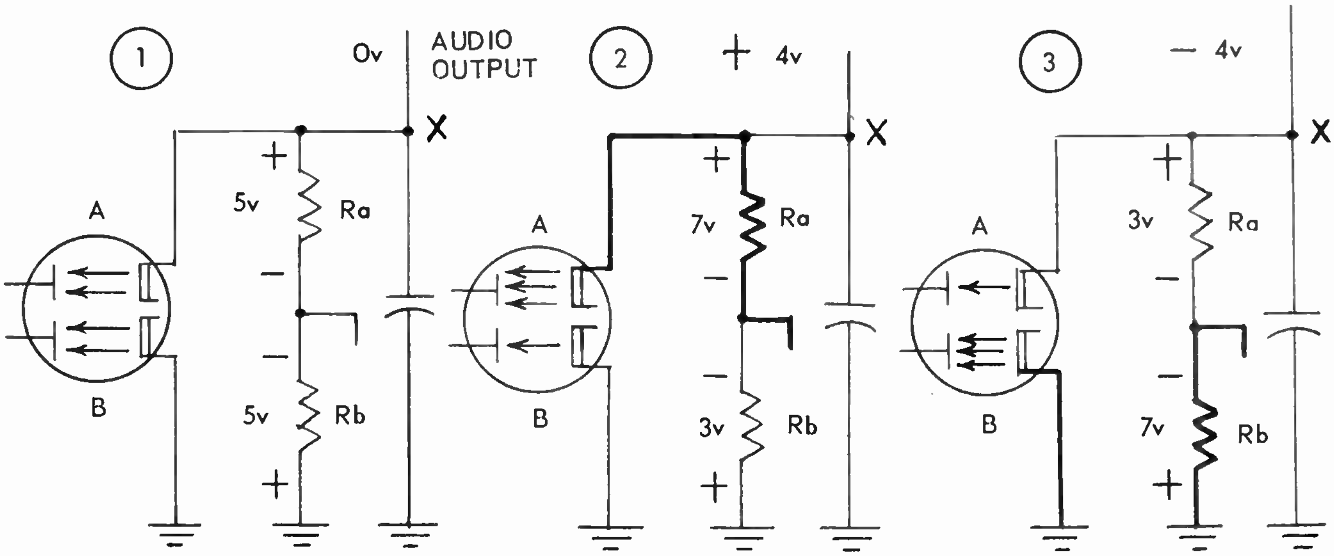


Fig. 37-3. Changes of voltage across load resistors and at audio output with the discriminator.

With frequency deviation swinging alternately higher and lower at the audio frequency of the signal modulation the audio-frequency output of the discriminator will alternately become positive and negative, with its amplitude proportional to the extent of frequency deviation. Thus the discriminator has accomplished its purpose of demodulating the f-m signal and recovering the audio signal.

Some of the variations found in discriminator circuits are illustrated by Fig. 37-4. The transformer is like that used for many ratio detectors, with inductively coupled primary and secondary and with primary voltage taken off through a third winding connected to the secondary center tap. On the output side of the diodes

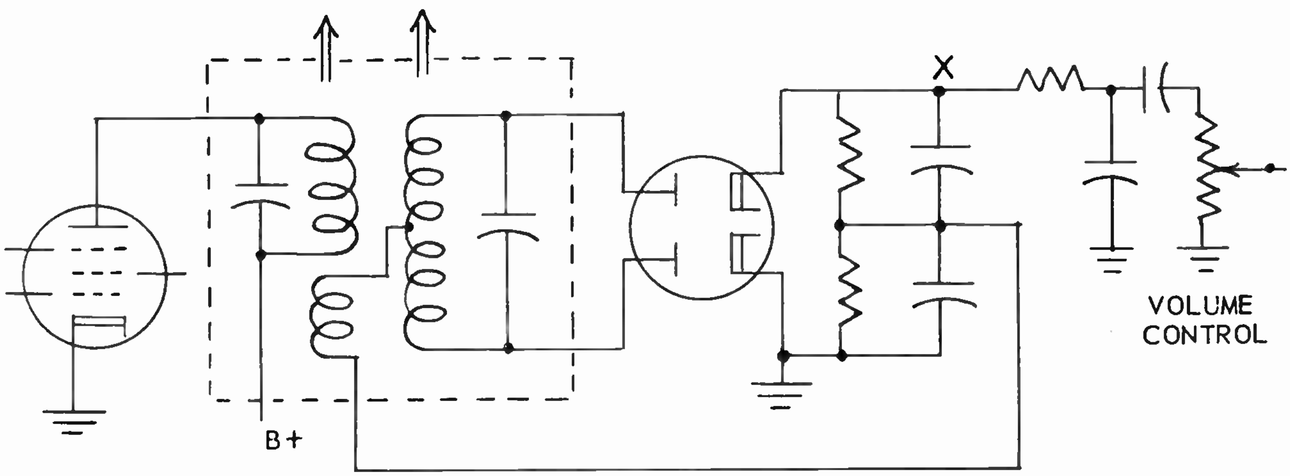


Fig. 37-4. Some of the modifications found in discriminator circuits.

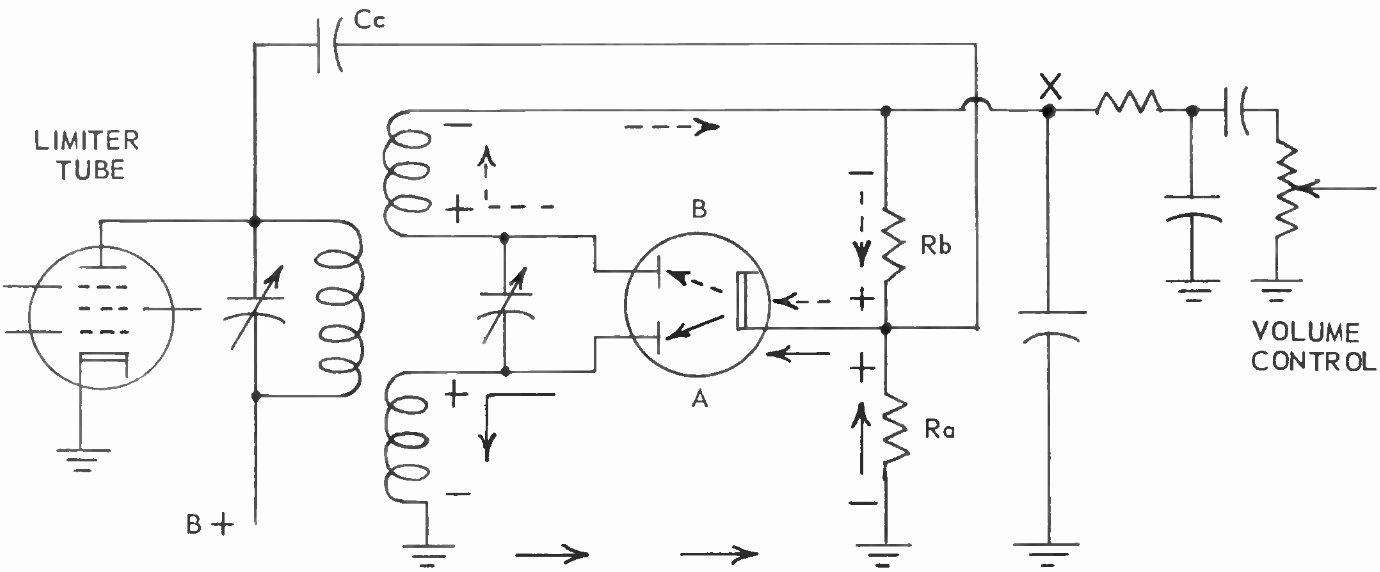


Fig. 37-5. Discriminator circuit in which the twin-diode has only one cathode.

are the two load resistors, but instead of having a single capacitor across both resistors there are separate capacitors across each resistor. The common line from the third winding of the transformer runs to a common junction point between the capacitors and the resistors. Demodulating action is the same as previously explained.

Fig. 37-5 shows a discriminator circuit in which the tube has two diode plates but only a single cathode. Full-line arrows show the path of electron flow when deviation is to a higher frequency. Broken-line arrows show the flow when deviation is to a lower frequency. In order to utilize only a single cathode it is necessary to have two separated secondary windings in the transformer, with both secondaries inductively coupled to the primary. Voltage from the primary is taken through capacitor *Cc* to the common cathode and thus is introduced into both secondary circuits, since the one cathode is in both these circuits. During frequency deviations there are unbalanced currents and unequal voltages across load resistors *Ra* and *Rb*. Polarities across these resistors oppose, and yield audio output voltage alternating positive and negative as deviation goes higher and lower than the center frequency.

In some of the older discriminator circuits there were two separately tuned secondary windings, with one made resonant at a frequency slightly higher than the center frequency and the other resonated at a frequency slightly lower than the center frequency. A higher deviation frequency would more nearly approach the resonant frequency of one secondary while getting farther from the resonant frequency of the other secondary. This would alter the relative values of inductive and capacitive reactances in the two secondaries, would cause shifting of phase of the induced secondary voltages and would increase the current in one diode while decreasing it in the other diode. Deviation to a lower frequency would, of course, have opposite effect on phase shift and on diode currents.

F-m demodulators may be constructed with crystal diodes instead of thermionic or hot-cathode tube diodes. Fig. 37-6 is the circuit diagram for a typical discriminator in which the diodes are of the germanium crystal type. Crystals for this service usually are a pair carefully matched for conductivity or resistance in both directions. A pair of matched germanium crystals is available under the type number 1N35. The use of crystals instead of tube diodes calls for no alterations in any other parts of the demodulator circuit.

PRE-EMPHASIS AND DE-EMPHASIS. In the signals radiated by f-m broadcasters the audio frequencies of modulation extend all the way from 50 to 15,000 cycles per second. This is a range of sound frequencies extending much higher than anything transmitted in the standard a-m broadcast channels, where maximum audio modulating frequency is limited to 5,000 cycles per second. The wide audio range of f-m reception permits improved tone quality or fidelity, provided the audio system of the receiver is capable of good reproduction.

When using an audio amplifier capable of reproducing high frequencies there is a tendency toward noise that results from slight irregularities of electron flow in tubes, resistors, and other circuit elements. These irregularities of electron flow are accompanied by minute voltages at the higher audio frequencies. These voltages, when amplified, cause hissing and "frying" noises from the speaker.

In addition, any amplitude modulation added to the carrier by static interference causes more noticeable sound from a wide range audio amplifier because voltages due to such interference extend through frequencies in the higher audio range. Although any of these amplitude modulation effects are greatly reduced by the ratio detector or the limiter-discriminator combination, it still is desirable to use additional means for reducing any noise voltages which may get through to the audio output of the demodulator circuits. This audio output point has been marked *X* in our circuit diagrams.

High-frequency noise is reduced without losing the desired higher frequencies of the audio modulation by a most ingenious method. The higher frequencies of the audio signal are intentionally over-amplified at the transmitter, then reduced to their normal values in the receiver. The process is called pre-emphasis at the transmitter and de-emphasis at the receiver.

Pre-emphasis at the transmitter consists of amplifying all audio frequencies above about 400 cycles per second at a gradually increasing rate, until voltages at the maximum of 15,000 cycles are made six or seven times as strong as those around 400 cycles per second. As a result, we have in the demodulated audio output of the ratio detector or discriminator an audio voltage whose amplitude increases with rise of audio frequency. The higher frequencies of the desired sound signal are over-emphasized.

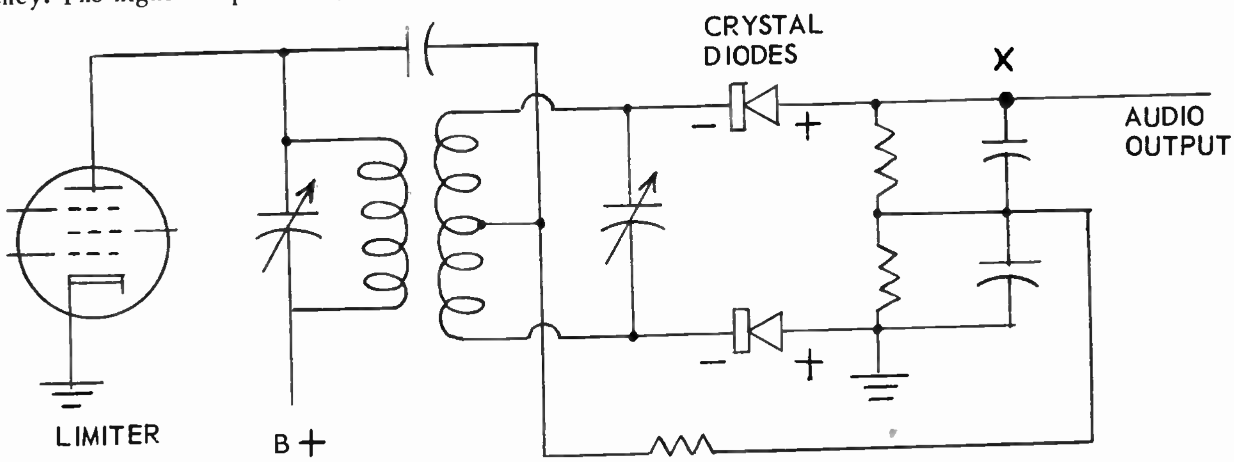
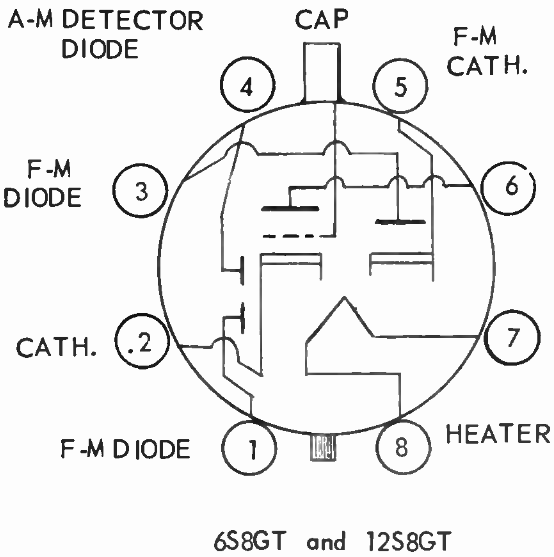


Fig. 37-6. Discriminator circuit employing crystal diodes.



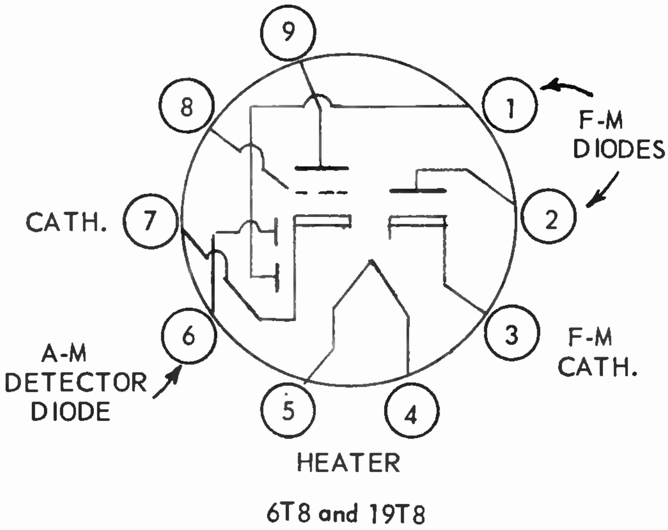


Fig. 37-7. Element connections of tubes having three diodes and a triode within one envelope.

In the output from the demodulator (point *X* in the diagrams) the high-frequency noise voltages will have whatever amplitude may have resulted from interference affecting the carrier and from irregularities of electron flow, so far as these effects have gotten through the demodulator. But amplitudes at the higher frequencies of the desired audio signal have been intentionally over-emphasized. The signal containing both noise and desired sound goes to the de-emphasis filter that always follows the demodulator in f-m receivers. In all our circuit diagrams the de-emphasis filter consists of the series resistor marked *Rf* and of a capacitor to ground, marked *Cf*.

This filter *is* designed to have its maximum attenuation at the highest audio frequencies, and gradually decreasing attenuation as audio frequencies become lower. Thus the de-emphasis filter compensates for over-emphasis at the transmitter. In bringing the higher audio-frequency voltages down to their normal amplitudes the noise voltages are reduced in the same proportions. We have then, at the output of the filter, all the desired sound voltages at their normal level, but have greatly attenuated noise voltages. This corrected audio signal voltage goes to the volume control and the audio amplifier section of the receiver.

For suitable de-emphasis the filter system should have a time constant of 75 to 100 microseconds, or 0.000075 to 0.000100 second. That is, the product of filter resistance in megohms and filter capacitance in microfarads should be something between 0.000075 and 0.000100. As an example, the filter resistance (Rf) might be 15,000 ohms or 0.015 megohm, and the filter capacitance 0.005 microfarad. The product of the two numbers is 0.000075, and the time constant is 75 microseconds.

COMBINATION DEMODULATOR-AMPLIFIER TUBES. There are a number of tubes which contain in a single envelope the diodes for both f-m and a-m demodulation and also a triode for audio amplification. These types are quite commonly used in combination fm-am receivers. Among the tubes of this class in general use are types 6S8-GT, 12S8-GT, 6T8, and 19T8. All these are called triple-diode high-mu triodes. All have the same elements and the same grouping of elements, but they have different base pin connections, and different heater ratings.

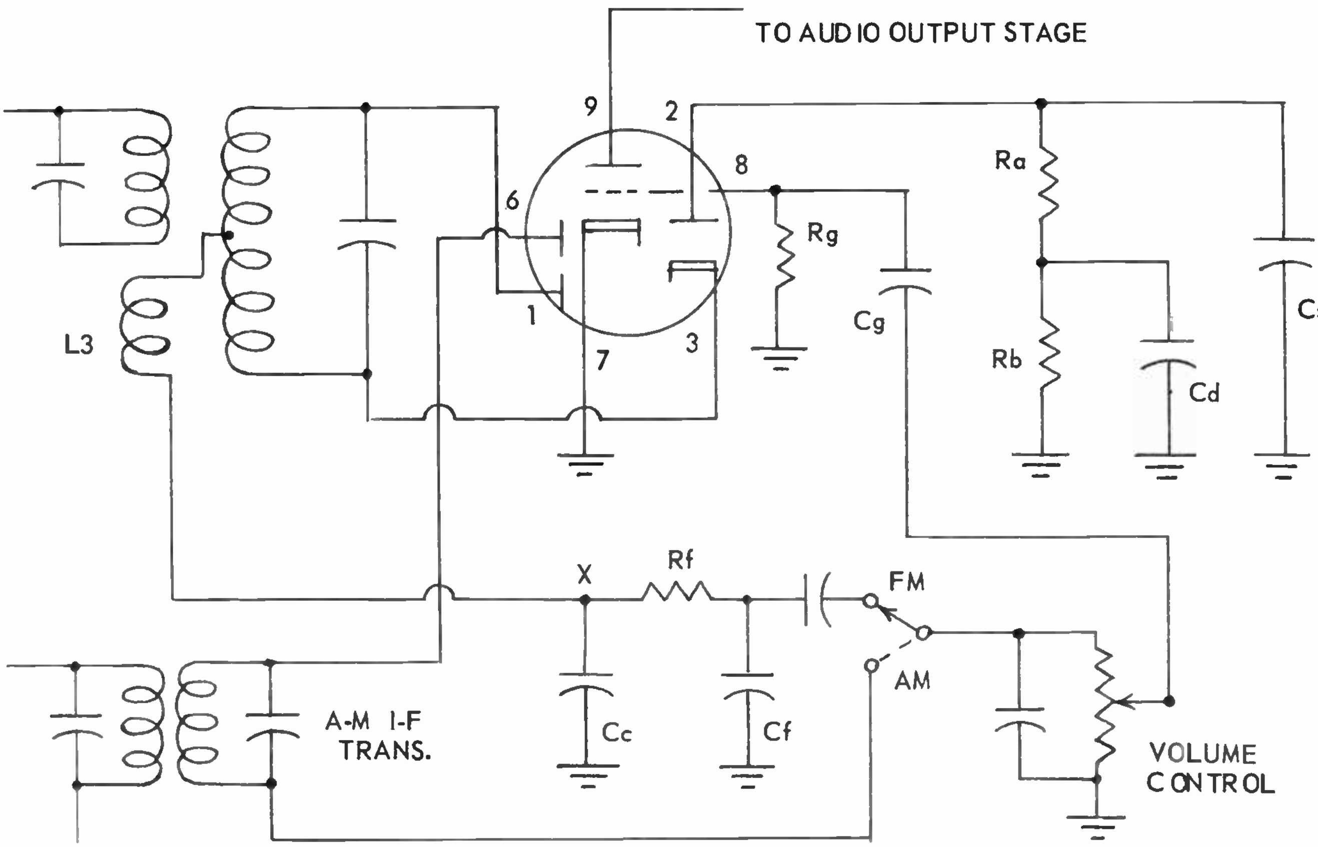


Fig. 37-8. Ratio detector using a single tube for f-m demodulation, a-m demodulation, and audio amplification.

Fig. 37-7 shows the internal elements and their connections when looking at the bottoms of the bases, or sockets, for the tubes mentioned. The 6S8-GT and 12S8-GT are alike except for having respective heater ratings of 6.3 volts at 0.3 ampère and of 12.6 volts at 0.15 ampere. Both have octal bases with GT-style glass bulbs and a top cap connection for the grid of the triode audio amplifier. The triode plate goes to pin 6.

On the same cathode that serves the triode are two diode plates, connected to pins 1 and 4. The diode plate on pin 4 and the cathode on pin 2 serve as the diode detector for a-m demodulation. On the same cathode, pin 2, is the first of the diode plates for f-m demodulation, the plate to pin 1. The second diode for f-m demodulation is entirely separate, consisting of the plate connected to pin 3 and the cathode connected to pin 5.

The 6T8 and 19T8 triple-diode high-mu triodes, whose symbols are at the right in Fig. 37-7, are alike except for respective heater ratings of 6.3 volts at 0.45 ampere and of 18.9 volts at 0.15 ampere. Both are miniature glass tubes with 9-pin bases that fit into the small button noval 9-pin socket. The internal elements and their grouping are like those for the 6S8 and 12S8 types, but element connections to the numbered base pins are entirely different. The connections are shown by the symbol.

In Fig. 37-8 are circuit connections for a ratio detector and first audio amplifier employing either the 6T8 or 19T8 triple-diode high-mu triode tube. If you trace the connections from the transformer secondary to diode plate 1 and cathode 3, and trace the connections to the load resistors (*Ra* and *Rb*) from cathodes on pins 2 and 7, you will find them essentially like connections in earlier diagrams for a ratio detector. The audio output from point *X* goes through the de-emphasis filter, resistor *Rf* and capacitor *Cf*, to the band switch leading to the volume control.

The diode plate on pin 6 connects to the high-side of the secondary in the last i-f transformer used for a-m reception. The low side of this winding goes to the band switch and to the volume control when the switch is in the position for a-m reception. The cathode (pin 7) for the a-m detector diode is grounded, so we have here the same diode detector circuit with which we became familiar in an earlier lesson. In order to permit grounding of the diode cathode the audio circuit of the ratio detector is completed through capacitors *Cc* and *Cd* to the point between the two load resistors. Note that the audio amplifier triode in the combination tube is biased by grid leak resistor *Rg* and capacitor *Cg*.

In Fig. 37-9 we have a discriminator for f-m demodulator and the first audio amplifier employing a 6S8-GT or 12S8-GT tube. There is nothing new about the connections for the discriminator. One end of the transformer secondary connects to diode plate 3, for which the cathode is connected to pin 5. The other end of the secondary connects to diode plate 1, whose cathode is connected to pin 2. The cathode on pin 5 is connected to the outer end of load resistor *Ra*, and from point *X* through the de-emphasis filter (*Rf* and *Cf*) to the f-m position of the band switch. This side of the circuit is completed through the volume control to the cap for the grid of the triode audio amplifier. Capacitor *Cg* and the resistor *Rg* provide grid lead bias for the triode amplifier.

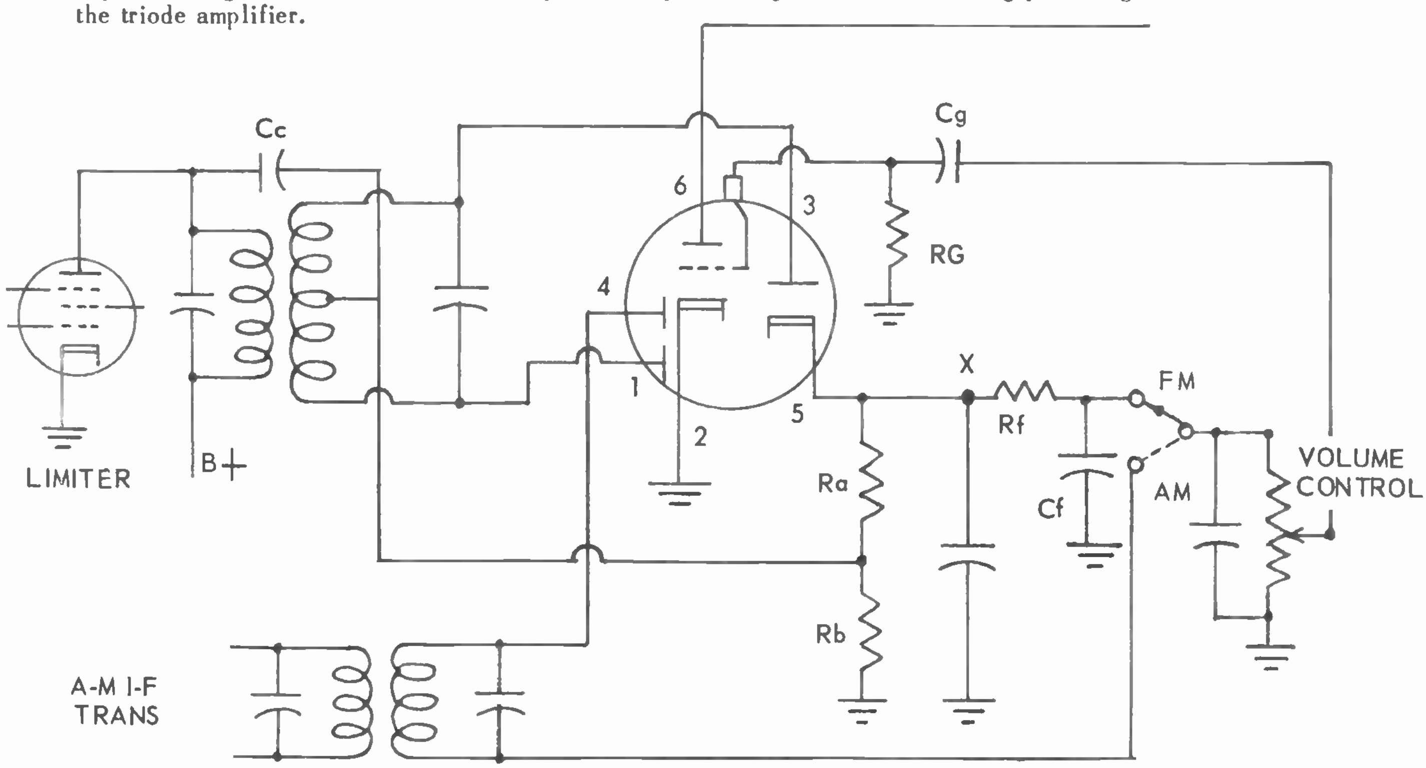


Fig. 37-9. Discriminator circuit with a combination tube for f-m and a-m demodulation and for audio amplification.

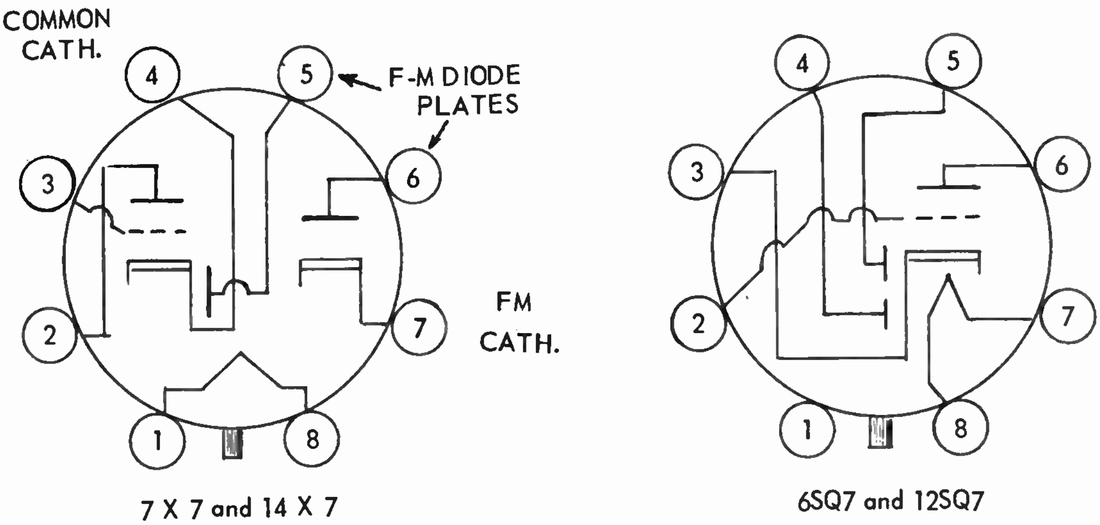


Fig. 37-10. A double-diode triode tube used in f-m receivers (left) and the type usually found in a-m receivers (right).

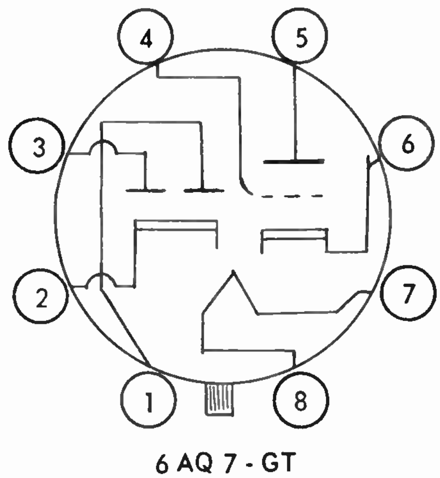
The diode detector for a-m demodulation consists of the diode plate on pin 4 and the cathode to pin 2, which is grounded. The diode plate goes to the secondary winding of the last i-f transformer for a-m reception, and the other end of this secondary is connected through the band switch to the top of the volume control. Grounding of cathode 2 not only completes the a-m detector circuit but also completes the discriminator load circuit through ground to the bottom of resistor *Rb*.

Were we to use a 6T8 or 19T8 tube in this discriminator circuit the only change in the diagram would be alteration of the pin numbering on the tube symbol. Similarly, using a 6S8-GT or 12S8-GT tube in the ratio detector circuit of Fig. 37-8 would require, so far as the circuit diagram is concerned, only a change of the pin numbering on the tube symbol.

Another type of tube that may be used as a combination f-m demodulator and first audio amplifier is the 7X7 and 14X7 whose symbol is shown at the left in Fig. 37-10. This is a lock-in tube with an 8-pin base and glass bulb. For one side of the demodulator circuit, either ratio detector or discriminator, the diode element consists of the plate on pin 6 and the cathode on pin 7. For the other side of the demodulator circuit there is diode plate 5 operating with the cathode on pin 4. This latter cathode is used also for the triode audio amplifier, whose grid is on pin 3 and whose plate is on pin 2.

Any tube having two diode plates operating with separate cathodes could be used in circuits for ratio detectors or discriminators where independent cathodes are required. For such circuits it is not possible to use the ordinary duodiode-triode tube like the 6SQ7 and 12SQ7 type shown at the right in Fig. 37-10. This is the tube, or the type of tube, most often employed as a diode detector for a-m demodulation in combination with the first triode audio amplifier. Here the two diode plates are on the same cathode. Both diode plates may be tied together to operate as a detector, or one of them may be used for detector service and the other for rectification of automatic volume control voltage.

A twin-diode high-mu triode having both diode plates working from the same cathode may be used in the discriminator circuit illustrated in Fig. 37-5. Tubes designed especially for this kind of service are shown by means of their symbols and base pin connections in Fig. 37-11. One is the type 6AQ7-GT with a standard octal base and glass bulb of the GT style. The other is the type 7K7, which is a lock-in tube with the lock-in style of 8-pin base and a glass bulb.



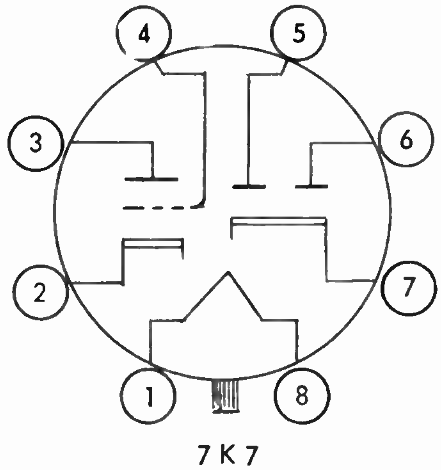


Fig. 37-11. Twin-diode triode tubes used in discriminator circuits having both diodes operated from a single cathode.

Both these tubes have a twin diode section operating with one of the cathodes as the f-m demodulator, and have a triode section with separate cathode for use as the first audio amplifier. There is no special provision for an a-m diode detector. The tubes of this style are employed chiefly in television sound sections where there is no a-m sound signal, but only the regular television sound which is handled by frequency modulation. These particular tubes are suitable also for straight f-m sound receivers in which there is no a-m band requiring a separate diode detector.

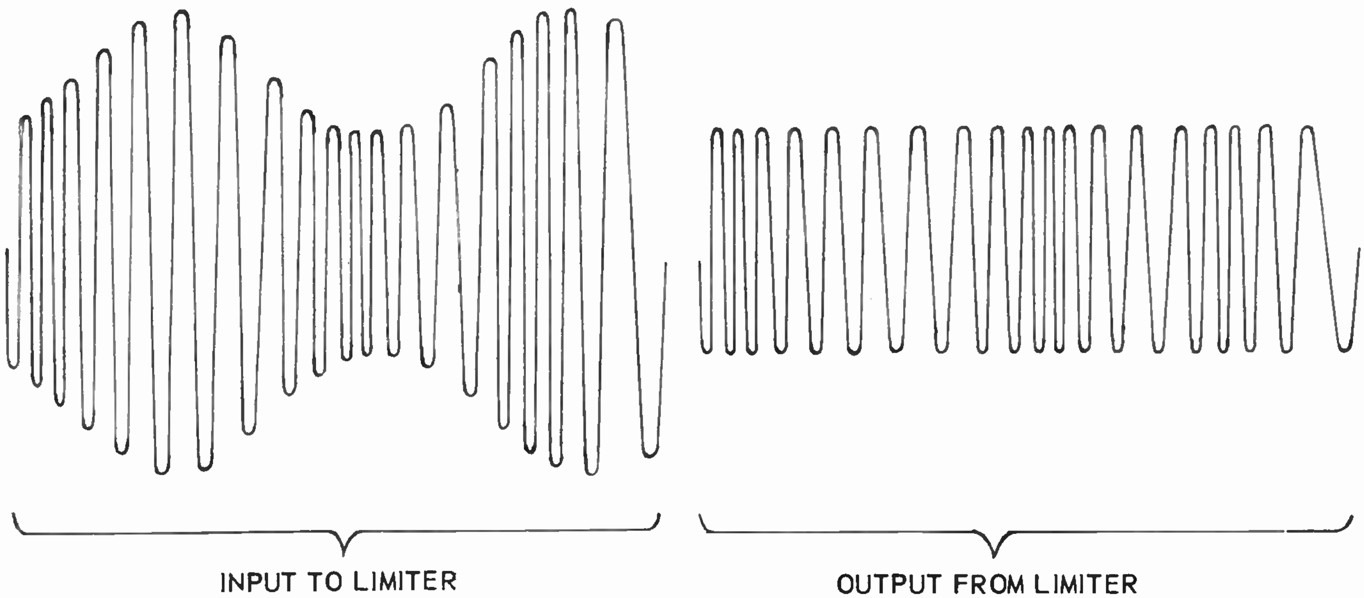


Fig. 37-12. How the limiter acts on a signal having amplitude modulation.

(4) *LIMITERS.* During our discussion of f-m demodulators it has been mentioned that a limiter stage is used ahead of a discriminator in order that the i-f voltage fed to the discriminator transformer may be free from amplitude modulation. We require this condition because the audio output of a discriminator will vary in accordance with any amplitude modulation that may be present while the applied signal deviates from the center frequency, or when the transformer windings are not tuned precisely to the center frequency. Then currents and voltages in the discriminator load resistors will be unbalanced by the changes of amplitude occuring at audio frequency, as well as by deviations of frequency, and the amplitude modulation will be demodulated to cause noise. There is nothing on the output side of the discriminator to absorb the changes due to amplitude modulation.

It is only while the i-f voltage is at the center frequency for which the transformer windings are resonant that the diode currents will be equal. Then, no matter how great may be the variations of amplitude in the applied voltage, there will be zero audio output.

The limiter stage might be omitted from a receiver using a discriminator. The f-m signals still would be demodulated and changed to audible sounds, but there might be an excessive amount of noise accompanying the desired sound. Although a limiter is not essential with a ratio detector, a limiter stage sometimes is used ahead of this type of demodulator to further improve the reproduction.

The purpose of a limiter stage is illustrated by Fig. 37-12. At the left is represented an input voltage which is frequency modulated and which also varies in amplitude. The limiter stage is intended to remove the peaks of amplitude, reduce the entire signal to a uniform amplitude, and leave only the frequency modulation. Such an output from the limiter is represented at the right. Although the action is not always so perfect as pictured here, the amplitude modulation should be reduced to no more than three to ten per cent of its original amount.

ACTION OF THE LIMITER. Typical connections for limiter stages are illustrated in Fig. 37-13. The principal difference between limiter connections and those in i-f amplifying stages is that the limiter tube is biased by the grid-leak method, with capacitor *Cg* and resistor *Rg*. One position for these biasing ele-

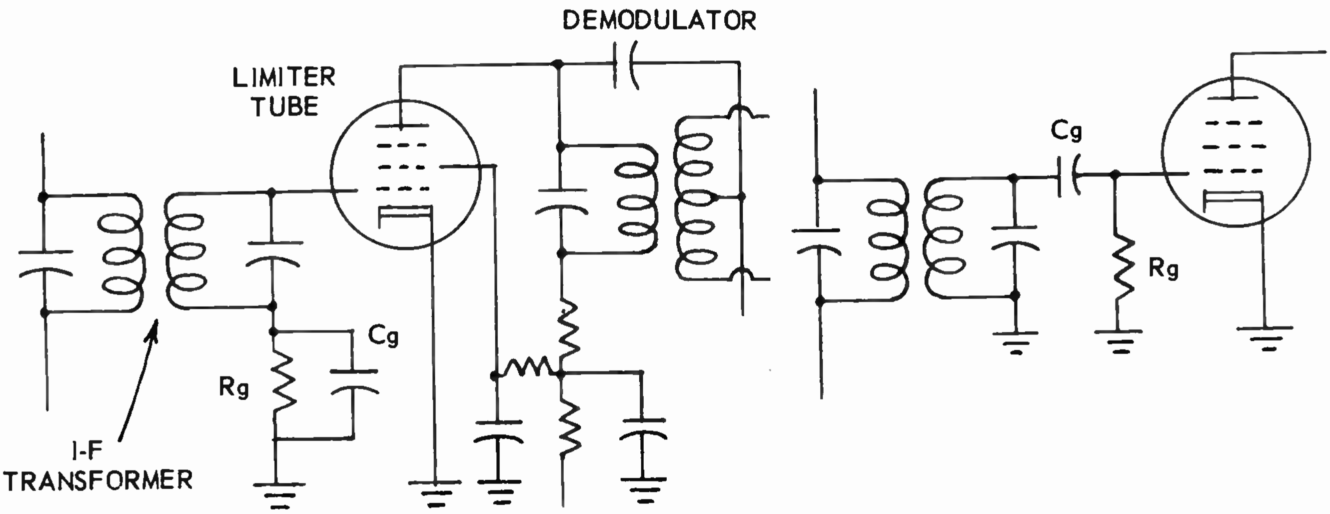


Fig. 37-13. Typical connections for limiter circuits.

(a) ments is shown in the left-hand diagram, another at the right. The action of biasing is the same with either arrangement. Another important difference is that the limiter tube is operated with plate and screen voltages much lower than those used for amplifiers. The limiter tube is a sharp-cutoff pentode.

So long as signal voltage coming to the limiter stage is of constant amplitude there is a negative grid bias of constant value. The bias voltage then depends on the average charge retained by grid capacitor *Cg* as this capacitor is repeatedly charged by each positive alternation of the incoming signal while continually discharging through grid leak resistor *Rg*. The bias will be sufficiently negative to hold plate current almost as low as the point of cutoff, were grid voltage only a little more negative it would cut off part of the plate current. This condition is represented at the left in Fig. 37-14.

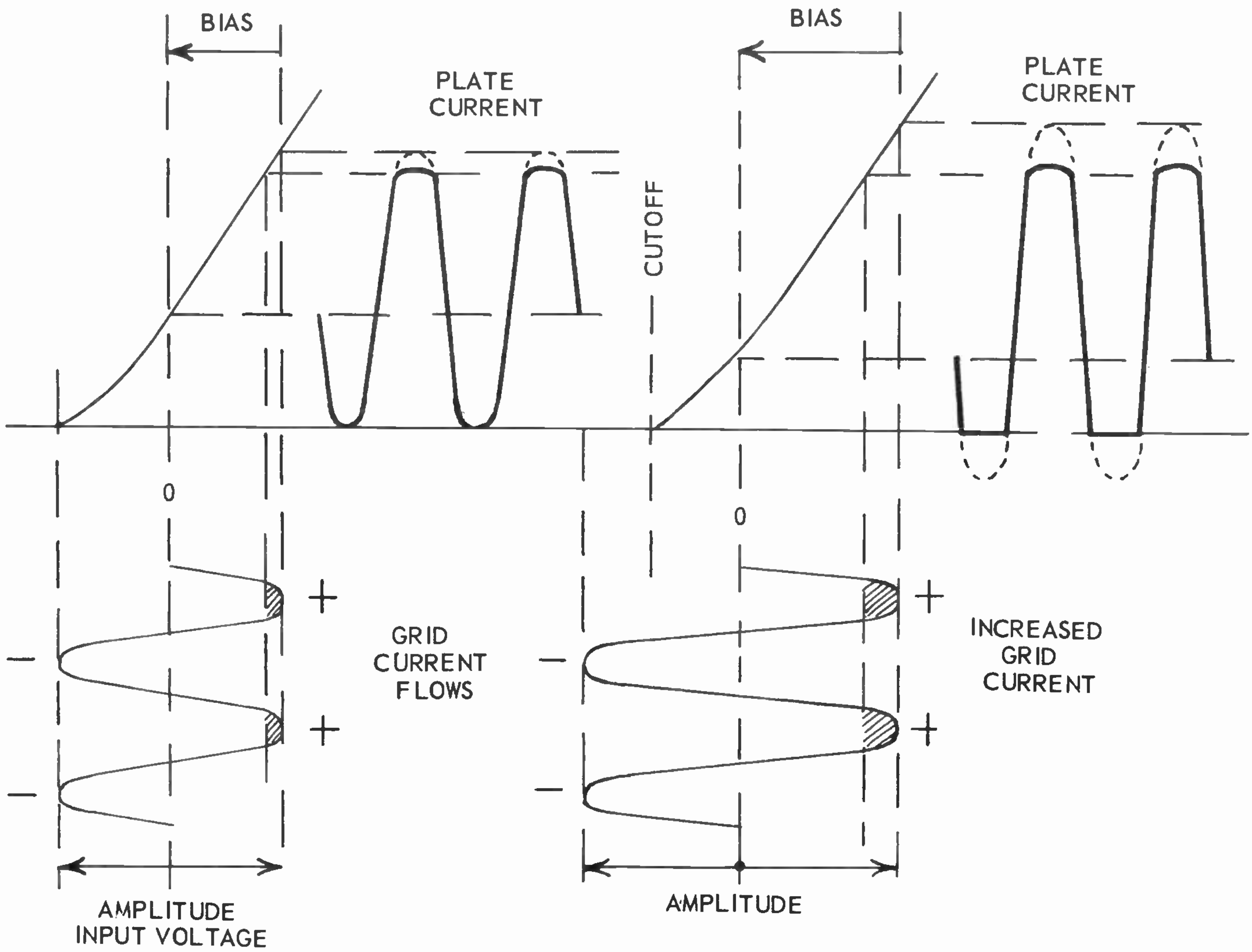


Fig. 37-14. How the limiter holds its plate current alternations at a constant amplitude.

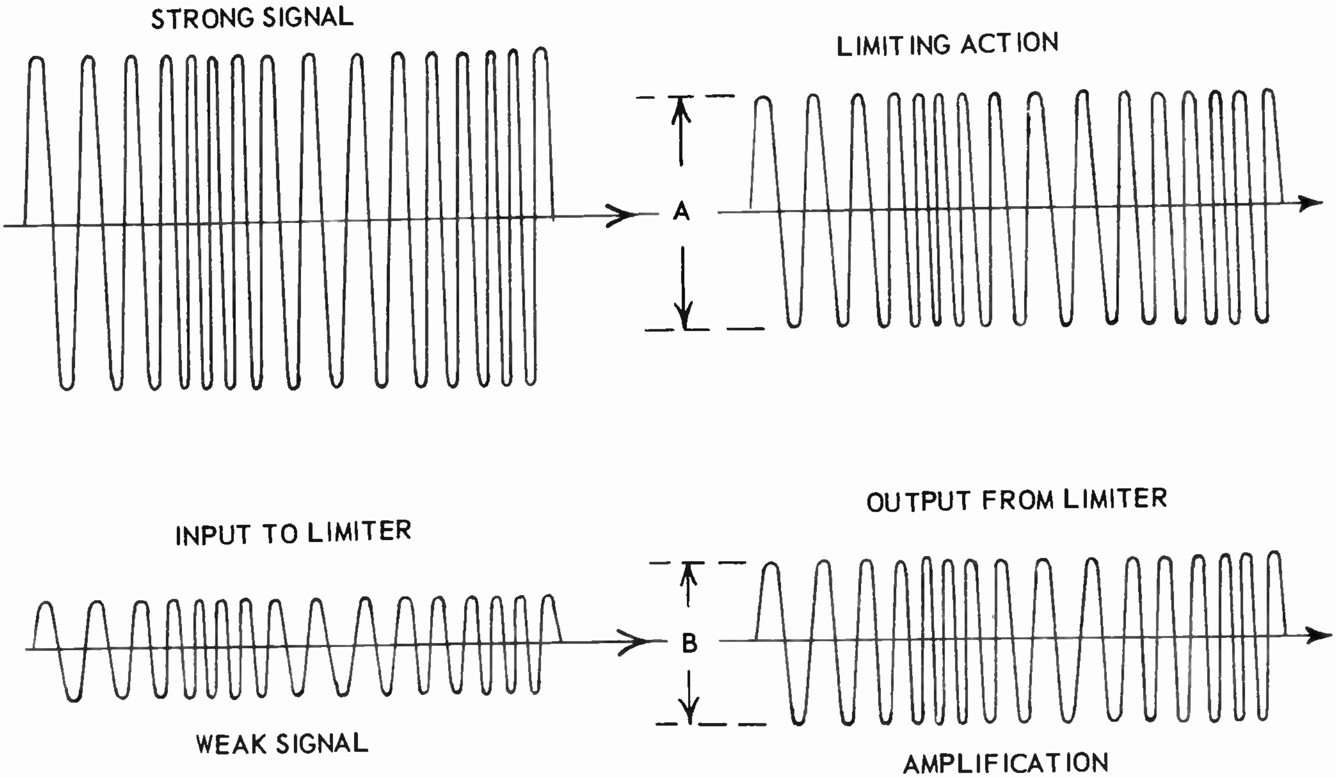


Fig. 37-15. Action of the limiter stage on strong and weak signal voltages.

At the most positive point of each positive alternation of input voltage there is a very small flow of grid current, just enough to restore the charge of capacitor *Cg* which has been lost through resistor *Rg* during the preceding cycle of voltage. During the brief periods during which the capacitor is recharged there is an increase of negative bias. These momentary increases of negative voltage on the grid prevent normal increase of plate current, and the positive peaks of current are flattened as shown on the graph.

Should there be an increase of amplitude in the input voltage the conditions will change as shown at the right in Fig. 37-14. The greater positive peaks of input voltage cause increased flows of grid current and of charging current into capacitor *Cg*. These increased pulses of grid current continue until the charge and voltage of the capacitor have increased to make the grid bias more negative than before. The more negative bias brings the zero line of the input voltage farther down on the curve of grid voltage and plate current. There is additional flattening of the positive peaks of plate current. The negative peaks of input voltage now go beyond the value which causes plate current cutoff, with the result that negative dips of plate current are flattened out along the line of zero current.

The net result of the increase of signal amplitude and the accompanying increase of negative grid bias is to limit the alternations of plate current to practically the same amplitude as existed with the original weaker signal voltage. Any greater increase of amplitude in the signal voltage would again cause increase of grid current pulses, a greater charge on the grid capacitor, a more negative bias, and the zero line of the signal would drop still lower on the characteristic curve. The plate current would show negligible increase of amplitude. This is the action of the limiter.

Limiting action is made effectively by using low voltages for the plate and screen circuits. With a tube that would be operated as an amplifier with voltages on the order of 135 to 150, limiter operation is at voltages between 40 and 60 for both the plate and the screen. With the resulting small current in the plate circuit only a small increase of negative grid bias is needed to reach the value for plate current cutoff.

High values of resistance are used between the B-power supply and limiter plate to drop the voltage. Small changes of plate current are accompanied by relatively large changes of voltage drop in this resistance. Consequently, during the portion of every signal cycle in which plate current is increasing there is a rapid decrease of voltage at the plate of the tube – because of increasing voltage drop in the load resistance. This decrease of plate voltage limits the increases of plate current and helps to cut off or flatten the positive peaks of current. No matter how positive the grid voltage might become, or how great might be its amplitude, there would come a point beyond which the plate current could show no further increase. Voltage remaining at the plate of the tube would have dropped too low to permit additional flow of current. This action may be called plate saturation.

LIMITER GAIN AND OUTPUT. While the limiter is performing its function there is severe distortion of waveform, as you may see from Fig. 37-14. The input voltages are shown as sine waves. The output, as shown at the right, is a succession of waves that have been flattened across both top and bottom. Instead of sine waves we have square waves.

Such waveform distortion would be very bad were we depending on changes of amplitude for carrying the desired audio signal. But in the f-m receiver we are depending on changes of frequency to carry the signal, and flattening or otherwise distorting the current waves does not alter their frequency nor does it alter the transmitted audio signal. It is true that a distorted waveform actually contains not only the original frequency but also many multiples (harmonics) of that frequency. But the demodulator transformer that follows the limiter has its windings tuned to the original frequency, which is the center frequency, and reactance in the transformer is so low to multiples of the center frequency that they have no effect.

At the top of Fig. 37-15 is illustrated the action of the limiter stage when input to this stage is a strong

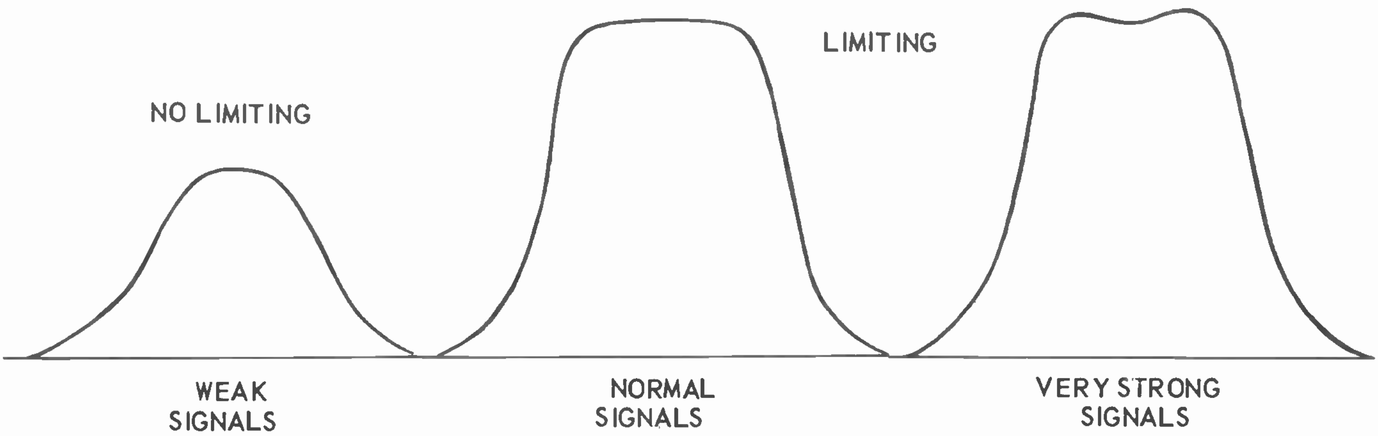


Fig. 37-16. Frequency responses before and after limiting occurs.

frequency-modulated voltage of constant amplitude. The limiter reduces the amplitude of this signal voltage just as it would reduce variations of amplitude which constitute amplitude modulation. The limiting action is the same as shown by Fig. 37-12. No matter how great may be the amplitude or strength of any incoming signal, within any reasonable values, the limiter will bring the amplitude down to the value determined by operating characteristics of the limiter stage.

Down below in Fig. 37-15 the input to the limiter stage is shown as a weak frequency-modulated voltage of constant amplitude. For this input the limiter acts as an amplifier. The output from the limiter is of greater amplitude than the input. But the limiter output for the weak incoming signal is not so great as for a strong signal. Output amplitude at *A*, when there is limiting action, is greater than output amplitude at *B*. For the weak signal there is no limiting. There is some amplification of the weak signal, but not a great deal, because with its low plate and screen voltages the limiter tube is not an efficient amplifier.

Had the weak signal contained amplitude modulation as well as frequency modulation the variations of amplitude would have been amplified by the limiter, and in the limiter output there would be amplified amplitude modulation. The limiter can perform its proper function of limiting, or getting rid of amplitude modulation, only when the input voltage to the limiter stage is strong enough to cause plate current cutoff and plate saturation as illustrated by Fig. 37-14. On incoming signals so weak that there is no limiting action, any amplitude modulation in these signals will be slightly amplified and passed on to the following demodulator stage.

During service work you often will look at the frequency response of a limiter, with the help of an oscilloscope. The frequency response shows relative voltage gains or amplifications at the center frequency and at frequencies above and below the center when there is deviation. Such responses are shown in Fig. 37-16. When you apply a weak signal from a signal generator there will be no limiting action, the peak of the response curve will rise and fall as you increase and decrease the voltage from the generator. When the input voltage is raised to the value that causes limiting the top of the response curve will flatten off. With very great input voltage the top of the curve may develop peaks, but its height will not be noticeably greater than when limiting commenced.

In order that the limiter may operate in the intended manner, and remove amplitude modulation, the gain in the i-f amplifying stages ahead of the limiter must bring signal strength up to the value that causes limiting. Always there will be some signals so weak that gain in the i-f amplifiers cannot bring the voltage up to the level for limiting. For such signals the sounds from the speaker always will contain noise, because there is no limiting to remove the noise pulses. How weak may be the received signals that are reproduced without noise will depend on how much gain there is in the i-f amplifiers, the converter or mixer, and the r-f stage if such a stage is used in the receiver.

The greater the total gain or amplification between the antenna and limiter the stronger will be the voltages applied to the limiter for all received signals, both weak and strong. Then the limiter may be operated with fairly high voltages for its plate and screen, and the limiter output will be proportionately greater for operating the following demodulator. When there is less total amplification ahead of the limiter it becomes necessary to reduce the plate and screen voltages in order that limiting action may take place with all signals which normally should be reproduced without noise. This, of course, reduces the output from the limiter and provides smaller voltages for operating the demodulator.

LIMITER TIME CONSTANTS. The time constant of the biasing resistor and capacitor in the limiter grid circuit determines how quickly or how slowly the bias voltage will follow any changes of amplitude in the applied signal voltage. As you know, the time constant in fractions of a second is equal to the product of resistance in megohms multiplied by capacitance in microfarads. Limiter time constants in general use range from 2 to 5 microseconds, or from about 0.000002 to 0.000005 second. These constants result from using such combinations as 0.047 megohm with 0.00005 mf (47K resistance and 50 mmf) or 0.047 megohm with 0.0001 mf, or 0.1 megohm with 0.000025 mf. These latter two combinations are equal respectively to 47K resistance with 100 mmf, and to 100K resistance with 25 mmf.

In order to maintain a reasonably steady average grid bias the time constant must be much longer than the intervals between successive i-f voltage cycles. With intermediate frequency of 10.7 mc this interval or period is somewhat less than 0.1 microsecond. Then a grid bias time constant of 2 microseconds extends over about 20 i-f signal cycles, and a constant of 5 microseconds extends over about 50 signal cycles. The longer the time constant the greater will be the range of signal voltages or amplitudes at which there is effective limiting.

On the other hand, the limiter time constant must be short enough to permit the bias to change fast enough to cut off sudden and brief noise pulses, such as are caused by sparking electrical devices, automotive ignition, and similar things. There will be effective cutoff or limiting, so far as speaker output is concerned, with any time constant considerably shorter than the period of the highest frequency to which human ears respond. If this high limit is taken as 20,000 cycles per second the interval or period is 50 microseconds, so when we get down to time constants around 2 to 5 microseconds there is effective cutoff of all ordinary noise.

In receivers which have a large number of tubes, with the intention of obtaining best possible results regardless of cost, we often find two limiter stages in cascade, one after the other. Usually the first limiter stage is operated with a short time constant for suppressing high-frequency noise pulses, while the second stage is operated with a longer time constant to permit handling a wide range of signal strengths.

TELEVISION

LESSON NO. 38

I-F AMPLIFIERS AND TUNERS FOR F-M

"Training Your Mind"

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind

C.T.I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON NO. 38

I-F AMPLIFIERS AND TUNERS FOR F-M

Now that we are familiar with demodulators or detectors for both f-m and a-m reception we shall examine some intermediate-frequency amplifiers and tuners which bring signals to these demodulators. We are chiefly interested in the f-m portions of the tuners and i-f amplifiers, but since nearly all receivers that cover the f-m broadcast band cover also the standard a-m band we shall use combination fm-am circuits for most of our examples.

Fig. 38-1 is a circuit diagram for a combination i-f amplifier system which is fairly representative in this field. The diagram extends from the converter at the left to the demodulators at the right. There are two i-f transformer units, one f-m demodulator transformer, and two i-f amplifier tubes. In each of the i-f transformer units there are two complete transformers. At the top are a primary and secondary tuned to 10.7 mc for f-m operation. Down below are another primary and secondary tuned to 455 kc, the intermediate frequency for a-m operation.

The lead from the converter plate goes to one section of the band switch. This switch, which has positions for f-m reception and for a-m reception, is shown in its f-m position by a full line and in the a-m position by a broken line. For f-m reception the converter plate is connected to the top of the f-m portion of the first i-f transformer. When the band switch is changed to its a-m position the converter plate is connected to the top of the primary of the a-m portion of the transformer unit.

In tuner circuits, which precede this i-f amplifier, are other sections of the band switch. One of these other sections of the band switch connects the r-f amplifier tube to either the f-m antenna and its tuned coupling or else to the a-m antenna and its tuned coupling. Another switch section not shown in the present diagram connects the plate of the r-f amplifier to the mixer grid of the converter through a coupling which is tuned for f-m reception or else through another coupling tuned for a-m reception. Consequently, when the band switch of our diagram is turned to its f-m position the converter plate will be delivering intermediate frequencies centered around 10.7 mc, which is the center frequency for f-m reception. When this switch is in its a-m position the converter plate will be delivering intermediate frequencies based on 455 kc.

The secondaries of both parts of the first i-f transformer are in series with each other, with the only connection to the grid of the first i-f amplifier tube taken from the top of the f-m secondary. When an i-f signal voltage centered at 10.7 mc is being delivered from the converter plate to the f-m primary winding there are induced in the f-m secondary the signal voltages of corresponding frequencies, because this secondary is tuned to 10.7 mc. The f-m secondary circuit is completed through the lower a-m secondary, but because this a-m secondary is tuned to 455 kc its reactance or impedance to currents at and around 10.7 mc is practically zero. The a-m secondary capacitor acts as though it were little more than an ordinary short in series with the f-m secondary.

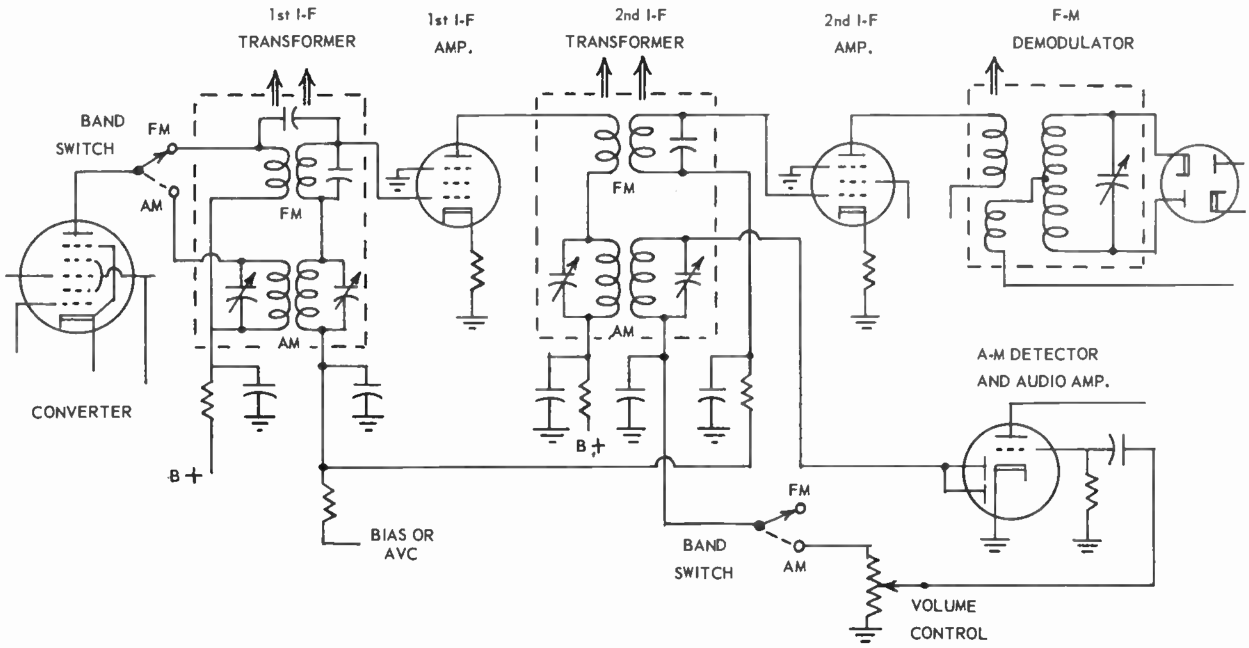


Fig. 38-1. Combination fm-am intermediate-frequency amplifier feeding a ratio detector and a diode detector.

When the band switch is set to allow the converter plate to deliver to the a-m primary a signal voltage based on 455 kc, for a-m reception, voltages at and near this frequency are induced in the a-m secondary, because this secondary is tuned to 455 kc. The f-m secondary, up above, now has negligible reactance or impedance to the currents around 455 kc, because this secondary is tuned to 10.7 mc. Then the f-m secondary winding acts practically like a plain conductor so far as the 455-kc signals are concerned, and these a-m signals pass freely through the f-m secondary to the grid of the first i-f amplifier tube.

We have seen how the position of the band switch determines whether f-m signals or a-m signals are applied to the grid of the first i-f amplifier tube. At the plate of this amplifier will appear either f-m signal voltages or else a-m signal voltages, according to the position of the band switch.

The plate of the first i-f amplifier tube is connected to the primary windings of the second i-f transformer. These primary windings are in series with each other. The upper primary with interelectrode and stray capacitances is tuned to 10.7 mc, for f-m reception, and the lower primary is tuned to 455 kc for a-m reception. The impedance of the a-m capacitance to signals at the high band, or the f-m winding at the low band is so low as to have no appreciable effect on the action of the tuned circuit at the selected frequency. Each tuned circuit responds only to frequencies for which it is tuned, and for frequencies in the other band, the untuned circuit acts essentially as a plain conductor.

Doubtless you will ask why we need no switching for the primaries of this second i-f transformer, yet we do need switching at the first i-f transformer? Switching is required in the plate circuit of the converter tube because in the output of the converter there otherwise might be voltages at sum frequencies and also at harmonic or multiple frequencies of the unused band. Currents and voltages at these frequencies would reach the following i-f amplifier tube and increase its load. When these effects are prevented by switching ahead

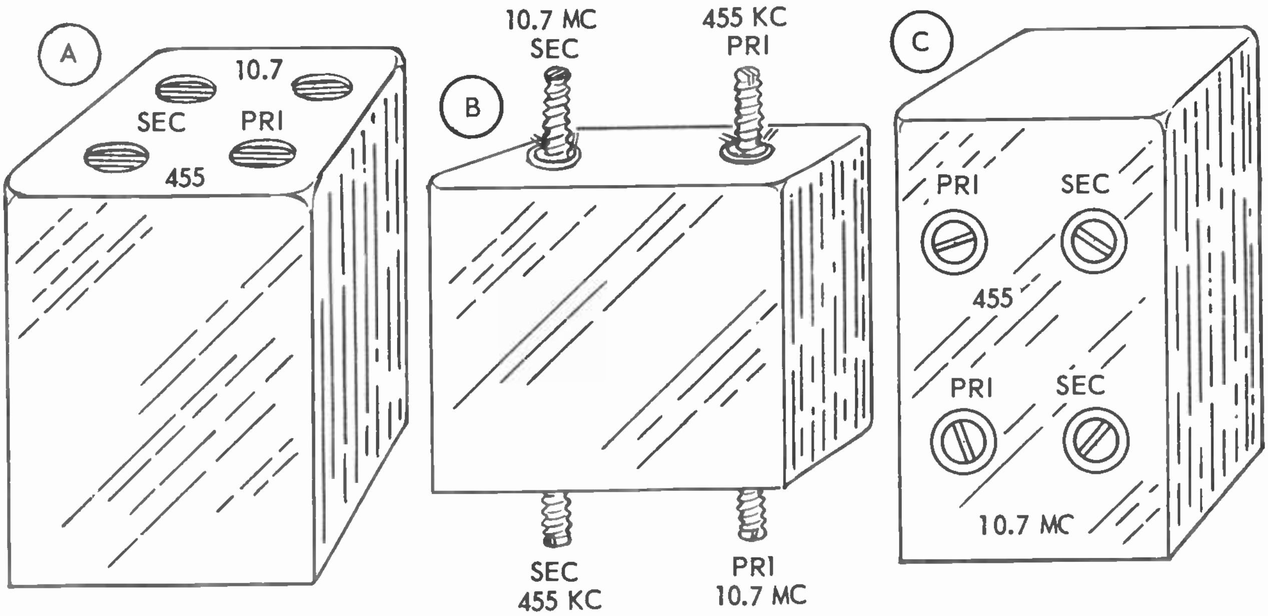


Fig. 38-2. Positions of alignment adjustments on i-f transformer units.

of the first i-f transformer there is no need for switching at any following transformer, since, with undesired voltages kept out of the first i-f amplifier they will not appear in following stages.

The transformers for both f-m and a-m reception usually, but not always, are inside of a single can or shielding enclosure which is mounted on top of the chassis, with terminals or lead wires coming out through the bottom of the can.

Alignment adjustments for primary and secondary transformer windings may all be of the movable core type, or all of the adjustable capacitor type, or both methods may be used in a single transformer unit as in the diagram of Fig. 38-1. At *A* in Fig. 38-2 all transformer tuning adjustments are reached from the top of the can. At *B* some adjustments are on top and others underneath, where they are reached from the under side of the chassis. At *C* all the adjustments are by means of screw heads extending through one side of the can. Usually there are markings for the frequencies which identify f-m and a-m adjustments, and often there are markings also for primary and secondary adjustments.

The two secondaries of the second i-f transformer of Fig. 38-1 are not connected in series with each other. The f-m secondary feeds the second i-f amplifier tube, which is used only for f-m reception. The output of this second i-f amplifier goes to the f-m demodulator transformer. The a-m secondary of this second transformer unit is connected to the plates of the a-m diode detector and to the volume control when the band switch is in its a-m position. When the band switch is in its f-m position the lower end of the a-m secondary is open circuited or disconnected, making this section of the transformer inoperative.

A somewhat different i-f amplifier system is shown by the circuit diagram of Fig. 38-3. Again there is switching at the mixer (or converter) plate for the primaries of the first i-f transformer. Now this switching short-circuits the a-m primary when the switch is in its f-m position (shown by the full line), while the f-m

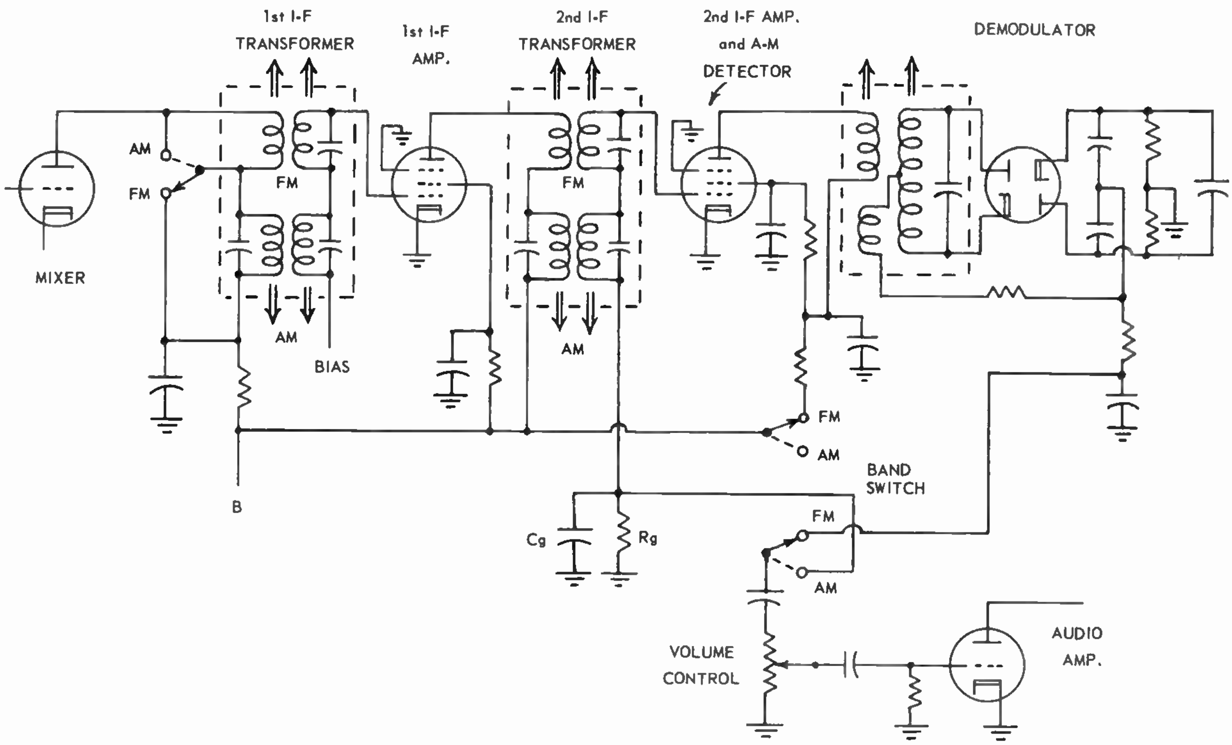


Fig. 38-3. An fm-am intermediate-frequency amplifier system operating into a pentode used as a diode for a-m detection.

primary is short circuited when the switch is in its a-m position (shown by a broken line). This switching leaves only one or the other of the primaries effectively in circuit at one time – depending on which band is being received.

Connections from the transformer secondaries to the grid of the first i-f amplifier tube, and connections from its plate to the primaries of the second i-f transformer, are the same as in Fig. 38-1. There is, however, an entirely different arrangement of secondary connections in this second i-f transformer. For reception in both hands the secondaries are in series with each other. The top of the f-m secondary is connected to the grid of the following pentode, which acts as an i-f amplifier and feeds its output signal to the f-m demodulator. The demodulator is shown as a ratio detector.

The f-m secondary circuit is completed through the a-m secondary, thence through capacitor *Cg* and resistor *Rg* to ground, and through ground back to the tube cathode. During f-m reception the output of the ratio detector goes through a section of the band switch to the volume control and from there to the grid of the audio amplifier.

When the band switch is shifted to its a-m position, B-voltage is disconnected from the plate and screen of the pentode that formerly acted as the second i-f amplifier. With plate and screen dead, only the grid and cathode of this tube remain active. The grid and cathode now act as the plate and cathode of a diode. This diode becomes the detector for a-m operation. The tube grid, now acting as a diode plate, is connected to the a-m secondary of the transformer through the negligible impedance of the f-m secondary. The lower end of the a-m secondary is connected through a section of the band switch to the volume control and audio amplifier tube.

In some receivers you will find a pentode f-m i-f amplifier transformed into a diode for a-m detection by disconnecting the screen of the pentode from its B-voltage. With no voltage on the screen there is no electron flow to the plate, and only the grid and cathode remain active. Receivers which employ the grid and cathode of a pentode as a diode appear, at first glance, to have no means for a-m detection.

① *IMAGE FREQUENCIES.* As has been mentioned several times, practically all sets designed in recent years use an f-m intermediate frequency of 10.7 mc. This standard intermediate frequency was chosen because it avoids reception of signals at image frequencies. You will recall that an image frequency is one that is as far above the r-f oscillator frequency as the received signal is below the oscillator frequency, or the image is a frequency that is higher than the received frequency by an amount equal to twice the intermediate frequency. Such frequency relations are illustrated by Fig. 38-4.

In the lowest f-m channel the center carrier frequency is 88.1 mc. The oscillator frequency is higher than this or any other received frequency by the amount of the intermediate frequency. Consequently, with an intermediate of 10.7 mc and a receiver tuned to the lowest f-m channel the oscillator frequency will be the sum of 88.1 and 10.7 mc, or will be 98.8 mc. The image frequency will be the sum of the oscillator and intermediate frequencies, the sum of 98.8 mc and 10.7 mc. This comes to 109.5 mc, and is higher than the 108-mc upper limit of the f-m broadcast band. Such a high frequency is beyond the normal tuning range of the receiver. For any higher f-m channel the oscillator and image frequencies will be proportionately higher, and the image will remain above the normal tuning range.

⑦ It always works out that any intermediate frequency equal to more than half the band width will throw all image frequencies out of the normal tuning range, or will make them higher than the highest frequency in the received band. The f-m broadcast band has a width of 20 mc, since it extends from 88 to 108 mc. Half this width is 10 mc, and any intermediate frequency greater than 10 mc will place all images above the high limit of the band.

In many f-m receivers manufactured before 1947 the intermediate frequency is 4.3 mc. This suited an earlier

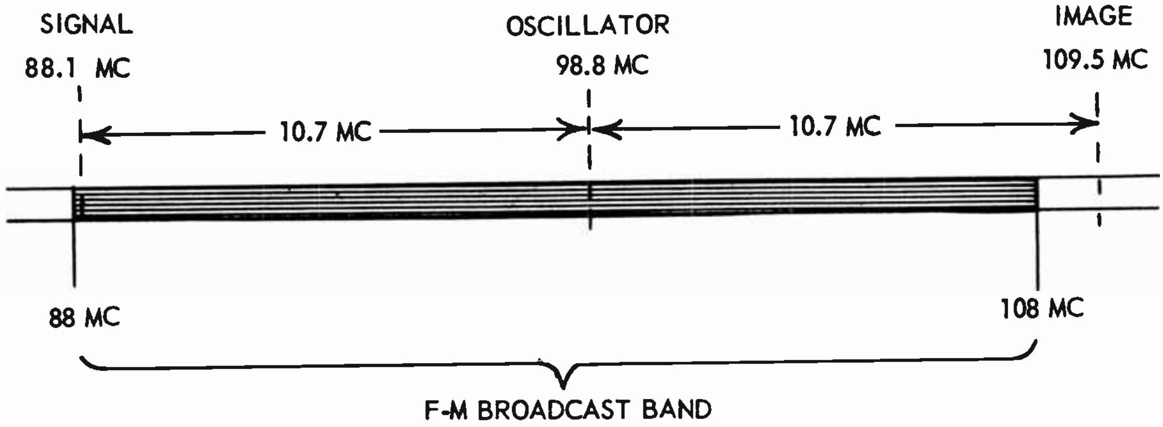


Fig. 38-4. A 10.7-mc intermediate frequency places all image frequencies above the high limit of the f-m broadcast band.

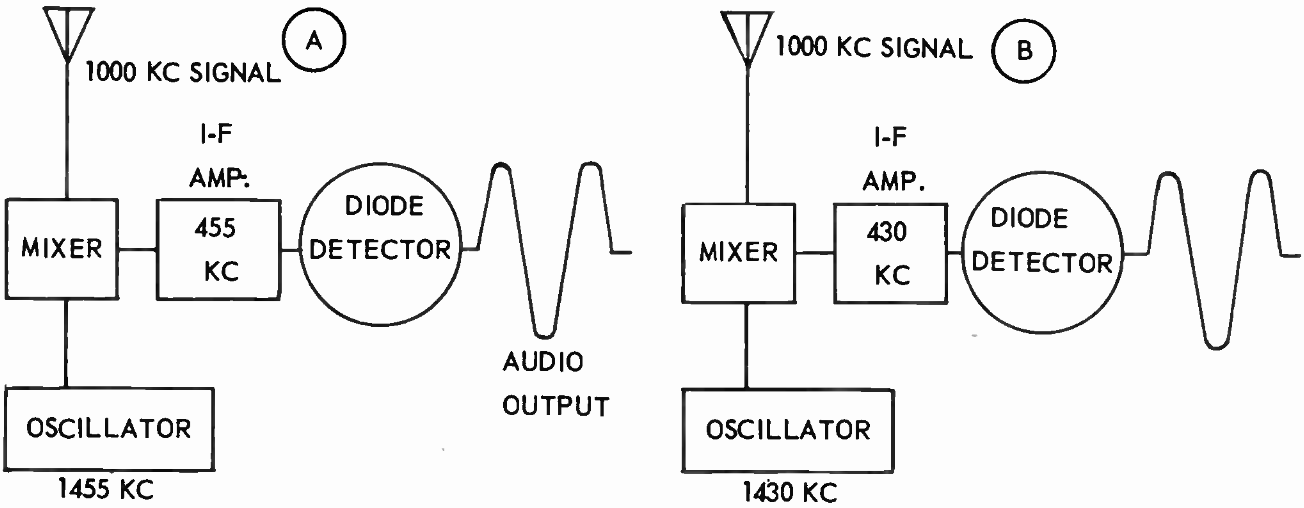


Fig. 38-5. Frequencies of the r-f oscillator and i-f amplifier for a-m reception.

f-m broadcast band which extended from 42 to 48 mc for a width of 6 mc. Then the 4.3 mc intermediate, which is more than half of 6 mc, got rid of the image frequencies. Still other receivers of early design use an intermediate frequency of 8.3 mc.

FREQUENCY RELATIONS. It is necessary to take much more care in the tuning or alignment of i-f amplifiers for f-m receivers than for a-m receivers. This is because, in a-m receivers, we require correct frequency relations between only the i-f amplifier and the r-f oscillator in order to permit satisfactory demodulation by the diode detector. In f-m receivers we must have correct frequency relations between the r-f oscillator, the i-f amplifier, and the tuned transformer of the discriminator or ratio detector. The differences between alignment requirements will become apparent from an examination of the processes of signal reproduction in the two methods of reception.

Detection of amplitude modulation is represented by Fig. 38-5. At *A* we have a 1,000-kc received signal and an i-f amplifier tuned to 455 kc. By tuning the r-f oscillator to 1,455 kc we obtain the difference between oscillator and signal frequencies as an intermediate frequency of 455 kc, with all the original amplitude modulation. This modulation is rectified by the diode detector.

In diagram *B* we still have the 1,000-kc received signal, but the i-f amplifier has been mistuned to 430 kc. However, by returning or realigning the oscillator at 1,430 kc we may obtain a 430-kc intermediate frequency with exactly the same modulation as before. The i-f amplifier will provide satisfactory gain because this amplifier is tuned to 430 kc. The diode detector demodulates this lower intermediate frequency and delivers exactly the same audio output as before.

A diode detector demodulates changes of amplitude regardless of the frequency, efficiently rectifying or detecting signals at all intermediate frequencies from a few cycles per second to many megacycles per second. The diode detects whatever frequencies are applied to it.

Fig. 38-6 shows how we may get into trouble with mistuning of either the i-f amplifier or the demodulator for f-m reception. In diagram *A* everything is correctly tuned or aligned. There is a received signal at 101.1

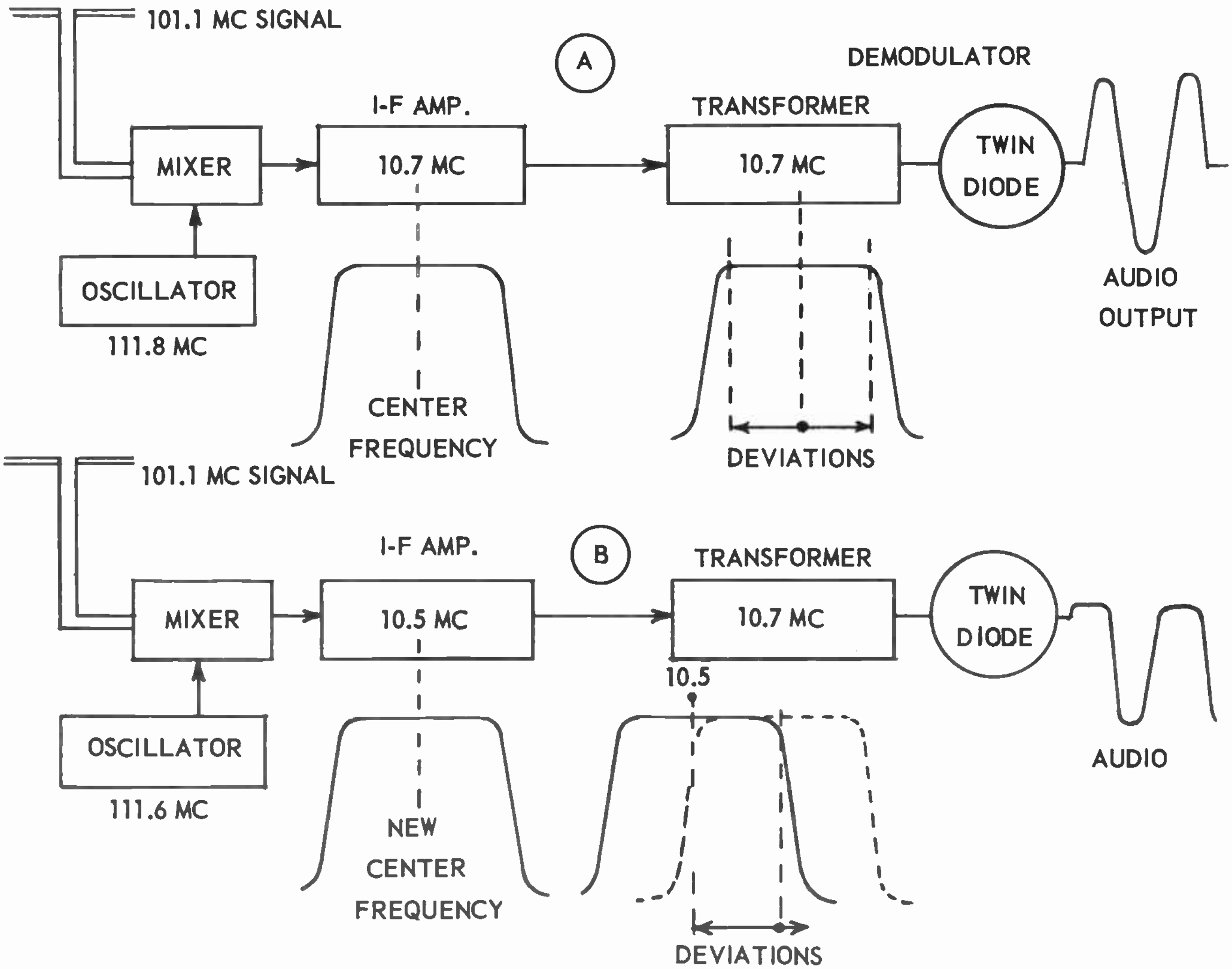


Fig. 38-6. Frequency relations of the r-f oscillator, the i-f amplifier, and the demodulator transformer for f-m reception.

mc. The i-f amplifier and the demodulator transformer are both tuned to 10.7 mc. With the r-f oscillator tuned to 111.8 mc for this particular received signal we obtain as the difference between oscillator and signal frequencies the required 10.7 mc intermediate center frequency. Deviations above and below the center are equally amplified, because the frequency response of the i-f amplifier extends equally above and below 10.7 mc. With the demodulator transformer correctly tuned to 10.7 mc, deviations of frequency above and below this center frequency cause equal changes of currents in the demodulator diodes, and the alternating audio output is equal above and below its zero value. This permits distortionless audio output.

In diagram *B* of Fig. 38-6 we have the same 101.1 mc received signal, but the i-f amplifier has been mistuned to 10.5 mc instead of 10.7 mc. The demodulator transformer still is tuned to 10.7 mc. By returning the r-f oscillator to 111.6 mc it is possible to obtain a modulated intermediate frequency centered on 10.5 mc, which is the difference between the signal frequency of 101.1 mc and the oscillator frequency of 111.6 mc.

The i-f signal voltages, now centered on 10.5 mc, suit the tuning of the i-f amplifier, and there are equal gains for deviation frequencies above and below 10.5 mc. But when this output of the i-f amplifier is applied to the demodulator transformer, tuned to 10.7mc, there is trouble. Deviations to frequencies lower than 10.7 mc will be reproduced as full audio output or possibly as excessive audio amplitudes. Deviations to higher frequencies can cause only very small audio amplitudes. This happens because the i-f amplifier is delivering practically no voltages at frequencies higher than 10.7mc, while delivering plenty of signal voltage at frequencies far below 10.7 mc. The result is severe audio distortion.

Were the i-f amplifier to remain tuned at 10.5mc, and oscillator at 111.6 mc for this particular channel, it would be necessary to retune the demodulator transformer to 10.5 mc to match the i-f amplifier. Quite likely it would be possible to do this, many transformers having enough tuning adjustment to allow it. Then the demodulator transformer and all the i-f transformers would be tuned to the same center frequency. The r-f oscillator could be aligned to produce this center frequency from signal frequencies in each channel.

Although mistuning of all transformers to some certain center frequency usually would produce acceptable audio output it is better to be on the safe side and use the standard intermediate of 10.7 mc. All modern transformers surely will tune to 10.7mc while having equal gains at higher and lower deviation frequencies.

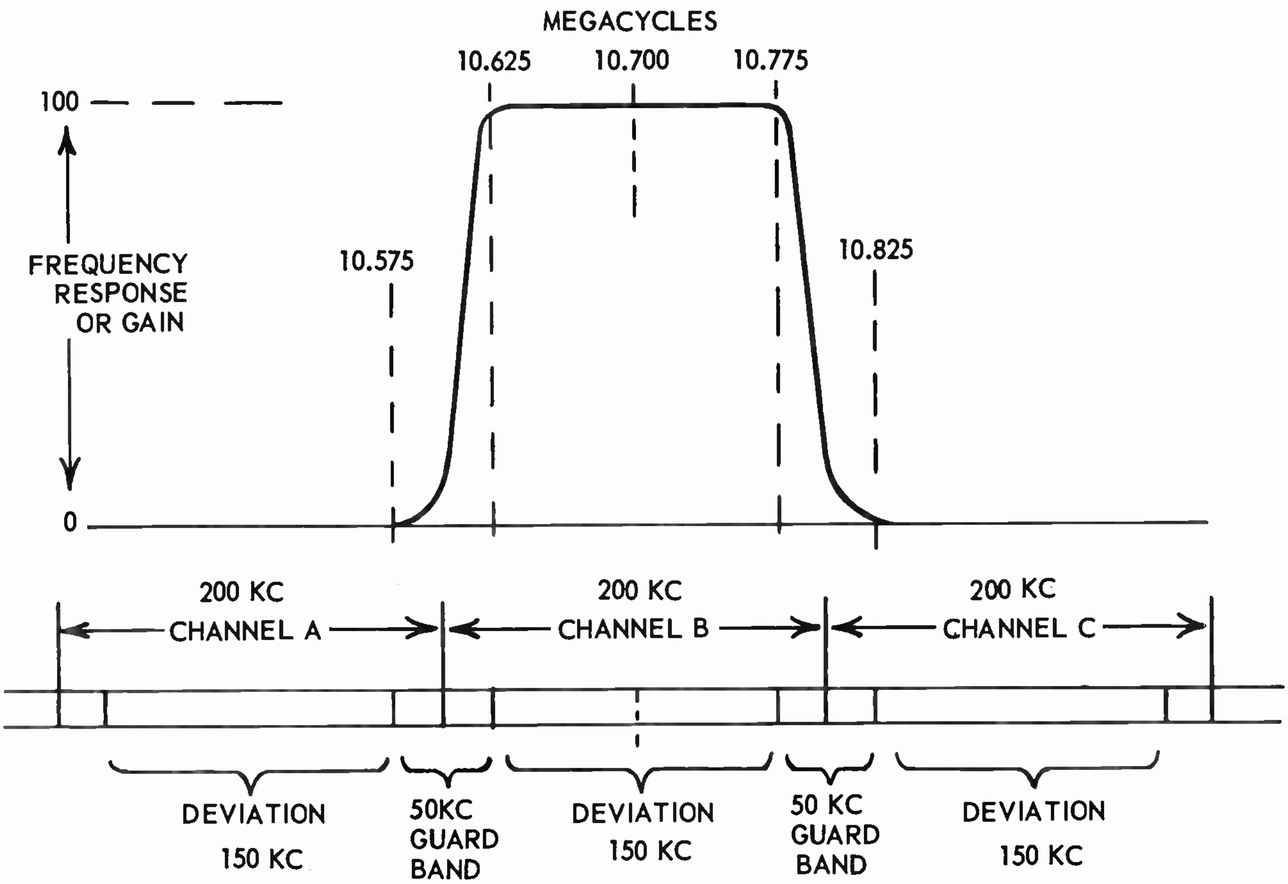


Fig. 38-7. The ideal frequency response of the i-f amplifier system for f-m reception.

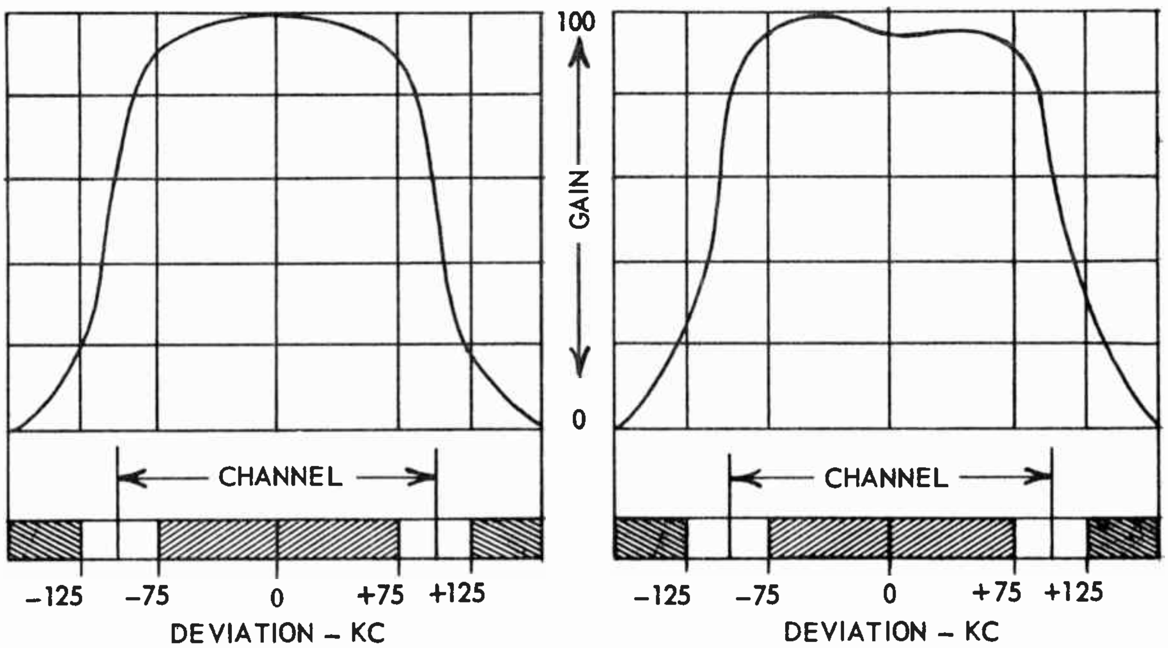


Fig. 38-8. Typical satisfactory overall responses of the i-f system for f-m reception.

When aligning an f-m receiver, or the sound section of a television receiver, it is customary to commence by tuning the demodulator transformer. Then we go to the i-f transformers, and finally to the tuner section. Whatever may be the center frequency used for demodulator alignment, this same frequency must be used for i-f amplifier alignment, and the r-f oscillator must be aligned to produce this center frequency on all received signals.

FREQUENCY RESPONSES. Fig. 38-7 shows how we determine the ideal frequency response or the relative voltage gains at various frequencies for an i-f amplifier tuned to a center frequency of 10.7 mc. At the bottom is represented the range of frequencies for three adjacent channels, marked *A*, *B*, *and C*. We are tuning with respect to channel *B*. Each channel extends over a total range of 200 kc 0.200 mc. Frequency deviations of signals in each of these f-m channels may go to 75 kc above and to 75 kc below the center frequency of that channel, making a total frequency swing of 150 kc. There is a 50 kc gap either side of adjacent channels, used as guard bands, wherein there are presumed to be no signal frequency deviations.

We should like to have equal gains or uniform response throughout the maximum deviations of 75 kc (0.075 mc). This would require a uniform gain or response all the way from 10.625 mc, representing a downward deviation of 75 kc, up to 10.775 mc, representing an upward deviation of 75 kc. But at maximum deviations in the two adjacent channels we should like to have zero gain or zero response. This would call for complete cutoff or zero gain at 10.575 mc on the low side and at 10.825 mc on the high side.

With response of this kind there would be perfect reproduction of all deviations in our own channel, and perfect selectivity or complete rejection of all signals in both adjacent channels. The ideal response, from zero gain below to zero gain above the center frequency, would extend from 10.575 mc to 10.825 mc. This is a range of 0.250 mc or 250 kc, with a flat top 150 kc wide.

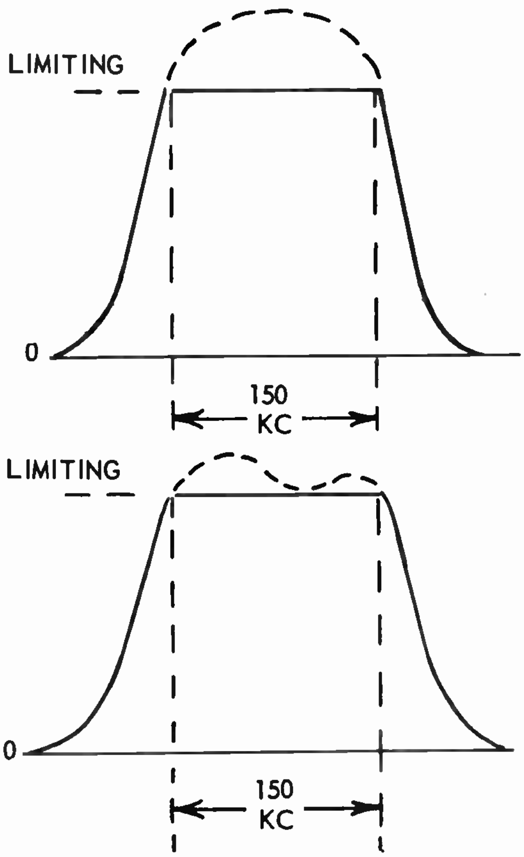


Fig. 38-9. Limiting action effectively flattens the top of the i-f response.

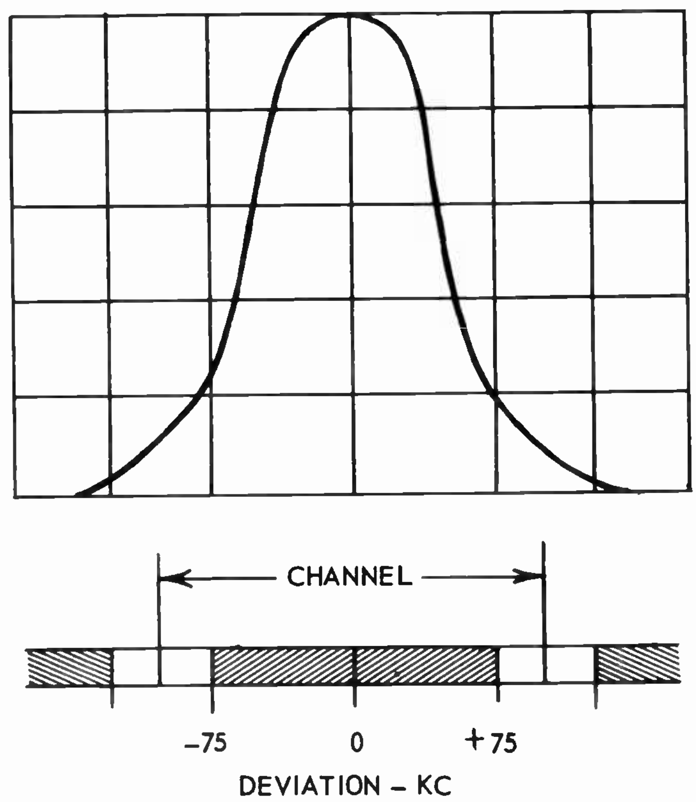


Fig. 38-10. Overall i-f response that is too narrow or too sharply peaked.

No matter how carefully you tune the i-f amplifier transformers in an ordinary receiver the response won't be of the ideal form. As a general rule, careful adjustment will produce a curve of gain versus frequency deviation somewhat like the responses in Fig. 38-8. Although there is considerable gain remaining at deviation frequencies in adjacent channels, this is not objectionable because f-m transmitters are geographically spaced far enough apart that signals from stations in adjacent channels should not interfere with each other. That is, stations operating in channels adjacent to any used in your locality will be of so many miles away that their signals will be too weak to cause interference.

When the top of a response curve is not flat there is greater gain at some frequencies than at others, for each point from left to right along the curve represents one frequency. This means that signal voltage amplitude will be greater for some deviations than for others. Such unequal amplitudes should cause no difficulty in f-m reception, because they will be reduced to a uniform value by the limiter used ahead of a discriminator or by the large capacitor in the load circuit of a ratio detector. This limiting effect or equalizing of amplitudes is shown by Fig. 38-9.

There is a peculiarity of f-m signal reception that improves selectivity. A strong received signal seems to "swamp" any weaker signals reaching the same set. When you correctly tune an f-m receiver to a reasonably strong signal, other weak signals seem to produce no response – as though the receiver circuits could work on only one set of frequency deviations at a time. With a-m reception you could hear the weaker signals in the background as interference with the stronger signal that you wish to hear. This peculiarity of f-m signals makes them undesirable for some emergency services where, oftentimes, it is essential to hear or to "read" a distant weak signal in spite of stronger signals from closer transmitters.

Selectivity is improved by having more tuned circuits in the signal path. The greater the number of amplifying stages, or the greater the number of tuned circuits of any kind which handle the signal one after another, the sharper will be the cutoff on the sides of the overall frequency response. A single i-f stage might have a rather broad frequency response. If followed by another similar stage there will be not only additional gain but also a narrower overall response. A third similar stage will give still more gain and a still narrower frequency response.

There are many i-f transformers with which, by incorrect adjustment, it is possible to obtain an overall frequency response that is too sharply peaked, as in Fig. 38-10. Here the voltage at maximum deviations of 75 kc either side of the center frequency drops to only about 20 per cent of the peak gain.

The maximum frequency deviations represent maximums of audio loudness, because amplitudes in the demodulated audio signal result from deviations of frequency in the i-f signal. The sharp peaking of i-f amplifier response will cut off maximums of audio loudness in the sound output. Sound will be flat and without liveliness. Of course, the entire sound reproduction may be made louder by manipulation of the volume control, but there will be little contrast in musical passages that should be alternately loud and soft, and speech will come through at a nearly constant level. All the audio frequencies will be there, because audio frequency results from the number of deviations per second rather than from the extent of deviation to higher and lower frequencies.

AUTOMATIC VOLUME CONTROLS. Earlier we learned how amplifications of r-f and i-f amplifier tubes in a-m receivers is subjected to automatic volume control to compensate for variations in strength of the received signal. The avc voltage usually is taken from the output of the a-m diode detector. Because we do not use a diode detector for demodulation of f-m signals it becomes necessary to use different methods of automatic volume control in f-m receivers.

In f-m receivers employing a discriminator and limiter the signal amplitude of the limiter output and the discriminator input is constant for all signals above a strength that operates the limiter. This constant amplitude would not produce a varying avc voltage if rectified. Consequently, in receivers having a discriminator, the avc voltage is taken from the limiter input or from some point ahead of the limiter. With a ratio detector for f-m demodulation the i-f signal voltage in the detector and its load circuit varies with average strength of the received signal. This detector load voltage can be used for automatic volume control.

Not all receivers which tune in the f-m broadcast band have automatic volume control for the f-m signals. Many designers have felt that amplification should not be reduced, but allowed to remain as great as possible so that even relatively weak received signals may be brought up to an amplitude at which they will be subjected to amplitude limiting by a limiter stage. Combination fm-am sets practically always have automatic volume control for a-m reception, using systems like those employed in single band standard broadcast a-m receivers.

We should keep in mind that when automatic volume control is used for f-m reception it really is a control for automatically increasing the amplitude of weak received signals and for automatically limiting the amplitude of strong signals. The latter function, of limiting the gain, supplements the action of a limiter stage in sets employing a limiter.

3 When automatic volume control is used for r-f amplifiers and i-f amplifiers in any receiver, these amplifier tubes are protected against overloading when signals are excessively strong. When signal amplitude becomes very great the positive peaks of voltage may exceed the normal grid bias and cause distortion. Overly strong signals also have an effect in amplifier tubes which broadens the tuning in connected circuits. This effect is reduced by automatic volume controls.

4 In f-m receivers which have ratio detectors we have ready at hand a voltage snitable for automatic volume control. This voltage exists at the negative terminal of the large capacitor that absorbs variations of signal amplitude on the load side of the detector. You will recall that voltage across this large capacitor varies only with changes of average signal strength. The capacitor voltage increases on strong signals and decreases when signals are weaker. Then the potential at the negative terminal of this capacitor becomes more negative with reference to ground when signals are strong, and becomes less negative when signals are weak.

One method of using this capacitor potential for automatic volume control on an r-f amplifier and on a second i-f amplifier is shown by Fig. 38-11. The avc bus connects to the negative terminal of capacitor *Cs*, and to the bus are connected the grid returns of the tubes to be automatically controlled. The varions capacitors marked *Cf* are bypasses for high-frequency fluctuations of signal voltage which exist at capacitor *Cs*. These bypasses, in connection with resistors marked *Rf*, ensure that only smooth direct biasing voltage reaches the grids of controlled tubes from the avc system, and that no signals from one grid circuit may pass to other grid circuits through the volume control system.

When a strong received signal increases the voltage across capacitor *Cs* the grids of the controlled tubes are made more negative with reference to ground and to the tube cathodes, and amplification or gain is reduced. Weak received signals decrease the voltage across capacitor *Cs*, with the result that grids of the controlled tubes become less negative, and amplification increases. The controlled tnbes may or may not have the additional cathode bias shown in the diagram.

The automatic control voltage may be applied to the grids of different tubes and combinations of tubes.

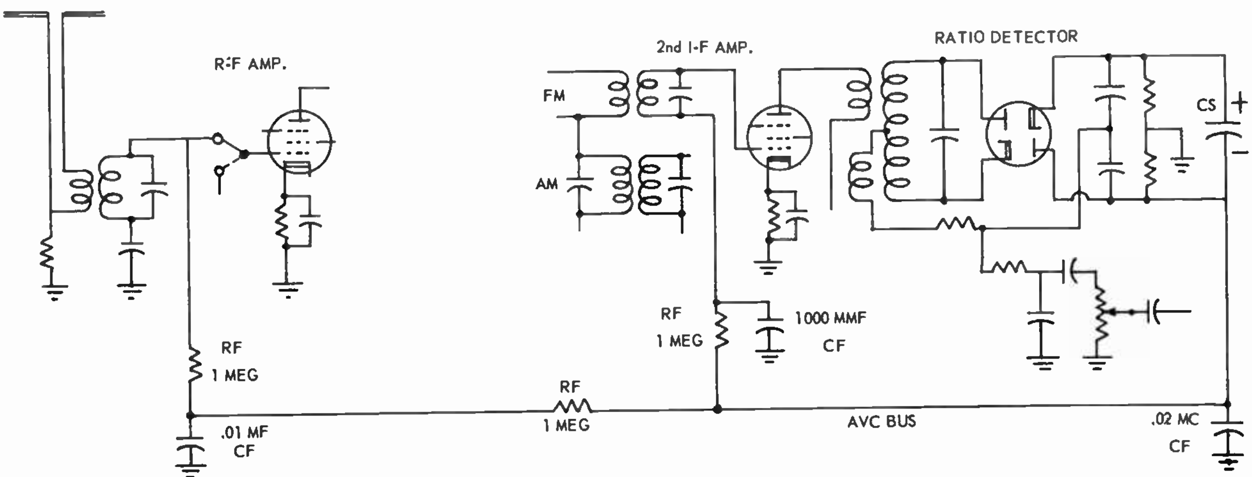


Fig. 38-11. **An automatic volume control taking its negative biasing voltage from the large capacitor on the load side of a ratio detector.**

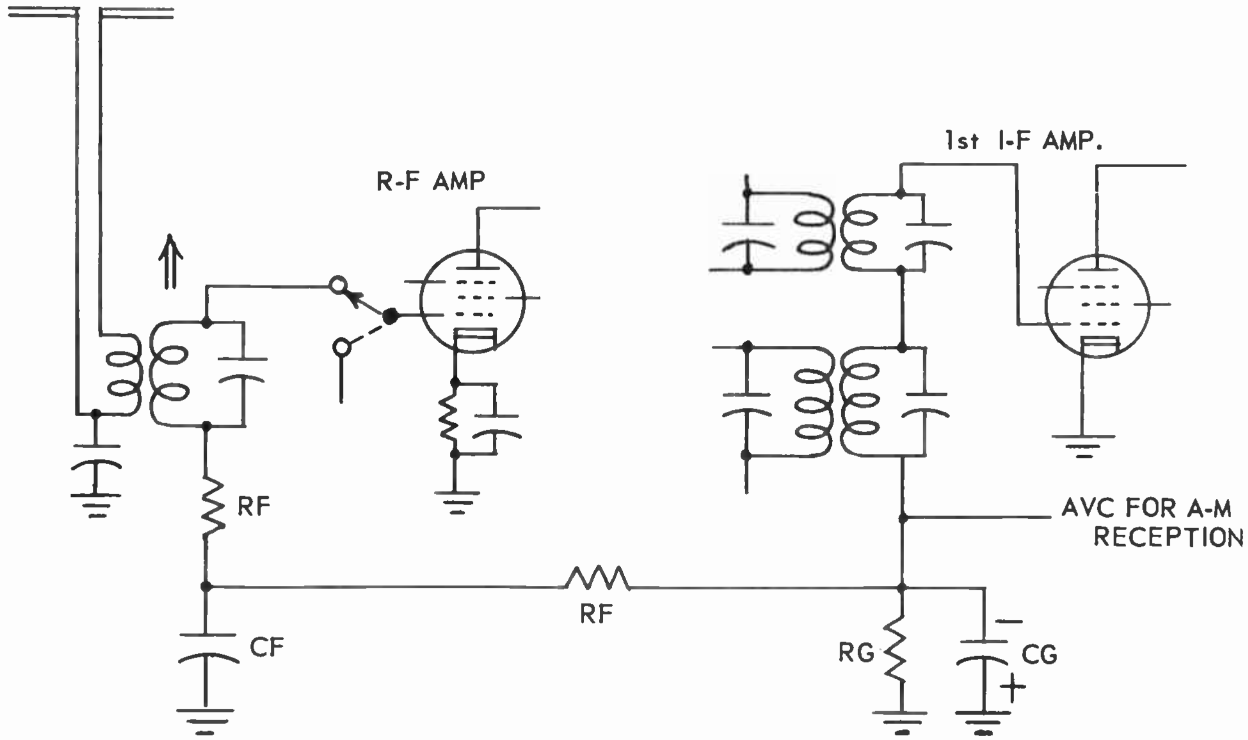


Fig. 38-12. How grid-leak biasing voltage may be used for automatic volume control.

In receivers having an r-f amplifier this amplifier usually has automatic volume control. When there is no r-f amplifier the automatic control usually is applied to the first i-f amplifier tube. In some sets there is automatic volume control on the r-f amplifier and also on both the i-f amplifiers used for f-m reception.

Another method of automatic volume control makes use of grid-leak bias for the controlled tubes. Fig. 38-12 shows a circuit in which grid-leak bias control is applied to the r-f amplifier and the first i-f amplifier. The biasing resistor or grid leak resistor is *Rg*. The biasing capacitor is *Cg*. Resistors *Rf* and capacitor *Cf* are for smoothing the direct biasing voltage, removing high-frequency variations, and for isolating the signal circuits from one another.

As with any system of grid-leak biasing there is rectification of the positive peaks of signal voltage alternations. The resulting pulses of grid current charge capacitor *Cg* in the marked polarity. This charge escapes slowly through resistor *Rg*. A strong received signal increases the charge on capacitor *Cg* and its upper end becomes more negative with reference to ground. This negative voltage at the upper ends of *Cg* and *Rg* is the biasing voltage for the controlled tubes. The more negative biasing voltage which results from strong signals reduces amplification of the controlled tubes. Weaker signals allow a bias voltage which is less negative, and there is greater amplification.

Grid-leak bias for automatic volume control is applied to various combinations of amplifying tubes. In Fig. 38-12 the automatic volume control acts on the r-f amplifier and the first i-f amplifier. It may be applied otherwise to one or more individual tubes, with separate grid-leak resistors and grid capacitors for each controlled tube.

In some f-m receivers which have a limiter and discriminator, an automatic volume control potential is taken from the grid end of the limiter grid resistor and applied to preceding i-f amplifiers, to an r-f amplifier, or to any combination of tubes which are to be automatically controlled. Connections would be essentially the same as in Fig. 38-12, except that the grid resistor and capacitor would be on the limiter instead of on an i-f amplifier tube.

When a limiter acts to reduce excessive amplitude of signal voltages applied to it there is an increase of charge on the grid capacitor. Then the end of the grid resistor which is toward the grid of the limiter tube becomes more negative with reference to the other end or grounded end. This negative voltage varies with strength of the received signal, becoming greater on strong signals and less on weak ones. Consequently, the limiter grid voltage is suitable for automatic volume control.

F-M TUNERS. In the great majority of recently designed f-m receivers and combination fm-am receivers the variable tuning is by means of variable capacitors and fixed inductors. This method of tuning is used also for the a-m sections of combination receivers. F-m and a-m tuning is handled by the same control knob. Usually there is a single pointer moving over a dial graduated for both bands. A tuning mechanism commonly used consists of a multi-section variable capacitor with groups of plates for both f-m and a-m tuning. One such design is illustrated by Fig. 38-13. In each a-m gronp of plates there are 12 stators and 13 rotors. In each f-m group there is only a single rotor plate that moves between two stators.

In the three-gang tuning capacitor pictured one of the gangs would be used for the coupling between antenna and r-f amplifier. Another gang would tune the coupling between r-f amplifier plate and grid of the mixer or converter. The third gang would tune the r-f oscillator. For a receiver having no r-f amplifier only two capacitor gangs would be needed, one for tuning the antenna coupling and the other for the r-f oscillator. A two-gang capacitor would be used also in receivers having an r-f amplifier but using an untuned coupling between antenna and this amplifier. Then one of the two gangs would tune the coupler between r-f amplifier and mixer or converter, while the second gang would tune the r-f oscillator.

Only small values of capacitance and inductance are needed for resonance at the f-m carrier frequencies between 88 and 108 mc. Typical variable tuning capacitors have maximum capacitances of 15 to 20 mmf and minimums of 5 to 6 mmf. Tuning inductances are on the order of 1/20 microhenry, which would be provided by a coil of three to five turns, widely spaced, and with diameter of 3/8 to 1/2 inch.

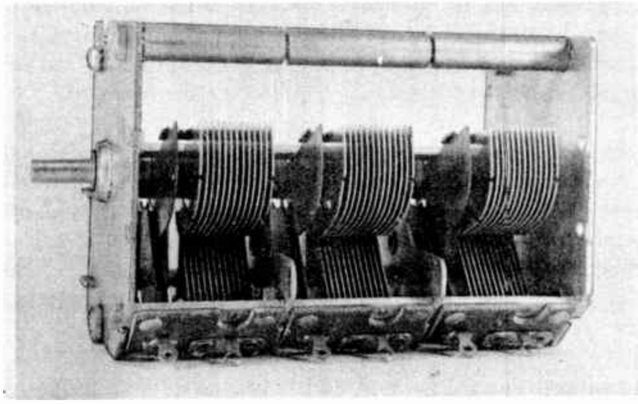


Fig. 38-13. A three-gang variable tuning capacitor for f-m and a-m reception.

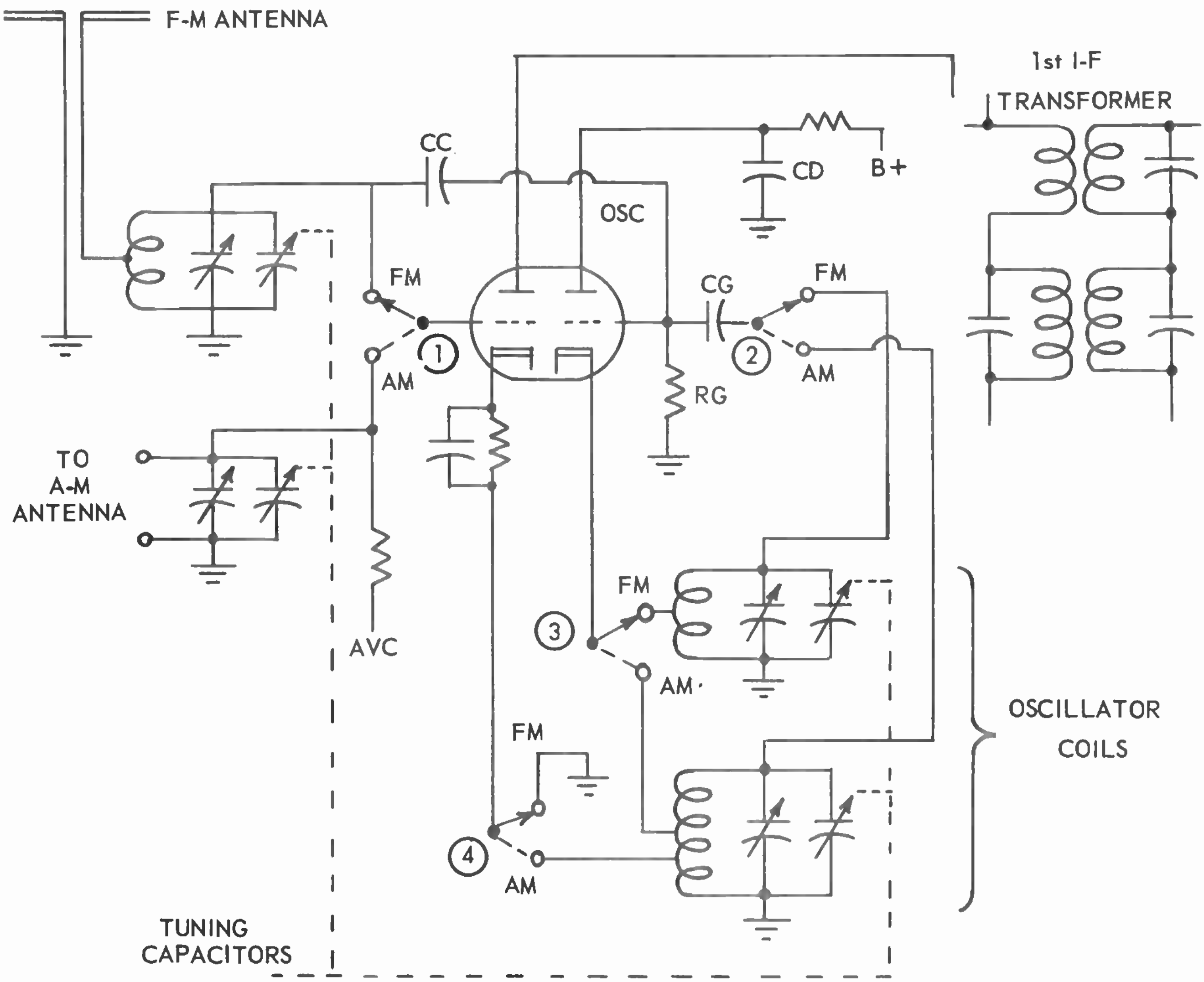


Fig. 38-14. Tuner circuits in a receiver having no r-f amplifier, and with the antenna coupled to the mixer grid.

Fig. 38-14 is a circuit diagram for a tuning system on a receiver having no r-f amplifier, using a twin-triode tube for mixer and oscillator, and having the antenna coupled to the grid of the mixer. The oscillator circuits for both f-m and a-m operation are of the Hartley type in this particular tuner. Other types of oscillators may be used for either or both bands.

Four sections of the band switch are shown in the tuner diagram. Section 1 connects the grid of the mixer to the tuned coupler for the f-m antenna or to another tuned coupler for the a-m antenna. Switch section 2 connects the oscillator grid to the tuned oscillator circuits for either band. Section 3 connects the oscillator cathode to either one or the other of the tuned oscillator circuits. Section 4 shifts the connection of the mixer cathode for reception in either band.

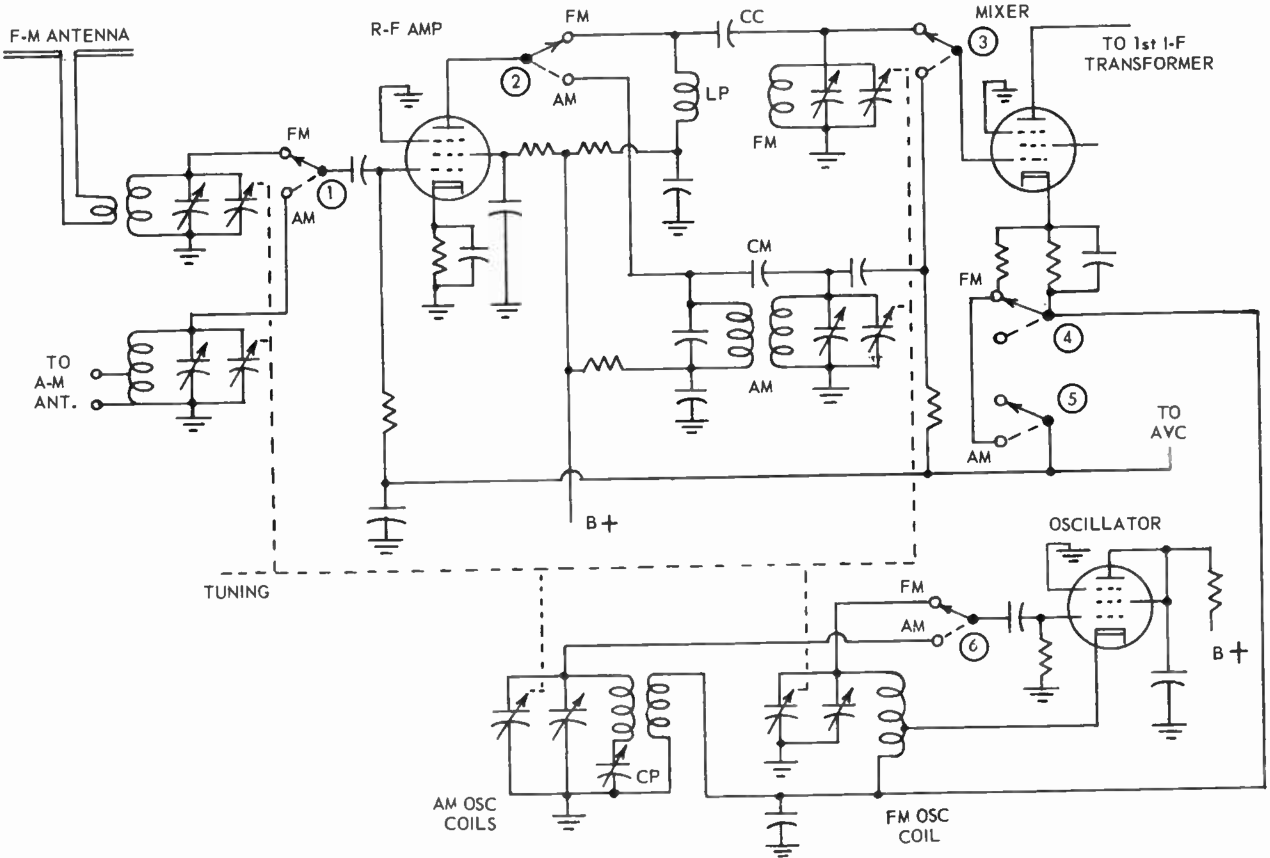


Fig. 38-15. Combination fm-am tuner in which parts of a ganged tuning capacitor control the antenna coupling, the r-f to mixer coupling, and the oscillator circuits.

The resistor and capacitor which are in series with the cathode lead of the mixer provide cathode bias during operation in both bands. This is the only bias for the r-f amplifier during f-m reception. But when switch section 1 is in its a-m position the grid of the r-f amplifier is connected to an avc bus, providing automatic volume control for a-m signals.

During f-m reception the high-frequency oscillator voltage is fed from the oscillator grid connection through capacitor *Cc* to the grid of the mixer while switch section 1 is in its f-m position. This oscillator-to-mixer connection is cut off when switch section 1 is moved to its a-m position. During f-m reception the cathode bias line of the mixer has been connected to ground through switch section 4. But when the band switch is moved over to the a-m position this section 4 of the switch connects the mixer cathode line through the lower part of the a-m oscillator coil and thence to ground. Then oscillator voltage is introduced into the mixer from the oscillator coil.

The variable or movable members of the ganged tuning capacitor are shown connected together by a broken line in Fig. 38-14. In parallel with each tuning capacitor is an adjustable trimmer capacitor. It should be noted that in this entire diagram we have seen nothing new except the method of changing connections by means of the band switch. Every separate circuit detail has been examined in earlier lessons.

In Fig. 38-15 is shown a combination fm-am tuner circuit in which there is an r-f oscillator, a mixer, and an r-f amplifier. There are tuned couplings leading to the grid of the r-f amplifier from the f-m antenna and from the a-m antenna. There are tuned couplings for each band between the plate of the r-f amplifier and the grid of the mixer. There are tuned oscillator coils for each band.

Tuning is handled by six variable capacitors, shown connected together by a broken line in the diagram to indicate that the rotors are ganged on one shaft, as in Fig. 38-13. In parallel with each of the six groups of plates in the tuning capacitor is an adjustable trimmer capacitor. There is also an adjustable padder capacitor *Cp* in series with one of the a-m oscillator coils shown at the bottom center of the diagram.

Six sections of the band switch are used in the tuner circuits. Switch section 1 shifts the grid of the r-f amplifier to either one or the other of the antenna couplers. Switch section 2 shifts the plate of the r-f amplifier between the f-m and a-m couplers leading to the mixer, while section 3 connects one or the other of these couplers to the grid of the mixer tube. This r-f to mixer coupling for f-m reception utilizes an r-f choke *Lp* to provide a load in the plate circuit of the r-f amplifier, and uses a variably tuned parallel resonant circuit for the mixer grid, with signal transfer through coupling capacitor *Cc*. For a-m reception the r-f to mixer coupling consists of a transformer in which the secondary, or the mixer grid side, is variably tuned. In addition to inductive coupling between the transformer windings there is coupling or signal transfer through the small fixed capacitor *Cm*.

Section 4 of the band switch connects two cathode bias resistors in parallel on the mixer circuit for f-m reception, and leaves only one of these resistors for a-m reception. When switch section 5 is in its f-m position the grid return line from the r-f amplifier is connected only to the automatic volume control voltage or the avc bus. With this switch section in its a-m position part of the cathode current of the mixer flows through connections not included in this diagram to furnish avc voltage for the r-f amplifier and mixer grids.

From the common terminal of switch section 4 a line goes to oscillator coils for both bands, thus providing for introduction of oscillator voltage in the cathode circuit of the mixer. Switch section 6 connects the oscillator grid to one or the other of the oscillator tuned circuits. The oscillator circuit for f-m reception is a Hartley type. For a-m reception the oscillator circuit is a tickler feedback type.

* * * * *

In this lesson we have traced all the circuits in a few i-f amplifiers and in two or three f-m tuners. Because no two receivers are exactly alike in every detail we might continue thus to trace individual circuits almost indefinitely. In doing so it would become evident that we are merely looking at different arrangements and combinations of a relatively small number of basic circuits. There would be wide variations in wiring connections and in methods of band switching. But you would find only a limited number of fundamental circuits between antennas and r-f amplifiers, mixers, or converters. There would be only a limited number of types of transformers or other couplings between r-f amplifiers and mixers or converters. There would be only a limited number of basic oscillator circuits.

Most of our instrument connections for testing, alignment, and trouble shooting will be made at points quite easily identified in any circuit arrangement. Nearly all these connections will be made at antenna terminals and at other points where signals normally enter the various sections of a receiver, or at the grids of certain tubes, at the plates or plate loads of other tubes, and at points where we have r-f, i-f, or audio output voltages from the various sections.

It is important to learn to trace given connections on circuit diagrams and also in receivers themselves so that you may identify certain tube elements or socket lugs, or identify some particular resistor, capacitor, or inductor in the circuit diagram and in the set itself. We learn to trace circuits so that we may make correct connections of service instruments, and so that we may determine the points between which some apparent trouble must exist.

Even with parts utilizing electrical principles so different as those in ratio detectors and discriminators you will discover that there is hardly any difference between methods of alignment or between the indications that you watch for during service operations. Wherever it happens that differences in electrical principles do require changes in service methods these changes will be brought out as we proceed with instructions for servicing.

TELEVISION

LESSON NO. 39

INSTRUMENTS FOR F-M ALIGNMENT

Training Your Mind...

C.T.I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

COMMERCIAL TRADES INSTITUTE

Training Your Mind

C.T.I.

Perfecting Your Skill

Chicago, Ill.

LESSON NO. 39

INSTRUMENTS FOR F-M ALIGNMENT

Now we are ready to align an f-m receiver. Each separate step, and the reasons for taking it, will be explained in detail — because the same methods will be used later for alignment of sound sections in television receivers. The principal differences between f-m receivers and television sound sections are in the use of different intermediate frequency and in that we have fewer adjustments in the television sound section than in the f-m receiver.

10 In Fig. 39-1 the rectangular blocks represent all the circuits which require tuning during alignment of an f-m receiver or the f-m portions of a combination fm-am receiver. These circuits will be aligned in an order that is in reverse to the direction of signal travel. The frequency-modulation signals travel from the f-m antenna to the demodulator, where they become audio signals. We commence alignment operations at the demodulator transformer and finish at the antenna coupler. After aligning the demodulator transformer we move to the preceding i-f transformer. Then we align, in order, whatever other i-f transformers may be used until completing adjustment of the transformer that follows the converter or mixer tube.

After completing alignment of the demodulator and i-f amplifiers we go to the tuner. Here the first job is alignment of the r-f oscillator coils and trimmers. If there is a tuned coupling between an r-f amplifier and the mixer or converter this coupling is aligned next. Finally we align the tuned coupling which is between the antenna and r-f amplifier or between antenna and mixer or converter, depending on how the receiver is constructed.

In Fig. 39-2 the blocks represent circuits which require alignment in the sound section of a television receiver. There is no separate tuner for sound in the television receiver, this function being cared for by the same tuner that handles picture or video signals. The sound intermediate frequency may be taken from the output of the television mixer tube, or from some point in the television i-f amplifier, or from the video amplifier. There sound signals, which are frequency-modulated, then pass through one or more sound i-f transformers to the sound demodulator. The demodulator may be a ratio detector or a discriminator.

In the television sound section we again carry out the alignment in an order that is in reverse to the direction of signal travel. First comes adjustment of the demodulator transformer. Then we work back through whatever i-f transformers may be used, and finally reach the point of sound takeoff. This completes the sound alignment in the television receiver.

Alignment may be needed in any receiver when tubes have been replaced in amplifier, oscillator, or demodulator circuits, or when any other parts of these tuned circuits have been replaced. If positions of wires or small parts in high-frequency circuits have been changed during service operations it may be necessary to realign these circuits. If an unqualified service man has "monkeyed" with the adjustments the set is certain to need alignment. There may be slow changes in values of capacitors, inductors, and resistors which eventually make realignment necessary.

Always remember that every receiver must have been correctly aligned to begin with. Poor performance is far more likely to result from defective tubes, small parts, and connections than from misalignment. Therefore, don't undertake the alignment of a receiver or sound section until you have fully investigated the possibility of other reasons for trouble.

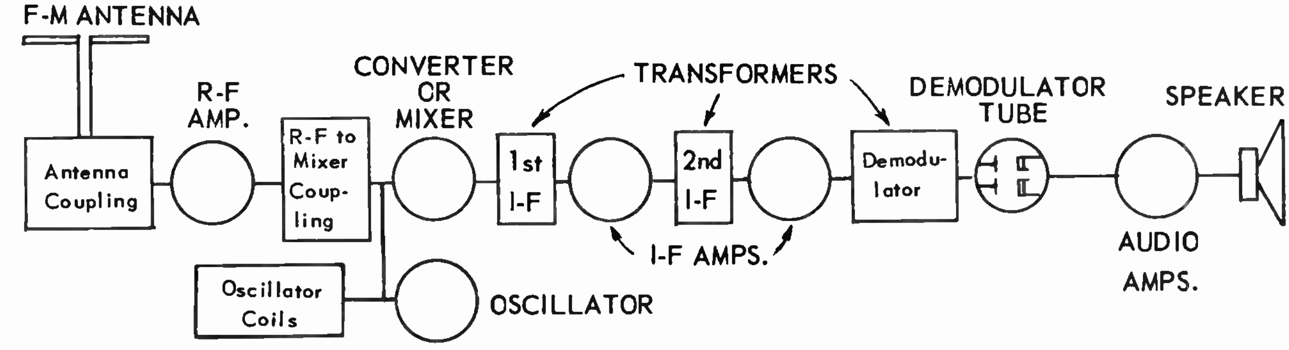


Fig. 39-1. The parts of an f-m receiver requiring alignment are shown by the rectangular blocks.

Nothing does more to improve the behavior of a receiver than correct alignment, provided there is real need for it. But to do a first class job of alignment requires skill that comes from practice as well as from a thorough knowledge of what you are trying to accomplish, and why. The first time you undertake alignment of a television receiver's video or sound section it probably will seem that everything you do makes matters worse instead of better, and you will wish you had let things alone. But you will keep on, and eventually bring about a big improvement. Soon you will get the "feel" of various adjustments, and then it will take only an astonishingly short time to bring ailing receivers up to top performance.

When you align a combination fm-am receiver always begin by adjusting the circuits that operate in the standard a-m broadcast band, then go on to the f-m adjustments. A-m alignment is made in the manner explained in an earlier lesson. If you complete the f-m alignment, then disturb any of the a-m adjustments, it will be necessary to realign the f-m sections. Tuned circuits for the two bands are so closely associated that each is affected by the other.

SIGNAL GENERATORS. The signal generator used for f-m alignment must furnish a frequency of 10.7 mc and other frequencies somewhat below and above this value. This range is needed for alignment of demodulators and i-f amplifiers. For older f-m sets it will be necessary to have frequencies at and near 4.3 and 8.3 mc.

In addition it is desirable that the signal generator furnish radio frequencies from a little below to a little above the f-m carrier band, say from 86 to 120 mc. If you do not have a signal generator furnishing such high frequencies it usually is possible to use harmonics of lower generator frequencies. All ordinary service type signal generators put out not only the frequency at which they are tuned, but at the same time deliver harmonic frequencies at many multiples of the tuned frequency.

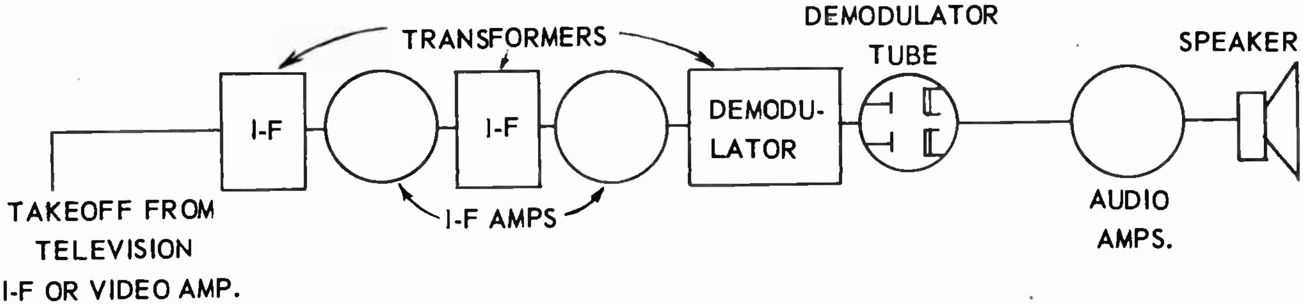


Fig. 39-2. The blocks show parts of a television sound section which are aligned.

For instance, when any ordinary signal generator is tuned to 10 mc it will be producing at the same time harmonic frequencies of 20, 30, 40, 50 mc, and even higher multiples of 10 mc. The frequency for which the generator is tuned is called the fundamental. Twice the fundamental is called the second harmonic, three times the fundamental is the third harmonic, and so on.

To obtain a radio frequency needed during alignment you may divide that frequency by some whole number that gives a quotient within the tuning range of your signal generator. If you want 100 mc, and your signal generator goes only to 30 mc, you might tune the generator to 25 mc and use the fourth harmonic (100 mc) or you might tune to 20 mc and use the fifth harmonic. You don't have to do anything special to obtain these harmonic frequencies, all of them are there all the time. In the accompanying table are listed generator dial frequencies whose various harmonics are at some of the radio frequencies commonly used for alignment of tuners in f-m receivers. Frequency settings enclosed by parentheses in the table are not exact divisions of the desired radio frequency, and it would be difficult to tune the generator to these fundamentals with sufficient accuracy for alignment.

SIGNAL GENERATOR SETTINGS WHICH FURNISH HARMONICS FOR ALIGNMENT

Number of the Harmonic Being Used	RADIO FREQUENCIES IN MEGACYCLES REQUIRED FOR ALIGNMENT These are harmonics of generator dial frequencies listed below. 86	88	90	98	106	108	110
	Frequencies At Which Generator Dial Is Set						
Second	43.00	44.00	45.00	49.00	53.00	54.00	55.00
Third	(28.67)	(29.33)	30.00	(32.67)	(35.33)	36.00	(36.67)
Fourth	21.50	22.00	22.50	24.50	26.50	27.00	27.50
Fifth	17.20	17.60	18.00	19.60	21.20	21.60	22.00
Sixth	(14.33)	(14.67)	15.00	(16.33)	(17.67)	18.00	(18.33)
Seventh	(12.29)	(12.57)	(12.86)	14.00	(15.14)	(15.43)	(15.71)
Eighth	10.75	11.00	11.25	12.25	(13.25)	13.50	13.75

The higher the harmonic frequency the less becomes the generator voltage. However, many service generators will furnish ample voltage up to at least the tenth harmonic.

Whatever error you make in tuning the generator to a fundamental frequency will be multiplied by the number of the harmonic frequency being used. For example, if you make an error of 0.25 mc on the generator dial, and use the fourth harmonic, that harmonic frequency will be in error by four times 0.25 mc, or will be 1.00 mc out of the way.

A signal generator must produce frequencies which correspond closely to the dial markings, which is to say that calibration of the generator must be accurate. When the generator dial is set for 10.7 mc, for alignment of the demodulator and i-f transformers, the actual frequency should not differ by more than 0.01 mc either way from the indicated value. For the methods of alignment which we shall use for f-m receivers the calibration on carrier frequencies need not be quite so close. If the actual frequency is within 1.0 mc. of the dial setting between 88 and 108 mc the results ordinarily will be acceptable.

More important than absolute accuracy of calibration is the ability to return the generator to a frequency that has been used before. Assume that you have been using an actual frequency of 10.71 mc, which might be furnished with the dial pointer at 10.70 mc. It is essential that the actual frequency come back to precisely 10.71 mc when the dial is reset at 10.70 mc after you have used some different frequency between times. Otherwise you wouldn't be sure that several transformers or couplers were aligned to exactly the same frequency.

Fig. 39-3. Signal generator designed especially for alignment of f-m receivers. Frequency scale at upper right. Output meter at upper left.

The attenuator of the generator, which is the control for output voltage, should be capable of dropping this voltage nearly to zero. Many attenuators fail to meet this requirement. It is absolutely necessary that the cable through which generator voltage is delivered to receiver circuits be shielded. The inner conductor that carries signal voltages must be enclosed within but insulated from an outer braided metallic covering that is connected to chassis ground within the generator and whose receiver end is fitted with a clip for connection to ground or B-minus at the receiver.

Fig. 39-4 is a picture of a shielded cable for use with a signal generator. At the left and uppermost is the insulated prod on the end of the central conductor through which generator voltage comes to the receiver. To the braided metallic shield of the cable is attached a spring clip. At the right is a screw connector that attaches to the signal outlet on the generator.

The signal generator preferably will furnish r-f and i-f voltages which may be either amplitude-modulated at some audio frequency or else unmodulated and of unvarying amplitude. Although we are working with frequency-modulation receivers and sound sections there are certain tests described in following pages which may be made only with an audio modulated signal.

The central conductor or the "high side" of the generator output cable will be connected to tube grids, antenna terminals, and other points through a capacitor or resistor as directed in following detailed instructions. Usual practice is to make this connection to tube socket lugs or other suitable points on the under side of the chassis, with the chassis removed from its cabinet.

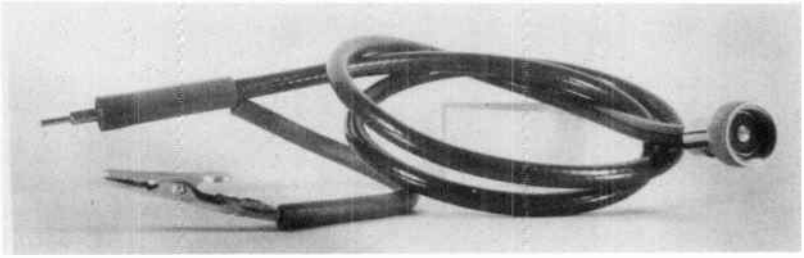


Fig. 39-4. Shielded output cable for a signal generator.

Sometimes it is more convenient to make test connections to grids or other tube elements from the top of the chassis. Many technicians feel that top connections are preferable because chassis metal then is interposed between possible radiation of signal fields from the generator cable and the circuit wiring that is underneath the chassis.

There are various ways of making top connections. One of the simplest is shown in Fig. 39-5, where connections have been made to two of the base pins on a tube. To make a connection, bare both ends of a short piece of insulated wire. Shape one end into a small loop just large enough to force onto a base pin. The insulation must come close to the pin so that when the tube is replaced in its socket the wire will remain insulated from chassis metal around the socket. The free end of the piece of wire then is used for an instrument connection. The wire should be left only long enough for convenient attachment of an instrument lead or clip.

A top connection may be made also with a test adapter such as made and sole for this express purpose. In Fig. 39-6 one of these test adapters is shown by itself at the left, and at the right is shown a similar adapter inserted in a socket, with a tube inserted in the top of the adapter. Adapters are available for octal-base tubes, lock-in tubes, and miniature tubes.

Extending out of the bottom of the adapter are pins just like those on the base of the corresponding tube. In the top of the adapter are openings just like those in a socket. The metal clip connector in each

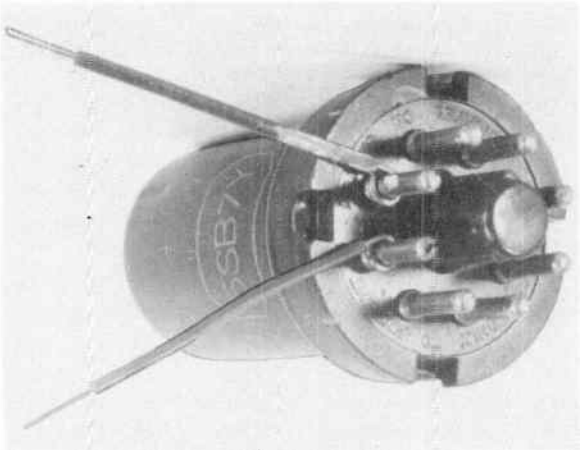


Fig. 39-5. Wires twisted around tube pins for top connections during tests.

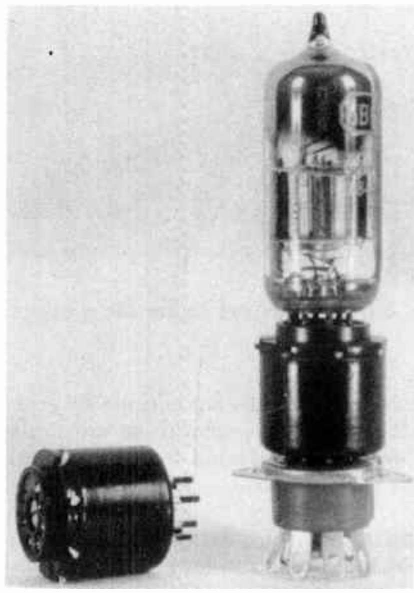


Fig. 39-6. A test adapter and how it is used with a miniature tube.

top opening is joined to the base pin directly below, and from each of these connections a short stiff wire extends out through the side of the connector. Test instruments may be connected to any one or more of these small extensions.

METERS FOR ALIGNMENT. During all processes of alignment we introduce voltages from a signal generator at some points in the receiver circuits and measure resulting output voltages at other points, thus determining the effect of various adjustments. Three types of meters may be used – an electronic voltmeter, a high-resistance d-c voltmeter, or an output meter which measures audio-frequency voltages.

Service types of electronic voltmeters, such as the one pictured in Fig. 39-7, are capable of making all measurements required during alignment. The cable and connectors shown attached to the left side of the instrument are for measuring direct voltages. At the right is attached another cable and a special "probe" for measuring alternating voltages at all frequencies from less than 60 cycles up to several megacycles, including all audio-frequency voltages.

The great advantage of electronic voltmeters over all other types is the very high resistance or impedance of the instrument itself. When making measurements of direct voltages the meter resistance ordinarily is something between five and twenty megohms, regardless of the range of voltage being measured. On a-c, a-f, and r-f measurements the instrument impedance decreases as the applied frequency rises, but the impedance always remains far greater than in any other type of meter used for similar service work.

When a meter having such high resistance or impedance is connected to or across any circuit other than one operating at high frequency the small meter current causes hardly any change in loading of the circuit, and indicated voltages are practically the same as with no meter connected. Meters of less resistance or impedance take more current from whatever source is feeding the measured circuit, and indicated voltages are lower than voltages which exist when no meter is connected.

In methods of f-m alignment which we shall use it is not necessary to make voltage measurements in

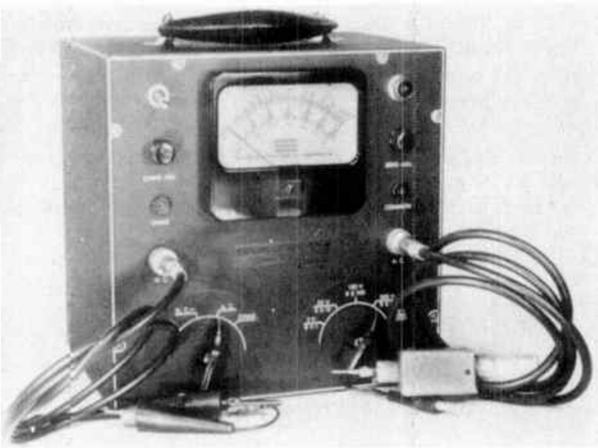


Fig. 39-7. Service type electronic voltmeter.

high-frequency tuned circuits while such circuits are operating. If, however, we did wish to make such measurements, an electronic voltmeter would be the only practicable type. Any other type would add so much capacitance, inductance, or both to the tuned circuit as to greatly alter the resonant frequency.

The second type of meter used for alignment, a high-resistance d-c voltmeter, is one whose internal

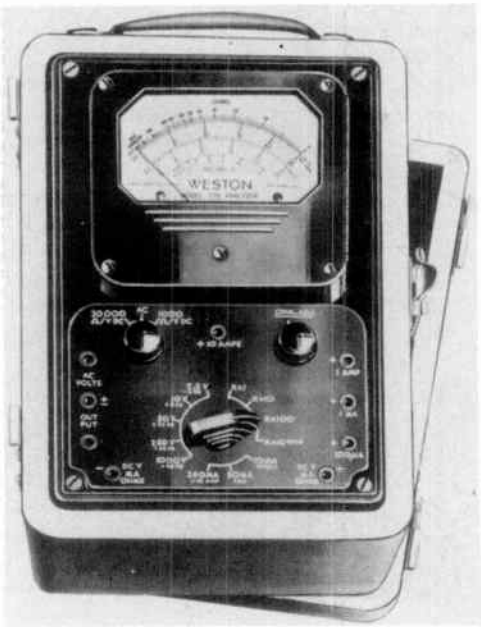


Fig. 39-8. High-resistance d-c voltmeter with provisions for also measuring d-c current and ohms of resistance.

resistance is 20,000 ohms or more per volt of maximum pointer deflection on each scale of the instrument. That is, when the meter is being used with a scale allowing maximum reading of 5 volts the internal resistance will be 5 times 20,000 ohms, or will be 100,000 ohms. On a 10-volt scale the internal resistance would be twice as much, or 200,000 ohms. On a scale reading up to 100 volts the internal resistance would be 100 times 20,000 ohms, or 2 megohms.

A meter having sensitivity of 20,000 ohms per volt always takes through itself a current of 50 microamperes at full scale deflection, regardless of which scale is being used. This you may prove by using our regular formula for current in relation to voltage and resistance. This formula is,

$$\text{Milliamperes} = \frac{1000 \times \text{volts}}{\text{ohms}}$$

If we use the meter values of 20,000 ohms and 1 volt the formula reads.

$$\text{Milliamperes} = \frac{1000 \times 1}{20000} = \frac{1}{20} = 0.05 \text{ ma} = 50 \text{ microamperes.}$$

Most of our alignment work is done with a meter scale of 3 volts, 5 volts, or whatever may be the lowest range of the meter. On these scales the internal resistances are rather low and the meter currents propor-

Fig. 39-9. Output meter for measuring voltages at audio frequencies.

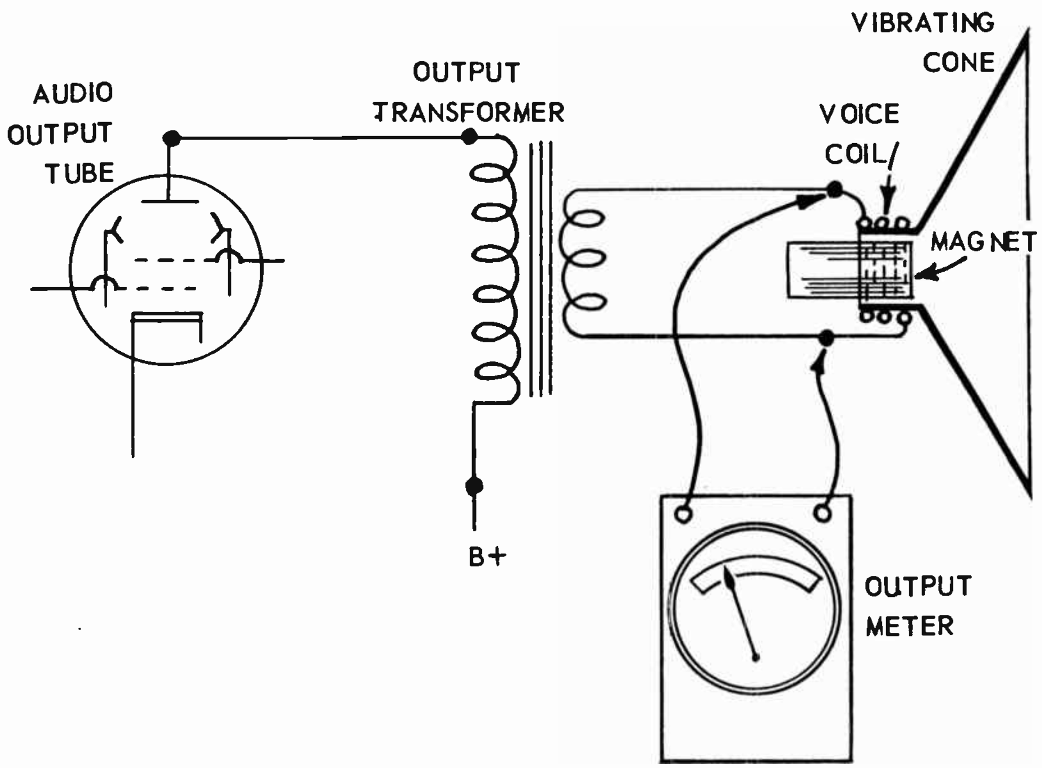


Fig. 39-10. Connections between an andio output tube and the voice coil of a speaker.

tionately high. Consequently, when using these low scales you should connect in series with one of the voltmeter leads a fixed carbon or composition resistor of something between 100,000 ohms and a half megohm. This will reduce the deflection of the meter pointer, and it will not indicate actual volts. But since nearly all alignment adjustments are for the purpose of obtaining either maximum or minimum readings, without reference to any particular number of volts, the operation of the meter is satisfactory under these conditions, and we will have increased the meter resistance to a value at which action of measured circuits will not be seriously altered by presence of the meter.

High-resistance d-c voltmeters for service work usually are incorporated in an instrument capable of measuring d-c voltages, a-c voltages, ohms of resistance, and sometimes currents in milliamperes or microamperes. Such a combination "analyzer" instrument is pictured in Fig. 39-8. Switches select the range of voltage or current to be measured, and also the dial scale to be used. An instrument which measures volts, ohms, and milliamperes usually is called a volt-ohm-milliammeter, abbreviated "VOM".

2 An output meter is a type designed especially for measuring audio-frequency voltages. One such instrument is illustrated by Fig. 39-9. The internal working parts or the "movement" ordinarily will be of the same type used in d-c voltmeters, or will be of the moving coil type. Inside the instrument are one or more small rectifiers of the contact type, called meter rectifiers. These rectifiers change the applied alternating audio-frequency voltages into one-way or direct voltages and currents that operate the meter movement.

The output meter illustrated has a terminal or jack at the upper left marked "Series Cond." Connected internally in series between this terminal and the rectifiers is a condenser or capacitor which allows only alternating currents and voltages to enter the meter even when the test leads are connected to a circuit containing both alternating and direct voltages. Such a circuit would be that connected to the plate of a tube, where there is a direct voltage from the B-supply and also alternating signal voltage. At the top center is a jack or terminal, marked positive and negative, to which may be connected a pure alternating voltage or any voltage in which it is desired to measure the effect of both direct and alternating components. The common terminal for the second side of any circuit is at the upper right. Down below the meter face is the selector switch.

Most output meters have one dial scale marked in decibels as well as other scales for a-f volts. A decibel is a unit of measurement of relative gain or loss of voltage, current, or power which may occur between different points in a circuit or between different circuits. Decibel measurements are used chiefly when working with audio-frequency amplifiers, speakers, and audio reproduction in general. We shall learn about these measurements later on. They are not used for alignment work.

For measuring the audio output of the receiver or sound section during alignment the output meter usually is connected across the voice coil of the speaker. This voice voil is shown in Fig. 39-10. It is a coil of a few turns of wire fastened onto the vibrating cone of the speaker. Inside the small end of the cone and the voice coil is the end of a permanent magnet or an electromagnet of diameter small enough to allow free movement or vibration of the cone. Reaction of the alternating (audio frequency) magnetic fields around the coil and the steady field of the magnet causes the voice coil and attached cone to vibrate at audio frequency.

3 Audio-frequency currents for the voice coil come from the secondary of an iron-core output transformer whose primary winding is connected between the plate of the audio output tube and the B-plus supply lines. The transformer has a large step-down ratio from primary to secondary. Consequently, there is high a-f voltage and small a-f current in the primary, while in the secondary there is very low a-f voltage but a large a-f current. The low voltage and high current of the secondary are also in the connected voice coil.

Should the audio output be so weak that voice-coil voltage will not give readable deflections of the output meter this meter may be connected across the primary winding of the output transformer, where voltage is many times greater than in the secondary circuit. Instead of connecting the meter leads to the ends of the secondary winding of the transformer one lead may be connected to the plate of the audio output tube and the other lead to B-plus or to B-minus or to chassis ground, wherever there are readable indications. With any of these connections to the tube plate or transformer primary it is absolutely necessary to have a capacitor in series with the meter. If no capacitor is built into the meter one must be connected externally. Use only a paper dielectric capacitor having a voltage rating well in excess of the B-supply voltage applied to the plate circuit. The capacitance should be 0.1 mf or preferably 0.5 mf.

Fig. 39-11 shows where to find voice coil connections on a typical speaker. The vibrating cone is at the left. Immediately back of the center of the cone is the winding for an electromagnet. Many speakers have a small permanent magnet in this position. On top of the magnet frame is mounted the output transformer, from which two wires go to lugs on an insulating support. One wire shows in the picture. From these lugs two other flexible wires go to the vibrating cone and to the voice coil. The lugs are the terminals of the voice coil to which you attach the leads of the output meter. Oftentimes there are long extension lugs provided for the express purpose of connecting an output meter to the voice coil. One such extension lug is shown in the picture, bent down so that its end is visible against the black background of the cone.

Whenever you use an output meter, no matter where it is connected, keep the volume control of the receiver at maximum. This allows maximum meter indications with minimum loading of the sound circuits.

5 No matter what kind of a meter you are using, always work with the lowest voltage scale or the scale showing the least maximum voltage. When you don't know how high the voltage may be to begin with it is advisable to first set the meter for one of the high-voltage scales. If the reading then is greater than could be measured on the lowest scale, reduce the voltage output of the signal generator until the lowest scale can be used. Always it is advisable to use a generator voltage output low enough that maximum readings on any meter do not come much if any above half scale.

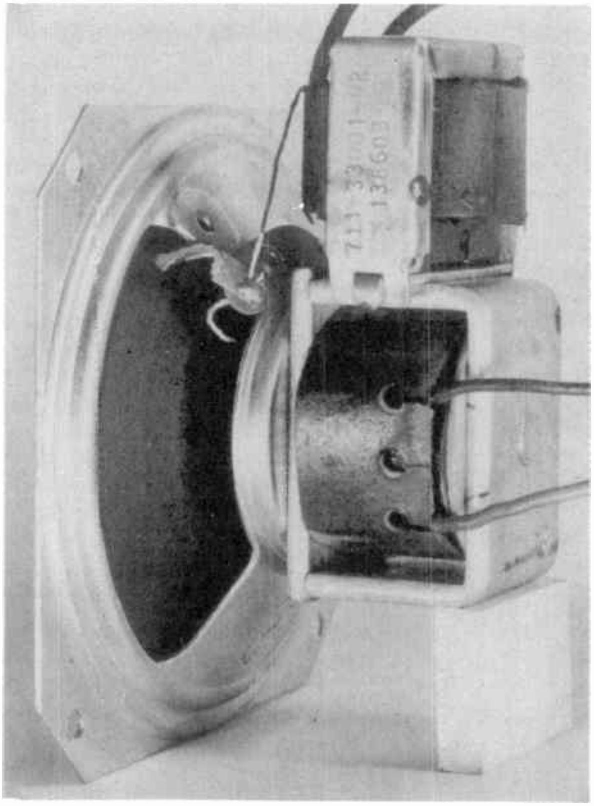


Fig. 39-11. Voice coil connection of a speaker. A similar connection is on the hidden side of the transformer.

DEMODULATOR OUTPUT VOLTAGES. Probably the most critical step in alignment of an f-m receiver or the sound section of a television receiver is adjustment of the demodulator transformer. The job becomes relatively easy, and good results will be assured, only when you have a clear understanding of what is happening to voltages in the load circuits of the demodulator.

To make some instructive measurements we shall use the hookup of Fig. 39-12. This is not a hookup for alignment, although it could be used for some steps, we merely are going to watch the performance in slow motion rather than with a signal rapidly deviating in frequency. A signal generator is connected through a capacitor *C* to the grid of the i-f amplifier or limiter tube preceding the demodulator transformer. A high-resistance d-c voltmeter or electronic voltmeter is connected from the audio output point *X* to ground. Note that nothing is being said as to whether we are working with a ratio detector or a discriminator, they both work the same so far as audio output is concerned. We may assume that primary and secondary of the demodulator transformer are accurately tuned to 10.7 mc.

First we tune the signal generator to some frequency such as 10.2 or 10.3 mc, well below the center frequency, and note the voltage indicated by the meter. This voltage will be small, and negative with reference to ground as at *A* on the graph of Fig. 39-13. The applied signal voltage is at a frequency so far from the tuned resonant frequency of the transformer that there is small response.

As the generator frequency is slowly increased the voltage will become more and more negative, as from

A to *B* on the graph. The voltage remains negative because, at frequencies below resonance or below the center frequency, one of the demodulator diodes is receiving a greater out-of-phase voltage than the other diode and is passing more current. This will be the diode that produces a negative swing of audio output voltage during normal operation. The voltage increases, or becomes more negative because the applied frequency is coming closer to resonance in the transformer, and the response is increasing.

When we reach the frequency at *B* on the graph the voltage reaches its negative peak and thereafter becomes less negative. Now the phase of voltages on the diodes is changing to cause the first diode to conduct less current while the other diode commences to conduct more current. These rectified currents in the diode circuits and the demodulator load resistors are direct currents, which is the reason why we may measure their effects with a d-c voltmeter.

By the time the applied frequency has been increased to 10.7 mc the diode currents and voltages are equal, with the result that audio output voltage is zero, as at *C* on the graph.

With further increase of frequency, from the signal generator the diode that formerly had greater conduction has less conduction, and the other diode now has the greater conduction. This, of course, reverses the polarity of the audio output voltage and this voltage being measured by the meter commences to increase in the positive direction from *C* to *D* on the graph. At *D* we have peak positive voltage. With increase of frequency beyond this point we are getting so far from resonance that the response commences to drop, and voltage gradually falls back to a very small value at *E*.

If the applied frequency is decreased the output voltage will go through all its changes in reverse, with

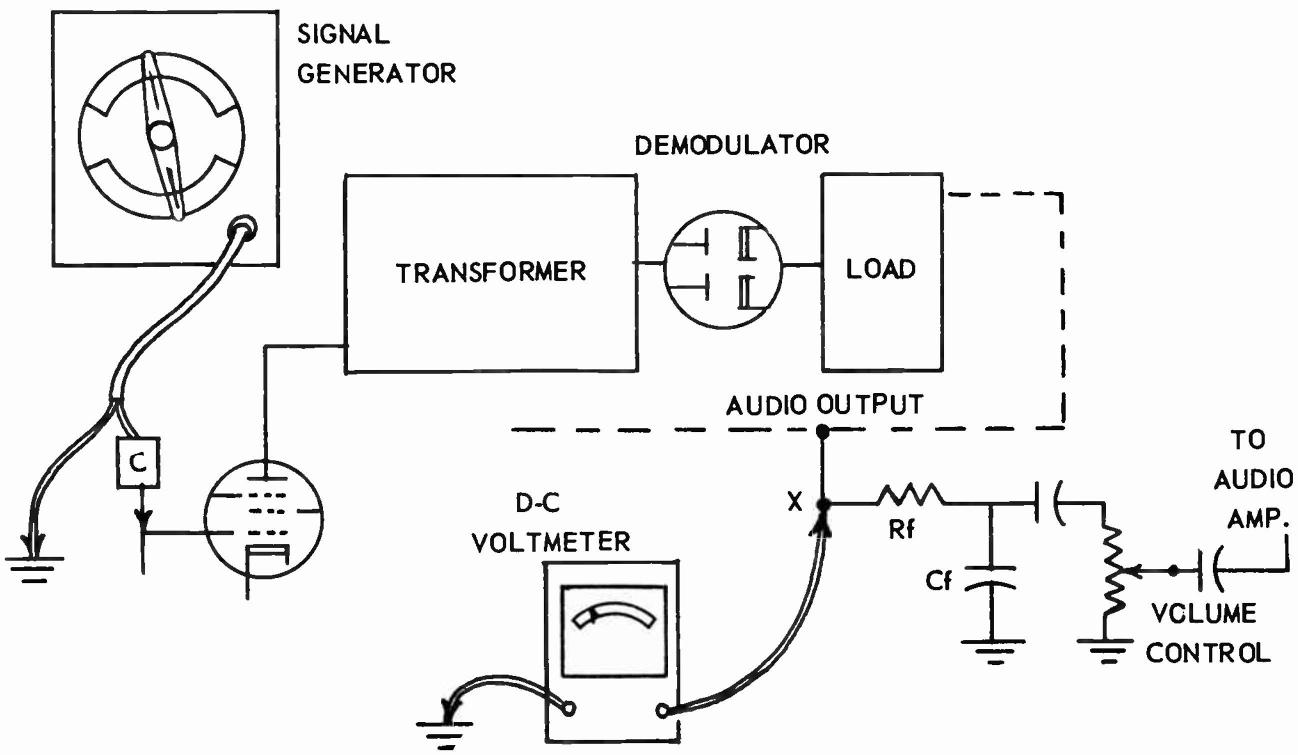


Fig. 39-12. Connections for measuring d-c voltage output of a demodulator.

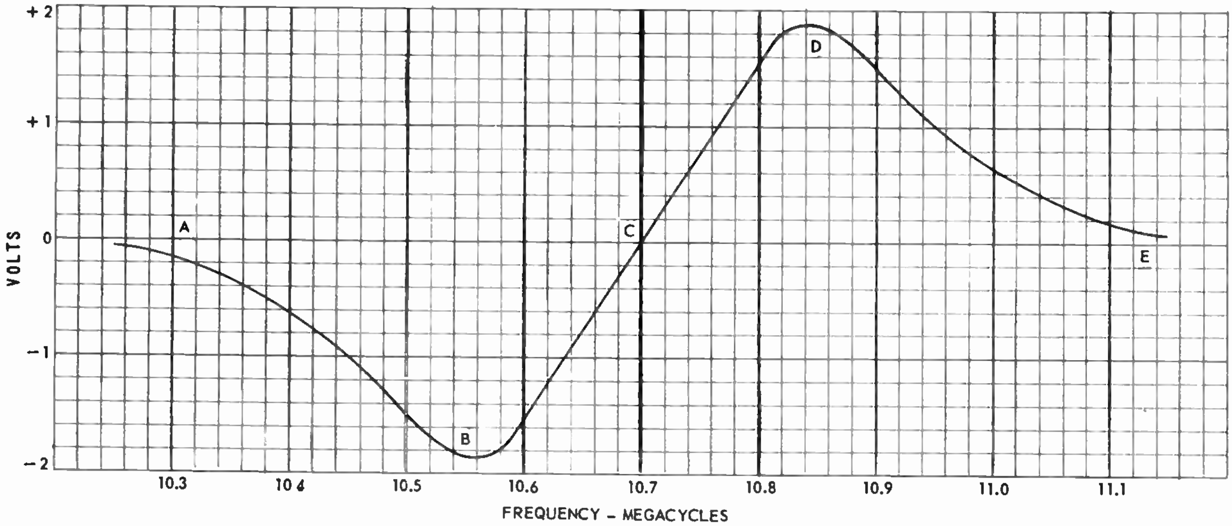


Fig. 39-13. An S-curve showing changes of demodulator output voltage when there is deviation of frequency.

variations as from *E* back to *A* on the graph. If there were to be frequency deviation at an audio rate the voltage would alternately swing through the zero, negative, and positive values at the same rate. We would have alternating (negative and positive) voltage at the audio frequency of deviation. The curve which we have developed in Fig. 39-13 is called an S-curve because it is shaped like a letter S lying on its side. S-curves for ratio detectors and discriminators are alike, you could not tell which kind of demodulator is being used by looking at its output curve.

The voltages at the two peaks of the S-curve will increase when generator output voltage is increased and will decrease when generator voltage is lowered. The same thing happens with change of received signal strength during normal reception. The peak voltages of audio output increase and decrease with greater and less amplitude of the f-m signal applied to the demodulator. This is the reason why automatic controls for amplification in r-f and i-f amplifiers often are called automatic volume controls. They do limit the amplitude of signals reaching the demodulator and thus limit the maximum peaks of audio output voltage.

The S-curve developed in Fig. 39-13 is perfectly straight throughout a frequency range of about 100 or more kilocycles (0.1 or more megacycles) each way from the center frequency. When this is true there will be equal audio responses for equal deviations all the way to maximum f-m broadcast values of 75 kilocycles each side of the center frequency. In checking the demodulator alignment for accuracy it is rather common practice to measure negative and positive voltages with the signal generator detuned to equal numbers of kilocycles below and above the center frequency. These voltages should be equal, or very nearly so.

It is desirable also that voltages be equal at the negative and positive peaks. This equality is not really necessary provided the center part of the response is straight as far as the limits for maximum deviations. When the peaks show decided difference or when the central part of the response is not straight, correction is made by careful adjustment of the trimmer or movable core of the transformer primary.

The secondary trimmer or movable core is adjusted to bring zero voltage, or the center of the S-curve, to the center frequency. With any type of meter measuring audio output the object is to obtain a zero reading when the applied signal is at the center frequency. You should note from Fig. 39-13, and keep in mind here-

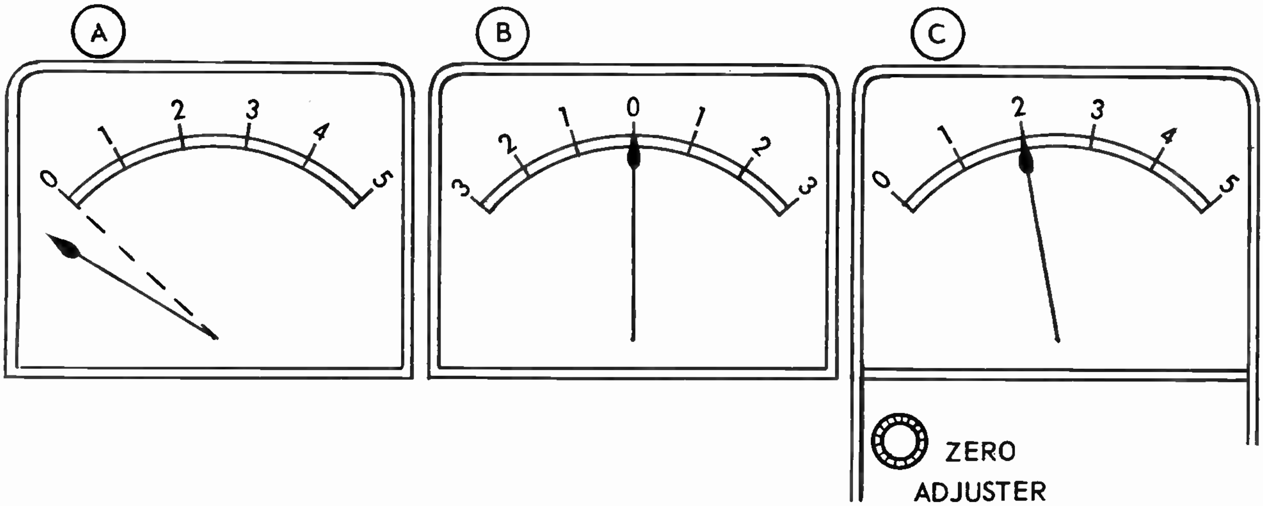


Fig. 39-14. How zero output voltages are indicated by various meters.

after, that voltage of zero or very near zero may be obtained at any of three different frequencies. At a low frequency, lower than point *A*, we would have zero voltage, and also at a frequency higher than point *E* we would have zero voltage. Neither would be the zero at which adjustment is aimed. The zero we always look for is the one at the center frequency, or at *C* on the S-curve.

When the primary of the demodulator transformer is tuned at or very close to the center frequency it becomes possible to adjust a trimmer or movable core for the secondary to obtain zero outputs at any of three different settings. One adjustment will be for resonance below the correct center frequency, another will be for resonance above the center frequency, and the third will be the correct one – for resonance at the center frequency. It is quite easy to determine whether you are using an incorrect adjustment or the correct one. At the correct setting the least change of adjustment either way will cause a rapid shift between negative and positive voltage outputs on a d-c meter. If you are using an a-c output meter the voltage will rise very rapidly on each side of the correct adjustment, and only slowly on the other adjustments.

There may be some difficulty in determining a true zero voltage with ordinary types of d-c voltmeters. One method is shown at *A* in Fig. 39-14. The pointer will rise from zero in the usual way on positive voltages, and will stand at zero for zero voltage provided it stands there when the meter is not connected to any source of voltage. On negative voltages the pointer can move a short distance to the left of zero. Then the pointer will pass the zero mark when applied voltage is at the center frequency, if alignment is correct. If negative voltage is to be measured it will be necessary to reverse the connections of the meter leads to the receiver circuits.

Some d-c voltmeters have a zero center scale, as at *B*. The pointer deflects to the left on negative voltages, and to the right on positive voltages. It is easy to precisely identify zero voltage, and to measure voltages in both polarities. Some electronic voltmeters provide zero-center readings with the selector switch in one of its positions.

A convenient method for use with any electronic voltmeter is shown at *C*. On meters of this type there is a zero adjuster knob that will move the pointer quite a ways up the scale when the instrument is turned on, but not connected to any voltage source. By initially setting the pointer at some easily recognized point on the scale this point may be considered as the zero reading and it becomes possible to read voltages which are either negative or positive.

COMMERCIAL TRADES INSTITUTE

Even when the pointer of an electronic voltmeter is not shifted from its normal zero position it is possible to identify the true zero by allowing the pointer to move slightly to the left of the regular zero point, as at *A* in Fig. 39-14. Nearly all electronic voltmeters have a polarity reversing switch. With this switch in one position the pointer reads up scale on positive voltages, and in another switch position the pointer reads up scale on negative voltages – all without changing the connections of the meter leads to the measured circuits.

a When making measurements with a d-c voltmeter it is necessary that no capacitor come between the audio output point of the demodulator and the meter, since a capacitor would not pass even the small direct current required by the meter. In earlier diagrams we have identified the audio outpoint as *X*. This point is shown again in the ratio detector circuit at the left in Fig. 39-15. In this particular modification of the ratio detector point *X* is conductively connected to the high side of the volume control, and d-c output voltages could be measured at the volume control. In many circuits there is a capacitor between point *X* and the volume control. Then no place on the volume control could be used for d-c output voltage measurement.

The d-c output point *X* for one style of discriminator circuit is shown at the right in Fig. 39-15. Here there is a capacitor between *X* and the top of the volume control, which would make it impossible to make d-c voltage measurements at any point on the control.

As mentioned before, the secondary of any demodulator transformer is aligned for zero voltage at the audio output point. The primary of the transformer is aligned for a maximum voltage when the applied signal is at the intermediate center frequency. In demodulator circuits having two resistors on the load side, the voltage for primary alignment may be measured with the voltmeter connected across either one or the other of these load resistors rather than at the audio output point.

In the ratio detector circuit at the left in Fig. 39-15 the voltmeter could be connected from chassis ground, or B-minus, to point *A* or to point *B*. in the discriminator circuit at the right the voltmeter could be con-

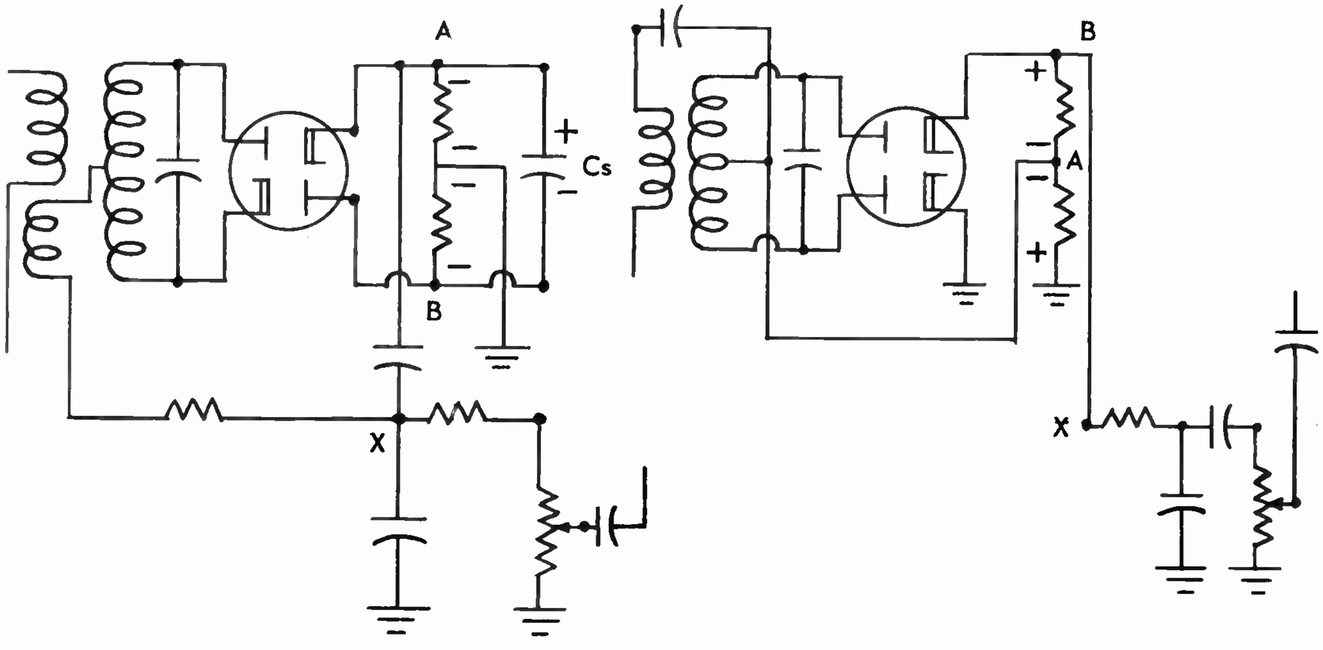


Fig. 39-15. Points at which meters are connected to demodulator load circuits during alignment.

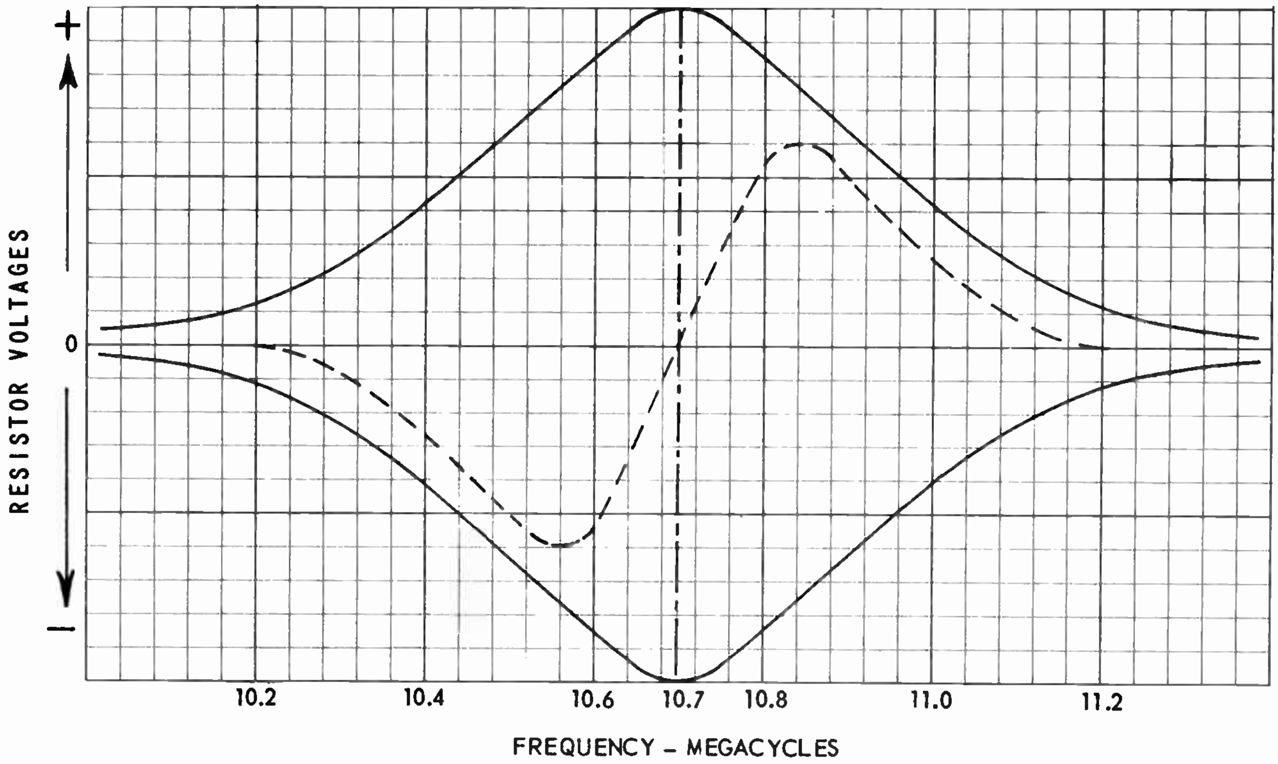


Fig. 39-16. Full-line curves show voltage changes across one load resistor when there is frequency deviation. An S-curve is shown for reference.

nected between ground and point *A*, across the lower resistor, or else between points *A* and *B* to measure voltage across the upper resistor.

The voltage across either load resistor should reach its peak value when the signal from the generator is at the intermediate frequency. Typical changes of voltage across load resistors are shown by Fig. 39-16. A positive voltage, with a positive peak, may be shown across one of the resistors while a negative voltage and peak are shown across the other one. The polarities will be as marked at the ends of the resistors in Fig. 39-15. An S-curve is included in Fig. 39-16 to show the relations between this audio output voltage and the separate voltages across the two load resistors.

INSTRUMENT CONNECTIONS. When you read instructions for alignment it will be specified that meters are to be connected to certain points, as across a voice coil. The same resnlts will be obtained with connections made to any other points from which there is a direct conductive lead or wire going to the specified points. You will use connections which are most convenient, this being a matter of receiver design and construction. For many adjustments on radio detector circuits it will be specified that the meter is to be connected across the large capacitor in the detector load, which is capacitor *Cs* in Fig. 39-15. The same results would be had with the meter connected to the cathode and plate on the load side of the diodes, or to the outer ends of the two load resistors, since these other points are directly connected to the two terminals of the large capacitor.

There is not so much leeway in making connections of the signal generator leads in f-m circuits, where

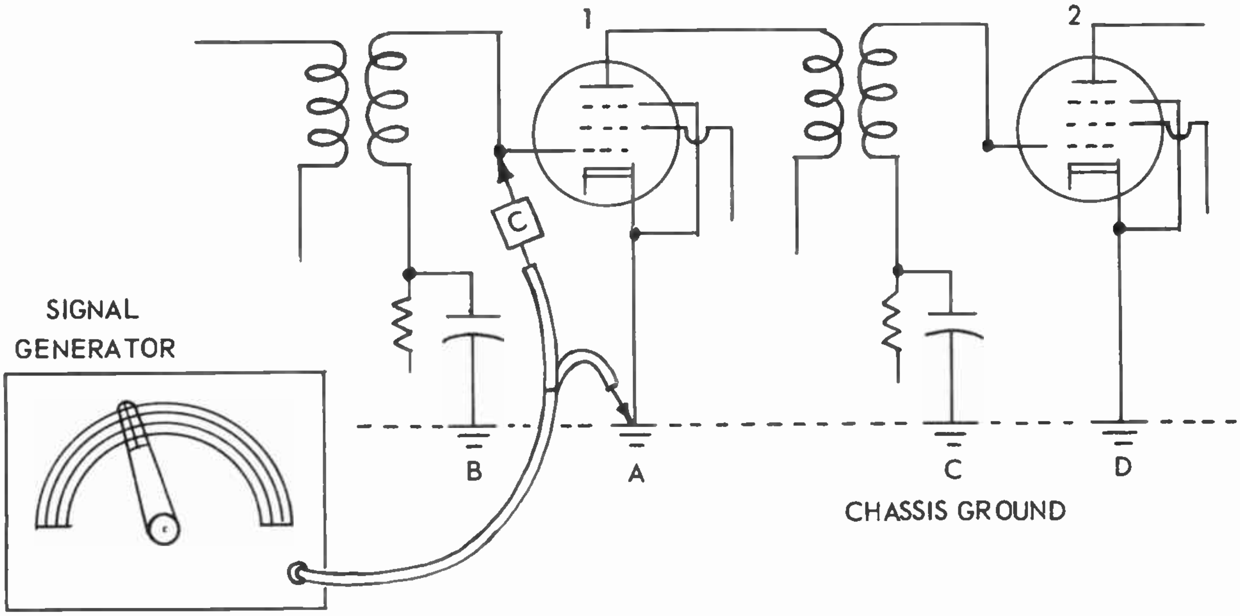


Fig. 39-17. How a signal generator should be connected during alignment.

we are working with frequencies of many millions of cycles. Even short lengths of conductors of any kind have appreciable inductive reactance at these frequencies, and the low capacitive reactances between small lengths of conductors allow considerable leakage of signal voltages.

Fig. 39-17 illustrates some principles to be observed when making connections of generator leads. Assume that the high side of the generator is to be connected to the grid of tube number 1. Make the connections as close as possible to the grid pin, at the pin itself or at the grid lug on the socket. Then connect the ground lead from the generator so that high-frequency signal voltages introduced from the generator follow the shortest possible path between the high side connection and ground. The ground connection may be made at *A*, where the cathode of tube number 1 connects to chassis ground, or it might be made at or near point *B*, where the bypass capacitor from the grid coil connects to chassis ground.

It would be poor practice to make the generator ground connection at points such as *C* or *D* when feeding a signal into tube number 1. Such connections would be far from the grid circuit where the signal is introduced, and high-frequency currents would have to flow through all the chassis metal from *C* or *D* *back* to *A* or *B*. This would allow the signal fields and voltages to go directly into many circuits where they are not wanted.

Keep the leads from the signal generator as far as possible from all receiver wiring, even though the high-side lead is protected by the grounded shield. Be sure to keep the leads for the signal generator and those for the meter well separated.

With all the instrument connections correctly made we still are not quite ready to follow the step-by-step instructions for aligning the demodulator, the i-f amplifier, and the tuner, as these instructions will be given in the following lesson. The one thing still to be done is to turn on both the receiver and the signal generator, also an electronic voltmeter if this type of meter is to be used, and let everything warm up for at least ten minutes and preferably more before attempting to make any adjustments. Never forget the warm-up period.

TELEVISION

LESSON NO. 40

METHODS FOR F-M ALIGNMENT

Training Your Mind
C.T.I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

DO NOT TEAR — CUT ALONG THIS LINE AND SEND IN FOR GRADING.

COMMERCIAL TRADES INSTITUTE

Training Your Mind

C. T. I.

Perfecting Your Skill

Chicago, Ill.

LESSON NO. 40

METHODS FOR F-M ALIGNMENT

The general procedure of alignment follows the same pattern whether we are working on a-m, f-m, or television receivers. Of course, we employ different frequencies and look for results appropriate to the type of receiver, but almost everything you learn about any one of the jobs proves directly or indirectly helpful on the others. This fact will become apparent as we proceed in this lesson with steps to be taken during alignment of f-m receivers, for many of these steps are only slight modifications of operations performed when aligning a-m receivers. Later, when we come to alignment of television receivers, your familiarity with adjustments on sound receivers will make it relatively easy to handle both sound and video sections of the television receiver.

No matter what kind of receiver you work on, if it is a transformerless type or has series heaters across the line, you should make connection to the power line through an isolating transformer if such a transformer is available. Otherwise you should take care to have the chassis side of receiver circuits connected to the grounded side of the power line, thus minimizing the effects of a hot chassis. Also, to avoid the possibility of wrecking the attenuator of your signal generator when working with receivers of this type it is advisable to connect the low side of the generator output cable to the chassis or B-minus through a fixed capacitor of at least 0.01 mf capacitance, rated at 150 volts or more, rather than making a direct conductive connection.

All alignment adjustments normally are made with signal voltages from a generator. Other signals coming into the receiver at the same time can cause nothing but trouble. Consequently, you should disconnect any outdoor, indoor, or built-in antenna regularly used with the receiver.

In addition to disconnecting the antenna, set the tuning dial of the receiver to some frequency on which no nearby f-m station operates. Setting the dial for either maximum or minimum frequency, or as far as the knob will turn in either direction, often proves satisfactory. After the generator and meter are connected for testing try moving the receiver tuning dial a little ways in either direction from this preliminary setting. If this causes any change in the meter reading, when connected for demodulator or i-f amplifier alignment, you are getting beat frequencies between the r-f oscillator and some external signal, and these beats will lead to incorrect alignment. If you can find no dial position with which slight movement does not affect the meter reading it is advisable to disable the r-f oscillator while aligning the demodulator and the i-f amplifiers.

There are various ways of disabling the r-f oscillator during alignment. If the receiver has a separate oscillator tube, not combined with the mixer, and if tube heaters are in parallel, simply remove the oscillator tube from its socket and leave it out until you are ready to align the tuner after finishing with the demodulator and i-f amplifiers.

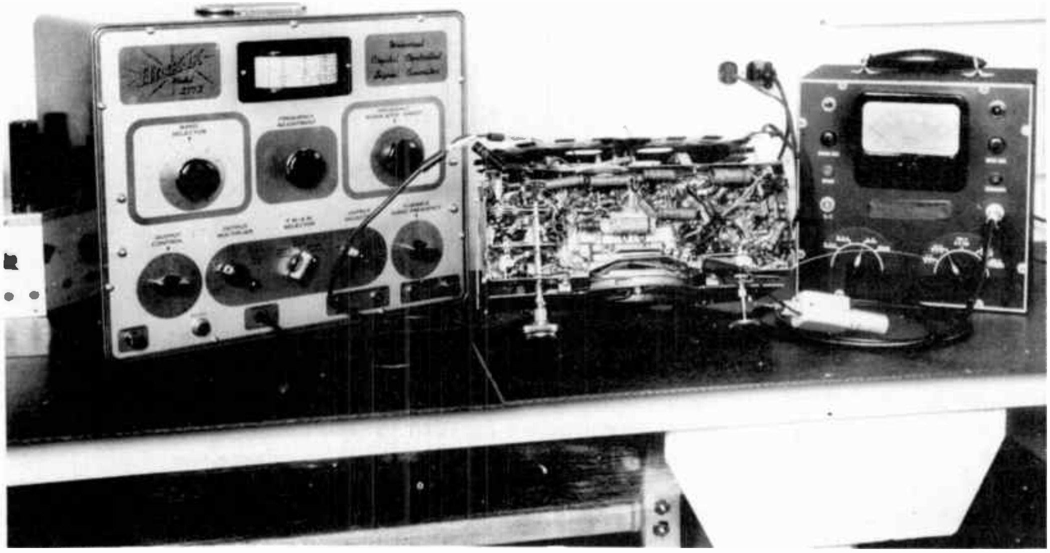


Fig. 40-1. Alignment setup for combination fm-am receiver, using signal generator and electronic voltmeter.

If there is a separate oscillator tube, but the tube heaters are in series, you can remove the oscillator tube and complete the heater circuit otherwise. One way of completing the heater circuit is to use an old tube in which the heater still lights but which has proven otherwise defective in operation. All pins except the two for the heater may be cut off short with diagonal cutting pliers, or you may cut off only the grid pin or only the plate pin. Of course, the old tube must have the same rating for heater volts and amperes as the tube it temporarily replaces.

Another way of completing the heater circuit is by use of a fixed resistor whose pigtail leads are pushed down into the socket holes that take the heater pins of the oscillator tube. It may be more convenient to fit the resistor pigtails into a test adapter, and insert the adapter in the oscillator socket. Resistance of the substitute resistor, in ohms, must equal the quotient of dividing the rated volts by the rated amperes for the heater of the oscillator tube, or a value within 10 per cent of this number of ohms. The wattage rating of the resistor should be at least twice the product of multiplying the rated volts by the rated amperes of the oscillator tube.

If the oscillator function is performed by one section of a twin triode or by a converter this tube must remain in its socket. Then you can connect the grid of the oscillator section to chassis ground or B-minus through a fixed capacitor of 0.001 mf or greater capacitance. This does not merely detune the oscillator, it stops the oscillation. The same result may be obtained by using the capacitor to ground the stator side of the oscillator tuning capacitor or the high side of an oscillator tuning coil.

5 If you are in doubt as to whether an oscillator circuit really is oscillating measure the grid voltage of the tube. While oscillation continues, the grid will be highly negative, usually about 10 volts negative with reference to the cathode or to ground. When oscillation stops the grid voltage will drop to near zero. Whenever you stop the r-f oscillator while aligning the demodulator and i-f amplifiers be sure to place it back in operation before attempting to align the tuner.

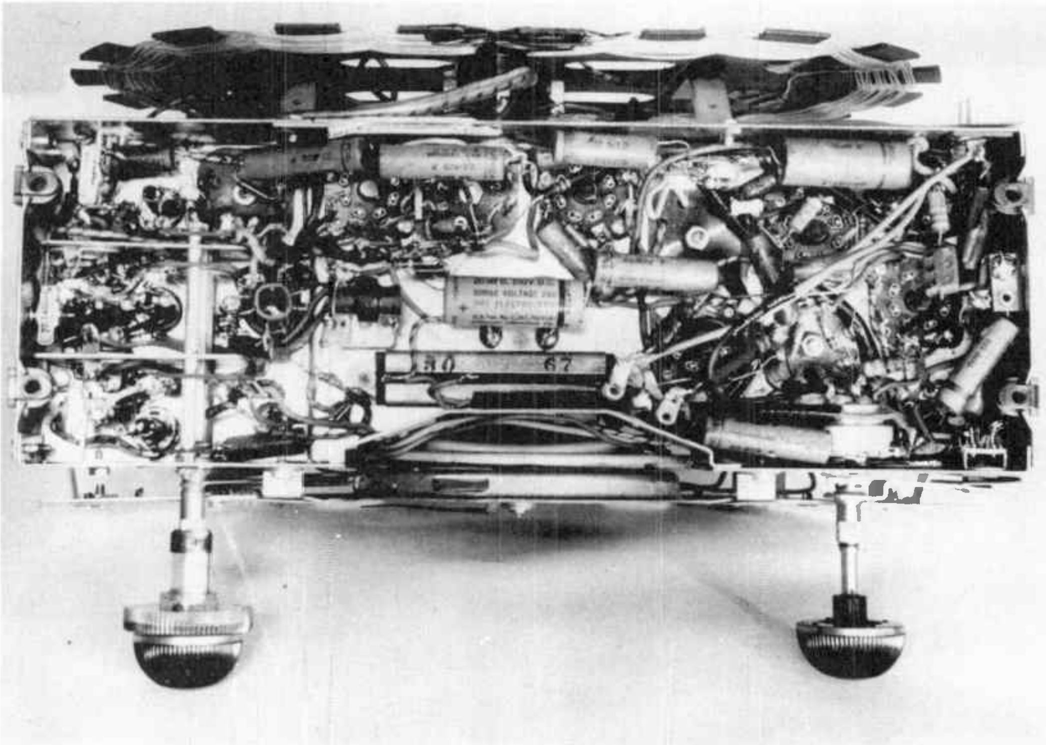


Fig. 40-2. Compact wiring underneath a chassis makes it necessary to carefully trace the wiring when making test connections.

DEMODULATOR ALIGNMENT. There are several ways of aligning the demodulator transformer by using a signal generator and either a high-resistance voltmeter or an electronic voltmeter. The same methods are used whether the demodulator is a ratio detector or a discriminator with limiter. The general scheme of instrument connections is shown by Fig. 40-3 for one of the methods in common use. Step-by-step procedure is as follows:

1. Use the signal generator without audio modulation. Tune the generator to the intermediate center frequency, which is 10.7 mc in all receivers of recent design. Connect the high side of the generator output cable through a fixed capacitor of 0.01 to 0.05 mf to the grid of the tube which precedes the demodulator transformer, as at *A* in the diagram. Connect the low side of the output cable to chassis ground or B-minus, as at *B*. Which of these latter connections is used depends on construction and wiring of the receiver. With B-power furnished through a power transformer, and heaters in parallel, make the connection to chassis ground. With transformerless receivers make the low-side connection to a B-minus point in the wiring, through a 0.01 mf capacitance.
2. To measure the demodulator output use either an electronic voltmeter or a high-resistance d-c voltmeter. When using the high-resistance d-c voltmeter connect in series with the ungrounded lead a fixed carbon resistor of 0.2 to 0.5 megohm resistance.
3. During alignment of the transformer primary connect the meter across either of the two load resistors. This connection would be from *C* to *D* or *C* to *E* in the diagram. If one end of either load resistor is grounded, connect the ground lead of the meter at the grounded end of the resistor or as near there as possible, and connect the high-side lead of the meter at or close to the other end of the same load resistor.

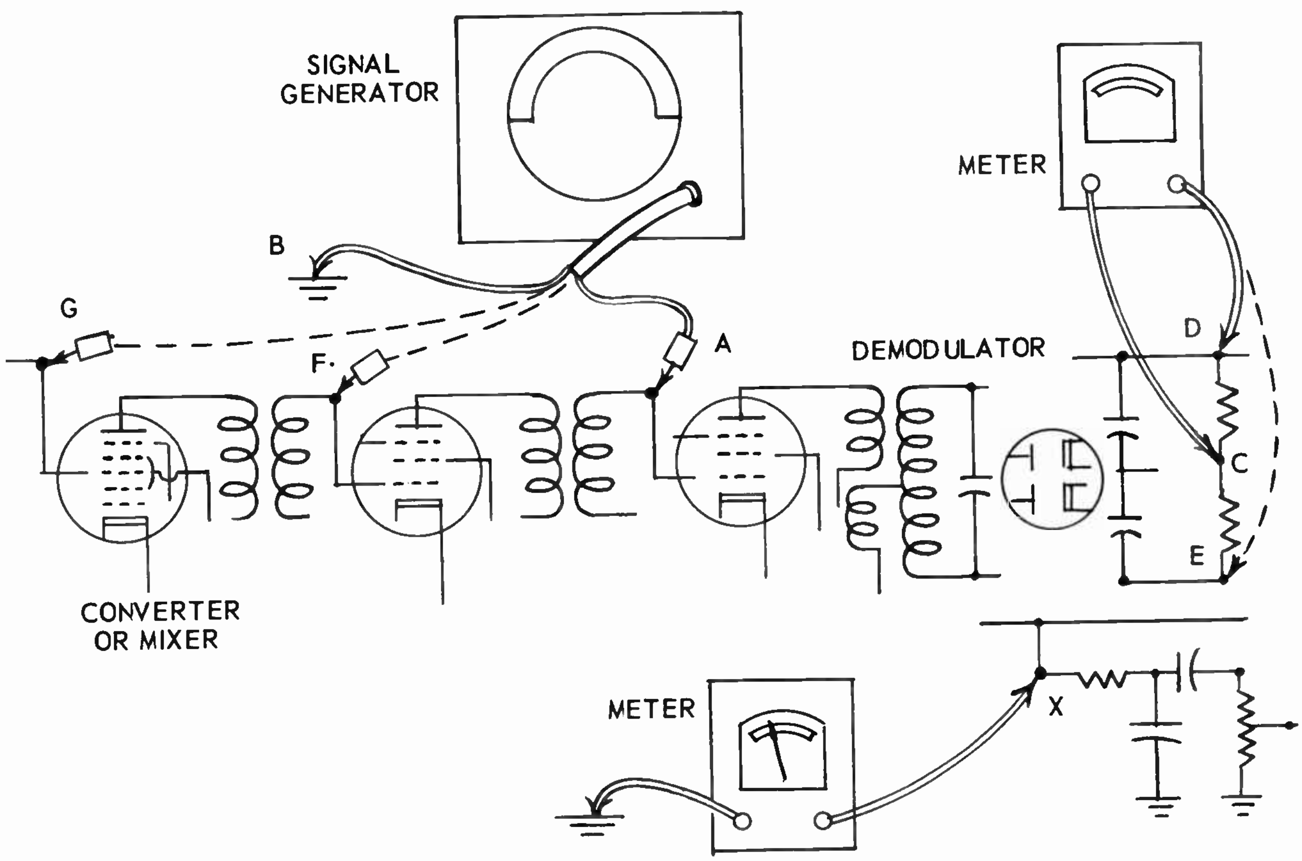


Fig. 40-3. Instrument connections for alignment of a demodulator transformer.

4. Proceed to align the primary trimmer or movable core to obtain maximum reading on the meter. Remember to use the lowest meter scale and the least generator output voltage that allows identifying peak voltage.
5. During alignment of the transformer secondary change the meter connections. Connect the high-side meter lead to the audio output point of the demodulator, point X in the diagram. Connect the low-side meter lead to chassis ground or B-minus in the wiring, according to how the receiver is constructed.
6. Align the secondary trimmer or movable core to obtain a zero meter reading which occurs in between positive and negative peak readings. This is a highly critical adjustment and requires care. Use enough voltage output from the signal generator to get distinct positive and negative peaks, then make the alignment for zero. This completes adjustment of the demodulator transformer.

Signal generator connection *A* would be to the limiter grid if the receiver employs a discriminator or limiter. Otherwise this connection would be to the grid of the last i-f amplifier. When used with ratio detectors this i-f amplifier often is called the driver.

If the i-f amplifier transformers are not far out of alignment, the high-side of the signal generator may be connected to the grid of a preceding i-f amplifier, as at *F*, or to the signal grid or r-f grid of the converter

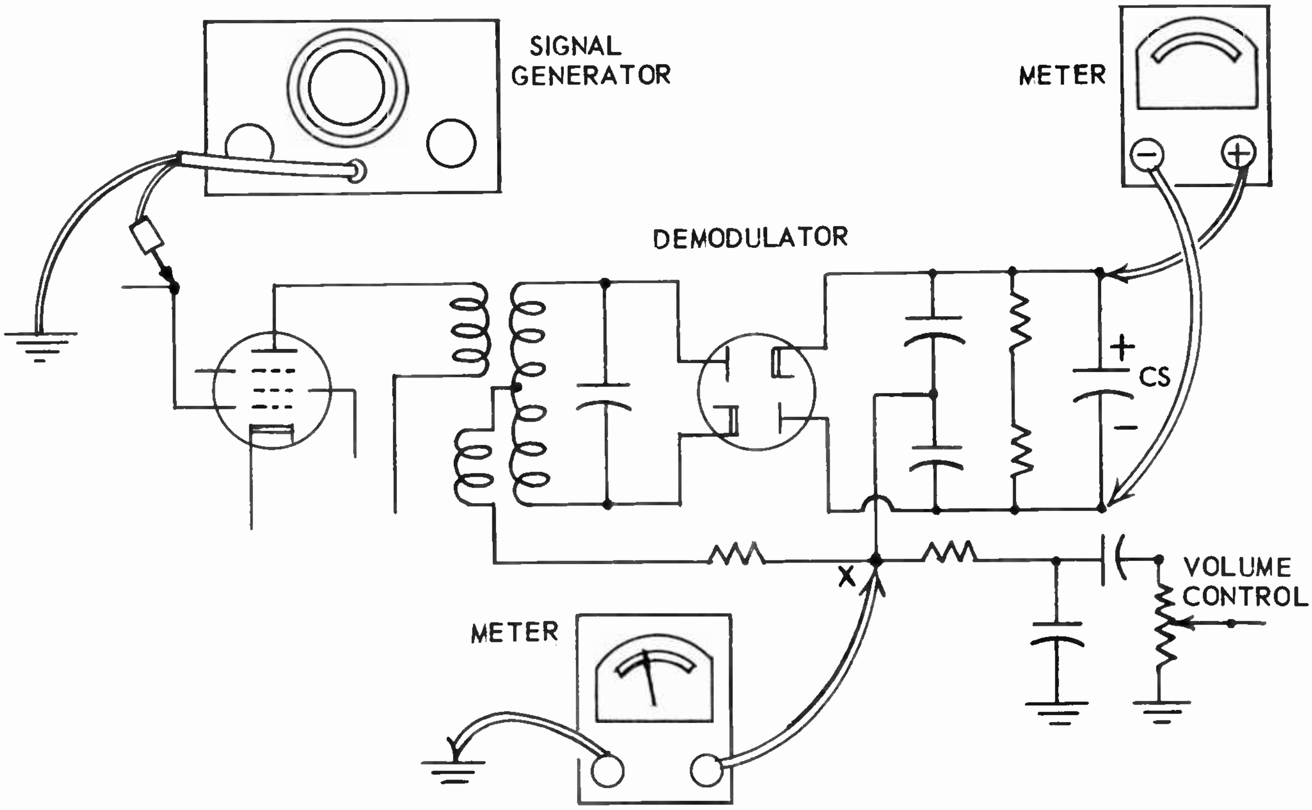


Fig. 40-4. Connections for alignment of a ratio detector transformer.

or mixer tube, as at *G* in Fig. 40-3. Either of these connections increases the strength of the signal reaching the demodulator for any given voltage output from the generator.

Oftentimes the signal generator connection is made to the grid of the converter or mixer tube and left there during alignment of the demodulator and also the i-f transformers. The signal generator will be connected to the grid of the converter or mixer for final alignment of the i-f transformers, and time is saved by using this connection for the demodulator also, provided the i-f stages will pass the signal voltage required at the demodulator.

A modification of the method just described is as follows.

1. Same as in first method.
2. Same as in first method.
3. For alignment of both primary and secondary in the demodulator transformer connect the meter to the audio output point X in Fig. 40-3, and to chassis ground or B-minus in the wiring.
4. If the meter reads zero, which it may do if the secondary is correctly adjusted, slightly detune the secondary until there is a distinct reading on the meter.
5. Adjust the primary trimmer or movable core for maximum meter reading.

6. Adjust the secondary trimmer or movable core for zero meter reading between positive and negative peak readings, just as in step 6 of the first method.

The next method of aligning the demodulator transformer can be used only with ratio detectors, not with discriminators. Fig. 40-4 shows the test connections.

1. Use the signal generator unmodnlated, tuned to the intermediate center frequency. Connect the high side of the generator output cable through 0.01 to 0.05 mf fixed capacitor to the grid of the tube preceding the demodulator transformer, or to the grid of the driver. Connect the low side of the cable to chassis ground or to B-minus.
2. For output measurement use a high-resistance d-c voltmeter or an electronic voltmeter on a d-c range.
3. Connect the meter across the large capacitor on the load side of the detector, with negative of the meter to the negative terminal of the capacitor and positive of the meter to the positive terminal. Align the primary of the transformer for maximum reading on the meter.
4. Change the meter connection to the audio output point of the ratio detector, or to point X in the diagrams. Align the transformer secondary for zero meter reading between two peaks. Be sure to get on the zero point where there is a rapid shift of the meter pointer with a small change of alignment adjustment.

After making adjustments by using a high-resistance d-c voltmeter or an electronic voltmeter it is good practice to make a check of response at frequencies below and above the intermediate center freqnency. With the meter still connected as for obtaining zero voltage, change the generator frequency to something between 75 and 100 kc (0.075 and 0.100 mc) below the center frequency. Note the voltage reading. Then change the generator frequency to exactly the same number of kilocycles or the same fraction of a megacycle above the center frequency. Again note the voltage. The two voltages should be equal or very nearly so. If there is any great difference you should repeat all the steps of adjustment for both primary and secondary.

The demodulator transformer may be aligned by using an output meter across the voice coil of the speaker with the connections shown by Fig. 40-5. If an electronic voltmeter has a range or a switch position for measurement of a-c volts it may be used instead of the output meter, employing the same connections. Do not forget that when using an output meter or any a-c meter on the voice coil the volume control of the re-

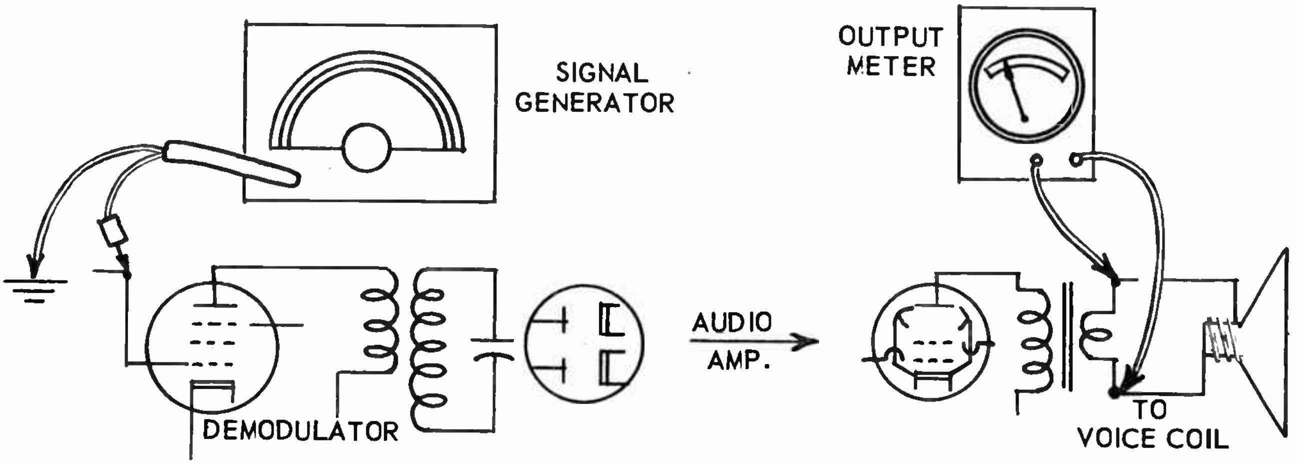


Fig. 40-5. Alignment connections with output meter on the speaker voice coil.

ceiver is to be adjusted for maximum volume. If such meters are used on the plate side of the output transformer it will not be necessary to set the volume control for maximum, but only high enough to give distinct readings on the meter.

The steps for alignment with this method are as follows.

(10) (a) (1)

1. The signal generator is to be used with audio modulation, in order to produce indications of audio-frequency voltage on the meter. Tune the generator to the intermediate center frequency, usually 10.7 mc. Connect the high side of the generator output cable through a fixed capacitor of 0.01 to 0.05 mf to the grid of the tube which precedes the demodulator transformer.
2. Connect the leads of the output meter or the a-c electronic voltmeter across the voice coil, as shown by the diagram.
3. Make a preliminary adjustment of the transformer secondary for minimum reading on the meter. Since the speaker remains connected to the output transformer, there will be sound which is relatively loud unless the signal generator voltage is at the center frequency. Tune for a minimum meter reading in a comparatively silent zone between two audible zones which you can hear with the transformer tuning slightly off resonance in either direction or with the generator detuned either way from the center frequency.
4. Align the primary of the demodulator transformer for maximum meter reading. In order to have a distinct meter reading it may be necessary to slightly detune the secondary, but do not get it further than necessary from the minimum voltage point at which the adjustment was set in step 3.
5. Realign the secondary for an exact minimum reading on the meter.

If the output meter is a fairly sensitive type, and is being used with a low-voltage scale, the reading will not drop all the way to zero nor will it pass through zero to bring the pointer on the other side. There is no steady positive or negative polarity in the alternating audio-frequency voltage being measured, and meter reading will increase as the secondary adjustment is changed either way from the point of resonance at the center frequency. It will be possible to have three very low readings, one for adjustment below the center frequency, a second at the center frequency, and a third for adjustment above the center frequency. Be sure to make the final adjustment for the middle minimum reading, at which the drop and rise will be much sharper than at either of the other minimums.

Should you have available an output meter and also a high-resistance d-c voltmeter or electronic voltmeter it is possible to do an excellent job in a short time by connecting the output meter across the voice coil and the other meter to the audio output point of the demodulator for making simultaneous readings. The test setup is as shown by Fig. 40-6.

The signal generator is connected as usual, to the grid of the tube preceding the demodulator transformer, through a capacitor of 0.01 to 0.05 mf. The generator is tuned to the intermediate center frequency. With the generator modulated at audio frequency read the output meter while aligning the transformer secondary for minimum voltage. Always work on the middle one of the three minimum readings which may be obtained by changing the adjustment.

Now use the signal generator without audio modulation while reading the d-c voltmeter or electronic voltmeter. Detune the signal generator first below and then above the intermediate center frequency to get peak voltage readings on each side of the center. Leave the generator tuning at whichever of these readings is the smaller, and align the transformer primary for maximum voltage reading.

Instead of connecting the d-c meter or electronic voltmeter to the audio output of the demodulator this meter could be connected across either load resistor, as from A to either B or C in Fig. 40-6. Then you

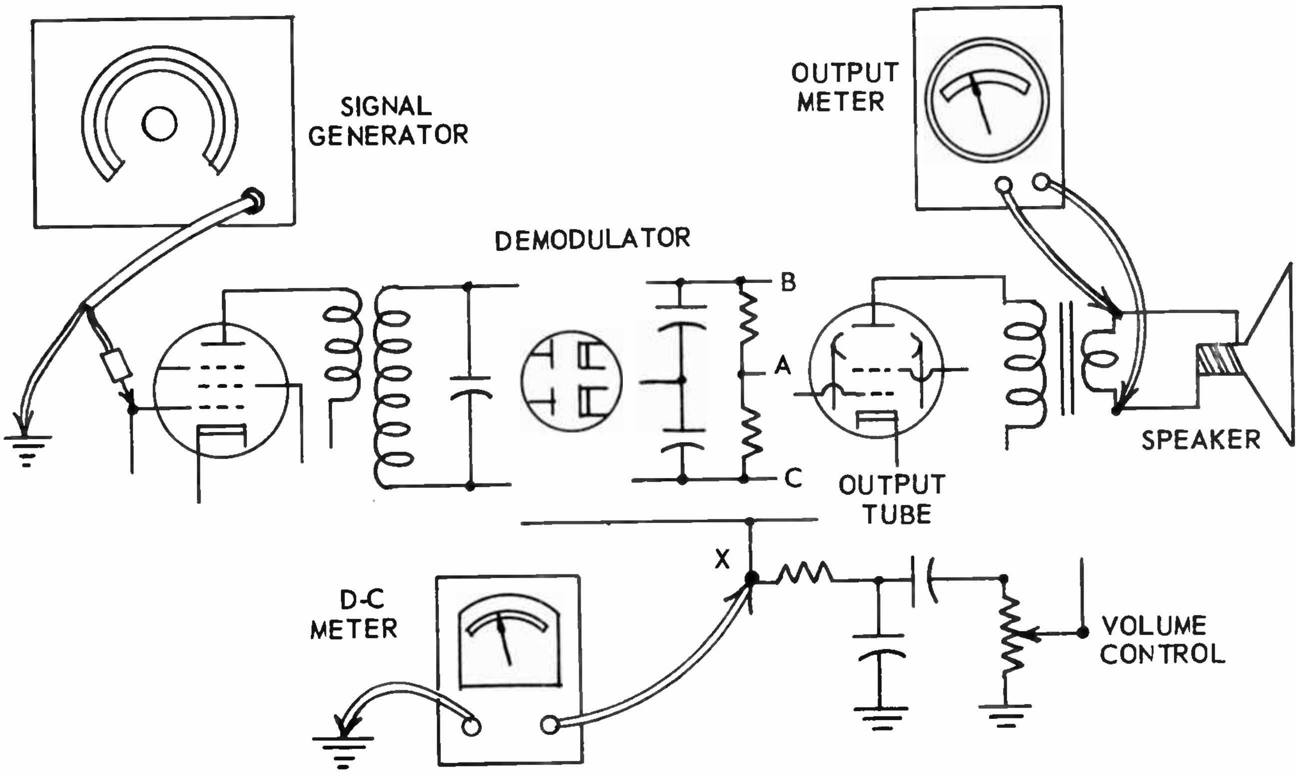


Fig. 40-6. Demodulator transformer alignment with output meter and d-c meter used together.

would not detune the signal generator, but would leave it at the center frequency while adjusting the transformer primary for a maximum voltage reading.

Possibly you are beginning to wonder why we describe so many ways of aligning an f-m demodulator. Why not describe the one best method and let it go at that? There are several reasons. First, you should realize that nearly all service jobs may be performed in any of various ways to produce the same or equivalent results. If you learn one way in these lessons, and some other technician uses a different method, don't assume that his method is wrong. Every method which has been described is recommended by one or more manufacturers of f-m and television receivers. The explanations have been changed only so far as is necessary to make all of them suitable for any receiver of construction allowing application of a given method.

Your choice may depend on the kind of equipment available in the service laboratory where you are working. It may depend on preferences of the superintendent of the laboratory. Some methods are more convenient than others for use on a particular receiver. At any rate, you should understand the operating principles of the receivers well enough to insert a signal voltage where it is needed and measure the results in any way that is appropriate to the job in hand. We won't always take time to explain so many different ways of handling one operation, but remember that usually there will be other possible methods.

Even the order of alignment may be changed. Several manufacturers recommend that you first align the demodulator primary and secondary in one of the ways previously described, then align the i-f transformers, and finally come back to make a precise setting of the demodulator transformer secondary for zero audio output at the center frequency for which the i-f transformers actually have been aligned. The reason is this: Unless you are working with test equipment of highest quality you won't be sure of making i-f alignments to within a few kilocycles of a specified center frequency, and of aligning the demodulator with equal precision. But even though you do align the i-f transformers a little bit off a specified frequency, everything will come out all right if you align the demodulator to that same frequency. It will later be easy to adjust the r-f oscillator to provide a beat which is the frequency you actually have used.

Fortunately, the average listener isn't too critical of musical reproduction, and will be well satisfied with the results of usual alignment methods. There are others whose natural perception or musical training lets them appreciate and demand the most nearly perfect reproduction, and they know that f-m reception is capable of giving such reproduction.

I-F TRANSFORMER ALIGNMENT. For methods of i-f transformer alignment to be described first we shall use the signal generator without modulation, connected as shown by Fig. 40-7. The high side of the generator output cable is connected through a fixed capacitor of 0.001 to 0.05 mf capacitance to the grid of the tube which precedes the transformer being aligned. That is, for aligning a transformer at *A* in the diagram the generator connection would be made at *A*, and for aligning a transformer at *B* the generator would be connected at *B*.

We commence i-f alignment with the transformer or coupler just ahead of the demodulator transformer, then work away from the demodulator until reaching the transformer or coupler whose primary is in the plate circuit of the converter or mixer tube. How many transformers or couplers are handled thus will depend on the number of i-f amplifying stages in the receiver.

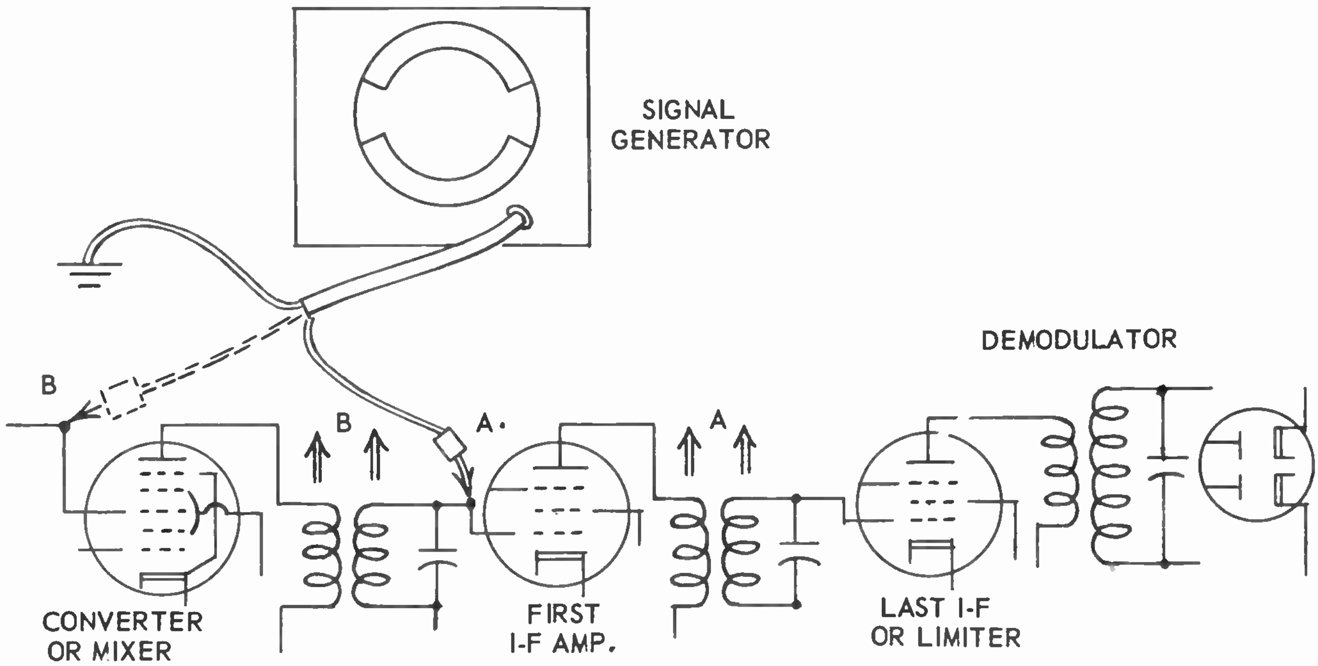


Fig. 40-7. Connections of the signal generator for aligning i-f transformers.

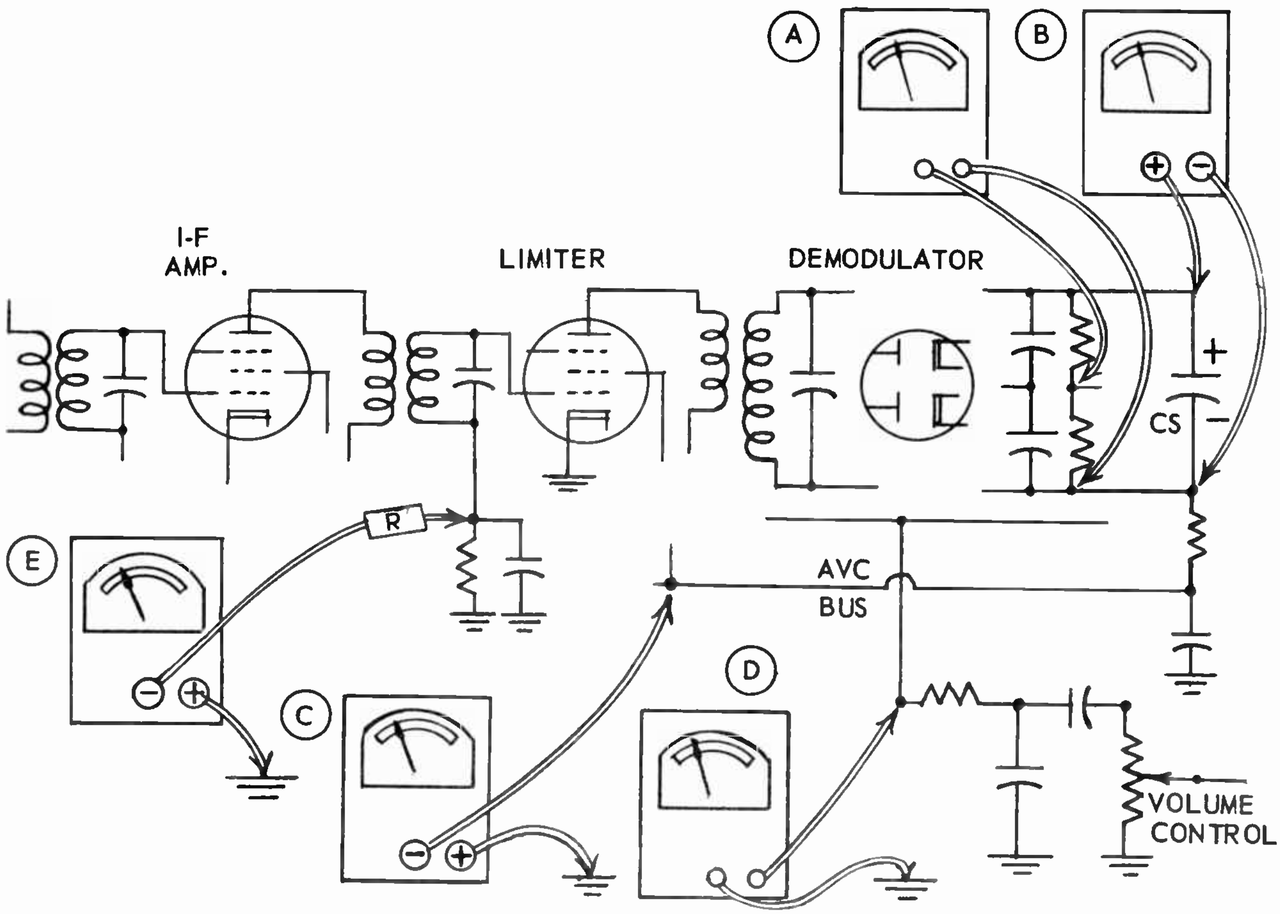


Fig. 40-8. Various points at which a d-c voltmeter or electronic voltmeter may be connected for alignment of i-f transformers.

If the transformers are not very far out of alignment to begin with, the entire job may be done with the generator high-side connected to the grid of the converter or mixer, instead of moving the connection back as the work proceeds. If the receiver has no r-f amplifier, but has the antenna coupled to the grid of the mixer or converter, the signal generator may be connected to one of the antenna terminals through a 300-ohm carbon resistor instead of through a capacitor.

No matter where the generator is connected, even if it is to the antenna terminal, the frequency is adjusted for the intermediate center value, usually 10.7 mc. In all alignment of i-f amplifier stages the generator output voltage must be kept as low as will give distinct readings on the meter. Otherwise, if there is automatic volume control on any of the i-f amplifiers, this control will prevent getting peak voltages at the center frequency. Either the peaking will become very broad or there will be two peaks instead of only one.

Various meter connections for measuring voltage output during i-f transformer alignment are shown in Fig. 40-8. In every case the meter is a high-resistance d-c voltmeter or an electronic voltmeter used on the d-c range. At *A* the meter is connected across one of the demodulator load resistors. With a ratio detector the connection may be made across either resistor. With a discriminator it usually is found that smoother and more reliable voltage indications are obtained across one load resistor than across the other. If one end of one discriminator load resistor is grounded, this resistor should be used for measurement. Correct polarity for connections depends on which resistor is used and on the type of demodulator. If the first trial doesn't give readings up scale on the meter, reverse the connections or change the polarity switch on the meter.

The connection shown at *B* is across the large capacitor on the load side of a ratio detector. There is no equivalent connection for a discriminator.

The meter shown at *C* is connected between the avc bus and chassis ground or B-minus. The negative side of the meter must go to the bus line and the positive side to ground or B-minus. The avc voltage need not necessarily be taken from the ratio detector capacitor, as in the diagram. This voltage may come from any source which is farther along in the amplifying and detecting system than the transformer being aligned in accordance with the avc voltage.

At *D* the meter connection is made between the demodulator audio output point *X* and ground. To employ this connection it is necessary to detune the secondary of the demodulator transformer far enough off resonance to allow distinct meter readings.

Unless the demodulator secondary is thus detuned the meter would read zero with i-f transformers correctly aligned. This zero indication would not assure correct adjustment of i-f stages. Of course, the demodulator secondary then will have to be readjusted for a zero reading after the i-f alignment is completed. This does not constitute an objection to the meter connection at *D* in Fig. 40-8, since the demodulator secondary adjustment always should be checked after i-f alignment is finished.

The meter connection shown at *E* is preferred whenever the receiver employs a limiter ahead of the demodulator. Limiters are used with a few ratio detector systems, but as a general rule only with discriminators. The negative terminal of the meter is connected to the grid end or the high side of the limiter grid resistor. The positive terminal is connected to chassis ground or B-minus. Unless the meter is an electronic type, connect in series with the negative lead a carbon resistor of 0.1 to 0.5 megohm. If you are in doubt as to whether the positive side of the meter should go to chassis ground or to B-minus, make the connection to neither of these points but to the cathode of the limiter tube. Then you have grid-to-cathode voltage, which is certain to give correct indications.

The meter connection such as shown at *E* could be made to the high side of the grid resistor on any i-f amplifier tube having grid-leak bias, provided this tube follows the transformer being aligned. You will remember that this type of bias often is used for automatic volume control on one or more tubes. With any meter connection to a grid resistor on a limiter or any other tube it is important to keep the generator output voltage just as low as will give readable indications of voltage. Otherwise the avc action always will flatten the response and prevent making correct peak adjustments.

With the signal generator and meter connected in any of the ways just described it remains only to adjust the secondary and then the primary of each transformer for peak voltage reading. Always change the adjustment one way and the other a few times to make certain that you recognize the peak voltage, then align precisely on this peak.

With another method of i-f transformer alignment we use an output meter across the voice coil of the speaker. This output meter connection was shown by Figs. 40-5 and 40-6. The signal generator is connected to the grid of the tube preceding the transformer being aligned in each step of the process, as shown by Fig. 40-7. The generator is used with audio modulation, as always is necessary when employing an

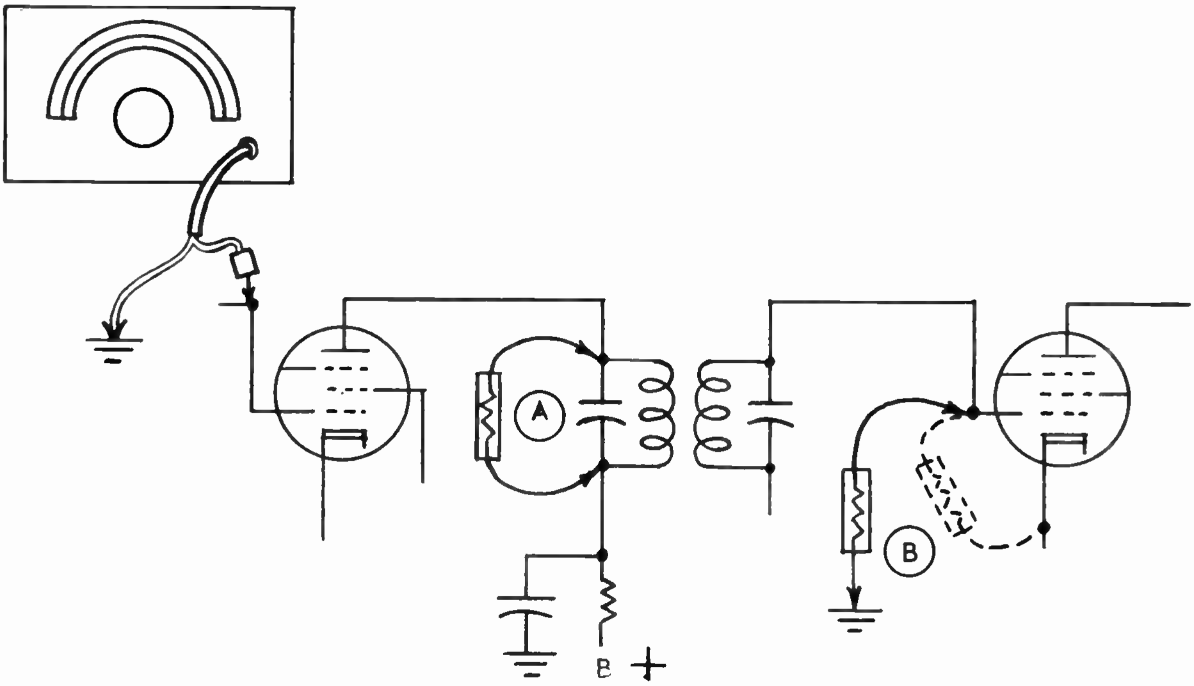


Fig. 40-9. Resistance loading of primary, secondary, or both, on an i-f transformer.

output meter on the speaker or anywhere on the audio amplifier system. The generator is tuned to the intermediate center frequency. It is essential that generator output voltage be kept just as low as will give distinct indications on the output meter.

With signal generator and output meter connected as described, the next step is to detune the secondary of the demodulator transformer enough to allow good readings on the meter as the generator frequency is temporarily shifted below and above the center frequency. Now the secondary and then the primary of each i-f transformer is aligned at the center frequency for maximum output as shown by the meter. Work on the transformers in order from the one ahead of the demodulator back to the one following the converter or mixer. Finally, readjust the secondary of the demodulator for minimum reading on the output meter, with the generator tuned at the center frequency.

LOADING OF OVERCOUPLED TRANSFORMERS. In order to broaden the tuning by providing a double-peaked frequency response many i-f transformers for f-m receivers use overcoupled windings. The mutual inductance is great enough to have a material effect on frequency at which resonance occurs. This added inductance, combined with the broad frequency response, makes it difficult or impossible to correctly align each winding to the same frequency by feeding a signal to the tube ahead of the transformer and measuring resulting voltage at points beyond the transformer.

To get around this difficulty it is the usual practice to temporarily connect a fixed resistor across the secondary winding, or else to use a resistor across the secondary while aligning the primary and then to use a resistor across the primary while aligning the secondary. One or the other of these expedients will allow a single resonant peak sharp enough for accurate alignment.

A resistor connected across a transformer primary is represented at *A* in Fig. 40-9. This resistor should

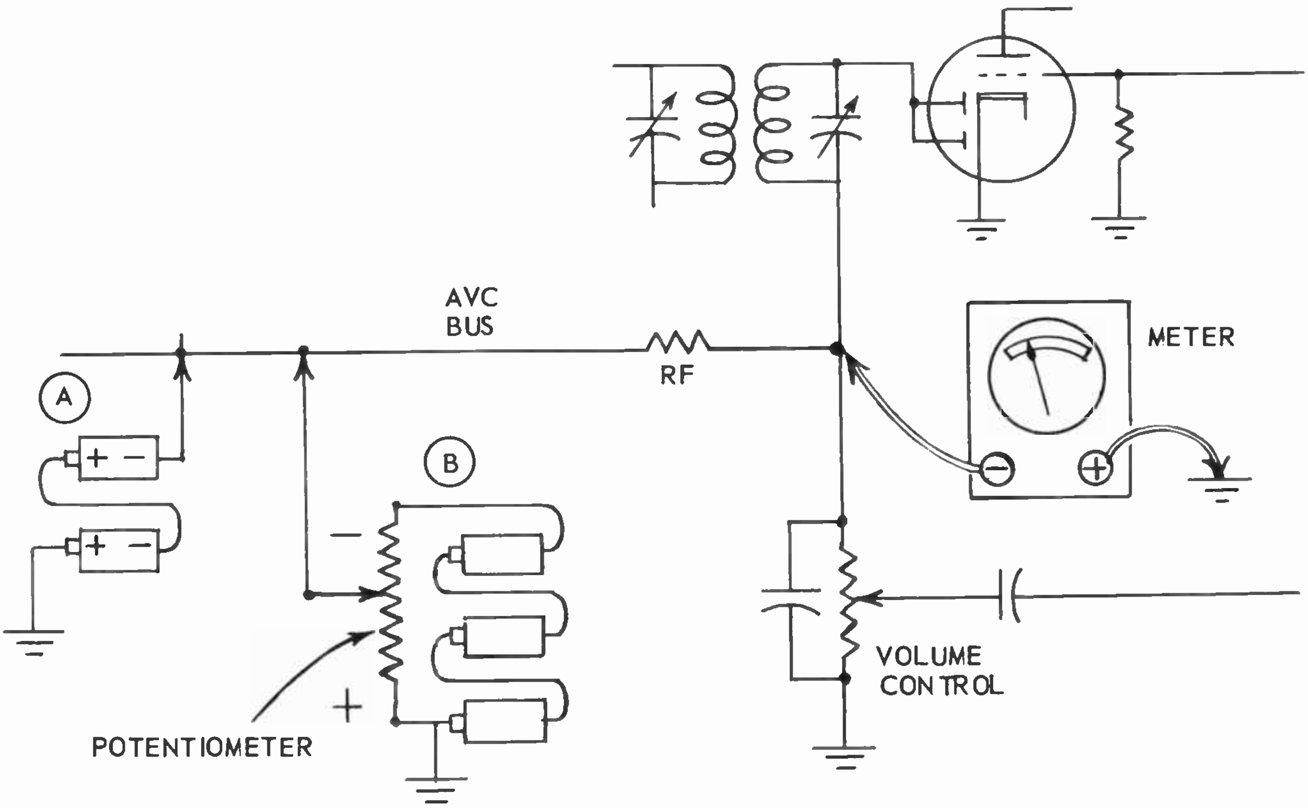


Fig. 40-10. Overriding an avc voltage with dry cells connected to the avc bus.

be connected only across the primary winding, not from the plate of the preceding tube to ground, B-minus, or B-plus, where the resistance would be subjected to plate supply voltage. A resistor may be connected similarly across the secondary if this is convenient. Otherwise the secondary may be shunted by connecting the resistor from the grid of the following tube to ground or to the cathode of that tube, as at *B*. Often a resistor across the secondary will be sufficient while aligning both the primary and the secondary windings. The resistor should be used on only one winding of one transformer at a time, on only the transformer being aligned.

The smaller the value of the shunting resistor the easier it will be to obtain a peak reading provided the voltage output of the signal generator can be made great enough to give a distinct meter reading. Suitable resistances range all the way from 300 ohms to 15,000 ohms or more. If some small value of resistance prevents good meter indications, try a greater resistance. Another way of raising the voltage indicated by the meter is to connect the signal generator to the grid of a tube farther ahead of the transformer being aligned, unless you already have the generator on the grid of the converter or mixer tube.

The only practicable way of determining whether or not a transformer is overcoupled is to measure the shape of its frequency response or to observe the response with the help of an oscilloscope. It would not be easy to determine the degree of coupling, even with the shield or can removed, merely by looking at the windings. A transformer probably is overcoupled if you have unusual difficulty in obtaining a sharp peak

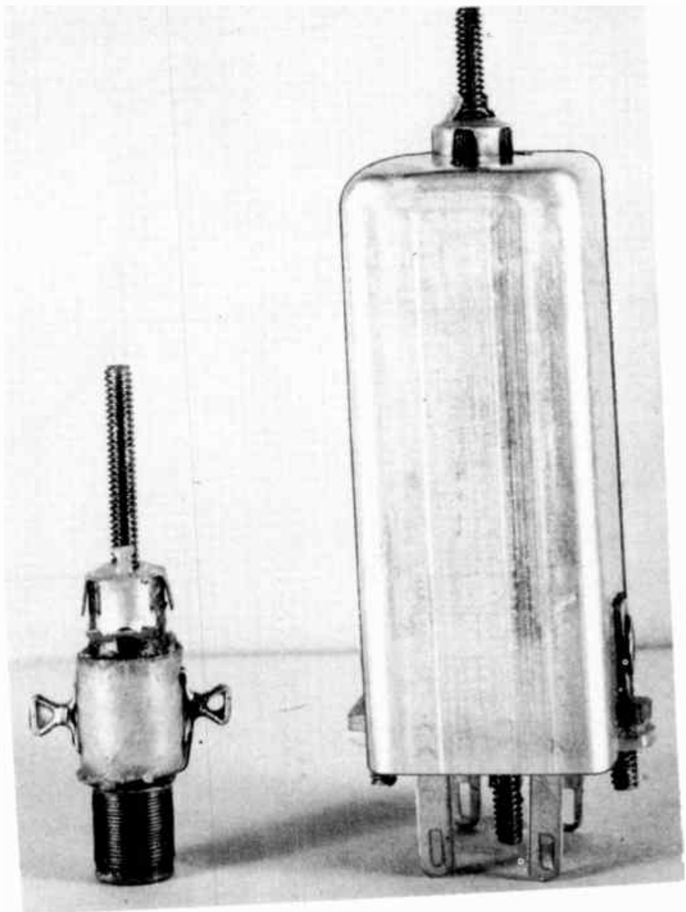


Fig. 40-11. Alignment adjustment screws on an i-f impedance coupler (left) and on an i-f transformer (right)

of voltage when moving the alignment adjustments, or if there seem to be two voltage peaks when you shift the signal generator frequency very slowly through the normal tuning range of the transformer. The transformer most likely to be overcoupled is the one which immediately follows the converter or mixer.

OVERRIDE OF AUTOMATIC VOLUME CONTROL. When aligning the i-f transformers of receivers having automatic volume control for one or more of the amplifier tubes it often is difficult to keep the signal generator voltage low enough to allow sharp peaking of the response. This is especially true when the generator is connected to the grid of the converter or mixer tube, for then there is a large total of amplification between this signal input point and anywhere that a meter may be connected. The difficulty occurs not only during alignment of f-m receivers, but also when working on a-m sets for standard broadcast reception, and it always has to be guarded against when aligning the video i-f amplifiers of television receivers.

The automatic bias may be overridden or made inoperative by using some dry cells connected as in Fig. 40-10. This diagram applies specifically to automatic volume control voltage taken from the high side of the volume control on a diode detector in an a-m receiver, but the same connection of dry cells may be used for f-m receivers.

At *A* there are two dry cells connected in series between the avc bus and ground, with the negative terminal of the cells to the bus and the positive to ground. This arrangement places a constant 3-volt negative bias on all tubes formerly controlled by the avc voltage. This much constant voltage is ample provided the output of the signal generator is kept fairly low. A single dry cell would provide 1½ volts of constant bias, or three cells would provide 4½ volts, always at the rate of 1½ volts per cell.

An adjustable constant bias may be provided with dry cells connected as at *B*. Three or more cells are in series across a potentiometer whose resistance is anything between 5,000 and 50,000 ohms. The positive end of the series cell group and potentiometer is connected to ground. The slider of the potentiometer is connected to the avc bus. Moving the slider toward the negative end of the cell group makes the constant bias more negative. The slider allows making the bias anything between zero and a number of volts equal to 1½ times the number of cells. To prevent needless discharge of the cells, disconnect one end of the potentiometer when the device is not in use.

Be sure to connect the source of constant bias, the dry cells, to the avc bus at a point beyond the first high-resistance unit, *Rf*, that is connected to the top of the volume control or to any other source of avc voltage. This places the constant bias on the grid return side of this resistor. A high-resistance voltmeter or an electronic voltmeter may be connected as usual to the top of the volume control for indicating the

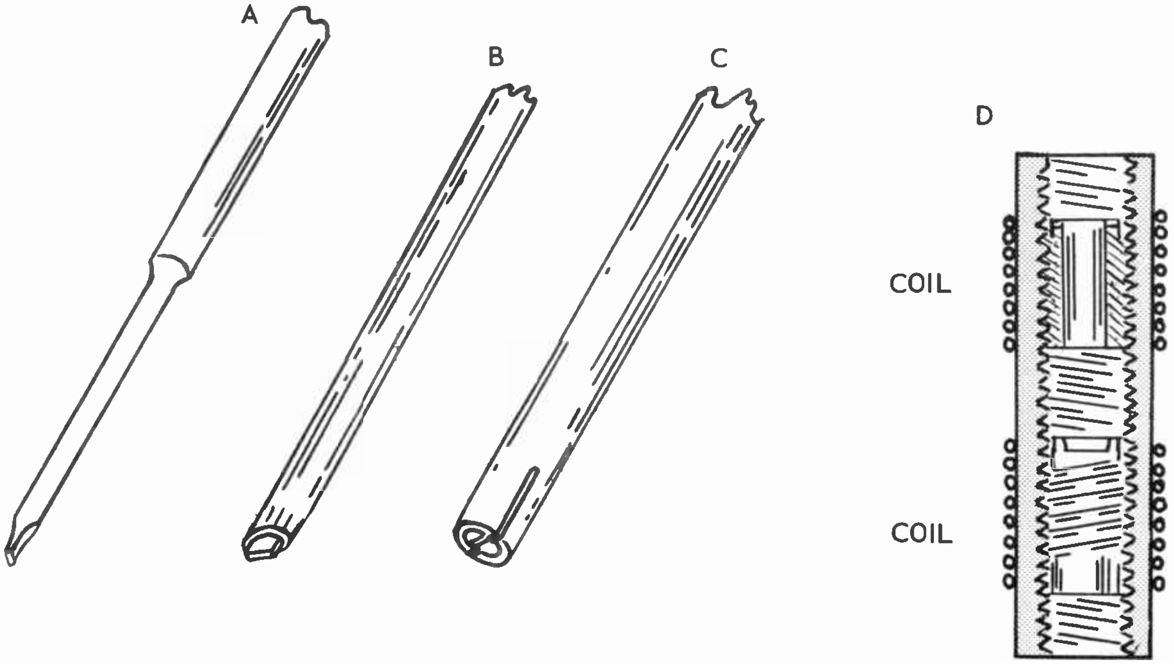


Fig. 40-12 Alignment tools with screw driver tips (left), and one type of double core adjustment (right).

effects of alignment adjustments. When using dry cells in a generally similar way on the avc bus in an f-m receiver the signal generator and any meter used for output measurement are connected in ways that have been described. So long as the constant bias is applied to the avc bus so that a high resistance is between this bias voltage and the source of avc voltage there will be no interference with readings of any meter used for output measurement.

OUTPUT METER FOR A-M ALIGNMENT. In the lesson which described alignment of a-m receivers for the standard broadcast and short-wave bands we obtained indications of output voltage from a meter connected to the high side of the volume control, as in Fig. 40-10. It is possible also to make all alignment adjustments on a-m receivers by using an output meter across the voice coil of the speaker. The output meter is connected exactly as shown by Figs. 40-5 and 40-6.

It is more general practice to use an output meter than either a high-resistance d-c voltmeter or an electronic voltmeter for alignment of a-m receivers. Nearly all volt-ohm milliammeters or analyzers have one a-c voltage range that is suitable for use as an output meter.

ADJUSTMENTS OF TRANSFORMERS AND COUPLERS. As has been mentioned before, any interstage transformer or coupler may be tuned to resonance with either adjustable capacitors or movable cores in the windings. Exactly the same methods of alignment are used with either type of adjustment.

Many transformers and couplers have screw type adjustments such as pictured by Fig. 40-11. These are easily aligned with a screw driver made entirely of hard insulating material or with one having a very small metallic blade carried on a shaft and handle of insulating material.

Greater care is necessary when the movable cores themselves are slotted at one or both ends, and when the alignment tool must be inserted through an opening in the shielding can. There is danger of breaking out the slots in the rather brittle core material, and there is danger of stripping the threads on the cores or in the parts that carry the cores. Always use an alignment tool on which the tip is flat ou both sides and on which the extreme end of the tip is square across the entire width. If the tip is too wide it will cut the threads of many core supports. A width of 1/8 inch is safe for any adjustment which requires inserting the tool into the case or can.

A thin blade and tip entirely of insulating material is shown at *A* in Fig. 40-12. At B is a small metallic tip mounted in an insulating tube for the handle or shank. At C there is a small metallic tip recessed in the tubular insulating handle. There are dozens of other styles of alignment tools available.

There are unusual mechanical constructions used in some transformers. As an example, at D in Fig. 40-12 there are upper and lower movable cores or slugs in a single insulating tube, with coil windings around the outside of each core position. The lower core is solid and has a screw driver slot in its upper end. This slot is reached by inserting a narrow shank alignment tool through an opening in the upper core. The upper core is adjusted by using a tool with a tip wide enough to reach across the center opening and engage notches on either side. In a few cases the adjustments are so inaccessible or otherwise are made in such manner as to call for the use of special alignment tools. Such tools, singly and in kits, are sold by radio and television supply houses.

Adjustments quite often are secured against accidental shifting and against unauthorized tampering by coating or filling them with some kind of cement. Some of these cements may be loosened by bringing the tip of a hot soldering iron close to them while applying moderate pressure on the screw or nut which is to be turned. Other cements may be dissolved by lacquer thinner, obtainable at hardware stores, or by special cement solvents sold by radio and television supply stores.

41-245

COMMERCIAL TRADES INSTITUTE

KNOWLEDGE + AMBITION = SUCCESS

LEARN . . .

ACT . . .

"The heights by great men reached and kept
Were not attained in sudden flight,
But they, while their companions slept,
Were climbing upward in the night."

ENJOY . . .

COMMERCIAL TRADES INSTITUTE

CHICAGO, ILLINOIS

TELEVISION

LESSON NO. 41

ALIGNMENT OF TUNERS

Training Your Mind...

C.T.I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind
C.T.I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON NO. 41

ALIGNMENT OF TUNERS

5 While aligning the tuner section of an f-m receiver, or the f-m tuning circuits of a combination receiver, the signal generator is connected to the f-m antenna terminals of the receiver. The results of adjustments are indicated by a d-c voltmeter or electronic voltmeter connected to the demodulator output. The signal being introduced at the antenna terminals has to pass from the tuner through all the i-f amplifiers and the demodulator before reaching the meter. Consequently, it is essential that the i-f amplifier and demodulator transformers shall have been aligned before undertaking adjustment of the tuner.

We desire maximum transfer of r-f signal from generator to tuner circuits, while avoiding distortion of input signal voltages. This requires that the impedance of the generator output cable, where connected to the antenna terminals of the receiver, be practically equal to the input impedance of the receiver. Input impedance of modern f-m receivers is standardized at 300 ohms. Output impedances of service type signal generators usually range between 20 and 50 ohms.

5 An impedance match satisfactory for tuner alignment ordinarily is provided by connecting in series with the high-side lead of the output cable a carbon resistor of 240 to 300 ohms, as at A in Fig. 41-2. It is presumed that the output impedance of the signal generator plus the series resistance is approximately 300 ohms. The same or better results may be had as at *B*, with a carbon resistor of 120 to 150 ohms in series with each of the output cable leads. Any resistor or resistors used in this general manner may be called a dummy antenna.

Some signal generators are provided with a special output cable on the receiver end of which is a matching pad for bringing the impedance to approximately 300 ohms at the clips or prods on the end of the cable. At *C* in Fig. 41-2 is shown such a pad for matching a 50-ohm signal generator to a 300-ohm receiver. No additional series resistance is used when there is a regular matching pad.

The meter used for measurements during tuner alignment is a high-resistance d-c voltmeter or an electronic voltmeter used on a d-c range. This meter is to be connected at some point where it will indicate peak voltage when the tuning capacitors or inductors of the receiver are set for resonance at the f-m carrier frequency being furnished by the signal generator. Any of the following meter connections might be used:

a Across either resistor on the load side of the demodulator, with connections of such polarity that the meter reads up scale.

To the avc bus for f-m reception, with negative of the meter to the bus and positive to ground or B-minus. Be sure you connect to the bus for f-m operation, and be sure that the connection to this bus is at a point between converter or mixer and the demodulator.

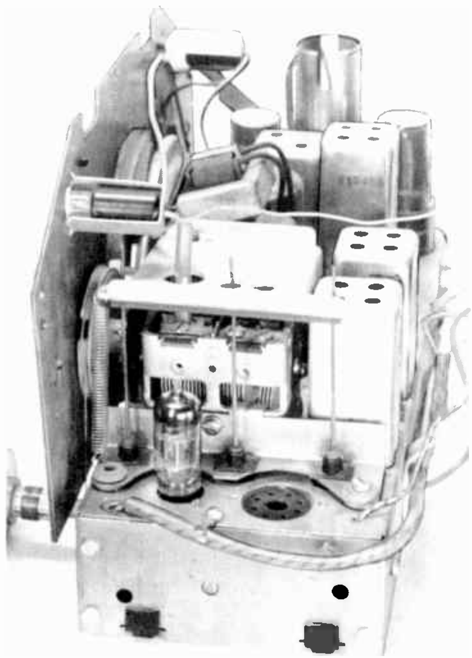


Fig. 41-1. The tuner end of a combination fm-am receiver.

If the receiver uses a ratio detector the meter may be connected to the large capacitor in the demodulator load circuit. Be sure to get the negative of the meter to the negative terminal of this capacitor.

If the receiver uses a discriminator and limiter the meter may be connected to the grid resistor of the limiter tube. Connect the negative of the meter to the high side or grid end of the resistor, with the positive of the meter to ground, B-minus, or the limiter cathode.

The signal generator is used without audio modulation. The generator and receiver are tuned to the same frequency except when specified otherwise in following instructions. For adjustment of the r-f oscillator, the r-f to mixer coupler, and the antenna coupler this frequency usually is somewhere between 106 and 108 mc, preferably at a frequency where no nearby f-m station operates. Alignment sometimes is made at the mid-band frequency of 98 mc.

The first step in alignment is adjustment of the f-m oscillator trimmer or movable core for maximum meter reading. With most receivers it is possible to obtain maximum readings at either of two settings of the oscillator adjustment. One is the correct frequency, which is higher than the signal generator frequency. The other is below the generator frequency. If there is any doubt about this frequency, commence work with the adjustment set for the lowest possible frequency, with a trimmer capacitor closed or with a movable core all the way into the coil. Then, as you change the adjustment, the first peak will be at the frequency which is too low. The second peak will be at the correct frequency. We used this method for alignment of the oscillator in a-m receivers. After identifying the correct oscillator frequency, shift the adjustment a little ways back and forth to make sure you recognize the true peak of voltage, then leave the adjustment there.

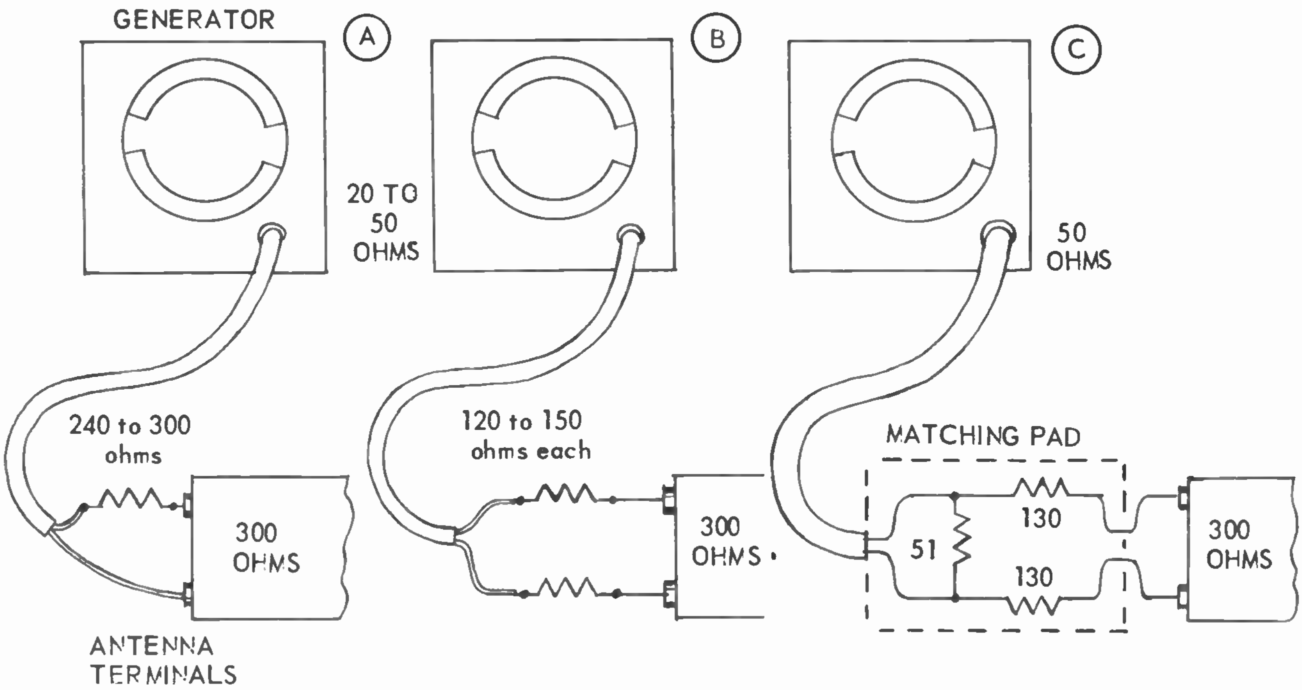


Fig. 41-2. Matching generator output impedance to antenna input impedance of an f-m receiver.

If there is an r-f amplifier, and if there is a tuned coupling between this amplifier and the converter or mixer, the next step is alignment of this coupling. Tuned frequencies of the generator and receiver remain as when aligning the oscillator. The r-f to mixer coupler is adjusted for maximum reading on the meter. While making this adjustment with the alignment tool in one hand try varying the receiver tuning capacitor back and forth to find the point at which there is maximum voltage. This is advisable because any tuning in the grid circuit of the converter or mixer may have some effect on oscillator frequency. It may be that the pointer of the receiver tuning dial is not at the exact signal generator frequency when there is maximum voltage. This is not particularly objectionable, but you must make the adjustment for maximum voltage regardless of the dial setting.

Fig. 41-3 is a circuit diagram which includes a tuned impedance coupler between the r-f amplifier and the r-f signal grid of the converter. High impedance in the plate circuit of the r-f amplifier is provided by the radio-frequency choke coil RFC. R-f signals go to the converter grid circuit through coupling capacitor *Cc*. A parallel resonant variably tuned circuit is between the converter grid and ground. The tuning capacitor is marked *C*. Alignment adjustment is made by the movable core *L* in the coil.

Another style of tuned coupling between an r-f amplifier and the signal grid of a converter is shown by Fig. 41-4. Here the coupling is a transformer with variably tuned secondary and untuned primary. Variable tuning is by means of the movable core in the secondary winding. The tops of three such movable cores may be seen in Fig. 41-1. One core is to the left of the small tube in the foreground, another core is to the right of this tube, and the third core is still farther toward the right.

Long threaded screws extend upward from each core to a cross bar which is raised and lowered by the tuning knob. Alignment adjustments are made by turning the threaded screws in their attachments to the tuning bar, thus moving each core up or down as may be required for resonance with the tuning bar in a position determined by setting of the tuning knob and dial. The three adjustable cores are for tuning the r-f oscillator, the coupling between r-f amplifier and converter, and the antenna coupling to the r-f amplifier. Only two such adjustable cores are shown in Fig. 41-4, one for the r-f to converter coupling and the other for the f-m oscillator coil. The antenna coupling here is untuned.

When the antenna coupling to the r-f amplifier is untuned, as in Figs. 41-3 and 41-4, alignment of the r-f to converter coupling completes the work on the tuner at the frequency first applied from the signal generator. There are a few f-m receivers which employ an untuned coupling between r-f amplifier and converter. Then there is a tuned antenna coupling, which is aligned following the r-f oscillator adjustment.

Whenever there is a tuned antenna coupling its alignment is the final step when using the frequency first applied from the signal generator. The receiver tuning dial remains at the frequency being furnished by the generator, and the generator still is used without audio modulation. The antenna coupling is aligned to produce maximum reading on the meter. It is advisable to vary the receiver tuning dial back and forth to find the position which allows maximum meter reading. The adjustment should be turned through the position for peak voltage several times to identify the actual peak. Tuning of the antenna coupling will be somewhat broader than that for the r-f to converter coupling, and much broader than that of the r-f oscillator.

The next step is to change the signal generator frequency to something between 88 and 90 mc, near the low end of the f-m band. Then operate the receiver tuning knob and dial to obtain maximum voltage on the

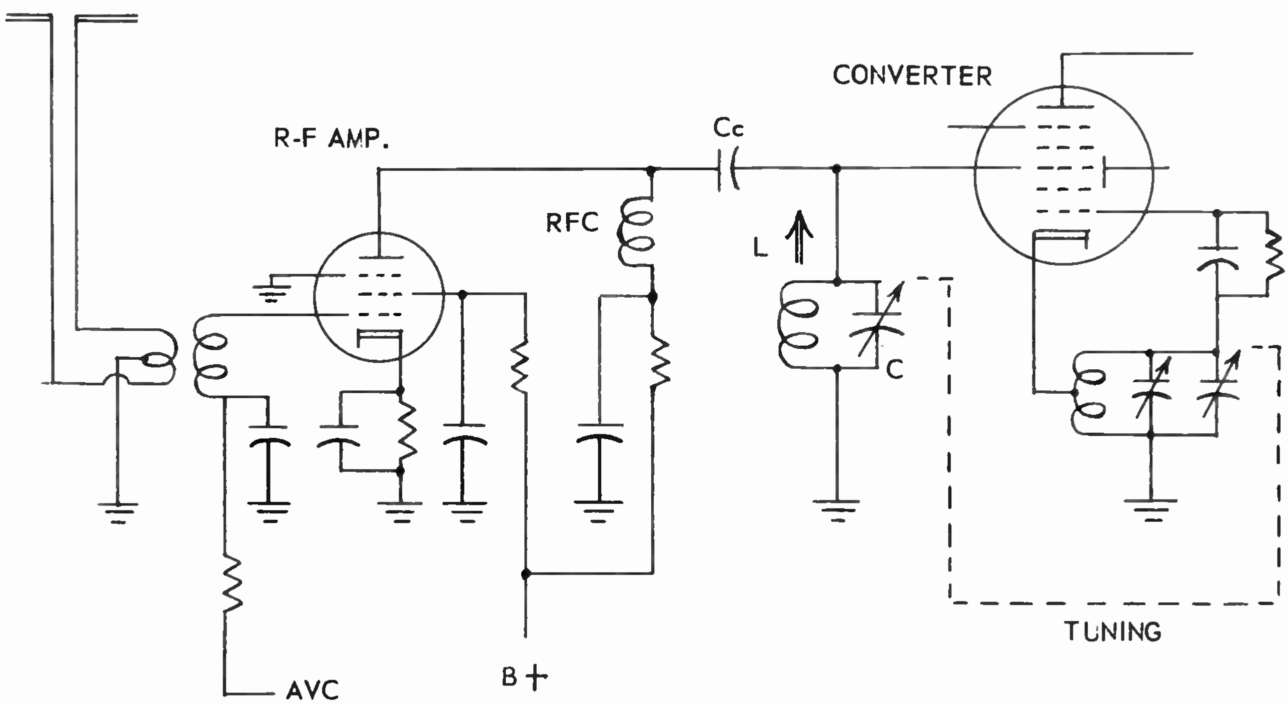


Fig. 41-3. R-f to converter coupling by means of tuned impedance.

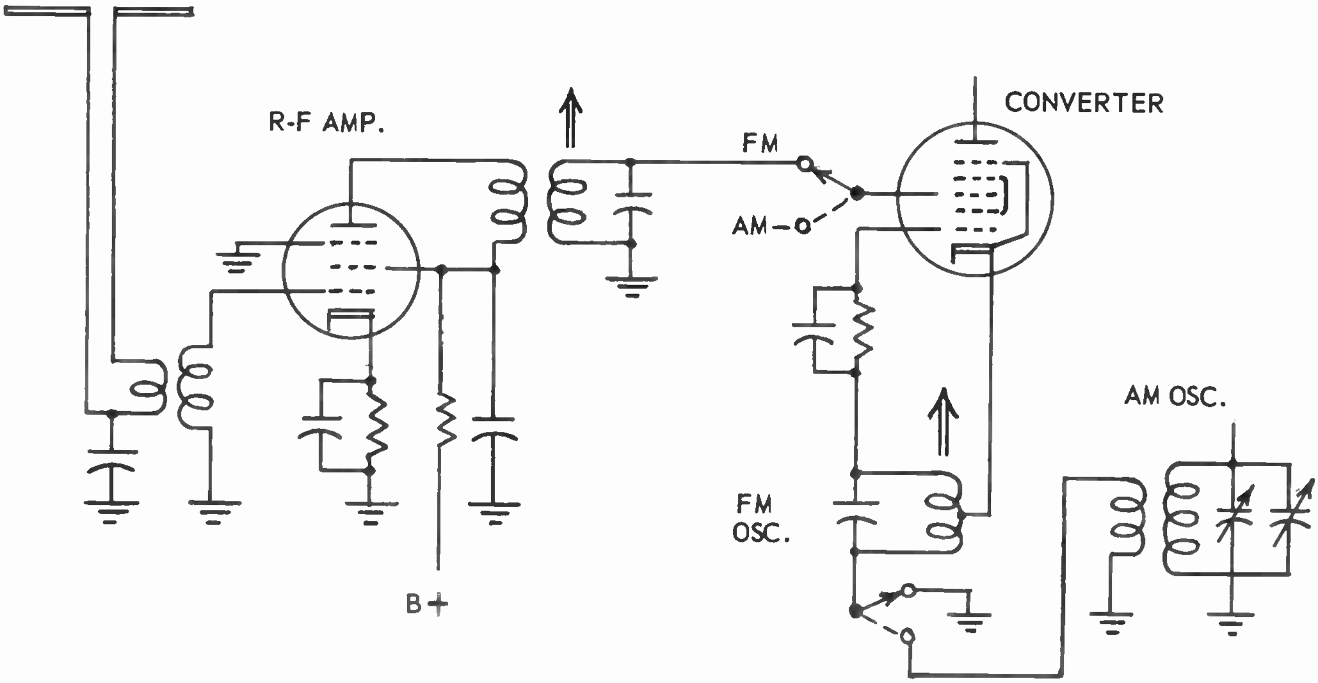


Fig. 41-4. R-f to converter coupling through a slug-tuned transformer.

meter. If the frequency indicated by the receiver dial is fairly close to the actual generator frequency the processes of alignment may be considered complete, and satisfactory. In a few f-m receivers there is a padder capacitor in the r-f oscillator circuit. This capacitor may be adjusted to obtain correct tracking or agreement between generator and dial frequencies at the low end of the dial. As with all padder adjustments it will be necessary to work back and forth between this adjustment and those made at the high-frequency end of the band to get the best possible overall results.

7 There are some receivers having adjustable trimmer capacitors across the tuned coils and having also movable cores in one or more of these coils. The capacitor trimmers are used for alignment at the high end of the band, and the movable cores are used for alignment at the low end of the band. The cores are equivalent in action to padder capacitors, and alignment must be checked alternately at the high and low ends of the frequency range until obtaining the best settings at both ends.

There will be some cases in which dial indications at the low-frequency end of the band are far out of line after making correct high-frequency adjustments, and in which there are no padders or movable cores to allow independent alignment at the low frequencies. This situation often may be handled by either spreading or squeezing the turns of one or more tuning coils. This can be done with all self-supported coils and also with many coils which are wound on smooth forms. Spreading the turns a little farther apart will make it possible to reach higher frequencies, while pressing the turns closer together will allow bringing in lower frequencies.

Change of coil inductance by altering the spacing between turns will change the high-frequency tuning more than the low-frequency tuning so far as number of mega-cycles is concerned. The process may be be-

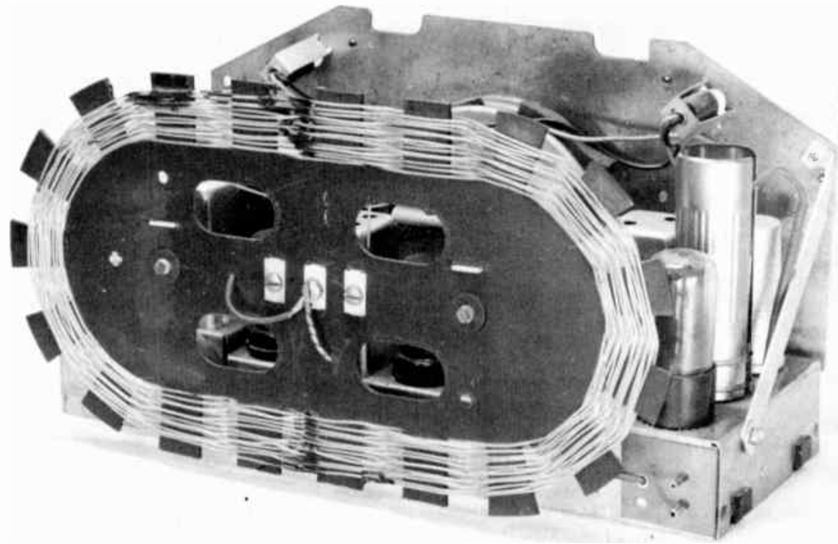


Fig. 41-5. A loop antenna used for a-m reception.

gun with the high-frequency trimmers in their center positions while altering the coil turns to get agreement between generator and dial frequencies at the low end. The trimmers then may allow adjustment at the high frequency end. It is necessary to continue working between the upper and lower frequencies to obtain the best possible average tuning.

The relative positions of all parts and all wires in f-m tuners are critical. The least change in the position of a resistor, capacitor, or inductor, or moving a wire connection, may completely upset your alignment adjustments. All wire connections carrying r-f currents are short, and usually are run as directly as possible. Grid wires, plate wires, and coupling capacitors should be kept away from chassis metal. Cathode wires, screen wires, and bypass capacitors often are close to chassis metal. Wires going to the ungrounded sides of bypass capacitors are to be considered as in r-f circuits.

9 After completing an r-f alignment job all the way from the demodulator to the antenna you should re-check all the adjustments. Make certain that the demodulator primary is aligned for maximum output and the secondary for zero output at the center frequency. Check the peaking of i-f transformers at the center frequency. Finally check the tuning of the r-f oscillator and of r-f to mixer couplings and of antenna to r-f couplings at two or more positions of the tuning dial.

LOOP ANTENNAS. In an earlier lesson explaining the alignment of a-m receivers it was specified that the antenna is to be disconnected and the signal generator connected to the antenna terminals during tuner adjustments. Those instructions apply specifically to receivers operated with outdoor antennas or indoor antennas which are not housed within the receiver cabinet. Such external antennas are used with some receivers for the a-m standard broadcast band and with many short-wave a-m receivers.

The majority of small a-m standard broadcast receivers and combination fm-am receivers are fitted with a permanently attached or built-in loop antenna for a-m reception. Such an antenna is not disconnected for a-m tuner alignment, but is used for signal pickup as will be explained later. Sig. 41-5 is a picture of a loop antenna permanently mounted on the rear of the chassis of a combination fm-am set. This loop antenna is used for a-m reception. A separate antenna of the "dipole" type is used for f-m reception.

Fig. 41-6 is a circuit diagram showing an a-m loop and an f-m dipole connected through a band switch to the grid circuit of an r-f amplifier. The loop antenna is essentially an inductance coil of large dimensions. This loop inductance is tuned to resonance at the received carrier frequency by an adjustable trimmer capacitor and a variable tuning capacitor, just as any other inductance might be tuned.

The loop conductors or turns are in the moving magnetic fields which are part of the radiated electromagnetic signal waves coming from a-m transmitters through the space in which the receiver and its loop are located. The magnetic lines cut through the conductors which form the right-and left-hand sides or ends of the loop, and emf's at signal frequencies thus are induced in these conductors.

The electromagnetic signal wave from the a-m transmitter moves outward in all directions. Any small portion of the wave front may be represented as in Fig. 41-7. There are electric or electrostatic lines of force which move up and down vertically at the carrier frequency. There are magnetic lines of force which move one way and the other horizontally at the carrier frequency. While all this is happening, the wave as a whole moves forward at approximately the speed of light, about 186,000 miles per second.

If the plane of the loop is in line with the direction of wave travel, as in Fig. 41-7, the magnetic field

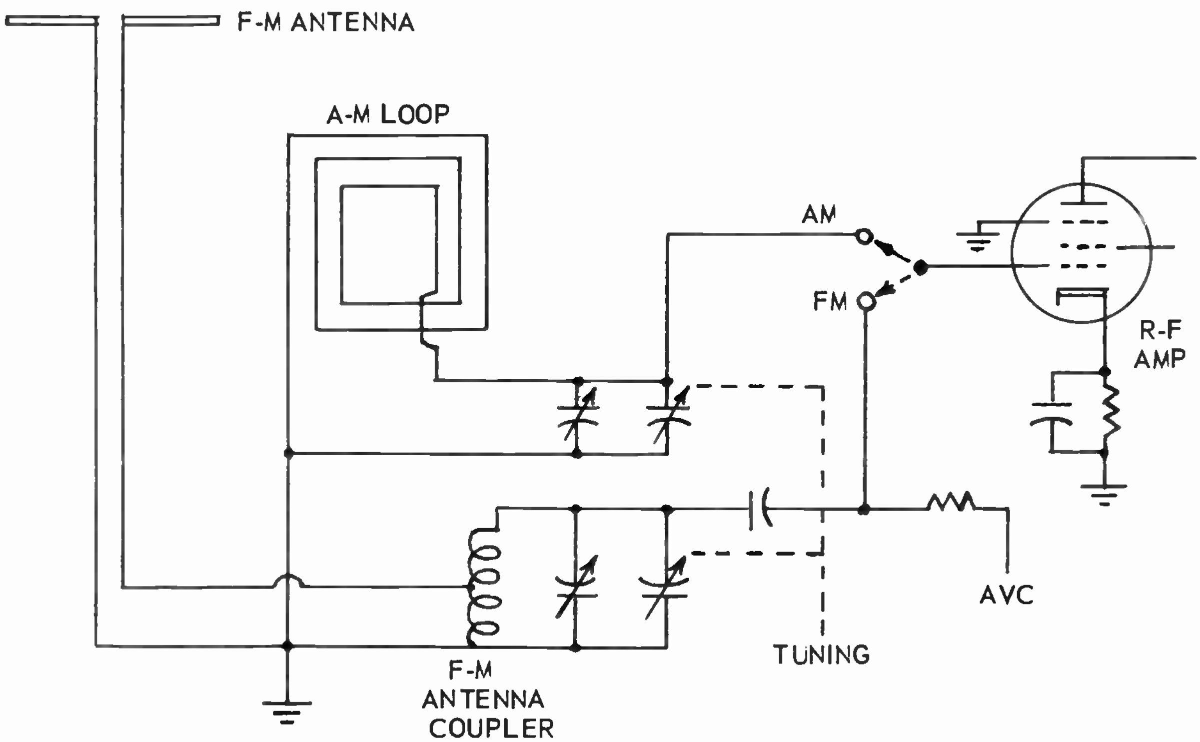


Fig. 41-6. Connections of an a-m loop and an f-m dipole antenna to the r-f amplifier.

COMMERCIAL TRADES INSTITUTE

TELEVISION

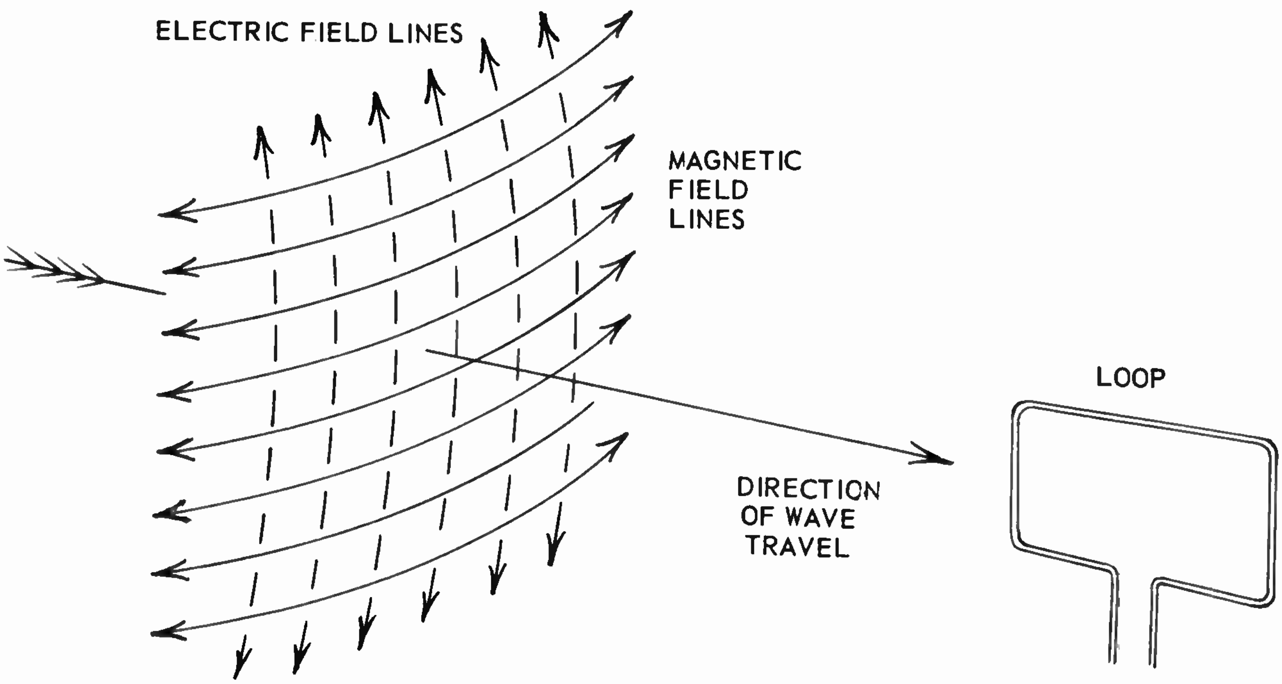


Fig. 41-7. Electric and magnetic fields of an a-m broadcast electromagnetic signal wave.

lines cut first through the conductors on the side of the loop which is toward the transmitter. A minute fraction of a second later the magnetic lines cut through the side of the loop which is away from the transmitter.

The length of conductor cut by the magnetic lines on one side of the loop is the same as the length on the other side. Consequently, the induced emf's are equal on both sides of the loop. The width of the loop from side to side is so little, and the velocity of the wave front so great, that emf's induced in both sides will be of the same polarity except at the instants in which there are reversals of field line direction.

Now we have the conditions represented at the left in Fig. 41-8. During one instant the induced emf's act in an upward direction in both sides of the loop, as at A. When the direction of the magnetic lines reverses in the carrier wave the induced emf's act downward in both sides, as at B. So far as the output terminals of the loop are concerned, the emf's in the two sides oppose each other in direction. Since the two emf's are equal it might seem that they should cancel and leave zero output.

The slight difference in times at which the oncoming wave front cuts the two sides of the loop saves the day. There is a slight phase difference between the two induced voltages. This is shown at C in Fig. 41-8. The emf induced in the side of the loop which is nearer the transmitter has a slight lead over the emf induced in the far side. When these two voltages combine at the output of the loop they do not quite cancel, but produce a net output voltage as shown at D. Unless the width of the loop were very great in relation to the wavelength in space of the carrier signal the net output voltage from the loop would be nowhere near as great in relation to the separate voltages as shown by Fig. 41-8, but it is great enough to satisfy the needs of a reasonably sensitive radio receiver.

10 Signal voltage obtained from a loop increases with greater width of the loop, because the greater the distance traveled by the signal wave between opposite sides the greater is the phase difference of emf's induced in the two sides. Output signal voltage is increased by making the loop higher, for then there are greater lengths of conductor being cut by magnetic lines of force on both sides, and induced emf's are proportional to the cutting. Output voltage is increased also by using more turns of conductor in the loop, for this too increases the total length of conductor being cut by the magnetic lines.

Inductance of the loop is increased by more turns, by greater height, and by greater width. Only so much inductance can be used, because with more and more inductance we must use less and less tuning capacitance to reach given frequencies of resonance, and soon the tuning capacitance becomes too small to allow covering the broadcast band of frequencies.

Signal voltage from a loop increases at higher received frequencies. The higher the frequency the shorter becomes the wavelength of the signals in space. For a loop of any given width this means a greater phase difference between emf's induced in the opposite sides. Were it possible to make the loop as wide as a half wavelength of the received carrier the emf's in opposite sides would be 180 degrees out of phase or in phase. Then the emf in one side of the loop would act upward while the emf in the other side is acting downward, and the two emf's would add together in the output rather than opposing each other as in Fig. 41-8. A half-wavelength loop for 1,000 kilocycles would be 492 feet wide.

A loop has rather pronounced directional properties. When the plane of the loop is in line with the direction of signal wave travel, as at *A* in Fig. 41-9, there is maximum signal output voltage from the loop

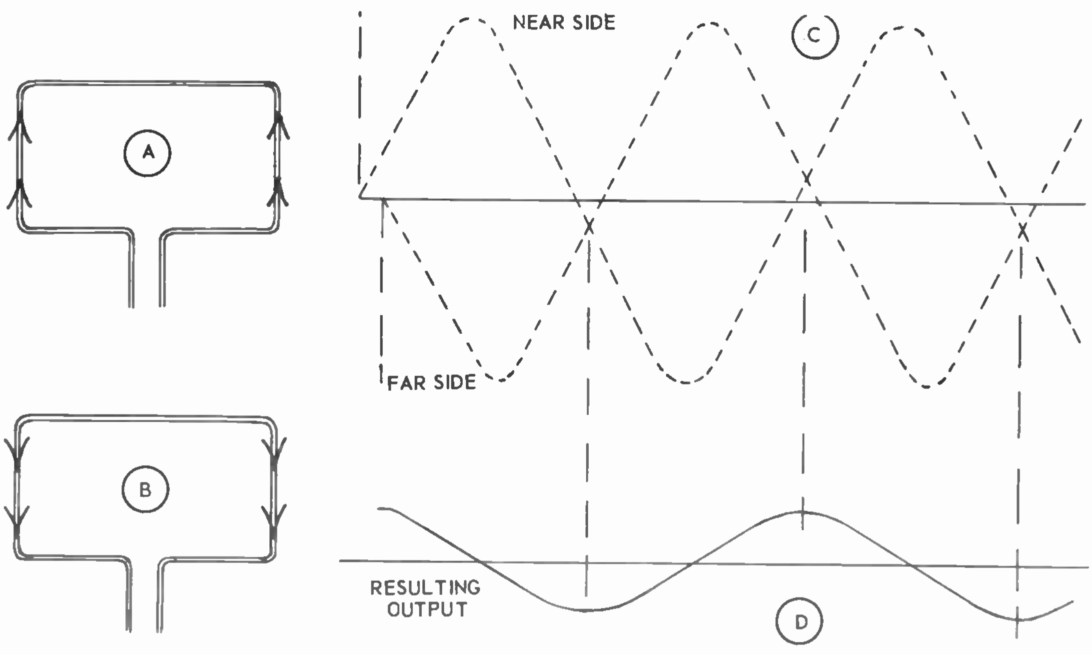


Fig. 41-8. Effects of emf's simultaneously induced in the two sides of a loop.

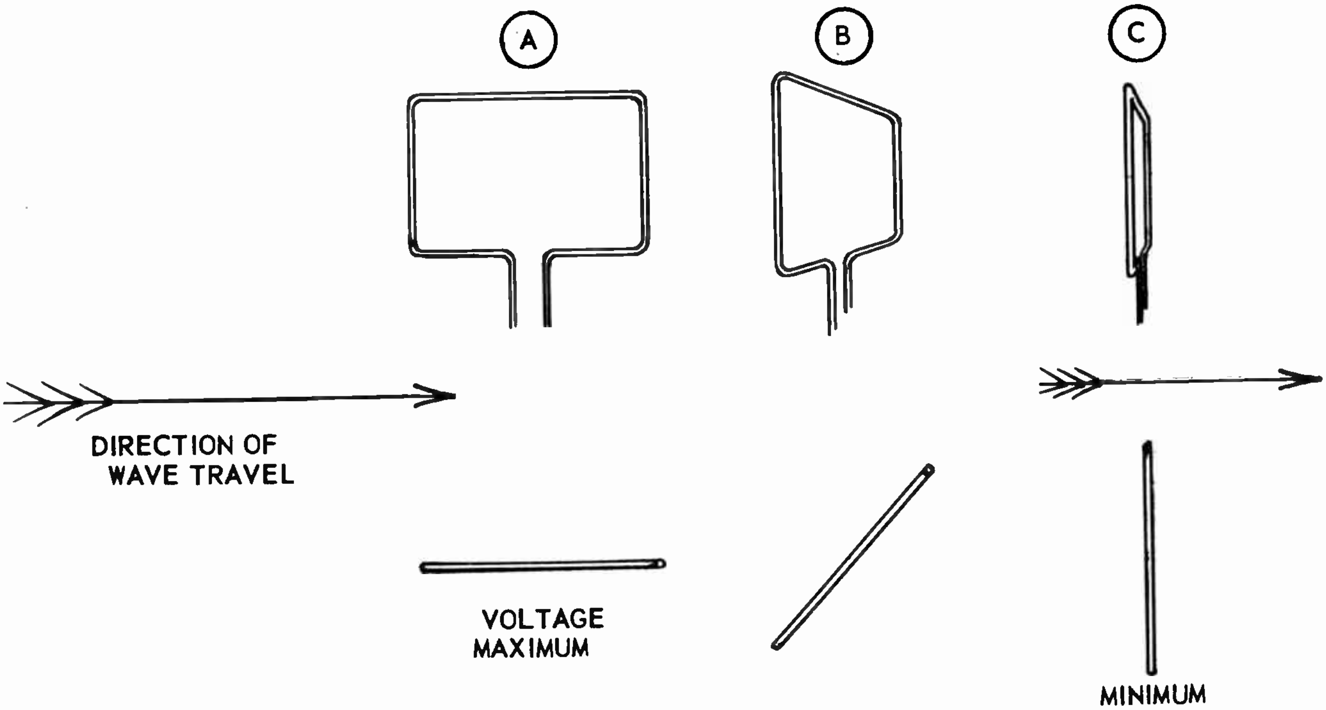


Fig. 41-9. Directional properties of a loop.

terminals. This is because, with the loop in this position, the signalwave travels the maximum possible distance between opposite sides of the loop, there is maximum phase difference between emf's in opposite sides, and maximum resulting voltage.

With the loop turned at an angle to the direction of wave travel, as at *B*, there is a shorter time interval between instants at which the wave front cuts opposite sides of the loop. Then there is less phase difference between the emf's, and a lesser output voltage. With the loop turned broadside to the direction of wave travel, as at *C*, the wave front cuts both sides of the loop at the same instant, there is zero phase difference between the two emf's, and the loop pickup drops to zero. Even with the loop broadside to the direction of wave travel the signal input to the receiver circuits does not drop to zero, for the loop still acts in the manner of any ordinary outdoor or indoor antenna and picks up some signal effect from the electrostatic field of the signal wave.

Nearby broadcasting stations are received almost equally well with the loop in any position except very nearly broadside to the direction of wave travel. With the exact broadside position there is a fairly sharp minimum of reception. Distant stations are satisfactorily received only with the loop edgewise or nearly edgewise to the direction of wave travel, as at *A* in Fig. 41-9. Signal pickup falls off very rapidly as the loop is turned to any other position.

A receiver having a built-in loop often is provided with an additional terminal for connection of an outdoor or indoor antenna which may be used in connection with the loop. This extra antenna terminal may be connected through either a capacitor or a resistor to any point on the grid circuit of the r-f amplifier which connects to the loop through the band switch in the a-m position.

Fig. 41-10 shows another method of using an external antenna with a built-in loop. There are two or three additional turns of conductor around the outer edge of the loop. These turns are connected between the external antenna and ground. Then these external antenna turns act like the primary winding of an antenna coupling transformer. The turns of the loop act as the secondary winding of this transformer, which is tuned to resonance by the usual trimmer and variable tuning capacitors.

The diagram of Fig. 41-10 shows a loop connected to the grid circuit of a converter tube. Fig. 41-6 shows a loop connected to the grid circuit of an r-f amplifier. One diagram applies to a combination fm-am set, and the other applies to a straight a-m set. The different diagrams are used to illustrate that a loop may be used for a-m reception with any type of receiver.

ALIGNMENT WITH LOOPS. When aligning the tuner sections of receivers equipped with built-in loop antennas the signal from the generator is introduced through the loop, which remains connected to the input grid circuits. There may be enough signal pickup through the loop when you merely place the high-side end of the signal generator cable within a few inches of the loop, with the low side end left free. There will be greater pickup if the low-side end of the cable is connected directly or through a fixed capacitor to the receiver chassis.

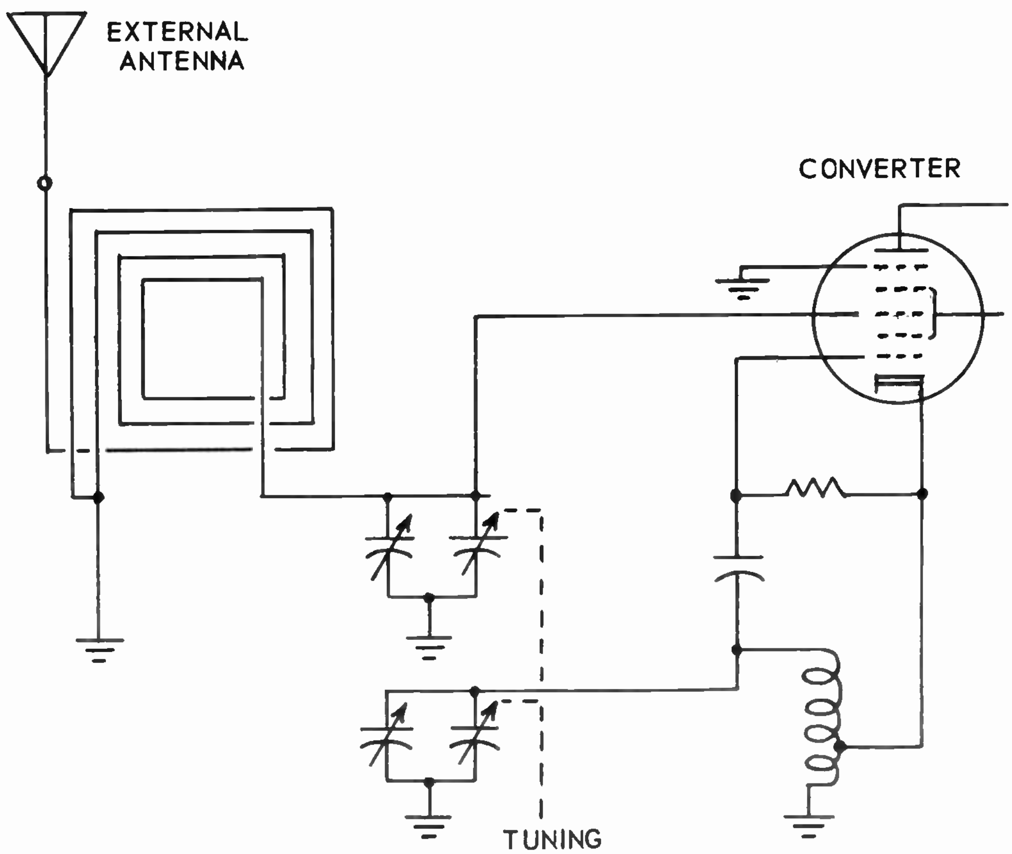


Fig. 41-10. An external or Marconi antenna coupled to a loop by means of added turns on the loop support.

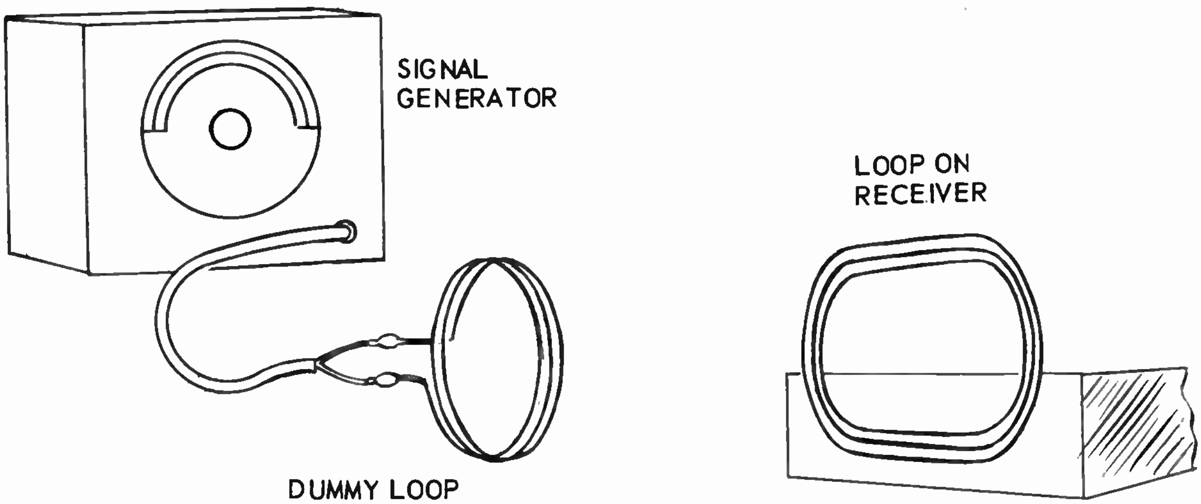


Fig. 41-11. Using a dummy loop for radiation of signals to a receiver loop.

If the loop support carries additional turns for an external antenna, as in Fig. 41-10, the signal generator may be connected to the external antenna terminals, with the external antenna disconnected. For alignment in the standard broadcast band the high side lead from the generator may be connected to the antenna terminal through fixed capacitance of about 1,000 mmf, or the connection may be direct. For short-wave alignment the high-side lead should be connected through a carbon resistor of 270 to 300 ohms.

The usual way of introducing a signal into the receiver loop is to use a dummy loop on the end of the generator output cable, as shown by Fig. 41-11. This loop may be made of any insulated wire that will be self-supporting, such as hookup wire or bell wire. Use two to four turns wound to a diameter somewhere between six and twelve inches. This dummy loop is placed one to three feet from the receiver loop, with the planes of the two loops parallel with each other. The ends of the generator output cable may be connected directly to the ends of the dummy loop, or a fixed capacitor of about 1,000 mmf may be used in series with the high-side lead.

The same dummy loop is used for all frequencies in the standard broadcast band, also for the short-wave bands. Place this loop as far from the receiver as will allow distinct readings on the meter used for alignment measurements. The farther apart the two loops are placed the weaker will be the signal picked up by the receiver, but the sharper will be the peaks of meter readings, and the adjustments are more likely to prove satisfactory for normal reception. Whatever may be the initial separation between the two loops it must not be altered during the entire process of tuner alignment. If you have to change the separation in order to increase or decrease the signal pickup, start the alignment all over again.

Instead of using either a loop or an external antenna, some receivers are fitted with rigid rods, metal plates, wire mesh screens, and other special devices for signal pickup. A dummy loop usually will be satisfactory for furnishing a signal through any of these special antennas, although it may be possible to obtain easier peaking with a dummy antenna of the same general form as the one used on the receiver. That is, for a receiver using a rod antenna, connect a similar rod to the high-side lead of the signal generator, for a metal plate antenna use a similar plate on the generator, and so on.

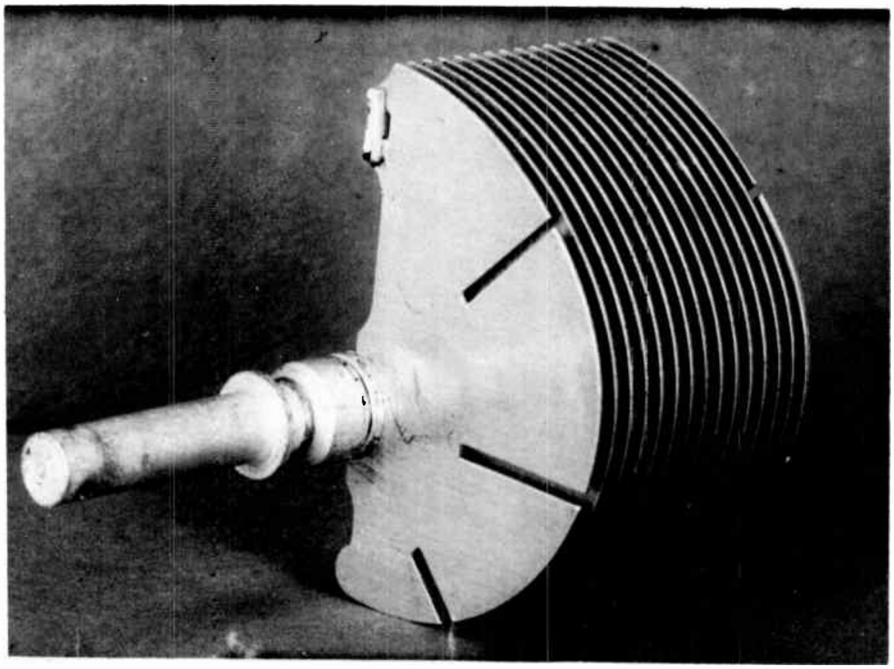


Fig. 41-12. Slotted end plates on the rotor of a tuning capacitor used for a-m reception.

You may run into difficulty in obtaining tracking of signal generator frequencies and receiver dial frequencies at both ends of the reception band when using a loop. The trouble usually could be overcome were it possible to alter the inductance of the loop to obtain tracking at the low-frequency end, while using the regular capacitance trimmers for adjustments at the high-frequency end. A few loops are constructed in such manner that one or more turns on the outside or inside may be moved farther from or closer to the other turns to vary the inductance.

Some tuning capacitors used for the standard broadcast band have slotted rotor plates at both ends of the group for making small changes of capacitance at various points in the tuning range. Fig. 41-12 is a picture showing such slotted rotors. When the capacitor is assembled these rotor plates are on the outside of the end stator plates. By bending the rotor sections between slots away from the stators the capacitance is reduced and the resonant frequency is raised over whatever portion of the tuning range has been covered by the sections that are bent.

Only the section of the rotor plate which is at the bottom in the picture would be in mesh with a stator plate for the high-frequency end of the tuning range, so bending this section would affect high-frequency tuning quite materially. It would affect tuning at lower frequencies only to a limited extent, for then there would be greater areas of the plates in mesh and the bent section would be a smaller fraction of the whole. Bending the following section would have a material effect on tuning just below the high end of the band. The next section would affect tuning just below the center of the range, and the last section (at the top in the picture) would chiefly affect the low-frequency tuning.

The regular types of adjustable trimmer capacitors usually are provided in addition to the slotted rotor plates. The trimmers ordinarily are relied upon for high-frequency alignment, while the rotor plates may be bent to improve tracking at the middle and low frequencies. The rotor plates can be bent far enough from the stators to cause considerable reduction of capacitance, but they cannot be bent much closer to the stators (to increase the capacitance) without danger of short circuits between the two kinds of plates.

OUTDOOR AND INDOOR ANTENNAS. Were we discussing radio receivers and reception in some year around 1925 the subject of outdoor antennas for a-m signals would be of major importance. Receivers of that day had little gain in themselves, and plenty of signal voltage from the antenna was needed to produce anything worth while from the speaker. Modern amplifier tubes have such high transconductances that present day standard broadcast receivers need only a few microvolts of signal from any type of antenna to deliver all the sound volume that can be used when transmitters are not too distant. Although outdoor and indoor antennas are not used for many broadcast receivers, we should examine the performance of such antennas as it differs from that of loops. Outdoor and indoor antennas are of the Marconi type, more often spoken of as open antennas or capacitance antennas. These names refer to antenna systems of which the earth or ground is an essential part.

(2) Some of the characteristics of a-m signal waves were mentioned in connection with Fig. 41-7. Radiation of these waves is carried out in such manner that the electric field lines are vertical. Such a wave is said to be vertically polarized. The term wave polarization refers to the direction of the electric or electrostatic component of the electromagnetic wave. For energizing the loop antenna we utilize the magnetic field lines, which act horizontally, but for energizing the Marconi antenna we utilize the vertically acting electric field lines. Incidentally, for f-m broadcast and for television video transmission we use waves that are horizontally polarized, with the electric field acting in a horizontal direction.

A Marconi antenna system is illustrated in principle by Fig. 41-13. The antenna consists of an elevated conductor, usually a long wire, which is connected to the earth or ground through a lead-in conductor, an inductance in the receiver circuits, and a ground wire. The elevated condnctor and the conductive earth act like the two plates of a large capacitor. Electric field lines of the moving signal wave act vertically

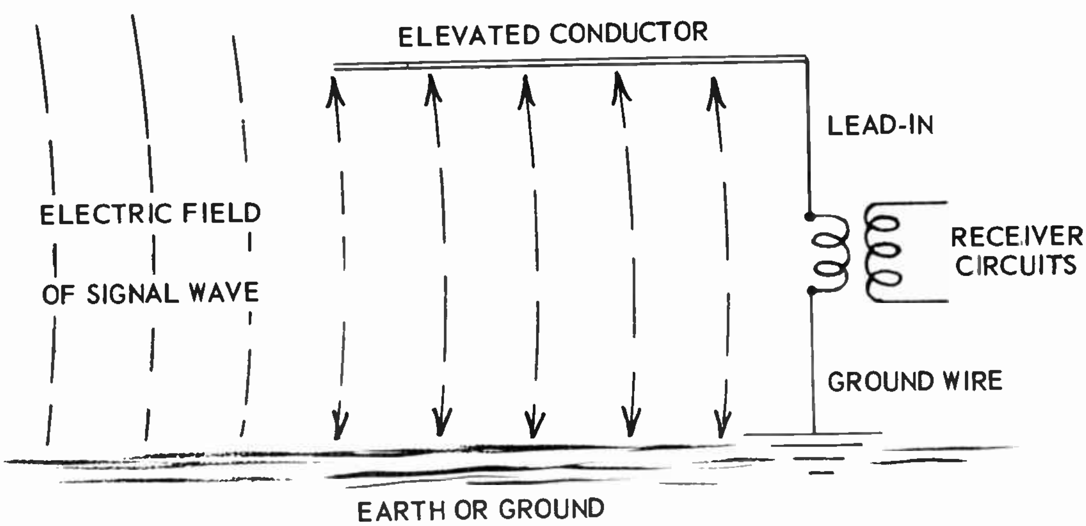


Fig. 41-13. How the electric field of the a-m signal wave charges the Marconi antenna.

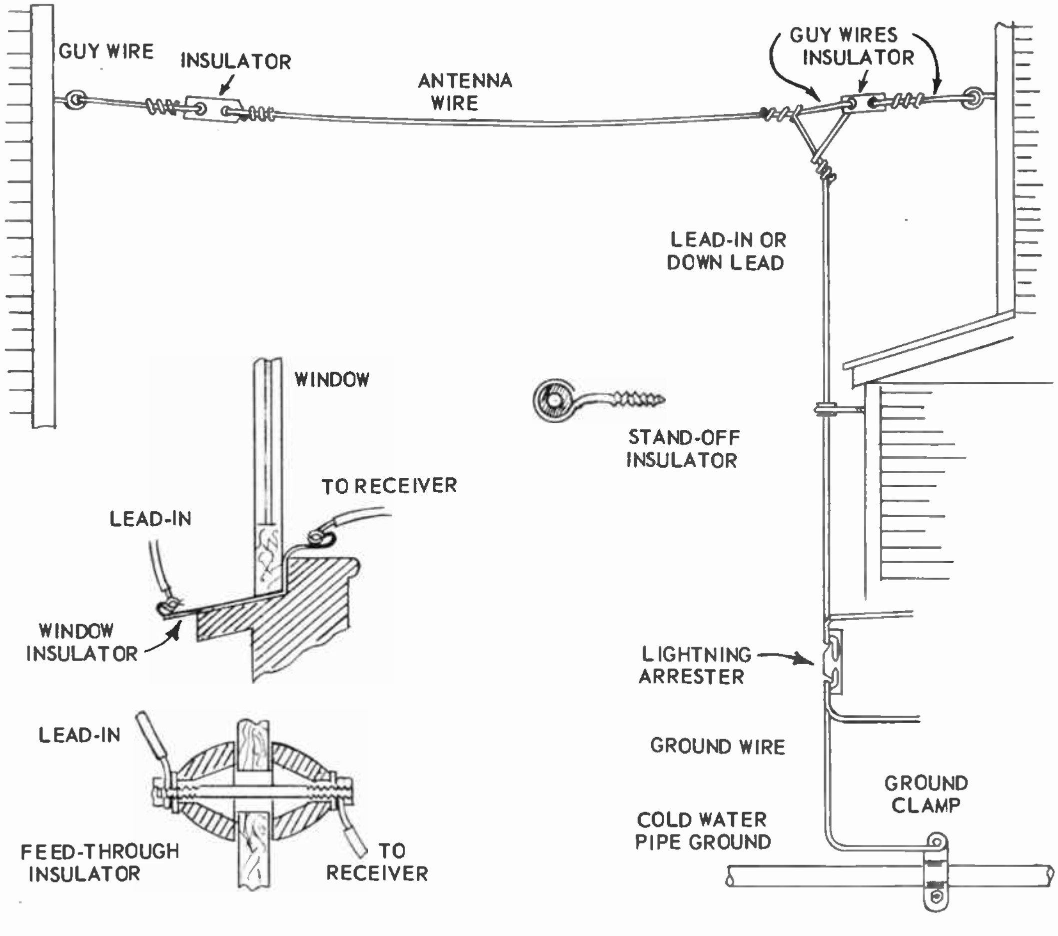


Fig. 41-14. An outdoor Marconi antenna with its fittings and accessories.

between these capacitor plates, charging the plates alternately in opposite polarities at the frequency of the signal. The charges induced on the elevated conductor and the conductive earth have potential differences that cause current at the signal frequency to flow through the inductance which is in the receiver circuits. This is all we need to introduce signal currents and voltages into the receiver.

The elevated conductor may be a wire 50 or more feet in length supported high above the ground out of doors, or it may be only a few feet of wire or of any other conductor in the room where the receiver is located. The ground connection may be made to cold water pipes which extend from the building into surrounding earth, or it may be to a metal rod driven a few feet into moist earth, or it may be to the receiver chassis and thence directly or through capacitors to the power and light wiring, in which one side always is grounded.

When the elevated conductor of a Marconi antenna is no more than 10 to 15 feet above the ground level it is subjected to electric and magnetic fields radiated from all manner of electrical devices. These electrical impulses are picked up by the antenna and amplified in the receiver to produce noise at the speaker. The higher the elevated conductor is raised the farther it is above all kinds of electrical interference. Furthermore, the radio signal waves are much stronger at greater elevations, they have not been so attenuated by passing through many obstructions.

4 To have a satisfactorily high ratio of signal to noise the elevated conductor must be high up. Otherwise there is no particular advantage in using this type of antenna. A long antenna wire will bring in more signal than a shorter wire, but the long wire may also bring in a disproportionate amount of noise and give poorer results than a relatively short wire. Much of the success in using a Marconi antenna depends on having connections of low resistance throughout the system and on getting the ground wire connected to some pipe or other conductor that goes down into earth that remains permanently moist the year around.

Since the lead-in wire from the elevated conductor to the receiver has to come down through a region of rather strong electrical interference in most localities this wire may pick up a lot of interference. Such trouble may be greatly reduced by using shielded wire for the lead-in, with the outer shield connected to ground at the receiver end.

Several features of antenna installations are illustrated by Fig. 41-14. The elevated conductor, called the antenna wire, is supported by insulators which, in turn, are supported from any immovable solid parts of buildings or other structures. The antenna wire usually is bare copper of seven strands, each consisting of a size between gauges number 22 to 26. Solid enameled wire may be used in gauge numbers 12 or 14.

1 Antenna insulators most often are of glazed porcelain or of glass whose smooth surfaces do not encourage dirt deposits. The sections or forms may be straight cylindrical or square, or may be ridged to increase the length of current leakage path. Diameters vary from 5/8 to 1 inch and lengths run from about 3 to 12 inches.

The lead-in or down-lead conductor in the illustration is a continuation of the antenna wire. A separate lead-in ordinarily is of number 18 stranded copper wire with rubber or composition insulation. If the lead-in is not a continuation of the antenna wire the joint between them should invariably be soldered. Otherwise there will be corrosion from long exposure to air and atmospheric impurities with resulting severe loss of signal strength.

The lead-in is prevented from making contact with building or supporting structural elements by means of stand-off insulators, of which many types are available. The one illustrated consists of a loop on the outer end of a wood-screwshank, with a rubber or composition insulating bushing held in the loop and enclosing the lead-in wire. These stand-offs are used at as many points as may be necessary to keep the conductor in the clear.

A lightning arrester approved by the Underwriters' Laboratories always should be used between the lead-in and ground wire for an elevated outdoor antenna. Some arresters consist of a small air gap between two metal points mounted on or inside of insulation. Lightning discharges will jump the gap and go to ground, but signal voltages will not do so. The gap may be in open air or in a vacuum. Other arresters have a high resistance element which diverts signal voltages through the receiver but allows discharge through it of atmospheric electricity. Still others make use of a small neon bulb which forms an open circuit for signal voltages but breaks down and conducts at potential differences exceeding about 100 volts.

The ground wire which connects to the receiver and to one side of the lightning arrester may be of the same material as the lead-in wire or may be uninsulated. The ground wire is preferably of number 12 gauge or larger in diameter. Aluminum wire often is used for grounding. The ground wire usually is fastened to a ground clamp that fits on or around any cold water pipe in the building. Cold water pipes are conductively continuous with piping that goes underground to water mains. Hot water pipes and gas pipes sometimes are interrupted by insulating joints, and do not form satisfactory grounds. If a cold water pipe is not accessible its place may be taken by a ground rod. This usually is a copper plated steel rod 3/8 to 1/2 inch in diameter, about four feet long, pointed at one end and having a wire clamp at the other end. The ground rod is driven down to permanently moist earth and the ground wire is fastened to the upper exposed end.

At the lower left in Fig. 41-14 are two styles of insulators for bringing the lead-in conductor indoors so that connection may be made to the receiver. The upper illustration shows a window strip consisting of an insulated flexible copper strap which is very thin, about 1/2 inch wide, and about 10 to 12 inches long. This strip is laid across the window sill and the window closed down on it. A spring clip on one end holds the lead-in conductor while a similar clip on the other end holds the wire going to the receiver. A clip of this style used out of doors does not long maintain a good electrical joint, it corrodes and then has high resistance.

The feed-through insulator consists of a metal stud, threaded on both ends, which is passed through an opening 1/4 to 1 inch diameter in a thin wall, partition, or panel. The opening is closed on both sides with insulating housings held in place by nuts and washers on the ends of the stud. The lead-in is clamped

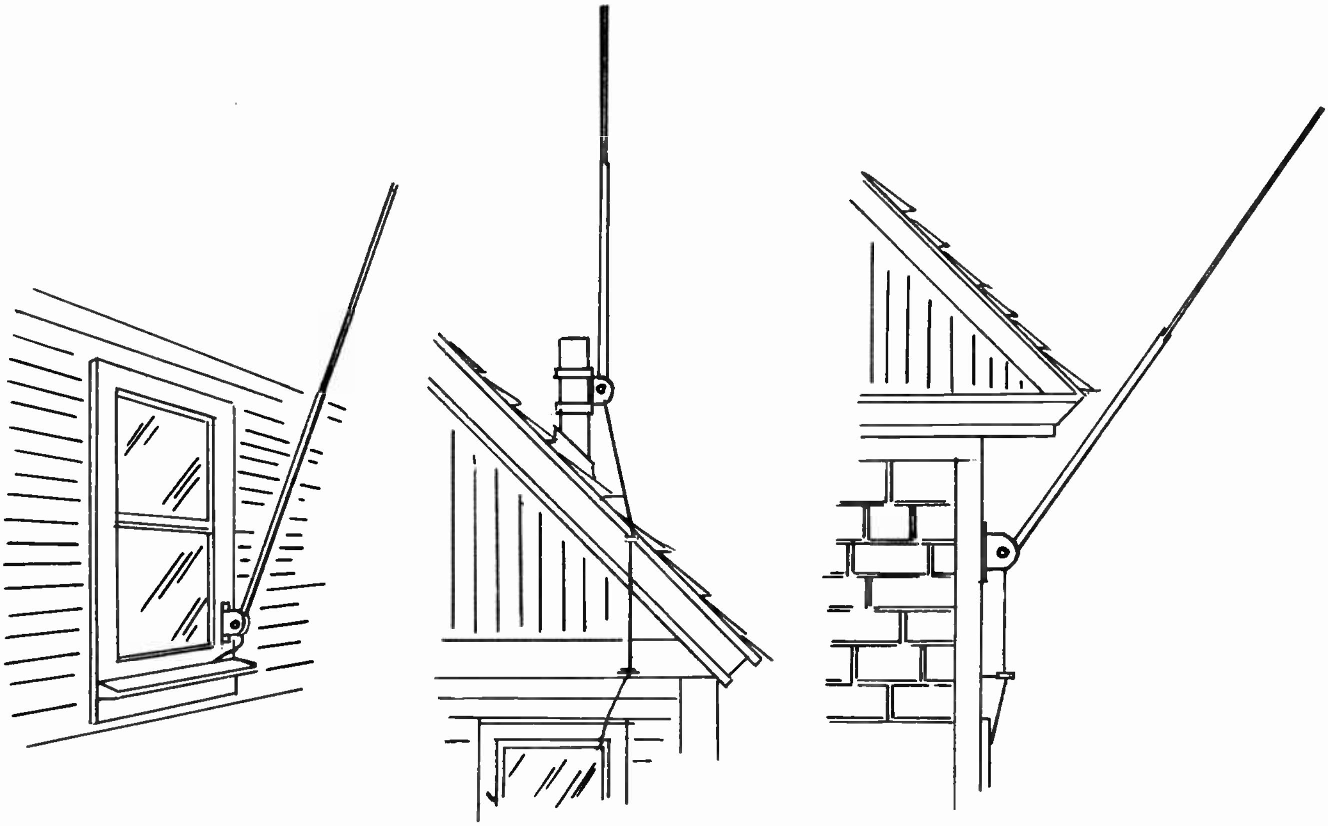


Fig. 41-15. Some of the mounting positions which may be used for a rod antenna.

under the outer washer and the wire to the receiver under the inner washer. The ends of both wires should be fitted with solder lugs.

ROD ANTENNAS. Vertical or inclined rod antennas which are somewhat similar to the whip antennas for automobile radios may be used for standard broadcast and short wave reception. A few of the many possible installation positions are shown by Fig. 41-15, at the left, on a window frame; at the center, on a vent stack; and at the right, on a side wall. Parapet walls, roof ridges, and chimneys may be used. Rod antennas usually are carried by an adjustable bracket which allows rotation and any angle between horizontal and vertical. The position will depend on the kind of support which is available, on necessary clearance from surrounding objects, and on suiting the position to the best possible reception.

The rods usually are of telescoping construction, with maximum lengths of 6 to 20 feet or even more. Materials are steel, aluminum, or bronze, sometimes plated with cadmium, nickel, or chrome. Rod antennas may be used on receivers equipped with loops. The two antennas may give increased signal strength from certain directions in which the response of the loop along would be near its minimum.

TELEVISION

Lesson No. 42

THE TELEVISION SIGNAL

Training Your Mind

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON 42 THE TELEVISION SIGNAL

We are well acquainted with the principles of tuning, amplification, and detection employed in both a-m and f-m sound receivers. All these principles are used in television receivers. The television tuning system is a superheterodyne type. Television sound is f-m sound, with i-f amplifiers and demodulators like those in f-m receivers. Television picture signals are amplitude modulated, and after being amplifier these signals are demodulated by the familiar diode detector. All our study of A-M and F-M receivers has provided a foundation for study of television receivers, which we now are ready to undertake.

The best possible introduction to the whole subject of television receivers will be for you to read over again the first three lessons, where we talked about the requirements for picture reproduction and about how these requirements are met. Your reviewing of those lessons will make it possible to omit all the preliminary explanations which ordinarily would be needed right here, and we can get down to the real business in hand without delay.

SCANNING. According to present standards for black and white picture transmission one complete picture is traced during each 1/30 second of time. One complete picture is called a frame. There are 30 frames per second, so the period for one frame is 1/30 second. The frame frequency is 30 per second or 30 cycles per second.

During the first half of each frame period, or during the first 1/60 second, only half of the horizontal lines required for a complete picture are transmitted and traced on the screen of the picture tube. As shown at the upper left in Fig. 42-2, only every alternate line is traced during this period. The electron beam in the picture tube is started from the upper left-hand corner of the picture space and traces the first horizontal line, from 1 to 1. This is called an active trace, because the electron beam is causing the phosphor of the picture tube screen to become luminous.

Next the electron beam is blanked or shut off. Voltages or currents which cause horizontal movement of the beam return to the values which place the beam at the left side of the picture space. Were the beam not blanked, these voltage or current changes would have moved the beam from the end of line 1 to the beginning point for line 3 of the picture. This is called a horizontal retrace. When it comes time to start the next luminous line, number 3, the blanking ceases and the beam traces the luminous line from left to right on the screen. Active traces and intervening retrace periods continue for the odd numbered lines in the picture until the beam reaches the lower right-hand corner of the picture space, at *A* on the diagram. To simplify the diagram only a few traces and retraces are shown. Actually there would be about 244 traces during the period of time here represented.

While the electron beam is being deflected or swept horizontally by one current or voltage it is being moved gradually downward by another current or voltage. We have been examining the first downward deflection, which has consumed a total time of about 15,493 microseconds, or millionths of a second.

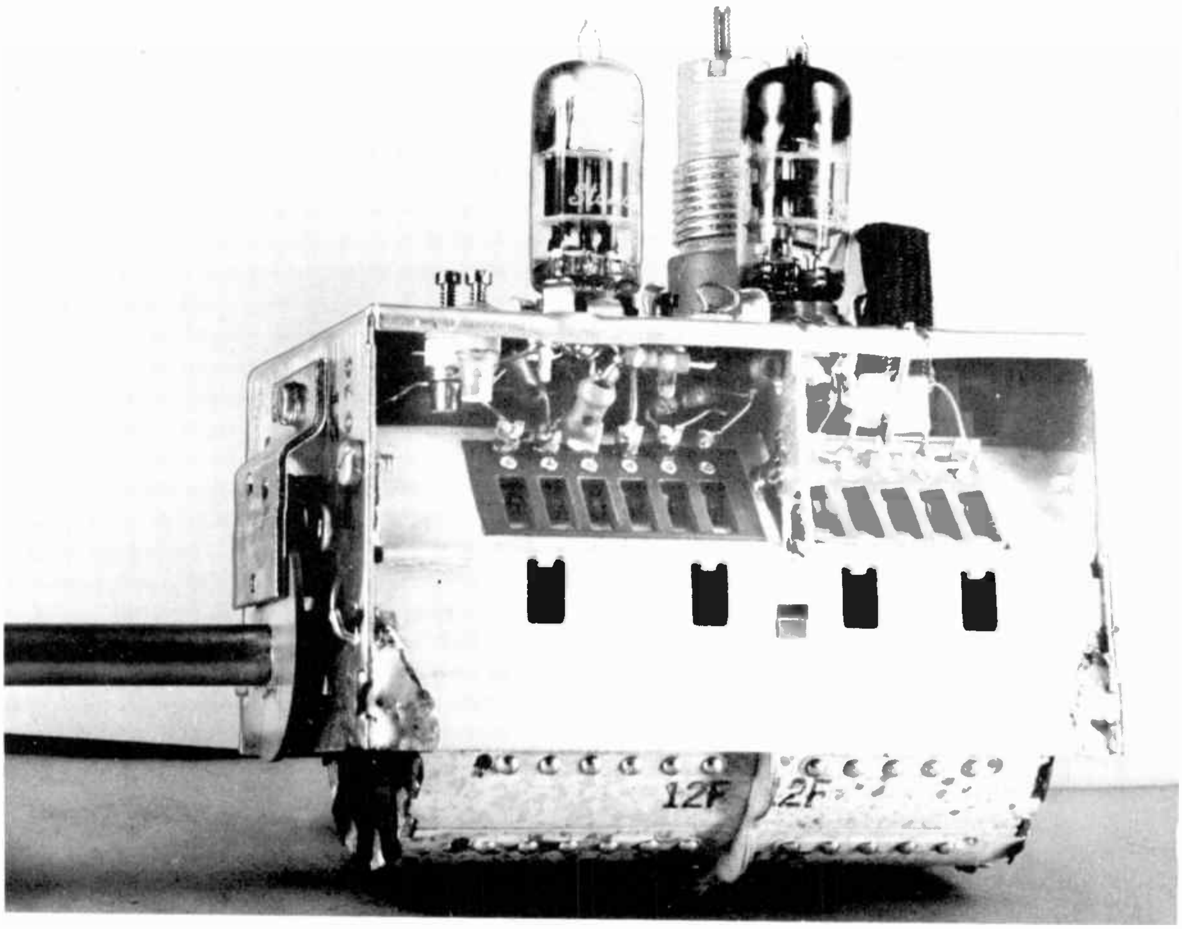


Fig. 42-1. A turret tuner for television receivers. The left-hand tube is a twin-triode oscillator and mixer. At the right is the r-f amplifier.

At point A the beam is blanked. Almost immediately the current or voltage which has caused downward movement reverses its polarity, and, were the beam not blanked, it would move the beam upward. The first vertical blanking period is shown at the upper right in Fig. 42-2. While one current or voltage is of polarity to move the beam upward, the current or voltage which has caused horizontal deflection continues to act. The vertical and horizontal currents or voltages act in a manner which would move the beam from A to B and thence on upward in a zig-zag path to point C, which is half way across the picture space. This vertical blanking period, during which vertical retrace occurs, takes up about 1,174 microseconds of time.

The downward deflection during which only alternate picture lines are formed is called one field. The field plus the following vertical blanking period, consume a total time of 16.667 microseconds, or 1/60 second. This is just half the time period for one frame. There are two fields or field periods in every frame. Field frequency is 60 times per second, or 60 cycles per second.

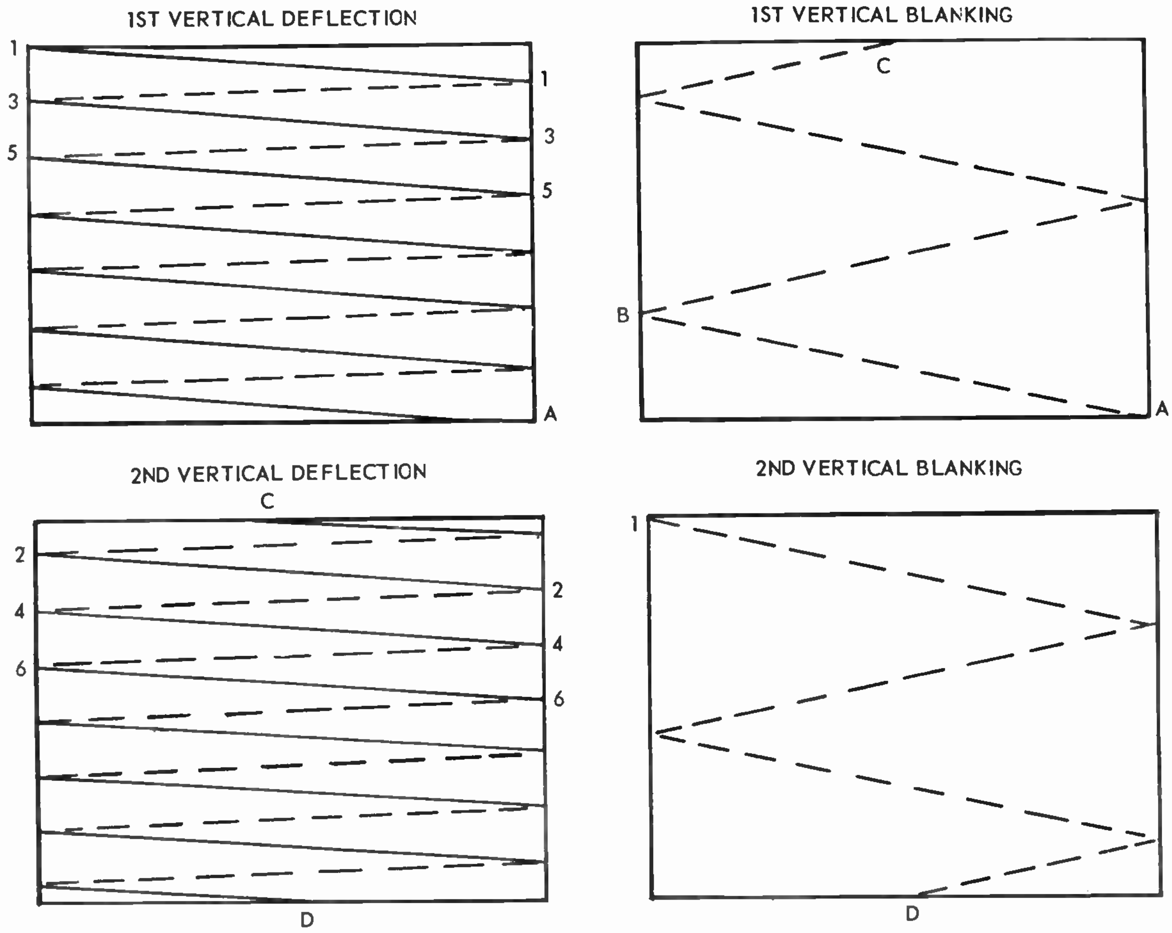


Fig. 42-2. Traces and retraces during scanning of the two fields in one frame.

When the deflecting currents or voltages reach values which would bring the beam to point C at the end of the first vertical blanking period the beam is again turned on. Then the beam starts from point C for the second downward vertical deflection and traces a half line to the right. Here we have begun the second field. This second field must begin with a half line, otherwise there would not be correct interlacing. Interlacing is the action by which active lines of the second field are traced midway between active lines of the first field in order to complete the picture or the frame.

If you examine the two diagrams at the left in Fig. 42-2 it is apparent that active line 2 will fall midway between lines 1 and 3, that line 4 will fall midway between 3 and 5, and so on. Because of persistence of vision we see all the lines of both fields at once, and detail in the picture is just as good as though the lines were traced consecutively instead of alternately. We notice no flickering effect, because each field forms a picture complete enough to be recognizable, and the result is as though we had 60 pictures per second instead of only 30. If we actually did have 60 complete pictures per second, with all lines traced consecutively, the rate of line transmission would have to be twice as fast. That is, instead of having 244 active lines in 1/60 second we would have to have 488 lines. There is plenty of difficulty in handling the lower line frequency allowed by interlacing, and there would be much more difficulty were this frequency to be doubled.

The second vertical downward deflection is completed with active lines and retraces as shown by the diagram. The last active line in this field ends half way across from left to right. Then the beam is cut off for the next vertical blanking period. The vertical blanking commences at *D*, and ends with the deflecting currents or voltages of such values as would bring the beam at the upper left-hand corner of the picture space, at point 1. Now the beam is again turned on and commences tracing the first active line of the next field, as in the diagram for the first vertical deflection. So the action continues so long as the receiver operates.

The entire action illustrated by Fig. 42-2 occurs during one frame period or during 1/30 second. In each field there have been 244 active lines and accompanying retraces, also enough horizontal changes of deflecting currents or voltages occur to have produced 18½ additional lines and retraces during one vertical blanking period. If you add these figures, and multiply by 2 for the two fields of a frame, the total will be 525 horizontal lines and retraces for the frame. The time required for one active line and one retrace is called a line period. There are 525 line periods, usually spoken of as 525 lines, per frame for standard black and white picture transmission and reception in television. In 1935 the standard was 343 lines per frame. Later it was raised to 441 lines. Since 1940 the standard has been 525 lines. Many people think it will go as high as 1,000 lines in the future.

With 525 lines, or line periods, per frame and with 30 frames per second, the line frequency must be the product, or 15,750 lines or cycles per second. For correct adjustment of many service instruments you must remember these frequencies.

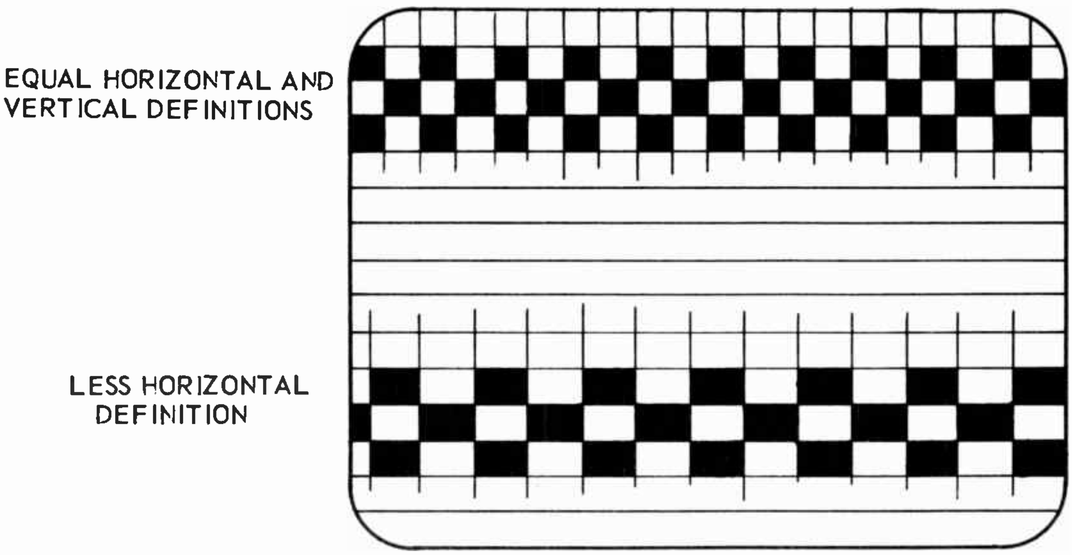


Fig. 42-3. Ideal and practical degrees of horizontal definition.

Frame frequency	30 cycles per second
Field frequency	60 cycles per second
Line frequency	15,750 cycles per second

When there are 525 line periods per frame the number of luminous horizontal lines in the picture will be close to our assumed value of 488. This is true regardless of the picture tube size. On a big tube the lines are farther apart, and on a small tube they are closer together. To have pictures appear in equal detail or definition you have to sit farther from a large tube than a small one, if the number of lines actually is the same. Reproduction systems in receivers having small picture tubes usually are not as good as in big sets, and the possible advantage of very close line spacing may not be fully realized in the small job.

10½ As originally viewed in the camera tube the image of every televised scene is of such proportions that the ratio of width to height is 4 to 3 or 4/3. This is called the standard aspect ratio. To show everything viewed in the original scene the reproduction on the picture tube screen must have this same aspect ratio. The picture might be 4 inches wide and 3 inches high, or 8 inches wide and 6 inches high, or 16 inches wide and 12 inches high, or of any other dimensions conforming to the standard ratio. We shall have more to say about actual picture sizes when discussing the subject of masks for picture tubes.

TELEVISION CHANNELS. For a-m standard broadcast transmission the sideband frequencies for audio modulation may extend as much as 5,000 cycles or 5 kilocycles below and above the carrier frequency, necessitating channels 10 kilocycles wide. For f-m broadcast the frequency deviations may go to 75 kilocycles below and above the center frequency, which is provided for by channels 200 kilocycles wide. Now let's determine the frequency width required for a television broadcast channel.

In Fig. 42-3 a picture screen area is ruled off into a few horizontal spaces which may represent the path travelled by the electron beam to form a large number of horizontal traces actually used on the screen. No object can be shown in sharp detail if its horizontal edges or boundary lines cover less space than one horizontal trace. If this degree of detail reproduction is to be equalled for vertical edges or boundaries, the vertical lines may be no wider than a trace is high. To attain such definition the "elemental areas" of the picture would be considered as squares whose sides equal the height of one trace. If there are 488 traces from top to bottom, and an aspect ratio of 4 to 3, there would be 650 areas crosswise.

In practice we do not need such good horizontal definition but, as shown lower in the diagram, may have minimum horizontal changes half again as great as the vertical changes from trace to trace. Instead of 650 possible changes of shading horizontally we then would have only 433 such changes. Each change from light to dark and back to light, or from dark to light to dark, would require one cycle of picture signal voltage. The 433 horizontal changes would require 216½ or, to avoid fractions, 217 cycles of picture signal voltage for each horizontal trace. One horizontal trace takes up only 5/6 of a line period, which includes a retrace, so figuring on the time per line period we would have a maximum of 260 cycles of picture signal

There are 525 line periods per frame. Even though some are used during vertical retraces, we still must multiply by 525 to determine the maximum frequency. There are 30 complete frames each second. Then the maximum signal cycles per second would be the product of 260 (cycles per line) by 525 (lines per frame) by 30 (frames per second). This comes to 4,095,000 cycles per second or about 4 megacycles per second. This would be the frequency for one sideband in the transmitted signal. Two sidebands would require 8 megacycles. In the same channel would have to be room for the accompanying sound signal and small guard bands above and below all modulation. Yet television channels are only 6 megacycles wide, including pictures, synchronizing, sound, and guard bands.

VESTIGIAL SIDEBAND TRANSMISSION. For ordinary double sideband transmission the frequencies in a television channel would have to be arranged as at the top of Fig. 42-4. Each of the sidebands for am-

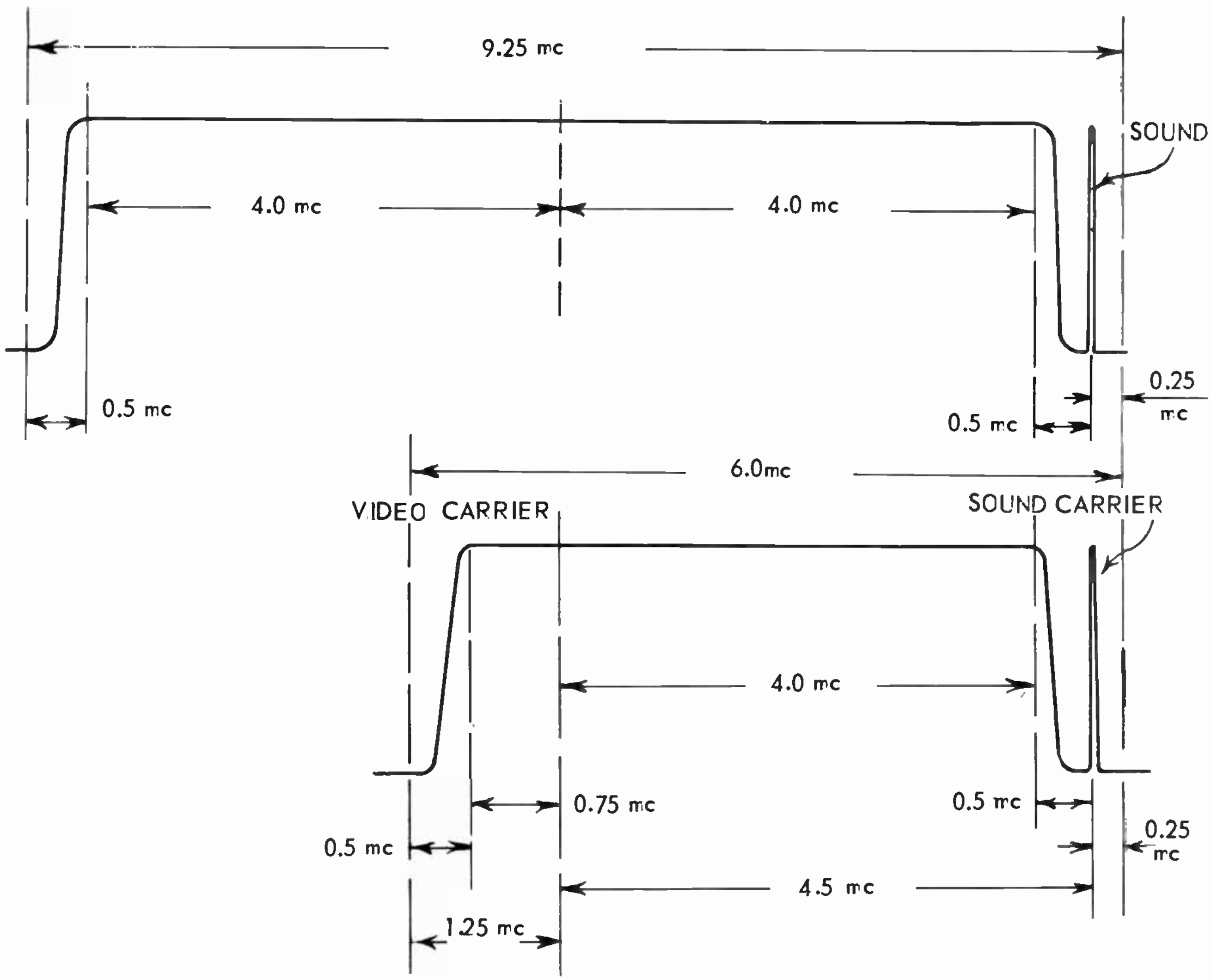


Fig. 42-4. The difference between double sideband transmission and vestigial sideband transmission.

plitude-modulated picture and sync signals would be 4 mc wide. F-m sound deviation for television is a maximum of only 25 kc below and above the sound carrier frequency, for a total of 50 kc or 0.05 mc. This sound carrier is 0.5 mc above the highest picture frequency. Above the sound carrier there is a 0.25 mc guard band, and below the lowest picture frequency is a 0.5 mc guard band to prevent interference with adjacent channels. The total width of the required channel would be 9.25 mc.

The lower diagram shows how channel width is reduced while preserving the signals. We must realize, to begin with, that amplitude modulation frequencies in a lower sideband are the same as in the accompanying upper sideband. The same modulating frequencies are simultaneously transmitted in both sidebands. If we omitted one of the sidebands completely the carrier still would be transmitting exactly the same modulation.

For television we do not cut off an entire sideband, but do cut off all but 0.75 mc of the lower-frequency

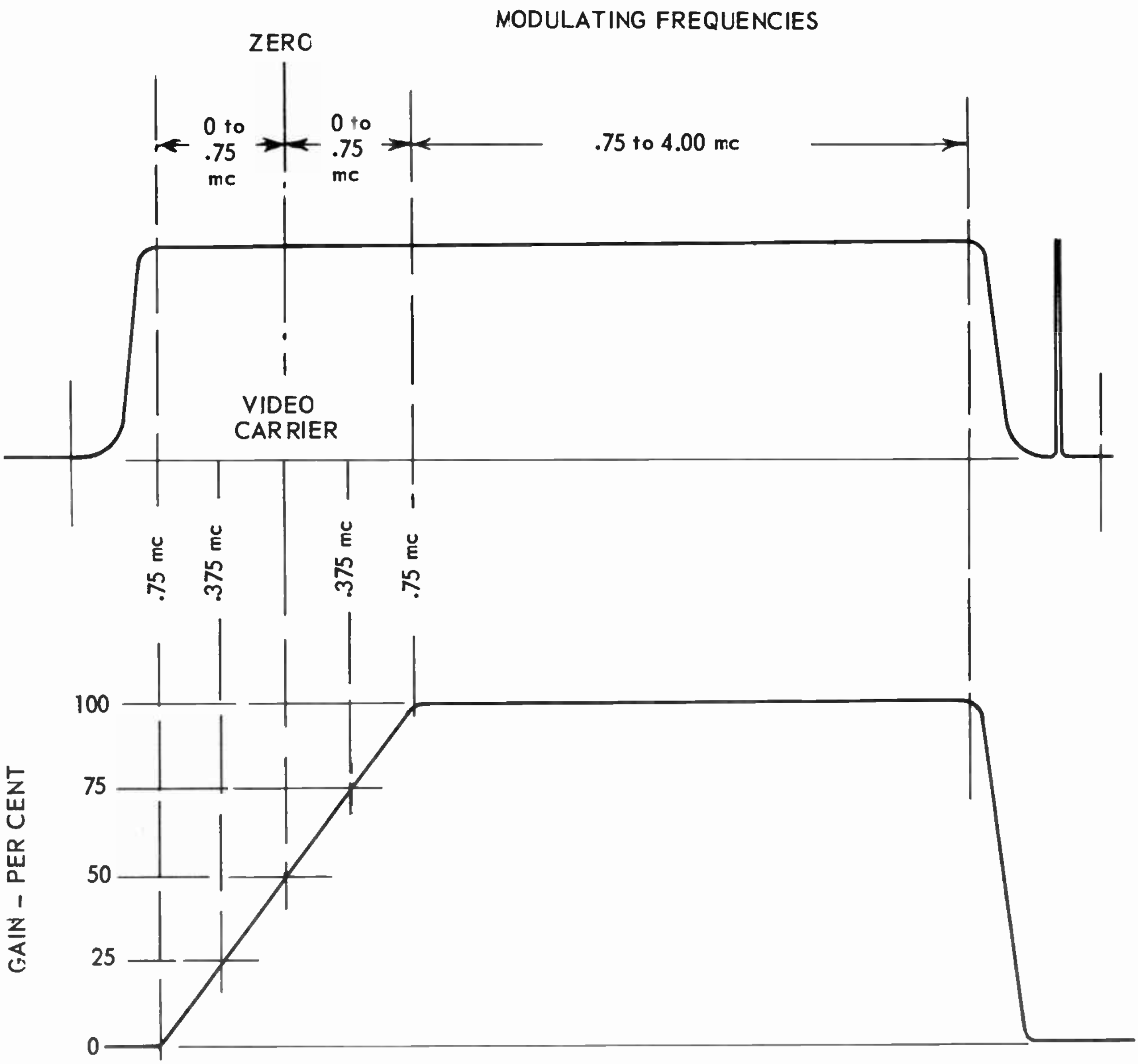


Fig. 42-5. How double strength of low-frequency transmission is evened out in the i-f amplifier response of the receiver.

(6) sideband while leaving the upper sideband in its full 4.0 mc width. The partial lower sideband is called a vestigial sideband. Placement of the sound carrier, and the widths of lower and upper guard bands are the same as before. Certain of the frequency differences shown here must be remembered during service work. (10d) First, and most important, the sound carrier always is exactly 4.5 mc higher than the video carrier, which is the carrier for picture and sync signals. The video carrier always is 1.25 mc above the lower limit of the channel. The sound carrier always is 0.25 mc below the upper limit of the channel.

All the modulating frequencies between zero and 0.75 mc are transmitted in both sidebands, as is made evident in the curve of transmission characteristic at the top of Fig. 42-5. This much of the modulation is transmitted and received in double strength. The remainder of the modulation, from 0.75 to 4.00 mc, is transmitted and received in single strength. Were the response of the receiver to be similar in shape to this transmission characteristic the signals delivered from the detector of the receiver would have twice as much amplitude for the low frequencies as for the high frequencies. This would distort the pictures.

The unequal strengths at various frequencies in the received signal is evened out in the receiver by making the frequency response of the i-f amplifier for video signals as nearly as practicable like that shown down below in Fig. 42-5. Gain applied for the video carrier is 50 per cent of maximum. At a frequency 0.75 mc lower than the carrier the gain is zero, and at 0.75 mc above the carrier the gain is 100 per cent. Gain increases at a uniform rate between these limits, and remains at 100 per cent for all frequencies from 0.75 mc above the video carrier to 4.00 mc.

Examine the gains applied to frequencies equally below and above the video carrier, which will be the same frequency so far as signal modulation is concerned.

At 0.75 mc below.	Gain 0	At 0.375 mc below	Gain 25%
At 0.75 mc above.	Gain 100%	At 0.375 mc above.	Gain 75%
Total gain.	100%	Total gain	100%

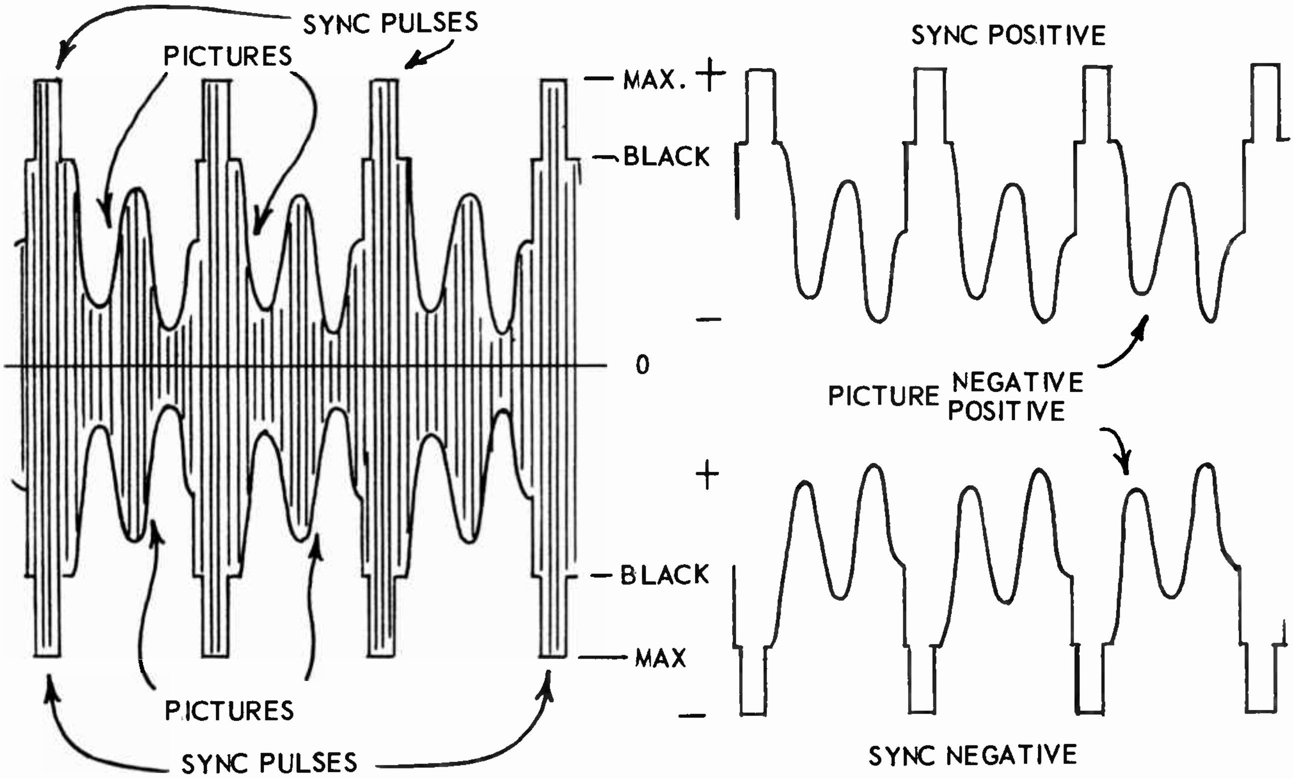


Fig. 42-6. The composite signal is amplitude modulation on the carrier.

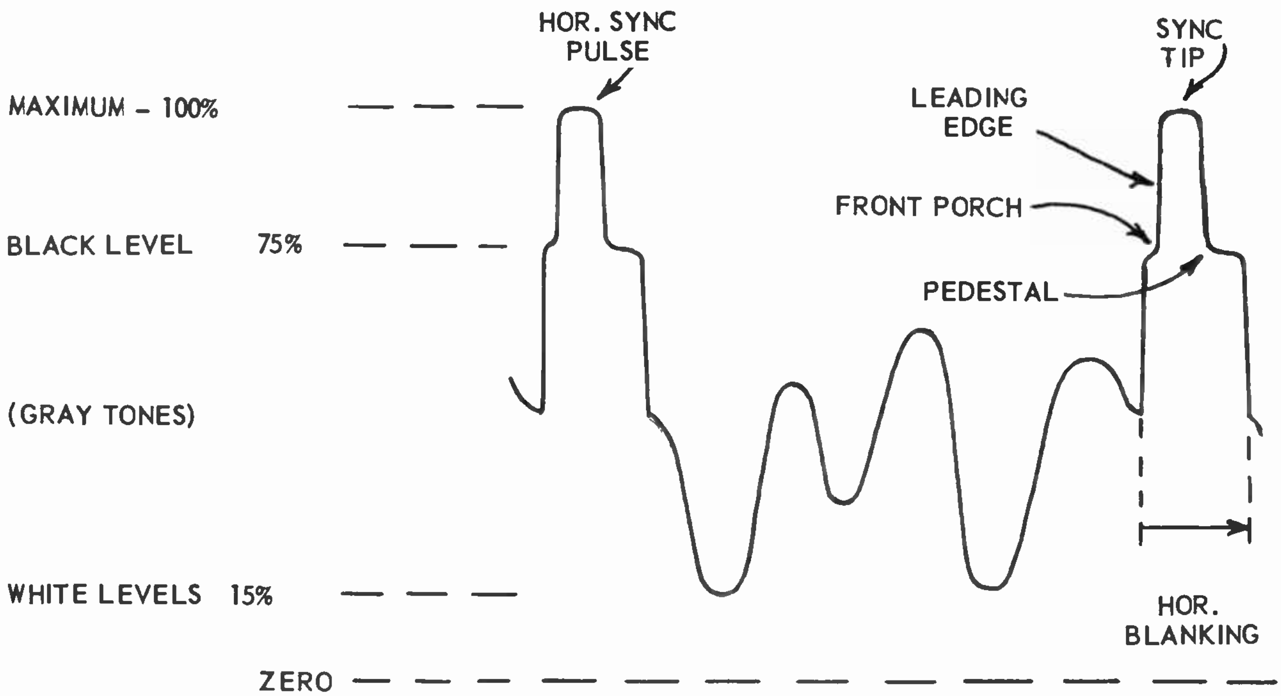


Fig. 42-7. Voltage levels and features of horizontal sync pulses.

No matter what modulation frequency you examine, the total of gains at this frequency below and above the carrier frequency always will be 100%. At the smallest frequency which could be measured below and above the carrier the two gains would be approximately 50% each, and the total would be 100%. For all modulation frequencies higher than 0.75 mc the response is 100%. These are the frequencies transmitted and received in single strength.

The response shown at the bottom of Fig. 42-5 is the one you will strive to attain when aligning the i-f amplifier for video frequencies in television receivers. In this i-f amplifier the video carrier frequency will have been changed to the video i-f frequency by action of the mixer tube. The chief objects of i-f alignment will be to get the video intermediate frequency at or near the point of 50 per cent gain, to have a cutoff before reaching the sound intermediate frequency at one end of the response, and to have a cutoff no more than 1.25 mc from the video intermediate frequency at the other end of the response.

THE TELEVISION SIGNAL. The complete television signal transmitted in any one channel includes amplitude modulation for pictures and synchronizing, also frequency modulation for sound. Other than being transmitted within the same 6-megacycle channel the sound signal is an entirely separate thing and is not related to the signal carrying pictures and synchronization. This latter amplitude-modulated signal may be called the composite television signal or simply the television signal. Still another name is video signal. The Latin word video means to see and is commonly used as a prefix to denote anything related to television parts, circuits or pictures.

The composite television signal forms the modulation envelope of the carrier amplitudes, as shown at the left in Fig. 42-6. It forms also the envelope of the intermediate-frequency signal, after the composite signal has passed through the mixer tube. When the i-f signal is demodulated by the diode detector or video detector of the television receiver we recover the signal voltages themselves, as shown at the right.

Depending on whether the i-f signal is applied to the plate or cathode of the diode detector we may recover either side of the envelope. With the i-f signal applied to the detector plate the composite signal taken from the detector cathode has the sync pulses positive and picture signals negative. With reversed detector connections we have picture signals positive and sync pulses negative. Both methods are used, as we shall see later on.

Note that in the modulated carrier the tips of the sync pulses are at maximum amplitude, while the black level and all picture-forming signals are at lesser amplitudes. The greatest signal strength or greatest amplitude is used for synchronizing, which helps keep the reproduced pictures in time with viewing at the camera. The lightest parts of the pictures are carried by the weakest amplitudes in the transmitted signal. This method is called negative transmission. Were the sync pulses to be at weakest amplitudes, and white parts of pictures at maximum amplitudes, we should have positive transmission.

Fig. 42-7 illustrates the meanings of the various "levels" which designate relative voltages of the composite television signal. The tips of all sync pulses are at maximum or 100% voltage above zero. This voltage might be either positive or negative with reference to zero, depending on whether the sync tips are positive or negative.

The black level is at approximately 75% of maximum voltage. It may vary by no more than 2½% either way from 75%. Note that the black level corresponds to a definite amplitude of modulation on the carrier, it does not vary with changes of tone or of light and shade in the picture. This level corresponds to carrier amplitude produced when the television camera is scanning a black area in the original scene. A black level voltage in the receiver does not necessarily cause black at the picture tube, but it should cause the least brightness of any area in the picture. The black level sometimes is spoken of as the blanking level, because this and any greater voltage cuts off or blanks the electron beam in the picture tube. The voltage range of the sync pulses, from 75% to 100% of maximum, often is called "blacker than black".

Some parts of the signal related to horizontal sync pulses are shown at the right in Fig. 42-7. The sync pulse occurs during the first part of the horizontal blanking interval. First there is an exceedingly brief front porch at the black level, followed by the sudden increase of voltage called the leading edge of the pulse. It is this leading edge, not only of horizontal pulses but of all other sync pulses too, that times the action of the sweep oscillators and thus times the horizontal or vertical sweeps of the electron beam in the picture tube. The part of the voltage waveform on which the horizontal pulse appears to be standing is the pedestal. For this reason, the black level sometimes is called the pedestal voltage.

Fig. 42-8 shows some of the things which must be done to maintain synchronization during the vertical blanking interval between the end of one field and the beginning of the next field. At the left-hand end of the waveform at the top is shown the signal for the last three lines in a field. Then comes the relatively long vertical blanking interval. At a time equal to three line periods after the beginning of vertical blanking comes the vertical sync pulse. This pulse lasts as long as three line periods, since its voltage must continue long enough to charge a rather large capacitance.

Although there are no picture signal voltages between the end of one field and the start of the next one, we continue the horizontal sync pulses right through the entire vertical blanking period. This is necessary in order to keep the horizontal sweep oscillator synchronized or timed with the received signal. If these horizontal pulses were to stop during vertical blanking, the horizontal sweep oscillator would get so far out of time that the appearance of many fields and frames would be ruined before the signal could restore synchronization.

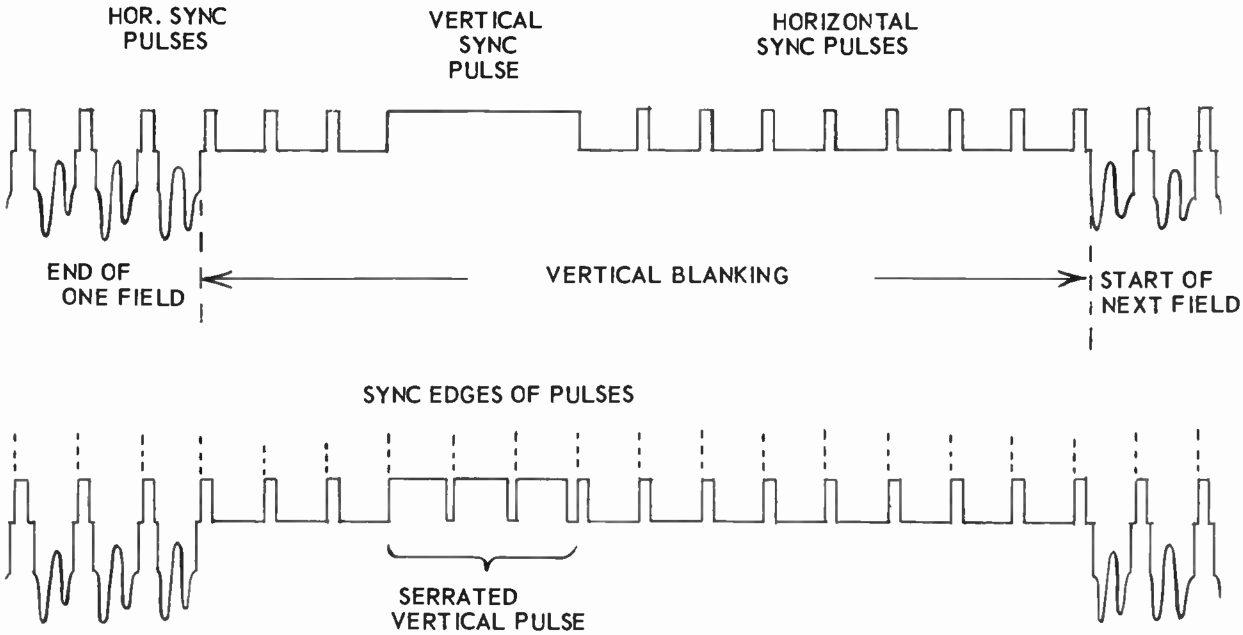


Fig. 42-8. Horizontal synchronization is continued throughout the vertical blanking period.

We wish to maintain horizontal synchronization through the long vertical sync pulse, so we serrate this pulse as shown by the lower waveform. To serrate anything means to put notches in it. In the serrated vertical pulse have been inserted three gaps, after each of which comes a leading edge to operate the horizontal sweep oscillator. The leading edges are identified by short broken lines. You can see how horizontal synchronization continues all through the vertical blanking interval.

Even though the composite signal as so far developed provides for both horizontal and vertical synchronization, it still would not permit starting alternate fields at the beginning of a line and at the center of a line, or it would not provide for interlacing. To meet this requirement we must make the further additions and modifications shown by Fig. 42-9. A series of six equalizing pulses has been inserted between the end or bottom of the first field and the vertical sync pulse, and a similar series of equalizing pulses follows the vertical pulse. We also have increased the number of serrations in the vertical sync pulse. The equalizing pulses occur at half-line intervals, as do also the serrations in the vertical sync pulse.

The upper waveform shows the signal for the last three lines at the bottom of the first of two fields, then the first vertical blanking interval during which there is a vertical retrace, and then the lines at the top of the second field. Note that the first field ends with a full line, and the top of the second field commences with a half line. The lower waveform shows the signal for the bottom of the second field, which ends with a half line, then shows the second vertical blanking interval, and finally the top of the first field or next field, which begins with a full line. The change from a full line to a half line in the upper waveform is made by using a half-line interval between the final equalizing pulse and the following regular horizontal sync pulse. The reverse change, from a half line to a full line in the lower waveform, is made by using a full-line interval between the final equalizing pulse and the following regular horizontal sync pulse.

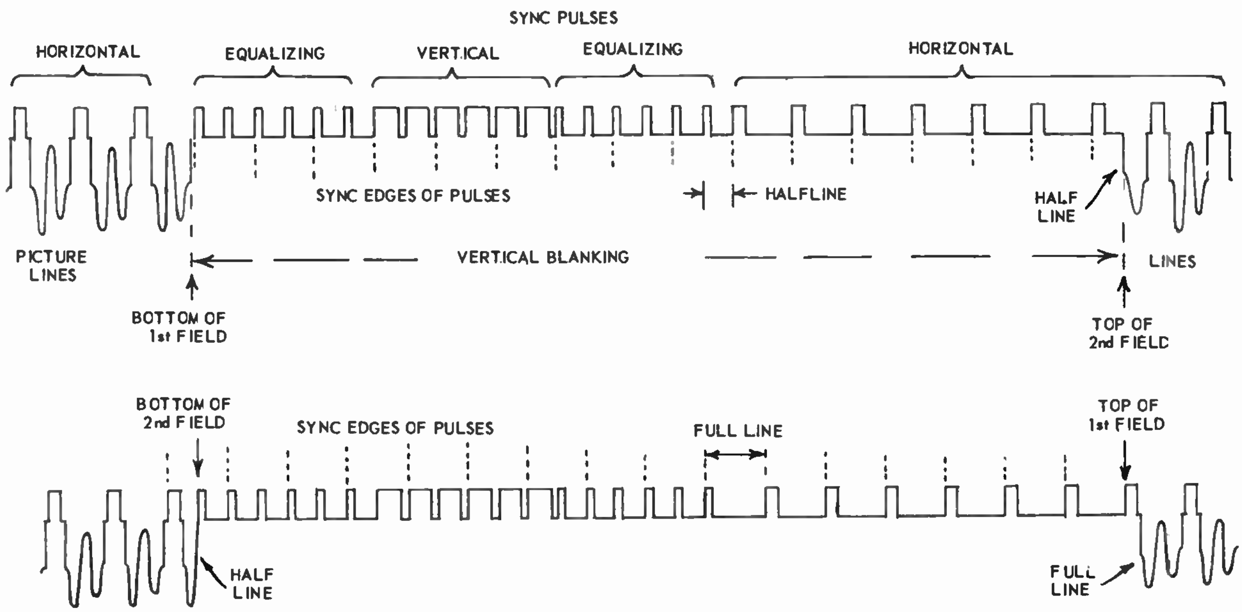


Fig. 42-9. The complete composite signal during the two vertical blanking periods of one frame.

Short broken lines identify the leading edges of all the pulses that maintain uninterrupted synchronization of the horizontal oscillator. These leading edges continue the horizontal sync timing through both vertical blanking intervals in spite of the interlacing. Of the equalizing pulses preceding the vertical sync pulse, only the first, third, and fifth control the horizontal oscillator during the upper waveform. Only the second, fourth, and sixth equalizing pulses control the horizontal oscillator during the lower waveform. The intervening equalizing pulses, also the unused serration edges in each of the vertical sync pulses. have no effect on the horizontal sweep oscillator. This is because at these intervening times the oscillator circuit is so far from conditions suited for oscillation that nothing is accomplished by these other leading edges.

Note that all the sync pulses, including horizontal, vertical, and equalizing, are at signal voltages higher than the black level. All the picture signals are at voltages below the black level. It is this difference between voltages that allows separating picture signals from sync pulses and using the picture signals for control of beam current in the picture tube. It also allows separating the sync pulses so that they may be used for control of the sweep circuits. These separations are accomplished by plate current cutoff and plate current saturation in tubes of the sync section of the receiver.

Note also thatthe duration of a horizontal sync pulse is much less than that of a complete serrated vertical sync pulse. In addition, the horizontal pulses occur at a frequency of 15,750 cycles per second while vertical pulses have a frequency of 60 cycles per second. The difference in frequency and in duration allows separating the two kinds of pulses in suitable filter circuits of the sync section. Then voltages from the horizontal pulses are fed to the horizontal sweep oscillator, while those from the vertical pulses go to the vertical sweep oscillator.

THE TELEVISION RECEIVER. What the receiver does with the composite signal is illustrated by Fig. 42-10. Carrier-frequency signals come from the antenna to the tuner. In the tuner are one or more r-f amplifiers, an r-f oscillator, and a mixer, all of which act in the same general way as similar tubes in any other superheterodyne tuner. There is some amplification. Carrier frequencies are changed to intermediate frequencies by action of the oscillator and mixer, and the intermediate frequencies retain the original f-m modulation for sound and the a-m modulation for pictures and synchronization. The tuned circuits are made resonant for frequencies to be received by means of the channel selector knob or dial attached to the tuner. Many receivers have an additional tuning control for the r-f oscillator circuit. This usually is called a fine tuning control. It allows the operator to compensate for slight variations of oscillator frequency, as may occur while the set is in operation.

The f-m sound signal may be taken off by a suitable coupler which immediately follows the tuner, or this signal may be given further amplification in i-f stages and then taken off at any of various points farther along in the signal path. The parts of the sound section are just like the i-f amplifiers and the demodulators we studied in connection with f-m receivers, except for operating at different frequencies.

Modulated intermediate frequencies go from the tuner to the i-f amplifier, in which may be three, four, or more i-f amplifier pentodes and necessary couplers. If the sound signal is taken off immediately after the tuner, the i-f amplifier stages are designed to handle only the video intermediate frequency. If the sound

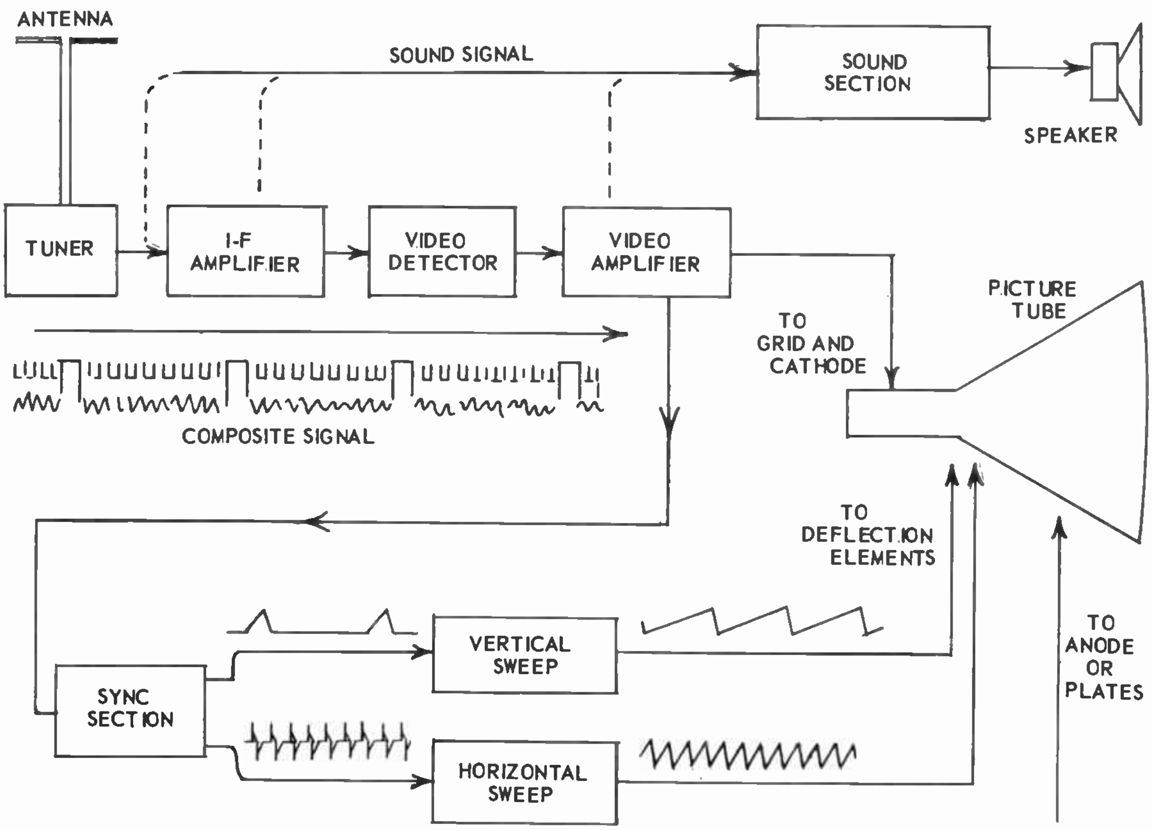


Fig. 42-10. Travels of the composite signal and its parts through the television receiver.

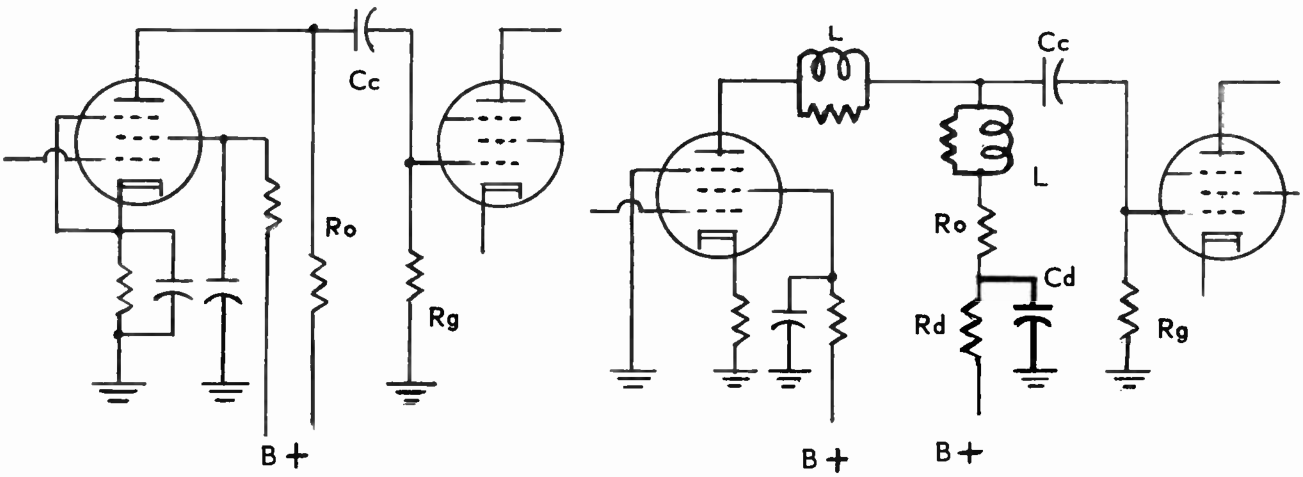


Fig. 42-11. Resistance-capacitance coupling as used in audio amplifiers (left) and in video amplifiers (right).

signal passes through some i-f stages before being taken off, these stages are designed to handle both sound i-f and video i-f signals while following stages are designed for only the video i-f signals. If the sound signal is taken off after the video detector all the i-f stages are designed for both video and sound coverage. The video detector most often is one section of a twin-diode tube, although in a few receivers this detector is a crystal diode.

In the video amplifier section are one, two, or possibly three pentode amplifier tubes. This amplifier must handle an extremely wide frequency range, all the way from the frame frequency of 30 cycles per second up to the highest picture element frequency of 4 megacycles per second. This requires that the video amplifier be of the broad band type, with special design and construction to allow reasonably uniform gain at all the frequencies which must be handled.

The basic resistance-capacitance coupled circuit for video amplifiers is shown at the left in Fig. 42-11. This is the same circuit used for audio-amplifier stages and other untuned couplings. The plate circuit load is resistor *Ro*. Coupling is through capacitor *Cc*. The grid return is through resistor *Rg*.

Parts added for broad band response are shown in the right-hand diagram. Between the first plate and the coupling capacitor is a peaking coil *L*. Another peaking coil is in the B+ lead for the plate circuit. These "peakers" help maintain uniform response through the higher video frequencies. Below the regular plate load resistor *Ro* are resistor *Rd* and capacitor *Cd*, which are selected of such values as to maintain low-frequency response in connection with the operation of *Cc* and *Rg*.

The entire composite signal has been brought from antenna through the video amplifier and now is applied to the grid-cathode circuit of the picture tube. This is the circuit that controls the rate of electron emission from the cathode and regulates the rate of electron flow in the picture tube beam.

5 Either side of the modulation on the i-f signal may be recovered by the video detector, as shown at the left in Fig. 42-12. But the signal at the grid of the picture tube must be of such polarity that the picture variations are positive and the sync pulses negative. This is accomplished by connecting the detector to the i-f amplfiers in such a way as gives a detector output which, combined with whatever inversions occur

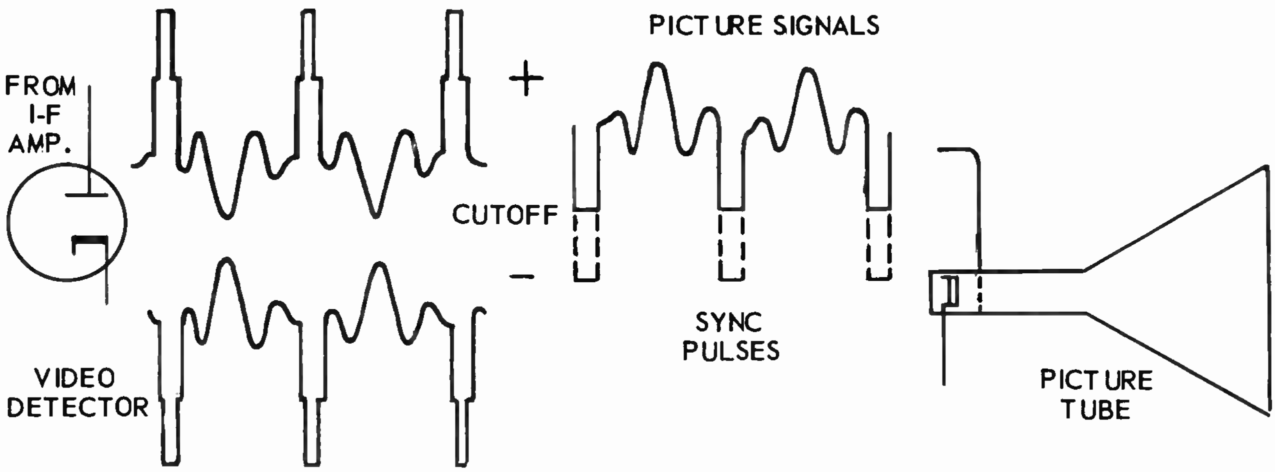


Fig. 42-12. Pictnre signal variations must be positive at the grid of the picture tube.

in the video amplifier, makes the signal of correct polarity at the picture tube. The picture tube grid is biased with reference to its cathode at a voltage causing beam current cntoff at the black level of the signal. Thus the negative sync pulses are cut off while the positive picture signals control the electron beam in the picture tube.

Looking back at Fig. 42-10, yon will see that the composite signal from the video amplifier is fed to the sync section as well as to the picture tube. Many operations may be performed on the signal between the instants of entering the sync section and passing on to the sweep sections. Some of these operations are illustrated by Fig. 42-13.

If, as in diagram *A*, the signal is of such polarity that sync pulses are positive and picture variations negative it is possible to separate and retain the pulses while cutting off the picture portion of the signal. This may be done in a sync separator tube having grid bias sufficiently negative for plate current cutoff at the black level voltage of the applied signal.

If the signal is too weak it may be amplified, as at *B*. Amplification may be applied to the sync pulses after they have been separated, or it might be applied to the incoming composite signal, or at any other point in the sync section where the signal needs strengthening.

Should the signal be of an unwanted polarity at any point it can be inverted, as in diagram *C*. Inversion of polarity requires nothing more than passage of the signal through a tube from control grid circuit to plate circuit. Inversion may be applied to the entire signal, as illustrated, or to only the sync pulses after they have been separated.

Sometimes the sync pulses become of non-uniform strength or voltage, as at *D*. This may happen when certain kinds of electrical interference enter the receiver and get all the way through to the sync section. Then the unequal pulses are clipped or limited to bring them to uniformity. This can be done by plate current cutoff, plate current saturation, or both, much as in the limiter stage of an f-m receiver.

Finally, as at *E* in Fig. 42-13, we have horizontal, vertical, and equalizing pulses all suitably separated, amplified, clipped, and limited if necessary. All the pulses are fed into two electrical filters. One of these filters gets rid of the horizontal and equalizing pulses, and from the serrated vertical pulse produces a peak of voltage. These peaks occur at the field frequency of 60 cycles per second. The other filter is unaffected by the long portions in the serrated vertical pulse, but from the horizontal pulses, equalizing pulses, and leading edges of the vertical serrations it produces sharp "pips" of voltage at the line frequency of 15,750 cycles per second. The 60-cycle voltage peaks go to the vertical sweep section. The voltage pips at 15,750 cycles go to the horizontal sweep section.

Each of the sweep sections contains a sweep oscillator. These oscillators operate on principles quite different from those employed in the r-f oscillator of the tuner. Whereas the r-f oscillator produces a steady output voltage of sine-wave form, the sweep oscillators produce intermittent voltages of sawtooth waveform. The sawtooth voltage results from a gradually increasing charge in a capacitor and the very sudden discharge of this capacitor. The sawtooth capacitor charges from the B-power supply. Discharge is controlled by the sweep oscillator. Every time a voltage from the sync section acts on the grid of the sweep oscillator, the oscillator acts to discharge the sawtooth capacitor. Then from the vertical sweep oscillator we have sawtooth waves at the vertical deflection frequency of 60 times or cycles per second, and from the horizontal sweep oscillator we have sawtooth waves at the horizontal deflection frequency of 15,750 cycles per second.

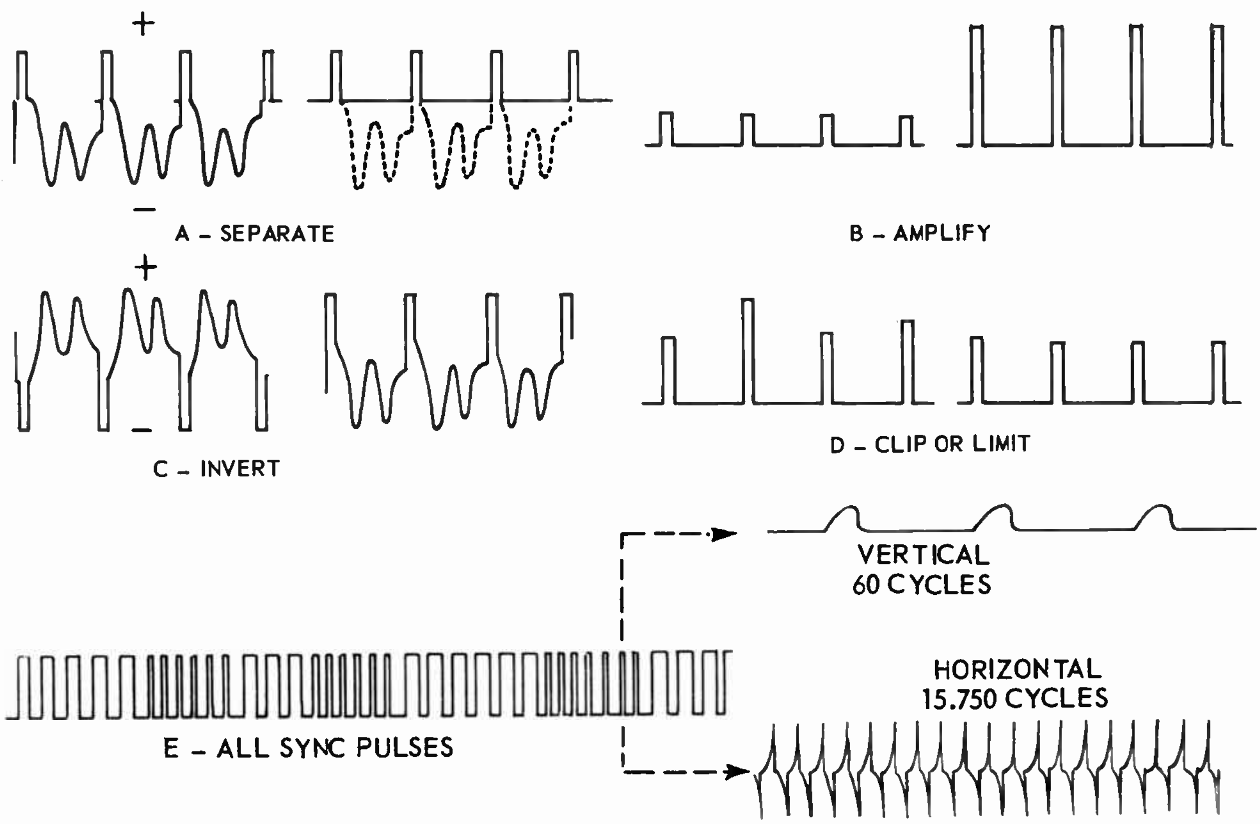


Fig. 42-13. What the sync section does with the composite television signal.

Fig. 42-14. Some of the parts in a high-voltage power supply.

In the sweep sections of most receivers the oscillators are followed by sweep amplifiers which increase the amplitude of the sawtooth voltages. When the picture tube is of the magnetic deflection type the sweep section includes transformers and coils which utilize the sawtooth voltages to control sawtooth currents required for this method of beam deflection. In magnetic deflection systems we find a damper tube whose purpose is to limit oscillation which will occur in the transformer and coil circuit at a frequency determined by its inductance and distributed capacitance.

In the horizontal sweep section for magnetic deflection picture tubes there is an automatic control for oscillator frequency which keeps this frequency precisely in time with the horizontal sync pulses even when interference and other electrical disturbances tend to upset the synchronization. Similar automatic controls sometimes are used in vertical magnetic deflection systems, and sometimes with electrostatic deflection.

The oscillator automatic frequency control compares the actual frequency of the sawtooth output with the frequency of the horizontal sync pulses from the signal. Any variation is made to produce a voltage which corrects the timing of the oscillator, to bring it back into synchronization with the pulses. Some of these correction systems utilize discriminators. Others use phase detectors, which are related in their action to ratio detectors and discriminators. Also, in these automatic frequency control systems, you will find many principles used nowhere else in radio or television.

The outputs of the vertical and horizontal sweep sections go to the deflection plates of an electrostatic picture tube or to the deflection coils of a magnetic picture tube. To the picture tube elements, or anodes, which accelerate the electrons in the beam, are applied voltages which range from 3,500 to 13,000, depending on the type and size of picture tube. These voltages come from the high-voltage power supply of the receiver.

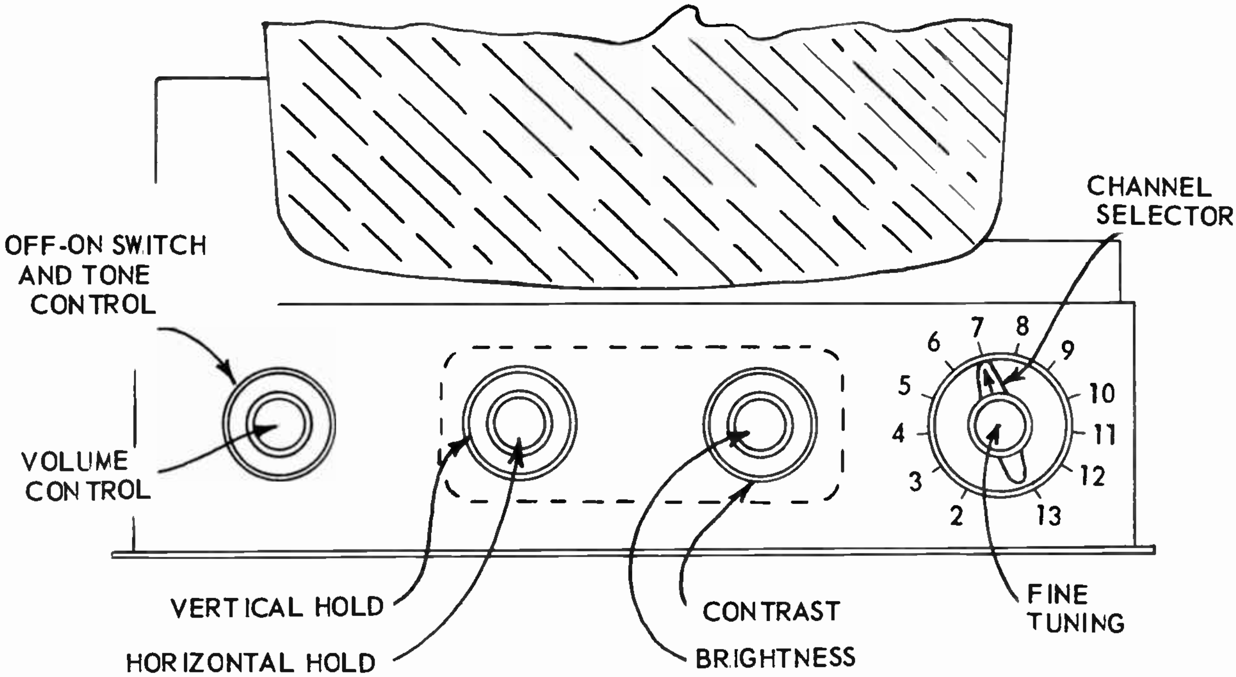


Fig. 42-15. Typical operating controls for a television receiver.

Fig. 42-14 is a picture of parts in a "flyback" 9,000-volt power supply used with magnetic deflection tubes. This unit is fed with about 400 volts from the low-voltage B-supply, it uses a very high emf induced during the horizontal retrace period, then steps this emf up in an auto-transformer whose high-potential a-c voltage is rectified and filtered. Other high-voltage supplies use an oscillator to produce radio-frequency voltages that are stepped up in an air-core transformer, then rectified and filtered.

TELEVISION CONTROLS AND ADJUSTMENTS. Controls which usually are made accessible to the operator of a television receiver are shown by Fig. 42-15. Shafts for the knobs protrude through the front of the chassis. Potentiometers in the controlled circuits are twin types, with concentric shafts and large and small concentric knobs for two controls in each position. At the left is a combined off-on switch, tone control, and volume control. The tone control is not used in all receivers. Some of the operator's controls may be concealed with a small panel, as indicated by the broken line around the four controls at the center of Fig. 42-15. Purposes of the controls are as follows.

HOLD CONTROLS. The vertical hold control adjusts the frequency of the vertical sweep oscillator to permit its synchronizing with the vertical pulses. This control holds the picture stationary in a vertical direction, preventing continual or intermittent movement upward or downward. The horizontal hold control adjusts the frequency of the horizontal sweep oscillator so that horizontal pulses can maintain synchronization. This control prevents the picture moving sideways in either direction, either slowly or rapidly.

Contrast Control. This control varies the amplification applied to the signal. Correct operation allows a full range of shadings from black to white. When adjustments of contrast and brightness controls are suited to each other, objects which should appear black will actually be black rather than dark gray, and objects which should be white will be so, rather than light gray.

(1) *Brightness Control.* This control adjusts the grid biasing of the picture tube for cutoff at the black level. Correct adjustment prevents sync pulses from affecting the picture, while not cutting off any of the picture signal variations. Brightness adjustment affects the all-over tone of the picture, making it either too light, too dark, or in correct proportion to the scene being viewed.

Channel Selector. A knob, dial, or series of push buttons tunes the r-f amplifier, r-f oscillator, and mixer circuits for reception of any desired channel.

Fine Tuning. This control will make slight variation of r-f oscillator frequency as may be required for best reproduction of pictures, sound, or both.

Service adjustments which are not accessible to the operator usually are on the rear of the chassis, as in Fig. 42-16, or they may be on top or underneath the chassis. In addition to the service adjustments shown here there are numerous alignment adjustments for the tuner, i-f amplifier, and sound sections. These alignment adjustments are reached from the top, bottom, or front of the chassis. The controls of Fig. 42-16 are described as follows.

Centering. Vertical centering brings the center of the picture to the center of the opening for the picture tube measured in a vertical direction. This control moves the picture up or down. Horizontal centering serves a similar purpose in moving the picture to the right or left. Centering for magnetic-deflection tubes often is accomplished by adjustments of parts which are on the neck of the tube rather than by circuit adjustments located on the chassis.

Focus. The focus adjustment permits forming the smallest possible spot of light at the point where the electron beam strikes the picture tube screen. The picture then has good definition or good reproduction of small details.

Linearity. This word refers to shapes and relative sizes of lines and objects in the reproduced picture these lines and objects are related to the original scene or its image in the camera. When the reproduced picture is linear it is not distorted in form, shape, or proportions. A non-linear picture is distorted in one or more ways. Horizontal linearity adjustments are intended for correction of sidewise distortion, such as

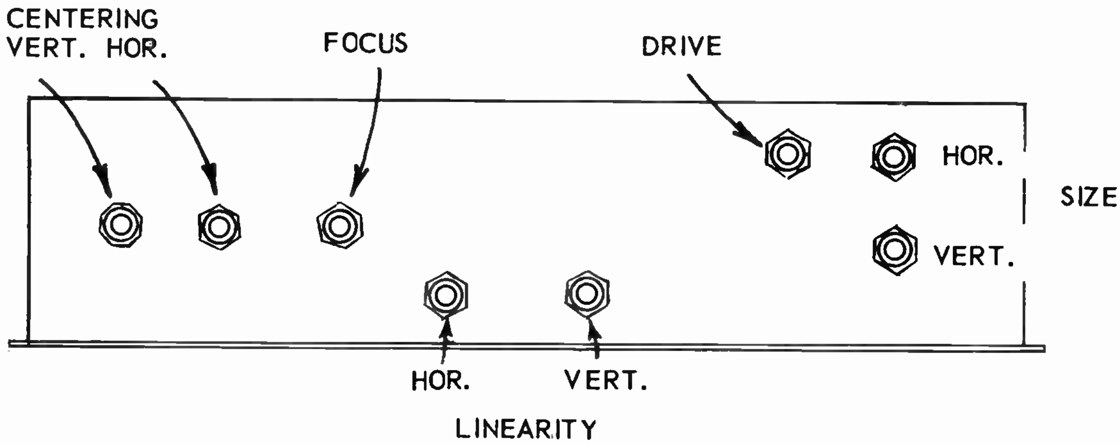


Fig. 42-16. Service adjustments, other than those for alignment, which may be on the chassis.

stretching or compressing in a horizontal direction. Vertical linearity adjustments correct for distortion in an up and down direction. Linearity adjustments most often act to improve or correct the form of sawtooth voltages and currents in the sweep section.

Drive. A drive control acts on the amplitude of sawtooth voltages or currents. This affects both linearity and size of reproduced pictures.

Size. The horizontal size control makes the picture either narrower or wider, while the vertical size control makes the height either greater or less. These two adjustments may be called the width control and the height control. Size, centering, and linearity controls are used together to make the picture just fill the opening of the mask placed in front of the picture tube.

Now that we know something about the relations of television receiver parts to one another, and what these parts are supposed to do with the television signal, we shall commence at the tuner and make detailed examination of everything that happens until we reach the picture tube.

DO NOT TEAR — CUT ALONG THIS LINE AND SEND IN FOR GRADING.

TELEVISION

LESSON NO. 43

TELEVISION TUNERS

"Training Your Mind"

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind
C. T. I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON 43
TELEVISION TUNERS

The functions of the television tuner are the same as those of the tuner in any other superheterodyne receiver; it selects the carrier having the desired program and changes the carrier frequencies into intermediate frequencies which retain the signal modulation. In the television tuner there always is an r-f oscillator and a mixer tube. Also, in all except a few early designs, there is at least one r-f amplifier and often there are two.

The wide range of carrier frequencies to be covered in television introduces some problems not present in a-m and f-m sound broadcast reception. At present there are five channels, numbered 2 through 6, in the range of television carrier frequencies called the low band. There are seven more channels, 7 through 13, in the range called the high band. Channel limits, also frequencies of the video and sound carriers, are listed in the accompanying table.

TELEVISION CHANNEL FREQUENCIES

Channel Number	Frequency Limits, mc.	Video Carrier, mc.	Sound Carrier, mc.
2	54 to 60	55.25	59.75
3	60 to 66	61.25	65.75
4	66 to 72	67.25	71.75
5	76 to 82	77.25	81.75
6	82 to 88	83.25	87.75
7	174 to 180	175.25	179.75
8	180 to 186	181.25	185.75
9	186 to 192	187.25	191.75
10	192 to 198	193.25	197.75
11	198 to 204	199.25	203.75
12	204 to 210	205.25	209.75
13	210 to 216	211.25	215.75

In the wide frequency gap between the top of the low-band and the bottom of the high-band television channels there are f-m broadcasting and many other radio services. If we were to tune uninterruptedly over the entire range of frequencies from channel 2 through channel 13, using "straight line" variations of capacitance or inductance, the tuning dial would be divided as at the left in Fig. 43-1. With straight-line frequency tuning we should have a dial as at the right.

Tuners of recent design employ either of two methods to avoid tuning all through the frequency gap between low and high band television channels. With one method there are tuned circuits which may be made

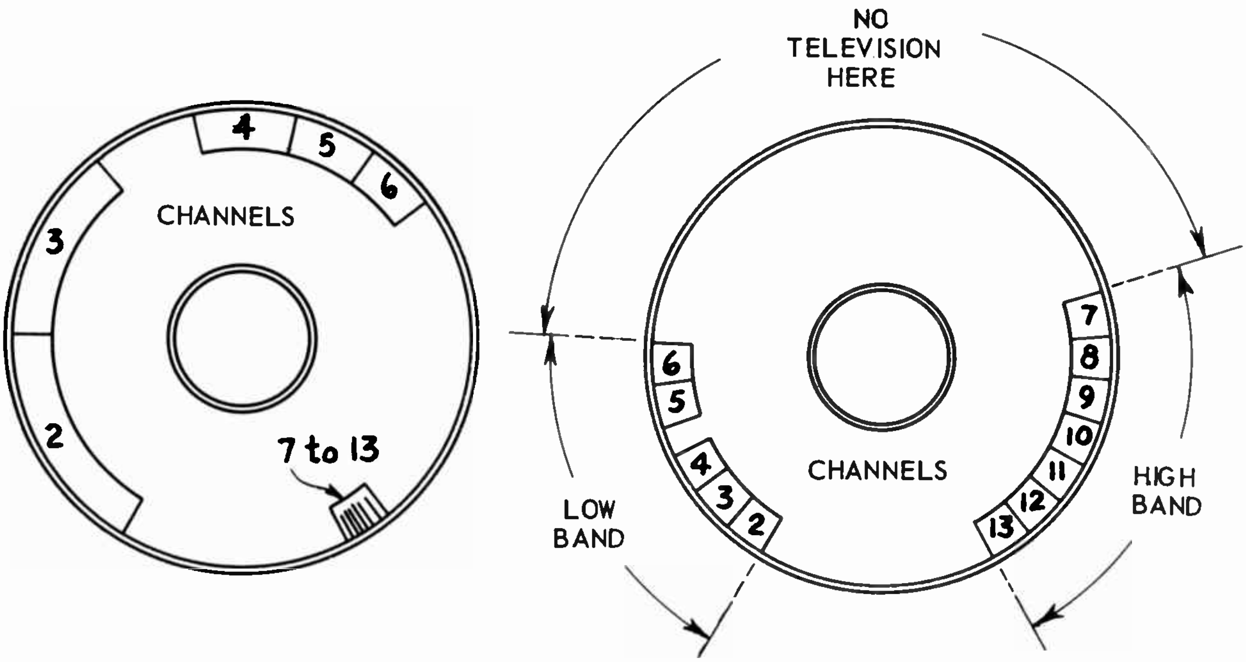


Fig. 43-1. Distribution of channel frequencies on dials.

resonant only to the particular frequencies of each channel, and not to any intermediate frequencies. This is done by means of switches having positions for each channel. With the other method there is continuous tuning throughout the low band, then a switching to circuits for the high band, followed by continuous tuning through the high band.

Any frequencies between 30 and 300 megacycles are classed as "very-high" frequencies, a term which often is abbreviated VHF, v-h-f, or vhf. Television channels 2 through 13 are in this very-high frequency range. Any frequencies from 300 through 3,000 megacycles are classed as "ultra-high" frequencies, abbreviated UHF, u-h-f, or uhf. Television channels 14 through 83 cover frequencies between 470 and 890 megacycles, which is the "ultra-high" frequency range. A television station operating in this frequency range is referred to as a UHF station. A thorough discussion on UHF fundamentals, antennas, and tuners will be given later in the course.

Although tuners for a-m and f-m sound broadcast receivers nearly always are an integral part of the chassis construction, the majority of television tuners are separate devices which may be removed as self-contained units from the chassis after disconnecting the wiring leads.

Literally there are dozens of different types of television tuners in use. Before proceeding to a detailed examination of some of the more popular designs it will be helpful to look at the relatively few fundamental electrical features and mechanical constructions which are put together in various combinations to form all the different types.

Most tuners have a single r-f amplifier tube, a mixer, and an r-f oscillator, as at the left in Fig. 43-2. The oscillator circuit *A* always is variably tuned in one way or another for reception in each channel. As a general rule there is variable tuning also for both the mixer grid circuit *B* and for the r-f amplifier plate circuit *C*, although in a few cases only one or the other of these circuits is variably tuned, with the other circuit coupled to it but untuned. The r-f amplifier grid circuit or antenna coupling circuit *D* may or may not be

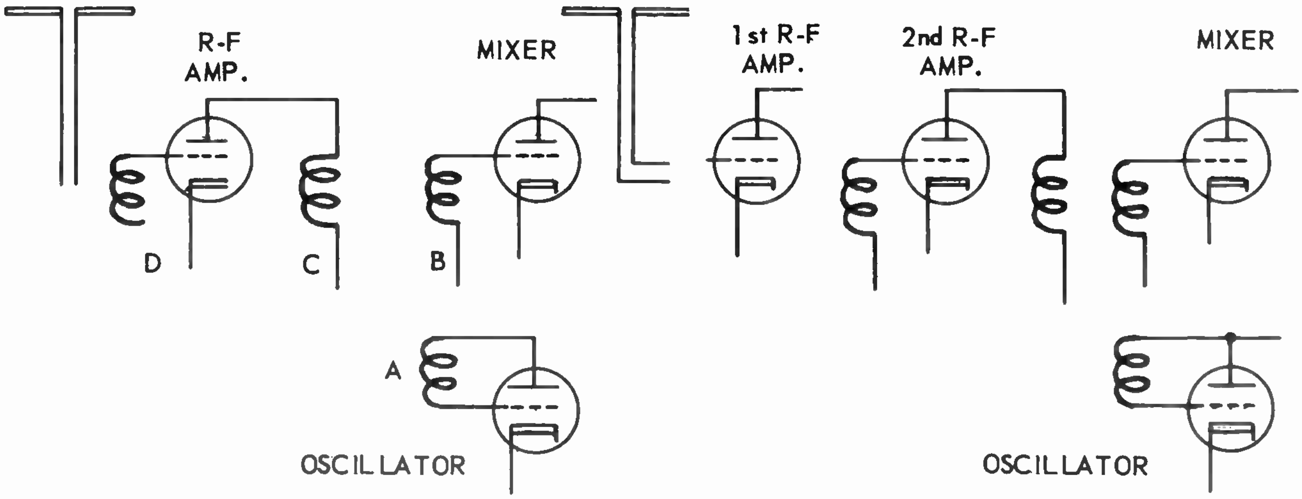


Fig. 43-2. Circuits which are tuned between antenna and mixer.

variably tuned; there are just about as many receivers built one way as the other. There may be three separate tubes for r-f amplifier, mixer, and r-f oscillator, or else there may be one r-f amplifier tube and one additional twin tube for mixer and oscillator.

Some television receivers have two r-f amplifier stages. Then, as at the right in Fig. 43-2, it is common practice to have an untuned coupling between antenna and grid circuit of the first r-f amplifier. This is followed by a single variably tuned impedance coupler between the two r-f amplifiers, then a variably tuned circuit for the plate of the second r-f amplifier and one for the grid of the mixer. As always, there is a variably tuned r-f oscillator circuit. The two r-f amplifiers are separate tubes, with usually a twin tube for combined mixer and oscillator.

TUNER DESIGN AND CONSTRUCTION

VARIABLE FACTOR	OPERATION BY MEANS OF	TYPE OF TUNING ELEMENTS	SELECTION IN
	Rotary selector switch	Continuous inductors Separate inductors	Each channel
Inductance	Rotary turret	Separate inductors	Each channel
	Sliding contacts	Continuous inductors	Two bands
	Movable cores	Inductors for each band	Two bands
	Movable rotors	Variable capacitors	Two bands
Capacitance	Rotary selector switch	Separate trimmer capacitors	Each channel
	Push-button switches	Separate trimmer capacitors	Each channel

A classification of designs and constructions commonly used in television tuners is shown by the accompanying table. The first column shows that either the inductance or the capacitance of the tuned cir-

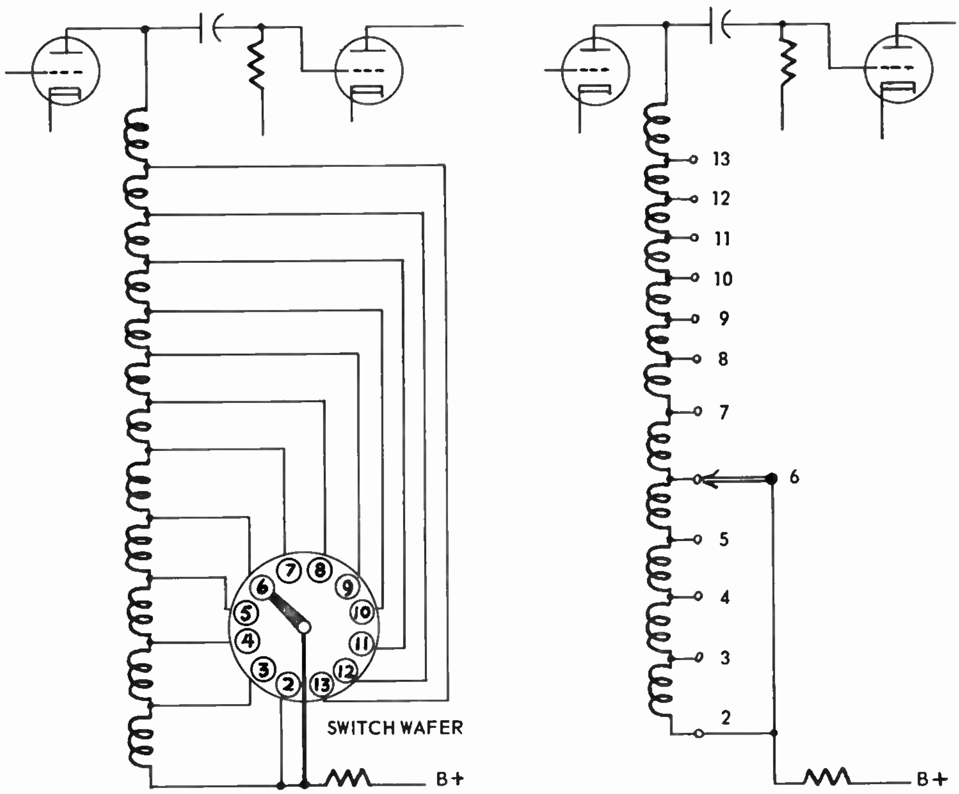


Fig. 43-3. Rotary selector switch for short circuiting portions of an inductor through tap connections.

cuits may be variable. The second column shows that inductance may be varied by either a rotary selector switch, rotary turret, sliding contacts, or movable cores in inductance coils. Capacitance may be varied either by movable capacitor rotors as in usual types of tuning capacitors or by making connections to any of a number of pre-adjusted trimmer type capacitors through a rotary selector switch or push-button switches. The third column lists inductor and capacitor designs. The fourth column lists methods of channel selection.

The design listed across the upper line of the table includes a rotary selector switch and continuous inductance. To illustrate the principle of this method of tuning we shall look at the single impedance coupling between an r-f amplifier and a mixer as at the left in Fig. 43-3. Tuning inductance is here provided by a series of coils or inductors which are electrically continuous. Various portions of this total inductance are short circuited by means of taps connected to terminals on one of the wafers of a rotary selector switch. There is one switch position for each channel. The unshorted portion of the inductance remains active in the tuned circuit. Capacitance for tuning is provided by distributed capacitance in the inductors, internal capacitances of the tubes, and stray capacitance in wiring and switch parts.

Shorted inductor construction often is shown on service diagrams as at the right in Fig. 43-3. The switch rotor usually is represented by an arrowhead which very apparently is intended to move along the tap connections. Both diagrams in the figure represent exactly the same construction.

The switch is shown in position for reception of channel 6. All inductance below tap number 6 is short circuited through the switch rotor and the connection for channel 2 and B-plus at the bottom. All inductance above the tap for channel 6 is actively in the plate circuit and, with capacitance of coils, tube, switch, and wiring, this inductance provides tuned impedance in the plate circuit at a resonant frequency suited for reception of channel 6. When the switch rotor is moved to its position for channel 13 only the top section of the inductance remains active. With the switch in position for channel 2 all the inductance is actively in the plate circuit, with none shorted out.

Tuning capacitance is nearly the same for all channels, since the greater part of it is found in the tube, the switch, and the wiring. The least active inductance is needed when tuning to the highest-frequency channel, number 13, More and more inductance must be brought into the circuit for tuning to lower and lower frequencies of channels having lower numbers, down to channel 2.

If we assume an unvarying total tuning capacitance of 10 micro-microfarads it will require inductance of about 56 thousandths of a microhenry to produce resonance at the center frequency of channel 13. This is a frequency of 213 mc. To change to the center frequency of channel 12, at 207 mc, it is necessary to add only about 3.3 thousandths of a microhenry inductance. To change the tuning all the way through the high-band channels and arrive at 177 mc, the center frequency of channel 7, we need add a total of only 25 thousandths of a microhenry to the amount needed for tuning channel 13. This means that 81 thousandths of a microhenry brings us to channel 7. These changes of inductance are shown to scale at the top of Fig. 43-4, starting from the left.

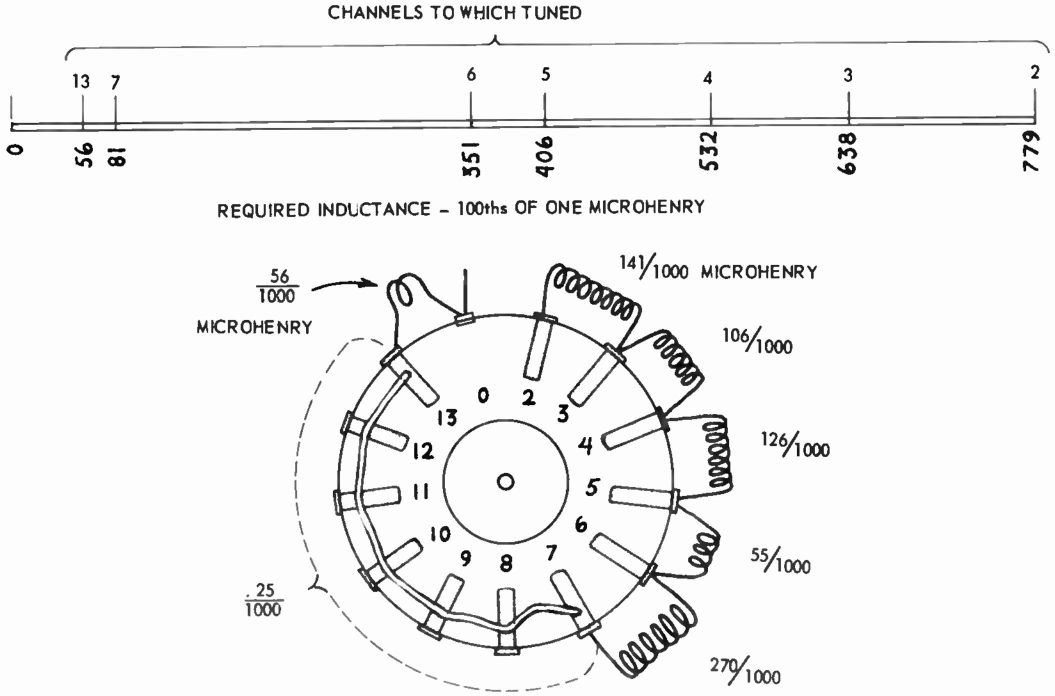


Fig. 43.4. Inductors for tuning high-band and low-band channels, as mounted on a wafer in a selector switch.

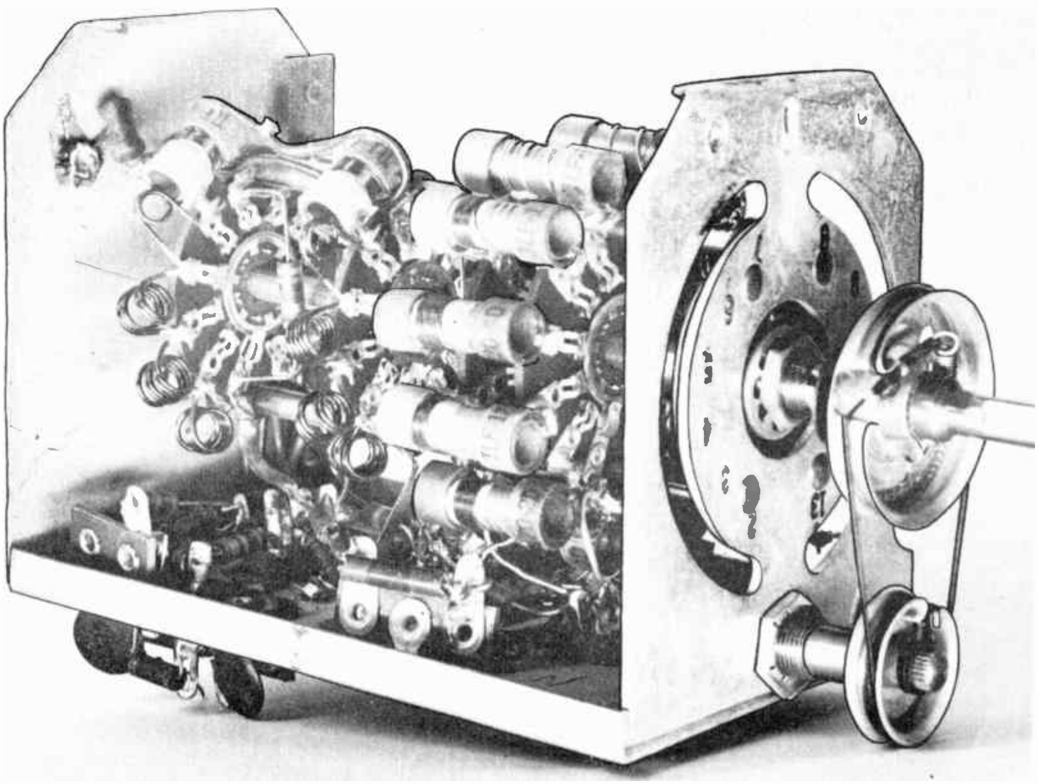


Fig. 43-5. Rotary selector switch with connections to taps on inductors.

There is a big frequency gap between channels 7 and 6. To bring our tuning down to the center of channel 6, at 85 mc, we must add about 270 thousandths of a microhenry inductance, making the total 351 thousandths. To reach the center of channel 5 takes another 55 thousandths of a microhenry. Now, if you look back at the table of television channel frequencies, you will see that channels 5 and 4 are not adjacent. There is a 4-mc gap between the bottom of channel 5 and the top of channel 4. Because of this gap, and because frequency is becoming steadily lower, it is necessary to add 126 thousandths of a microhenry in order to tune for channel 4, making a total of 532 thousandths. Then, to reach channel 3, the total inductance must be made 638 thousandths, and to reach channel 2 it must be 779 thousandths of a microhenry.

In tuners built with rotary selector switches it is common practice to mount the tuninginductance around the edges of the stationary switch wafers. Such an arrangement is illustrated at the bottom of Fig. 43-4. At the upper left is a small coil of two or three turns providing the inductance for tuning channel 13. The additional very small inductances needed for tuning through remaining high-band channels are furnished by short lengths of straight or slightly bent wire connected between successive contacts of the switch. For the frequency jump between channels 7 and 6, or from high band to low band, there is a rather larger coil of possibly 10 to 12 turns. Then, in order, come other coils having diameters and numbers of turns which provide the required additional inductances for tuning the low-band channels.

Fig. 43-5 is a picture of a tuner wherein a rotary selector switch makes tap connections to indnctances for the r-f amplifier plate circuit, the mixer grid circuit, and the r-f oscillator circuit. Just ahead of the rear plate of the frame or shield you can see a number of the small self-supporting coils which are in the r-f amplifier plate circuit for low-band tuning. Near the middle of the tuner are similar coils for low-band tuning of the mixer grid circuit. Toward the front are oscillator coils wound on tubular forms.

Fig. 43-6 illustrates another type of rotary selector switch which short circuits portions of the tuning inductance. The inductance is shown connected to the grid of a mixer tube, but it might be used in any other tuned circuit. On the front and back of the switch wafer are rotors having shorting segments that engage a number of contacts at the same time, rather than having a narrow tongue that engages only one stationary contact at a time.

Both rotors turn together in the direction indicated by the arrows. Both rotors always remain connected to ground through the long stationary contacts whose inner ends rest on the inner parts of the rotors. Thus all the contacts engaged by the rotor segments at any one time are shorted together and connected to ground. Were this switch to be used in a plate circuit the connection here made to ground would go to B-plus.

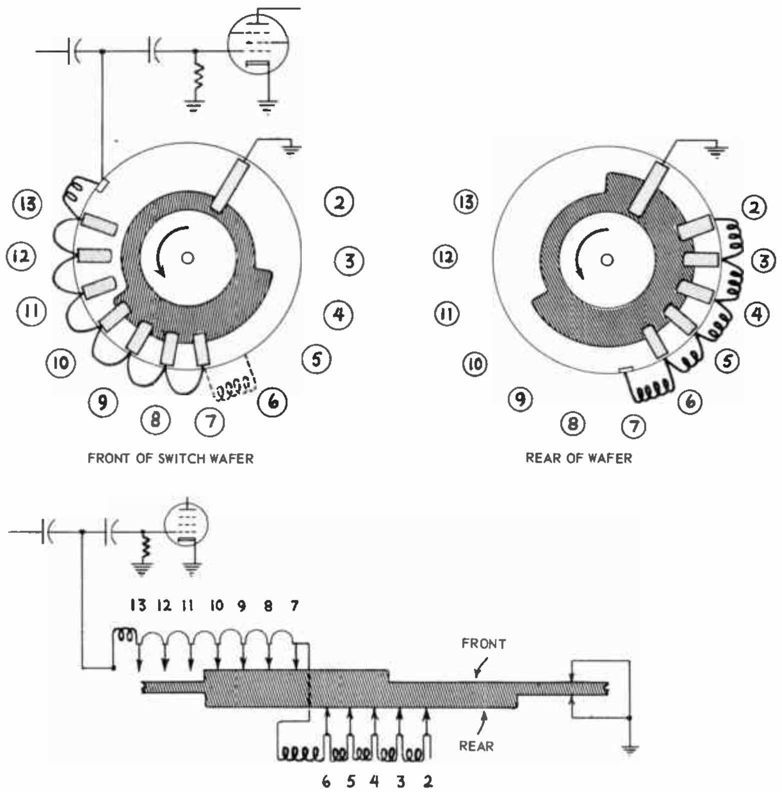


Fig. 43-6. Short circuiting of portions of an inductor by means of an extended rotor on a switch wafer.

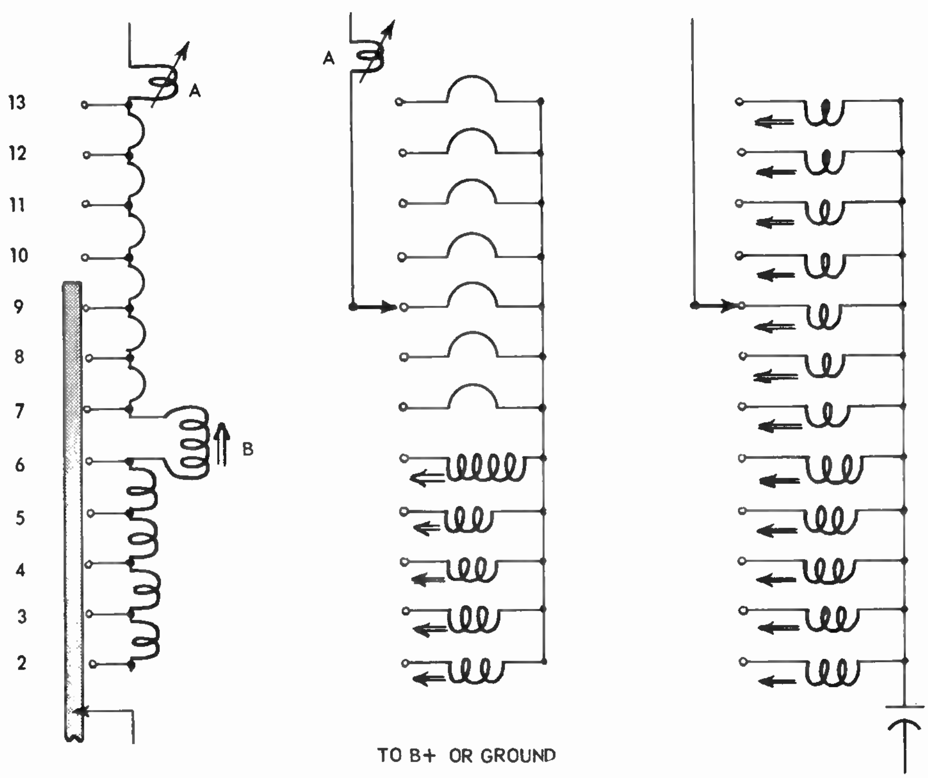


Fig. 43-7. Alignment adjustments for bands and for individual channels as used on selector switches.

Switches of this general type may be shown on service diagrams as at the bottom of Fig. 43-6. Since both rotors are connected together they are indicated by a single long bar. The rotors are shown in position for reception of channel 10. The inductance actively in the grid circuit includes the portion between the grid and the switch contact numbered 13, also the portions between this contact and contact 12, between contacts 12 and 11, and between contacts 11 and 10. The front segment is engaging contact number 10 and all other lower numbered contacts on this side of the wafer, shorting them together and connecting them to ground. All contacts on the rear of the wafer are shorted together and connected to ground.

As the switch is rotated in the direction of the arrows, the shorting and grounding segments will leave more and more portions of the inductance active in the grid circuit until, for reception of channel 2, the entire inductance will be active. The length of the shorting segment on the front of the switch must be such that it will not re-engage contact number 13 until after the switch has been turned through all the lower channel numbers. The shorting segment on the rear of the switch must be long enough to short and ground all the low-band contacts while all the high-band contacts are being uncovered or disengaged. These requirements mean that switches of this particular shorting type must have the contacts on the wafers close enough together to leave part of the circle without contacts. This open part is needed for free movement of the extended rotor segments after they leave the contacts at one end of their travel and before they engage contacts at the other end.

The number of channels for which alignment adjustments are provided varies with different tuners. In many cases there are only two such adjustments in a plate circuit or grid circuit. This is shown in principle at the left in Fig. 43-7. For alignment of all high-band channels reliance is placed on the adjustable inductor at A. Adjustment at this point in the circuit has its chief effect on channels from 13 through 7.

When the shorting contact of the switch is moved down as far as the connection for channel 6 the second adjustable inductor at *B* is brought into the tuned circuit. This inductor has no effect on tuning of the high band, for it is not then in the active circuit. But it does have a large effect on tuning of all channels from 6 through 2. Because values of iuductance for the low-band channels are so very much greater than the total inductance used for the high band, any adjustment which has been made at *A* has but little effect on resonant frequencies for the low band, and what little effect this upper adjustment does have is easily compensated for by a slight change at *B*.

Other tuners employ the principle illustrated by the middle diagram of Fig. 43-7. Here there are separate inductors for each channel. Those for the high-band channels are not individually adjustable, but there is an adjustable small inductance at *A* which serves for overall alignment of channels 13 through 7. In each of the coils for low-band channels 6 through 2 there is a movable core for individual alignment of its channel.

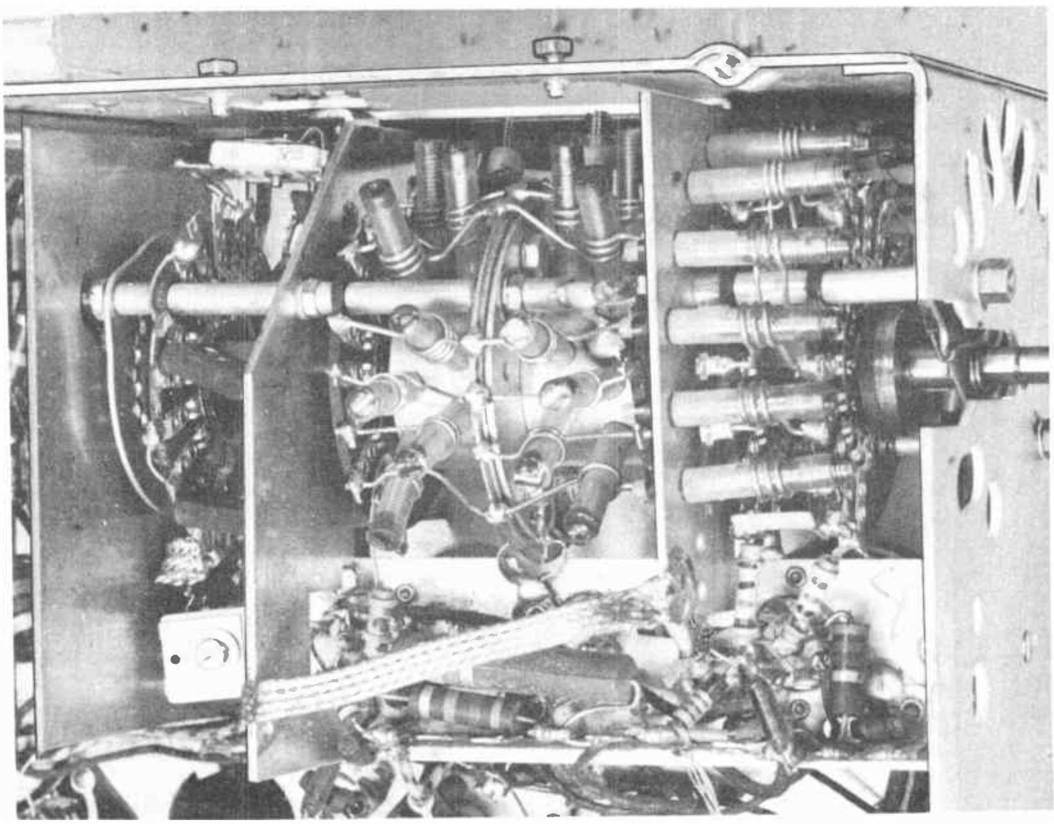


Fig. 43-8. Tuner having individual coils for each channel in circuits for r-f plate, mixer grid, and r-f oscillator.

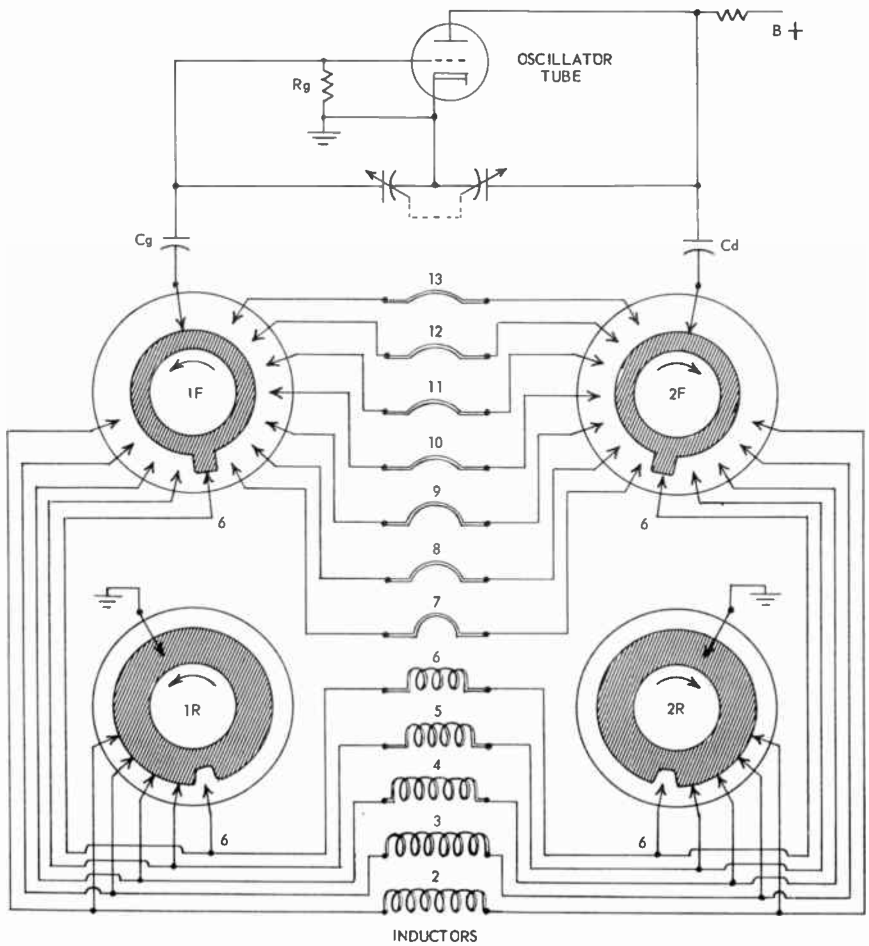


Fig. 43-9. Selector switch connections for shorting and grounding both ends of unused r-f oscillator inductors.

At the right in Fig. 43-7 there are separate inductors for each channel, and in each inductor or coil there is an adjustable core for alignment. This general principle is employed for the tuner pictured, as mounted in a chassis, by Fig. 43-8. The coils at the right tune the r-f oscillator, with connections from the two ends of each coil going to contacts on two switch wafers. In the center compartment are coils for the mixer grid, connected through a single switch wafer, and also the coils for the plate of the r-f amplifier, connected through another switch wafer. At the left-hand end of this tuner are capacitors used for adjustable tuning of the r-f amplifier grid circuit on the low-band channels.

R-f oscillators commonly are tuned with separate coils or inductors for each channel. Each oscillator coil may be provided with a movable core to allow precise adjustment of oscillator frequency for its channel. Sometimes only the oscillator coils for the low-band channels have movable cores. Oscillator inductors or coils which are not in the active circuit may or may not be shorted to ground.

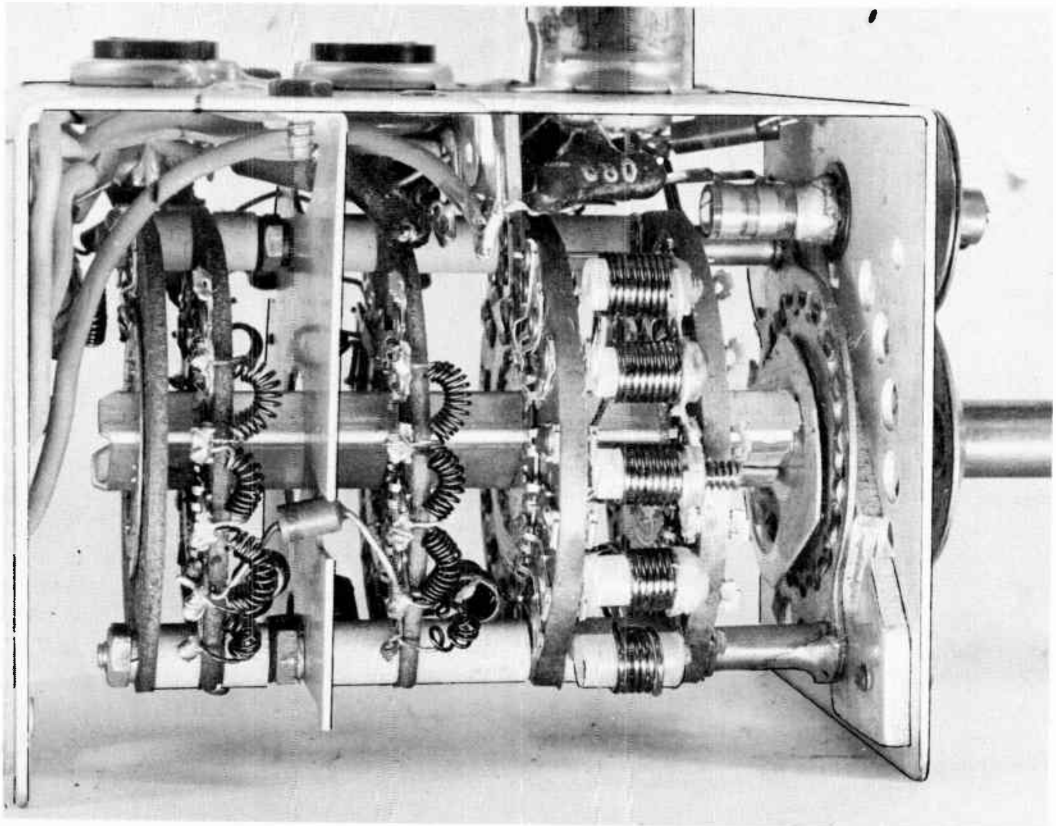


Fig. 43-10. Coils for tuning the low bands mounted on and between switch wafers in a tuner.

Fig. 43-9 shows switching connections with which unused oscillator coils for low-band channels are shorted together and grounded. The inductors or coils for each channel appear in the center of the diagram. These elements are mounted between the outer edges of two switch wafers shown at the left and right. Rotor connections on the front of number 1 wafer are shown at 1F and those for the rear of this same wafer at 1R. Connections for the front of number 2 wafer are shown at 2F, and those for the rear of this wafer at 2R. All the rotors turn together in the directions indicated by curved arrows.

The rotor at 1F has a narrow extended tongue that connects the grid of the oscillator tube, through capacitor *Cg*, to one end of whichever inductor is to be used for a selected channel. The other end of that inductor is connected by means of the tongue on rotor 2F to the plate circuit of the oscillator tube.

The shorting and grounding rotors are on the rear of the switch wafers, at 1R and 2R. The switch is in position for reception of channel 6. All the low-band inductors except the one for channel 6 are shorted together at one of their ends, and are connected to ground by rotor 1R. The same inductors are shorted together at their other ends, and are grounded, by rotor 2R.

The principle just discussed is employed in the tuner pictured by Fig. 43-10. Toward the right may be seen the five coils for oscillator tuning in the low-band channels. These coils are supported by the two switch wafers that are on either side of the coils. The movable core of the center coil has been screwed nearly all the way out, to make it visible in the picture. Toward the left are small coils for tuning the mixer grid circuit and the r-f amplifier plate circuit to the low-band channels.

The oscillator tuning system of Fig. 43-9 might be shown on a service diagram as in Fig. 43-11. Circuit connections are easier to follow in the service diagram, but the switch construction is not shown so clearly. Service diagrams are drawn in whatever way the manufacturer or service organization feels will best indicate the connections, and, in a general way, the type of switch construction. When you have both the service diagram and the tuner or receiver in front of you, the relations between parts and their wiring always become clear.

It would be possible to short circuit and ground all the high-band inductors which are not active, as well as those for the low band. The low-band inductors are shorted, or shorted and grounded, because their rather large inductance in combination with their distributed capacitance and stray capacitances could form circuits which would be resonant in themselves at frequencies in the high band. Power absorbed in such self-resonant circuits might interfere with high-band reception. In the high-band tuning elements the inductances and capacitances are too small to cause self-resonance at frequencies within the television carrier range, so shorting and grounding are not considered as necessary in most designs.

Not all tuners which are constructed with rotary selector switches have the inductors or coils mounted around the edges of the switch wafers or mounted between two wafers. The inductors sometimes are carried by a supporting framework of metal or of insulation placed either around or between the switch wafers. The coils in the center compartment of the unit in Fig. 43-8 are thus mounted. In other tuners all the coils are carried on a flat metal plate which is mounted directly above the rotary selector switch. Fig. 43-12 is a view from underneath a chassis on which is mounted a three-gang rotary selector switch for tuning or channel selection. The various coils are mounted through the chassis. They appear around the inner end of the switch. These coils are adjustable either from above or below the chassis.

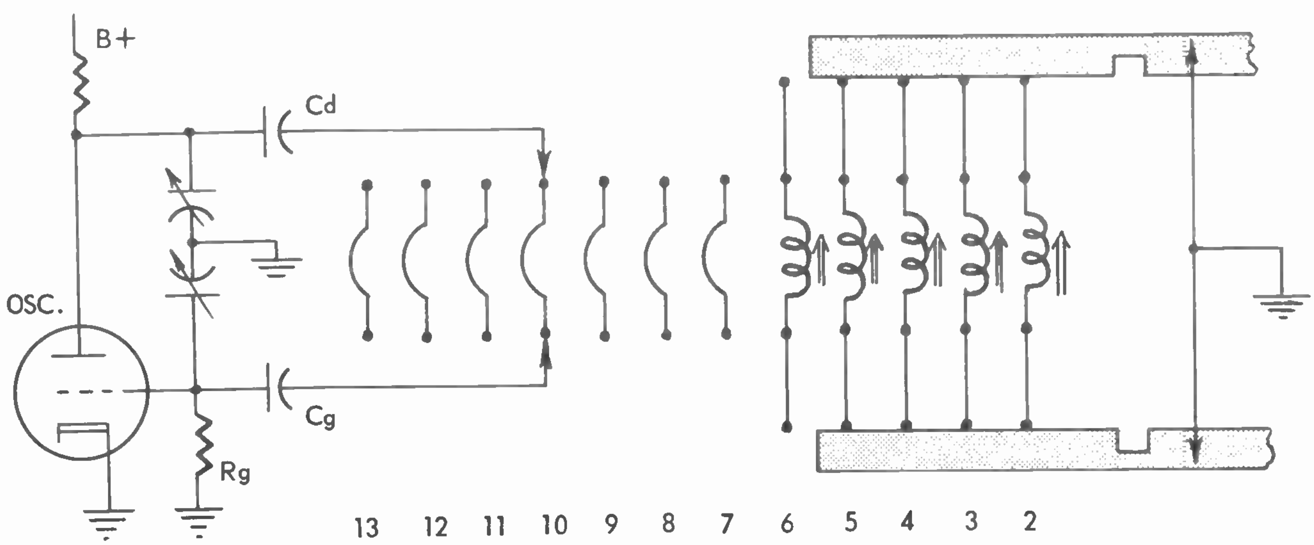


Fig. 43-11. How short-circuiting connections and tap connections may be shown on service diagrams.

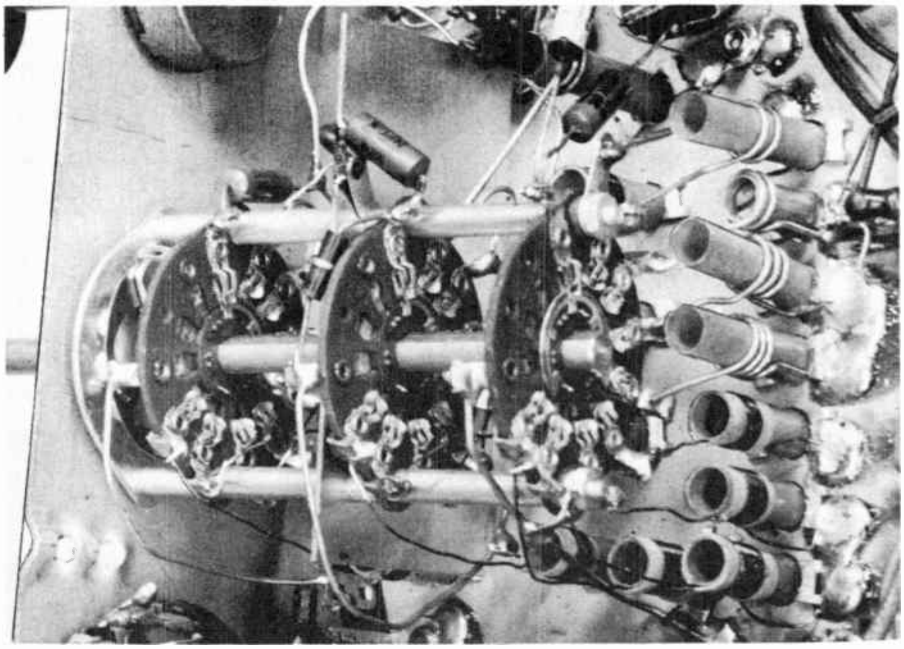


Fig. 43-12. Rotary selector switch with separately mounted coils to form a tuner.

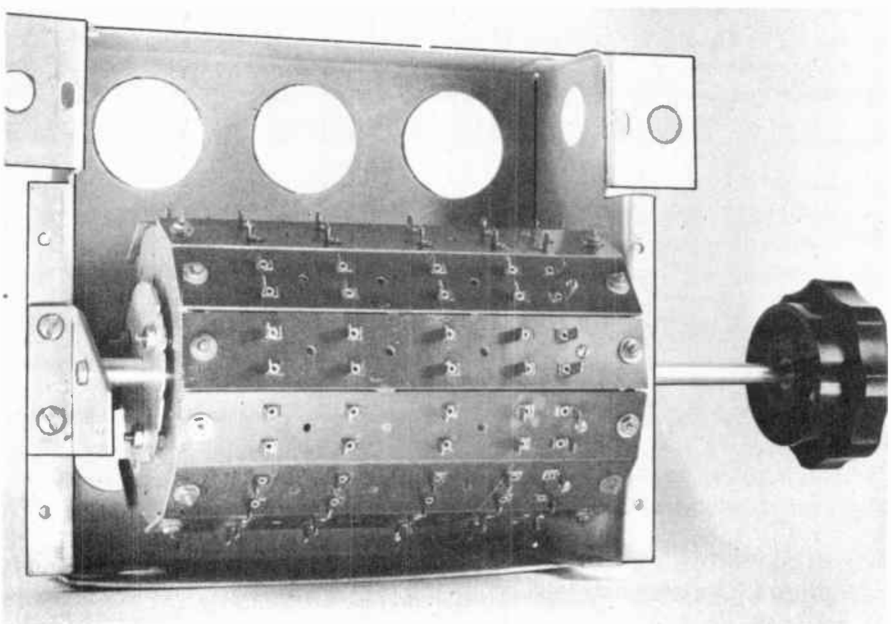


Fig. 43-13. The rotary drum of a turret tuner.

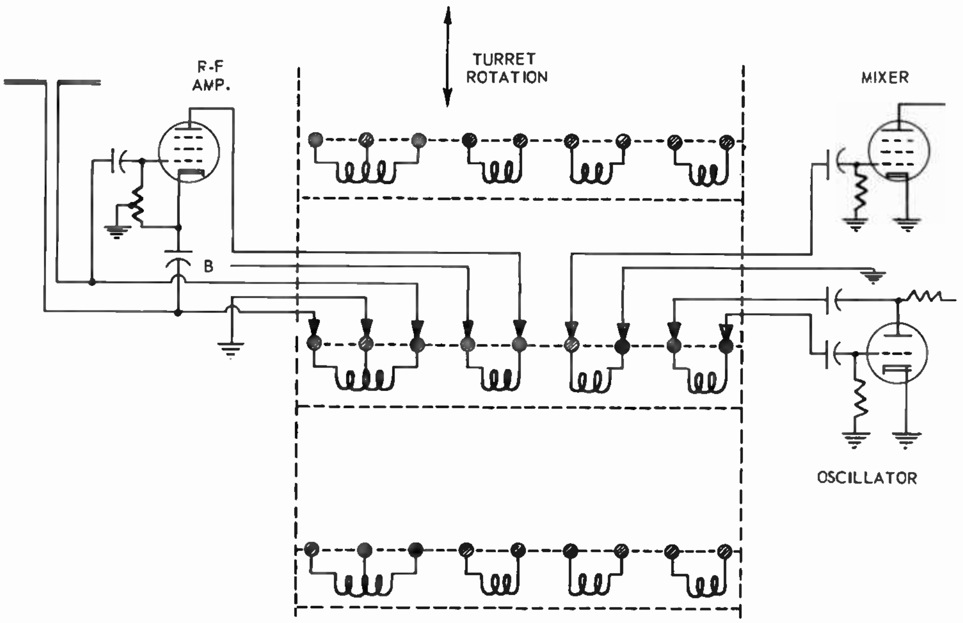


Fig. 43-14. The principle of channel inductor switching in a turret tuner.

TURRET TUNERS. We have been looking at tuners which are constructed with rotary selector switches. In all these types the inductors or coils are stationary, and channel connections are changed by movement of the switch rotors. In another widely used class of tuners, the turret tuners, the inductors are moved successively into positions where they make contact with stationary switch elements that are connected into the tube and antenna circuits.

Fig. 43-13 is a picture of a turret unit for a 12-channel tuner. The turret is a cylindrical drum around the outside of which are mounted strips of insulation, one strip for each channel. On the exposed outer surfaces of the strips are metallic contacts which as the turret is rotated to various channel positions, come into engagement with stationary contacts not shown in this illustration. On the concealed inner surface of each strip are all the inductors or coils required for tuning of one channel. The turret is rotated by the channel selector knob.

The principle of turret tuning is illustrated by Fig. 43-14. Broken-line enclosures down through the center of the diagram represent inductor strips for any three adjacent channels. On each strip are mounted, from left to right, a center-tapped antenna coil, a coupling transformer primary winding for the r-f plate circuit, the transformer secondary for the mixer grid circuit, and an oscillator tuning coil.

The middle one of the three strips is in the position where its coil terminals engage a series of stationary contacts indicated by arrowheads. From these stationary contacts there are leads to tube sockets and other circuit elements. When the turret is rotated in either direction the inductor strips move around with the drum. Then the strip whose terminals now engage the stationary contacts will be moved away from these contacts, and the strip for another channel will be brought into the active position. The inductors mounted on each strip are of such values as tune correctly for reception of one channel.

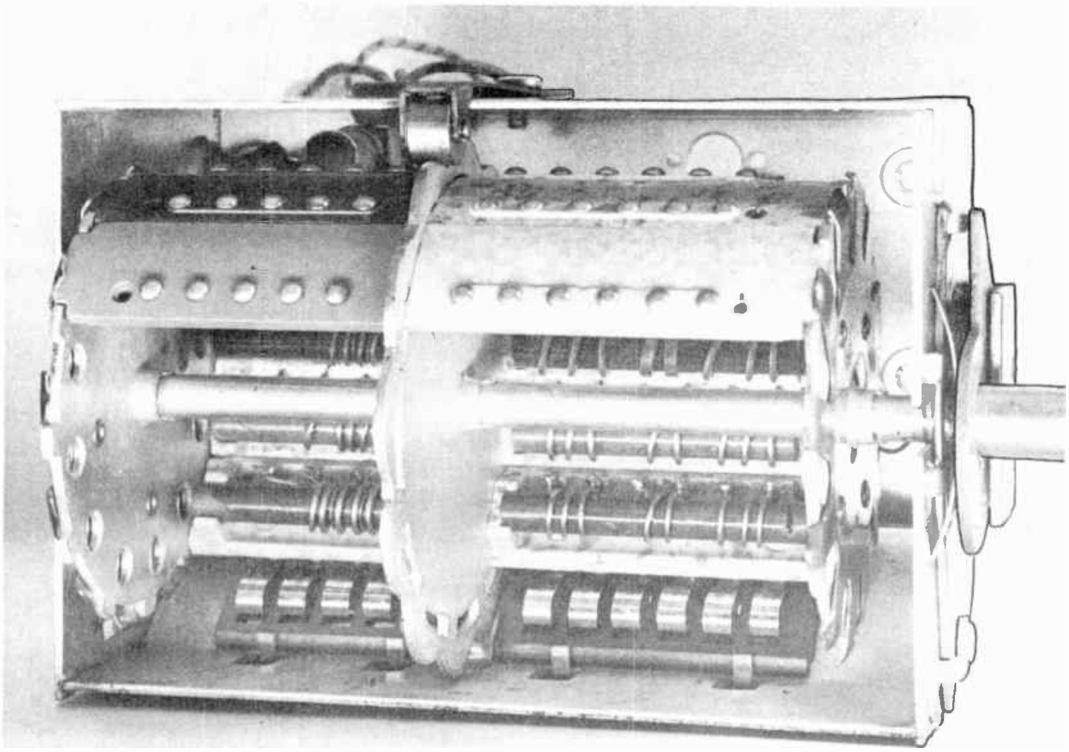


Fig. 43-15. Turret tuner from which some of the inductor strips have been removed to show the construction.

The turret inductor strips are quite easily removed and replaced on the rotary drum. Some strips are held with screws, or studs and nuts, but most are attached with some form of spring clip. Fig. 43-15 shows a turret tuner from which a number of inductor strips have been temporarily removed to expose the inside of the drum. At the top of the drum may be seen the outsides of several remaining strips, on which are the terminal studs or buttons arranged in a row from left to right. Inside the drum appear the coils or windings which are carried on the inner surfaces of several strips. Down below the drum, mounted on the inside of the tuner frame or housing, appear the stationary contacts which are engaged by the strip terminals as the drum is rotated.

On the tuner pictured by Fig. 43-15 the inductor strip for each channel is made up of two parts, or there are two strips in line for each channel. On the strips toward the left are five terminal studs, and on those toward the right there are six studs. In Fig. 43-16 are shown pairs of strips for two channels. These strips have been removed from the drum and turned to expose the coils which normally are toward the inside of the drum.

The pair of inductor strips at the top of Fig. 43-16 are for one of the high-band channels. The pair down below are for one of the low-band channels. The five-terminal strips at the left carry the two windings for a transformer that couples the antenna to the grid circuit of the r-f amplifier. The antenna winding has a center tap which is grounded, while the grid winding has only two end terminals.

The six-terminal strips at the right carry a transformer for coupling the plate circuit of the r-f amplifier to the grid circuit of the mixer, and carry also the tuning coil for the r-f oscillator. The two windings of the coupling transformer are connected to four of the terminals, and the single winding for the oscillator is connected to the remaining two of the six terminals on these strips.

Fig. 43-17 is a complete circuit diagram for the turret tuner whose construction we have been examining. The five-terminal inductor strip for antenna to r-f coupling appears at the upper left in the diagram. The six-terminal strip for r-f to mixer coupling and oscillator tuning is shown between the r-f amplifier tube and the twin-triode mixer-oscillator tube. The stationary contacts are represented by arrowheads. The adjustable capacitors at a, b, c and d are used during alignment, as will be discussed when we come to the general subject of alignment for television tuners. The capacitor at *E* is a fine tuning control which allows the set operator to make such changes of oscillator frequency as permit best reception.

TUNER CIRCUIT DETAILS. The most generally used types of tuners are those having rotary selector switches with stationary inductors and those of the turret type having inductors mounted on a movable drum. The other types listed in the earlier table of tuner design and construction will be examined presently, but first it will be well to discuss some features which affect all tuners.

In all tuners of recent design the tubes are miniature types. R-f amplifiers most often are sharp cutoff pentodes. There are, however, many tuners in which the r-f amplifier is a twin-triode. Remote cutoff pentodes are used in a few designs, but not many. Mixer tubes which are not in the same envelope with the oscillator most often are sharp cutoff pentodes. R-f oscillators practically always are triodes. The triode oscillator may be a separate tube or it may be one section of a twin-triode whose other section is used as a triode mixer.

In some designs there is no variable tuning for separate channels in the coupling between antenna and r-f amplifier grid circuit. Then this coupling must cover the entire range of television carrier frequencies,

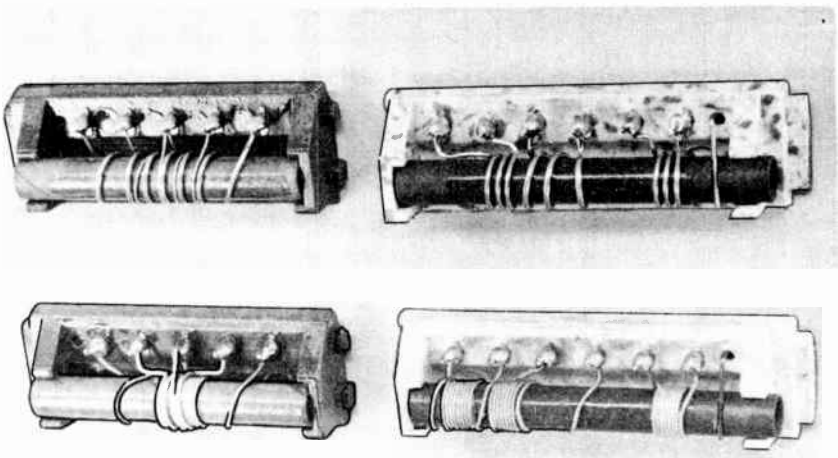


Fig. 43-16. Inductor strips for one type of turret tuner.

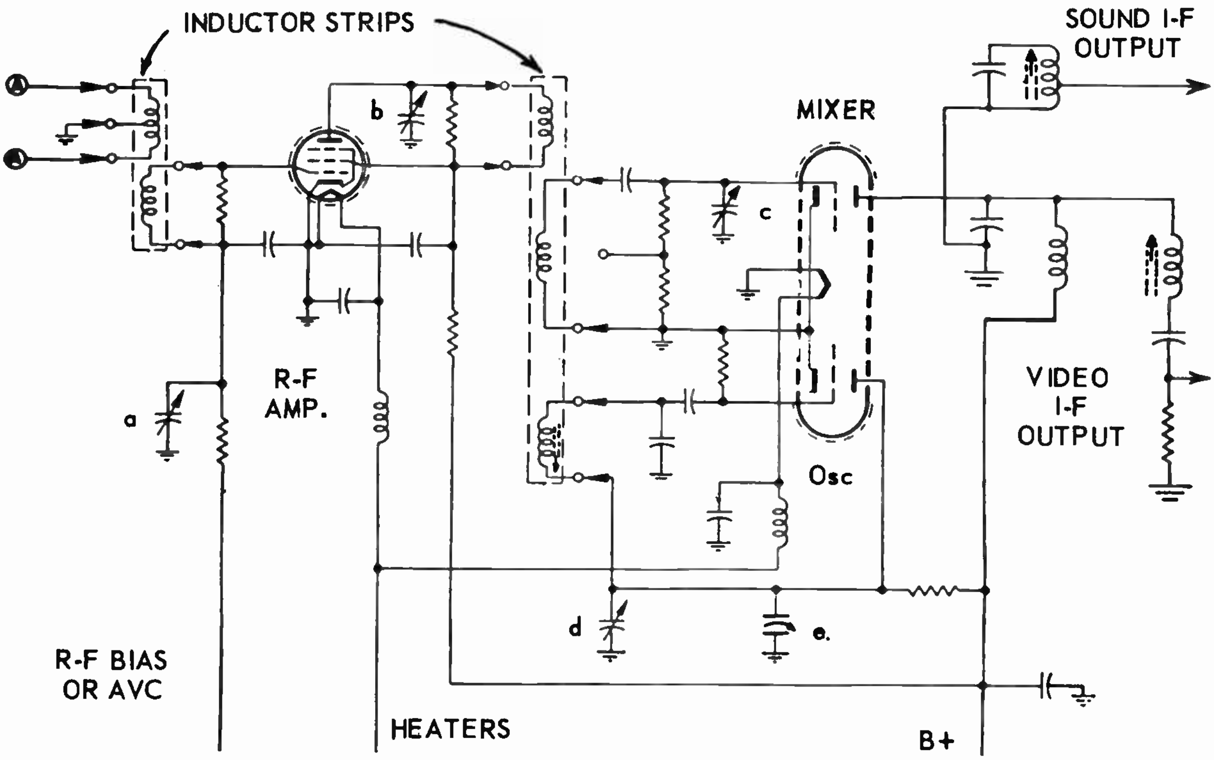


Fig. 43-17. Circuits from antenna to video and sound i-f outputs in a turret tuner.

and there can be little or no gain in this portion of the circuit. It always is true that the wider the range of frequencies to be covered the less may be the gain. In extending the frequency range we unavoidably lose the advantage of high peaking or high gain at resonance.

In other designs of antenna to r-f coupling there is one value of inductance for low-band channels and a smaller value for high-band channels. This permits somewhat greater gain in each band. The greatest r-f gain is possible when there are different inductors or different values of inductance for each separate channel.

(4) Along with any increase of gain from antenna through the r-f amplifier there is an increase of selectivity against unwanted high-frequency signals and many forms of interference. There is also a decrease of radiation of the r-f oscillator frequency through the antenna. Such oscillator radiation is proving to be one of the most serious troubles in television reception. A poorly designed or constructed receiver can radiate with enongh strength to interfere with normal reception of many other receivers in the neighborhood.

When it is desired that a tuned r-f circuit cover some limited range of frequencies it is usual practice to design the inductance and capacitance elements for the highest practicable Q-factor (least high-frequency loss) and then to broaden the response by connecting a fixed resistor across the tuned circuit. Such a resistor is used across the r-f grid winding of the antenna coupler in Fig. 43-17, and another is used across the primary of the r-f to mixer coupling transformer in the plate circuit of the r-f amplifier.

(4)

For any given frequency of resonance and a fixed ratio of inductance to capacitance in the tuned circuit, the band width is inversely proportional to resistance connected across the tuned circuit. Using twice as much resistance decreases the band width to one-half, and using half the original resistance doubles the band width. The band width is measured as the difference between low and high frequencies at which the voltage gain drops to 70 per cent of its maximum or peak value. When shunting resistances are used in this manner it is possible to raise the gain and narrow the response with more resistance, and to drop the gain while securing a wider response by using less resistance. Such changes sometimes are made during service work when it is necessary to raise the gain for certain channels in which the received signal is weak.

ANTENNA COUPLINGS. Fig. 43-18 shows some of the many circuits which are used for coupling the television antenna to the grid-cathode circuit of the r-f amplifier. All the couplings illustrated in this figure are used in receivers whose input impedance at the antenna terminals is 300 ohms. At *A* there is a center-tapped inductor across the antenna terminals, with the tap grounded and the outer ends connected directly to the tube cathode and to the grid through a blocking capacitor that isolates the grid and its agc (automatic gain control) bias voltage from the ground connections.

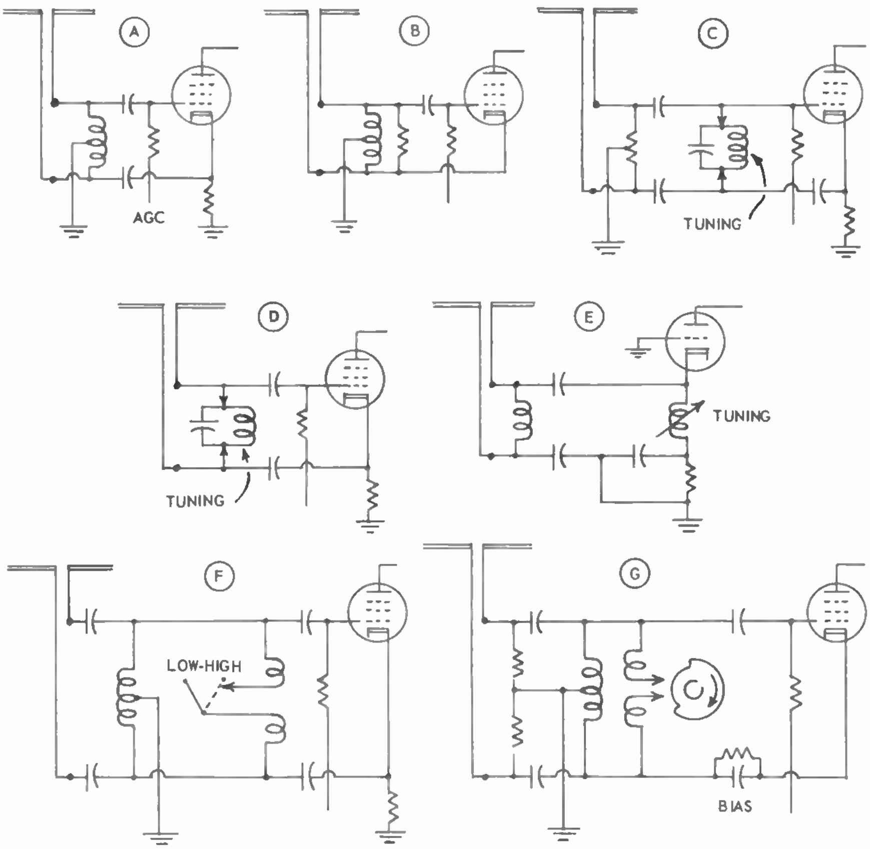


Fig. 43-18. Couplings which may be used between the antenna terminals and the grid-cathode circuit of the r-f amplifier tube.

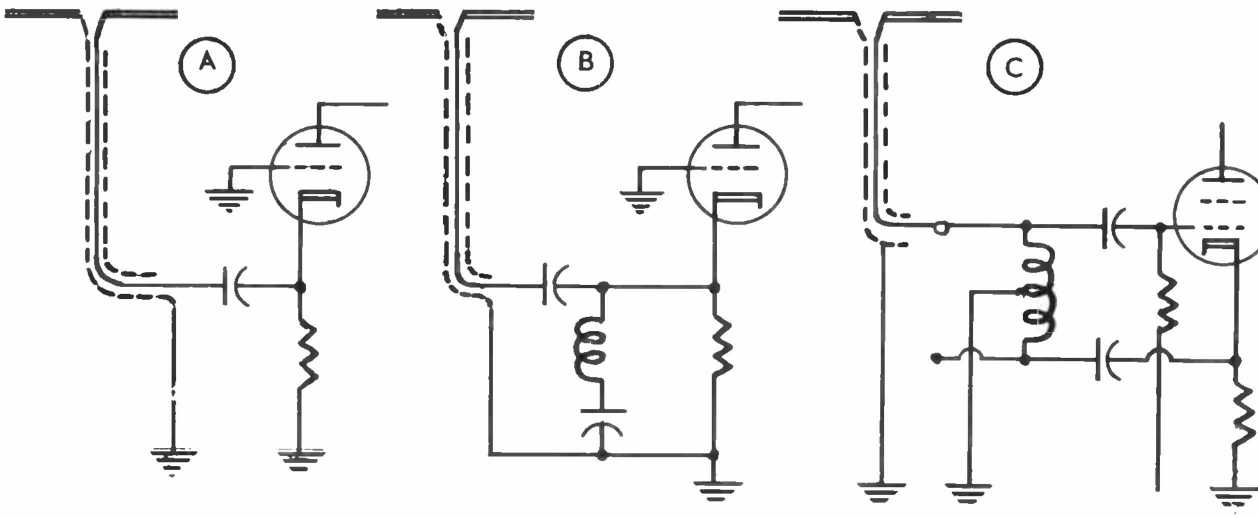


Fig. 43-19. Couplings for 72-ohm coaxial transmission line to r-f amplifiers.

The connection from the agc bus through a resistor to the tube grid is shown also in most of the other diagrams. This is to illustrate how this method of gain control may be applied to the r-f amplifier tube, but it should be understood that many r-f amplifiers are operated with other methods of biasing.

In diagram *B* there is again a center-tapped inductor across the antenna terminals. A resistor for widening the frequency response is connected across this inductor. The cathode line returns to ground through the lower part of the tapped inductor, whereas in diagram *A* the cathode line goes to ground through a biasing resistor in series with the cathode.

Diagram *C* shows a center-tapped resistor instead of an inductor across the antenna terminals. Across the antenna circuit is a tuning element consisting of a capacitor and inductor in parallel with each other. This element may be mounted on the drum of a turret tuner, or it may be mounted between wafers of a rotary selector switch, depending on the type of tuner construction. In any case this is one of a number of tuning elements brought into the circuit for the various channels. Diagram *D* shows a generally similar tuning element connected directly across the antenna terminals. Again this is one of the tuning elements brought into the circuit for the several channels.

At *E* the r-f amplifier tube is of the grounded grid type with signal input to the cathode circuit and signal output from the plate as usual. This is a method which permits using a triode tube in a very-high frequency circuit while presenting feed-back from plate to grid due to the fact that the grid is the element connected to ground. Connections from the two antenna leads are made across an inductor in series with the cathode. The antenna circuit is tuned, or its impedance is matched to that of the antenna itself, by varying this cathode inductance for the several channels. A single inductor at this point may be varied in value by shorting portions of it by means of one rotor on a rotary selector switch, or different inductors may be switched into the circuit for each channel.

At *F* in Fig. 43-18 is shown one method of changing the input impedance of the antenna circuit to suit either the high-band or low-band channels. At the left is a center-tapped inductor which always remains connected between the antenna terminals. Toward the right is a two-section inductor which is open circuited with the switch in the low-band position and is closed between the inner ends with the switch in the high-band position. With the switch closed this auxiliary inductor is effectively in parallel with the first one. The self-inductance of two inductors in parallel with each other is less than that of either inductor alone. Consequently, closing the switch for high-band reception reduces the inductance in the antenna circuit. At the higher carrier frequencies this reduced inductance has approximately the same inductive reactance or impedance as has the single inductor at the left for reception at the lower carrier frequencies of the low-band channels.

The diagram at *G* again shows an auxiliary inductor whose circuit is opened for low-band reception and closed for high-band reception. The switching element here is represented as one of the rotors on a selector switch. In this diagram the cathode line is returned to ground through the lower part of the fixed inductor across the antenna terminals. In series with the cathode line are a capacitor and resistor for biasing the tube.

Input impedances at the antenna terminals of television receivers are fairly well standardized at 300 ohms. This particular impedance allows a good "match" and satisfactory power transfer from the most generally used types of antennas, whose impedances are approximately 300 ohms at the carrier frequency in the middle of the band over which the antennas are designed to operate. With 300-ohm impedances of antenna and receiver the "transmission line" that connects the two together should be of a type having this same value of impedance. Incidentally, in some later lessons we shall become well acquainted with antennas, transmission lines, and all their peculiarities and problems. At present we are interested only in connections to receivers.

Other types of antennas, which may be used to meet special requirements of reception, have impedances of 72 ohms. These antennas are connected to receivers through transmission lines whose impedance is 72 or 75 ohms. Many receivers have been designed with input impedances of 72 ohms to match these 72-ohm antennas and lines. In many other receivers the input circuits are arranged for connection of either 300-ohm or 72-ohm antennas and lines.

At *A* and *B* in Fig. 43-19 are shown connections of 72-ohm antennas and 72-ohm transmission lines to receivers having input impedances of this value. Both diagrams show grounded grid r-f amplifiers. The 72-ohm transmission line commonly used has a central conductor covered with insulation around which is a braided metallic shield protected with an outer coat of insulation. This is called coaxial cable or coaxial line. The insulated central conductor is connected to the cathode input circuits of the r-f amplifiers. The shield is connected to ground.

When there are center-tapped input inductors, as at A, B and F of Fig. 43-18, it usually is possible to have satisfactory reception from a 72-ohm antenna and line by connecting the line across either half of the inductor instead of across the ends as for 300 ohm matching. This connection allows impedance matching, for reasons which follow.

If the impedance across an entire inductor or coil is 300 ohms the impedance across either half will be close to 75 ohms. This relation exists because inductive reactance, the principal factor in impedance, varies as the square of the number of turns. Should there be 6 turns in all, the total impedance will be proportional to the square of 6, which is 36. Then the impedance of half this winding, or of 3 turns, will be proportional to the square of 3, which is 9. Since 9 is one-fourth of 36, the impedance of half the coil will be one-fourth that of the whole coil.

Diagram C of Fig. 43-19 shows a 72-ohm antenna and transmission line connected across one half of the input inductor of a tuner primarily designed for a 300-ohm antenna and line. The insulated conductor of the line is connected to one end of the center-tapped inductor, usually to the grid end. The shield of the line is connected to ground and through ground to the center tap of the input inductor. The terminal at the lower end of the antenna inductor remains open or unconnected.

TELEVISION

LESSON NO. 44

TELEVISION TUNER DESIGN AND CONSTRUCTION

Training Your Mind...

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

DO NOT TEAR — CUT ALONG THIS LINE AND SEND IN FOR GRADING.

Training Your Mind
C.T.I.
Perfecting Your Skill

COMMERCIAL
TRADES
INSTITUTE

Chicago, Ill.

LESSON NO. 44

TELEVISION TUNER DESIGN AND CONSTRUCTION

In the preceding lesson we examined the operation of television tuners built with rotary selector switches and those having rotary turrets. In both types the resonant frequency is varied by changes of inductance. In a third general type of television tuner which operates with changes of inductance, the variable elements consist of spiral conductors, around the turns of which travel small contact brushes to short circuit more or less of the spiral. A three-element variable inductance unit of this type is pictured by Fig. 44-1.

Each spiral conductor is of flat section with one edge molded into a support of low-loss insulation, leaving exposed an edge around which travels a contact brush that is supported from and rotated by a shaft of insulation material. Some tuners have three of these spiral elements, as illustrated, while others have four such elements. The contact brushes for all elements are rotated together by the common shaft. The brushes are pivoted on their rotating support arms so that the brush ends follow the spiral edge of the conductor as the shaft is rotated.

Connections from the wiring terminals to the spiral and the brush are shown by Fig. 44-2. One of the external terminals is connected to the outer end of the spiral conductor and to the contact brush. The other external terminal is connected to the inner end of the spiral. These connections for each inductor element usually are shown on service diagrams by some symbol like the one at the right in the figure.

Whatever portion of the inductance spiral is between the outer end and the contact brush is short circuited by the brush connection, and is inactive. The remaining portion of the inductance, between the brush contact and the inner end of the spiral, is not shorted and is active in the connected circuit.

Clockwise rotation of the shaft and brush brings the brush contact to the inner end of the spiral. Here all of the turns are shorted and there is minimum possible inductance. This remaining inductance, due to lengths of conductors, amounts to about 0.025 microhenry in each element. Counter-clockwise shaft rotation brings the brush contact to the outer end of the spiral, leaving none of the turns shorted and all of the inductance active. This maximum inductance is slightly less than 1.0 microhenry in each element.

Note that we have here a ratio of maximum to minimum tuning inductance of nearly 40 to 1. This very great change of inductance allows tuning throughout the entire range of television carrier frequencies, all the way from 54 megacycles at the bottom of channel 2 through to 216 megacycles at the top of channel 13, and including the f-m broadcast band and other radio services between the low-band and high-band television carriers.

The full range of tuning for the unit of Fig. 44-1 requires six turns of the shaft. Television channels 2

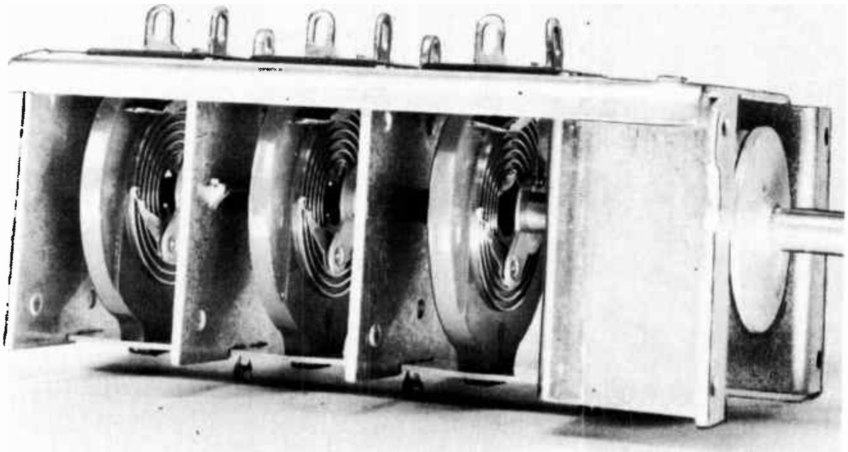


Fig. 44-1. Variable inductance tuning element of the type employing spirally wound inductors.

through 6 come in during the first two and one-half turns. The next three-quarters of a turn covers the f-m broadcast band. The high-band television channels tune within the final one and one-quarter turns of the shaft. In the DuMont Inputuner, which uses the inductance elements being discussed, there is a frequency skip between the top of the f-m band and the high-band television carrier frequencies. This tuner covers the television and f-m channels with only four turns of its shaft.

Fig. 44-3 is a diagram of a tuner circuit employing the variable spiral inductors. Inductor element *La* is in the plate circuit of the r-f amplifier. Inductor *Lb* is in the grid circuit of the mixer. Inductor *Lc* tunes the oscillator. At *Ld* is a small diameter coil of possibly two or three turns in series with spiral inductor *La*. This small added coil is changed in length to vary the total minimum inductance when the brush is in position to short circuit all of the spiral inductor. This is done during alignment. Similar small coils at *Le* and *Lf* are in series with the other two spiral tuners for adjusting minimum inductances in their circuits. At *Rb* and *Rb* are fixed resistors in parallel with inductors for the r-f amplifier and mixer circuits. These resistors are for the purpose of broadening the tuning band. Band width is adjusted also by the variable capacitors and the inductor at the top of the diagram, these elements forming the coupling between r-f amplifier and mixer circuits. At *Cb* and *Cb* are blocking capacitors in the r-f amplifier plate circuit and the mixer grid circuit.

The spiral inductor element *Lc* which is used for the r-f oscillator is just like the other inductor elements, having the same limits of inductance and the same rate of change of inductance with shaft rotation. But in order to have the oscillator frequency remain higher than that of the receiver carrier it is necessary to have less inductance in the oscillator circuit than in the r-f amplifier and mixer circuits. Oscillator circuit inductance is reduced by connecting coil *Lg* in parallel with the variable inductor *Lc* and end inductor *Lf*. The paralleled inductances have less combined inductance than exists in either one alone.

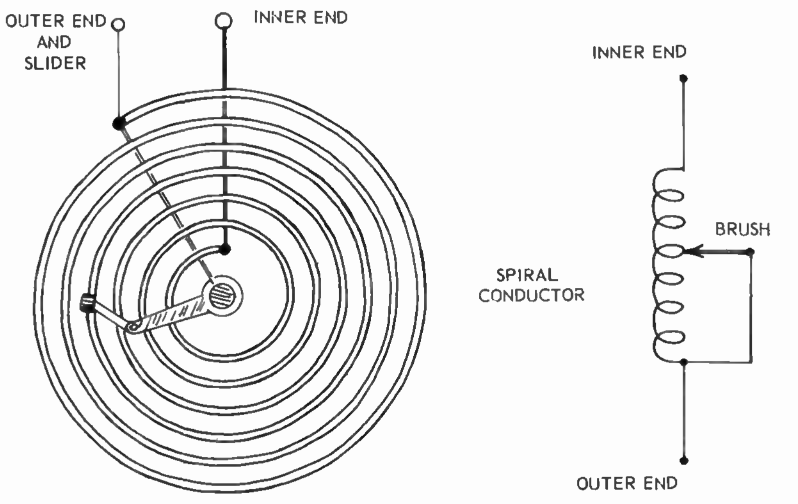


Fig. 44-2. Connections to one spiral element of a variable inductance tuner.

In this inductance-tuned oscillator circuit the paralleled inductance *Lg* serves much the same purpose as a padder capacitor in some capacitance-tuned oscillator circuits. Just as a padder capacitor in series reduces the effective capacitance of an oscillator tuning capacitor, so the added inductance in parallel reduces the effective inductance of the oscillator tuning inductor.

Capacitor *Cf* of Fig. 44-3 is an adjustable fine tuning control by means of which the set operator can make slight changes in oscillator frequency. Capacitor *Cg* and resistor *Rg* are the grid capacitor and resistor for biasing the oscillator. Capacitor *Cc* couples the oscillator frequency into the mixer grid circuit.

In earlier models of slider-type variable inductance tuners the elements are coils wound with spaced turns on a long rotating shaft of low-loss insulation. Bearing on the conductor of each coil is a contact brush that maintains connection between the coil and a fixed conductive plate. Rotation of the coil form by means of the tuning shaft moves the brushes along the turns to short circuit more or less of the inductance. Electrical characteristics are like those mentioned for the spiral elements. Ten turns of the shaft are required to change the inductance from minimum to maximum. The inductor unit with elements in line is used in the same tuner circuits as the later spiral type.

6 *TUNERS WITH MOVABLE CORES.* A fourth general method of varying the inductance in television tuners makes use of movable cores in the inductors. The basic principle of these tuners is similar to that employed for tuning in many a-m and f-m sound broadcast receivers, where the tuning dial operates a mechanism that moves the cores farther into or out of several coil windings which are in the tube circuits. The cores are of molded powdered iron. Moving such a core farther into its winding increases the inductance and lowers the resonant frequency, while moving the core out of the winding lessens the inductance and raises the frequency.

Fig. 44-4 shows the inductor portion of a television tuner employing this method of inductance change and frequency variation. There are three pairs of coils having spaced ribbon windings on tubular forms. The movable cores are inside these forms. Each core is attached to a threaded stud which may be turned to raise or lower the core to its correct position in relation to the plate which is underneath all the coils.

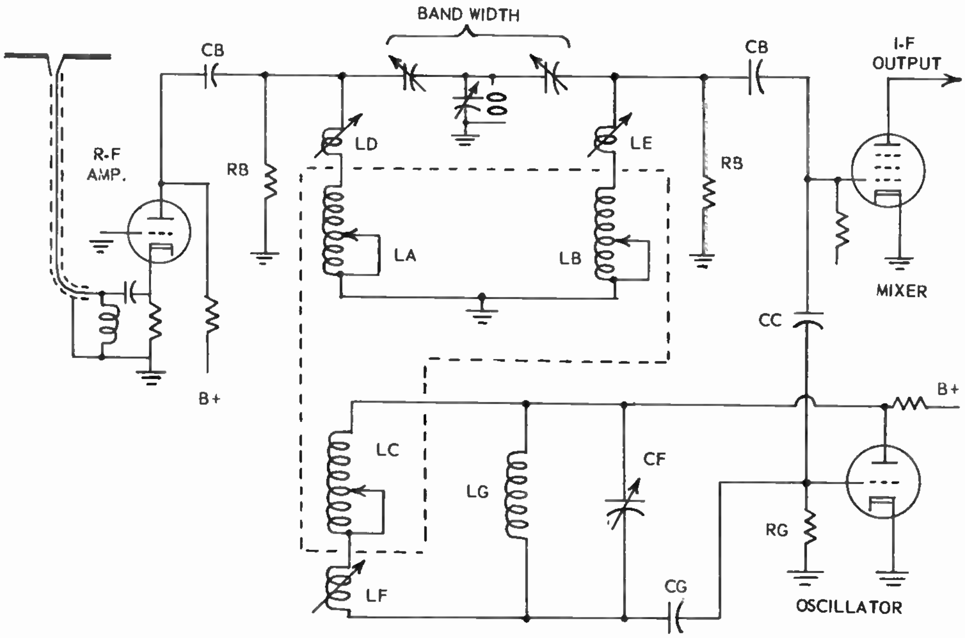


Fig. 44-3. Circuit for variable inductance tuner using either spiral or cylindrical continuously variable inductors.

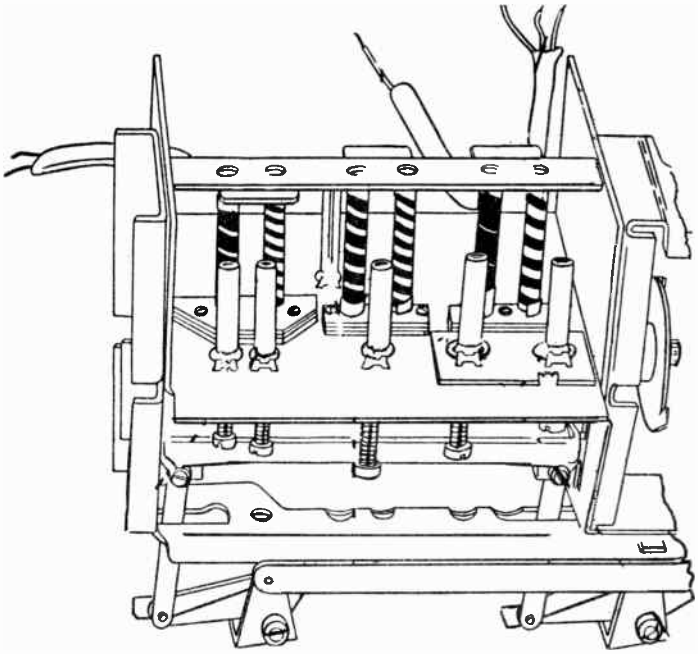


Fig. 44-4. Inductors and trimmer capacitors on a two-band tuner using continuously variable inductance in each band.

This plate is raised and lowered by mechanism attached to the tuning dial, so that all the cores are moved up and down together.

In each pair of coils the one with more turns is used for tuning the low-band channels, and the one with fewer turns tunes the high-band channels. There is continuous tuning through each band, rather than tuning to each channel individually as in tuners having selector switches or turrets. As the tuner dial is turned through its positions for channels 2 to 6 in the low band the cores are moved out of their coils to reduce the inductance and raise the frequency. Continued turning of the dial operates a switch that shifts tube circuit connections from the low-band coils to those for the high-band. As the dial is turned still farther in the same direction the cores are moved farther back into the coils to tune through channels 13 to 7 as inductance is increased and frequency lowered.

Fig. 44-5 is a fairly typical circuit diagram for a television tuner constructed with movable-core inductors. There are four two-position band switches shown in their high-band positions by full-line arrows and in low-band positions by broken-line arrows. High-band positions are marked *H*, low-band positions are marked *L*. One of these switches connects either of two antenna coupling transformers to the grid-cathode circuit of the r-f amplifier. The remaining switches shift the connections between the tuning inductors of each pair.

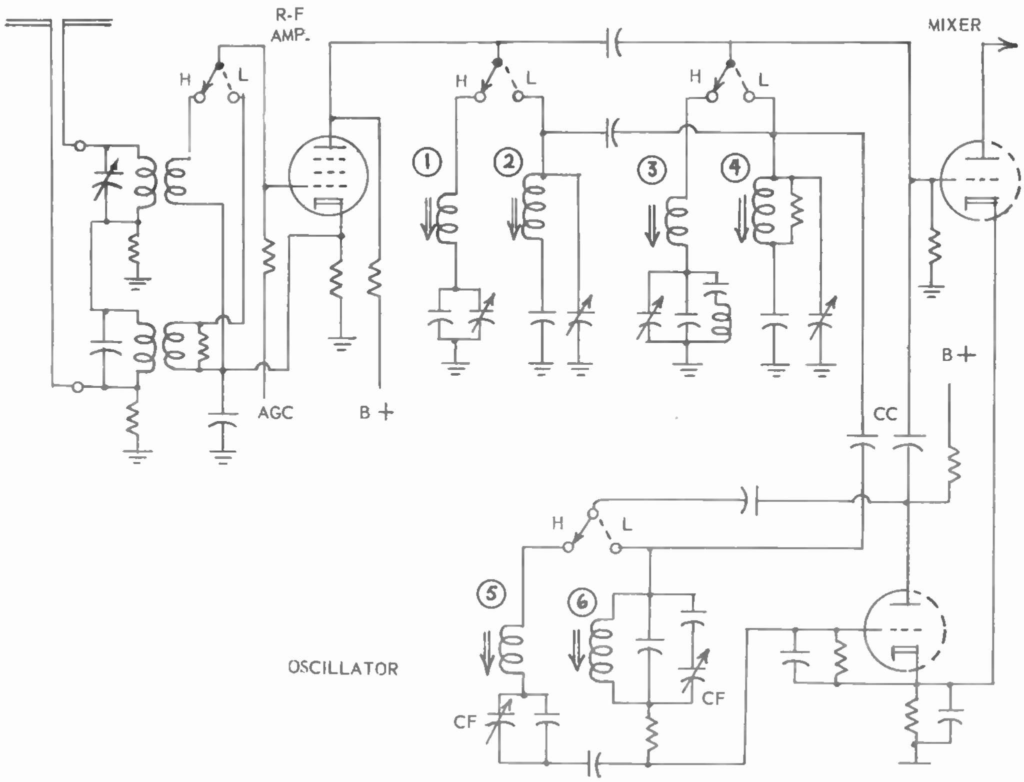


Fig. 44-5. Circuit for two-band tuner using continuously variable inductances in both bands.

Two of the adjustable inductors are in the plate circuit of the r-f amplifier. In series between the high-band inductor and ground are a fixed capacitor and an adjustable trimmer capacitor. In series with the low-band inductor is a fixed capacitor and in parallel with this inductor is an adjustable trimmer capacitor. Two more of the adjustable inductors are in the grid circuit of the mixer. In series and in parallel with inductors of this pair are frequency alignment circuits containing adjustable trimmer capacitors. These trimmers are used during the process of alignment to obtain a frequency response of correct band width, with a symmetrically shaped curve, and to obtain the greatest possible voltage gain while satisfying the other requirements.

In the r-f oscillator circuit shown in the lower part of the diagram are the inductors of another pair. Capacitors marked *Cf* adjust the oscillator frequency. These two capacitors may be operated together as a fine tuning control used by the set operator, or they may be service adjustments. The symbols for mixer and oscillator indicate that they are the two sections of a twin triode. This twin triode may be of a type having a single common cathode, since the two cathodes are shown connected directly together. Mixer grid bias is provided by the paralleled resistor and by-pass capacitor which are in series between the cathode and ground. Oscillator bias is of the grid-leak type, provided by the capacitor and resistor which are in parallel between the grid and cathode of the oscillator section. At *Cc* are coupling capacitors for feeding the oscillator frequency to the mixer grid circuit. (Note that the left-hand one of these two capacitors is in circuit only during low-band reception, thus providing greater oscillator voltage injection for this band than for the high band.)

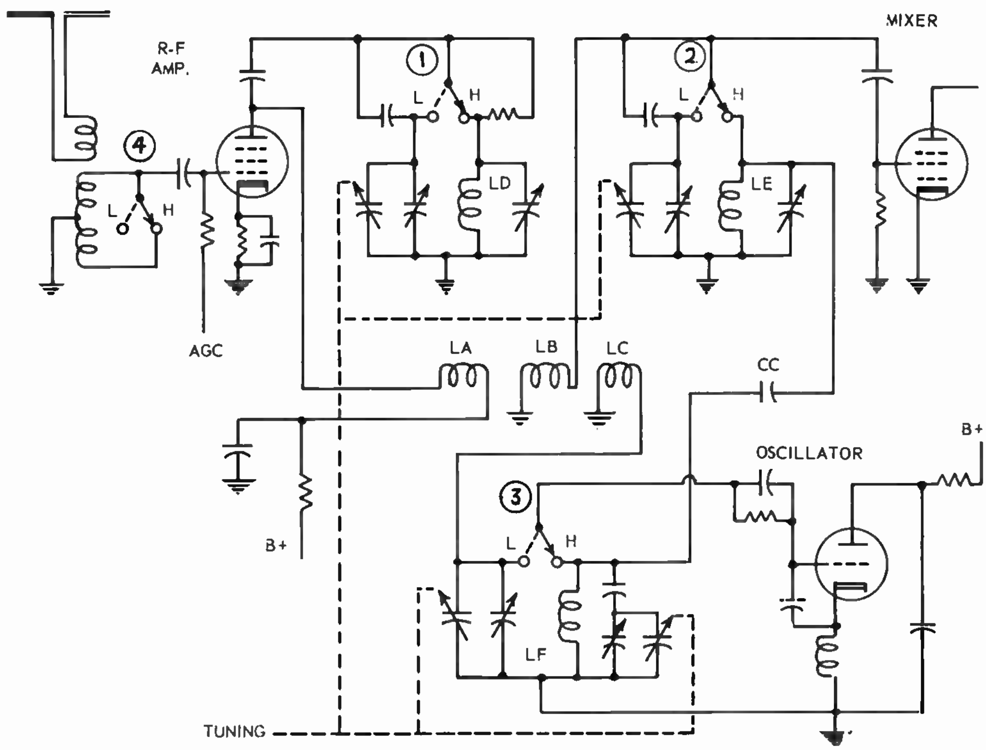


Fig. 44-6. Circuit of a tuner using variable capacitors for channel selection.

The double biasing of a twin tube having a single cathode is not a feature peculiar to this type of tuner, it might be used anywhere. The same is true of the two capacitors that couple the oscillator to the mixer. You must learn to watch for any such unusual features as you examine circuit diagrams. Your knowledge of the principles of television and radio should be enough to explain the significance. In some tuners of the movable core type there is a third band, for f-m broadcast reception. This range of tuning is provided by a third switch position at which additional coils are connected in parallel with the low-band inductors. Your knowledge of principles should allow you to recognize that the two inductances in parallel would lessen the effective inductance, raise the range of resonant frequencies, and allow tuning through the f-m band.

TUNING WITH CAPACITANCE. The tuner whose circuit diagram is shown by Fig. 44-6 operates with four variable capacitors whose rotors are turned together by the channel selector dial. These four capacitors are joined by broken lines in the diagram. One is in the plate circuit of the r-f amplifier, another is in the grid circuit of the mixer, and the remaining two are in the oscillator circuit. In each of these circuits there are tuning elements for the low-band channels and other elements for the high-band channels. These elements are cut into or out of the circuits by two-position switches which are shifted one way or the other as the selector dial is turned between its positions for channel 6, at the top of the low-band, and channel 7 at the bottom of the high-band. The switch positions are marked *L* for low band, and *H* for high band.

In addition to the three band switches in circuits for the r-f plate, mixer grid, and oscillator, there is a fourth band switch for the coupling between antenna and r-f grid. This antenna band switch operates on the secondary of the antenna coupling transformer, which is in two sections. With the switch in its high-band position the two parts of the secondary are connected in parallel with each other, to reduce the effective inductance. With the switch in the low-band position one section of the secondary is open circuited, leaving the effective inductance of the other section.

In the plate circuit of the r-f amplifier is inductor *La*, and coupled to this inductor is another one, *Lb*, in the grid circuit of the mixer. These two inductors provide coupling between the two tubes and at the same time are made resonant for received carrier frequencies by the capacitance tuning elements. When band switch number *1* is moved to its high-band position the inductor marked *Ld* is placed in parallel with inductor *La*, thus reducing the effective tuning inductance for the high-band channels. When band switch number *2* is in its high-band position, inductor *Le* is placed in parallel with inductor *Lb*, providing lessened inductance for high-band tuning of the mixer grid circuit.

In the oscillator circuit low-band tuning is accomplished with the variable capacitor which is connected when the band switch is in its low position and with inductor *Lc*. This inductor also couples the oscillator frequency into the mixer grid circuit because of its inductive coupling to inductor *Lb*.

With the oscillator band switch in its high position the variable capacitor and the inductor *Lc* are disconnected from the oscillator grid, and then are inactive because they are connected to ground at their other ends. High-band oscillator tuning is by means of inductor *Lf* and the capacitors connected across this inductor. During high-band tuning the oscillator frequency is coupled into the mixer grid circuit through capacitor *Cc*, since inductor *Lc* no longer is performing this function.

In the tuned circuits of Fig. 44-6 which are connected through band switch sections *1*, *2*, and *3* are various adjustable capacitors in addition to the tuning capacitors. These adjustable capacitors are identified

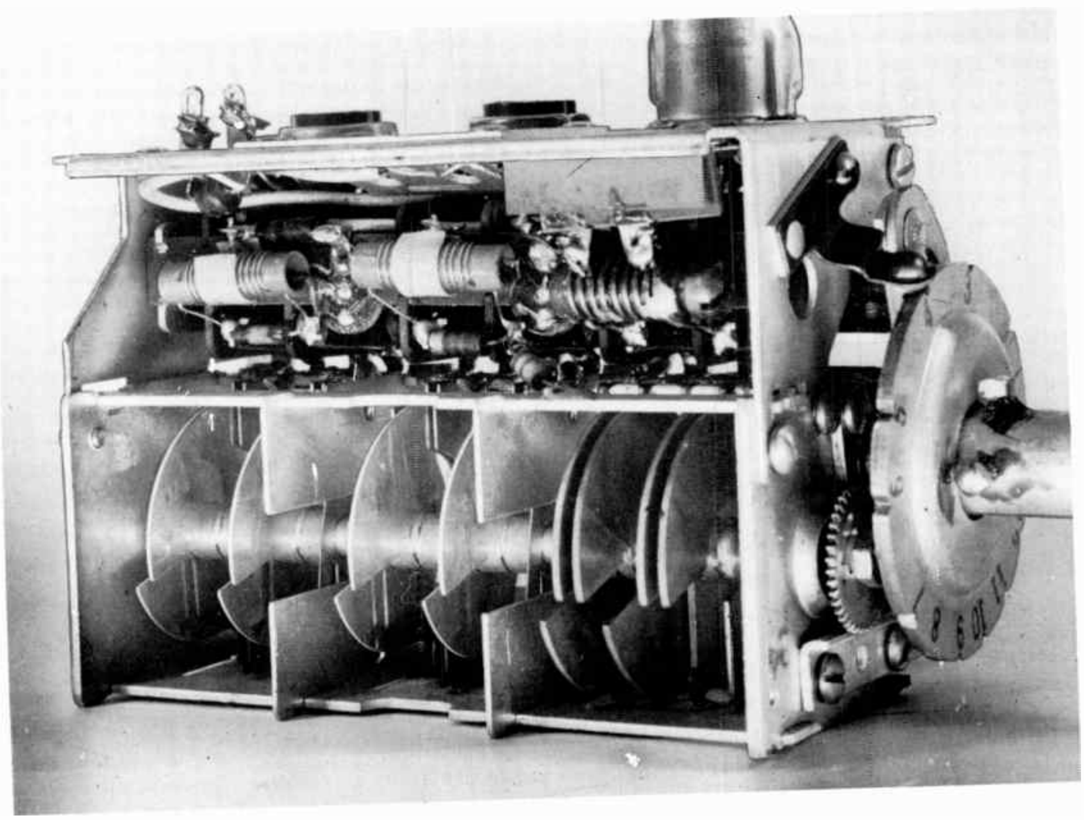


Fig. 44-7. Variable capacitors and other circuit elements of a tuner with which channel selection is by change of capacitance.

by arrows through their symbols. All of them are trimmer capacitors used during alignment of this tuner. In the circuit for the r-f plate there are two trimmers, one in the low-band side and the other in the high-band side. Similarly there are low-band and high-band trimmers in the mixer grid circuit and in the oscillator tuning circuits.

Fig. 44-7 is a picture of a television tuner operating with variable capacitors. The several tuning capacitors are in the lower shielded compartments. Directly above are the inductors and resistors which complete the circuits. The tube shelf is at the top of the picture. At the right-hand end of this tuner is the mechanism driven from the selector dial for turning the shaft and rotor plates of the variable capacitors. The trimmer capacitors are on the far side of this tuner, the side which is not visible in the photograph. The tuner pictured here is not the one for which a circuit diagram is shown by Fig. 44-6. The unit in the photograph employs push-pull circuits, whose action we shall examine in later pages.

PUSH-BUTTON TUNERS. There are several television tuners with which any one channel may be selected when the operator presses the appropriate one of a series of push buttons which take the place of the more common selector dial on the front panel. Fig. 44-8 is a circuit diagram for push-button tuning of

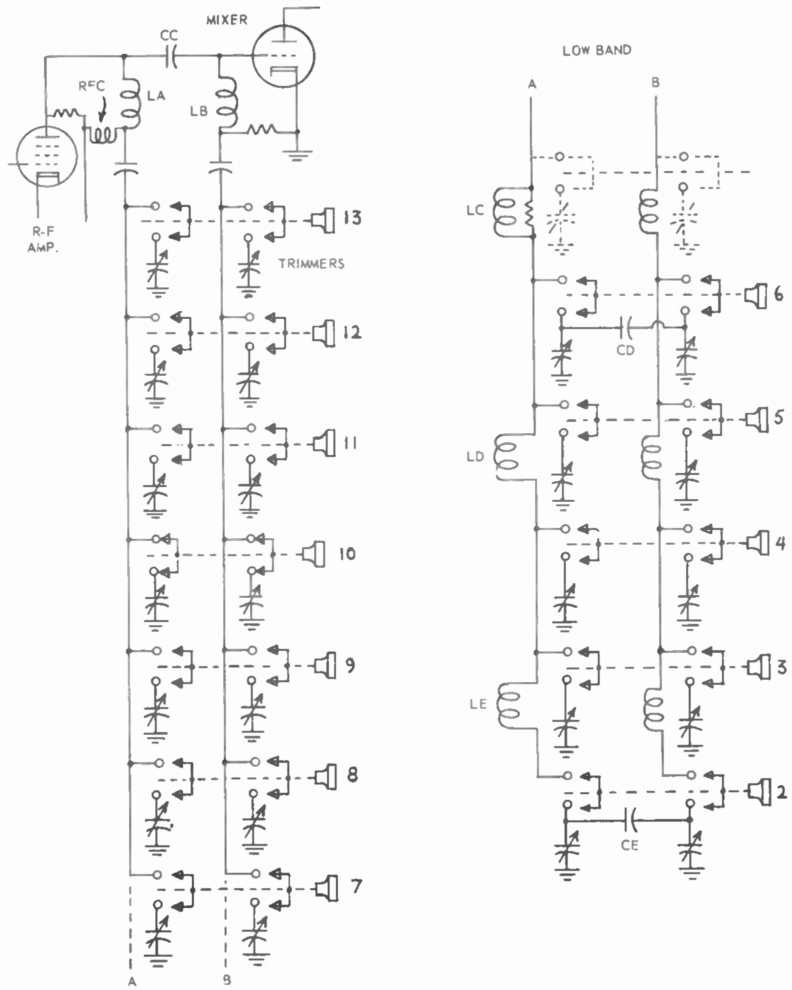


Fig. 44-8. Connections of r-f to mixer coupling in a tuner using push-button switching of capacitors for channel selection.

the coupling between an r-f amplifier plate and the grid of a mixer. A similar series of push-button switches would be used for tuning the r-f oscillator.

High-band connections are shown at the left in the figure, and low-band connections are at the right. Each of the numbered push-buttons operates all the switches for one channel, which would include the two switches shown for r-f amplifier and mixer, and also a third similar switch for the oscillator. When a switch is closed it connects to the tube circuit one side of an adjustable trimmer capacitor whose other end is grounded. Inductors *La* and *Lb*, also an oscillator inductor in a complete system, then become resonant at a frequency determined by their inductance and the capacitance of the trimmer. The trimmers are adjusted during the process of alignment to provide resonance at the frequencies which are correct for each channel. The switch sections for tuning channel 10 are shown in their closed positions. All other switch sections are shown in their open positions.

In some push-button tuners the low-band switching sections would be a continuation of the high-band sections. That is, the points marked *a* and *b* at the bottom of the left-hand string would connect to the similarly lettered lines at the top of the string on the right. In other tuners of this general type there are separate tubes or twin tubes for r-f amplifier, mixer, and r-f oscillator, with one tube or one part of a tube used for high-band reception and the other tube or part used for low-band reception. Then the high-band switching would be connected to one set of tubes, as in the left-hand part of Fig. 44-8, and the lines at the top of the low-band switching units would be connected to the second set of tubes. Then, in effect, we have separate tuners for the two bands.

In the low-band tuning section there are features not found in the high-band section. There are additional inductors between channels 7 and 6, as at *Lc*. These are necessary to bridge the frequency gap between the two bands. Additional inductors are used also between channels 5 and 4, as shown at *Ld*, where there is a considerable change of frequency. Extra inductors may be used also at *Le* between channels 3 and 2 where a rather great inductance is needed because of the relatively low frequencies at which tuning is being carried out.

Any of the tuning inductors may have band broadening resistors in parallel. This is shown at inductor *Lc*, but may be found also anywhere else on the tuning string as well as on the inductors directly connected to the tube circuits at the top of the left-hand part of the figure. In addition to the coupling capacitor *Cc* between the r-f plate and the mixer grid, coupling capacitance may be added as at *Cd* and *Ce* in the low band switching connections.

Instead of capacitance tuning as shown by Fig. 44-8 it is possible to operate movable cores in tuning inductors by means of push buttons. In one such tuner the push buttons have adjustable stops which allow coil cores to be moved to certain definite positions as each button is operated. There is one coil with its movable core for each tuned circuit. Alignment consists of setting the adjustable stops for each button and of adjusting the trimmer capacitors which are connected across each of the tuning inductors.

In a recent paragraph it was mentioned that some push-button tuners are designed to operate as two independent units, one for each band. There are a number of other designs employing separate tuning mechanisms and either separate tubes or else twin tubes for tuning in the low and high television bands. Some double tuners are made with movable cores working in the two sets of inductors for the two bands. There are others made with variable tuning capacitors, one three-gang capacitor being used for the low band and a second three-gang capacitor for the high band. All the movable cores or all the variable capacitors are operated from a single channel selector dial.

R-F OSCILLATORS. Although all parts of the tuner must operate correctly to have acceptable pictures and sound, it is the r-f oscillator which usually gives the service technician the most concern. For this reason we shall now examine certain features which are of especial interest.

For one thing, we must keep in mind that it is the purpose of the oscillator to produce a single sharply defined or sharply peaked frequency and not a band of frequencies. A sharply defined oscillator frequency will beat with the received carrier frequencies to form intermediate frequencies corresponding exactly to the signal modulation. Were the oscillator to produce even a narrow band of frequencies every signal frequency would form a corresponding band of frequencies, and the i-f output from the mixer would be far from satisfactory. For this reason we find oscillator circuits designed for small energy losses or designed to have high Q-factors.

We must not forget that the same oscillator frequency beats with both the video carrier and sound carrier frequencies and their modulations. The carrier frequencies have certain separations. By using the same oscillator frequency with both carriers we produce video and sound intermediate frequencies having the same separation as in the carriers.

In nearly all television receivers the oscillator frequency is tuned above the received carrier frequencies by the amount of the intermediate frequency. Then the oscillator frequency is equal to the sum of the video carrier and video intermediate frequencies, and also to the sum of the sound carrier and the sound intermediate frequencies. Here is an example for channel 4. Any other channel and any other intermediate frequencies might be used in similar examples.

	Video	Sound
Carriers (channel 4)	67.25 mc	71.75 mc
Intermediate frequencies	26.10	21.60
Oscillator frequency (the sum)	93.35 mc	93.35 mc

It is important to note this fact: The video carrier frequency is lower than the sound carrierfrequency, yet the video intermediate frequency is higher than the sound intermediate frequency. This happens be-

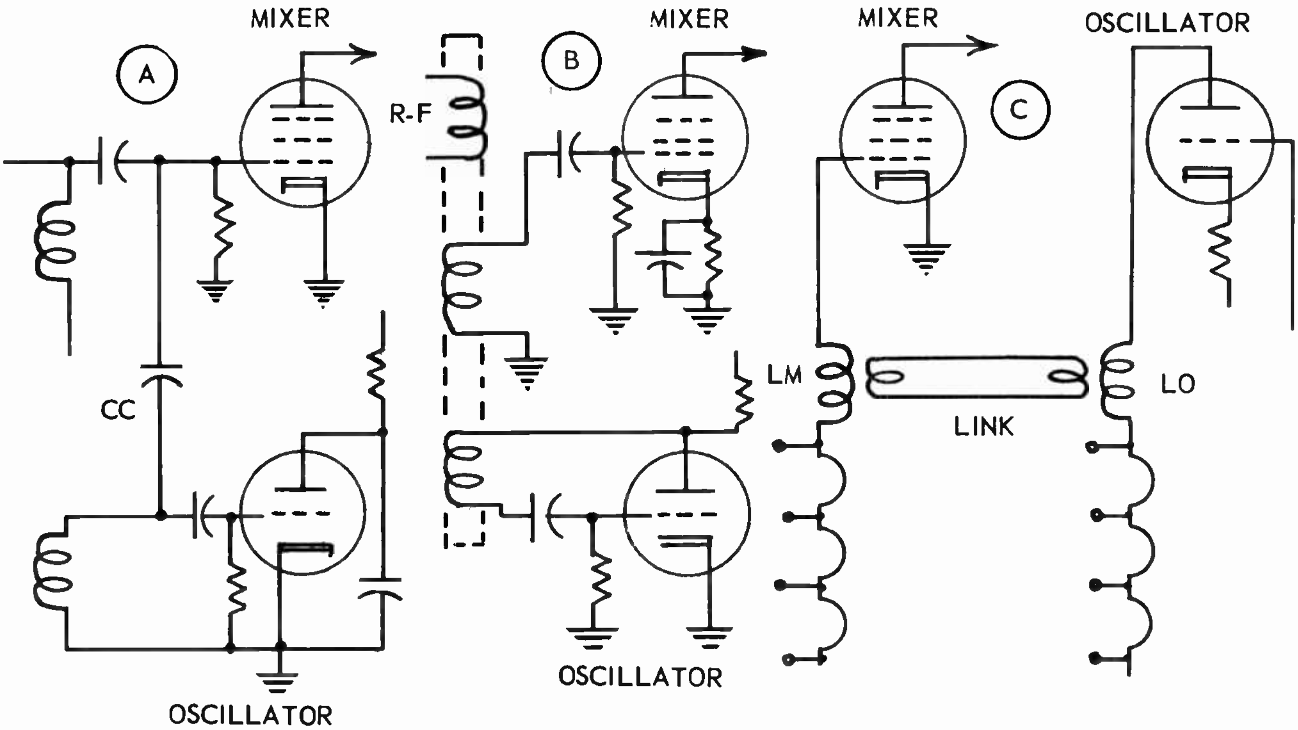


Fig. 44-9. Methods of injecting voltage at oscillator frequency into the grid of the mixer.

cause whenever you subtract lower and higher numbers from the same number the differences come out reversed; the lower is higher and the higher is lower. Here is an example in which we subtract two carrier frequencies from the same oscillator frequency to determine the intermediate frequencies.

	Video	Sound
Oscillator frequency	201.00 mc	201.00 mc
Carriers (channel 7)	175.25	179.75
Intermediate frequencies	25.75	21.25

There are various ways of feeding oscillator voltage, at the oscillator frequency, into the grid-cathode circuit of the mixer tube. The most common method, as shown in principle at *A* in Fig. 44-9, is by means of a small capacitor, *Cc*, connected from the oscillator tuned circuit to the mixer grid. The capacitance of *Cc* ranges from less than one mmf to as much as 4 mmf, but most often is between 1 and 2 mmf. This method of feeding oscillator frequency to the mixer may be used not only with the tube circuits illustrated, but with practically any other oscillator and mixer circuits.

Another fairly common method is shown by diagram *B*. Here the coupling or tuning coil in the mixer grid circuit is wound on the same supporting form as the oscillator coil. Voltage is transferred from oscillator to mixer by the inductive coupling between the two coils. The diagram shows the form as carrying also a coil which is connected in the plate circuit of the r-f amplifier.

A third method is illustrated by diagram *C*. In the mixer grid circuit and in the oscillator plate circuit are tuning inductors *Lm* and *Lo*. Between them is a link coupling. The link consists of two small coils or loops connected together through conductors. Inductive coupling from coil *Lo* to the small coil on one end of the link, induces emf at the oscillator frequency in the link circuit. Link current at the oscillator frequency induces emf of this frequency in the mixer grid coil *Lm*.

At the high frequencies of oscillator and carrier currents the fields around almost any conductors in the respective circuits will cause transfer of oscillator voltage provided the conductors are fairly close together. There is always some coupling and voltage transfer between any tuned circuits of oscillator and mixer unless these circuits are well separated or are carefully shielded from each other.

When the r-f oscillator and the mixer tubes are the two sections of a twin-triode having a single cathode for both functions, both plate currents flow in the cathode line. Whatever inductance that may be in this line has enough inductive reactance to form an impedance coupling, or any resistance in the cathode line forms resistance coupling between oscillator and mixer.

FREQUENCY DRIFT IN OSCILLATORS. When there is variation of the frequency supplied by the r-f oscillator, the beating of this changed frequency with the constant carrier frequencies will cause corresponding variations in the video and sound intermediate frequencies. Since the i-f amplifier circuits are aligned for best response at fixed frequencies there will be trouble with both picture and sound reproduction when there is variation of oscillator frequency. This trouble is especially bad in receivers using what are called dual or split sound systems, where operation is at the sound intermediate frequency and where sound takeoff is ahead of the video detector. The passband of these sound systems is only a few hundred kilocycles at most, and small variations of oscillator frequency will throw the actual sound intermediate frequency completely outside the passband. The result is a picture without accompanying sound.

10) Variation of oscillator frequency is called frequency drift. There are four principal causes for drift. First are changes of temperature as the receiver warms up. Second are changes of voltages at elements of the oscillator tube. Third are variations in load on the oscillator, which are not very likely to occur during normal operation. Fourth comes mechanical vibration or shock.

Change of temperature, with resulting expansion of all heated parts, alters the internal capacitances of the oscillator tube and also the capacitance of tuning capacitors and the stray capacitances of wiring and parts. Expansion of wire and supporting forms of coils will change the tuning inductance.

1) Among the many ways of lessening the effects of temperature change, one is to use as part of the tuning capacitance a fixed capacitor having negative temperature coefficient of capacitance, which is connected across all or part of the tuning inductance, or is connected in series with other tuning capacitors. These compensating capacitors drop their capacitance with rise of temperature, counteracting the lowering of frequency which is the usual result of temperature rise. Compensating capacitors should be located where their temperature will change at about the same rate as other parts of the oscillator circuit. Should you replace one of these capacitors with an uncompensated type the result will be excessive oscillator drift.

Tube heating is lessened by designs in which there is relatively small plate current, as may be obtained with highly negative bias through grid leaks of higher resistance. Heating of circuit parts is reduced

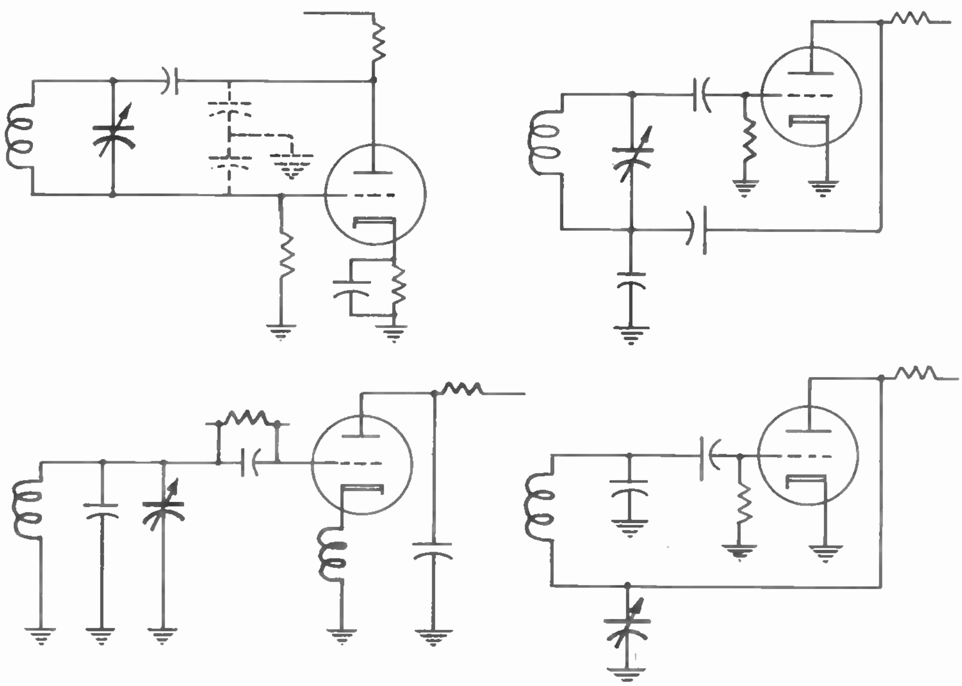


Fig. 44-10. Some of the possible locations for fine tuning capacitors.

by using large conductors, and by employing high-Q construction to lessen the transfer of energy into heat. Insulating supports are made of material having a small temperature coefficient of expansion, such as the high-grade ceramics. There must be good ventilation. The oscillator tube, and other tubes, usually are mounted above other circuit parts so that heat rising from the tubes will not raise the temperature of these parts.

5 Oscillator circuits often are designed with relatively large tuning capacitance and small inductance. Then changes of capacitance in the tube, wiring, and other parts make a relatively small percentage change in the total tuning capacitance, and a correspondingly small change of oscillator frequency.

Any change of d-c voltage at the plate, screen, or grid of the oscillator tube will alter the effective plate resistance, the transconductance, and the frequency. D-c voltages are held nearly constant by good filtering at the power supply and also in the leads to the tube elements. A few power supplies are constructed with voltage regulating tubes or voltage regulating power transformers, although this is not common practice in receivers.

Sudden shocks or continued vibration of parts will alter the spacings between elements in the oscillator tube, will alter the internal capacitances, and will change the frequency in time with the vibration or shocks. Parts of the tuner which are outside the oscillator tube are similarly affected, and stray capacitances then vary to change the frequency. These effects are lessened by good solid construction and firm supports, which accounts for the seemingly excessive mechanical strength in tuner frames, shields, and supports. Vibration of the oscillator tube often is held down to slow rates by enclosing this tube in a shield that grips the envelope and clamps onto the tuner frame, or by using a shield with a heavy lead liner.

FINE TUNING CONTROLS. A fine tuning control or sharp tuning control consists of a small variable capacitor connected into the r-f oscillator tuned circuit in such manner as to allow slight variations of oscillator frequency. The fine tuning control is used by the operator of the receiver to correct any frequency drift which may occur in the oscillator. When used for this purpose the control requires changing chiefly during the warm up period.

The fine tuning control is used also to compensate for slight discrepencies in oscillator alignments for different channels. When oscillator tuning adjustments have not been set for the exact frequency required in each channel the reception may be improved by operation of the fine tuning control. When used for this purpose the control is changed when switching from one channel to another.

Fig. 44-10 shows a few of the many possible connections for fine tuning capacitors in oscillator circuits. The control capacitor is shown in each case by a symbol with an arrow through it. The capacitance at minimum setting of the control usually is something between a small fraction of a mmf and about two mmf. At maximum the capacitance ranges from about two to five or more mmf.

7 Moving the fine tuning control from end to end of its capacitance range makes the least change of oscillator frequency in the lowest-frequency channel, number 2, and makes the greatest change of oscillator frequency in the highest-frequency channel, number 13. Some controls are designed for total frequency variation of less than one megacycle, while others will change the oscillator frequency by two to four megacycles. When a receiver uses a dual or split sound system the fine tuning control must be accurately adjusted in order to have satisfactory reception of both picture and sound.

The fine tuning control usually is operated by a shaft that is concentric with the shaft for the channel selector. Either of the shafts may be the solid member at the center, with the other a tubular member around the outside. The knobs for channel selection and for fine tuning then are together on the panel ends of their respective shafts.

Fine tuning capacitors seldom are the conventional variable types with rotor and stator plates semi-circular and meshing together. Often the capacitor consists of a tube of insulating dielectric with the stationary metallic plate around the outside and an inner metallic plate that is moved end wise by the control shaft turning a threaded screw. Sometimes the two capacitor plates are disc-shaped, with one moved toward or away from the other by screw threads turned with the control shaft. In still other designs the two plates are stationary with a sheet of solid dielectric material moved into and out of the space between the plates. In a number of receivers or tuners which have separate alignment adjustments for oscillator frequency on each channel there is no fine tuning control for use by the operator.

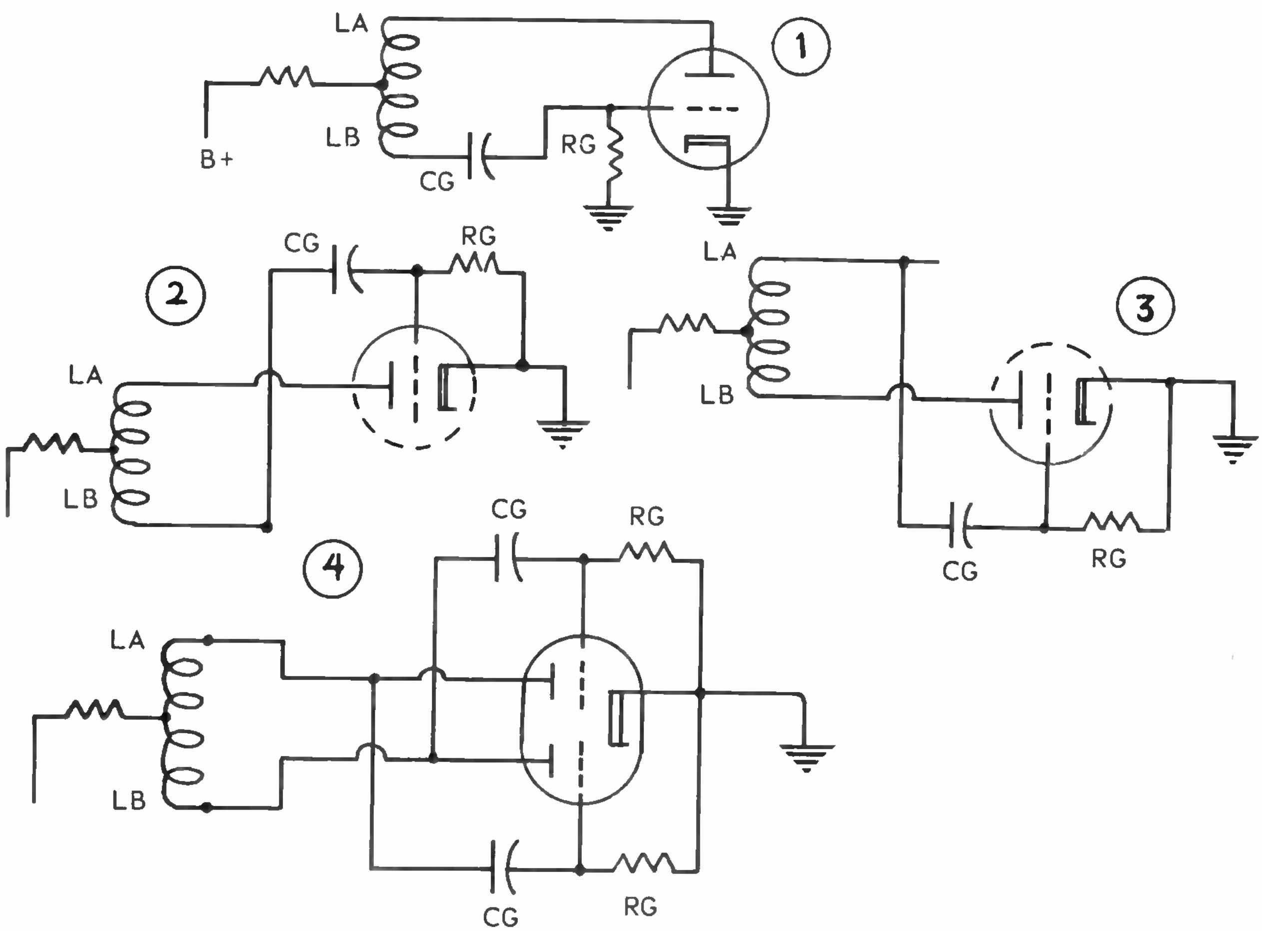


Fig. 44-11. Two Colpitts circuits combined as a push-pull oscillator on a twin-triode.

PUSH-PULL OSCILLATORS. The r-f oscillators in quite a few tuners consist of two Colpitts circuits combined with a twin-triode tube in a manner which lessens the effective tube capacitances, allows each section of the tube to act as a sort of amplifier for the other section, and in general, provides quite stable oscillation at the high frequencies used for television. This arrangement is called a push-pull oscillator, because while one section "pushes" on the electron flow the other section "pulls" the flow after oscillation commences.

The development of a push-pull oscillator circuit is illustrated by Fig. 44-11. In diagram *1* we have an ordinary Colpitts oscillator with which tube and circuit capacitances are used for tuning. The only new feature is center-tapping of the inductance coil, with B-voltage applied through the tap.

Without making any electrical changes in connections, but by altering only the appearance of the diagram, we may redraw the Colpitts oscillator as in diagram *2.* Again with no electrical changes we may redraw the original circuit as at *3.* The same inductance coil appears in both these latter diagrams. The plates and grids of the two diagrams are the elements in the two sections of a twin-triode which has a single cathode.

By putting together the diagrams shown at *2* and *3* we have the connections of diagram *4.* This is our push-pull oscillator. Fig. 44-12 is a circuit diagram for a television tuner in which the oscillator is of the push-pull type. Connections are essentially the same as just described except that there is an added biasing resistor between the oscillator cathode and ground, and a fine tuning capacitor is shown connected to

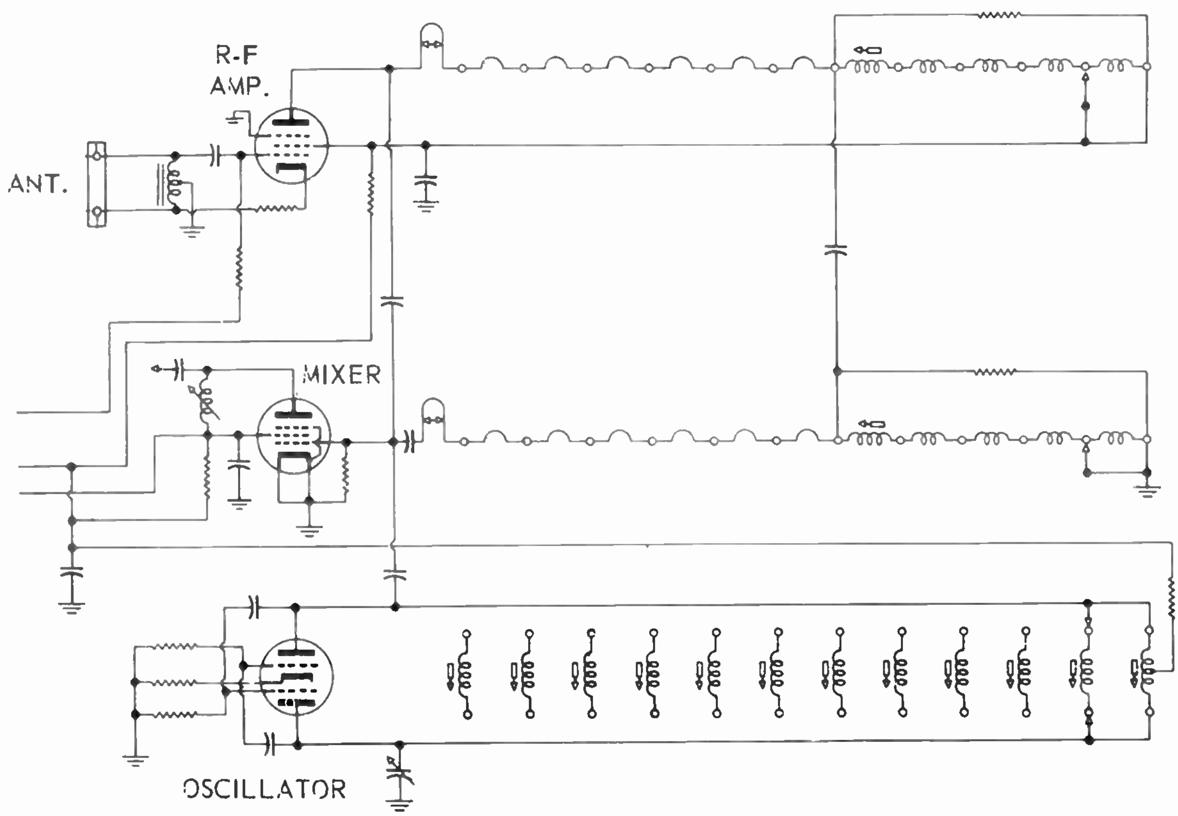


Fig. 44-12. Circuit diagram for a tuner in which the oscillator is a push-pull type.

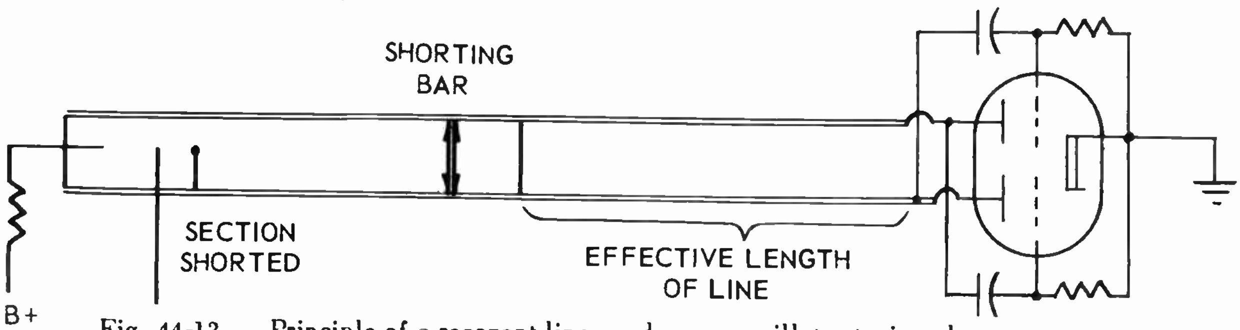


Fig. 44-13. Principle of a resonant line used as an oscillator tuning element.

the bottom oscillator line. The center-tapped coil for channel 2, at the extreme right, remains in the circuit at all times. Contacts on the rotors of two selector switch wafers connect the other coils successively in parallel with the channel 2 coil for tuning the remaining channels from 3 through 13. The r-f plate and mixer grid circuits are tuned by shorting contacts on two other switch wafers.

RESONANT LINE OSCILLATORS. There is still another way of tuning television r-f oscillators, we may use a quarter-wave shorted resonant line. A shorted resonant line consists of two parallel conductors which are shorted together at one end and connected to some radio or television circuit at the other end. If the lengths of the two conductors are made equal to one-quarter of a wavelength of radiation in space, the line will behave exactly like a parallel resonant circuit at the frequency corresponding to the wavelength. The subject of resonant lines will be discussed at more length in a later lesson. It is possible to take short circuited and open circuited conductors of certain lengths and have them behave like parallel resonant circuits or series resonant circuits at certain frequencies, and like either capacitance or inductance at other frequencies.

Fig. 44-13 illustrates the principle of an oscillator tuned by means of a quarter-wave shorted line. At the right is a twin-triode oscillator tube with the same grid and cathode connections as in the final diagram, number *4* of Fig. 44-11. But instead of having a tuning coil connected between the plates there are two long parallel conductors with a shorting bar which is a conductor arranged to move along the conductors while maintaining electrical contact with both of them. The effective length of this resonant line is the distance from the plates to the shorting bar. The remainder of the line, at the left, is simply shorted out of the circuit and has no effect on performance.

If the effective length of the resonant line is made about 13.9 inches, by moving the shorting bar, the line is equivalent to a parallel resonant circuit which is tuned to the mid-frequency of channel 13. Moving the bar farther from the plates will lower the resonant frequency. With an effective length of about 16.7 inches the line would be resonant at the mid-frequency of channel 7. A length of about 34.8 inches would be resonant for channel 6, and a length of about 51.8 inches would tune for channel *2*.

The lengths of line for tuning in the low-band channels are prohibitively great for a television tuner. So we replace the straight lines with suitable coiled conductors whose electrical lengths are those required for the various frequencies. Then we have an oscillator system whose connections are as shown at the bottom of Fig. 44-14, which is a circuit diagram for a tuner using shorted resonant lines for not only the oscillator but also for the r-f amplifier plate circuit, at the top, and the mixer grid circuit, through the center.

The shorting bars of this tuner are shown all the way to the right, in position for tuning channel *2*. These bars really are the movable contacts on the rotors of two wafers in a rotary selector switch. The

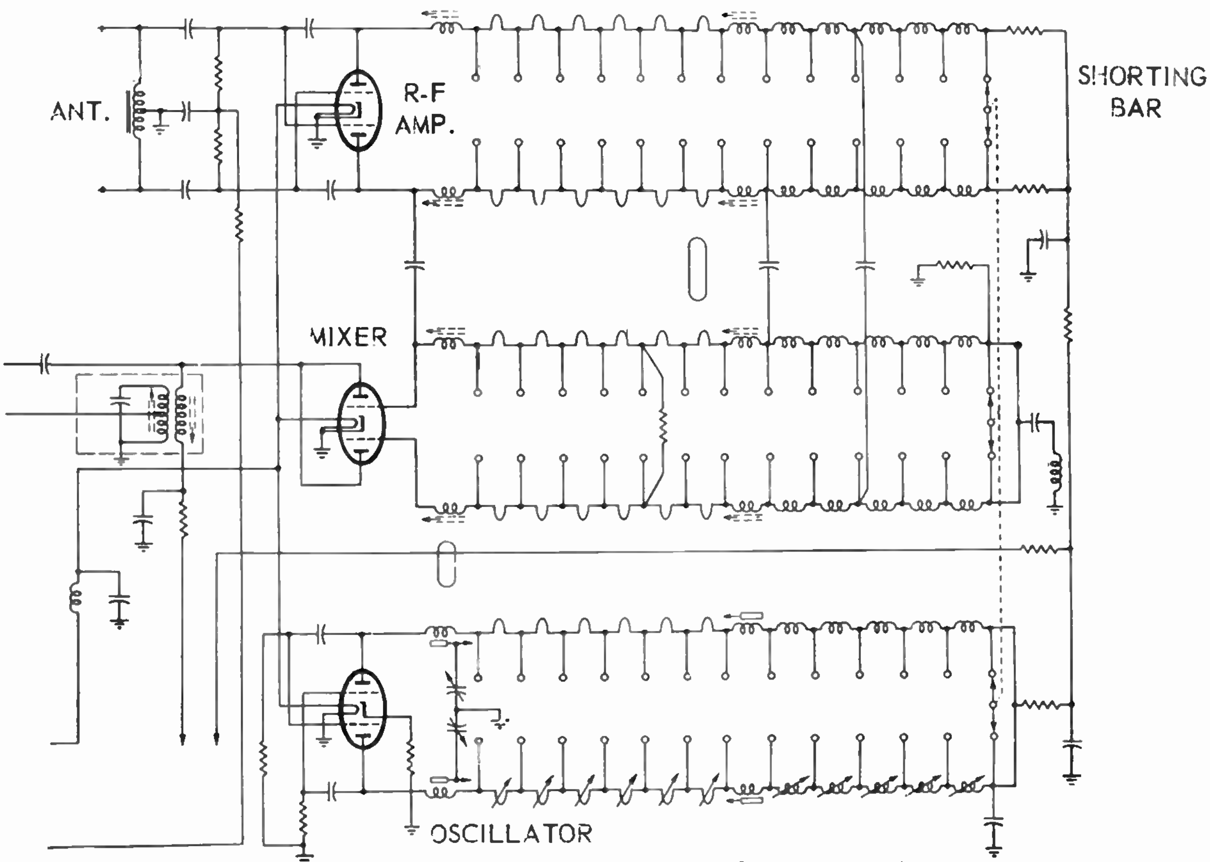


Fig. 44-14. Circuits of a tuner in which quarter-wavelength shorted resonant lines are used for tuning the r-f plate, mixer grid, and oscillator circuits.

sections of resonant line for low-band channels are small coils, with larger coils (of greater electrical length) between the positions for channels 6 and 7. Line sections for high-band channels are straight or bent wires, depending on the required changes of electrical length for each successive channel. Between the oscillator plates and the switch positions for channel 13 are coils or loops whose effective length is adjusted for tuning channel 13 during the work of alignment. Alignment adjustment is provided also for the coils between positions for channels 6 and 7, just as with many other tuners.

This completes our examination of the principal features, both electrical and mechanical, of television tuners which are in general use. There are minor variations which have not been shown, but by now you should have enough understanding of design and performance in general that such matters will cause no difficulties. As an example of what you may find, there are turret tuners and also selector switch tuners designed to handle only seven or eight channels instead of all twelve. With turret tuners of this class it is necessary to insert inductor strips suited for the particular channels which are used in your locality. With selector switch tuners of this class each alignment adjustment may be varied over a frequency range wide enough to suit either of two channels. You then make the alignments for whatever channels can be received where the receiver is used.

We are not yet ready to undertake the alignment of television receivers, not even the tuners. First it is necessary to become acquainted with all the parts and circuits which follow the tuner, so that you will appreciate the dependence of each one on all the others. In the following lesson we shall proceed to the i-f amplifiers, which come right after the tuner as we follow the signal.

DO NOT TEAR — CUT ALONG THIS LINE AND SEND IN FOR GRADING.

TELEVISION

LESSON NO. 45

TELEVISION I-F AMPLIFIERS

"Training Your Mind"
C. T. I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind

C.T.I.

Perfecting Your Skill

COMMERCIAL

TRADES

INSTITUTE

Chicago, Ill.

LESSON NO. 45

TELEVISION I-F AMPLIFIERS

The television i-f amplifier consists of two or more pentode tubes with their necessary interstage couplers, all connected between the plate of the mixer and the input to the video detector diode.

Fig. 45-1 is a picture of a typical string of four i-f amplifier tubes and couplers as they would appear when removed from a chassis and arranged in order between a tuner and a video detector tube. The tubes from left to right on the tuner are the r-f oscillator, the mixer, and the r-f amplifier. To the plate of the mixer would be connected the coupler which is placed in front of the tuner and slightly to its right. This coupler feeds into the first i-f amplifier tube.

The output of the first i-f amplifier goes to a second coupler. After this second coupler comes the second i-f amplifier tube, which is followed by another coupler mounted within a shielding can. This third coupler feeds into the third i-f amplifier tube. The output of this amplifier goes to the next coupler, which is a small unshielded type. Then follow the fourth i-f amplifier tube, another small unshielded coupler, and finally the video detector tube at the extreme right.

The i-f amplifier tubes and intervening couplers most often are arranged in a fairly straight line on the receiver chassis. This line of tubes and couplers will commence at or near the tuner and will extend toward the rear of the chassis. Quite often the first coupler or transformer, the one connected between mixer and first i-f amplifier, is mounted on the tuner.

The parts which we call couplers, when used between successive tubes in an amplifying system, often are called transformers. We call them couplers because a coupler is any device which transfers signal energy by any means from one tube or circuit to another tube or circuit. Couplers may operate with inductive, capacitive, or resistive coupling, or with combinations of these methods. Some couplers have a single winding or coil, others have two or three windings, and still others have no windings at all, but have resistors, capacitors, or both.

A transformer is a device having two or more windings with a single or common magnetic circuit. Transformers are couplers utilizing mutual induction. All transformers are couplers, when used between tubes or tube circuits, but all couplers are not transformers – just as all cows are animals, but all animals are not cows. In spite of this it is rather common practice to apply the name transformer to any kind of coupler.

SOUND TAKEOFF POINTS. At *A* in Fig. 45-2 the takeoff for sound i-f signals is shown as immediately following the mixer tube. All the i-f stages then are designed for or intended to handle only the picture

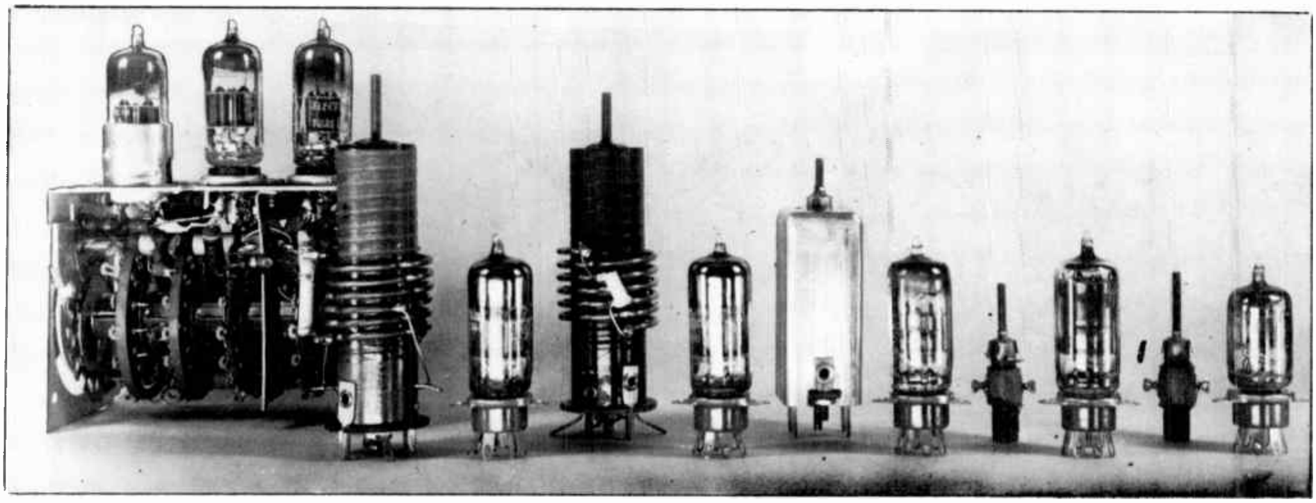


Fig. 45-1. Couplers and tubes of an i-f amplifier system as they are used between the tuner and the video detector.

and sync signals, or only the video signals. The entire amplifier shown here may be called a video i-f amplifier, as distinguished from the sound i-f amplifier which receives sound i-f signals from the mixer.

In diagram *B* both video i-f and sound i-f signals go from the mixer through the first i-f amplifier. Then the sound signals are taken off and fed to the sound section, while picture and sync signals (video signals) go through the remainder of the i-f amplifier to the video detector. In some receivers the sound i-f signals and video i-f signals go through more than one of the i-f amplifiers before coming to the sound takeoff.

The amplifiers and takeoffs illustrated at *A* and *B* are called dual sound systems or split sound systems. The sound i-f amplifiers of receivers using this system operate at the sound intermediate frequency, which is the difference between the r-f oscillator frequency and the sound carrier frequency.

In diagram *C* of Fig. 45-2 both kinds of i-f signals go from the mixer through all the i-f amplifiers and into the video detector. The two i-f frequencies beat together in the video detector, which is acting like a second mixer at the same time that it acts as a detector. The result is a beat frequency centered at 4.5 megacycles. The beat always is of this exact frequency because video and sound intermediate center frequencies always are exactly 4.5 megacycles apart, just as are the video and sound carriers. This is called an intercarrier sound system.

The 4.5 mc beat frequency carries the frequency modulation of the sound signal. This modulated f-m signal is taken off at some point following the video detector and is fed to the sound section. The i-f amplifying portion of the sound section then operates at a center frequency of 4.5 mc, not at the sound intermediate frequency which is the difference between oscillator and carrier frequencies. The intercarrier sound system is used in most makes and models of recently designed receivers.

VIDEO I-F RESPONSE. In an earlier lesson we learned that the video intermediate frequency is higher than the sound intermediate frequency, although the video carrier frequency is lower than the sound carrier frequency. This always is true when the r-f oscillator frequency is higher than received carrier frequen-

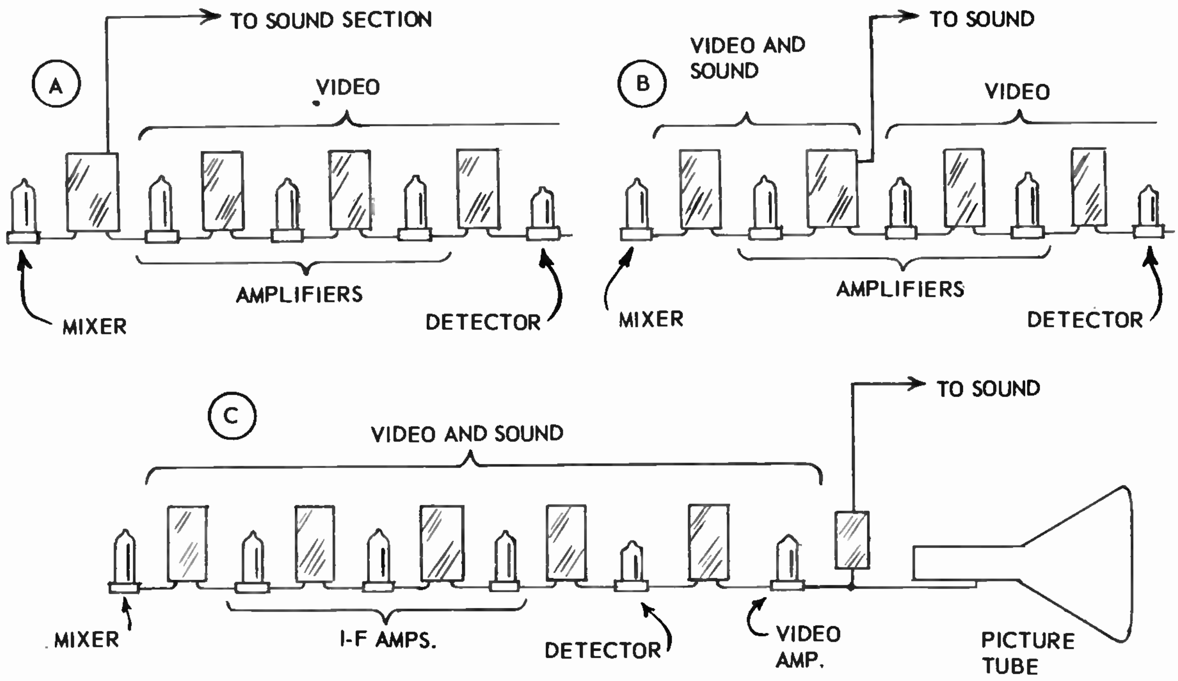


Fig. 45-2. Various points of sound takeoff.

cies. However, were the oscillator frequency to be lower than the carrier frequencies, then the oscillator frequency would be subtracted from the carrier frequencies to arrive at the intermediate, and the sound intermediate frequency would be higher than the video intermediate.

We learned also that the i-f amplifier must have a frequency response that compensates for unequal strengths at which low and high modulation frequencies are transmitted by the vestigial sideband method. When first we discussed vestigial sideband transmission and the form of video i-f amplifier response required to receive such transmission, the ideal video i-f response was shown as at *A* in Fig. 45-3. One side of this response is sloped, and the video intermediate frequency is located on this sloped side at a point where the gain or amplification is at 50 per cent of maximum.

When doing service work you often will see the video i-f response traced out on the screen of an oscilloscope. The form of the response curve will not be exactly like the ideal, because commercial amplifier circuits cannot produce this ideal response of gain versus frequency. The actual form will be more like the curve shown at *B*. On the curves at *A* and *B* the video intermediate frequency is to the left of the sound intermediate frequency. The video intermediate is a higher frequency than the sound intermediate, so in these two curves the frequency is being shown as increasing from right to left.

Most oscilloscopes are so designed that their traces show frequency increasing from left to right. Consequently, the trace on the oscilloscope screen usually would appear as at *C*. This is the same response as at *B*, but is reversed with respect to right and left, or with respect to frequency.

The response curves at *B* and *C* are "right side up", meaning that maximum gain is at the top and zero gain at the bottom. With some oscilloscopes and with some connections to receiver circuits the traces will be "upside down", with zero gain at the top and maximum gain at the bottom. Then the trace first shown at *B*, where frequency is increasing from right to left, will be changed to the one at *D*, where frequency still increases in the same direction. The trace first shown at *C*, with frequency increasing from left to right, will be changed to the one at *E*, where the direction of frequency increase is unchanged but the polarity is inverted. All curves of Fig. 45-3, except the one at *A*, show exactly the same frequency response.

In theory the video intermediate frequency should be exactly 50 per cent down on the i-f response. In actual practice it often is advantageous to make alignment adjustments which bring the video intermediate frequency higher up, even to a point where this frequency is only 25 per cent down or is at 75 per cent of maximum gain. A good many receivers give best performance with the video intermediate frequency at 55 to 60 per cent of peak gain. In no case should this frequency be below the 50 per cent point.

Incorrect adjustment of r-f oscillator frequency can displace both the video and the sound intermediate frequencies on the i-f response curve, and cause serious difficulties in reception. This is shown at the top of Fig. 45-4. In diagram *A* we have an i-f response of reasonably good form and have satisfactory placements of the two intermediate frequencies on this response. The two intermediates are separated by 4.5 mc, which is something you never can change.

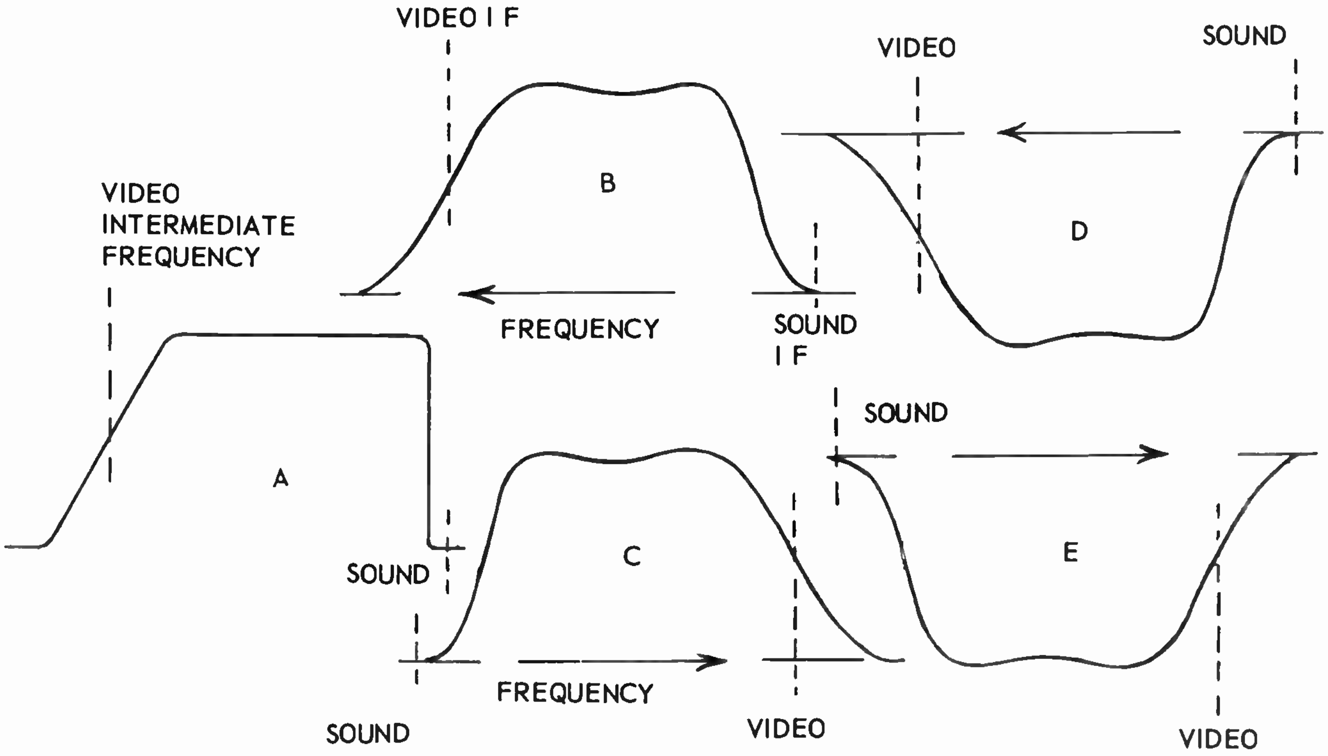


Fig. 45-3. How the overall response of an i-f amplifier may appear on the screen of an oscilloscope.

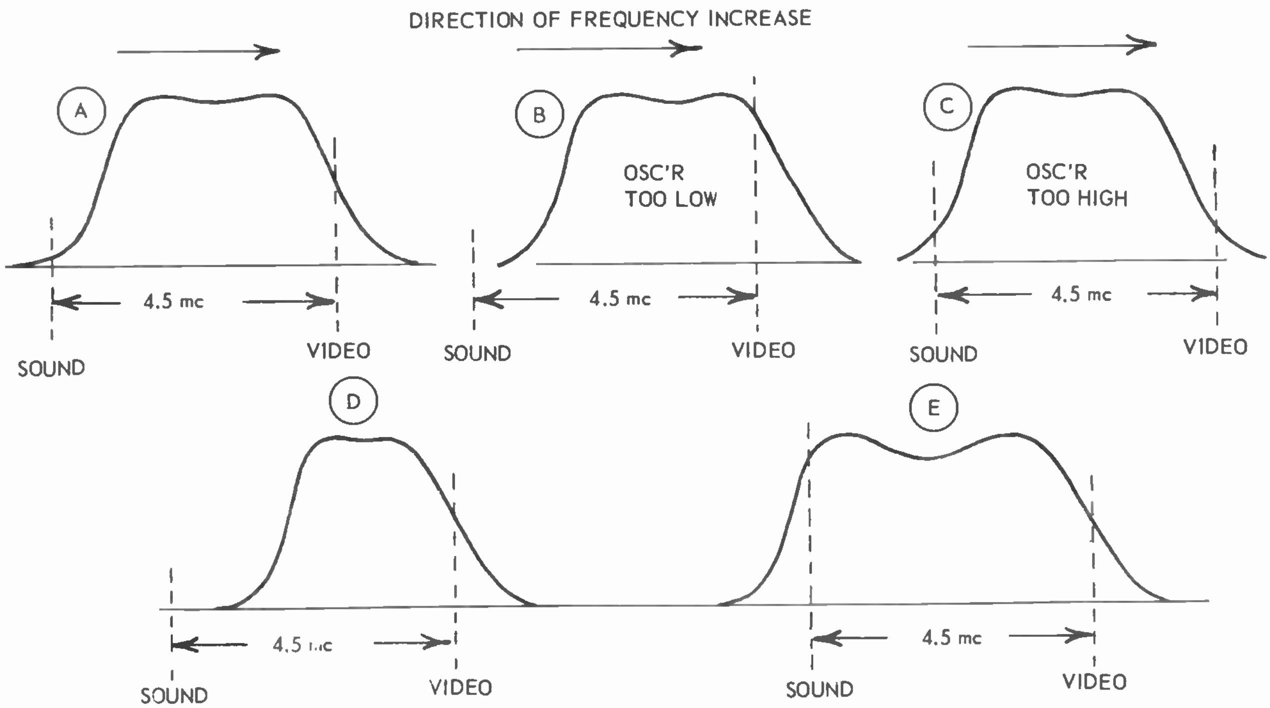


Fig. 45-4. Effects of incorrect adjustment of r-f oscillator frequency, and of incorrect shaping of the over-all response curve.

In diagram *B* the r-f oscillator frequency is too low. Since the same oscillator frequency beats with both the video and sound carriers, both the intermediates will be lowered in frequency. Lowering the video intermediate frequency brings it higher up on the response curve, and lowering the sound intermediate moves it clear off the response, or to a point where response is zero.

To determine the results of such lowering of the r-f oscillator frequency we must consider the relation of modulation frequencies to the video intermediate frequency. Low frequencies of modulation lie nearest the intermediate frequency, high modulation frequencies are farther from the intermediate. In the curve shown at *B* the highest frequencies of video modulation are at points where gain is very low or is zero, and these frequencies are not suitably amplified. The high video frequencies carry all the fine details of the pictures, so we now will have lack of detail or poor definition in the pictures. With a dual sound system there will be no sounds from the speaker, because the sound intermediate frequency is moved outside the pass-band of the sound i-f amplifier as well as being at zero gain. With an intercarrier sound system there will be no sound reproduction if the sound intermediate really is at a point of zero gain. However, with this type of sound system there is adequate reproduction in most sets even when the sound intermediate is at a point where gain is as little as two or three per cent of maximum, so unless the r-f oscillator is away off its frequency on the low side we still may have good sound.

Now look at curve *C* of Fig. 45-4. Here the r-f oscillator frequency is too high. The video intermediate is moved to a point of weak gain, and the sound intermediate is at a point of fairly high gain. We have weakened the amplification of low video frequencies, which lie close to the video intermediate. All our sync pulses are in this low-frequency region, so the picture may fail to remain synchronized and may move either vertically or horizontally on the picture tube screen. We have also dropped the amplification for the lower frequencies of modulation, which carry the bold outlines of the pictures. Consequently, the pictures will lack brilliance, they will appear dull and lifeless.

With the oscillator frequency too high we again will have no sound reproduction from a dual sound system, because the sound intermediate frequency again is outside the bandpass of the sound i-f amplifier. With intercarrier sound there will be a loud buzz from the speaker, for reasons which we shall talk about when coming to the subject of trouble shooting in general. With either type of sound system it may be possible for sound modulation to get to the picture tube, where it can produce dark horizontal bars across the pictures. These are called sound bars.

Troubles similar to those caused by wrong oscillator frequencies can happen when you or someone else aligns the i-f amplifier in a way to produce incorrect forms of response curves, as at the bottom of Fig. 45-4. At *D* the frequency bandpass has been made too narrow. Even though the oscillator is adjusted for correct placing of the video intermediate frequency, the sound intermediate is away beyond the limit of the curve, it is zero. At *E* the alignment has been such as to make the passband too wide. Now, with correct placement of the video intermediate frequency, the sound is far too high on the gain curve.

If you adjust the r-f oscillator to bring in the sound with a response as at *D*, the video intermediate will drop to zero gain, for separation always remains 4.5 mc. If you adjust the oscillator to lessen the sound gain with the curve at *E*, the video intermediate will come up to peak response, and there will be over-amplification of low modulating frequencies.

INTERMEDIATE FREQUENCIES. There are a number of different intermediate frequencies used in var-

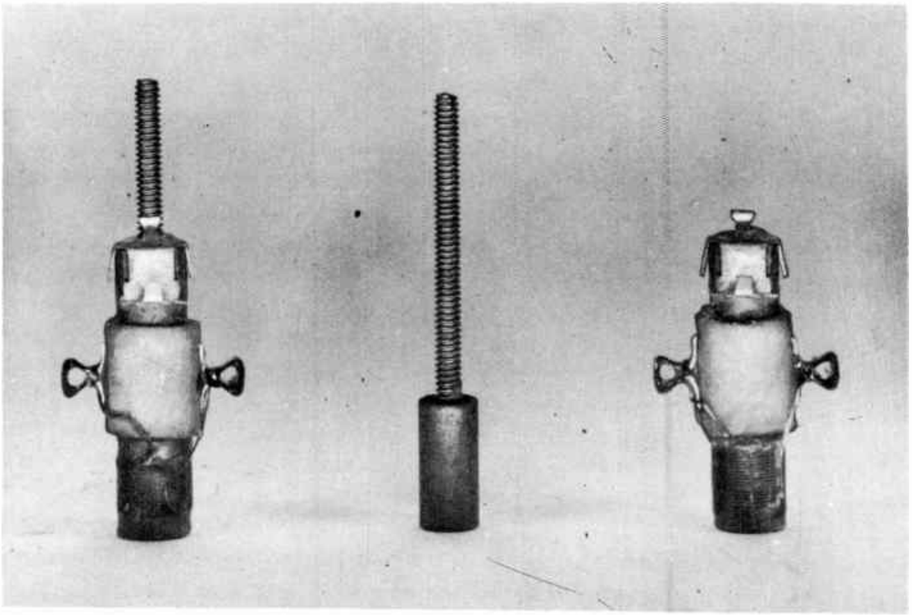


Fig. 45-5. Impedance couplers, one with its core taken out of the winding form.

ious makes and models of television receivers. The video and sound intermediate frequencies are specified for each receiver by its manufacturer. Correct response at these frequencies is obtained by alignment of the video i-f couplers to correctly shape the response curve of the amplifier system.

Intermediate frequencies originally are chosen to avoid, so far as is practicable, pickup of interference frequencies by the i-f amplifier circuits, and formation of the receiver intermediate frequencies by beating of the many kinds of interference signals that may come through the tuner. These interferences may result from transmissions by radio amateurs and by other radio services whose carrier frequencies are within the range commonly used for television intermediate frequencies. They may result also from image frequencies of television channels other than the one being received, or from f-m broadcast transmitters, or from oscillator radiation by other television receivers in the neighborhood, and from harmonics formed by many kinds of interference.

Sound and video intermediate frequencies which are in general use are listed by the accompanying table.

TELEVISION INTERMEDIATE FREQUENCIES

In Megacycles

SOUND	VIDEO	SOUND	VIDEO
21.25	25.75	22.1	26.6
21.3	25.8	22.25	26.75
21.6	26.1	31.5	36.0
21.75	26.25	31.63	36.13
21.9	26.4	32.8	37.3
22.0	26.5	41.25	45.75

TYPES OF I-F COUPLERS. The simplest coupler used in i-f amplifier systems is the tuned impedance type consisting of a single winding with a movable core. A complete coupler of this type is shown at the left in Fig. 45-5. At the center is one of the movable cores attached to a threaded stud or screw used for adjustment. At the right is a coupler winding and supporting form from which the core was removed.

A tuned impedance coupler may be connected between the mixer tube and the first i-f amplifier, or between two amplifiers, as shown by Fig. 45-6. In all diagrams except number *2* there is a blocking capacitor *Cc* which insulates the grid of the second tube from the high positive voltage in the plate circuit of the first tube. This blocking capacitor is of such value that its reactance at usual operating frequencies is less than 100 ohms. Consequently, so far as transfer of signal voltage or current from the first to second tube is concerned, this capacitor offers practically no opposition; it serves only to isolate the second grid and allow maintaining a negative bias while allowing signal currents to pass from tube to tube.

In diagram *1* the coupler is connected between the plate of the first tube and the source of B-plus voltage. Although there is no capacitor across the winding we nevertheless have here a paralleled resonant circuit acting as a plate load. The capacitance necessary for parallel resonance is furnished in part by the distributed capacitance of the winding. More is furnished by stray capacitance of wiring and tube sockets. The remainder is the internal output capacitance of the first tube and the internal input capacitance of the second tube. All these capacitances are represented by the broken-line symbols *Cp* and *Cg* of diagram *2.*

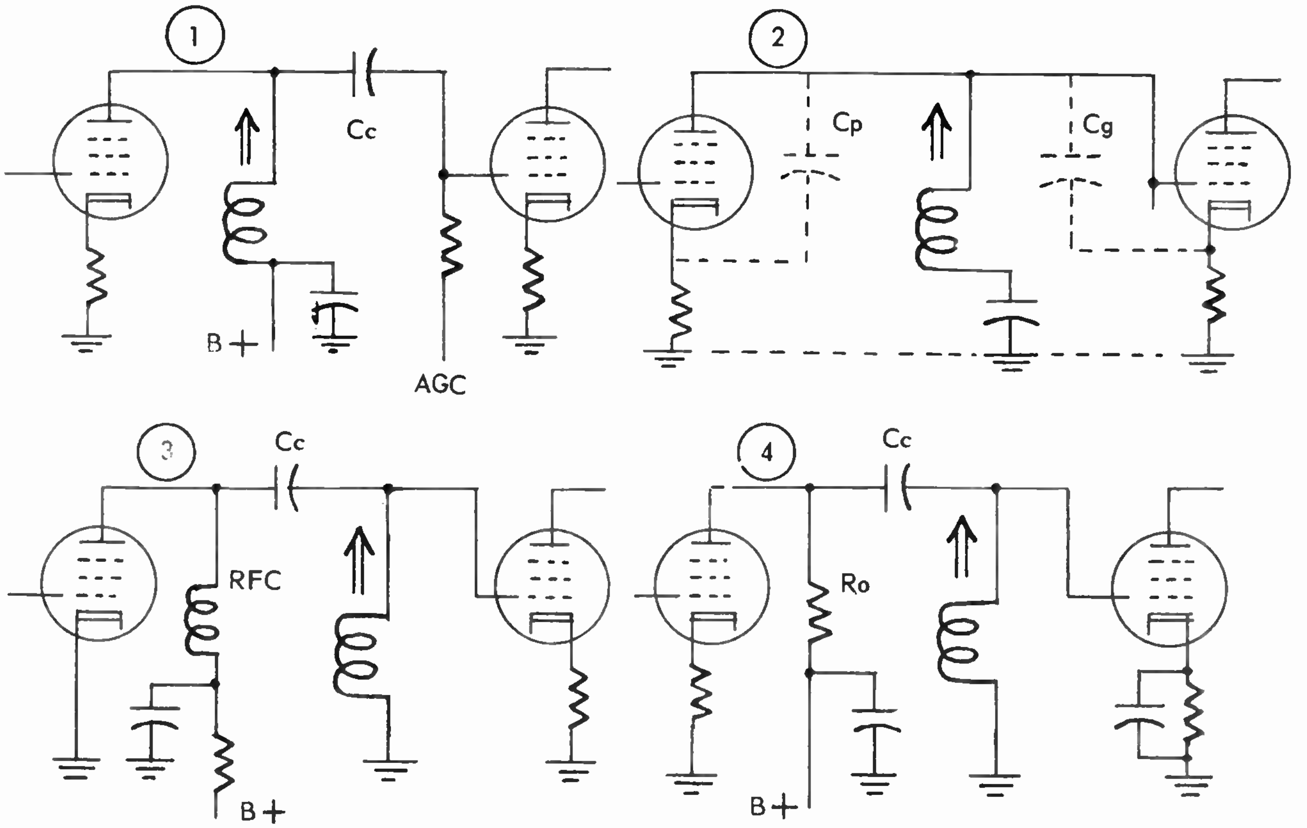


Fig. 45-6. Connections of impedance couplers in i-f amplifiers.

They are effectively in parallel with the coupler inductance. Capacitor *Cc* is omitted from this diagram illustrating principles because, as explained, this capacitor has negligible effect on tuning.

The coupler is tuned to resonance by moving its core to vary the inductance, and with the fixed parallel capacitances, to produce resonance at a frequency within the range of desired video i-f response. The tube whose plate load is formed by the coupler will have maximum amplification at the tuned frequency, for at this frequency there is maximum impedance. The amplified signal voltage appearing across the coupler is applied to the grid of the second tube.

As shown by diagram *3* there may be shunt feed, with the tuned impedance coupler connected on the grid side rather than on the plate side of capacitor *Cc*. D-c plate voltage for the first tube is here furnished through the radio-frequency choke coil *RFC*. This choke has high impedance at all the video intermediate frequencies to be amplified, and retards escape of signal current and voltage to and through the B-supply lines. Keeping in mind that capacitor *Cc* merely isolates the second grid while allowing passage of signal current, we have in diagram *3* essentially the same action as in diagram *2*.

In diagram *4* a high resistance at *Ro* replaces the radio-frequency choke of diagram *3*. This resistor re-

COMMERCIAL TRADES INSTITUTE

TELEVISION

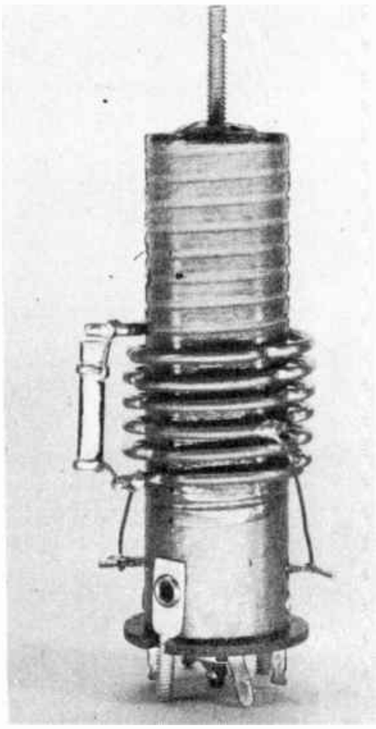
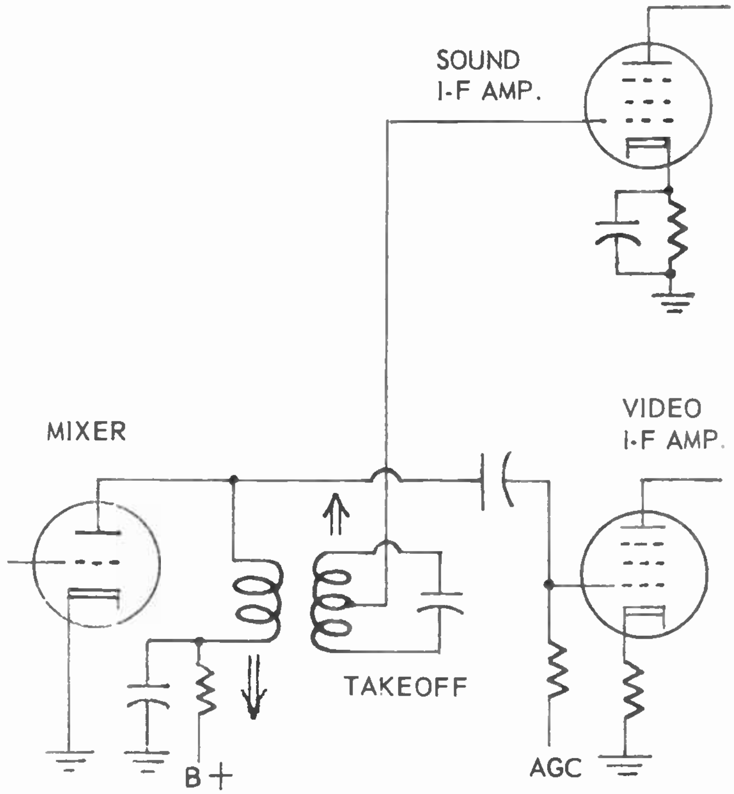



Fig. 45-7. Impedance coupler unit with sound takeoff winding on the outside.

tards escape of signal current and voltage through the B-supply, and forces them to go through capacitor *Cc* to the impedance coupler and the grid of the second tube. Otherwise the action is the same as in diagram *2*.

At the left in Fig. 45-7 is a picture of a unit having a large-diameter tubular form inside of which, down close to the bottom, is a small impedance coupler similar to those illustrated by Fig. 45-5. The adjustment screw for the movable core of this coupler extends downward through the bottom of the large form. The coil of heavy wire which is around the outside of the form is a winding for sound takeoff. A movable core for tuning this takeoff winding is adjusted by a threaded stud which extends up through the top of the large form.

The circuit for this combined impedance coupler and sound takeoff is shown at the right. The impedance coupler is connected between the plate of a mixer and the grid of the first video i-f amplifier tube. The action of this coupler is exactly the same as described in connection with Fig. 45-6. There is enough inductive coupling between the impedance coupler inside the large form and the sound takeoff winding on the outside that the takeoff winding acts as the secondary of a transformer. Tuning capacitance for the takeoff winding is provided chiefly by a fixed ceramic capacitor which is plainly visible on the left of the form in the photograph.

Many i-f couplers consist of a transformer with its primary winding in the plate circuit of the first tube and its secondary in the grid circuit of the second tube, as at *A* in Fig. 45-8. If the two windings are tuned to the same frequency there will be maximum energy transfer and maximum response or gain at the tuned frequency. In order that the two windings may be resonant at the same frequency the product of inductance and capacitance on the primary side must be equal to the product of inductance and capacitance on the secondary side.

Across the primary winding there is the output capacitance of the first tube and the other capacitances in the plate circuit. Across the secondary there is the input capacitance of the second tube and other capacitances in the grid circuit. These primary and secondary capacitances ordinarily will not be equal, and even were they made so to begin with, the equality would be upset when replacing either of the tubes. Two tubes, even if the same type, seldom will have exactly the same internal capacitances.

For tuning plate and grid windings to the same frequency an adjustable trimmer capacitor could be connected across either winding, or trimmers could be across both windings, as at *B* in Fig. 45-8. Were only one of the windings variably tuned in this manner, the adjustable trimmer could be set to combine its capacitance with that of the tube and circuit so that there would be a match for the fixed tube and circuit

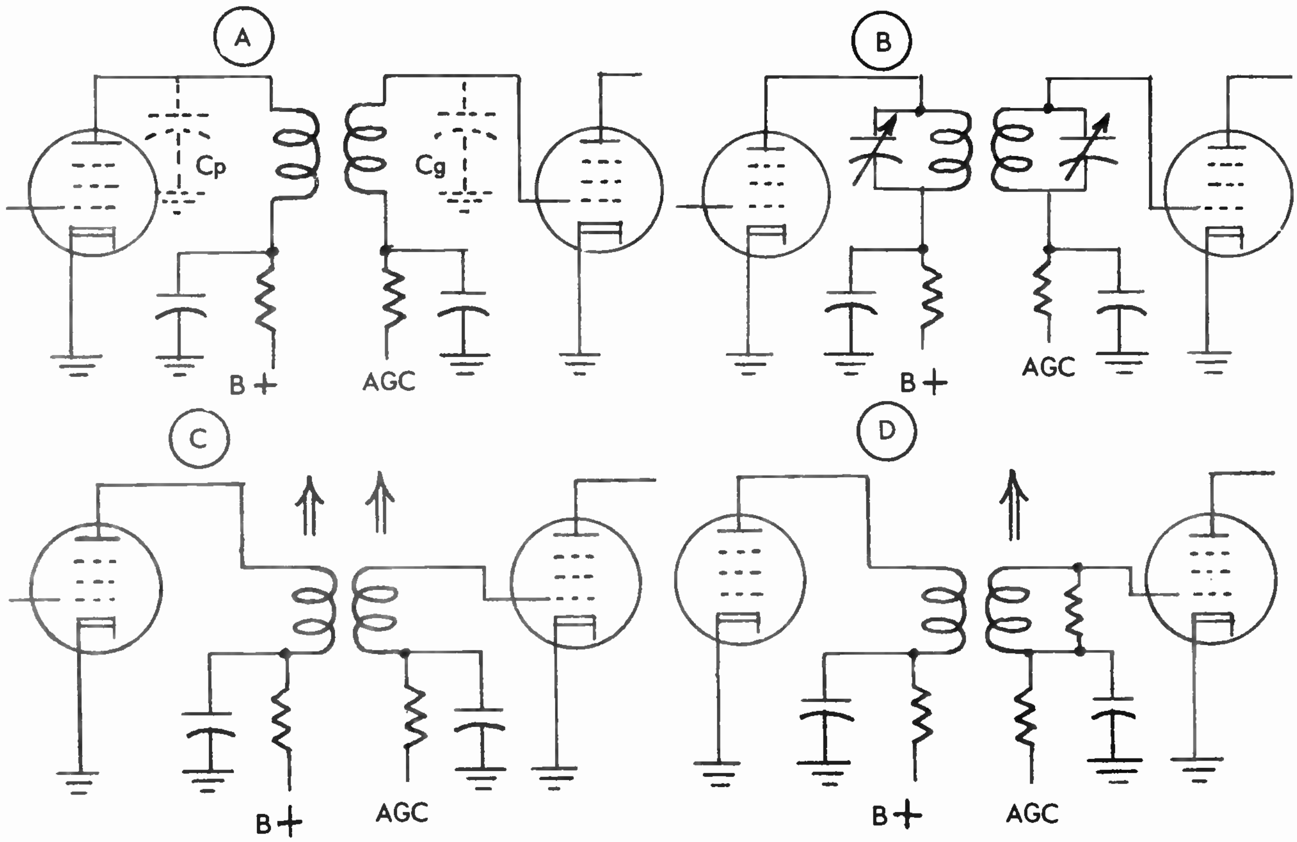
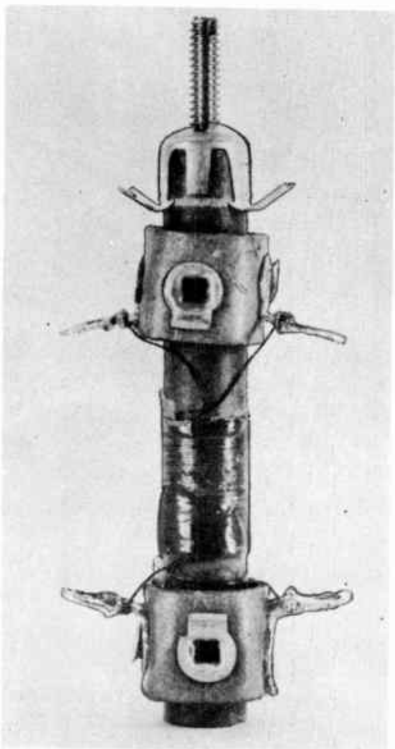


Fig. 45-8. Tuning of transformer windings to the same frequency.



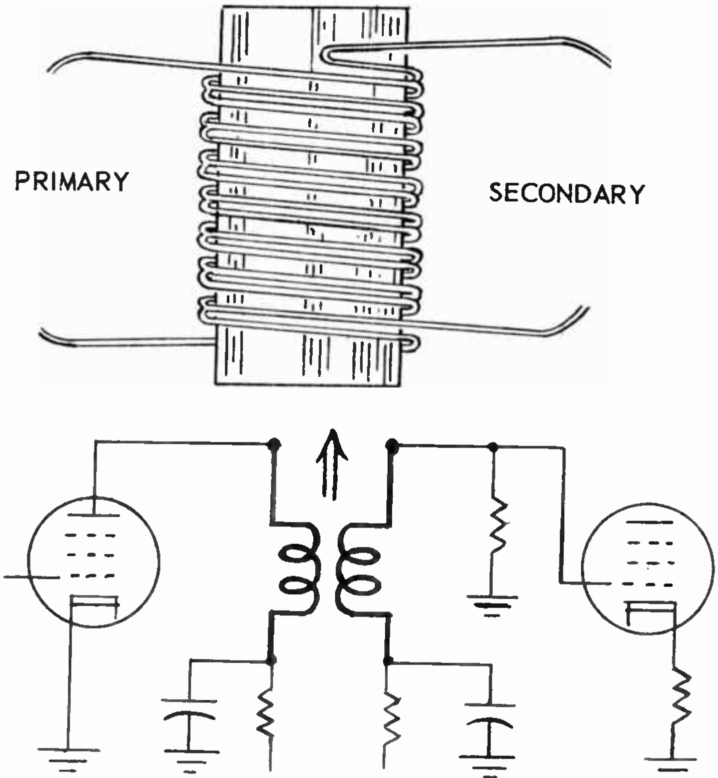


Fig. 45-9. Coupling transformer constructed with a bifilar winding.

capacitance on the other side. Because one of the major problems in design and construction of i-f amplifying stages is reduction of circuit capacitance, this method of adding still more capacitance for tuning is not often used.

The primary and secondary inductances can be made adjustable for tuning to the same frequency, with whatever circuit capacitances may exist, by providing movable cores in both windings. This is shown by diagram *C*. This method allows retaining minimum possible circuit and tube capacitances while still tuning both windings to resonance at the same frequency.

Still another method of using transformer coupling is shown at *D* in Fig. 45-8. The primary winding is tuned by the circuit and tube capacitance on the plate side. The secondary is adjustably tuned by a movable core. Across the secondary winding is connected a resistor which broadens the frequency response sufficiently to cover the band required in the one coupling.

A transformer made with what is called a 'bifilar' winding is illustrated by Fig. 45-9. The wires for primary and secondary are laid side by side and are wound thus for the entire length of the coil. This results in very close coupling between plate and grid circuits, and a high rate of signal energy transfer. There is a single movable core which alters the inductance of both windings at the same time. The inductances of primary and secondary windings are practically equal, since both have the same number of turns, same diameter, and same length. In most amplifiers constructed with this type of transformer there is a band broadening resistor connected across either the primary or secondary. One of these resistors, connected from grid to ground, is shown in the circuit diagram of Fig. 45-9.

STAGGER TUNING. All of the couplers which we have so far discussed provide the familiar form of frequency response having a peak at the tuned resonant frequency and sides sloping down to zero at frequencies below and above resonance. None of them, by itself, provides a response such as required from the entire video amplifier system. As you will recall, this overall response must be down about 50 per cent at the video intermediate frequency and must not extend far, if any, beyond the band covered by a channel.

By combining in the video amplifier system the responses of three or more couplers having peak gains at suitable frequencies it is easily possible to obtain an overall response satisfying all requirements. Fig. 45-10 is a circuit diagram of an i-f amplifier system in which there are four single-peaked impedance couplers. The first two couplers are in the plate circuits of the mixer tube and the first i-f amplifier tube. The third coupler is in the grid circuit of the third amplifier tube, and the fourth coupler is in the video detector circuit. The plates of the second and third amplifier tubes are connected to the B-supply through radio-frequency chokes. The grids of the first and second amplifier tubes are connected to a source of automatic gain control, and have minimum negative biases provided by cathode resistors. The third amplifier operates with only cathode bias.

In a particular receiver containing this i-f amplifier system, the sound intermediate frequency was to be 21.6 mc and the video intermediate frequency 26.1 mc. The first and third impedance couplers were aligned to have their peak responses at 25.6 mc. The second and fourth couplers were aligned for peak response at 23.4 mc. The frequency responses of the individual couplers from first to fourth are shown in the first

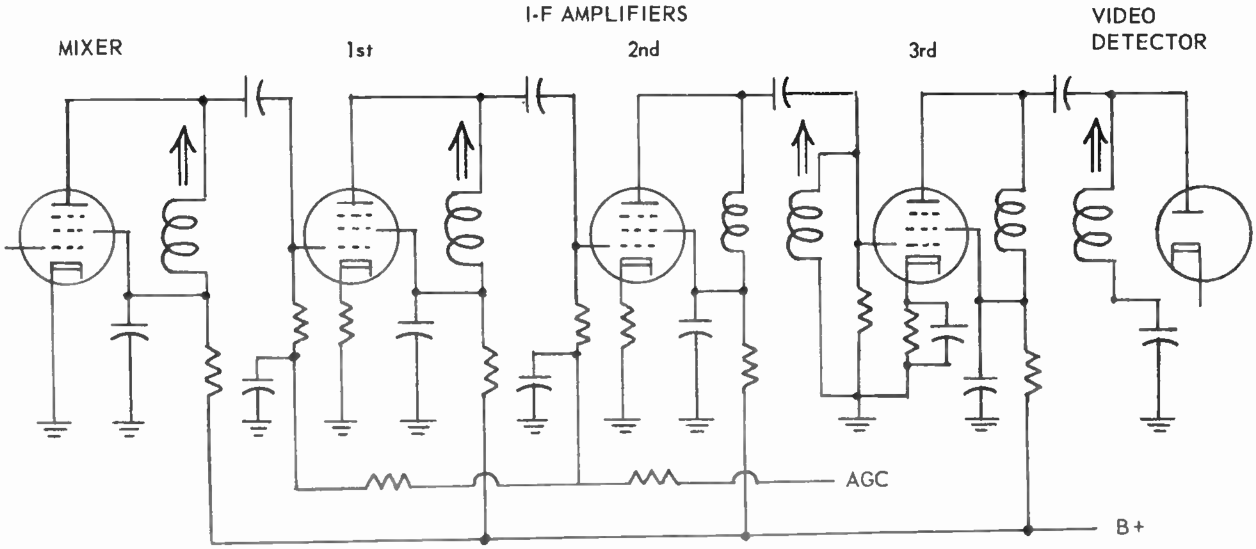


Fig. 45-10. An i-f amplifier circuit in which there are four impedance couplers.

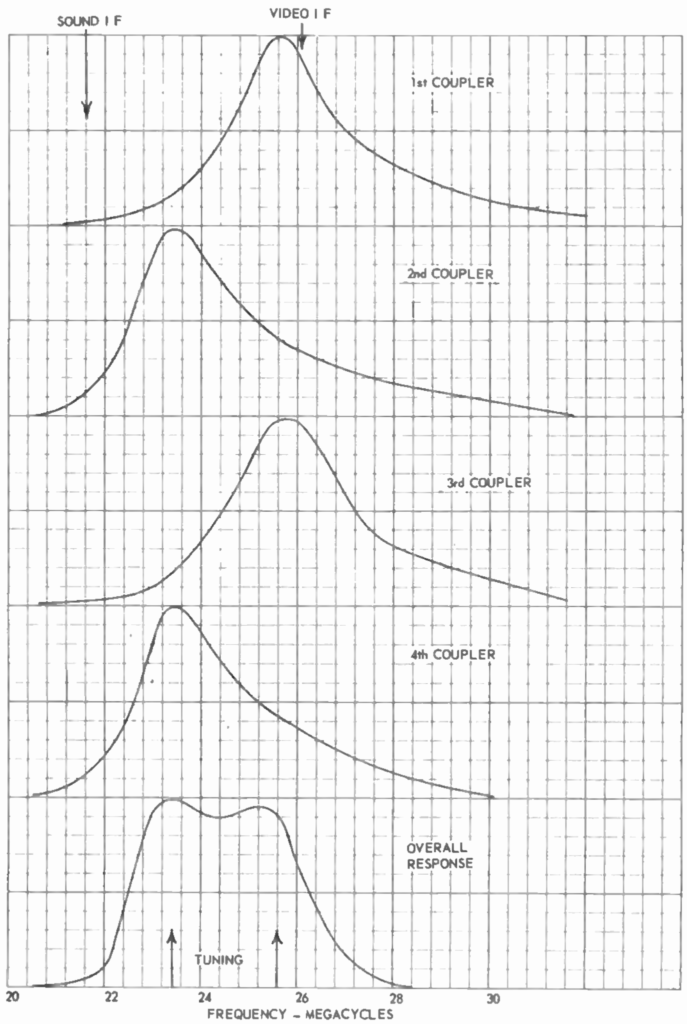


Fig. 45-11. How the responses of several single-peaked couplers may be combined by stagger tuning.

four curves of Fig. 45-11. These four separate stage responses produce an overall response as shown by the bottom curve. The video intermediate frequency is at about 60 per cent of maximum gain, and the sound intermediate is down around 3 per cent of maximum. This is a satisfactory gain for the sound intermediate frequency, since the receiver uses an intercarrier sound system.

This general method of obtaining an overall i-f response of suitable shape is called stagger tuning. It is possible to obtain a desired overall response by using any of many different combinations of peak frequencies for the several couplers. It is essential that the two couplers preceding and following any one tube be peaked for different frequencies. Were the couplers on both sides of the same tube to be resonant at the same frequency there nearly always would be enough feedback of signal energy through the tube capacitance to cause oscillation. The tube would act as an oscillator rather than as an amplifier.

We should note that the gains or amplifications in successive stages do not add together at a particular frequency, instead they multiply. For example, were there a gain of 5-to-1 at 25 mc in one stage, and a gain of 4-to-1 in a following stage at this same frequency, the total gain for the two stages at 25 mc would be 20-to-1, which is the product of 5 and 4. Because of this fact it is not only the peak frequencies of the various couplers which determine the shape of the overall response, it is the shapes of the responses of all the couplers. That is, it is not only the gains at the peak frequencies which are of importance, but it is the product of the gains at all frequencies which determine the overall gains at all frequencies. Should there be zero gain at some certain frequency in any one stage, that frequency would be zero in the overall response no matter how much amplification it might receive in other stages. Any amount of gain multiplied by zero becomes zero gain.

When there are four couplers and three amplifier tubes in the i-f system it is general practice to tune the first and third couplers for the same frequency, and to tune the second and fourth couplers to another frequency, as in the example which has been illustrated. As a general rule the peak frequency for two of the couplers will be a fraction of a megacycle lower than the video intermediate frequency, and the peak frequency for the other two couplers will be somewhat less than 2 megacycles above the sound intermediate frequency. In some receivers all four couplers are peaked at different frequencies. If there are only three couplers and two amplifier tubes the couplers will be peaked at three different frequencies.

When there are five couplers and four i-f amplifier tubes it is most common practice to tune all the couplers to different peak frequencies. As a general rule two of the couplers will be peaked slightly below the video intermediate frequency, two more will be peaked around a half to one megacycle above the sound intermediate. The fifth coupler will be peaked somewhere in between to prevent formation of a " valley" between the two overall peaks, as occurs in the response shown at the bottom of Fig. 45-11. There are some receivers in which two of the four couplers are peaked at the same frequency, with the remaining couplers peaked at three different frequencies.

In i-f amplifying systems containing fewer than five couplers it is somewhat of a problem to obtain an overall frequency response that is sufficiently wide without having an excessively deep valley between two peaks at higher and lower frequencies. The passbands or frequency responses of the several peaked couplers often are broadened by connecting resistors across the coupler windings or across the parallel resonant circuits of which the couplers are a part. Broadening resistors often are of such values as 5,600 ohms, 8,200 ohms, or 10,000 ohms, but they may have resistances as low as 2,000 ohms or as high as 50,000 ohms.

A broadening resistor is easily recognized in a circuit diagram when it is connected across a transformer winding as at *R* in diagram *1* of Fig. 45-12, or across the winding of an impedance coupler. But a resistor may be located anywhere in a resonant circuit and still broaden the response. We might have connections such as in diagram *2*, where resistor *R* affects the passband because it is in the high-frequency circuit completed through capacitors from the top of *R* to the top of the impedance coupler and from the bottom of *R* to ground and the bottom of the coupler.

In diagram *3* the broadening resistor is in the high-frequency circuit completed through three capacitors; one to ground from the bottom of *R*, another from ground to the bottom of the impedance coupler, and a third between the top of the impedance coupler and the grid. Resistors marked *R* in diagrams *2* and *3* are part of the d-c grid return paths to the automatic gain control voltage sources. From the appearance of the circuit connections these broadening resistors and the capacitors connected to the grids could be mistaken for

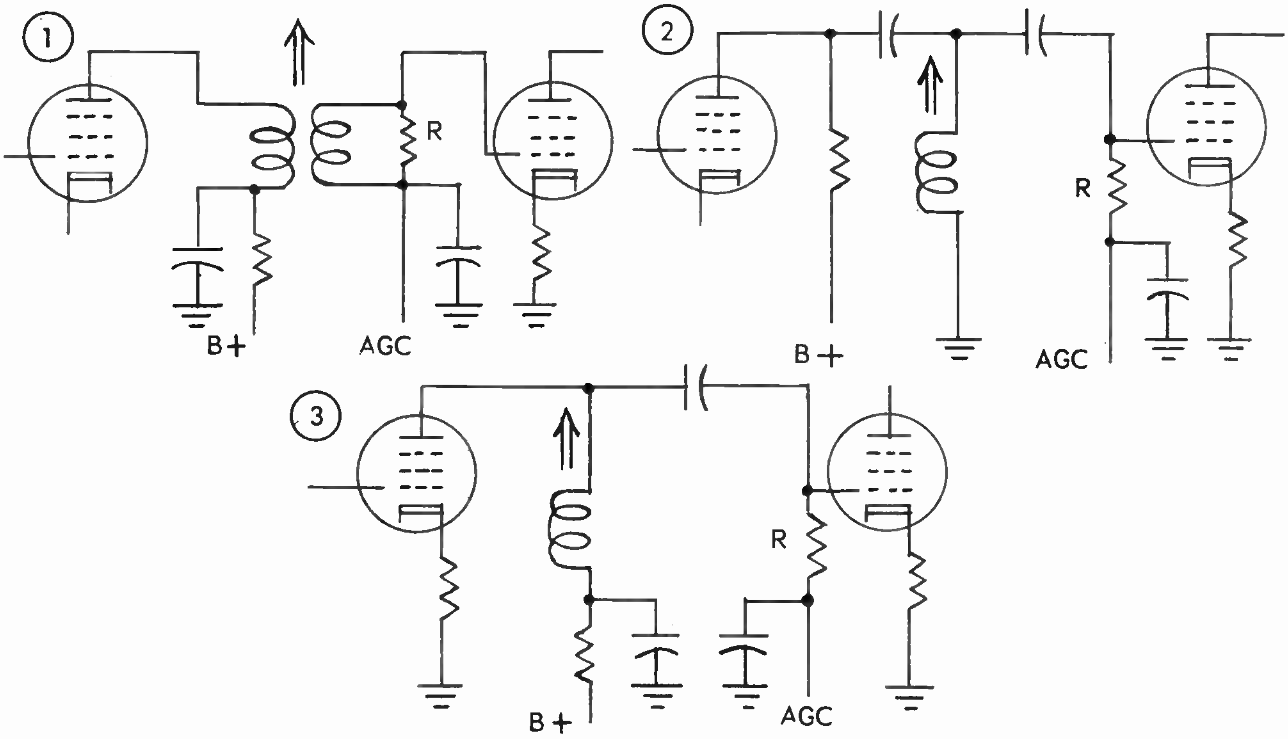


Fig. 45-12. Some locations for band-broadening resistors.

grid-leak biasing circuits. However, the values of resistance always will be much less than used for this kind of biasing.

Resistors which are in series with amplifier tube cathodes, and which have no bypass capacitors, cause the action which is called degeneration. Degeneration broadens the frequency response, as we shall learn when studying broad band amplifiers a little later. When a tuned circuit of any kind is connected to a diode video detector the response is relatively broad, because the diode "loads" the circuit in a manner which has the same effect as a paralleled resistance.

COUPLERS WITH DOUBLE-PEAKED RESPONSE. If the primary and secondary of a coupling transformer are separately tuned to two slightly different frequencies the response will have two peaks, one for each frequency. Although this broadens the passband, it is a method not commonly employed because the energy transfer and resulting gain is considerably less than when both windings are tuned to the same frequency.

If the degree of coupling in a transformer or in any other type of coupler can be increased beyond what is called 'critical coupling', the single resonant peak will be replaced with two separated peaks when both parts of the transformer or coupler are tuned to the same frequency. The greater the degree of coupling or the closer the coupling, the greater will be the frequency separation between the two peaks. Thus an increase of coupling beyond the critical value will broaden the response while maintaining high gain.

There are numerous ways of obtaining a coupling sufficiently close to produce a double-peaked response. Some are shown in diagram form by Fig. 45-13. With all of these methods the plate and grid circuits may be independently tuned to the same frequency using the existing circuit and tube capacitances with the adjustable inductances to produce resonance. Also, with all methods illustrated here, the coupling is adjustable so that frequency separation between the two resonant peaks may be varied to secure the necessary width of pass band. All the couplings are shown connected between two pentode amplifiers. They might be used also between a mixer and an i-f amplifier or between an i-f amplifier and a diode detector.

In diagram *1* inductances *Lp* and *Lc* are in the plate circuit. They are tuned together for resonance at the desired frequency. Inductances *Lg* and *Lc* are in the grid circuit. With any value of inductance to which *Lc* has been adjusted, the inductance of *Lg* may be aligned for resonance at the same frequency for which the plate circuit is tuned. The greater the inductance and inductive reactance at *Lc*, which is in both the plate circuit and the grid circuit, the closer will be the coupling and the greater the separation between resonant peaks. Thus the inductor at *Lc* determines the coupling and also affects the tuning, while the other two inductors have their chief effect on tuning.

In diagram *2* the plate circuit is tuned to resonance at the desired frequency by alignment of inductor *Lp*. The grid circuit is tuned to resonance at the same frequency by alignment of inductor *Lg*. Coupling is var-

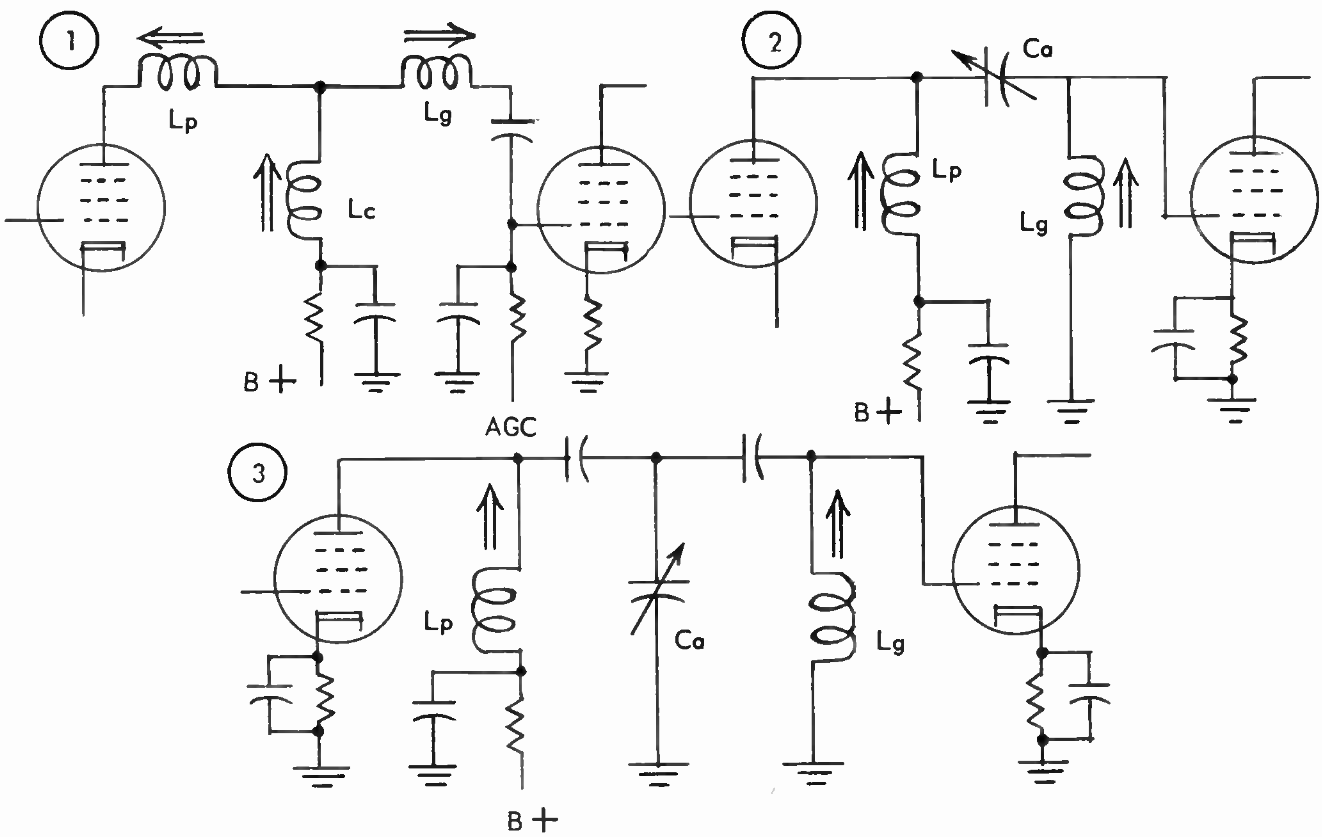


Fig. 45-13. Types of couplers which give double-peaked responses and have adjustable degrees of coupling for peak separation.

ied by alignment of capacitor *Ca*. The smaller the capacitance and the greater its reactance the weaker or looser becomes the coupling, while more capacitance and less reactance increases the coupling. In order that this capacitor may vary the coupling its reactance always must be great enough to offer considerable opposition to passage of signal currents through it. If the capacitance is too great and the resulting reactance too small, the unit acts merely as a blocking capacitor to isolate the grid bias from the plate voltage.

In diagram *3* the plate circuit again is tuned to resonance by inductor *Lp* and the grid circuit is tuned to the same frequency by inductor *Lg*. The degree of coupling is determined by variable capacitor *Ca*, which is acting as a sort of shunt between ground and the top connection between plate and grid circuits. The less the capacitance and the greater the capacitive reactance at *Ca* the closer becomes the coupling and the greater is the frequency separation between resonant peaks of the response. More capacitance and less reactance at *Ca* weakens the coupling.

The response of a single coupler of the double-peaked variety might be made broad enough to cover both the video and the sound intermediate frequencies, but the sides or skirts of the curve would have such gentle slopes as to extend into adjacent channels. But when several such couplers are used, one after another in the i-f stages, the skirts become steeper while there is an increase of gain.

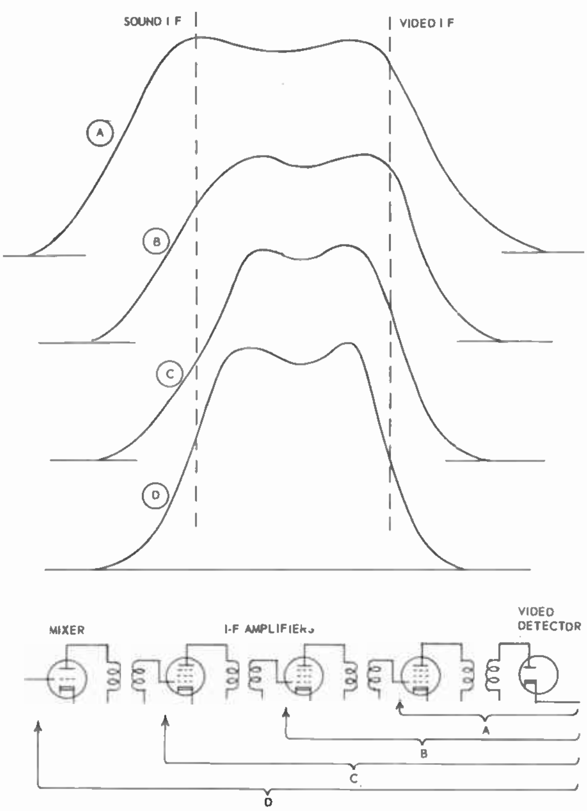


Fig. 45-14. Responses when including successive stages coupled through double-peaked units.

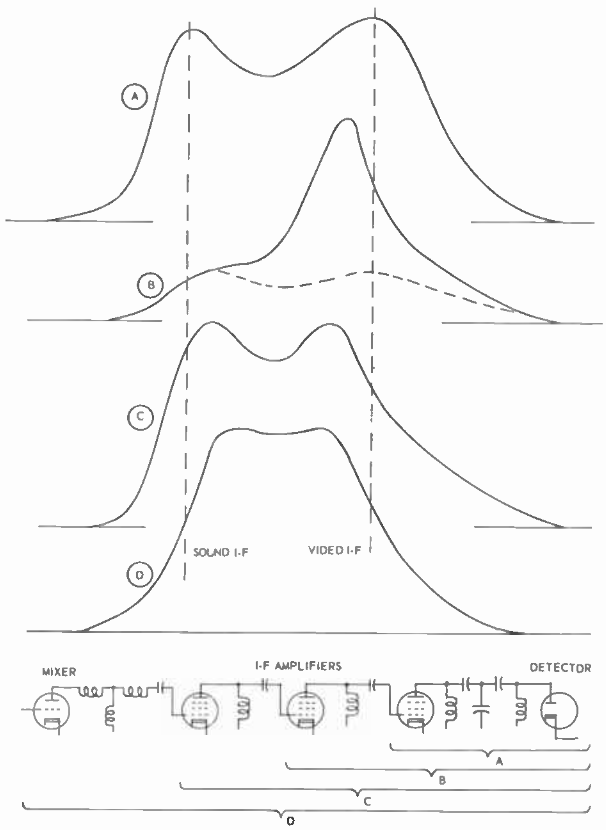


Fig. 45-15. Responses of successive stages having double-peaked and single-peaked couplers.

Fig. 45-14 illustrates successive responses as a signal is put through four couplers, each of which has a very broad pass band when used alone. The response curves are taken with an oscilloscope connected to the output of the video detector, while the signal generator is moved back step by step to include additional couplers between this signal source and the detector output. Curve *A* is the response of one coupler, with generator and oscilloscope connected as indicated by bracket *A* at the bottom of the figure. Curve *B* is taken with the generator moved back to pass the signal through one more coupler, as shown by bracket *B* down below. Curve *C* shows the response when three couplers are included, and curve *D* shows the overall response of the entire video amplifier system, with the generator feeding into the mixer grid and the oscilloscope still at the detector output.

Couplers providing double-peaked response and others giving single-peaked response often are used together in the same i-f amplifier system. Fig. 45-15 shows frequency responses of a system in which the couplers, from mixer to video detector, consist of a double-peaked unit, then two single-peaked impedance couplers, and finally another double-peaked unit. The responses are shown by curves marked *A* to *D*. As shown by the diagram at the bottom of the figure the oscilloscope remains connected to the output of the video detector while the signal generator is moved successively back to include more and more couplers. Letters on the brackets under the diagram show the couplers included for similarly lettered response curves.

The response marked *A* is that for one of the double-peaked couplers. When the signal generator is moved back to include one single-peaked coupler, in addition to the double-peaked unit, we have the response of curve *B*. The single-peaked impedance coupler adds a high response peak at the frequency to which this coupler is tuned, at a point slightly lower than the video intermediate frequency. The broken-line curve which has been added to response *B* corresponds to the original response at *A* with its height lowered enought to accommodate the new peak. Although all our response curves are shown of the same height, the gain actually would increase as more and more stages are included, and the curves would get higher and higher. To keep the responses within the size of the oscilloscope screen we keep lessening the output of the signal generator as the tests continue.

Response curve *C* shows what happens when the signal generator is moved back to include another single-peaked impedance coupler. This added coupler is tuned to a frequency somewhat higher than the sound intermediate frequency, and it adds a second peak at its tuned frequency.

The final step is to move the signal generator back to the grid of the mixer tube, thus putting the signal through all four couplers, including the double-peaked unit that follows the mixer. The resulting response is shown by curve *D*. The effect of the added coupler is chiefly in narrowing the passband and in making steeper skirts on the response.

There are numerous other combinations of single and double-peaked couplers in the same amplifier system. In some receivers there will be two or more couplers whose total effect is a response with a dip or valley between two resonance peaks. Response at the valley frequency may be raised by using one single-peaked coupler tuned to this frequency, thus leveling off the valley or removing it entirely. In other cases you will find one double-peaked coupler used with two or more single-peaked impedance couplers. The double-peaked unit helps to broaden the upper part of the overall response and increases the steepness of the skirts.

In many i-f amplifier systems it is desirable or necessary to reduce the response at certain frequencies while leaving the gain as high as possible at other points. This is accomplished with a variety of wave traps, which we shall investigate in the following lesson.

46→50

TELEVISION

LESSON 46

VIDEO DETECTORS AND TRAP CIRCUITS

Training Your Mind...
C. T. I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind

C.T.I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON 46

VIDEO DETECTORS AND TRAP CIRCUITS

At the output of the television i-f amplifier we have a signal at the intermediate frequency, with amplitude modulation for sync pulses and picture variations. This signal is demodulated by the video detector. The output of the detector contains the sync pulses and picture variations or picture signals. This detector output goes through the video amplifier system and from there to the grid-cathode circuit of the picture tube.

The video detector could be any type which is capable of demodulating amplitude-modulated signals. It is, however, practically always a simple diode type. The diode detector is used for the same reasons as in a-m sound receivers, because it is capable of handling strong signals with a minimum of distortion. In the majority of television receivers the detector is one section of a twin-diode tube, with the other section used for automatic gain control or for any of various other purposes. Instead of this thermionic or vacuum tube type of video detector quite a few sets have a crystal diode in this position.

Fig. 46-1 shows two of the many video detector circuits which are in common use. The detector diode is the one section of the twin-diode that is connected into the circuits. Input to the detector is from signal voltage developed across the last impedance coupler in an i-f amplifier, but might just as well be be brought from the plate circuit of the final i-f amplifier tube in any other suitable manner. The i-f signal may be represented as at the lower left. When this signal is applied to the plate of the detector diode, as in the circuit diagrams, there is conduction through the diode only on the i-f alternations which make the plate more positive than the cathode, and there is no conduction on the opposite alternations. This action cuts off the negative side of the i-f signal and leaves the envelope of the positive side as the input to the video amplifier tube which follows the detector.

In the circuit shown by diagram 1 there is a choke coil or inductor L in series with the detector output. This inductor has high reactance and impedance at the video intermediate frequency, but has relatively low impedance at the picture and sync frequencies. Consequently, the inductor helps to prevent the high intermediate frequencies from getting into the video amplifier, but allows relatively free passage of the lower sync and picture frequencies.

The detector load resistor is marked Ro. Across this resistor appears the demodulated signal voltage consisting of the sync pulses and picture variations which form the video signal. This video signal is applied to the grid of the following video amplifier through coupling and blocking capacitor Cc. The greater the resistance or impedance of the detector load the stronger is the signal voltage fed to the video amplifier.

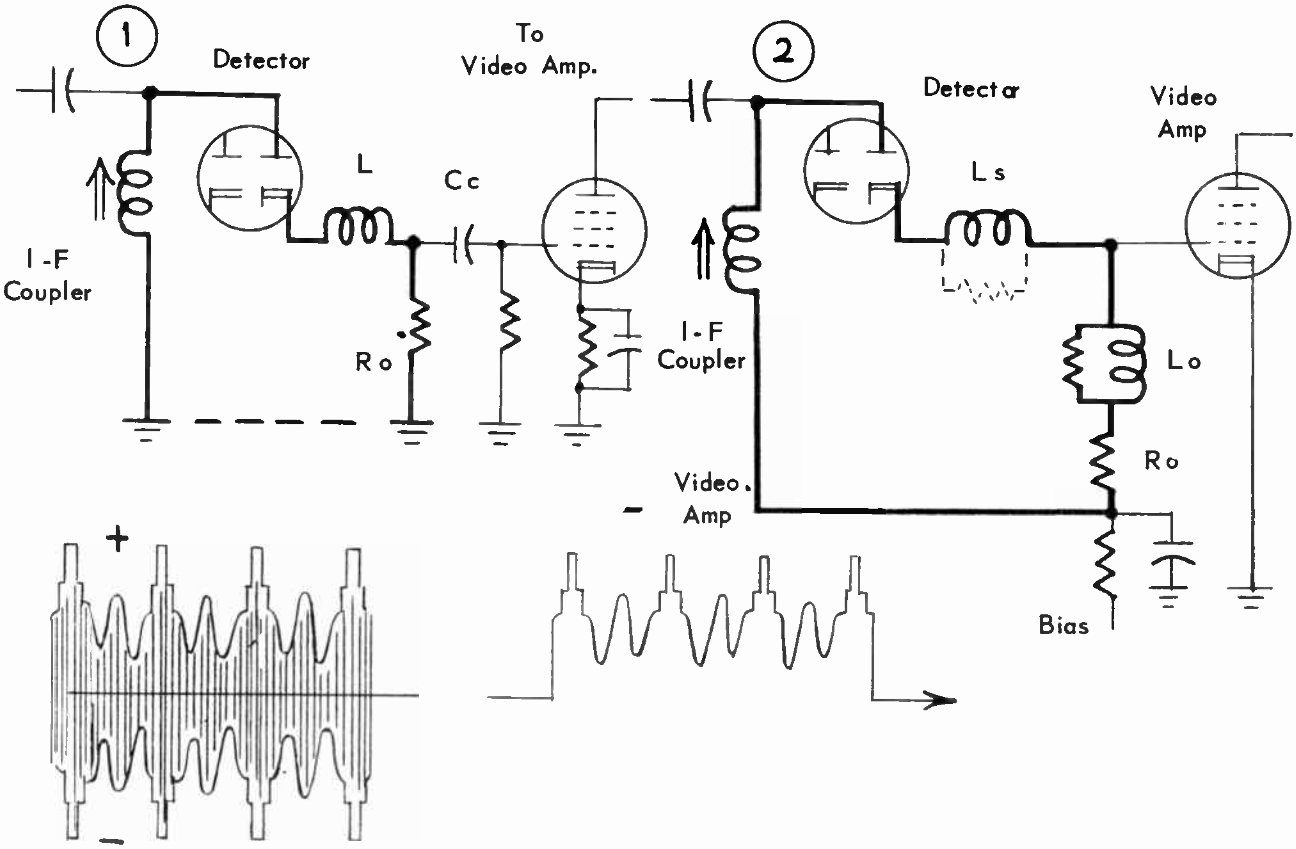


Fig. 46-1. Video detector circuits which provide a negative output signal.

The video detector circuit as shown by heavy lines on diagram 1, includes the i-f coupler, the diode tube, inductor L, load resistor Ro, and the ground connections leading back to the coupler.

In the circuit shown by diagram 2 the inductor in series between the detector diode and video amplifier is marked Ls. Sometimes this inductor is shunted by a broadening resistor. Here we have added to the detector load, in series with resistor Ro, a second inductor or coil Lo which is shunted by a broadening resistor. In parallel with this detector load are the input capacitance of the video amplifier, the capacitance from cathode to ground of the diode detector, and various stray capacitances in the circuit. At the higher video frequencies of picture signals the reactances of these paralleled capacitances become quite low. Since these capacitances are in parallel with the load they tend to make the effective impedance of the entire load become lower and lower as frequency goes up.

The purpose of inductor Lo is to counteract the decrease of load impedance caused by tube and circuit capacitances. At the low video frequencies the inductive reactance of Lo is so small as to add very little to the resistance or impedance of Ro. But this reactance increases with rising frequency. At the higher video frequencies, where tube and circuits capacitances are acting to drop the load impedance, the reactance of inductor Lo is increasing, to raise the impedance. The result is a fairly uniform load impedance at all video frequencies.

In diagram 1 of Fig. 46-1 the video amplifier tube is provided with cathode bias voltage which is insulated from the detector circuit by capacitor Cc. In diagram 2 there is a direct conductive connection between the detector cathode and the grid of the video amplifier, also through the detector load and i-f coupler to the detector plate. The entire detector circuit is connected to a source of bias voltage suitable for the amplifier grid. Various other amplifier bias methods might be used without affecting the operating principles of the detector.

In the video detector circuit of Fig. 46-2 the signal voltage from the last coupler in the i-f amplifier system is applied to the cathode of the detector diode rather than to the plate. Inductor Ls still offers relatively high impedance to intermediate frequencies which otherwise would get to the grid circuit of the video amplifier. Inductor Lo and load resistor Ro perform the same functions as earlier explained. There is a bypass capacitor Cb for carrying intermediate frequencies to ground as these frequencies come through the detector and are opposed in farther travel by inductor Ls. This bypass capacitor usually is used when detector output is from the plate, but usually is not employed when detector output is from the cathode, as in Fig. 46-1. There is enough capacitance from the detector cathode to ground to serve the same purpose. Capacitance at Cb usually is about 10 mmf.

When the i-f signal voltage is applied to the detector cathode, as in Fig. 46-2, there is a conduction through the diode only when its cathode is more negative than its plate, or only during negative alter-

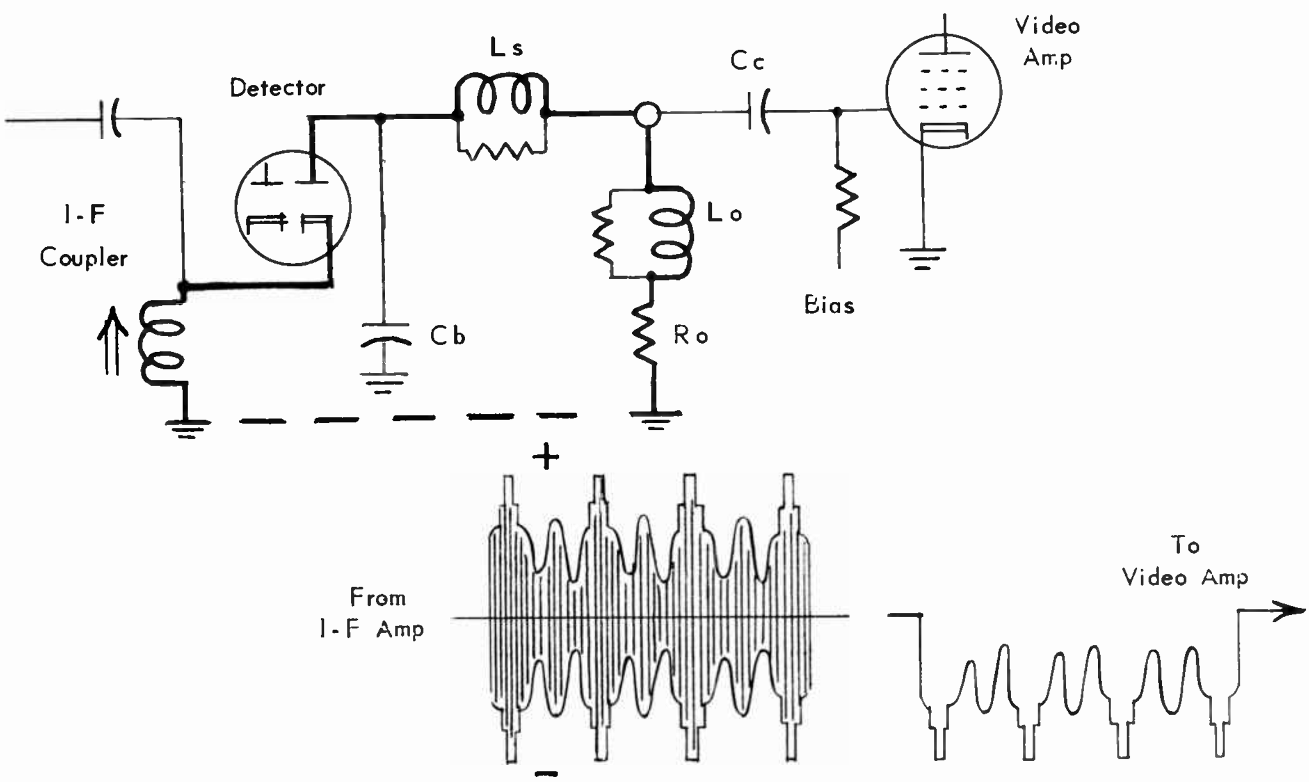


Fig. 46-2. A video detector circuit which provides a positive signal.

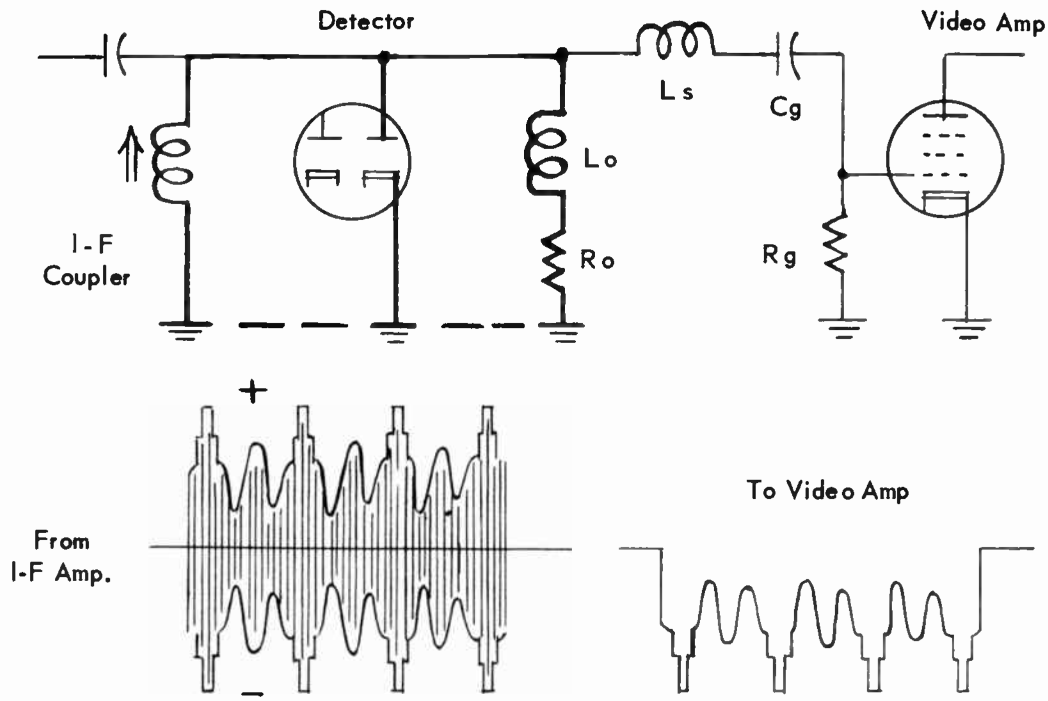


Fig. 46-3. A shunt connected video detector.

nations of the i-f voltage. This results in cutting off the positive side of the input signal and having in the demodulated output of the detector only the negative side, which is applied to the video amplifier grid.

In an earlier lesson it was mentioned that the last coupler in the i-f system tunes rather broadly because of the load formed by the video detector. This loading varies directly with the value of the video detector load, which is the impedance of resistor Ro, or of Ro in series with inductor Lo. The loading, in ohms, is equal to about one-half the detector load impedance. There is more or less loss of signal energy in the detector circuit. The peak output amplitude always will be lower than the peak amplitude of the applied i-f signal voltage.

Still another connection for a diode used as a video detector is shown by Fig. 46-3. Here the diode is in parallel with the i-f coupler and the detector load, which consists of inductor Lo and resistor Ro. I-f voltage from the high side of the coupler is applied to the detector plate. The detector conducts when the plate is made more positive than the cathode, or conducts during positive alternations of the i-f voltage. But this conduction current goes to ground, not to the following video amplifier. During negative alternations of the i-f signal there is no conduction through the detector, for its plate then is negative with reference to the cathode. Consequently, the negative alternations are the ones demodulated, and in the detector output we have the negative side of the input signal envelope. So far as polarity of the detector output is concerned this circuit is equivalent to the one shown by Fig. 46-2.

Fig. 46-4 shows one video detector circuit employing a germanium crystal diode instead of a thermionic tube type diode. The circuit itself is exactly the same as the one shown by Fig. 46-2. The only difference is in substitution of the crystal for the tube. The output polarity with the crystal connected as shown is just the same as with a tube similarly connected with reference to cathode and plate or anode. You will recall that the anode of a crystal diode is the part which corresponds to the plate in a tube. A crystal diode may be used instead of a thermionic diode in any of the detector circuits.

DETECTOR OUTPUT POLARITY. If the output of the video detector as shown by Fig. 46-1 were passed through a coupling capacitor it would become an alternating current and voltage. In this alternating signal voltage the sync pulses would be positive and the picture signals would be negative. Then we say that the detector output is negative, because we refer to output polarity with reference to picture signals.

Were the detector output of Fig. 46-2 to be made an alternating current or voltage the picture signals would be positive and the sync pulses would be negative. In this case the detector output is said to be of positive polarity, because picture signals are on the positive side.

Now let's go to the picture tube and see what signal polarity is required in order that light parts of the original image may be light in the reproduction, and dark parts may be dark. In diagram 1 of Fig. 46-5 the video signal is applied to the grid of the picture tube. In order that the screen of the picture tube may be made brighter when the signal changes from the black level to the white level the signal must be positive. That is, the picture variations must be positive with reference to the sync pulses whenever a video signal is applied to the picture tube grid.

We might obtain a positive video signal directly from one of the video rectifiers whose circuits have been examined. But this signal would not be strong enough to operate the picture tube or to control the rate of electron flow in the picture tube beam. It is necessary to have at least one video amplifier stage between the detector and the picture tube so that the signal may be sufficiently strengthened. The video amplifier tube will invert the polarity of the signal. Consequently, at the grid of the amplifier we must have a negative signal. Such a signal would be furnished by the detector circuits shown in Fig. 46-1.

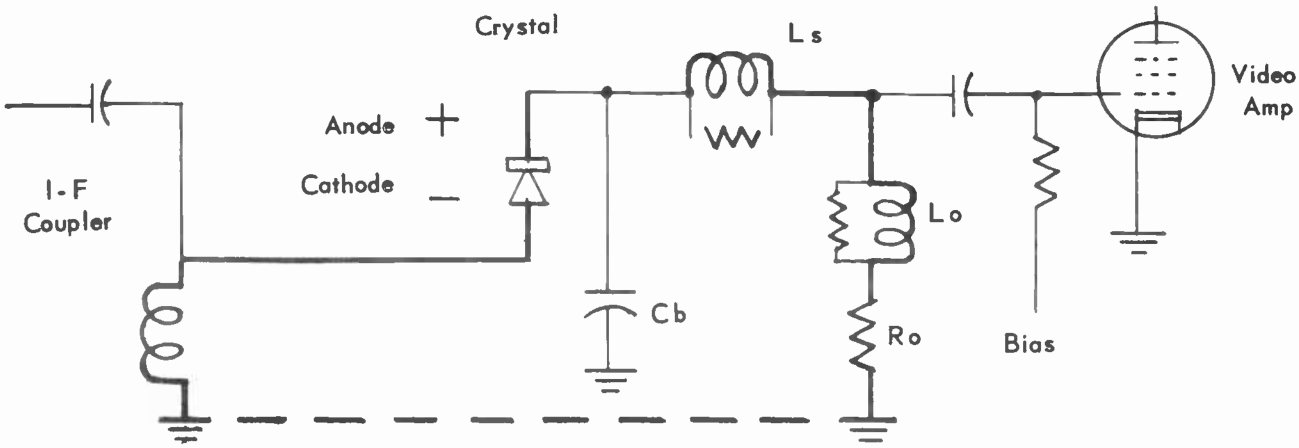


Fig. 46-4. Circuit in which the detector is a crystal diode.

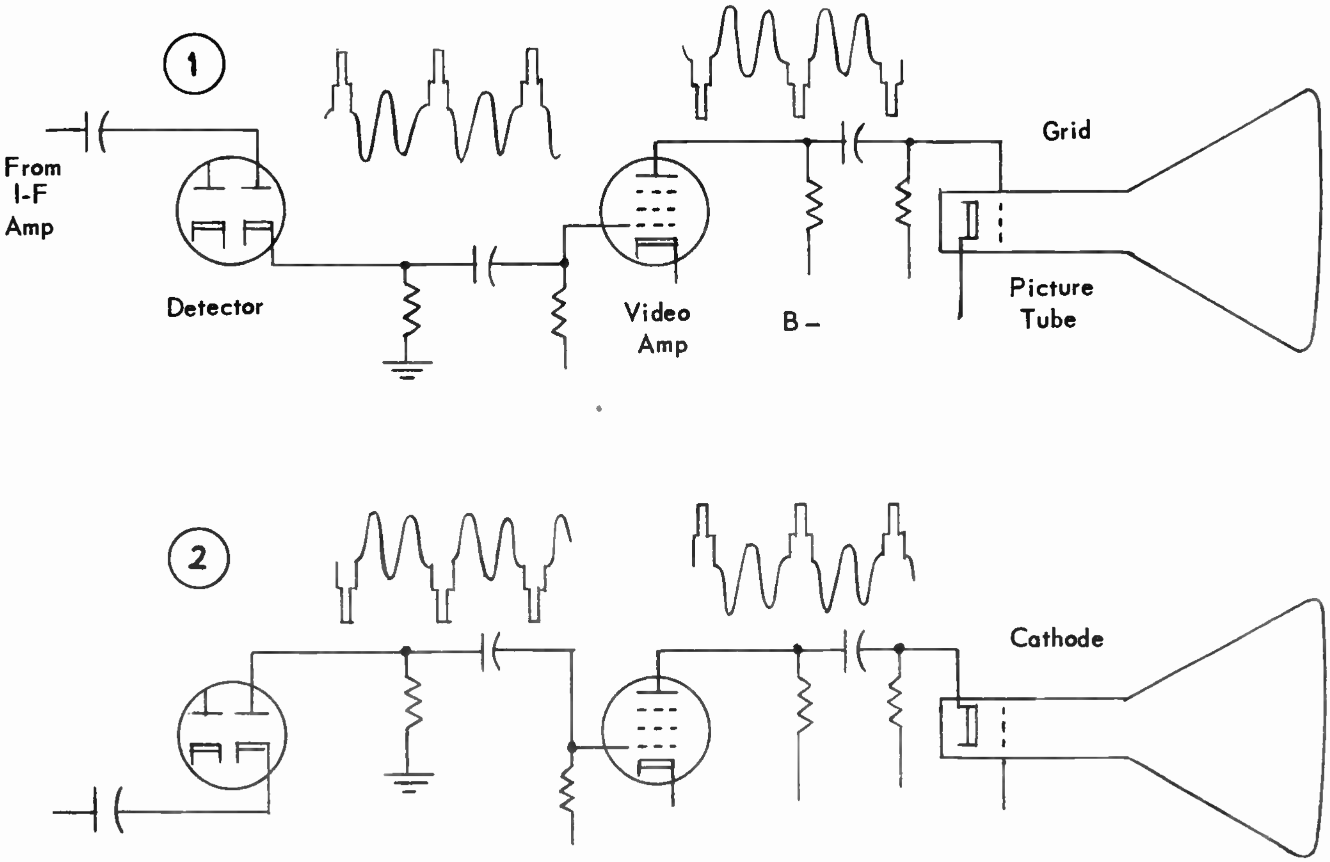


Fig. 46-5. Signal polarities when there is one video amplifier.

In diagram 2 of Fig. 46-5 the video signal is applied to the cathode of the picture tube. Now the signal must be of negative polarity, because making a cathode more negative has the same effect on plate current in an amplifier or on beam current in the picture tube as though we made the grid more positive. The negative video signal at the picture tube cathode produces the same relative lights and shadows on the screen as are produced by a positive video signal at the grid. If we still use a single video amplifier between detector and picture tube it is necessary now to have a positive signal at the grid of the amplifier, or a positive signal from the detector. This signal would be furnished by the detector circuits shown in Figs. 46-2 and 46-3.

Fig. 46-6 shows polarities for the video signal when there are two video amplifiers between the detector and picture tube. The polarity is inverted by the first video amplifier and is inverted again by the second amplifier. We still must have a positive video signal for application to the picture tube grid, and a negative signal for application to the picture tube cathode. Since there is one additional inversion of polarity as compared with the single stage video amplifier, we have to use detector circuits providing for this extra inversion. The detector used for picture tube grid input with a single video amplifier now is used for picture tube cathode input with two amplifiers. The detector first used for cathode input to the picture tube now is used for grid input.

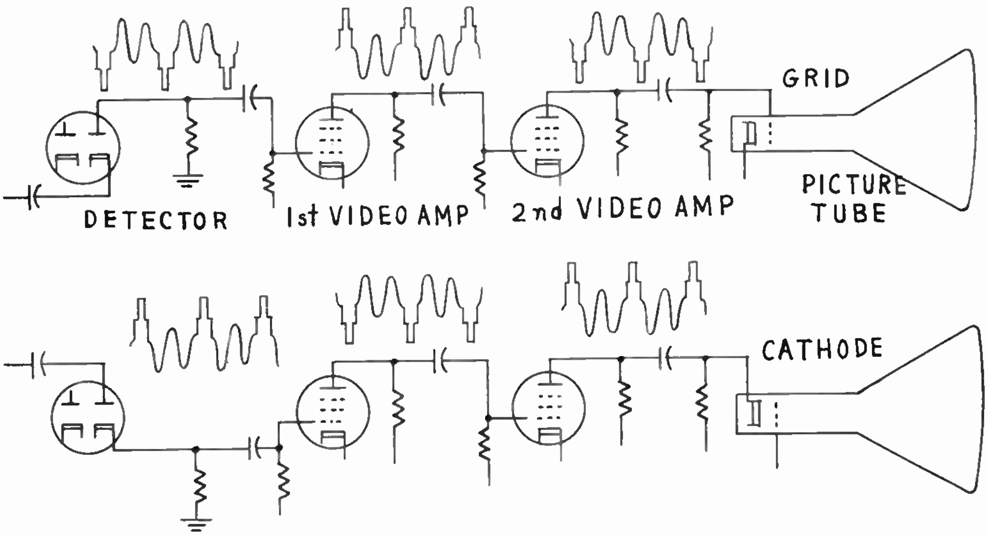


Fig. 46-6. Signal polarities required with two video amplifiers.

The easiest way to remember the signal polarity relations is to keep in mind that picture variations must be positive at the grid of the picture tube. If you don't forget this one fact there will be no trouble remembering that opposite polarity is needed for cathode input. From there it is easy to figure back to the detector output, for the signal must be inverted once in each video amplifier. The detector output polarity required in any receiver is determined by the number of video amplifier stages and by whether the signal is applied to the grid or the cathode of the picture tube.

INTERFERENCE FREQUENCIES. Now we shall examine some undesired signal frequencies which may enter the video i-f amplifier system and cause trouble unless we take steps to reduce or eliminate these frequencies before they reach the video detector. First, let's see how the undesired frequencies are produced.

Any signal frequency coming to the antenna and getting through the r-f amplifier to the grid of the mixer will beat with the r-f oscillator frequency to produce sum and difference frequencies at the mixer output. If these beat frequencies are within the response range of the video i-f amplifier they will be amplified and fed to the picture tube. We wish to amplify certain of the beat frequencies, but not others.

5 In receivers having intercarrier sound systems we wish to amplify the beat which is the video intermediate frequency up to 50 per cent of maximum gain, and to amplify the beat which is the sound intermediate up to something like 2 to 5 per cent of maximum. With dual or split sound systems we again want 50 per cent gain for the video intermediate, but want no sound intermediate frequency at all in the output of the video i-f amplifier.

To learn what actually happens in a particular case assume that the receiver is tuned for reception in channel 8, and that the i-f amplifier is designed and aligned for a video intermediate frequency of 25.75 mc and sound intermediate of 21.25 mc. The r-f oscillator will be tuned for a frequency 25.75 mc higher than the received video carrier frequency. The video carrier in channel 8 is at 181.25 mc, so our r-f oscillator frequency for this channel will be the sum of 181.25 mc and 25.75 mc, or will be 207.00 mc.

Fig. 46-7 shows the beat frequencies which result from mixing of this oscillator frequency with video and sound carrier frequencies of channel 8, also with the carriers in adjacent channels 9 and 7. Each carrier frequency is shown subtracted from the oscillator frequency. The differences form the several beat frequencies. Down below is an overall frequency response curve such as might be had with an i-f amplifier. The broken lines show where the various beat frequencies fall on this response, or in relation to it.

The sound intermediate for the received channel, which here is number 8, is called the accompanying sound frequency or may be called the associated sound frequency. The beat frequency resulting from the video carrier in the next higher adjacent channel (here number 9) is called the adjacent video frequency. It is 6 mc lower than the received video intermediate frequency. The beat frequency resulting from the sound carrier in the next lower adjacent channel (here number 7) is called the adjacent sound frequency. It is 1½ mc higher than the received video intermediate.

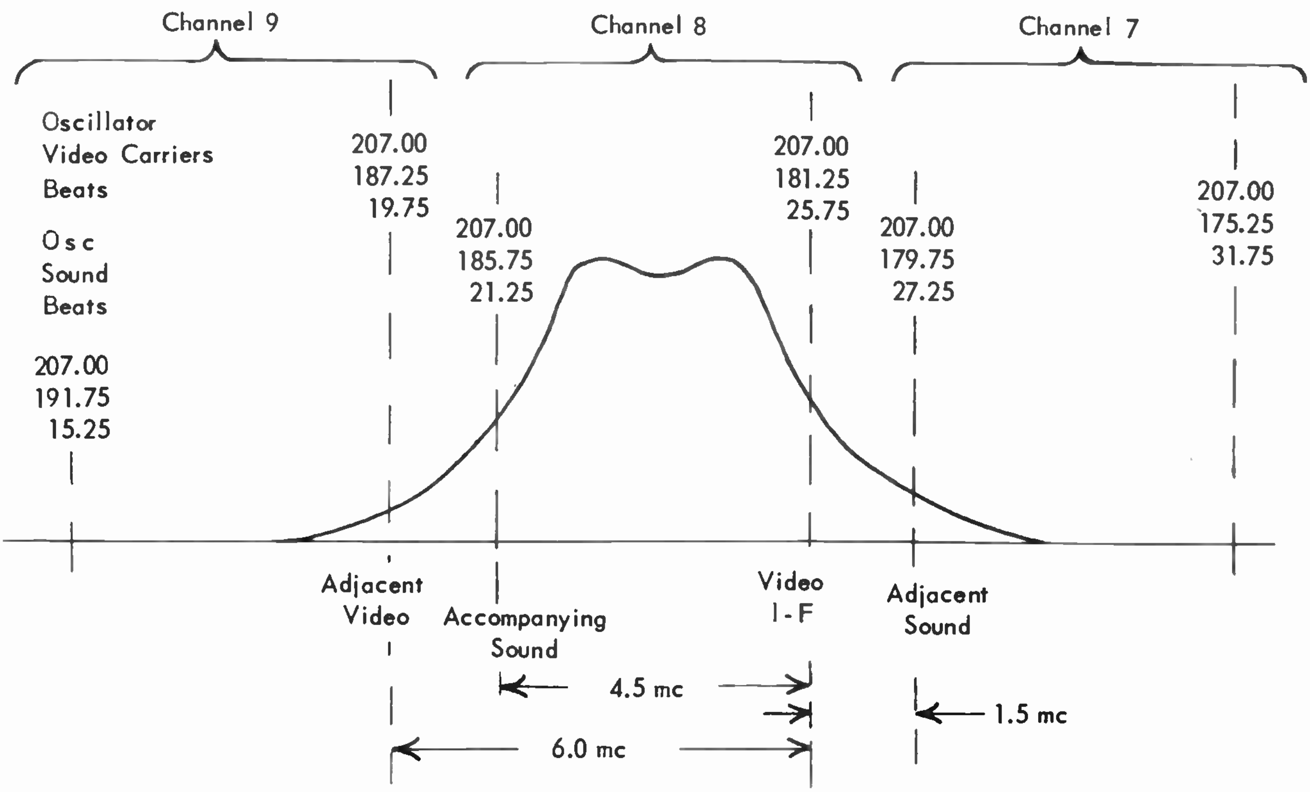


Fig. 46-7. Relations between video intermediate frequency and accompanying and adjacent frequencies.

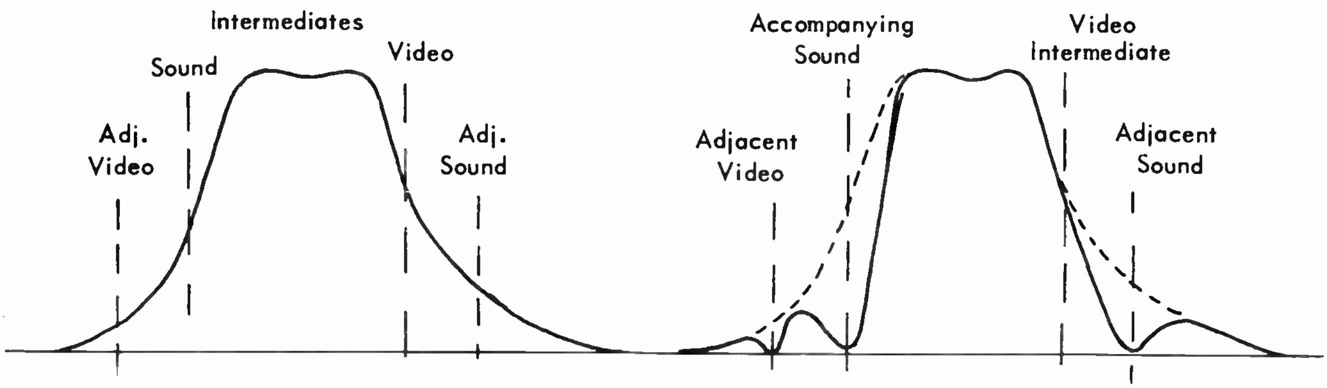


Fig. 46-8. How a response curve may be altered by traps operating on the i-f amplifier.

Adjacent video and adjacent sound frequencies are quite likely to fall on a point of the i-f response where there is considerable gain or amplification. For any ordinary design of i-f amplifier the selectivity is sufficient to completely cut off the sound beat frequency in the higher adjacent channel (15.25 mc on the graph) and to cut off the video beat frequency in the lower adjacent channel (31.75 mc on the graph).

At the left in Fig. 46-8 is the overall response of an i-f amplifier in a receiver having a dual sound system. There should be no amplification of the accompanying sound frequency, nor of adjacent sound or video frequencies. Actually there is gain at all these unwanted frequencies. From zero at the low end to zero at the high end the frequency range is about 11 megacycles.

Gain at the sound intermediate frequency is nearly half of maximum. Even though the sound takeoff follows the mixer in this receiver, fairly strong sound signal voltage will pass on into the i-f amplifier and the sound signals will be strongly amplified. These sound signals would cause dark horizontal bars, called sound bars, on the screen of the picture tube. To prevent such trouble the gain of the i-f amplifier must be dropped almost to zero at the sound intermediate frequency. There is considerable gain also at the adjacent sound and video frequencies. Adjacent sound frequency, when amplified, can cause the same kind of trouble as accompanying sound. Adjacent video frequency can cause unsatisfactory picture reproduction. The gain at both adjacent frequencies should be brought down practically to zero. When all these reductions are made, the overall i-f response will be as shown by the full-line curve at the right in Fig. 46-8.

3 *WAVE TRAPS.* The dips where gain is dropped almost to zero in the response at the right are put there by means of wave traps. A wave trap is a parallel resonant circuit or a series resonant circuit connected at some point in the i-f amplifier system where the trap will reduce signal strength at an unwanted frequency.

The elementary principles of wave traps are illustrated by Fig. 46-9. In diagram 1 there is a parallel resonant trap in the line between the plate of one tube and the grid of another. A parallel resonant circuit has maximum impedance at the freqnency for which it tuned. Therefore, this trap will greatly reduce the strength of signals at its resonant frequency when such signals try to pass from one tube to the other.

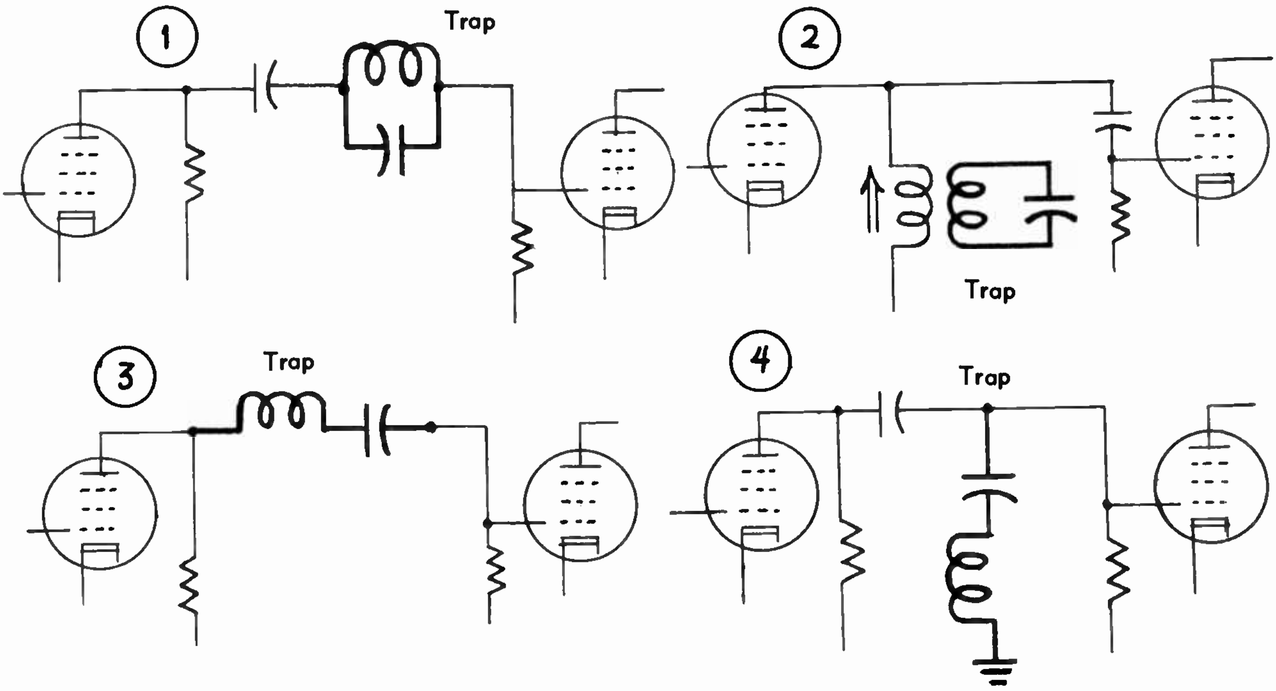


Fig. 46-9. Connections which illustrate the principles of trap circuits.

Signals at other frequencies pass through the parallel resonant trap circuit with relative freedom. A trap of this general type may be inserted in any line which would carry the frequency which is to be weakened or eliminated.

In diagram 2 a parallel resonant trap circuit is inductively coupled to a tuning coil which is between the plate of one tube and the grid of a following tube. This coupling will allow induction of emf's in the coil and capacitor of the trap. At the frequency to which the trap circuit is tuned its induced currents will be large, and the energy to maintain these currents will be taken from the tuning coil and dissipated in the trap. Signal currents at the frequency to which the trap is tuned thus will be greatly weakened and and will reach the grid of the second tube with little or no strength. Other signal frequencies will suffer little absorption, and will pass from tube to tube with almost full strength. This type of parallel resonant trap may be coupled anywhere in the signal-carrying circuits.

Fig. 46-10 pictures one way of constructing an inductively coupled trap. The bottom winding on the tubular form is that for an ordinary impedance coupler, which is tuned by a movable core whose threaded stud extends down below the mounting. The upper winding is that of the trap. To the ends of this trap coil is connected the small ceramic capacitor which appearsonthe left-hand side of the form. Thus we have, in practical form, the trap circuit illustrated by diagram 2 of Fig. 46-9. In addition, the actual trap has a movable core for tuning it to whatever frequency is to be weakened or eliminated. The threaded stud for tuning the trap extends from the top of the winding form.

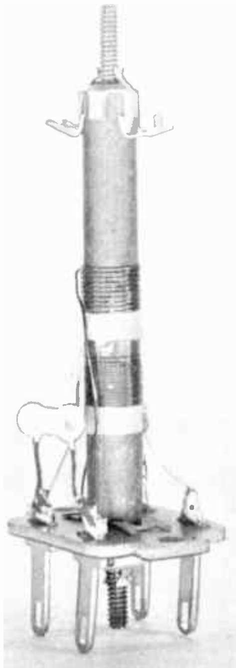


Fig. 46-10. Impedance coupler with inductively coupled trap.

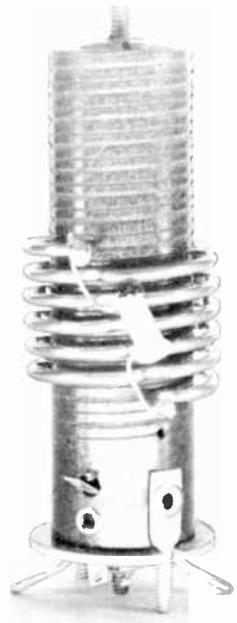


Fig. 46-11. The large winding with its connected capacitor is a trap circuit.

Fig. 46-11 shows another way of constructing the same type of inductively coupled trap. Inside the large tubular form is the usual kind of impedance coupler, with a movable core whose adjusting stud comes out through the bottom of the support. The coil of large wire around the outside of the form is the trap winding. A small ceramic capacitor is connected between the ends of the trap winding. The trap is tuned to its own frequency by a second movable core whose adjusting stud extends from the top of the form. Fig. 46-12 is a picture of the parts of this trap. At the left is the small impedance coupler which goes into the bottom of the large form. At the right is the movable core that tunes the trap to whatever frequency is to be attenuated.

Now we may go back to Fig. 46-9 and look at some other elementary trap circuits. In diagram 3 there is a series resonant circuit in the line from the plate of one tube to the grid of a following tube. The impedance of a series resonant circuit is minimum at the frequency for which the circuit is tuned, and this frequency will pass through with hardly any opposition, while all other lower and higher frequencies will be attenuated in greater or less degree. This type of trap would be tuned to the frequency which is not to be weakened. In a video i-f amplifier the series resonant circuit would be tuned to the video intermediate frequency.

In diagram 4 of Fig. 46-9 a series resonant trap circuit is connected from the line between two tubes to ground. Since the series resonant circuit has minimum impedance at its tuned frequency this trap grounds its tuned frequency, or short circuits this frequency to ground and greatly weakens or attenuates it. The trap offers relatively high impedance at all other lower and higher frequencies, which then pass from the

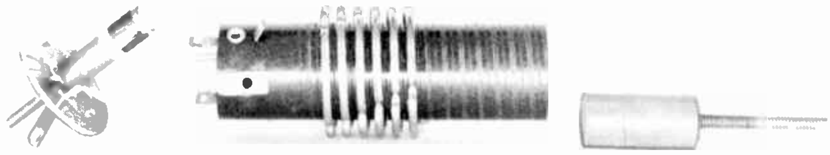


Fig. 46-12. The small impedance coupler is mounted inside the form that carries the trap coil.

plate of the first tube to the grid of the second tube. This grounding trap may be connected to any other points in the tube circuits from which a frequency is to be removed.

Variably tuned traps often are constructed with an adjustable capacitor and a fixed inductance. One method of making a parallel resonant trap is shown by Fig. 46-13, where a small self-supporting coil of enameled wire is across the terminals of a trimmer capacitor having mica dielectric. The coil and capacitor could be connected end to end to form a series resonant trap circuit. A trap may be constructed also with a coil such as illustrated connected across a fixed capacitor. A limited range of tuning adjustment is possible by spreading the coil turns farther apart for less inductance and higher frequency, or by squeezing the turns closer together for more inductance and a lower tuned frequency.

Fig. 46-14 shows some actual television circuits in which there are parallel resonant traps offering high impedance and causing maximum attenuation of their tuned frequencies. The trap circuits are drawn with heavy lines. In diagram 1 the trap is between the plate of the first tube and the grid of the second tube. The trap is tuned to resonance by a capacitor. For signal coupling between the tubes there is a

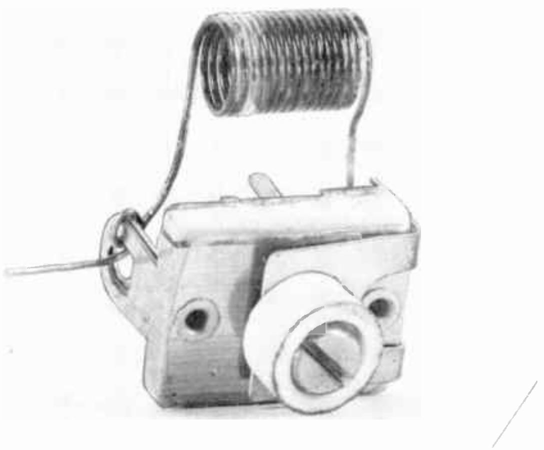


Fig. 46-13. A trap may be assembled from a coil and a trimmer capacitor.

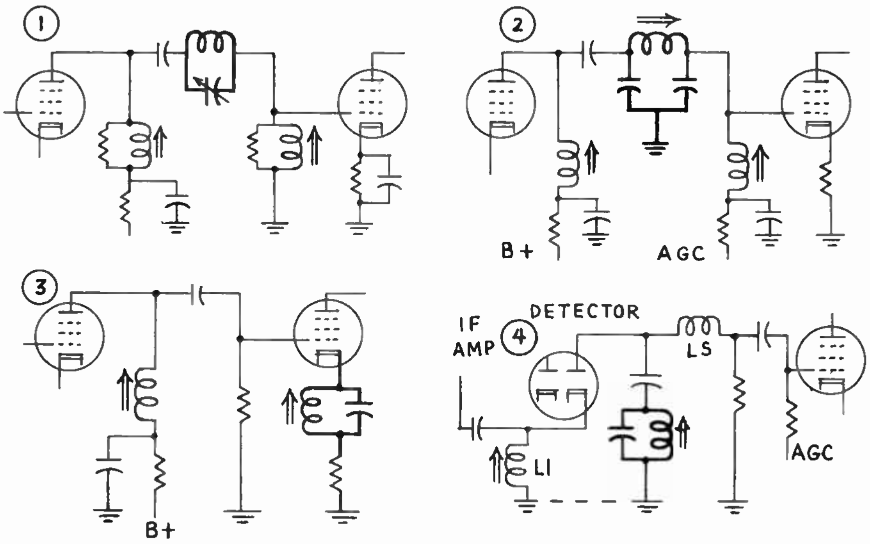


Fig. 46-14. Connections for parallel resonant traps.

tuned impedance coil in the plate circuit and another one in the grid circuit, with each coupling coil shunted by a broadening resistor.

In diagram 2 the parallel resonant trap circuit is of different construction. The trap coil is tuned by two capacitors in series. From between the capacitors there is a connection to ground. This ground connection drains away signal energy at the undesired frequency which is to be attenuated by the trap.

Diagram 3 shows a parallel resonant trap in the cathode line of an amplifier tube. Signal currents which are in both the grid circuit and the plate circuit of an amplifier must flow also in the cathode circuit. Therefore, with this trap tuned to an undesired frequency that particular frequency is attenuated in the grid circuit and also in the plate circuit of the amplifier.

In diagram 4 a parallel resonant trap is connected to the output of a video detector. Intermediate frequencies coming through the detector are opposed by the high impedance of inductor Ls. These intermediate frequencies, which are being rectified by the detector, must pass through the i-f coupler Li, the diode tube, the trap, and the ground connections between trap and coupler. Any intermediate frequency to which the trap is tuned will be greatly attenuated. For example, if the trap is tuned to the accompanying sound frequency, this frequency will be attenuated, will not remain to be rectified or detected, and will be kept out of the following video amplifier.

Fig. 46-15 illustrates several ways in which traps may be coupled to tube circuits in which some unwanted frequency is to be attenuated. Diagram 1 shows two traps which are inductively coupled to impedance couplers in two successive stages of an amplifier. The trap in the first stage is on a coupler connected into the plate circuit of the tube at the left. The trap in the second stage is on a coupler connected into the grid circuit of the tube at the right. One of these traps might be tuned to the accompanying sound and one to the adjacent sound frequency, or both might be tuned to the same frequency.

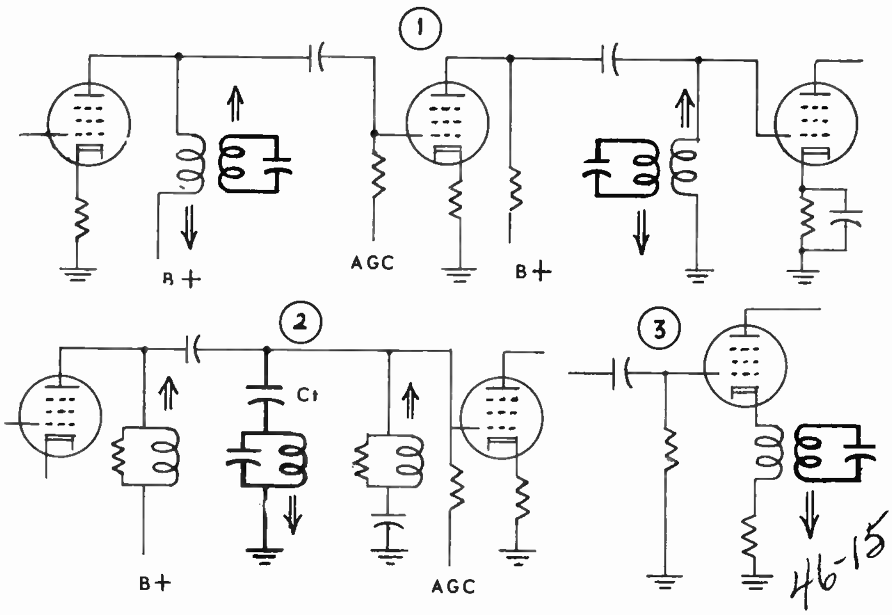


Fig. 46-15. Traps inductively or capacitively coupled to grid and plate circuits.

The trap of diagram 2, shown in heavy lines, looks quite like a parallel resonant circuit connected between the high side of the two tuned couplers and ground, but it does not behave as a parallel resonant circuit. The trap is coupled to the plate-grid signal line through capacitor Ct, and takes energy from the signal line at the frequency for which the trap is tuned. This energy causes large current to flow in the coil and capacitor of the trap, and the energy in this current is carried away to ground through the connection from the bottom of the trap. Thus we have great attenuation of any frequency to which the trap is tuned.

In diagram 3 a tuned trap circuit is inductively coupled to an inductor or coil in the cathode lead of an amplifier tube. Energy taken from the cathode circuit induces emf and current in the coil and capacitor of the trap at the frequency for which the trap is tuned, and this energy is dissipated in maintaining flow of the induced current. The effect is much the same as with the parallel resonant trap of diagram 3 in Fig. 46-14. In that earlier trap circuit there is high impedance at the trapped frequency, while with the inductively coupled trap there is absorption of energy at the trapped frequency.

Fig. 46-16 shows two ways of using series resonant trap circuits. In diagram 1 the series resonant circuit, shown by heavy lines, is for the purpose of keeping out of the i-f amplifier all frequencies except the range of intermediate frequencies which should be amplified. The series resonant circuit is tuned to have minimum or low impedance in this range. It then has relatively high impedance at other frequencies coming from the mixer. These other frequencies include the r-f oscillator frequency, the carrier frequency, and the sum frequencies.

The impedance coupler Lp is tuned to the desired intermediate frequencies to give them maximum gain,

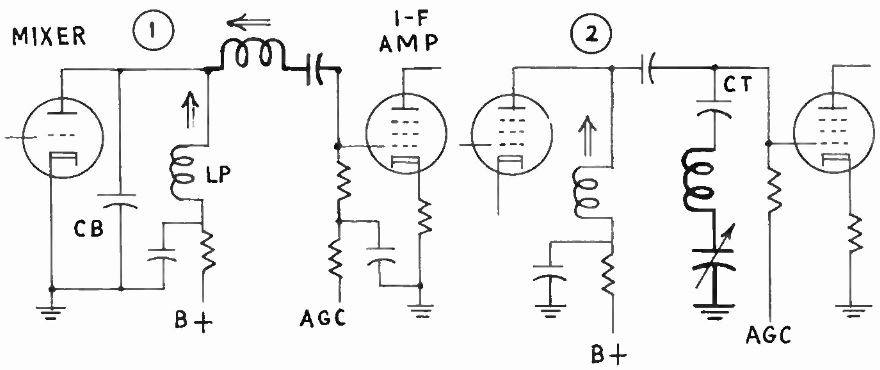


Fig. 46-16. How series resonant traps may be connected.

while the series resonant circuit is tuned to freely pass these frequencies from mixer plate to amplifier grid. The frequencies which are opposed by the series resonant circuit are bypassed to ground and the mixer cathode through capacitor Cb.

In diagram 2 of Fig. 46-16 there is a series resonant trap circuit from the grid of the second amplifier tube to ground. This circuit is coupled to the grid line through capacitor Ct. The frequency to which the trap is tuned meets very low impedance in the series resonant circuit, and is shorted to ground. The trap circuit has relatively high impedance to lower and higher frequencies, so they go to the amplifier grid. The trap is shown as being tuned by a trimmer capacitor, but it might equally well be tuned by a movable core in the coil.

WHERE TRAPS ARE USED. Unless the overall response of an i-f amplifier is wide enough to cover sound and video frequencies developed from adjacent channel signals there is no need of traps for these frequencies. Some video i-f amplifiers, even in the larger receivers, do not have a response wide enough to reach the adjacent frequency, and no adjacent sound trap is needed. In some receivers having only two or three i-f amplifier stages handling video signals the response curve is rather sharply peaked and has steep skirts with relatively sharp cutoff of frequencies. Then there is no danger that the response will reach the adjacent video frequency, and no adjacent video trap is used.

Receivers having 7-inch picture tubes ordinarily have an i-f band width covering no more than three megacycles, and the response has sufficiently high peaking and sufficiently steep skirts to prevent reaching any adjacent frequencies. The i-f amplifiers of these receivers usually have no traps.

The i-f amplifiers of receivers employing intercarrier sound systems have no traps for accompanying sound, because the sound intermediate frequency must go all the way through the i-f amplifier system to the detector, where it beats with the video intermediate frequency. As a general rule these receivers have no traps for either of the adjacent frequencies, sound or video.

When the response of an i-f amplifier is normally wide enough to cover the full video frequency range up to four or more megacycles there would be amplification of the adjacent sound frequency. One or more traps for accompanying sound then are required. Quite often these broad i-f responses would reach the adjacent video frequency, and a trap is needed for this frequency.

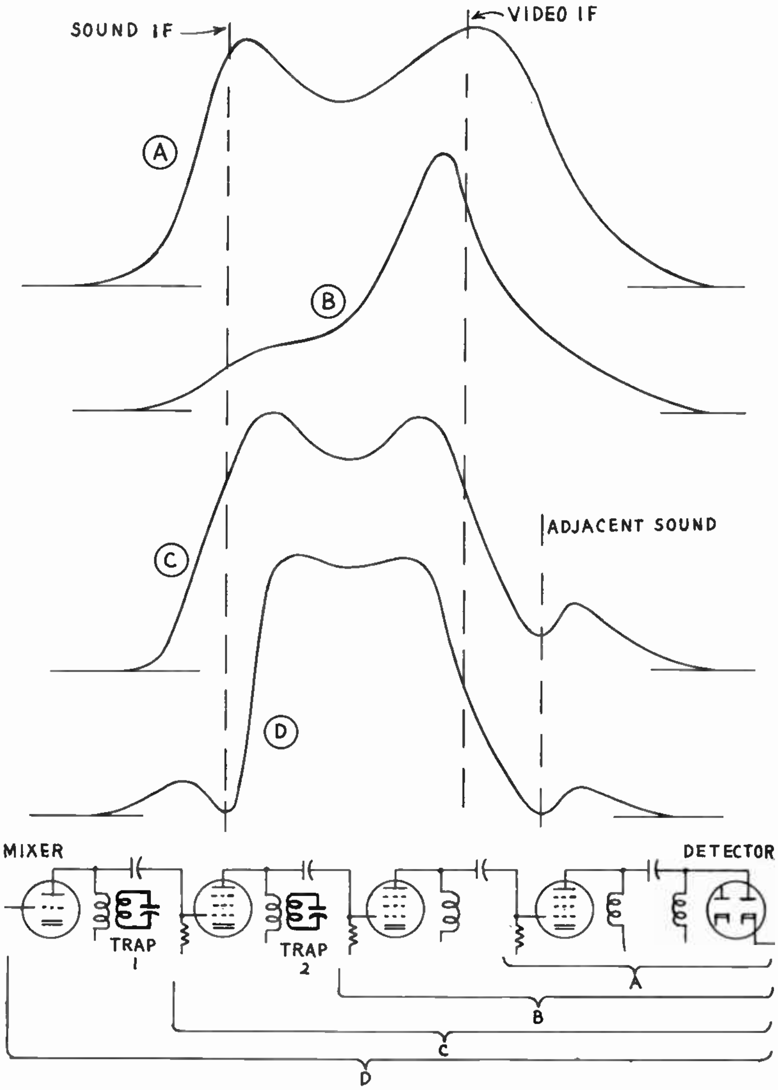


Fig. 46-17. How traps for adjacent and accompanying sound affect the shape of an i-f response curve.

Where the sound takeoff does not immediately follow the mixer, but is placed after one or more i-f stages, there must be no accompanying sound trap in the stages which carry both sound and video intermediate frequencies. Accompanying sound traps, if used at all, will be in i-f stages which are between the sound takeoff and the video detector. Traps for adjacent sound and for adjacent video may be placed anywhere in these i-f amplifiers.

If you look back at the table of television channel frequencies in an earlier lesson you will see that there is no other channel immediately adjacent below channels 2, 5 and 7, nor above channels 4, 6, and

13. With a receiver tuned to these channels there would be no adjacent sound interference or else no adjacent video interference. But when tuned to other channels there might be either or both of these interferences, and since a receiver may be tuned to any channel at all, the absence of immediately adjacent channels at some points has no bearing on the use of traps for the adjacent frequencies.

Traps are not needed in i-f amplifiers for sound because such amplifiers have a response which is very narrow in comparison with a video i-f response, and the resulting selectivity is enough to keep our video and adjacent channel frequencies.

The exact frequencies, in megacycles, to which traps in any receiver are adjusted or aligned depends on the intermediate frequencies used in that particular receiver. This will become apparent from an examination of Fig. 46-7. Were the intermediate frequencies to be anything other than those used for illustration, the r-f oscillator frequencies for each channel would be correspondingly changed. The r-f oscillator frequency always must differ from received carrier frequencies by the amount of the intermediate frequencies. With different oscillator frequencies there would be different beat frequencies to which the various traps would have to be tuned. Trap frequencies always have the following relations to video and sound intermediate frequencies used in the receiver.

Adjacent sound trap	= video intermediate freq.	+	1.5 mc
or	= sound intermediate freq.	+	6.0 mc
Adjacent video trap	= video intermediate freq.	–	6.0 mc
or	= sound intermediate freq.	–	1.5 mc
Accompanying sound	= video intermediate freq.	–	4.5 mc

Fig. 46-17 shows the effects of traps on frequency responses of a particular receiver. The response curves are shown at A, B, C, and D. Locations of couplers and traps in the i-f amplifier are shown by the simplified circuit diagram at the bottom of the figure. The oscilloscope remains connected to the output of the video detector for all tests, while the signal generator is connected to the grid of the last i-f amplifier when taking curve A, to the grid of the preceding amplifier for curve B, and so on back until the generator is connected to the grid of the mixer for curve D, which is the overall response of the amplifier. In an earlier lesson we looked at frequency responses from an amplifier generally similar to this one, except for having no traps. Here we have, between mixer and first video amplifier, a trap tuned to the accompanying sound frequency. We may call this trap number 1. Between first and second i-f amplifiers is another trap, number 2, which is tuned to the adjacent sound frequency.

Curve A is the response of the coupler between the last i-f amplifier and detector. This is a broad band coupler giving a double-peaked response. Curve B shows the response when we add the single-peaked impedance coupler which is between the last two i-f amplifiers. This coupler adds a high peak at its tuned frequency. So far we have included neither of the two traps in the signal-carrying circuits.

Curve C shows the response when we have included the effect of the single-peaked coupler between the first two i-f amplifiers. This coupler adds a second peak at its own frequency. Here we have also the effect of the adjacent sound trap, number 2, which is taking energy from the circuit at the trap frequency. The result is a pronounced dip in the response at the adjacent sound frequency.

Curve D is taken with the signal generator connected to the mixer grid. Now we have the effect of the accompanying sound trap, number 1, which puts a sharp dip in the response at the accompanying sound frequency, which is the same as the sound intermediate frequency. On the other side of the response, on the higher-frequency side, we still have the dip caused by the adjacent sound trap. This overall response is like the one we set out to obtain away back in Fig. 46-8, except that we have no adjacent video trap effect. To obtain a dip in the response at the adjacent video frequency we could add a third trap on one of the other couplers, and tune it to the adjacent video frequency.

TRAPS FOR SPECIAL PURPOSES. Traps are most commonly used to attenuate certain frequencies in the overall response of video i-f amplifiers. When used for this purpose a trap must be tuned or aligned at the exact frequency it is supposed to attenuate or weaken, regardless of how this may affect the form of the response at other frequencies.

It is true, however, that traps may alter the general form of the response in ways that are desirable while still performing their primary purpose of cutting down on the trapped frequencies. As one example, in Fig. 46-17, the gain at the video intermediate frequency would come too high were it not for the effect of the adjacent sound trap in making its side of the response come down more sharply. Also, the accompanying sound trap practically eliminates the sound intermediate frequency or accompanying sound frequency while allowing the gain to remain as high as possible on its side of the response. It is in this part of the response that we have the higher video frequencies, and maintaining high gain in this region helps the fine detail or the definition of pictures.

In some receivers there are trap circuits for the express purpose of helping to shape the response in some desirable manner. Such traps are not tuned to a frequency which is to be removed from the response, but to whatever frequency causes the desired result. In some receivers a trap of this nature is used for the video intermediate frequency to a point of 50 to 60 per cent gain on the response.

The manner in which a trap alters the shape of the overall response depends largely on the Q-factor of the trap and on how it is coupled to an amplifier circuit. The higher the Q-factor of the trap itself the sharper and narrower will be the dip that it produces. The looser the coupling of the trap to the signal-carrying circuits the sharper will be the dip and the steeper will be the slopes of the response on both sides of the trapped frequency. Coupling which is very loose also will lessen the degree of attenuation at the trapped frequency.

In some i-f amplifiers there may be undesirable peaks of high gain over very narrow ranges of frequency. Such peaks may result from beat frequencies, or harmonic frequencies, or from feedback of energy from one stage to a preceding stage at a frequency for which some plate or grid circuit has high impedance. Traps sometimes are provided for the purpose of reducing these peaks of gain so that the response may be made more nearly uniform throughout the whole range normally covered by the i-f amplifier.

Wave traps quite often are coupled to the r-f amplifier circuits of a-m and f-m sound receivers. Many of these traps are for the purpose of attenuating the intermediate frequency of the receiver itself, which would be a frequency of 455 or 456 kilocycles for a-m sound receivers, or a frequency of 10.7 megacycles for f-m sound receivers. Traps on the r-f amplifier circuits may be provided also for attenuating the carrier frequencies of other services. For instance, an a-m sound receiver may have a trap for attenuating the carrier frequency of some nearby f-m transmitter.

DO NOT TEAR — CUT ALONG THIS LINE AND SEND IN FOR GRADING.

TELEVISION

LESSON NO. 47

BROAD BAND AMPLIFIERS

Training Your Mind...

C.T.I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind
C.T.I.
Perfecting Your Skill

COMMERCIAL
TRADES
INSTITUTE

Chicago, Ill.

LESSON NO. 47

BROAD BAND AMPLIFIERS

5 Now we are ready to examine the video amplifier of the television receiver. The video signal, which is the output of the video detector, goes through the video amplifier to the picture tube and there controls the rate or intensity of flow of electrons in the beam. This produces picture lights and shadows to look at. The video amplifier of the television receiver has much the same relation to other parts of this receiver as has the audio amplifier of a sound receiver to other parts of the sound receiver. The audio signal, which is the output of the sound detector, goes through the audio amplifier to the loud speaker and there controls motion of the speaker cone or diaphragm that produces sound waves to listen to.

The similarity between video and audio amplifiers goes even further. In the great majority of sound receivers and in the sound sections of television receivers the audio amplifier is a resistance coupled type, and in practically all television receivers the video amplifier is basically a resistance coupled type. Up to this time we have made no detailed study of the audio amplifiers of sound receivers or of television receivers. Since these are quite similar to the video amplifiers of television receivers we may learn about all three amplifiers in the least possible time by studying them right along together, and that is what we shall do.

At the top of Fig. 47-1 are the parts of a simple resistance coupled amplifier stage as set up for testing the performance. Down below is the corresponding circuit diagram. Input voltage for the grid of the amplifier tube comes from a resistor which is fed with signal current through a capacitor at the left. This resistor and capacitor are shown by broken lines because they are not parts of the amplifier stage to be examined. The amplifier stage is considered as commencing at the grid of the tube.

Connected between the plate of the amplifier tube and the B-supply is a load resistor Ro. Amplified signal voltage developed across this resistor causes signal current to flow through coupling capacitor Cc. This signal current goes through resistor Rg, where it is accompanied by a corresponding amplified signal voltage which is applied to the grid of the following tube. Resistor Rg would be the grid resistor connected to this following tube. Our amplifier stage is considered to have its output, and to end at the top of resistor Rg. We commence at the amplifier grid with one signal voltage, and end at the grid of the following tube with this signal voltage amplified.

7 When designing and constructing a resistance coupled amplifier we are chiefly concerned with the voltage gain which may be obtained without distortion of the signal to any greater extent than can be tolerated in the particular application. The gain will vary with frequency. It will be least at the low frequencies, will rise to maximum through a range called mid-band frequencies, and will drop off at the high frequencies.

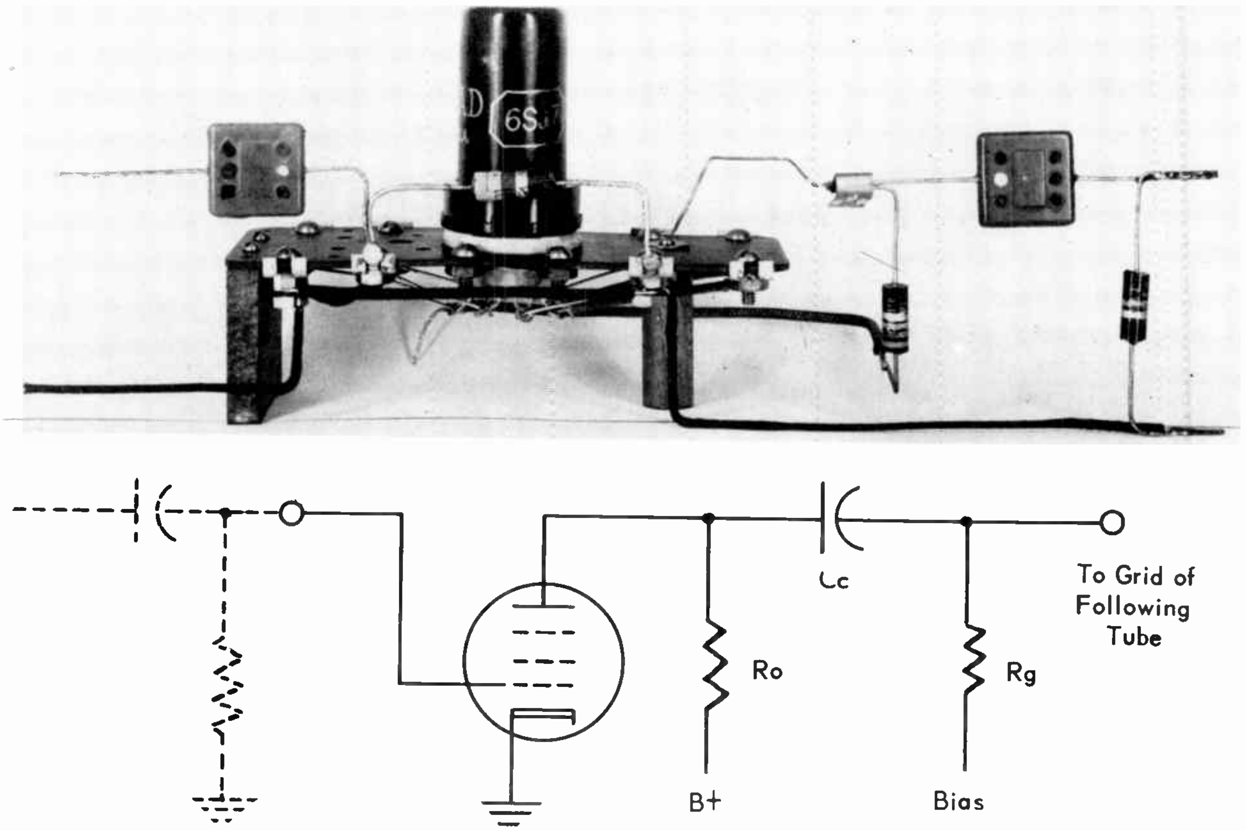


Fig. 47-1. The parts of a resistance coupled amplifier and the corresponding circuit diagram.

The lowest frequencies handled by usual types of audio amplifiers are on the order of 80 to 100 cycles per second, and the highest are around 5,000 cycles. In high-fidelity audio amplifiers the frequency range may be from as low as 50 cycles to as high as 10,000 or 15,000 cycles per second, or even higher. High-quality television video amplifiers provide fairly uniform gain over a tremendous ratio of frequencies, from as low as 30 or 40 cycles all the way up to 4 million cycles or 4 megacycles per second. In other video amplifiers we may have fairly uniform gain from 50 or 60 cycles up to about 2½ or 3 megacycles per second, which still is a very great ratio of highest to lowest frequencies.

All the parts shown by Fig. 47-1 have more or less effect on gain and distortion at all frequencies, but performance of resistance coupled amplifiers is affected also by a number of other things which are represented in Fig. 47-2. In diagram 1 we have the amplifier circuit as usually shown. From the bottom of load resistor Ro there is a bypass capacitor Cb to ground, and a connection through ground back to the tube cathode. Reactance of this bypass capacitor at amplified frequencies is quite small, and we may consider the audio or video signal circuit on the plate side as being completed through the bypass back to the tube cathode rather than through the B-supply wiring. A similar low-reactance capacitor is connected from the bottom of grid resistor Rg to ground. This capacitor completes the signal circuit on the grid side or output side of the amplifier.

At mid-band frequencies the reactance of coupling capacitor Cc is so low in comparison with the circuit resistances that we may disregard it so far as signal paths are concerned, and may draw the amplifier as in diagram 2 with no coupling capacitor. The bypass capacitors are omitted because, for signal currents, they form short circuits to ground and the tube cathode. We consider the tube as consisting of two parts. One part is the internal plate resistance which is here represented as Rp. The other part includes all that provides amplification or transconductance, which now is represented by the symbol Gm. In this "equivalent circuit" the part of the tube which puts out the amplified signal may be considered as the signal source or as an a-c generator of signal current.

Diagram 2 shows that plate resistance, load resistor resistance, and grid resistor resistance are in parallel with one another. Then the effective plate load for the amplifier tube is the combined parallel resistance of these three resistances. This is the load which we must consider as existing for mid-band frequencies. Its value is less than that of any one of the separate resistances. The voltage gain at these

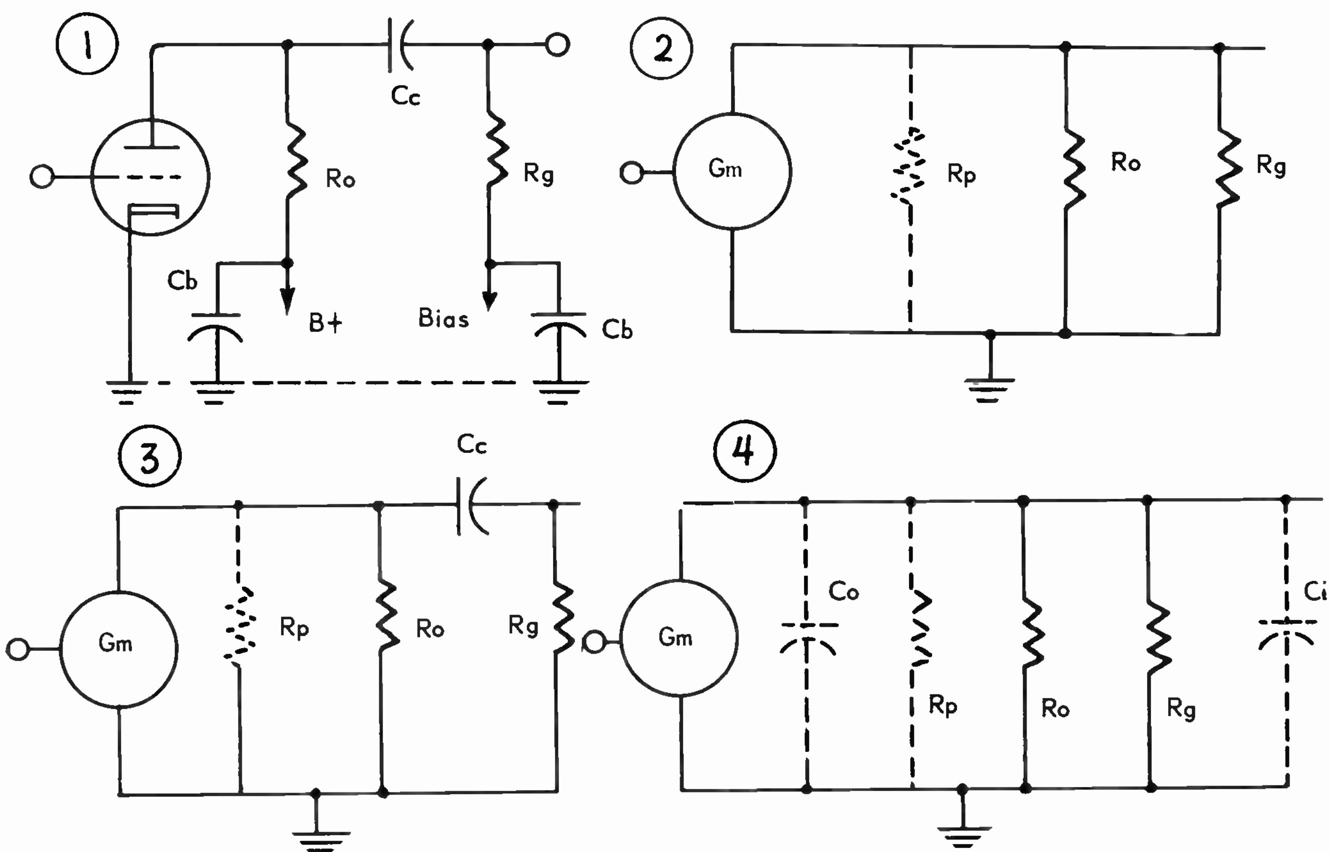


Fig. 47-2. Equivalent circuits for a resistance coupled amplifier.

frequencies is equal approximately to transconductance or mutual conductance of the tube multiplied by combined or parallel load resistance. It follows that gain will be increased by doing any or all of three things; we may use a tube with greater plate resistance, use a higher value of load resistor, or use a higher value of grid resistor.

When amplifying the lowest frequencies the equivalent circuit is as shown by diagram 3 of Fig. 47-2. We still have the three parallel resistances as before, but now the reactance of coupling capacitor Cc cannot be neglected. As frequency goes down this reactance goes up, and offers more and more opposition to signal currents flowing to grid resistor Rg. Of the total signal voltage being developed across the plate load we are losing more and more in the reactance of Cc and leaving less to produce output voltage across resistor Rg. Voltage output or gain will decrease at low frequencies unless the reactance at Cc is reduced proportionately, which means using capacitance inversely proportional to frequency - the lower the frequency the greater must be this capacitance if gain is not to drop off.

Diagram 4 of Fig. 47-2 shows the equivalent amplifier circuit for high frequencies. Here we have the circuit elements which become of major importance in television video amplifiers. The three paralleled resistances remain as before. But on the plate side there is the output capacitance of the amplifier tube and all the stray and distributed capacitances of the plate circuit. On the grid side there is the input capacitance of the following tube, and other stray and distributed capacitances. These capacitances, represented at Co and Ci are in parallel with the load resistance.

As frequency goes up the reactance of the paralleled capacitances goes down. Since all the reactances and resistances are in parallel with one another the total effective impedance never can be greater than that of the two capacitances in parallel. Suppose, for example, each capacitance were only 30 mmf. The two in parallel would come to 60 mmf. Their reactance at 5,000 cycles would be somewhat more than a half megohm, which would not reduce the total effective plate load below a value giving good gain. But at a frequency of 4 megacycles the capacitive reactance would be about 667 ohms, which would bring the amplification away down. With such capacitance the effective plate load never could be more than 667 ohms even though all the load resistances could be made many megohms - for impedance of any untuned parallel circuit always is less than that of the least of its separate reactances or resistances.

The effective band width of a resistance coupled amplifier usually is taken as the range between low and high frequencies at which gain drops to 70 per cent of maximum. Frequency response curves for any and all simple resistance coupled amplifiers are of the general form shown by Fig. 47-3. The only parts of the curve included in the figure are the end portions in which frequency falls off. On the low end there is 70 per cent gain at a frequency of 100 cycles, and on the high end there is 70 per cent gain at 10,000 cycles. Therefore, the band width as here represented is from 100 to 10,000 cycles. With an amplifier designed for this band width the gain would rise to 100 per cent or maximum at about 500 cycles, would remain at approximately this value until about 2,000 cycles, and would drop as shown at higher frequencies. No matter how great or how small the band width, gain at low and high frequencies would drop off proportionately just about as shown by this typical curve.

EFFECTS OF RESISTORS AND CAPACITORS. Gain and distortion of the resistance coupled amplifier are affected in various ways by the load resistor, grid resistor, coupling capacitor, and, when cathode bias is used, by the bypass capacitor for the biasing resistor. Whether or not the load resistor, Ro in the diagrams, has much effect depends on the kind of tube used. If the tube is a medium-mu triode its plate

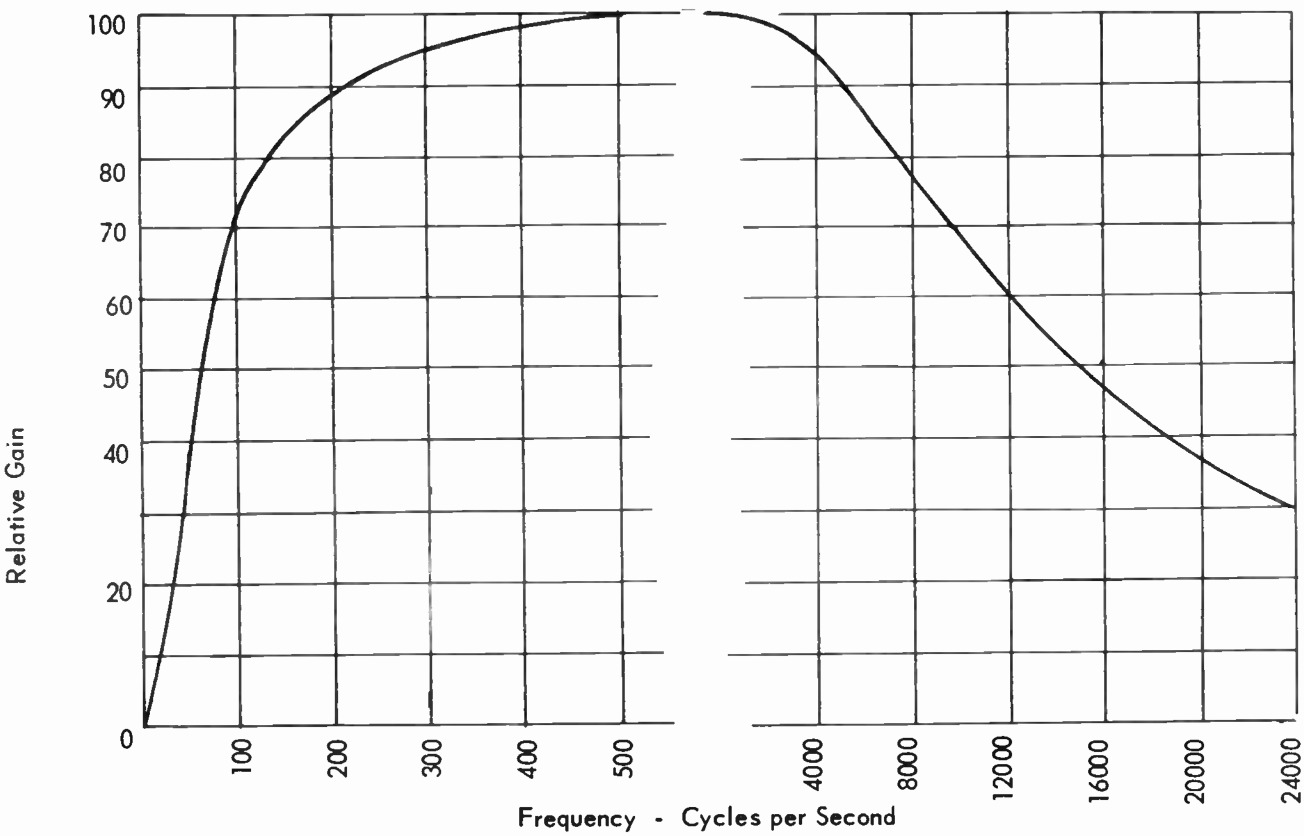


Fig. 47-3. Typical frequency response of a resistance coupled amplifier.

resistance will be something like 7,000 to 12,000 ohms, and no matter how great the load resistor is made it cannot bring the effective load resistance any higher. With high-mu triodes, having plate resistances of 50,000 to 100,000 ohms, the load resistor will be 100,000 to 500,000 ohms, and its value will have a decided effect on the total parallel load resistance. With pentodes there will be plate resistance of 200,000 ohms to one megohm, and the load resistor will largely determine the total load resistance.

These relations between plate resistances and load resistors are illustrated with typical values in Fig. 47-4. At the left we have a medium-mu triode with plate resistance of 9,000 ohms. Increasing the load resistor from 100,000 to 500,000 ohms raises the effective parallel resistance only from 8,260 ohms to 8,840 ohms, an increase of about 7 per cent. At the center the load resistor is changed in the same way with a medium-mu triode whose plate resistance is 80,000 ohms. The effective load is raised from 44,400 ohms to 69,000 ohms, an increase of about 57 per cent. At the right is a pentode having plate resistance of 500,000 ohms. The same change of load resistor increases the effective load by 300 per cent, from 83,400 ohms to 250,000 ohms.

A higher load resistor increases the gain only if plate current is kept high enough to maintain the trans-

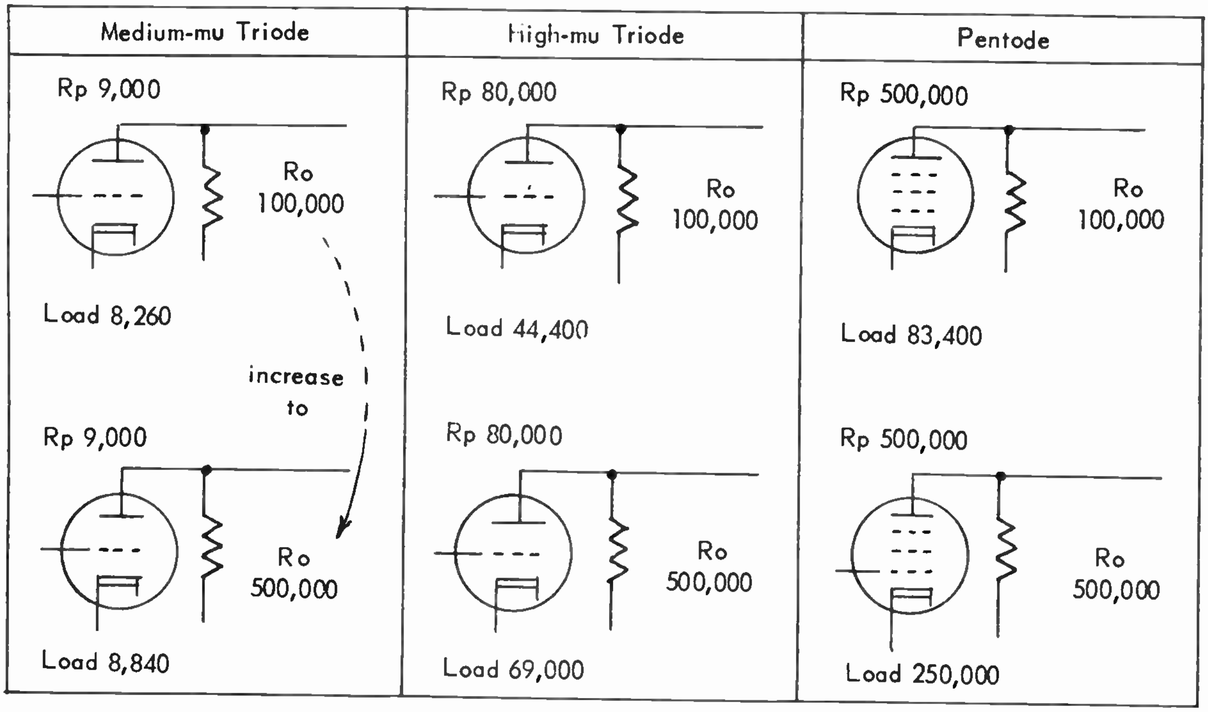


Fig. 47-4. How increasing the value of the load resistor alters the effective load resistance with various plate resistances.

conductance. Unless B-supply voltage is raised, the voltage drop through the higher load resistor will reduce plate voltage, plate current, and transconductance - thus wholly or partially neutralizing the advantage of the higher load resistor. With any given type of tube, variations of the load resistor tend to have much more effect at mid-band and high frequencies than at low frequencies. Using a smaller load resistor reduces the gain, but at the same time it allows fairly uniform gain over a wider range of frequencies or increases the band width. This is especially true with pentode amplifier tubes.

As mentioned earlier, the coupling capacitor marked Cc in the diagrams has a great effect on low-frequency gain or on the lowest frequency at which there is any given gain. Capacitances most often are somewhere between 0.005 and 0.02 mf. At a frequency of 30 cycles the reactance of 0.005 mf is more than a megohm, and a large part of the signal voltage would be used in getting through this reactance. Were the following grid resistor to have a value of one-half megohm, two-thirds of the signal voltage would be used up in the capacitor and only one-third left across the grid resistor. The greater the coupling capacitance the less will be its reactance and the greater will be the percentage of possible voltage developed across the following grid resistor and applied to the grid of the following tube.

In some receivers whose power supply is poorly filtered, to leave considerable ripple voltage at 60 or 120 cycles, the coupling capacitor is intentionally made rather small to limit amplification of these hum frequencies. It should be mentioned that the coupling capacitor must not allow appreciable leakage

Fig. 47-5. Test setup for checking the effect of coupling capacitors on gain at various frequencies.

through it of direct current from the B-supply connected to the amplifier plate. Leakage current would have to pass through the following grid resistor in such direction as to make the grid of the following tube positive, or at least less negative than it should be for correct biasing.

Relations between the coupling capacitor and following grid resistor may appear more clearly upon examination of Fig. 47-6. In diagram 1 the combination of tube transconductance, plate resistance, and load resistor is such that 1.20 volts of signal is developed across the two resistances. Coupling capacitor Cc and grid resistor Rg really are in series with each other across this developed signal voltage. The voltage divides between Cc and Rg inversely as their impedances. We assume that capacitance at Cc is such that capacitive reactance is 50,000 ohms. With 100,000 ohms at Rg there will be 0.40 volt of the signal across the capacitor and the remaining 0.80 volt across the resistor. This 0.80-volt signal is applied to the grid of the following tube.

In diagram 2 the developed signal voltage is unchanged, the grid resistor Rg is unchanged, but capacitance at Cc has been made 5 times as great - to drop the capacitive reactance to 1/5 its former value or to 10,000 ohms. Now we have only 0.11 signal volt across the capacitor, and have 1.09 signal volts across resistor Rg and on the grid of the following tube.

In diagram 3 the capacitor has been changed back to its original small size, with reactance of 50,000 ohms. Grid resistor Rg has been increased from 100,000 ohms to 500,000 ohms. We have the same

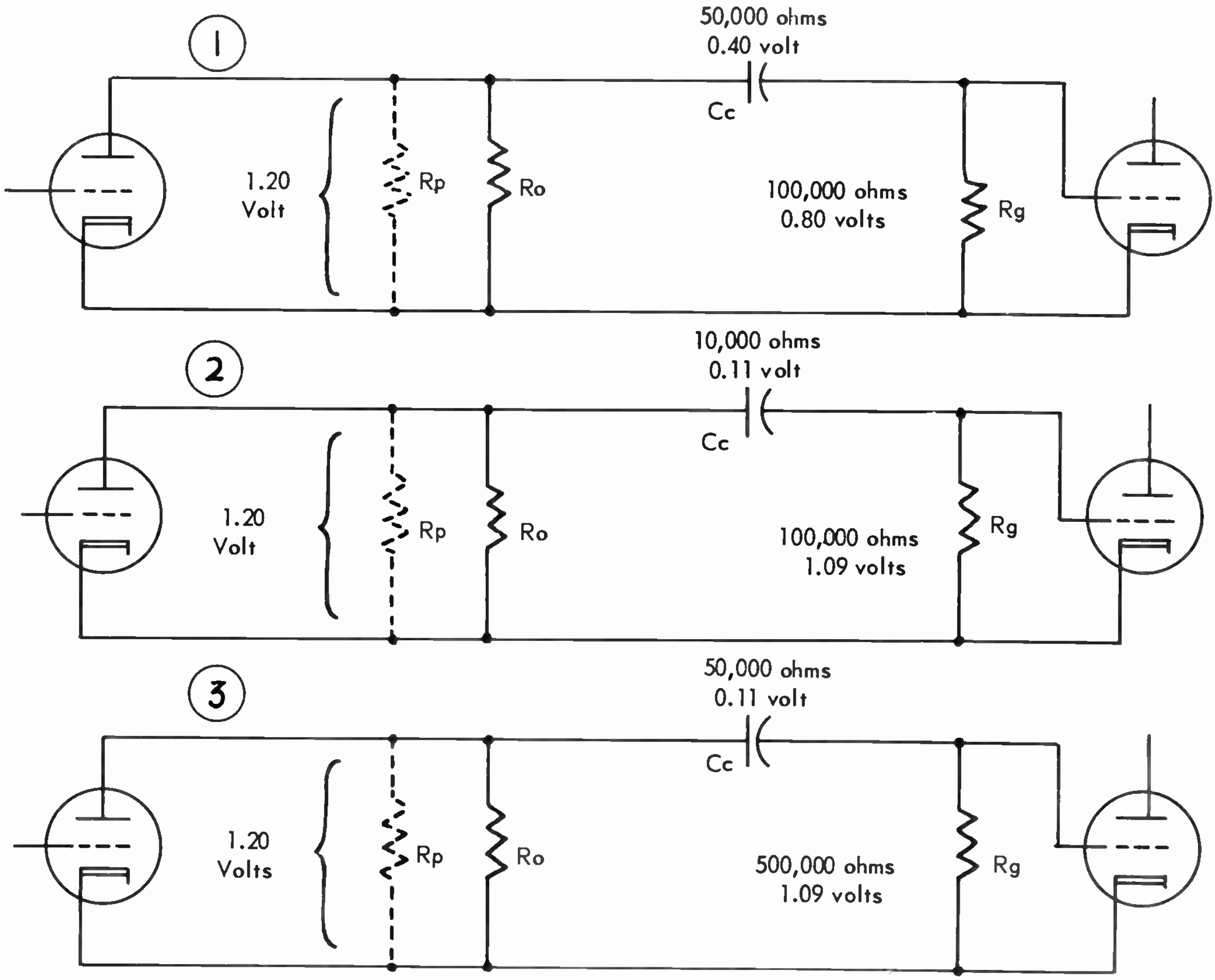


Fig. 47-6. How coupling capacitance and grid resistance affect the signal voltage applied to a following tube.

division of signal voltage between capacitor and resistor as in diagram 2. The reason is that the ratio of capacitive reactance to grid resistance is the same in both diagrams, it is 1 to 10, or 1/10.

Ordinarily we use the highest value of grid resistance which is permissible for the type of the following tube and the method used for biasing it. Higher grid resistance may be used with cathode bias than with fixed bias. Too much grid resistance will hold too many electrons on the grid of the following tube; they cannot leak off fast enough through the high resistance, and the grid is made too negative. The amount of resistance used at Rg has a decided effect on low-frequency gain, because this gain depends so much on the capacitance and reactance of Cc, and because there is such a close relation between the coupling capacitor and the grid resistor, as just explained.

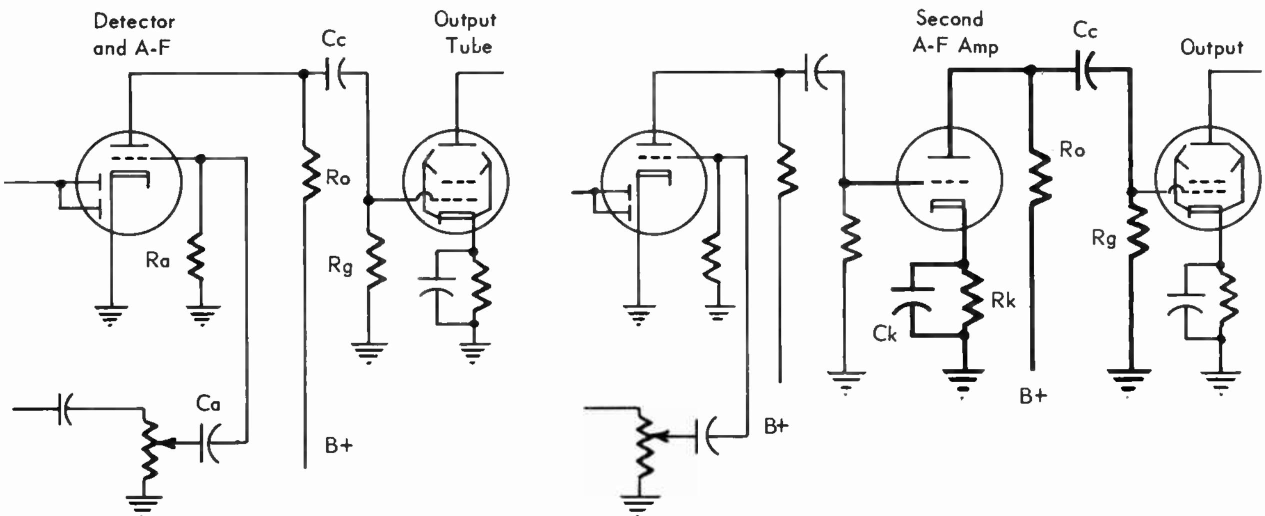


Fig. 47-7. Adding a resistance coupled stage in an amplifier.

In audio amplifiers, especially in those for a-m sound receivers, there ordinarily is only one resistance coupled stage. Such a stage is shown at the left in Fig. 47-7. The amplifier tube is a triode built into the same envelope with the diode detector. The output of this a-f amplifier is fed to the grid of the output tube which, in turn, feeds the loud speaker. The resistance coupling between a-f amplifier and output tube consists of load resistor Ro, coupling capacitor Cc, and grid resistor Rg. The audio amplifier triode has grid-leak bias, as provided by grid leak resistor Ra and grid capacitor Ca.

At the right in Fig. 47-7 we have added between the first a-f amplifier and the output tube a second a-f amplifier stage, whose parts and connections are shown by heavy lines. This added amplifier tube commonly would have cathode bias, by means of cathode resistor Rk and its bypass capacitor Ck.

When a resistance coupled amplifier tube is operated with cathode bias the value of capacitance in the bypass capacitor has much to do with the low limit of frequencies which will be amplified to a satisfactory level. A commonly used rule for determining a suitable capacitance for the bypass capacitor is to have its reactance at the lowest amplified frequency equal to 1/10 the number of ohms in the biasing resistor. That is, with a biasing resistor of 4,000 ohms, the reactance of the bypass capacitor should be no greater than 400 ohms at the lowest frequency to be satisfactorily amplified.

If this ratio of reactance to resistance is to be observed, the following simple formula allows computing the capacitance in microfarads.

$$\text{Bypass mf} = \frac{1\ 600\ 000}{\text{lowest frequency, (cycles)} \times \text{bias resistance, (ohms)}}$$

For an example, assume that the bias resistance is 4,000 ohms and that the lowest amplified frequency is to be 50 cycles. Here are the figures.

$$\frac{\text{Bypass}}{\text{mf}} = \frac{1\ 600\ 000}{50 \times 4000} = \frac{1\ 600\ 000}{200\ 000} = 8 \text{ mf capacitance.}$$

If you compute the reactance of 8 mf capacitance at 50 cycles per second it will come very close to 400 ohms, which is 1/10 of the biasing resistance in ohms. A capacitance giving this ratio of reactance to resistance usually is satisfactory for any type of tube with an a-c operated heater cathode. Much smaller capacitances are enough when using filament-cathode tubes with their filaments heated from batteries. Because of the large values of capacitance needed with a-c heaters the bypass capacitor usually is of the electrolytic type, with its negative terminal connected to ground and its positive terminal to the tube cathode pin on the socket.

When using pentode amplifiers with resistance coupling the low-frequency response is affected by the relation between resistance of the voltage dropping resistor for the screen and the reactance of the bypass capacitor for this resistor. These two parts are marked respectively Rs and Cs in Fig. 47-8. Capacitor reactance in ohms may be almost anywhere between 1/10 and 1/2 of the resistor value in ohms. The capacitor reactance is made less in circuits having relatively high resistances in load resistors and screen resistors, and having smaller coupling capacitances. In other words, the capacitance of the bypass is made larger under the conditions mentioned. Bypasses in common use range from 0.05 to 0.2 mf. These seemingly small capacitances, with their high reactances, are sufficient for the purpose because of the high values of dropping resistance which are being bypassed.

PHASE DISTORTION. In resistance coupled amplifiers, and for that matter, in all amplifiers which contain capacitance, inductance, or both, there is a rather peculiar action called phase distortion. This kind of distortion exists because signals do not pass instantly from the input to the output of an amplifier but actually require a measurable time for their travel. How long it takes a signal to get through an amplifier depends on its frequency and on what capacitances must be charged and on what inductances must be magnetized along the way.

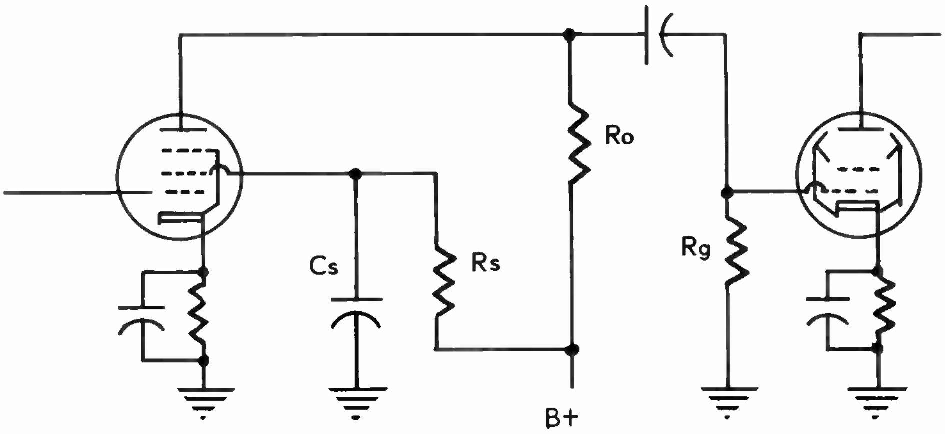


Fig. 47-8. Parts of the B-supply circuit that affect low-frequency gain.

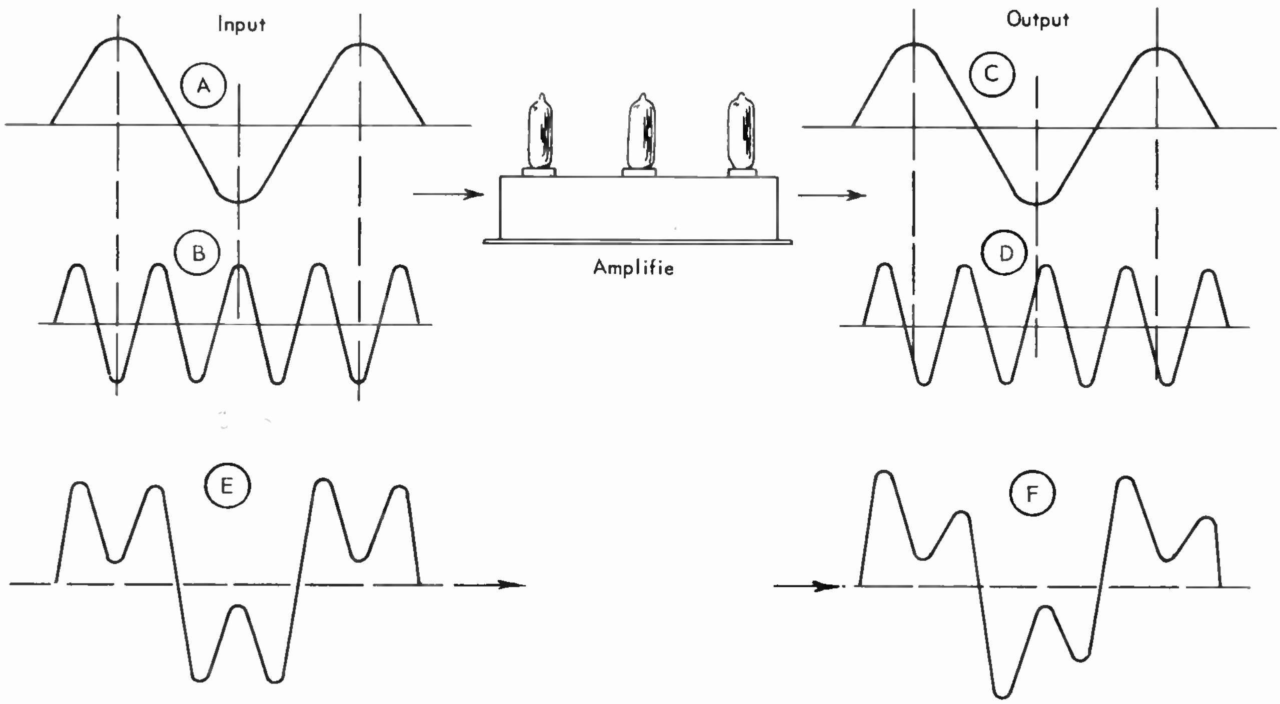


Fig. 47-9. How phase distortion alters the form of the output voltage wave from an amplifier.

As you know, capacitance causes current to lead its voltage, inductance causes the current to lag, and resistance causes neither lead nor lag but leaves current in phase with or in time with voltage. How great is the lead or the lag depends on the values of capacitive or inductive reactances in the amplifier. Reactance of any capacitance or inductance varies with frequency. Consequently, the lead or lag imparted to signal currents varies with their frequency.

The signal currents lead or lag with reference to the input signal voltage, which may be the voltage applied to an amplifier grid. These currents then may flow in a plate load resistor, where they cause signal voltage which is in phase with the current (since it is affected only by resistance) but which must be leading or lagging the original input signal voltage just as the signal current is leading or lagging that original voltage.

The result of all this, in a resistance coupled amplifier, is that a low-frequency signal voltage gets through the amplifier in less time than does a high-frequency signal voltage. The delay of high frequencies is illustrated in simple fashion by Fig. 47-9. At A is a low-frequency signal voltage and at B is a high-frequency sig-voltage fed to the amplifier input. There is a certain phase relation or time relation between cycles of one voltage and cycles of the other voltage.

At C is the low-frequency signal voltage as it appears at the amplifier output, and at D is the high-frequency output voltage. The high-frequency cycles now occur later with reference to the low-frequency cycles than they did at the input. The low-frequency signal has gone through the amplifier circuits faster than the high-frequency signal.

If the low and high frequencies are fed to the amplifier input together there will not be two separate voltage waves, rather there will be a combined or composite wave as at E. This wave is formed by adding the instantaneous values of the two separate waves at A and B. The composite wave is of a certain definite form. If instantaneous values of the low-and high-frequency waves at C and D are added together in the same manner the result is the composite wave shown at F. This output waveform is not at all the same as the input waveform at E. The change of waveform is the result of phase distortion.

High-frequency delay occurs in a resistance coupled amplifier because such an amplifier is made up almost wholly of capacitances and resistances. If we could construct an amplifier almost wholly with inductances and resistances, the low-frequency signals would be delayed more than high-frequency signals. If an amplifier could be constructed wholly of resistances there would be neither lead nor lag, and all frequencies would go through in equal times. We cannot build such amplifiers without capacitances, for capacitances always exist in the tubes and between wires and other parts of the circuits, even though there are no capacitors.

It is possible to construct a coupling circuit with such a combination of capacitances, inductances, and resistance that leads of current due to capacitance are counteracted by lags due to inductance. This would not be a simple resistance coupling, but would be a phase correcting network. Remedies of this general nature will be discussed when we come to the subject of compensated broad-band amplifiers such as used in video amplifiers.

Another way of preventing phase distortion may be explained with the help of Fig. 47-9. Note that the high-frequency at B is just three times the low-frequency at A. There are three high-frequency cycles to each low-frequency cycle. Suppose we could shift the high-frequency cycles either forward or back by exactly one-third of a low-frequency cycle or by exactly the time required for one high-frequency cycle. Then the phase relation between the two frequencies would not be changed in the least, every high-frequency cycle would occupy the same relative position that it has in the input. If phase shift or phase delay between input and output of an amplifier can thus be made proportional to frequency there will be no phase distortion. Still another possibility would be to have all frequencies delayed or advanced by the same times, which would mean no change of relative phase and there would be no phase distortion. These remedies cannot be completely applied in a practical amplifier, but they can be approximated.

Phase distortion is not very important in audio amplifiers. The signal at F in Fig. 47-9 would produce sound which hardly could be distinguished by ear from sound produced by the signal at E, although there is very great phase distortion.

in video a...

Phase distortion can cause serious troubles in television receivers. Picture signals are made up of frequencies from as low as 30 cycles to as high as 4 megacycles. There may be widely different frequencies along a single horizontal line and in successive lines. If the highest frequencies are much delayed the fine details of the picture will be shifted or blurred. On the face of a 16-inch picture tube the electron beam travels an inch in about 5 millionths of a second. A phase delay of only 1 millionth of a second would displace some details by 1/5 inch in the picture.

Phase distortion can cause trouble with synchronizing. A square-cornered sync pulse is really a composite voltage wave made up of a great range of frequencies, from very low to very high. High-frequency delay can round the corners of the pulse so that the leading edge does not have the sharp rise necessary for positive sync action.

3 If a resistance coupled amplifier is designed and constructed to have negligible phase distortion through the mid-band frequencies the output signal may lead the input at low frequencies and lag the input at high frequencies. This is illustrated by Fig. 47-10. At the low frequencies the reactance of the coupling capacitor Cc is increasing with drop of frequency. This puts too much capacitive reactance in the circuit, and the output voltage will lead the input more than at mid-band frequencies, where it is assumed that neither lead nor lag. At high frequencies all the capacitive reactances are decreasing, including the reactances due to tube and circuit capacitances. This leaves less capacitive reactance than for the mid-band frequencies and, in relation to the phase difference between input and output at those frequencies, the output lags the input. The changes of output lead and lag shown by Fig. 47-10 are such as really occur in resistance coupled amplifiers which handle a wide range of frequencies.

VOLTAGE GAIN IN AMPLIFIERS. Back in Fig. 47-3 it was shown how the gain of a resistance coupled amplifier drops off as frequency goes to lowest values, drops off toward the highest frequencies, and remains fairly uniform only through the mid-band frequencies. Frequencies for that curve were shown as extending to 24,000 cycles. The general form of the curve would be the same no matter what the range of frequencies, it applies to video amplifiers as well as to audio amplifiers.

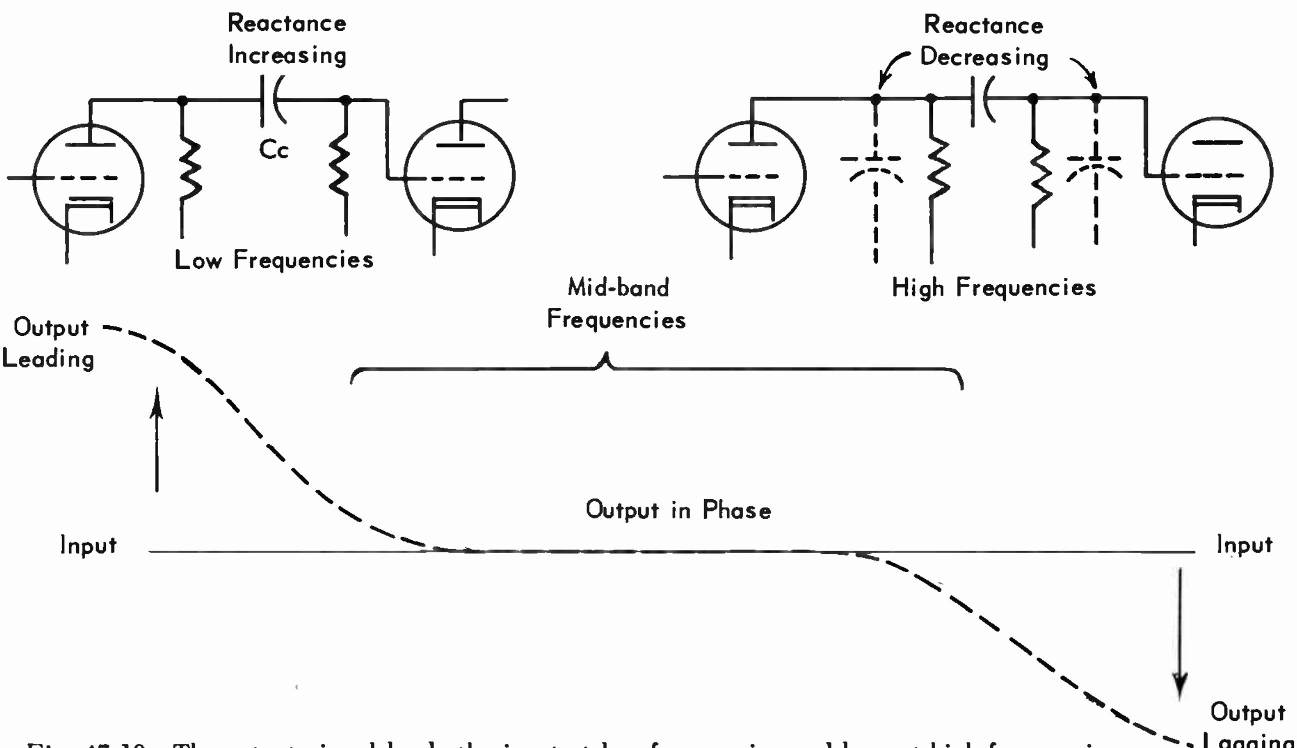


Fig. 47-10. The output signal leads the input at low frequencies and lags at high frequencies.

In Fig. 47-10 the broken line curve shows how the output signal leads the input signal at lowest frequencies, how output lags the input at highest frequencies, and how output and input are in phase only through mid-band frequencies. Those two curves, one for gain and the other for phase distortion, are shown together in Fig. 47-11. It is clear that phase shift, either leading or lagging, accompanies a drop of gain at both ends of the frequency range. If we do things to this amplifier which will make the gain more uniform throughout the frequency range, those things will not only improve the frequency response but also will greatly lessen phase distortion.

The theoretical gain of an amplifier may be computed by using any of various formulas which take into account the transconductance of the tube, its plate resistance, the values of load resistor and following grid resistor, and the coupling capacitance. These formulas apply at what we are calling the mid-band frequencies. Computed values of gain would be interesting, but not of any particular use in service work on broad-band amplifiers because of the effect of shunting capacitances (tube and circuit) on the actual impedance of the plate load.

When during service operations, we wish to know the gain of any amplifier stage the information isn't obtained from formulas. Rather we use an electronic voltmeter to measure the signal voltage at the grid of the following tube and at the grid input to the amplifier tube in the measured stage. Then dividing the the output signal volts by the input signal volts gives the gain.

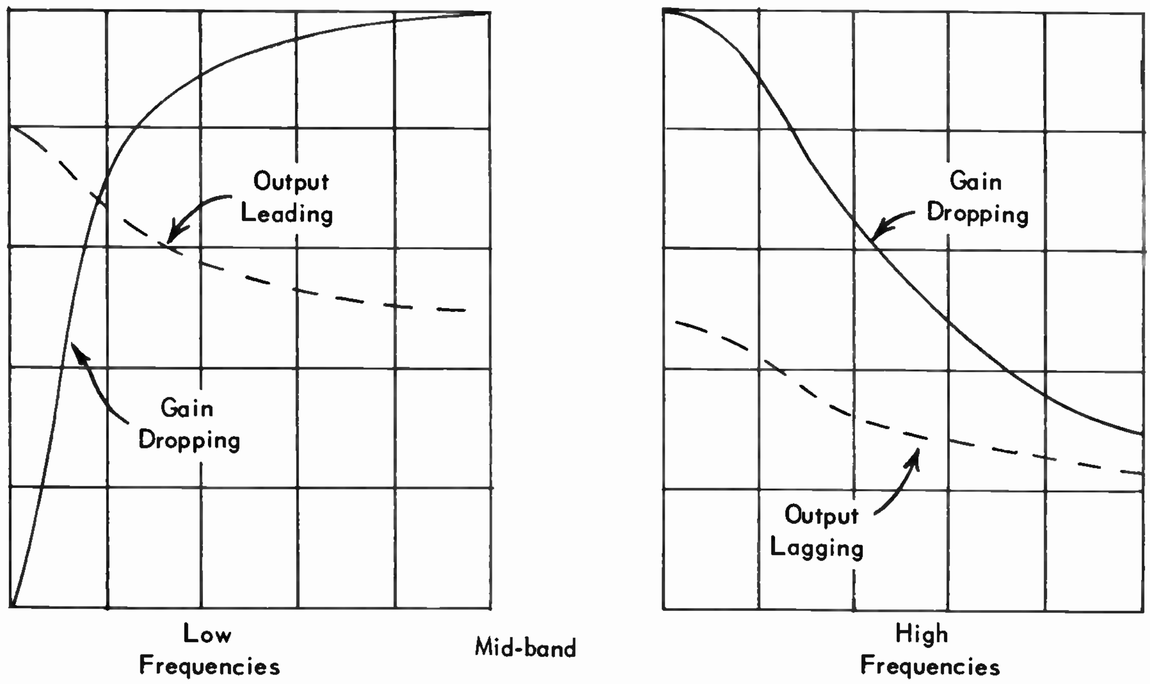


Fig. 47-11. Phase distortion accompanies changes of gain at low and high frequencies.

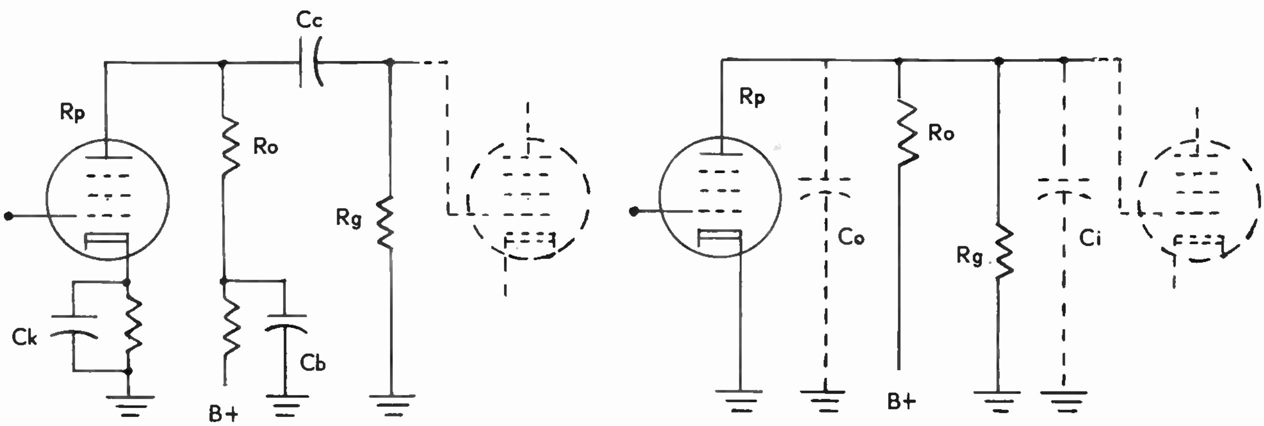


Fig. 47-12. Things which chiefly affect low-frequency gain (at left) and which affect high-frequency gain (at right).

There is, however, one simple formula which is quite useful in determining the effects of various tube transconductances and various actual plate load impedances on relative gains. It is this one.

$$\text{Voltage gain} = \frac{\text{transconductance,(micromhos)} \times \text{load impedance,(ohms)}}{1\ 000\ 000}$$

As an example, supposing the amplifier tube is operated to have transconductance of 5,000 micromhos, and the actual load impedance is 3,000 ohms. The gain figures out this way.

$$\text{Voltage gain} = \frac{5000 \times 3000}{1\ 000\ 000} = \frac{15\ 000\ 000}{1\ 000\ 000} = 15 \text{ times}$$

The actual load impedance at mid-band and low frequencies depends chiefly on the things marked on the diagram at the left 47-12. Resistances Rp, Ro, and Rg still are in parallel with each other. The cathode bypass and coupling capacitors now have such small reactances that they are neglected; being in series with other elements these reactances form what amounts to direct connections of negligible opposition to signal current. Shunting capacitances Co and Ci are in parallel with all the paralleled resistances. These capacitances are the internal output and input capacitances of the tubes plus all the distributed and stray capacitances in the plate and grid circuits. The things shown here are to be worked upon to improve high-frequency response.

LOW-FREQUENCY COMPENSATION. First we shall attempt to raise the gain and lessen the phase distortion at low frequencies. The grid resistor Rg should have the highest value permissible for the type of following tube and the method of biasing for that tube. These grid circuit resistances are specified by tube manufacturers in their listings of characteristics and ratings for various tubes. The grid resistor often is of 0.5 megohm, but may be as low as 0.1 megohm or as high as 1.0 megohm.

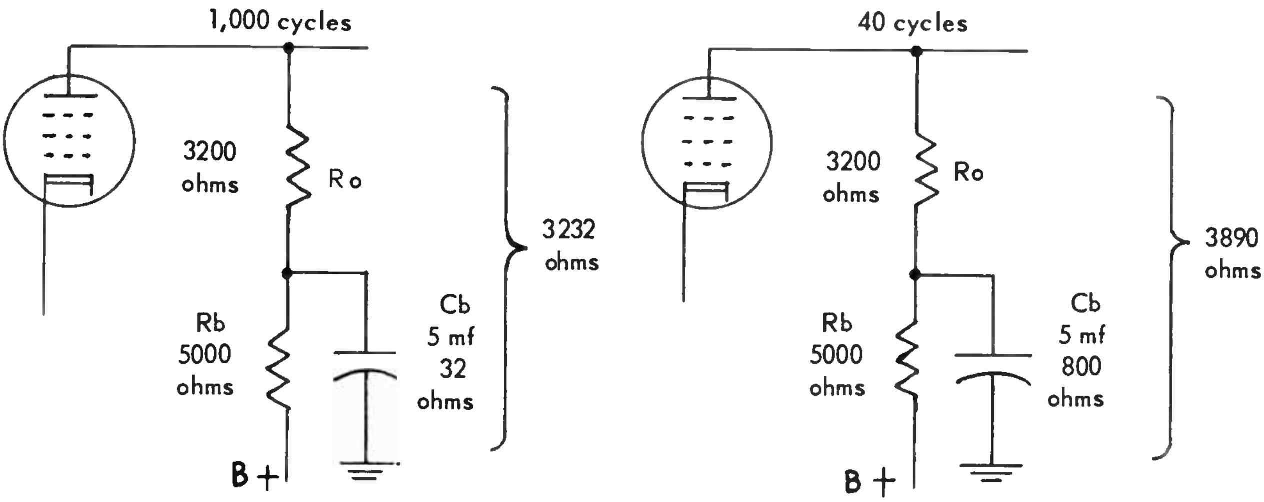


Fig. 47-13. A method of increasing plate load impedance and maintaining gain at low-frequencies.

Coupling capacitance at Cc should be large, in order to have low reactance at the low frequencies. Low reactance at Cc and high resistance at Rg causes most of the signal voltage to appear across Rg and thus to be applied to the following grid. The coupling capacitor should be of small physical dimensions, of small size as measured in inches, and kept away from chassis ground. A big capacitor too close to ground adds greatly to the shunting capacitance which reduces high-frequency gain. Capacitance of about 0.5 mf usually is the practical limit.

No matter how great the coupling capacitance may be made its reactance is going to rise at lower and lower frequencies. It will help matters if we can push more signal voltage through this capacitance and the grid resistor. This can be done only by getting more impedance into the plate load as the frequency goes down.

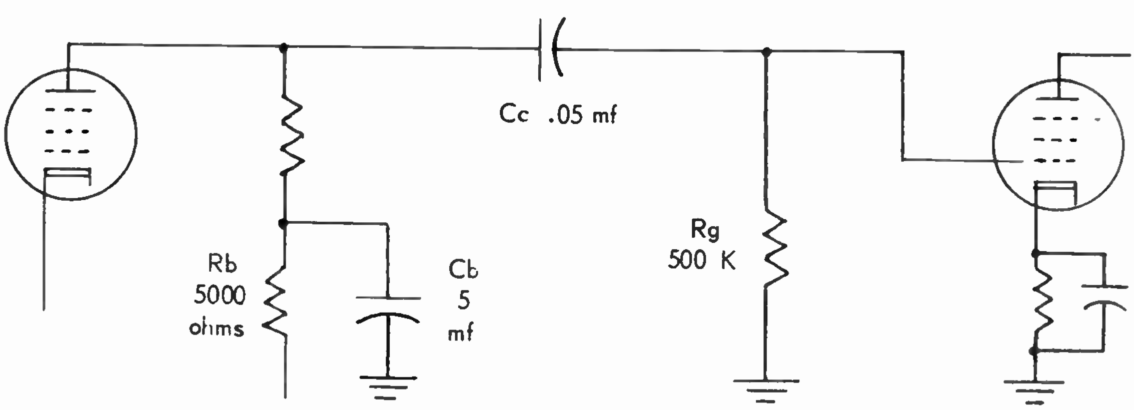


Fig. 47-14. Matching the time constants as a means for improving low-frequency gain.

COMMERCIAL TRADES INSTITUTE

TELEVISION

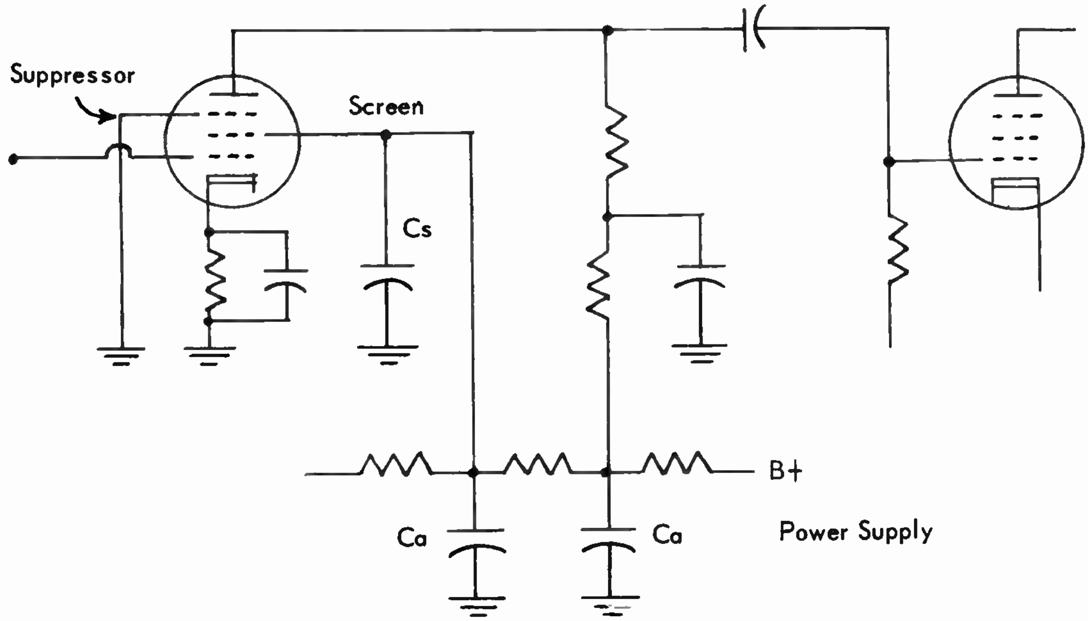


Fig. 47-15. Features commonly found in broad band amplifiers.

A method of increasing the load impedance at low frequencies is shown by Fig. 47-13. In order to work with actual figures we assume a load resistor Ro of 3,200 ohms, a dropping resistor Rb of 5,000 ohms, and a bypass Cb of 5 mf. Other values could be used without altering the principles. Conditions at a frequency of 1,000 cycles are shown in the left-hand diagram. Reactance of Cb is approximately 32 ohms. This reactance in parallel with 5,000 ohms at Rb gives a combined impedance of about 31.5 ohms, or practically the same 32 ohms as before. This combined impedance is in series with 3,200 ohms at Ro to give a total effective load of 3,232 ohms at 1,000 cycles.

In the diagram at the right the frequency has dropped to 40 cycles. Reactance of Cb now is about 800 ohms. In parallel with Rb this makes an impedance of about 690 ohms, which is in series with Ro for a total effective load of 3,890 ohms. The effective load resistance increases by about 20 percent between 1,000 cycles and 40 cycles. This effective load would go up to about 4,075 ohms at 30 cycles, because of increasing reactance of Cb. The greater the voltage dropping resistance at Rb, and the smaller the load resistance at Ro, the greater will be the percentage increase of effective load impedance as frequency drops.

Phase distortion at low frequencies will be lessened if the time constant of Cb and Rb in the plate line is approximately equal to the time constant of the coupling capacitor and the following grid resistor. An example is illustrated by Fig. 47-14. We are using the same dropping resistor and bypass capacitor as before. The coupling capacitor Cc is of 0.05 mf capacitance and the grid resistor is of 500,000 ohms (500 K) resistance. To check equality of time constants all you need do is multiply one of the capacitances and its accompanying resistance together, then multiply together the values of the other two elements - always using the same units of capacitance and resistance in both multiplications, like this.

$$Cb \times Rb = 5 \times 5000 = 25{,}000$$

$$Cc \times Rg = .05 \times 500{,}000 = 25{,}000$$

Very seldom will you find the two products exactly equal, but approximate equality, if it doesn't upset the uniformity or amount of gain at low, medium, or high frequencies, will help reduce phase distortion.

Several additional features often found in broad band amplifiers are illustrated by Fig. 47-15. The amplifier tubes practically always are pentodes of types having high transconductance, which directly increases the gain at all frequencies. Pentodes also have high plate resistances, which helps to raise the effective load impedance.

The suppressor usually is connected directly to chassis ground or to B-minus. It is not connected to the tube cathode when there is a biasing resistor on the cathode. The screen is bypassed to ground or B-minus through a large capacitance at Cs, or in the power supply at Ca, or at both places. The screen must be supplied with nearly pure direct voltage, otherwise it will act like a control grid and cause amplification of ripple voltage.

We already have discussed the plate circuit so far as the load resistor, voltage dropping resistor, and its bypass capacitor are concerned. In addition, there must be a large bypass capacitor to ground or B-minus at the plate supply end of the line, as with capacitor Cb in the power supply. This bypassing is necessary in order that plate signal voltages won't get into other circuits through the B-supply connections.

DO NOT TEAR — CUT ALONG THIS LINE AND SEND IN FOR GRADING.

TELEVISION

LESSON NO. 48

THE VIDEO AMPLIFIER

Training Your Mind

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON NO. 48

THE VIDEO AMPLIFIER

When we construct a resistance coupled amplifier having good gain at low frequencies, and fairly uniform gain up to frequencies of 10,000 to 20,000 cycles per second, we probably have a high-fidelity audio amplifier, but not a video amplifier. A video amplifier requires reasonably good gain at low and mid-band frequencies, and the gain must extend all the way to something between 2½ and 4 megacycles. Extending the gain with satisfactory uniformity to such high frequencies is not easy. It is done by obtaining the best possible gain at the highest required frequency. The things we do to obtain good gain at the highest frequencies will reduce the mid-band and low-frequency gain, but there will be reasonable uniformity throughout the whole range.

Response of a resistance coupled amplifier having good performance at low frequencies and through all audio frequencies, or well beyond them, would be about as shown by the broken-line curve of Fig. 48-1. We cannot maintain the high gain at higher frequencies, but we can extend the response with moderate gain as shown by the full-line curve. This is a typical response of a video amplifier.

The great obstacle in the way of obtaining high gain at high video frequencies is the shunting capacitance. This is the combined capacitance of the tubes, the wiring, and all the circuit parts. This capacitance is in parallel with the total load resistance. The total resistance is the parallel resistance of the load resistor, the grid resistor, and the plate resistance of the tube.

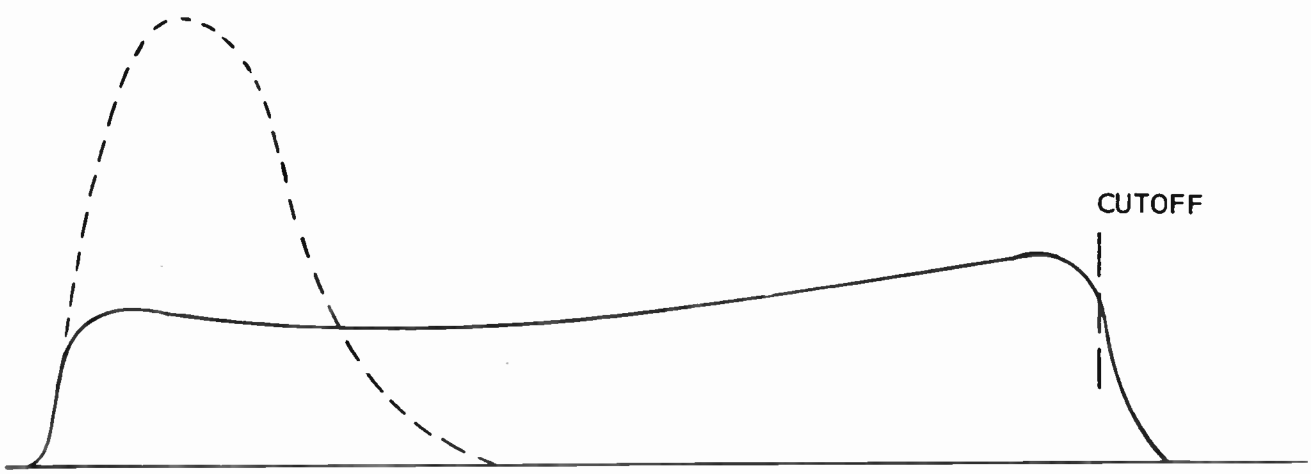


Fig. 48-1. Frequency response of audio amplifier (broken-line) and of video amplifier (full-line)

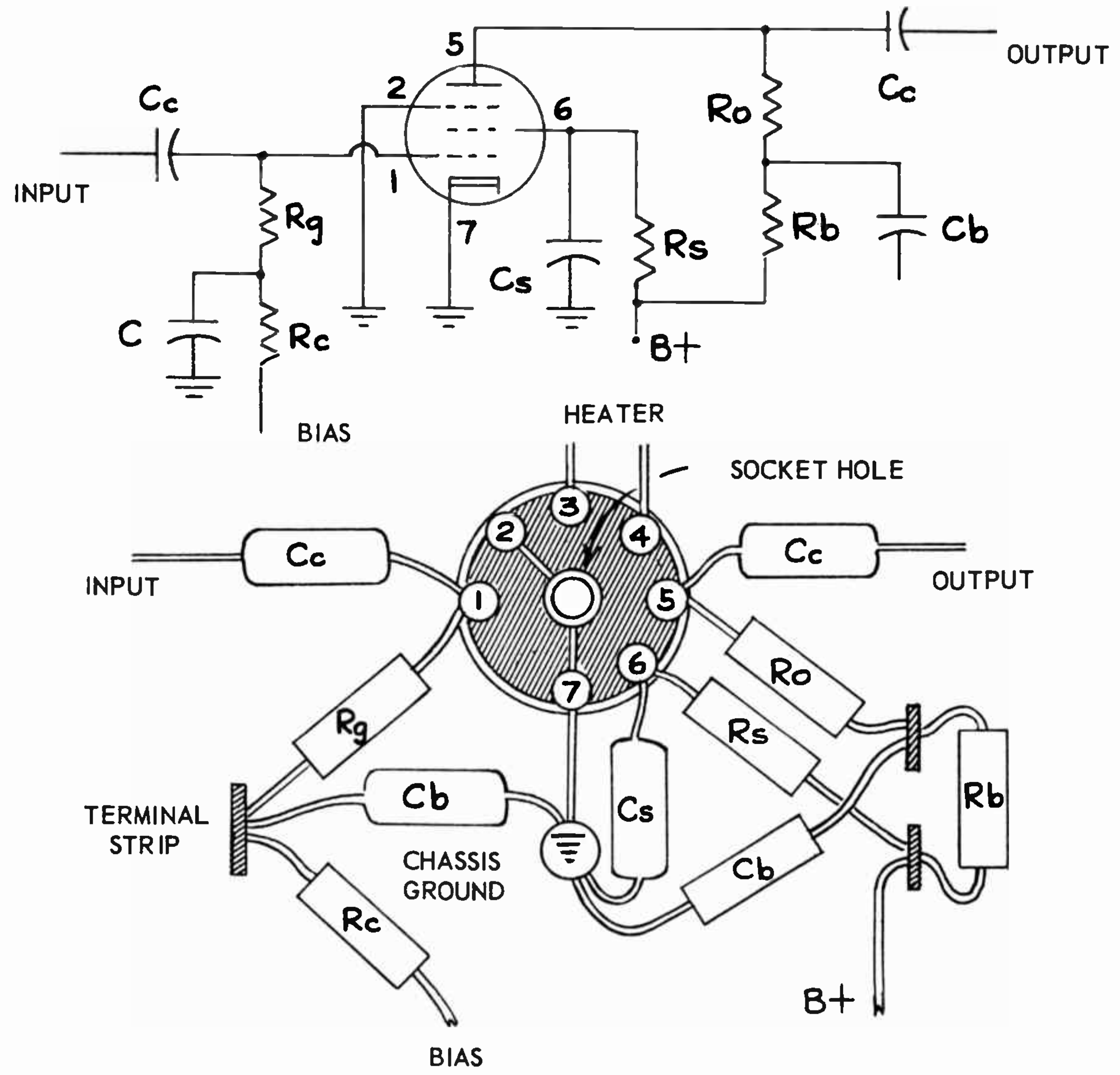


Fig. 48-2. Connections are short and parts close together in a high-frequency amplifier stage.

The grid resistor usually has resistance of a half-megohm or more. Plate resistance of a pentode video amplifier tube will be something between a half and one and one-half megohms. But the load resistor for a video amplifier nearly always is less than 10,000 ohms. If you figure out the parallel resistances of a few such combinations it becomes evident that varying the plate and grid resistanccs between their widest limits won't vary the combined parallel resistance more than two or three per cent. This combined resistance always is practically the same as that of the load resistor by itself.

Reactance of the shunting capacitances drops lower and lower with increase of frequency. This decreasing reactance, in parallel with the total load resistance, lessens the impedance in the plate circuit of the tube, and lowers the gain proportionately. To obtain satisfactory high-frequency gain we must do everything possible to lessen the shunting capacitances and their bad effects, thereby maintaining higher impedance at high frequencies in the plate circuit.

The frequency of cutoff for the video amplifier may be taken as the frequency at which the reactance of the shunting capacitances drops to the value of resistance in the load resistor. For example, if the load resistor is of 3,500 ohms, cutoff will occur at the frequency which brings reactance of the shunting capacitances down to 3,500 ohms.

With good design and construction the total shunting capacitance may be as low as 20 mmf. Let's assume this capacitance and figure on a high-frequency cutoff at 3½ megacycles. At this frequency the reactance of 20 mmf is about 2,430 ohms, which is the required value of load resistor for 3½-megacycle cutoff. If the amplifier tube has transconductance of 5,000 micromhos the approximate gain with this load resistor will be 12.15 for one stage. Were you to substitute a load resistor of 5,000 ohms it would raise the gain to 25 for one stage, but the cutoff frequency would come down to 1.6 megacycles. The only way to get around the situation is to reduce the shunting capacitances.

To begin with we use the shortest and most direct wiring connections. In Fig. 48-2 is a wiring diagram of circuit connections on grid and plate sides of an amplifier tube. Underneath the diagram is a sketch of how the various resistors and capacitors might be arranged around the tube socket. Capacitors, resistors, and wiring connected to grid and plate lugs should not come too close to chassis metal or B-minus connections, for we do not want capacitances to ground.

Tube sockets should be of types having only small capacitance between their lugs or inserts. Insulating supports should be of material having small dielectric constants. Tubes preferably are of types having small internal capacitances, although a tube with extra high transconductance may produce more gain in spite of somewhat greater input and output capacitances. After all this has been done we use a load resistor that allows the desired frequency of cutoff.

The accompanying table lists interesting facts about some of the miniature pentode tubes which may be used for video amplifiers.

VIDEO AMPLIFIER TUBES

Type Number	Capacitances, mmf. Output	Input	Plate Resistance, megohms.	Transconductance, micromhos	Plate Voltage
6AH6	2.0	10.0	0.5	9000	300
6AU6	5.5	1.5	1.5	4450	250
6BA6	5.0	5.5	1.5	4400	250
6BC5	4.3	4.0	0.8	5700	250
6BH6	4.4	5.4	1.4	4600	250
6CB6	1.9	6.3	0.6	6200	200

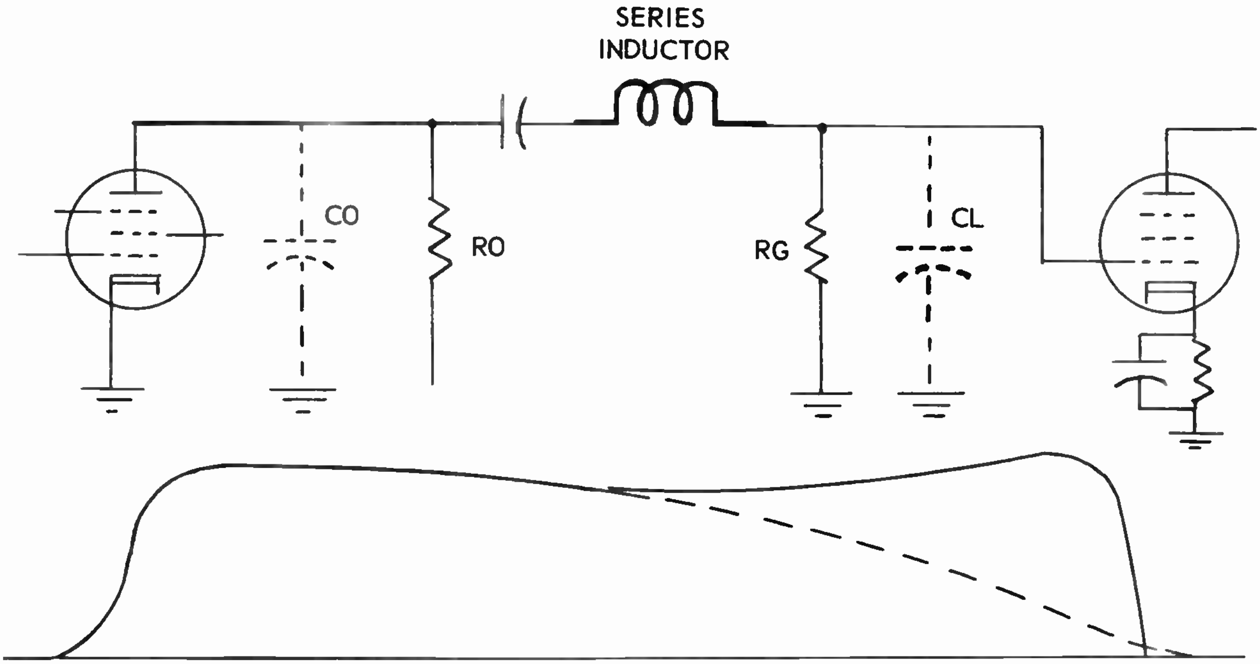


Fig. 48-3. Series compensating inductor and its effect on response.

HIGH FREQUENCY COMPENSATION. Having done all that we can to reduce the shunting capacitances the next step is to remove the effect of about half this capacitance from the load in the amplifier plate circuit. This is done by inserting in series between the plate of the amplifier and the grid of the following tube an inductor, as shown by Fig. 48-3. This series compensating coil has such value of inductance that it resonates with shunting capacitances *Ci*, on the grid side of the circuit, at a frequency close to or slightly below that desired for cutoff.

The result of adding this inductor is to practically remove the effect of capacitance *Ci* from the plate load, and to leave this load equal to the parallel impedance of the load resistor, the tube plate resistance, and the reactance of shunting capacitances represented by *Co*, on the plate side of the circuit. The reduction of shunting capacitance across the plate load allows using more ohms in the load resistor, which means more gain throughout the whole frequency range. In addition, we will have extended the high-frequency cutoff about as shown by the response curves below the circuit diagram. Without series compensation the gain would fall off as shown by the broken-line curve. With series compensation it is extended as shown by the full-line curve.

In Fig. 48-4 we have done something else that extends the gain into the high frequencies. An inductor *Lo* is connected in series with load resistor *Ro*. This inductor completes a parallel resonant circuit connected between the tube plate and ground, as shown by the heavy lines. The resonant circuit consists of the added inductor and the shunting capacitances represented at *Co*. Inductance of *Lo* is made of such value as causes resonance at a frequency slightly higher than the desired cutoff. Because the resistance of *Ro* is included in the resonant circuit the tuning is broad. That is, the resistance of *Ro*, together with other high-frequency losses, makes this a low-Q circuit.

Because inductor *Lo* is in parallel with or is shunting the shunting capacitances of the circuit this coil is called a shunt compensating coil. The impedance of any parallel resonant circuit becomes maximum at the resonant frequency. This shunt compensating circuit increases its impedance at frequencies approaching its resonant point, and thus increases the impedance in the plate circuit. This increase of load impedance counteracts the drop of impedance due to decreasing reactance of the shunting capacitances. The result is as shown by either of the full-line response curves down below, or by any response drawn in between those shown.

Sharper and higher peaking at the high-frequency end of the response could be had by connecting a small capacitor across the ends of compensating inductor *Lo*, which would make this inductor and the added capacitor a resonant circuit of narrow response. More often it is desirable to make the resonant peak less pronounced. This is done by connecting a fixed resistor of maybe 20,000 ohms across the ends of inductor *Lo*. A fixed resistor usually is connected across the ends of the series compensating inductor of Fig. 48-3 for the purpose of extending the high-frequency gain more uniformly and without high peaking.

When inductors of the types being examined have resistors connected across them the actual method of construction ordinarily is to wind the inductor, or at least mount it, on the resistor. Several such units are pictured by Fig. 48-5. You will see many of them in television receivers. The ends of the small wire in the coil are soldered to the heavier wire in the resistor pigtails. Then the whole thing is supported by and connected into the circuit by the pigtails. These coil and resistor combinations commonly are called peakers.

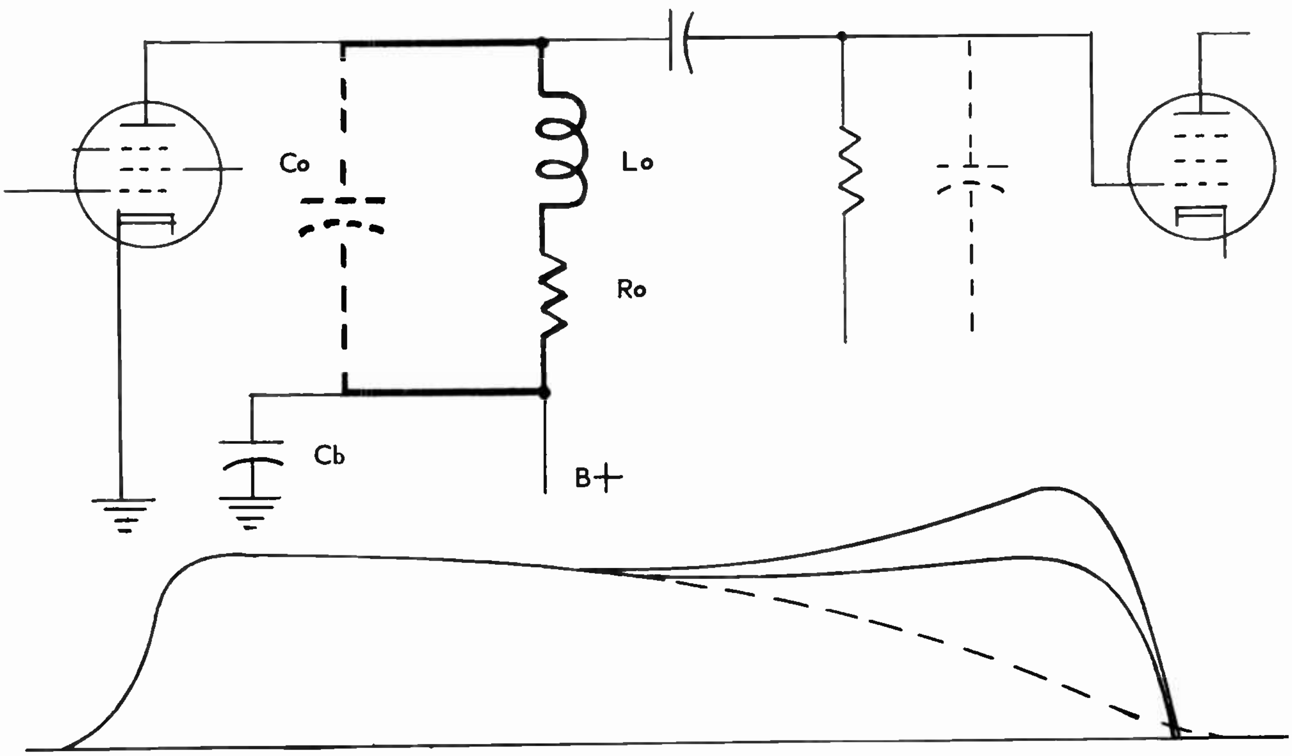


Fig. 48-4. Shunt compensating inductor and how it affects response.

It has been mentioned that the shunt compensating inductor of Fig. 4 resonates with shunting capacitances *Co* on the plate side of the circuit. This would be strictly true only when there is also a series compensating inductor (as in Fig. 48-3) to separate the capacitances on plate and grid sides of the circuit. When there is no series compensating inductor for thus "splitting the capacitances" the shunt compensating inductor resonates with the combined or paralleled capacitance of the shunting capacitances on both sides. The shunt compensating inductance at *Lo* ordinarily is somewhere between 30 and 100 microhenrys, depending on the amount of capacitance with which it is to resonate. The series compensating inductance of Fig. 48-3 usually is somewhat greater, often being between 80 and 150 microhenrys.

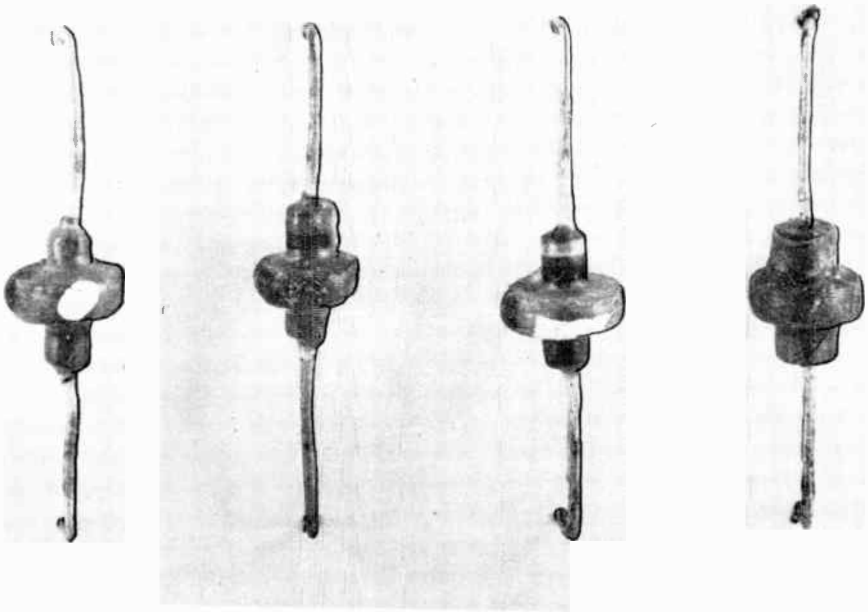


Fig. 48-5. Peaking coils mounted on fixed resistors with pigtail leads.

In Fig. 48-6 are circuit diagrams of video amplifier stages as used in two typical receivers. In both diagrams there are series compensating or peaking coils *Ls* and also shunt compensating or peaking coils *Lo*. In the upper diagram there are broadening resistors across both peakers, and in addition, on the series compensating coil, there is a small capacitor *Cp* which helps extend the gain to higher frequencies. In the lower diagram there is a broadening resistor across the series peaker, but none across the shunt peaker. Using both series and shunt compensation allows extending the band to much higher frequencies than using either of the compensating methods by itself.

Shunt compensation, series compensation, or both, are used in the coupling between the video amplifier and the picture tube as well as between the two video amplifiers when two are used.

Shunt compensation is used also in the coupling between video detector and video amplifier, as shown by Fig. 48-7. The shunt compensating coil *Lo* is in series with the detector load resistor *Ro*. As explain-

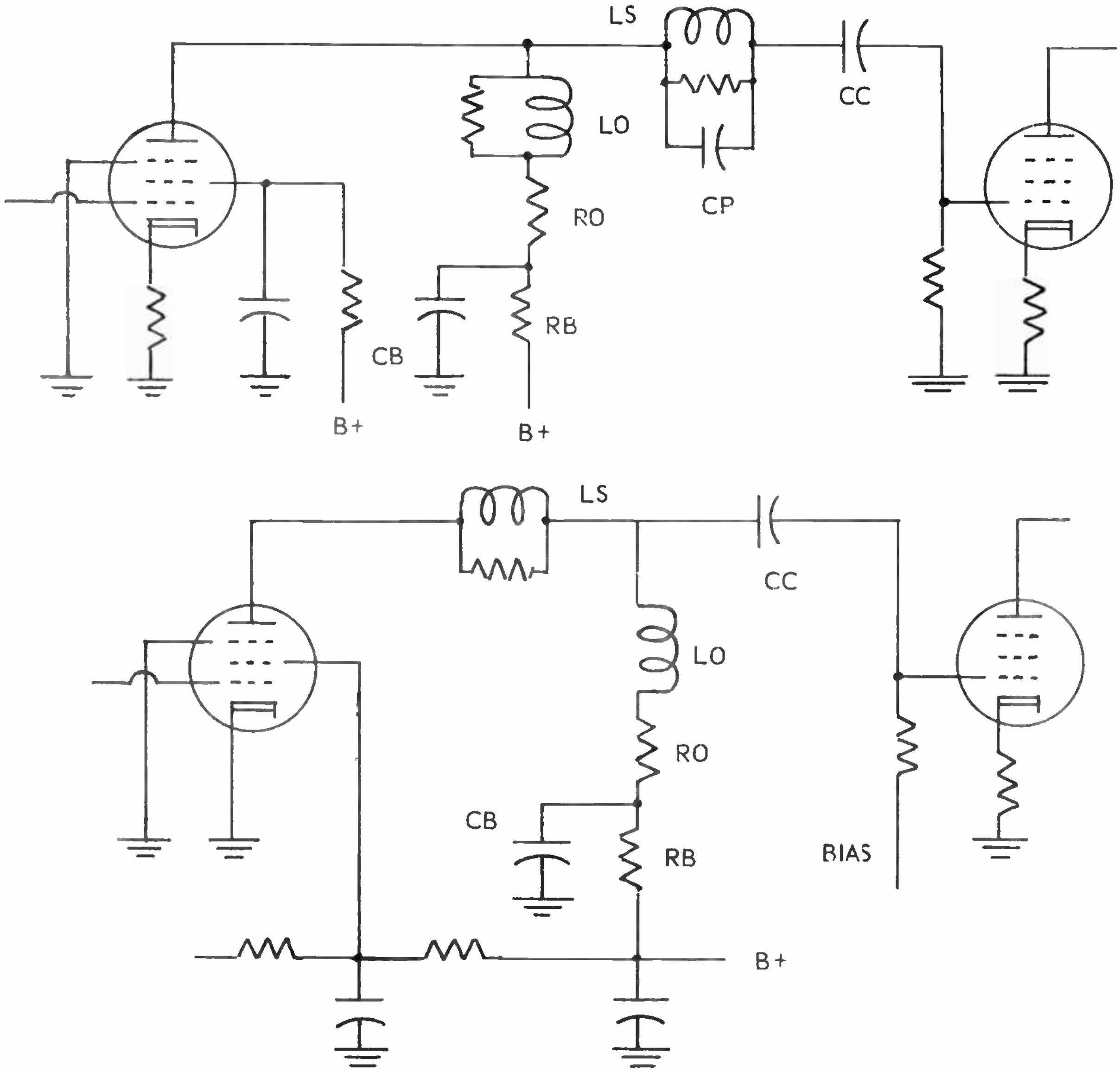


Fig. 48-6. Video amplifiers with series and shunt compensation.

ed earlier, the high impedance of inductor *Ls* to currents at intermediate frequencies, in combination with the low reactance of the detector output capacitance to ground at *Cd*, prevent intermediate frequencies from going to the video amplifier. Capacitance of coupling capacitor *Cc* is negligible at the higher video frequencies, so inductor *Lo* becomes part of a resonant circuit in which is the shunting capacitance *Ci* on the input side of the video amplifier. This resonant circuit includes detector load resistor *Ro*, and is paralleled by grid resistor *Rg*, and consequently has a rather broad resonance peak. The peaking may be further reduced by a broadening resistor connected across the ends of *Lo*.

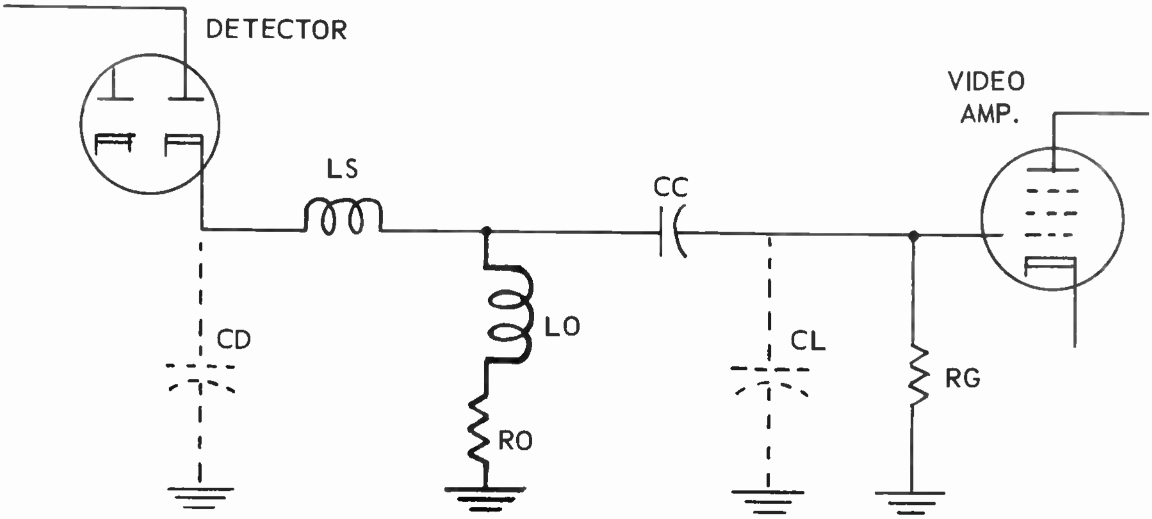


Fig. 48-7. Shunt compensation between video detector and amplifier.

(3) (10) *VIDEO AMPLIFIER REQUIREMENTS.* We may summarize some of the principal requirements of the video amplifier system as follows. These requirements apply to circuits which are between the video detector output and the input to the picture tube grid-cathode circuit. The band width should extend from a low frequency of about 30 cycles to a high of between 2.5 and 4.0 megacycles. The 30-cycle response is needed when one cycle of light and shade in the picture lasts through an entire frame. The high limit depends to some extent on diameter or face size of the picture tube, higher frequencies being desirable to furnish good detail or definition for large tubes. Frequencies above 3 megacycles are desirable for live programs received directly from the producing station, where there is good detail in the transmission. Frequencies somewhat less than 3 megacycles are satisfactory for programs transmitted by long distance cable and for those originating on motion picture film, since in neither of these cases are higher frequencies found in the received signal.

The band width of any particular amplifier is determined chiefly by transconductance of the tubes, by how small are the shunting capacitances, and by the gain desired. This assumes that design and construction are good enough to take full advantage of high transconductances and low shunting capacitances. The higher the transconductances, and the less the shunting capacitances, due to design and construction and the less the gain per stage, because of the plate load resistor the greater may be the band width. Band width at satisfactory gain is extended by using shunt or series compensation, or both.

The importance of having minimum phase distortion already has been explained. It is desirable also to have but little amplitude distortion, which means that weak signals of small amplitude and strong signals of large amplitude will receive equal amplification. Finally, we should have a high ratio of signal voltage to voltages which are due to tube and circuit "noise". When noise voltages are strong in an audio amplifier they cause actual noise from the speaker. The same voltages in a video signal cause small streaks, specks, flashes, and the appearance of snow in the reproduced picture. Amplitude distortion and noise will be considered at more length in connection with audio amplifiers.

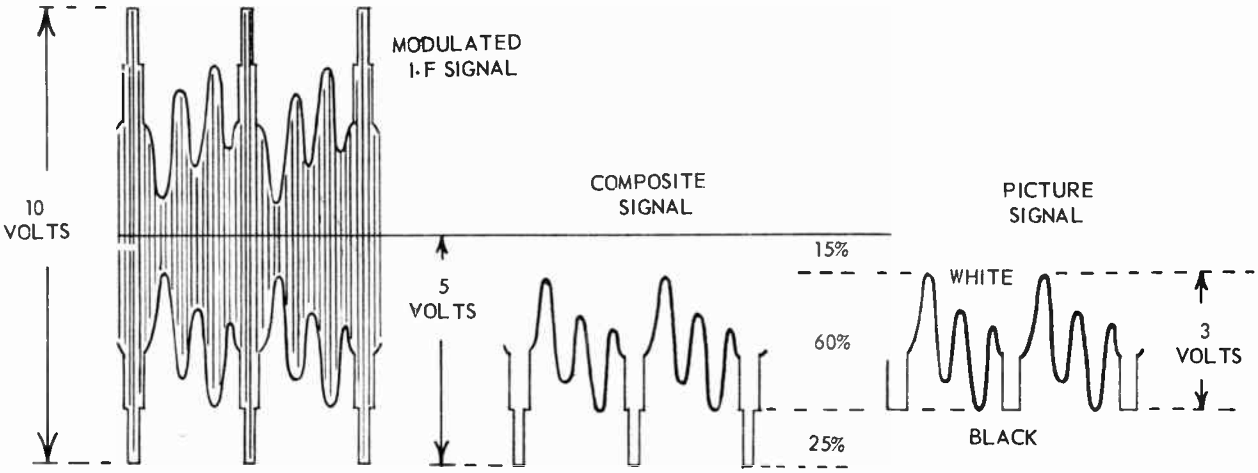


Fig. 48-8. Relations between video i-f signal and picture signal variations from detector.

The number of times the detector output voltage must be increased in the video amplifier system depends on this output and on what variation of voltage is needed to vary the intensity of the picture tube beam all the way from darkest to lightest shades. This picture tube voltage varies somewhat with the type and size of tube and with its operating voltages, but 50 volts swing may be taken as a fair average for purposes of explanation

The picture signal voltage at the video detector output is determined as shown by Fig. 48-8. At the left is represented the modulated voltage at the output of the i-f amplifier and at the detector input. We may assume a peak-to-peak strength of 10 volts. After rectification by the detector only one side of the i-f signal will remain, as shown by the center diagram. Here we have half the original voltage, or have 5 volts in the composite television signal.

In any composite signal the white level is about 15 per cent from zero and the black level is at about 75 per cent. In between white and black there is then 60 per cent of the total composite signal voltage. In our assumed case, 60 per cent of 5 volts would be 3 volts of picture signal swing from the detector. We have assumed that there is no voltage loss in the detector itself, which could be true if the effective load on the detector exactly matched the internal impedance of the detector. Ordinarily there would be a loss of about 10 per cent or more in the detector circuit, so our actual picture signal swing would be about 2.7 volts.

If the picture tube requires a picture voltage swing of 50 volts, and we have 2.7 volts at the detector, the required gain is the quotient of dividing 50 by 2.7, or is a gain of 18.5 times in the video amplifier system. With greater voltage from the i-f amplifier, or with less voltage needed at the picture tube, we would not need so much gain in the video amplifier. Less i-f output or more voltage needed at the picture tube would call for more gain. The i-f output will vary with changes of strength in the received signal, so we shall need a contrast control to maintain a reasonably uniform picture.

SIGNAL POLARITIES. When studying video detectors we learned that the composite signal must be of positive polarity when applied to the grid of the picture tube, and of negative polarity when applied to the picture tube cathode. Positive polarity means that picture variations are on the positive side of the sig-

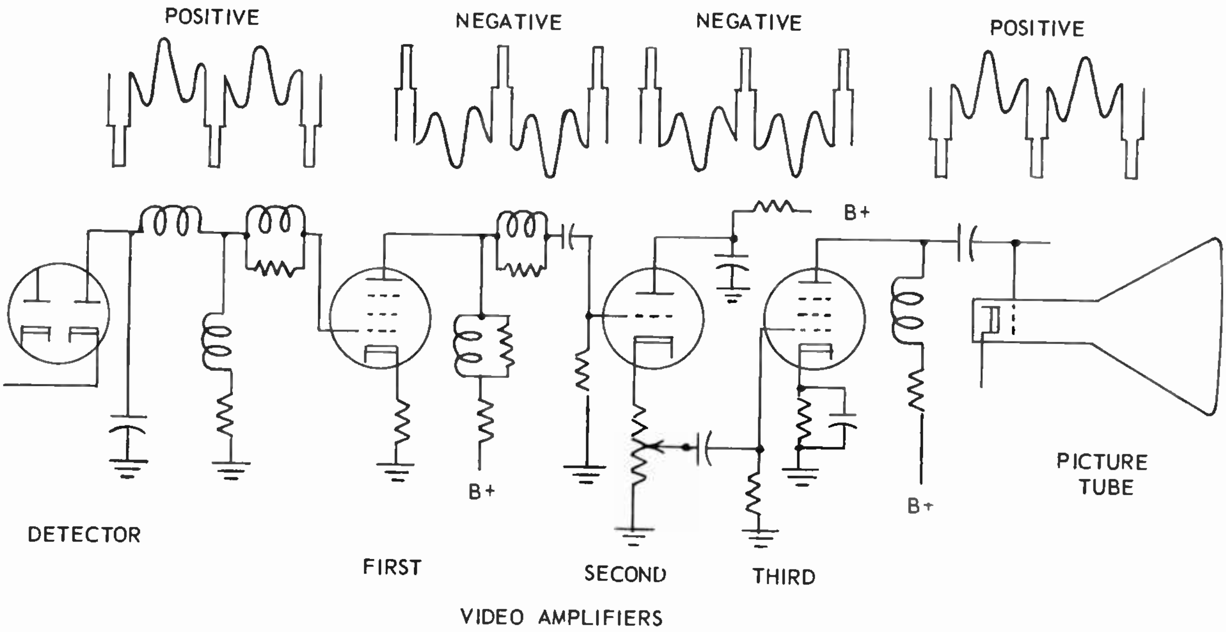


Fig. 48-9. Signal polarities in a video amplifier having stage connected as a cathode follower.

nal, with sync pulses negative. Negative polarity means the opposite relation between picture variations and sync pulses.

We learned also that the signal is inverted in polarity between grid and plate sides of each video amplifier tube. Finally, we learned that a signal of either polarity may be obtained from the video detector, depending on whether the detector plate or its cathode is connected to the i-f amplifier output. This detector connection to the i-f amplifier is chosen so that, with whatever number of video amplifier stages are used, the signal will be of required polarity at the picture tube.

In a few video amplifier systems there will be a stage that does not invert the signal. This is true of the second video amplifier in Fig. 48-9. Input to this tube is at its grid, and output is from the cathode resistor. This tube is connected as a cathode follower. There is no signal inversion in a cathode follower, so here there are negative signals on both sides of the second amplifier. This tube might not be called an amplifier, for there is no voltage gain in a cathode follower, and there usually is some loss. The follower connection is used in the circuit illustrated in order that the cathode resistor may be used as a low resistance contrast control, feeding the third video amplifier a signal whose strength is determined by adjustment of the control.

TRAP FOR INTERCARRIER BEAT. If both the video intermediate and sound intermediate frequencies reach the video detector they beat together and the detector acts as a mixer to form a 4.5-mc frequency,

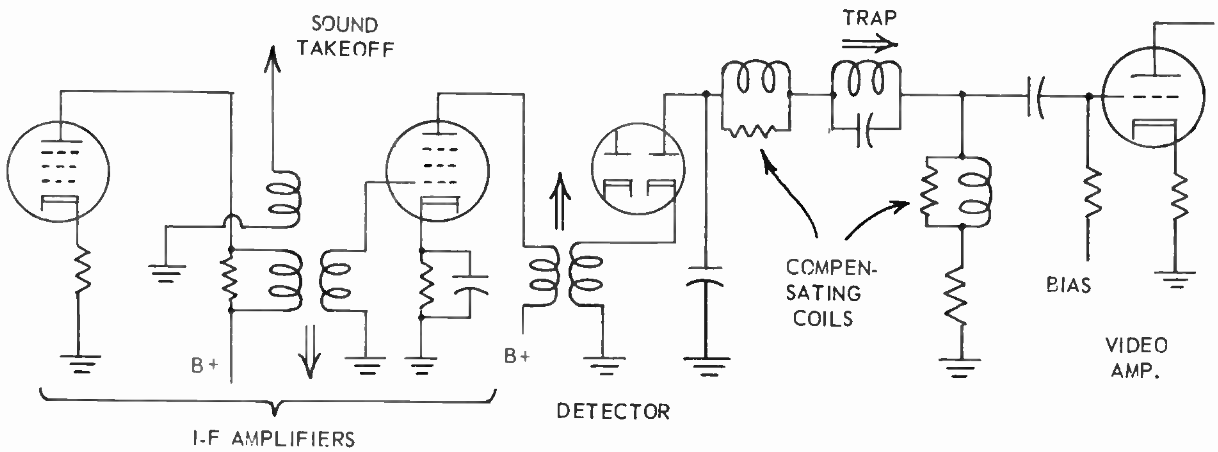


Fig. 48-10. A trap for intercarrier beat frequency between video detector and amplifier.

which is equal to the difference between the two intermediate frequencies. The 4.5-mc frequency is called an intercarrier beat frequency. In receivers employing the intercarrier sound system this is the frequency that forms the input for the sound section. It is almost completely removed from the video amplifier system by the sound takeoff.

If the intercarrier beat frequency goes through the video amplifier and reaches the picture tube of a receiver employing a dual or split sound system the result will be a pattern of narrow slanting lines on the pictures. If the sound intermediate frequency is completely attenuated by accompanying sound traps in the i-f amplifier this frequency will not remain to beat with the video intermediate at the video detector, and there will be no intercarrier beat to cause picture trouble.

If sound intermediate frequency remains at the video detector in a receiver using a dual or split sound system, the resulting intercarrier beat may be trapped out in any of the couplings which are between the detector and the input to the picture tube. Fig. 48-10 shows an intercarrier beat trap of the parallel resonant type connected in series between the detector and first video amplifier tube. The trap is tuned to 4.5 megacycles by a movable core in its coil, and then offers maximum impedance at this frequency. A similar trap might be connected between video amplifier tubes, or between the last video amplifier and the picture tube.

The 4.5-mc frequency of the intercarrier beat trap is enough higher than the highest required video frequency that the trap causes no reduction of band width in the video amplifier system. An incorrectly tuned trap may cut off some of the higher video frequencies to prevent good picture detail, and at the same time

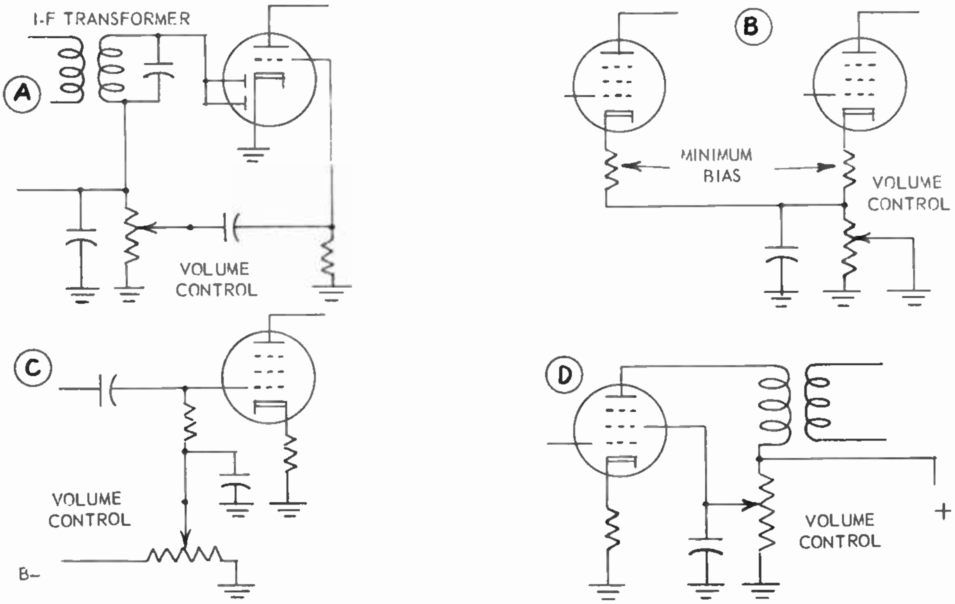


Fig. 48-11. Manual volume controls for a-m sound receivers.

allow the fine-line pattern in the pictures. Intercarrier beat traps seldom are needed and are not often found in receivers having intercarrier sound systems, the sound takeoff being relied upon to keep this frequency from the picture tube.

1 *CONTRAST AND AUTOMATIC GAIN CONTROLS.* Closely associated with the video amplifier, video detector, and i-f amplifier of the television receiver are two important controls. One is the contrast control, which is adjusted by the user while the set is in operation. The other is the automatic gain control, whose adjustment is a service operation.

The television contrast control is comparable in some ways to the manual or hand-operated volume control of a sound receiver. The volume control helps maintain a desired sound level with signals from stations of high or low power, or from nearby and distant stations. A television contrast control helps maintain desired contrast between light and dark tones of a reproduced picture from various stations whose received signals may be strong or weak.

3 The automatic gain control of the television receiver is directly comparable to the automatic volume control of sound receivers. Both these automatic controls help compensate for changes of strength which may occur either suddenly or gradually in the signal from a single station during one program. The automatic controls make it unneccessary for the operator to continually manipulate the manual volume or contrast controls, and they compensate for changes of signal strength which occur too rapidly to allow correction by any manual control.

In the majority of sound receivers the manual volume control is of a type generally similar to that shown at *A* in Fig. 48-11. A potentiometer is used as load resistor for the diode detector, with the adjustable slider taking off any suitable value of audio signal voltage, which becomes the input to the first audio amplifier tube.

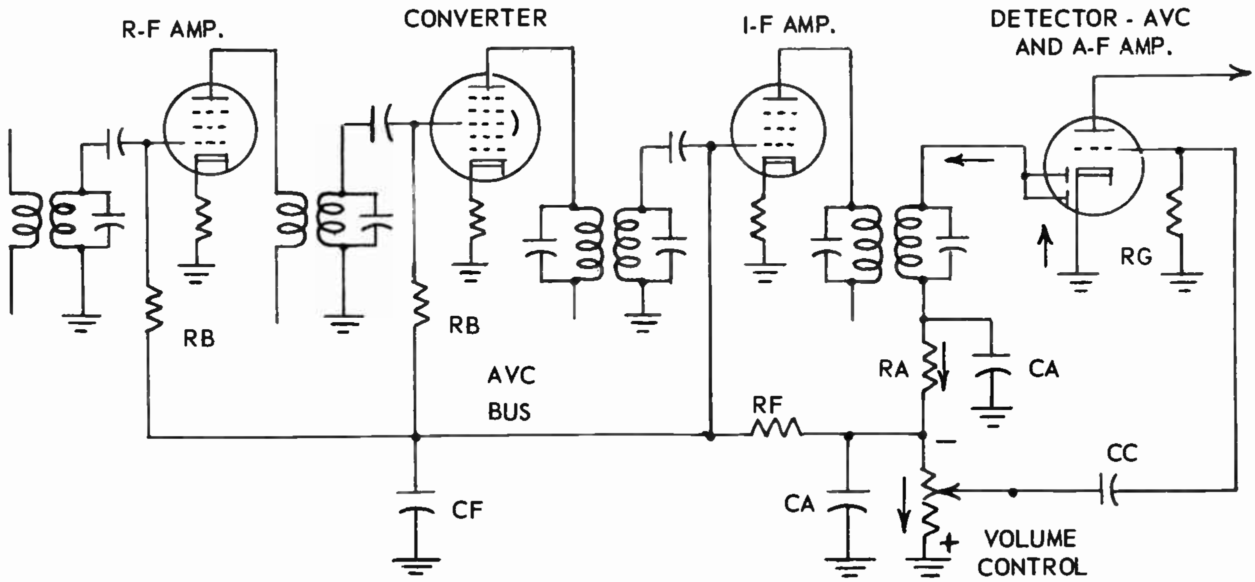


Fig. 48-12. Simple avc system as used in an a-m sound receiver.

In some present-day sound receivers, and in many of the older types, the manual volume control acts on the r-f or i-f amplifier tubes. One such control, shown by diagram *B*, varies the bias of one or more variable-mu amplifier. The biasing resistance is in two sections, one adjustable for volume control and the other non-adjustable to maintain some minimum negative bias regardless of how the control is adjusted.

Instead of using an adjustable cathode bias, the biasing voltage may be obtained from the negative side of the B-power supply as in diagram *C*. Some of the older sound receivers had volume controls consisting of a variable voltage for the screen of r-f amplifier tubes, as in diagram *D*. Still others had adjustable grid resistors, or adjustable resistors across the antenna coupler.

AUTOMATIC VOLUME CONTROL. Most of the automatic gain controls in television receivers are quite similar to automatic volume controls in sound receivers. Therefore, as an introduction to the subject of automatic gain control, we shall examine more closely than in earlier lessons some of the commonly used automatic volume controls for sound receivers.

All automatic volume controls employ the same basic principles, which are as follows. To begin with, the i-f output signal voltage from the i-f amplifier is proportional to strength of the signal received at the antenna. This i-f signal voltage is rectified by a diode, producing a direct current varying in accordance with the i-f signal voltage. The rectified current, which really is the audio signal, is filtered by means of capacitors and resistors to remove the audio variations and leave only a smooth direct current whose average strength is proportional to the voltage of the signal from the antenna and i-f amplifier.

The smooth direct current flows in a resistor, across which appears a voltage whose strength is proportional to the received signal and the i-f amplifier output voltage. The end of this resistor which is negative with respect to ground is connected to the grid returns of whatever amplifier tubes are to be automatically controlled. Then the stronger the received signal, and the i-f amplifier output voltage, the more negative will be the bias on controlled amplifiers. Thus their amplification is reduced to compensate in great measure for the increase of received signal voltage. With a weaker received signal the avc bias becomes less negative, and amplification increases.

Fig. 48-12 shows a simple avc system for an a-m sound receiver. The combined detector, avc tube, and a-f amplifier is a duodiode-triode in which the triode section is the a-f amplifier, with the diodes acting for detection and automatic volume control. Rectified electron flow is in the direction indicated by arrows. Capacitor *Ca* bypasses intermediate frequencies to ground. Its capacitive reactance at 455 kc ordinarily is 1/100 the resistance of *Ra*, making this resistor and the capacitor act as a filter to keep intermediate frequencies out of the parts below *Ra*.

Electron flow downward through the volume control resistor means that the top of the control is negative with respect to the grounded lower end, which is positive. The negative end of the control resistance is connected through resistor *Rf* to the avc bus line, to which come the grid returns of the r-f amplifier, the converter, and the i-f amplifier tubes. The cathodes of these tubes are connected through resistors to ground, these resistors maintaining a small minimum negative bias at all times. The automatic bias voltage then is the drop across the volume control resistor, whose upper negative end is connected to the grids and whose lower positive end is connected to the cathodes of controlled tubes, through chassis metal ground.

Since there is no electron flow in the grid circuits of negatively biased tubes there is no flow and no voltage differences in parts of the avc circuit on the left of the volume control. Connections shown to ground usually would be to B-minus in a transformerless receiver. The two diode plates of Fig. 48-12 are connected together to act like a single plate. Sometimes only one of the diode plates is used for detection and avc action, with the other plate grounded. The i-f filter formed by *Ra* and *Ca* is not always used, the avc bus line being connected to the bottom of the transformer secondary when it is desired to have stronger biasing voltage. Resistors *Rb* serve to isolate the several grid circuits and their signal frequencies from one another. These resistors carry no current and do not alter the biasing voltage.

In Fig. 48-12 the grids of controlled tubes are insulated from the secondary windings of the coupling transformers by blocking and coupling capacitors, with the grid returns for biasing voltage through resistors *Rb* or directly to the avc bus. A different method of connection is shown in Fig. 48-13. Here the grid returns to the avc bus are through the secondary windings. Insulating or blocking capacitors are connected from the lower ends of the secondaries, and the grid returns, to ground.

The control voltage applied to amplifier and converter grids is the voltage that appears across capacitor *Cf*, which is connected between the avc bus and ground. The satisfactory performance of the avc system depends greatly on the time constant of this capacitor and the resistances through which it charges and discharges. Charging current comes from the bottom of the transformer secondary through resistors *Ra* and *Rf*. Resistance at *Rf* usually is much the greater of the two. Discharge of *Cf* is through resistances at *Rf* and *Rv*, in series between the capacitor and ground. Resistance of the volume control, *Rv*, may be somewhat less than that of *Rf*, or it may be greater. The time constant for charge is the product of capacitance at *Cf* and the charging resistance, in microfarads and megohms, while the constant for discharge is the product of the same capacitance and the resistances carrying the discharge. Capacitor *Ca* is a bypass for intermediate frequencies around the volume control, and its capacitance is so small compared with that of *Cf* as to have negligible effect on time constants.

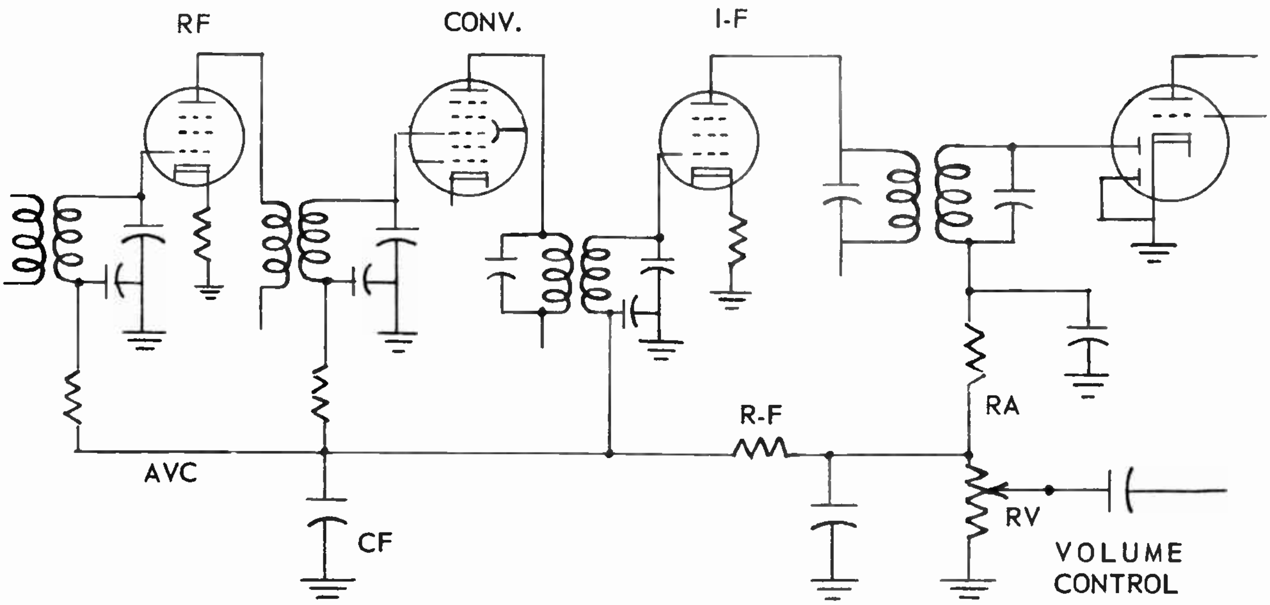


Fig. 48-13. Grid returns to avc bus through coupling coils.

The time constant for standard broadcast reception usually is about 0.1 second, or may be somewhat longer, possibly as great as 0.3 second. If the time is too short there will be a-f variations remaining in the avc voltage. These variations will act on amplifier biases to reduce the modulation of the signal in i-f stages and cause a form of distortion. There also will be poor reproduction of low audio frequencies or there will be poor bass response. High-fidelity audio systems often are preceded by avc systems with a time constant as long as 0.5 second.

If the time constant is too long there will be lack of volume control under such conditions as rapid fading of the received signal. Also, sharp pulses of interference, as from static, may drive the controlled grids far negative and audio output will stop until the excessive charge leaks away. This is the action called blocking.

2 In a receiver having one r-f amplifier, one converter, and one i-f amplifier the avc voltage usually is applied to all three. With no r-f amplifier the control is applied to the converter and i-f stages. This is true whether the amplifier tubes are remote cutoff or sharp cutoff pentodes. Remote cutoff tubes are used more than sharp cutoff types in both r-f and i-f amplifier stages. With more than one i-f amplifier there may be avc on the first one and cathode-bias on the second. Automatic volume control has its maximum effect on tubes closest to the antenna, and least effect on those nearest the detector.

Fig. 48-14 shows a method for using one of the two twin-diodes for avc and the other for detection. The complete diagram for both functions is at the left. The detector circuit is no different than others which have been examined. The upper end of the i-f transformer secondary is connected to one of the twin-diode plates and the lower end of the secondary is connected to the volume control resistor in the usual way. From the upper end of the secondary there is an additional connection through capacitor *Cd* to the avc bus

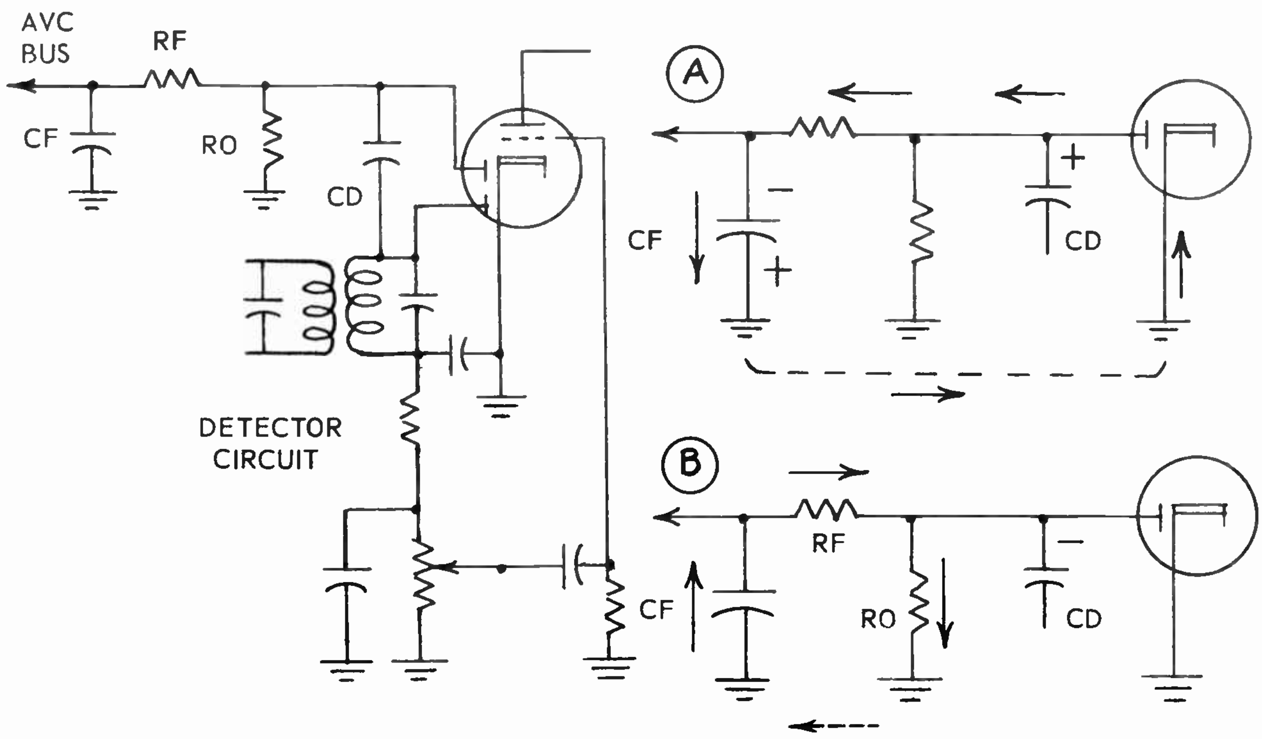


Fig. 48-14. Using one diode for detection and another for avc.

and the other diode plate. I-f signal voltage from the transformer is applied directly to the bottom diode plate, and is applied through capacitor *Cd* to the top diode plate. The purpose of this capacitor is to pass the i-f signal voltage to the top diode and the avc circuit while blocking the d-c voltage developed in the avc circuit so that this voltage does not affect the detector circuit.

The avc circuit is shown by itself in diagrams at the right. In diagram *A* the arrows indicate directions of electron flow when i-f signal voltage coming through capacitor *Cd* is of such polarity as to make the diode plate positive and to make the diode conductive. Capacitor *Cf* is here being charged in a direction which makes its plate connected to the avc bus of negative polarity. Diagram *B* shows electron flow when i-f signal voltage makes the diode plate negative, and the diode non-conductive. Here the capacitor *Cf* is discharging slowly through resistors *Rf* and *Ro*, and through the ground connections.

Fig. 48-15 illustrates other methods of using one diode for detection and the other for automatic volume control. In diagram *A* the voltage from the i-f transformer secondary is applied to the top diode plate and to avc capacitor *Cf* through a high resistance at *Ro*. When the diode plate is positive the capacitor charges through the diode. When the diode plate is negative the capacitor discharges through *Ro*, the transformer, the filter resistor, and the manual volume control resistor to ground.

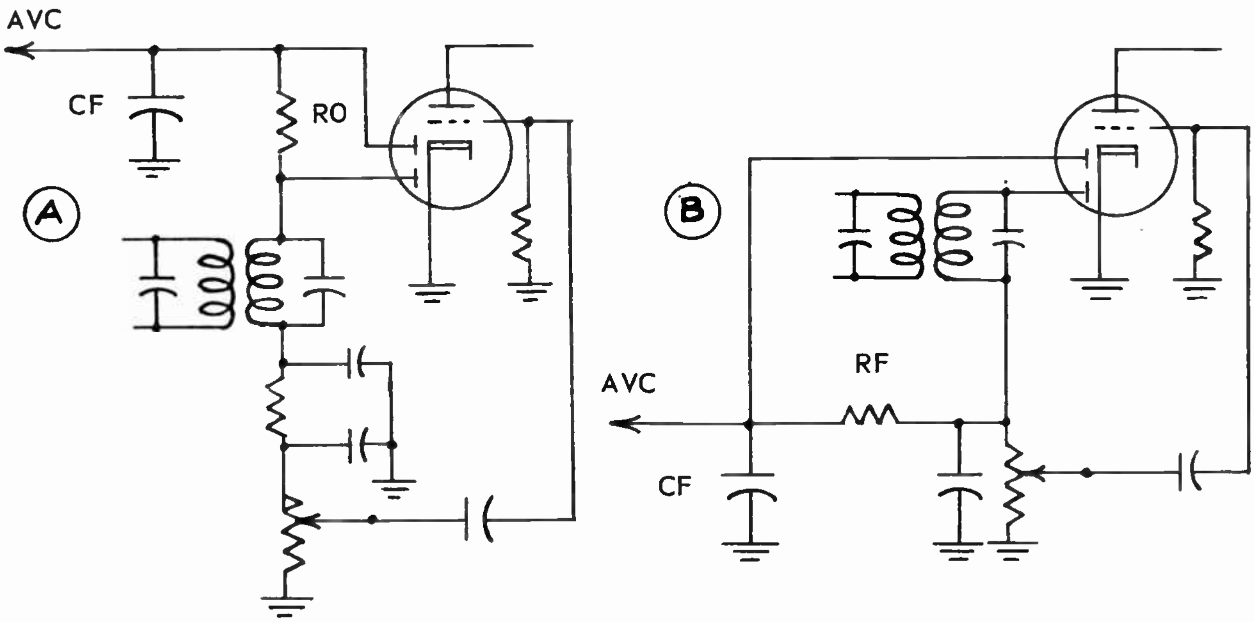


Fig. 48-15. Separate diodes for detection and avc.

In diagram *B* the i-f signal voltage is applied to the top diode plate and avc capacitor *Cf* from the bottom of the transformer secondary, for charging of *Cf*. This capacitor discharges through resistor *Rf* and the manual volume control resistor while the avc diode is non-conductive.

Fig. 48-16 shows, for a fairly typical receiver, the relations between antenna signal input in microvolts and audio output in volts. Assuming that amplification is uniform for all input voltages, frequencies, and waveforms, the audio output will increase in direct proportion to antenna input, as shown by curve *A*, which is a straight line on the graph.

Curve *B* shows how the audio output is cut down by automatic volume control systems such as those which have been examined. The output is reduced as amplification is lessened by the increasingly negative bias from the avc system. In order that this bias may become more and more negative with stronger antenna signals the signal voltage from the i-f amplifier must continually increase to some extent, which will increase the charge and voltage on the avc capacitor marked *Cf* in circuit diagrams. This continued increase of i-f amplifier output causes the audio signal to increase with stronger antenna signals, but the rate of increase (curve *B*) is much less than with no avc action (curve *A*).

The avc system whose effect is shown by curve *B* commences to act on even the weakest signal from the antenna. If good reception from far distant stations is wanted it is not desirable that the weakest signals be subjected to automatic control. Then it would be better to have an avc system acting as shown by curve *C*. Here there is no control until the antenna signal rises to about 20 microvolts, the audio output is the same as with no avc action (curve *A*). After this minimum input signal strength is reached the avc action commences and continues for all greater antenna signal strengths. Curve *C* represents the performance of what is called delayed automatic volume control.

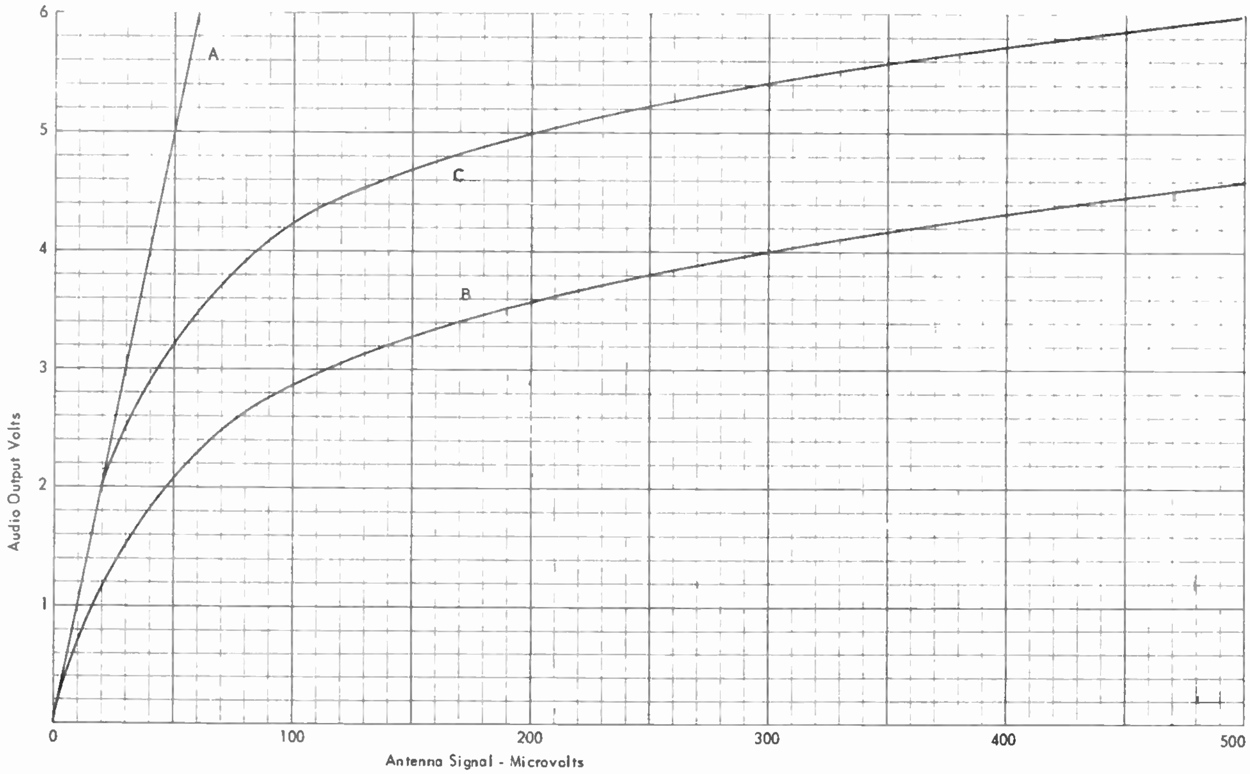


Fig. 48-16. Audio outputs with no avc (A), with simple avc (B), and with delayed avc (C).

A circuit commonly used for obtaining delayed avc is shown by Fig. 48-17. The principal difference between this delay circuit and the one without delay shown in Fig. 48-14 is that the duodiode-triode tube is operated with cathode-bias obtained from the cathode resistor at the bottom of the new circuit. All return connections in the detector circuit are to the cathode rather than to ground. Avc capacitor *Cf* and resistor *Ro* are connected on one side to ground, and through ground to the bottom of the biasing resistor.

The result of these connections is to place on the entire avc system, including the avc diode plate of the tube, a negative bias. That is, the avc diode plate is made negative with reference to the tube cathode by an amount equal to the voltage drop across the cathode-bias resistor. Then this avc diode cannot rectify and build up any biasing voltage on capacitor *Cf* until i-f signal voltage coming through capacitor *Cd* reaches a peak value equal to the bias voltage, and avc action is delayed until the i-f signal reaches this strength.

The bias voltage and delay voltage must be suitable for a-f amplifier action of the duodiode-triode tube, or the tube must be operated with plate voltage suited to the bias applied. Biasing and delay voltage usually is something between one and three volts. In the delay system it is necessary to use separate diodes for avc and detector functions, since a delay voltage in the detector circuit would cause serious distortion of the audio output.

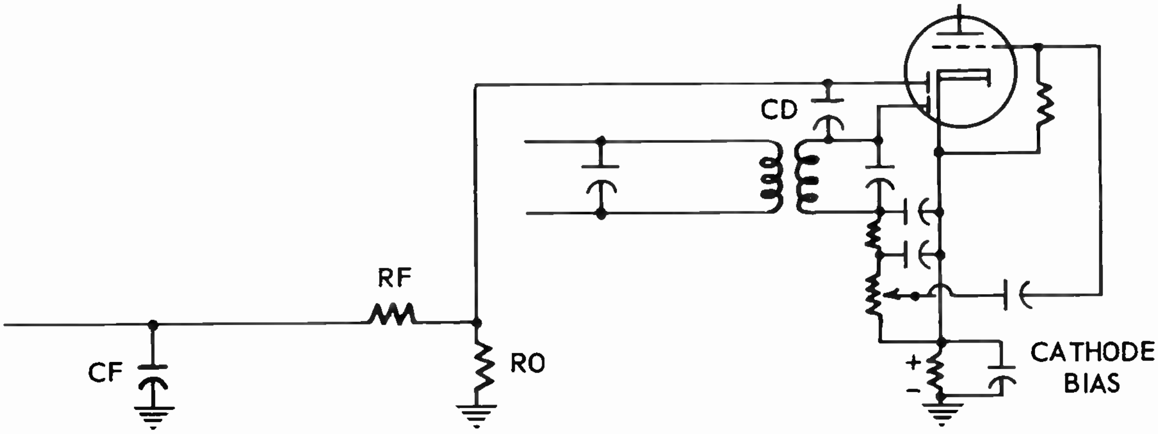


Fig. 48-17. Connections commonly used for delayed avc.

Not all detector diodes for a-m receivers are in duodiode-triode tubes. One variation is shown by Fig. 48-18, where a diode-pentode tube is used as the second i-f amplifier, detector, and avc tube combined. There is a single diode plate employed for detection and avc actiou. The pentode section is connected as an i-f amplifier, and a separate tube, over at the right, is provided for audio amplification.

The i-f transformer at the upper left brings the i-f signal to the grid of the combination tube. The plate of this tube connects to the primary of the next i-f transformer, shown below the combination tube. The secondary of this transformer is in the diode detector circuit. There is nothing new or unusual about the detector circuit. We have seen similar circuits many times before. The avc bus is connected from the top of the manual volume control resistor. The avc bias connection to the grid of the diode-pentode tube is included in the diagram. Avc leads to other controlled tubes would be taken from the bus farther to the left.

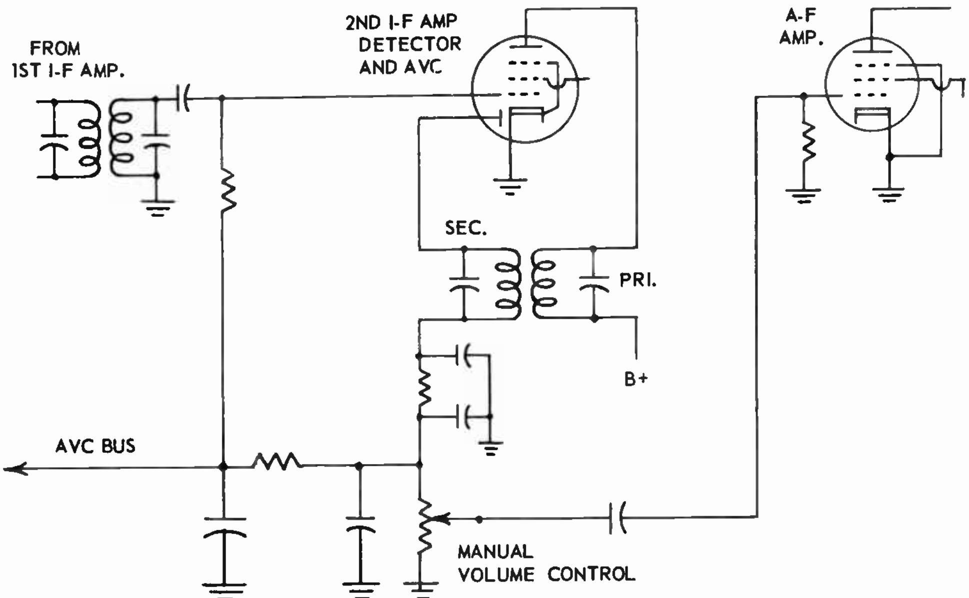


Fig. 48-18. Diode-pentode tube used as i-f amplifier, detector, and avc tube.

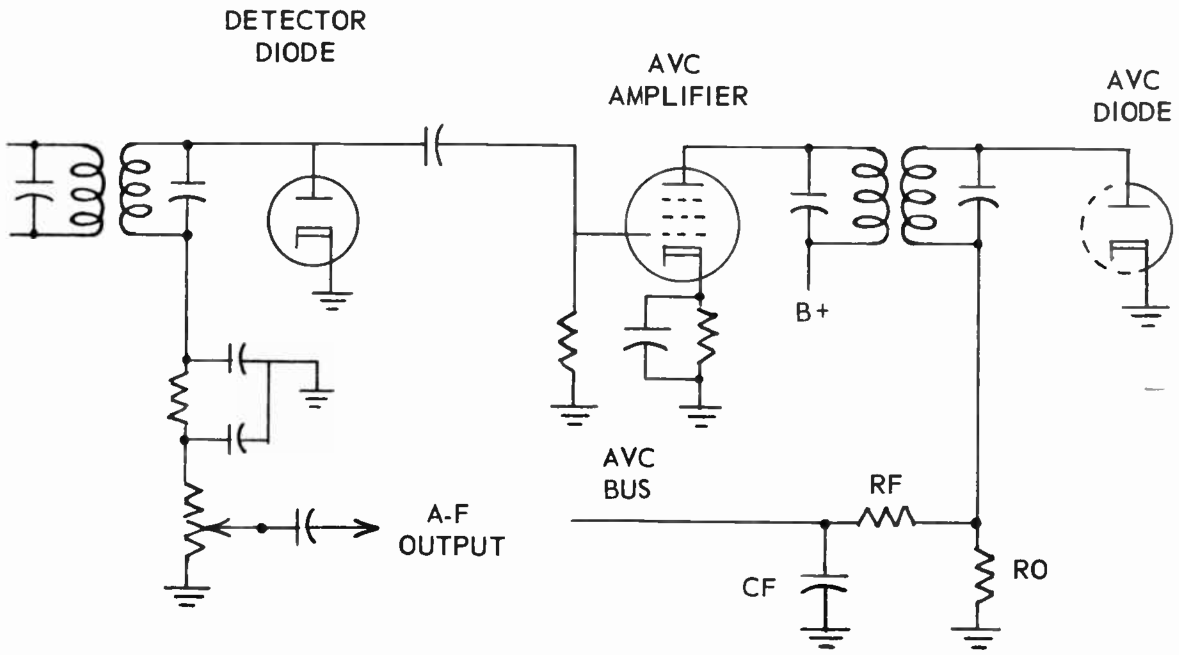


Fig. 48-19. Connections used for amplified avc.

In some receivers a diode-pentode tube is used with its diode functioning in detector and avc circuits, and with its pentode connected as the audio amplifier that follows the detector.

Avc action may be made to hold the audio output voltage more constant or on a flatter curve than shown back in Fig. 48-16 by employing what is called amplified avc. In such a system the secondary of the last regular i-f transformer feeds a separate diode that acts as the detector. Connected also to the secondary of this transformer is an additional i-f amplifier tube as shown in simplified form by Fig. 48-19. This is the avc amplifier. Following the avc amplifier is another transformer whose secondary feeds a second diode serving the avc function. The detector circuit and avc bus circuit are like others which have been examined. The two diodes ordinarily are sections of a twin-diode tube, or they may be elements in a duo-diode-triode or duodiode-pentode tube having suitable cathode connections for the several purposes.

In f-m receivers or combination fm-am receivers employing a ratio detector the avc voltage for f-m operation nearly always is taken from the negative side of the large capacitor which is across the ratio detector output. When f-m detection is by means of a discriminator the avc voltage is taken from the high side of the limiter grid resistor, or from the high side of the grid resistor on some amplifier tube in the i-f system.

When an avc voltage is to be measured, or its changes noted during service operations, the measurement is made with an electronic voltmeter or high-resistance d-c voltmeter connected between the avc bus and either ground or B-minus. The negative terminal of the meter is connected to the bus, where exists the negative avc voltage, and the positive terminal is connected to ground or B-minus.

DO NOT TEAR — CUT ALONG THIS LINE AND SEND IN FOR GRADING.

TELEVISION

LESSON NO. 49

CONTROLS FOR CONTRAST AND GAIN

Training Your Mind

C.T.I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

COMMERCIAL TRADES INSTITUTE

Training Your Mind... C.T.I. Perfecting Your Skill

Chicago, Ill.

LESSON NO. 49

CONTROLS FOR CONTRAST AND GAIN

The contrast control of a television receiver varies the strength of the signal voltage applied to the grid-cathode circuit of the picture tube. The stronger is this applied signal the greater is the difference between dark and light tones in the picture. That is, the darkest tones are more nearly black, the lightest tones are more nearly white, and all the intermediate gray tones are present in their correct proportions. Contrast, in any picture, is a measure of the difference between darkest and lightest tones. When there is a great difference we have high contrast, and with little difference there is low contrast.

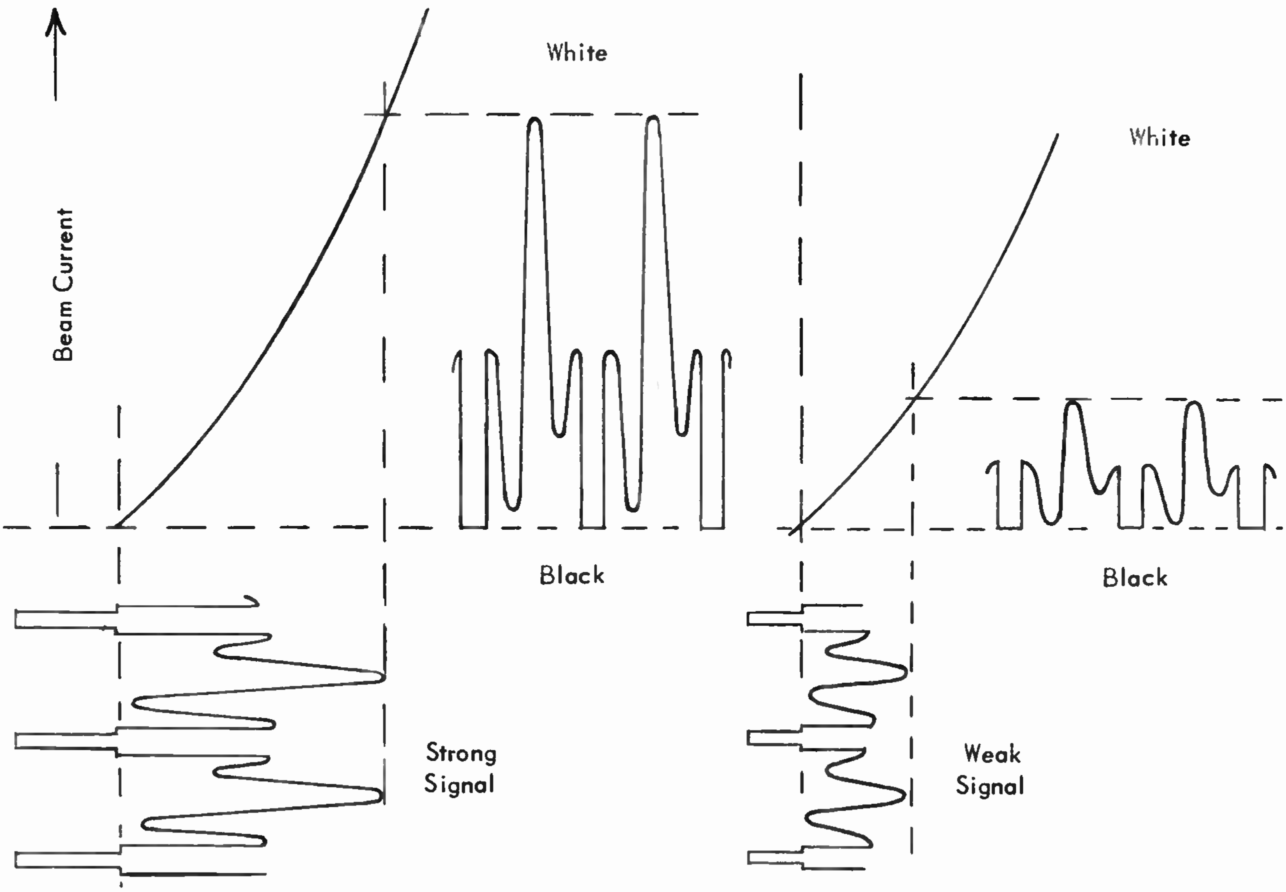


Fig. 49-1. How signal strength affects contrast in the picture.

The signal voltage applied between grid and cathode of the picture tube varies the electron beam current, just as signal voltage applied between grid and cathode of an amplifier tube varies the plate current. Light tones in the picture are produced by relatively large beam current, and dark tones result from small beam current. To say the same thing in another way, large beam current causes high brightness in the picture areas and small beam current causes low brightness. There is the same relation between signal voltage and beam current (or brightness) in a picture tube as between signal voltage and plate current in an amplifier.

The relations between signal voltage and brightness for a picture tube may be shown as in Fig. 49-1. This is the same type of "transfer characteristic" that we used earlier for showing amplifier performance. We shall assume that good picture reproduction results from the signal voltage represented at the left. Darkest tones are down almost to the black level, and the lightest tones are up nearly to the white level. If the strength of input signal is reduced, as at the right, the dark tones will remain nearly at the black level, but the lightest tones will drop far below the white level.

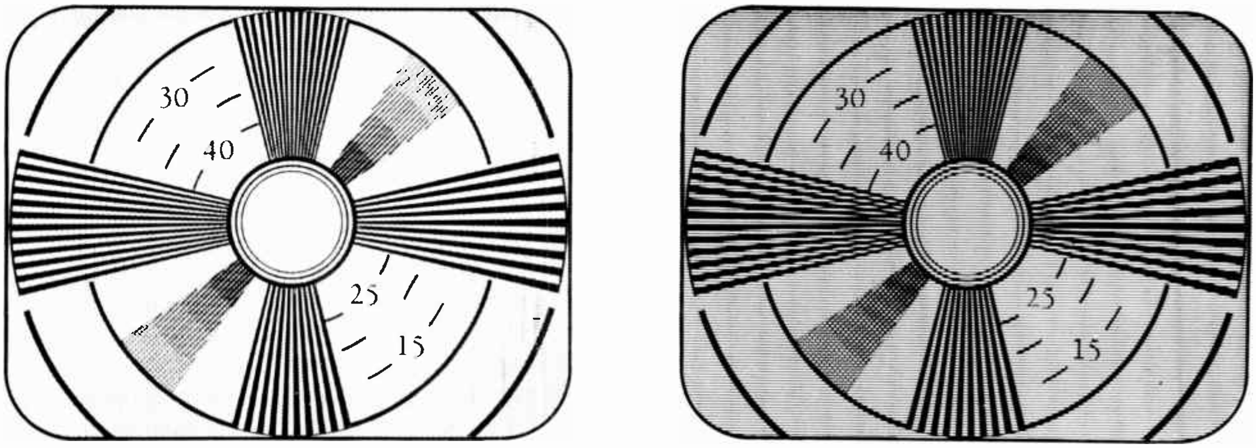


Fig. 49-2. Normal contrast in test pattern (left) and lack of contrast (right).

With picture tube signal voltage of the value assumed as satisfactory there might be reproduction of a television test pattern about as illustrated at the left in Fig. 49-2. The parts which should be black appear really black, and those which should be lightest are almost white. Intermediate gray tones also are well brought out, as is evident on the diagonal "tone wedges" of the pattern. With a picture tube signal voltage too weak the appearance of the pattern would change about as shown by the right-hand illustration. The darkest tones still are practically black, but the lightest tones are merely a dark gray instead of being nearly white.

The differences between appearances of the pattern, as illustrated, might result from changing the contrast control setting. The darker pattern might result also from a received signal so weak that even with maximum setting of the contrast control there still would not be enough signal voltage at the picture tube to produce a satisfactory pattern or picture.

It is a general rule that the contrast control varies the amplification or gain in the video amplifier, acting usually on one tube but sometimes on more than one when the video amplifier contains more than one amplifier. Amplification in the tuner and in the i-f amplifier ordinarily is regulated by an automatic gain control, which corresponds to the automatic volume control of a sound receiver.

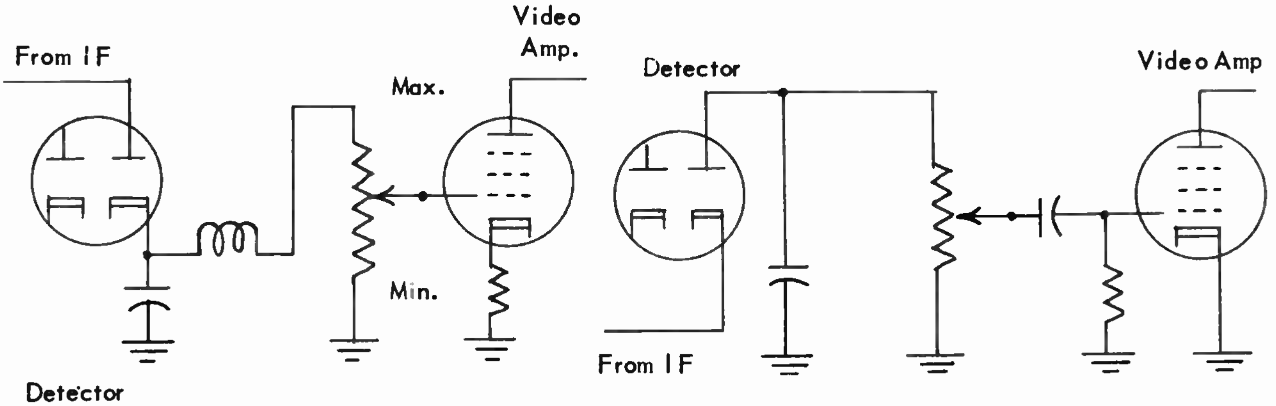


Fig. 49-3. Contrast controls on the video detector load resistor.

CONTRAST CONTROLS. It is possible to vary the gain of the video amplifier system in many ways; there is great variety in contrast controls. Two of the simplest methods are illustrated by Fig. 49-3. The load resistor on the video detector is a potentiometer whose ends are connected to the detector output and to ground. The slider of the potentiometer is connected to the grid of the following video amplifier tube. With the slider at the detector end of the potentiometer there is maximum signal voltage applied to the amplifier grid, for this voltage then is the total drop across the potentiometer or load resistor. With the slider at the grounded end of the potentiometer there is minimum or zero signal voltage for the amplifier grid, since the grid then is connected almost directly to ground and does not receive any voltage developed in the potentiometer.

Fig. 49-4 shows some contrast controls formed by an adjustable resistor in the cathode line of a video amplifier tube. This control might be on the first or second or any other video amplifier. The control of diagram A varies the cathode-bias of the amplifier and thereby varies the transconductance and gain of this tube. In diagram B the contrast control varies the bias on the amplifier to some extent, but the control resistance may be of such value as to materially alter the plate current. Reducing the plate current by increasing the contrast control resistance will lessen the transconductance and gain of the amplifier. An auxiliary negative grid bias voltage is connected to the grid circuit through one resistor and cross connected to the cathode through a higher resistance.

Diagram C of Fig. 49-4 is the circuit for a cathode type contrast control as used in a receiver having the plate-cathode circuits of the sound and video systems in series. We examined this type of plate-cathode connection when studying B-power supplies. With the contrast control setting as shown in the diagram the tube cathode would be more positive than the B+ line to which it connects through the control unit. There is grid-leak bias on the amplifier tube by means of grid capacitor Cg and grid resistor Rg.

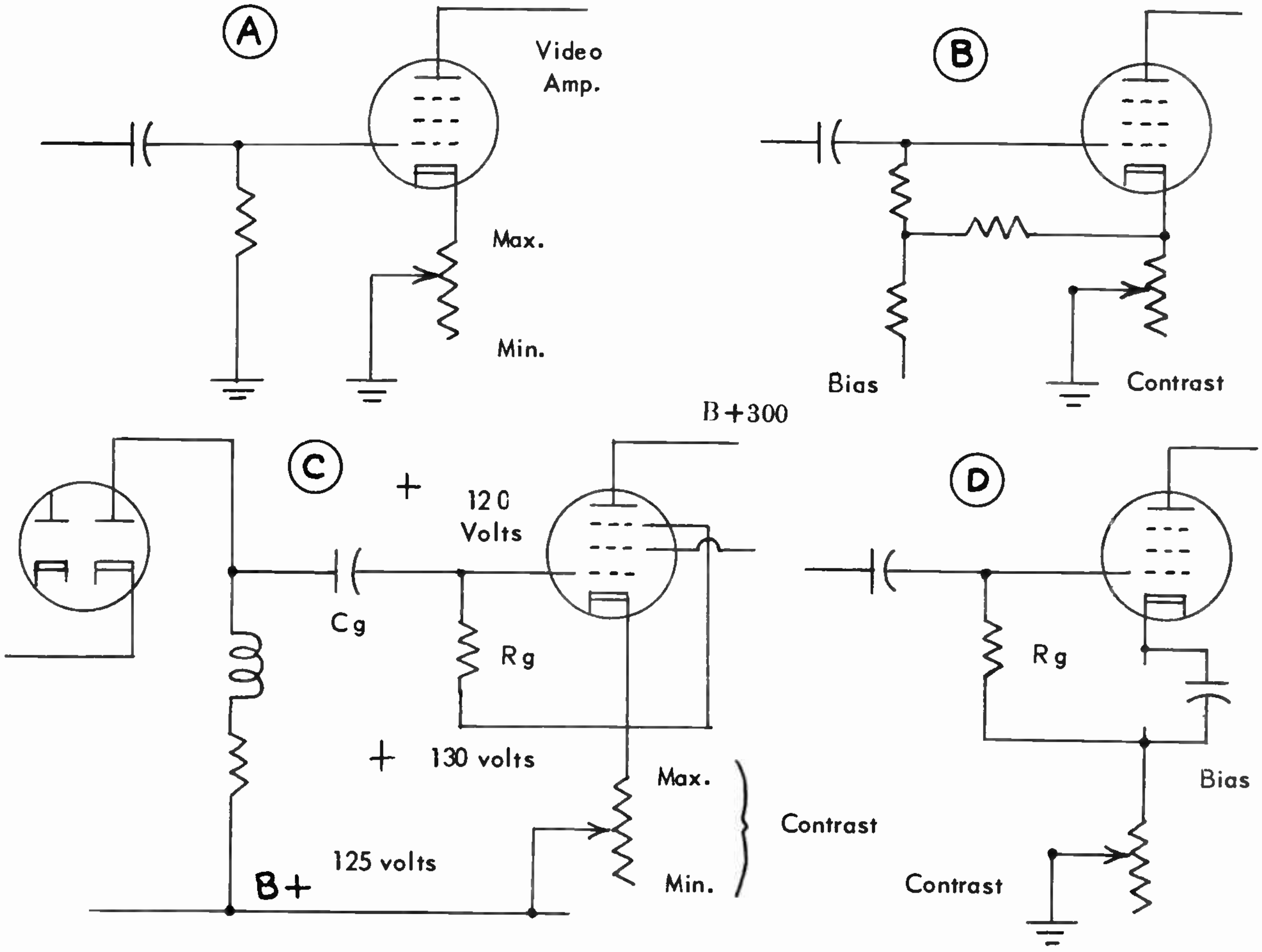


Fig. 49-4. Contrast controls on cathodes of video amplifiers.

This bias voltage is developed across resistor Rg, which is between grid and cathode, not between grid and ground or grid and B+. Adjustment of the contrast control varies the resistance and thereby the current in the plate-cathode and screen-cathode circuits, thus varying the transconductance and gain of the amplifier tube.

In diagram D the contrast control again acts chiefly in varying plate current and transconductance of the amplifier. There is cathode-bias for the grid by means of the cathode resistor and bypass capacitor connected directly to the cathode. Resistor Rg is merely a grid return connection, and does not act to give grid-leak bias. Another way of explaining the action of cathode resistors which are not used for variable grid bias is to consider that they change the amount of degeneration. Degeneration is a feed-back of signal voltage in polarity or phase opposite to that applied to the grid from the detector or a preceding amplifier. Increasing the cathode resistance increases the degenerative voltage, and thus reduces the gain of the amplifier tube.

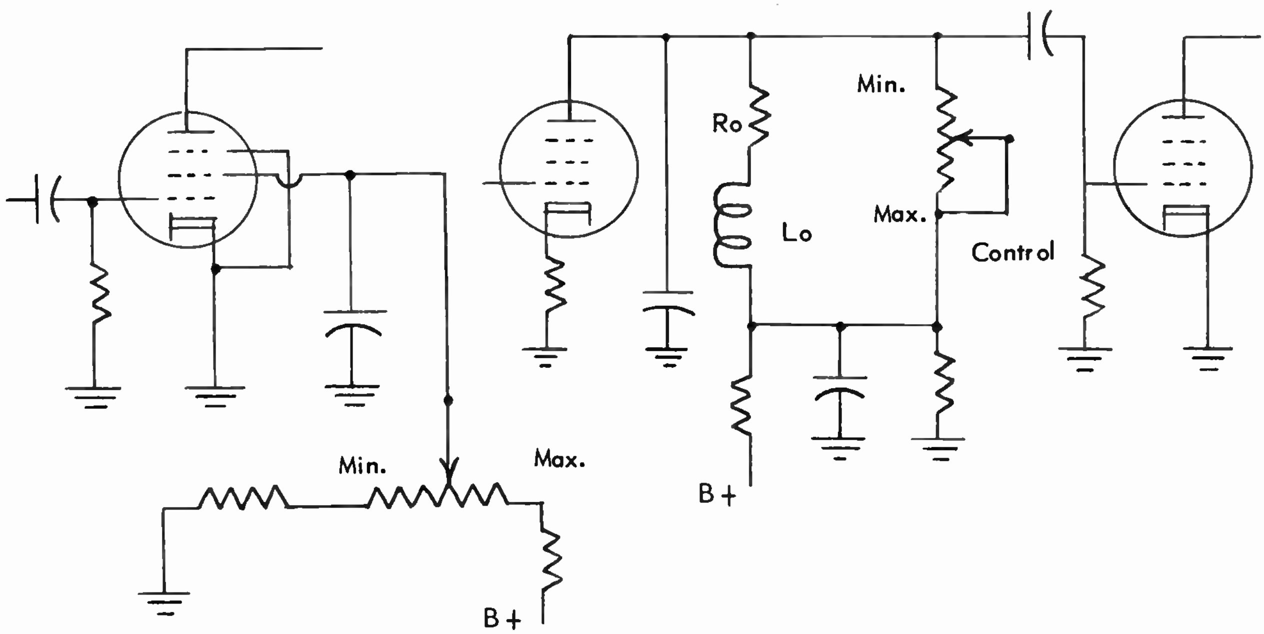


Fig. 49-5. Contrast control by adjustable screen voltage (left) and by adjustable plate load (right).

The transconductance of a pentode amplifier increases with higher screen voltage and decreases with lower screen voltage. Consequently, as at the left in Fig. 49-5, a screen voltage adjustment may be used for contrast control. A potentiometer is connected in a line between B+ and ground, with the slider connected to the amplifier screen. The screen is bypassed to ground with a large capacitor, usually on the order of 10 mf capacitance.

Earlier we learned that the gain in a video amplifier stage depends almost directly on the resistance or impedance of the load in the plate circuit. This fact is utilized for the contrast control shown at the right in Fig. 49-5. In parallel with load resistor Ro and compensating inductor Lo is the adjustable resistor of the control unit. Short circuiting more of the control resistance drops the plate load, for less gain and less contrast, while using maximum control resistance has opposite effects.

Contrast control may be had from variation of gain in one or more i-f amplifier tubes, or in both i-f and r-f amplifiers. Fig. 49-6 illustrates the principle of a method quite often employed for such contrast controls. Biasing voltage for the controlled tubes is provided by a rectifier and filter system used for this sole purpose. The rectifier may be one section of a twin-diode whose other section is used as the video detector, or the rectifier may be a selenium type such as used in power supplies. Either type of rectifier is fed alternating voltage taken from a tube heater circuit in the receiver.

Smooth direct voltage is obtained from the rectifier by a large bypass capacitance at Ca which is across the control resistor and one or more other resistors in series with the control. Additional filtering action

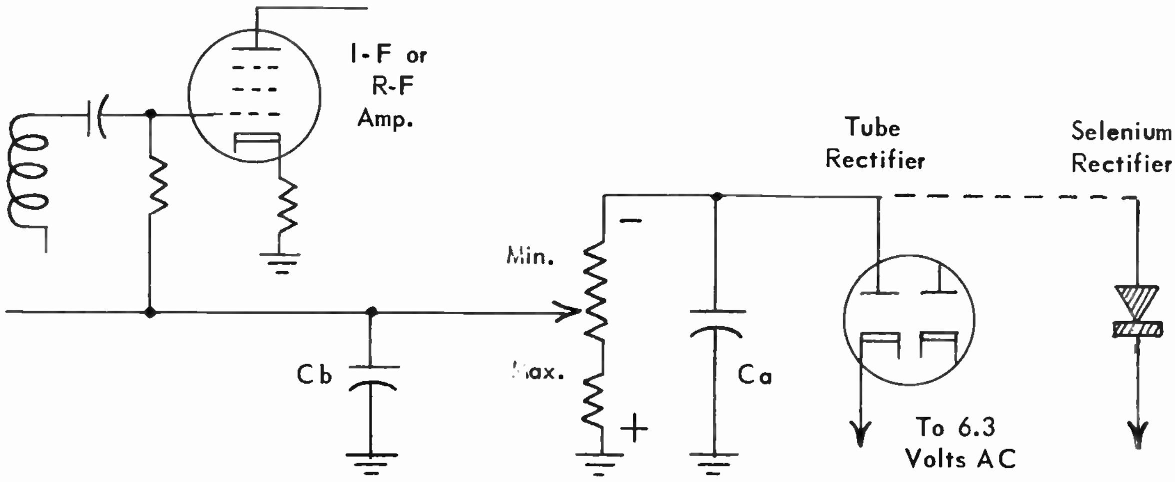


Fig. 49-6. Rectifiers providing direct voltage for variable bias used as contrast control.

is had from capacitor Cb. The slider of the control resistor is connected to the grid returns of the controlled tubes. The value of negative bias on these tubes then is equal to the voltage drop between the control slider and ground, plus any additional bias secured from cathode resistors on the controlled tubes. Making the bias more negative reduces amplification and contrast. Bias less negative, increases amplification and contrast.

A generally similar contrast control for i-f amplifier tubes may be had with voltage taken from the B-minus side of the plate power supply, much as a fixed negative bias voltage may be taken from the plate power supply for any tubes. This latter method requires no separate rectifier for contrast control.

AUTOMATIC GAIN CONTROL. Automatic gain control in the television receiver maintains fairly constant signal strength at the picture tube input after the contrast control has once been adjusted for satisfactory reception of a certain station. If the automatic gain control is more than usually effective, it may make unnecessary any great readjustment of contrast and brightness controls when switching from weak to strong stations or vice versa. There are advantages also in preventing such things as power line surges, swaying antenna and transmission line, or low-flying airplanes from making sudden changes of picture contrast.

The circuit for a simple automatic gain control (agc) system as shown by Fig. 49-7 looks quite like that for an automatic volume control in sound radio, except that one section of the twin-diode tube is used for detection and the other for agc. The output from the last i-f amplifier is fed directly to the detector diode and is fed through capacitor Ca to the agc diode. This capacitor blocks the direct voltage developed in the agc circuit from getting into the detector circuit, but passes the i-f voltage which is to be rectified for gain control. The functions of capacitor Cf and of resistors Rf and Ro are exactly the same as explained for similarly lettered parts in avc systems.

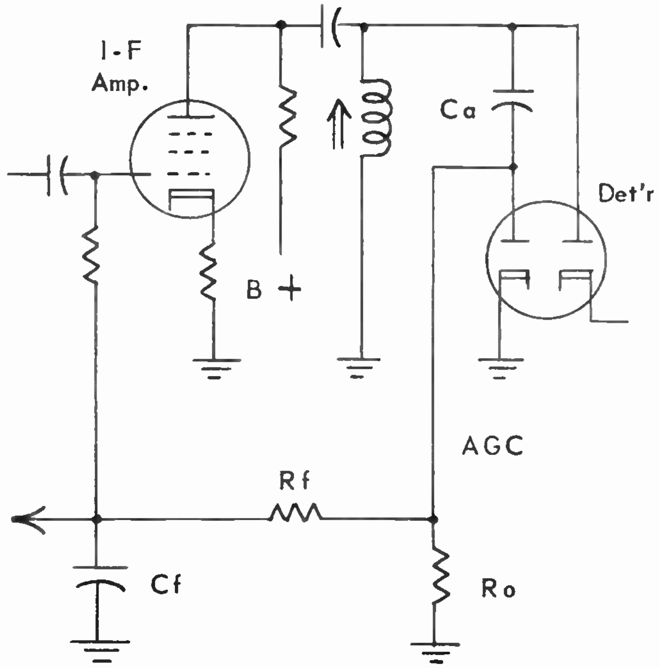


Fig. 49-7. A simple automatic gain control.

Avc voltage in a sound receiver is of a value proportional to average strength of the received signal and the i-f amplifier output. In a television agc system we cannot use the average strength of the received signal for reasons which become apparent when we examine Fig. 49-8. Up above is a rectified or detected video signal in which the picture is of generally light tone, most of the picture variations are near the white level. Down below is a signal for a picture of generally dark tone, with picture variations close to the black level.

For the light-toned picture the video signal (picture variations and sync pulses) is of much greater height or shows a much greater overall change of voltage than does the video signal for the dark-toned picture. When these two signals pass through one or more capacitors, and become alternating currents or voltages, the amplitude for the light-toned picture would be much greater than for the one of dark tone. In other words, the light-toned signal is stronger than the dark-toned one when both become alternating voltages or currents as applied to the agc rectifier. Then the agc voltage developed for a light-toned picture would be more negative than for a dark-toned picture. Amplification of controlled tubes would be reduced when pictures are of light tone, and increased for dark toned pictures. The resulting weaker output signal for light-toned pictures would make them darker (less beam current), and the stronger output signal for dark-toned pictures would make them lighter.

With action such as just explained the agc system would be lessening or destroying picture contrast. Lightening the dark tones and darkening the light ones would bring all of them to intermediate grays, which certainly is not the intended purpose of automatic gain control.

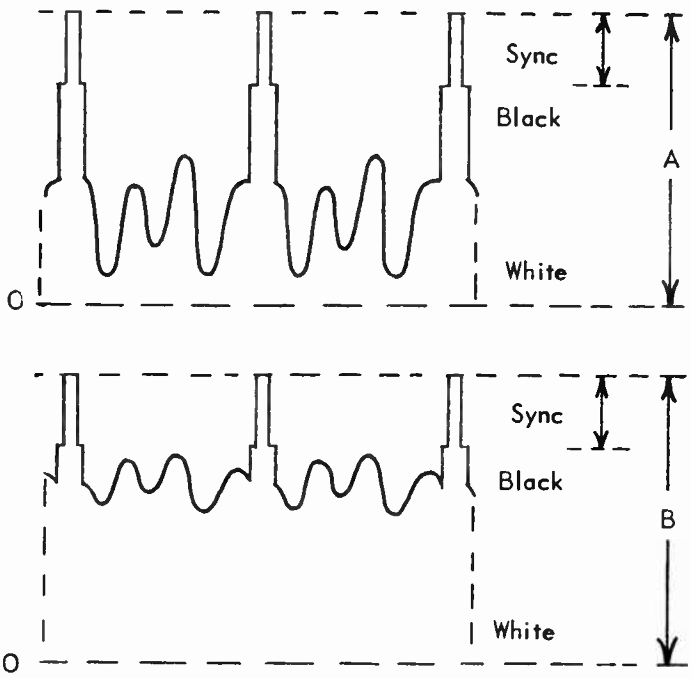


Fig. 49-8. Tone of a picture affects amplitude of its signal.

For satisfactory agc action the biasing voltage must be developed from some kind of signal voltage that varies only with changes in strength of signal received at the antenna and delivered from the output of the i-f amplifier, and one that is not affected by changes in tone of the pictures. Such a voltage is that between zero and the tips of the sync pulses, as marked A and B in Fig. 49-8. Voltage A in the light-toned picture signal is exactly the same as voltage B in the dark-toned picture signal.

Voltage at the sync pulse tips will vary only with changes in strength of the received signal. With a stronger received signal the sync pulse tips will be farther from zero, on a voltage basis, and with a weaker received signal the pulse tips will be closer to zero, or will be at a lesser voltage. Voltage at the pulse tips then should be satisfactory as the origin of the agc biasing voltage. The agc voltage across capacitor Cf of Fig. 49-7 will remain practically equal to the voltage at sync pulse tips if the discharge time constant of this capacitor and resistors Rf and Ro in series is made considerably longer than the time between successive horizontal pulses, so that very little charge can escape from Cf during the period of one horizontal line.

When the discharge time constant of capacitor Cf is long enough to maintain sync pulse peak voltage for agc action any pulses of interference or noise voltage reaching the agc system will add to the charge on Cf, and will make the agc bias voltage more negative. Continuing interference or noise voltage can make the agc bias so negative as to stop amplification in controlled tubes and shut off the picture. Prevention of such undesirable effects is a major problem in the design of television agc systems. Some of the methods will be considered a little later.

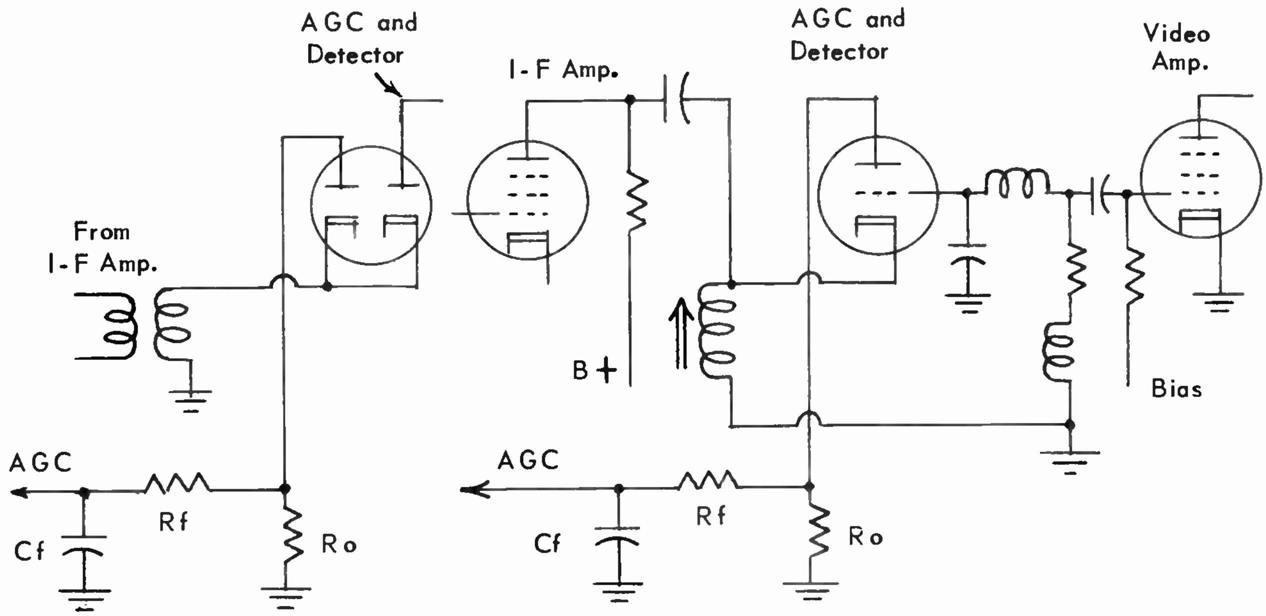


Fig. 49-9. Automatic gain controls with i-f input to rectifier cathodes.

The i-f output voltage which is to be rectified and filtered for agc purposes may be applied to the rectifier cathode as in Fig. 49-9. In the left-hand diagram a twin-diode tube serves as agc rectifier and video detector. The agc filter system, down below, is connected to the rectifier plate and is generally the same as other circuits which have been examined.

At the right in Fig. 49-9 a triode tube is used as agc rectifier and video detector. Input from the last i-f amplifier is to the cathode of the agc-detector triode. The grid of the triode acts as the plate of a diode, in connection with the cathode, for the detector functionl The triode plate acts as the plate of a diode for agc rectification. Again the agc filter system connected to the plate is like others with which we are familiar.

When automatic gain control is applied to i-f and r-f amplifiers of television receivers the variations of amplifier tube grid bias may cause an undesirable detuning of the controlled stages. This detuning usually is spoken of as related to the "Miller effect". The Miller effect is a change of capacitance which appears in the grid circuit of a tube when the plate load is a resistance or is partly a resistance. This capacitance is altered when there is a change of grid bias, as from agc voltage. The change of capacitance is not serious in low-frequency circuits where there is a tuning capacitance so large that the total capacitance is not affected but it will change the tuning or resonant frequency or frequency response where there is relatively small total capacitance, as in r-f and i-f amplifier circuits. A change in frequency response of controlled stages will change the points at which the video and sound carrier or intermediate frequencies fall on the curve, and will alter the gains at these important frequencies.

The Miller effect may be partially compensated for by using a cathode resistor across which is no bypass capacitor, which allows a degenerative feedback as mentioned earlier. Sometimes part of the cathode resistor is bypassed with a capacitor and the remainder is not bypassed. In some receivers an additional small capacitor is connected into the grid circuit to correct the detuning for reception under conditions where only a slight change would be harmful.

(2) *DELAYED AUTOMATIC GAIN CONTROL.* Delayed automatic gain control in television is used for the same purpose as delayed automatic volume control in sound receivers, to allow better reception of very weak signals while retaining automatic control for stronger signals. The delay most often is obtained by making the cathode of the agc rectifier somewhat positive. Then this diode cannot conduct until peaks of sync pulse voltage from the i-f amplifier become strong enough to make the diode plate still more positive than the cathode, or make the plate positive with reference to the cathode.

Fig. 49-10 shows a simple delay system. Positive delay voltage on the rectifier cathode is the potential difference across a resistor between the B+ line and ground. If, for example, this delay potential is two volts the diode will conduct only when sync pulse amplitude from the i-f amplifier exceeds two volts. Only when the signal exceeds this value will there be conduction in the diode, charging of agc capacitor Cf, and a negative voltage on this capacitor for the agc bus.

The delay voltage may be of fixed value, usually between one and three volts. In other cases the delay may be varied by adjusting the value of resistance across which is the delay voltage. This adjustment may be in steps, by means of resistor taps connected through a selector switch to the cathode of the agc diode.

We should note that the signal from the i-f amplifier may be applied to either the plate or the cathode of the detector diode, while applying this i-f signal to the plate or the cathode of the agc diode. In Fig. 49-7 both inputs are to the plates. In Fig. 49-9 both are to the cathodes. In Fig. 49-10 the i-f input is to the detector cathode and to the plate of the agc diode.

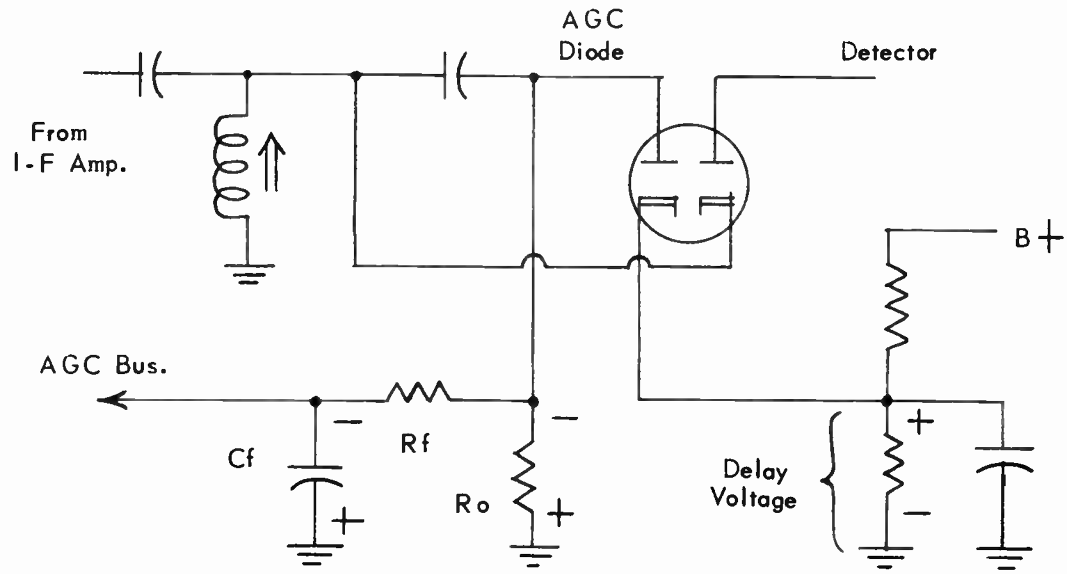


Fig. 49-10. Circuits for a simple delayed automatic gain control.

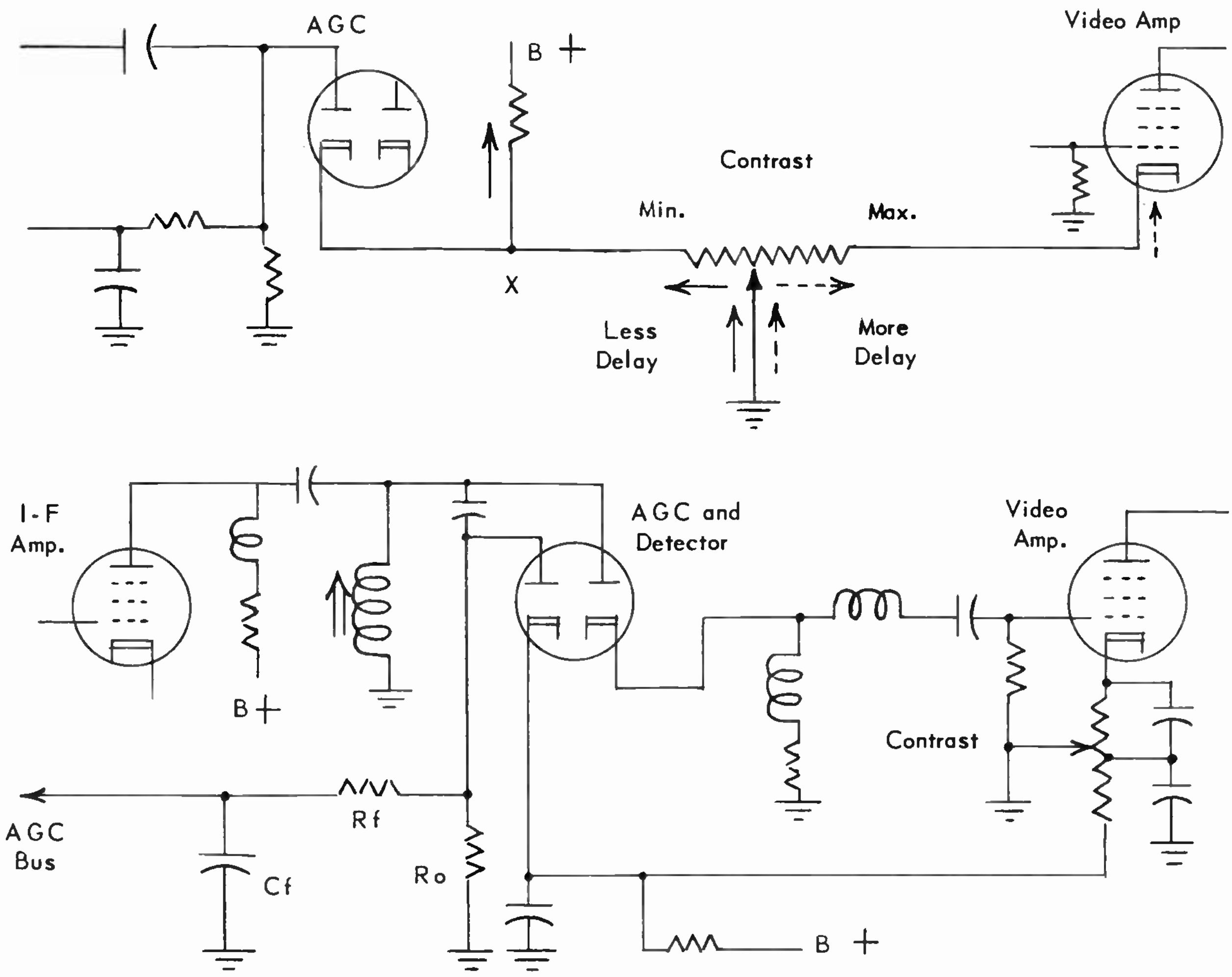


Fig. 49-11. A single adjustable resistor used for contrast control and for varying the agc delay voltage.

In a number of delayed agc systems a smaller delay voltage is applied to the r-f amplifier than to the i-f amplifiers. This means that agc action will be applied to the i-f amplifiers before it is applied to the r-f amplifier as received signal strength is gradually increased. The r-f amplifier continues to operate with its full gain after the gain of the i-f amplifiers has commenced to be reduced by agc action. The two different agc voltages, or more than two if desired, are obtained by connecting a resistor type voltage divider on the grid return side of the agc capacitor marked Cf in diagrams. Then different agc voltages may be taken from different taps on the voltage divider, but all these agc voltages will increase on strong signals and decrease on weak signals.

This general method of agc operation may improve the ratio of desired signal to noise voltage from the entire r-f and i-f amplifier system. Noise voltages caused by small and random electron flow in tubes,

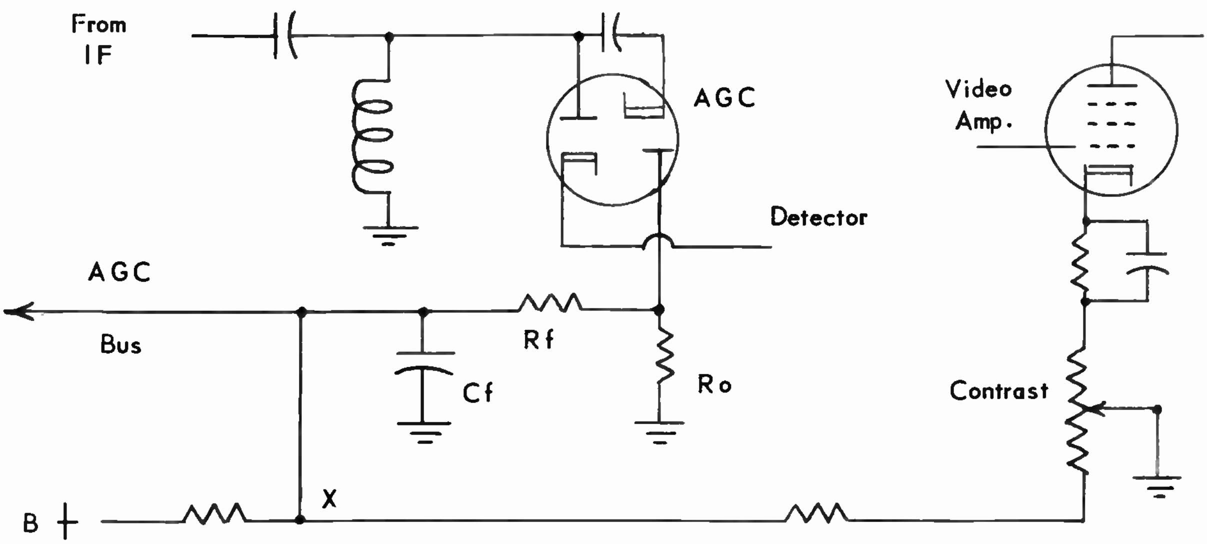


Fig. 49-12. A delay voltage from contrast control applied to agc bus.

resistors, and conductors generally are more or less independent of gain in the stage where noise originates. Noise voltages developed in the r-f stage are amplified by all the following tubes, and have much worse effect than noise voltages originating in the later stages, which are followed by less amplification. Then the stronger the signal at the output of the r-f amplifier the less will be the relative strength of noise voltages at this point. With both r-f signal output and noise voltage equally amplified in all following stages there still will be more signal than noise at the final output from the i-f amplifier.

Delay voltages usually are in such proportion that agc action on the r-f amplifier commences only after the signal from the antenna is strong enough that the r-f output might overload following i-f amplifiers. Were there no agc action on the r-f amplifier its output on strong received signals could overload the i-f amplifier, which would reduce the i-f gain and cause weak picture reproduction. In some receivers there is a switch for cutting out the agc action while tuned to very weak received signals.

There are many television receivers in which the agc delay voltage is altered by the same adjustable resistor that acts as a contrast control for the video amplifier. When reception is of a weak signal the operator naturally increases the contrast control, which increases the gain of the video amplifier. This action increases the delay voltage, thus allowing r-f and i-f amplifiers to operate without automatic gain control or with full amplification until stronger signals are received.

One of the methods quite commonly used for interconnecting the contrast and gain controls is shown by the simplified diagram at the top of Fig. 49-11. The contrast control unit is a potentiometer whose slider is connected to ground. One of the ends is connected to the cathode of a video amplifier tube and the other end is connected to the cathode of the agc diode, and also to the B+ supply through a resistor. Broken line arrows show the direction of electron flow from ground through part of the control resistance and to the cathode of the video amplifier. Full-line arrows show the direction of electron flow from ground (which also is B-minus) through the other part of the control resistor to the B+ line.

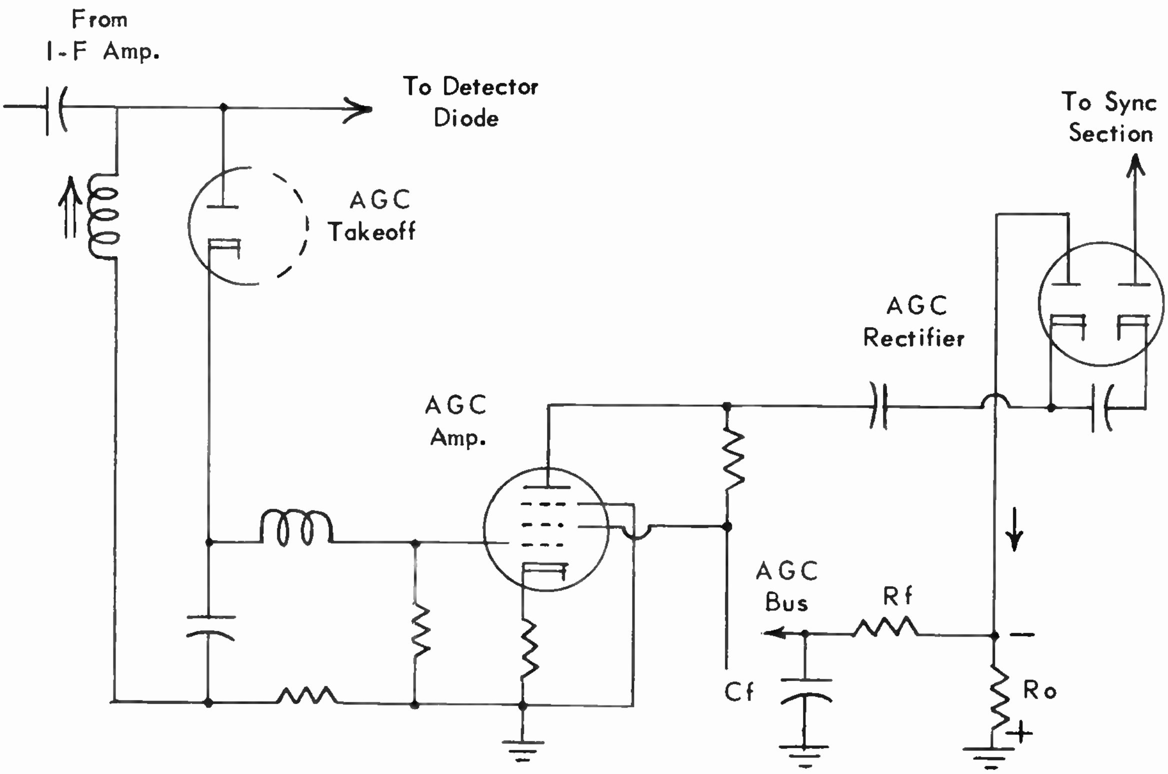


Fig. 49-13. One method for amplifying the agc voltage.

Agc delay voltage is the voltage drop in the portion of the resistor which is between the slider and the B+ connection. This is the voltage between ground and point X, which connects to the cathode of the agc diode. Moving the slider toward the agc cathode reduces the delay voltage, moving it toward the amplifier cathode increases the delay voltage.

Positive voltage on the video amplifier cathode, with respect to ground, is the voltage drop in the portion of the control resistor which is between the slider and the amplifier cathode. Positive voltage on an amplifier cathode is equivalent to negative bias on the grid when the grid is connected (through ground) to the slider of the resistor. Moving the slider toward the amplifier cathode reduces the negative bias on this tube and increases the gain or contrast. Moving the slider the other way reduces gain and contrast.

A control such as described would be shown on a service diagram somewhat as in the larger diagram at the bottom of Fig. 49-11. The description of control action in the upper diagram applies equally well to the lower diagram. The two bypass capacitors across the contrast control resistor in the lower diagram are of different values, possibly 100 mmf nearest the cathode and 300 mmf or more nearer the ground connection. This varies the degeneration to allow more amplification at high video frequencies than at low frequencies. We shall examine the whole subject of degeneration a little later on.

Fig. 49-12 shows another way of utilizing two voltage drops across a contrast control potentiometer, once for controlling gain in a video amplifier tube and the other for an agc delay voltage. Directions of

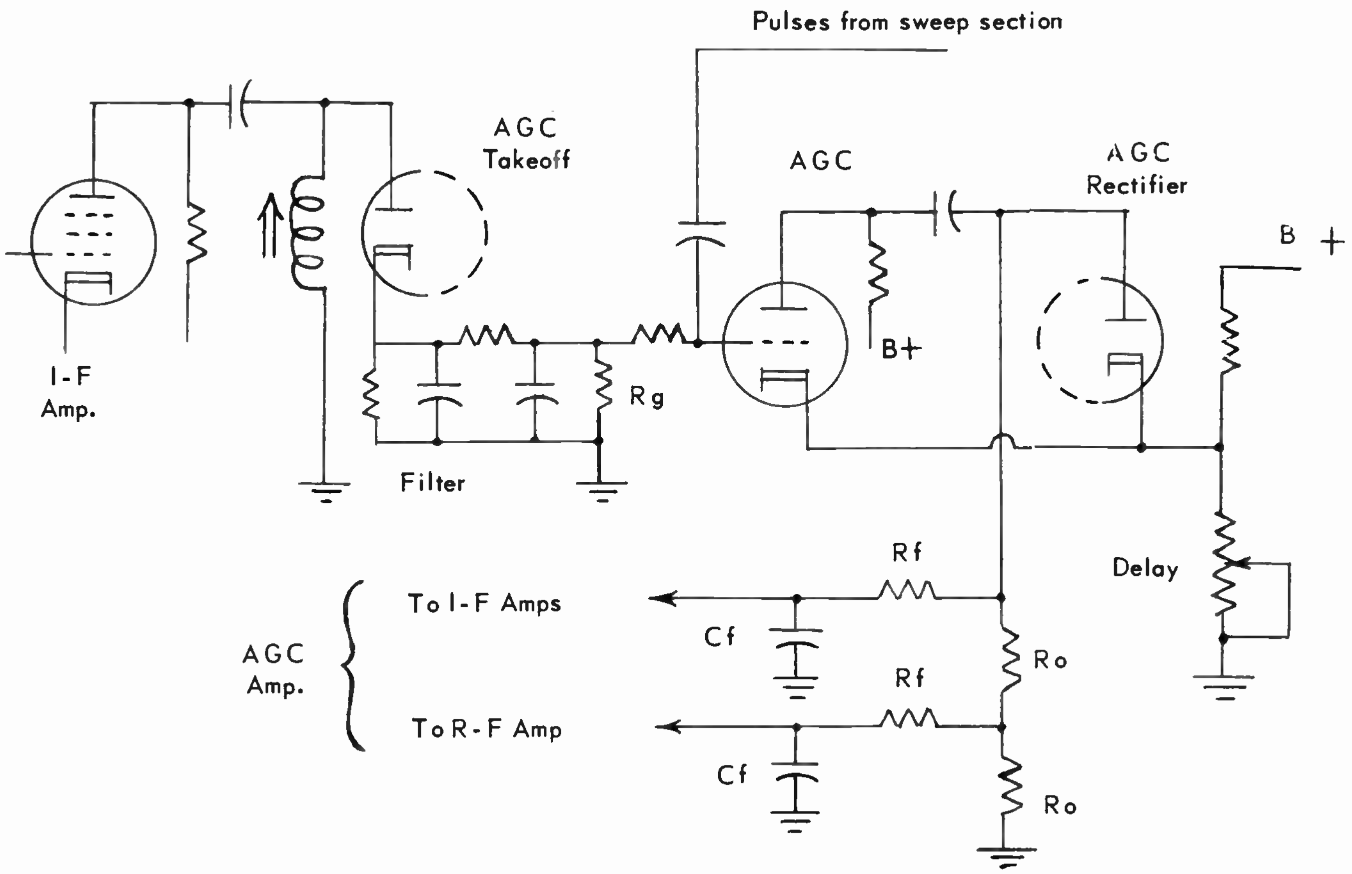


Fig. 49-14. I-f output signal voltage used for regulating the gain of an agc amplifier.

electron flow and divisions of voltage drop in the contrast control resistance are the same as previously explained. At point X there is a positive voltage with reference to ground, this voltage being applied directly to the agc bus rather than to the cathode of the agc diode. This positive delay voltage counteracts more or less of the negative agc voltage developed across capacitor Cf and thereby alters the signal strength at which the agc voltage becomes effective in limiting amplification of r-f and i-f amplifiers.

(a) *AMPLIFIED AUTOMATIC GAIN CONTROL.* Automatic gain control systems often include an amplifier tube which increases the agc or biasing voltage. The amplifier may precede or follow the agc rectifier. In agc systems which provide amplification, there practically always is delay action in addition.

Fig. 49-13 is a somewhat simplified circuit diagram for an agc system in which the amplifier comes ahead of the agc rectifier. The signal from the last i-f amplifier is applied to the plate of a diode acting as a takeoff for the agc system. Another line goes to a separate video detector diode. The rectified voltage from the takeoff diode is applied to the grid of the agc amplifier through a filter that removes remaining intermediate frequencies and delivers positive sync pulses to the amplifier. The amplifier output voltage is applied to the cathode of the agc rectifier diode, whose plate is connected to the usual

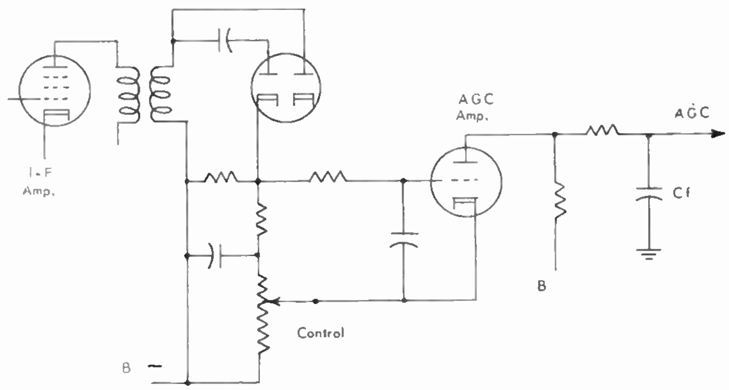


Fig. 49-15. A negative-going plate voltage may be used for automatic gain control.

type of agc filter and the agc bus. The second section of this final twin-diode carries amplified sync pulses to the sync section of the receiver.

Fig. 49-14 is a simplified circuit diagram for an amplified agc system in which are several somewhat unusual features. We should note first that the voltage which is amplified and then rectified to provide the final negative bias on controlled tubes is not obtained from the i-f output, rather it consists of voltage pulses taken from the sweep section of the receiver and applied to the grid of the agc amplifier. These pulses are at a frequency of 15,750 cycles per second, which is the horizontal line frequency. The signal voltage from the i-f amplifier is used to control the bias on the grid of the agc amplifier. This varies the amplification applied to the voltage pulses from the sweep section, and thus varies the value of the agc voltage for controlled tubes.

Signal voltage from the i-f amplifier is rectified by the takeoff diode. The rectified d-c voltage from this diode is passed through a rather elaborate filter of resistors and capacitors to make it a smooth direct voltage, which appears across resistor Rg in the amplifier grid circuit. This grid voltage is positive with reference to ground, but the cathode of the agc amplifier is made still more positive by connecting it to the B+ line that goes to the delay circuit over at the right. Thus the amplifier grid is effectively negative with reference to the cathode. A stronger signal from the i-f amplifier makes the amplifier grid more positive, or really less negative with reference to the cathode, and increases the amplification of this tube. This greater amplification works on the pulse voltages coming from the sweep section.

The amplified pulse voltages from the plate of the agc amplifier are applied to the plate of the following agc rectifier. The rectifier cathode is connected to an adjustable positive delay voltage, the principle of which has been examined earlier.

The plate of the agc rectifier is connected to the circuits, down below, in which the agc voltage is developed in the usual manner. Here, however, the total voltage drop is divided between two resistors marked Ro and between two capacitors marked Cf. The higher agc voltage from the top of this divider system goes to the grid returns of i-f amplifiers. The lower agc voltage from farther down on the divider goes to the grid return of the r-f amplifier. This allows the r-f amplifier to operate at full gain or without agc action until received signals are relatively strong, all as discussed in preceding pages.

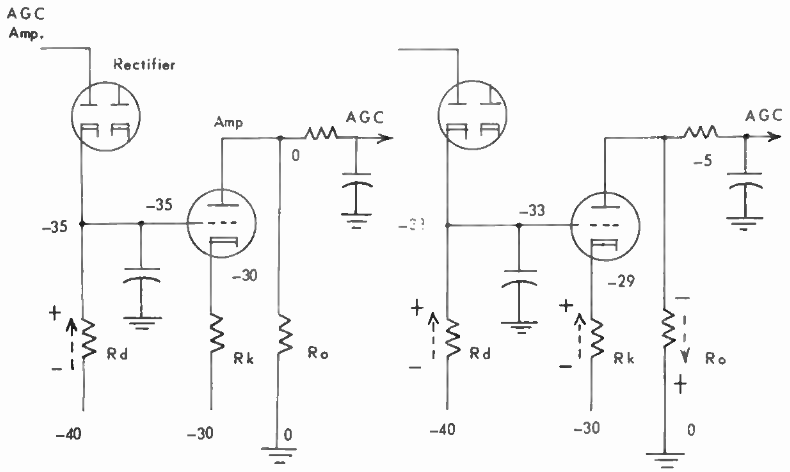


Fig. 49-16. Voltage changes in an agc system which operates below ground.

Fig. 49-15 shows the essential parts of an amplified agc system which utilizes a principle not heretofore examined. The i-f signal from the final i-f amplifier is applied to the plate of the agc rectifier diode whose d-c output, well filtered to leave smooth direct voltage, is applied to the grid of the agc amplifier. This direct voltage is positive with reference to B-minus at the rectifier output and at the amplifier grid, and it becomes increasingly positive with increase of signal strength from the i-f amplifier. The cathode of the agc amplifier is connected through the adjustable control resistor to B-minus. Negative cathode-bias voltage developed in the portion of this control resistor which is below the slider more than overcomes the positive voltage from the rectifier and makes the grid negative with reference to the cathode.

An increase of strength in the signal from the i-f amplifier, which is rectified by the agc diode, tends to makes the grid of the agc amplifier still more positive, and really does make the grid less negative. Between grid voltage changes and plate voltage changes at the agc amplifier there is inversion of polarity, such as occurs in every amplifier. Then a less negative voltage at the amplifier grid means a less positive voltage at the plate of this tube. This is equivalent to a plate voltage which is more negative. This plate voltage, which effectively goes more negative on stronger signals from the i-f amplifier, could be used as an increasingly negative agc voltage except for the fact that plate voltage in a circuit such as shown here actually would be positive with reference to B-minus.

To utilize the principle of a plate voltage that goes more negative on stronger signals the entire agc amplifier circuit must be operated at voltages which are negative with reference to ground. Agc systems in which all voltages on tube elements are negative with respect to ground may be said to operate "below ground", while systems in which tube elements are positive with reference to ground are said to operate "above ground".

When an agc system is operated below ground the action may be explained in an elementary way with the help of Fig. 49-16. The left-hand diagram represents conditions with which the received signal is weak. Then we wish to have no agc action or to have zero agc voltage. I-f signal voltage is being rectified by the diode at the upper left. The cathode of this rectifier is directly connected to the grid of the amplifier, and is connected through resistor Rd at the same time to a point in the power supply system which is 40 volts negative with reference to ground. Ground is considered to be zero voltage in the power supply system. The rectified diode current is flowing toward the cathode of the diode, upward in resistor Rd, and is making the top of this resistor and the amplifier grid 5 volts more positive or 5 volts less negative than –40 volts at the bottom. Thus we have –35 volts at the amplifier grid.

The amplifier cathode is connected through resistor Rk to a point in the power supply that is 30 volts negative with reference to ground. There is no current or electron flow in resistor Rk because the amplifier grid is so much more negative than 30 volts as to be in a condition of plate current cutoff. That is, the amplifier now is biased beyond cutoff. With no current in Rk there is no potential difference across this resistor, and the amplifier cathode voltage is 30 volts negative.

The amplifier plate is connected through resistor Ro to ground, which is of zero voltage. Since plate current is cut off, there is no current and no voltage drop across resistor Ro, and the amplifier plate is at zero voltage. The amplifier plate is connected to the regular agc filter system and the agc bus, so we have zero agc voltage.

Supposing now that i-f amplifier output voltage commences to increase, with a stronger received signal. There will be more rectified current through resistor Rd, more voltage drop across this resistor, and voltage at the amplifier grid will become more positive or, actually, less negative. However, the amplifier will not begin to conduct until its grid voltage rises to a value above the point of plate current cutoff. This means that we have a delayed agc system; there will be no negative agc voltage until the received signal and i-f output increases from minimum to a value that releases the agc amplifier from the condition of cutoff.

We next assume that i-f amplifier output has become strong enough to bring about the conditions shown by the diagram at the right in Fig. 49-16. There is more current and more voltage drop in resistor Rd, which brings amplifier grid voltage to –33. The amplifier plate is more positive, or less negative, than its cathode – because the plate is at zero volts and the cathode is 30 volts negative. This plate voltage causes electron flow or current in the amplifier. This current flows upward in resistor Rk, there is a 1-volt drop across Rk, and the cathode voltage becomes –29. With amplifier grid voltage at –33 and cathode voltage at –29 there is a 4-volt negative grid bias instead of the former 5-volt negative bias. Amplifier plate current flows downward through resistor Ro. The accompanying voltage drop in Ro makes the top of this resistor more negative than the bottom, which is at zero volts. Then the top of Rk and the agc bus becomes 5 volts negative. Thus the agc voltage changes from zero to 5 volts negative when there is a stronger received signal.

The voltages used for illustration of principles in Fig. 49-16 are merely typical values. Obviously, the amplifier tube has to be of such characteristics, the resistors of such values, and the applied negative voltages of such values as will produce desired results. This is a matter of design, and is not too difficult.

AGC LIMITERS OR CLIPPERS. Earlier it was mentioned that noise pulses or interference pulses can upset the action of an agc system. Such voltage pulses may be stronger than the sync pulses. Then the value of agc negative voltage becomes proportional to the noise rather than to the strength of the received signal. It is quite possible for noise pulses to make the agc bias so negative, and gain of r-f and i-f

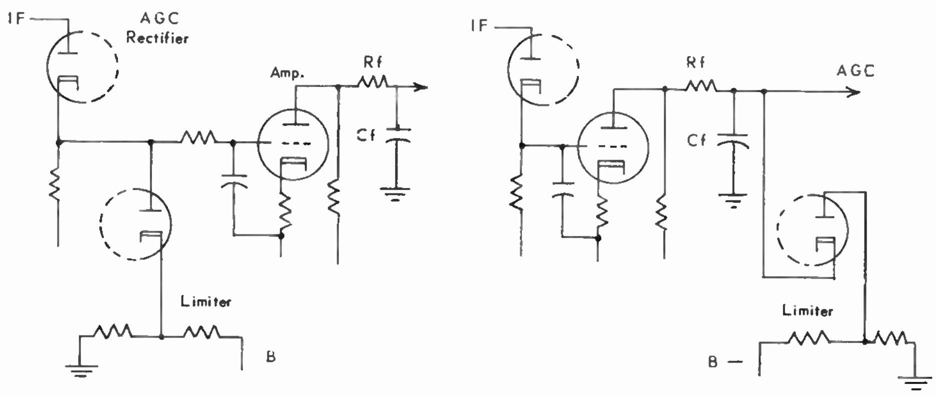


Fig. 49-17. How noise limiters or clippers may be connected to agc circuits.

stages so low, that the video signal disappears from the output. This will be true when there is a low ratio of signal to noise in the first place. There are numerous methods for reducing or preventing the bad effects of excessive noise.

At the left in Fig. 49-17 is an amplified agc system similar to those which have been explained, except for the addition of a limiter diode. The plate of this diode is connected to the cathode of the agc rectifier, at which sync pulses are positive and at which noise pulses also would be positive. The limiter cathode is connected to a point in the B-power system which is positive with reference to ground. This diode cannot conduct unless its plate is made enough more positive than the cathode voltage that the plate becomes positive with reference to the cathode.

The limiting B-voltage on the limiter cathode is made equal to the maximum peak voltages of sync pulses which are to be amplified and used for agc purposes. Then the limiter diode does not conduct with normal video signals. If noise pulse voltages are stronger than normal sync pulse voltages the limiter plate is made more positive than its cathode. Then the limiter conducts and acts like a short circuit to ground (through the B+ lines) for the excess voltage, and this voltage does not affect the agc output from the amplifier plate.

At the right in Fig. 49-17 a limiter diode is connected to the plate circuit of the agc amplifier and to the agc bus. The limiter cathode is connected to the bus, where there is a negative agc voltage. The limiter plate is connected to a point in the B-power system at which voltage is more negative than ground. This negative plate voltage on the limiter is made equal to the desired maximum negative agc voltage, and under this condition the limiter does not conduct. If agc voltage is made still more negative by noise pulse voltages coming through the amplifier, the limiter cathode becomes more negative than its plate. With the plate relatively positive the limiter diode conducts and shorts the excess noise voltage to ground through the B-power lines. Either or both these limiter methods may be used in an agc system. The limiter often is called a noise clipper.

Other and more elaborate methods which prevent noise pulses from affecting the agc system and also the sync section of the receiver will be examined in the next lesson. These other methods are called keyed or gated automatic gain controls.

TELEVISION

LESSON NO. 50

RELATIONS BETWEEN BRIGHTNESS AND CONTRAST

"Training Your Mind"
C. T. I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

COMMERCIAL TRADES INSTITUTE

Training Your Mind — C.T.I. — Perfecting Your Skill

Chicago, Ill.

LESSON NO. 50

RELATIONS BETWEEN BRIGHTNESS AND CONTRAST

At the end of the preceding lesson we had an automatic gain control in which there are the desirable features of delay on weak signals, of amplification for more uniform signal output, and of limiting or clipping any existing noise pulses so that they would not cut off the desired signal.

Our agc system, as so far developed, operates with a biasing or control voltage proportional to amplitude of sync pulse peaks in the received signal. This insures a change of agc bias only when there is a change of signal strength, not when there is a change of picture tone or brightness. Charging of the agc capacitor results from voltage at the peaks of horizontal sync pulses and also from peaks of the much longer vertical sync pulses. The vertical sync pulses occur 60 times per second.

To prevent the agc biasing voltage from varying at the rate of 60 times per second, due to the relatively long rectified currents for each vertical sync pulse, the time constant of the agc capacitor and associated resistors must be long enough to hold the charge practically constant for at least 1/60 second. If, with this long time constant, a noise pulse lasts for less than 1/60 second the agc voltage cannot change fast enough to reduce amplification during the noise pulse. Then the noise voltage can pass on through the controlled amplifiers.

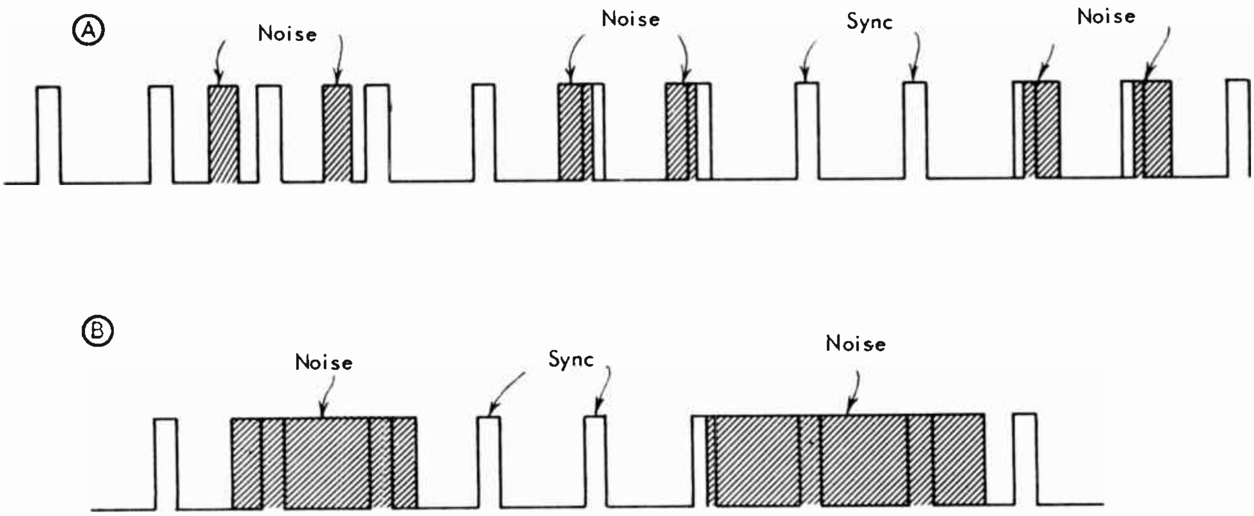


Fig. 50-1. Noise pulses may upset synchronization, even when no stronger than sync pulses.

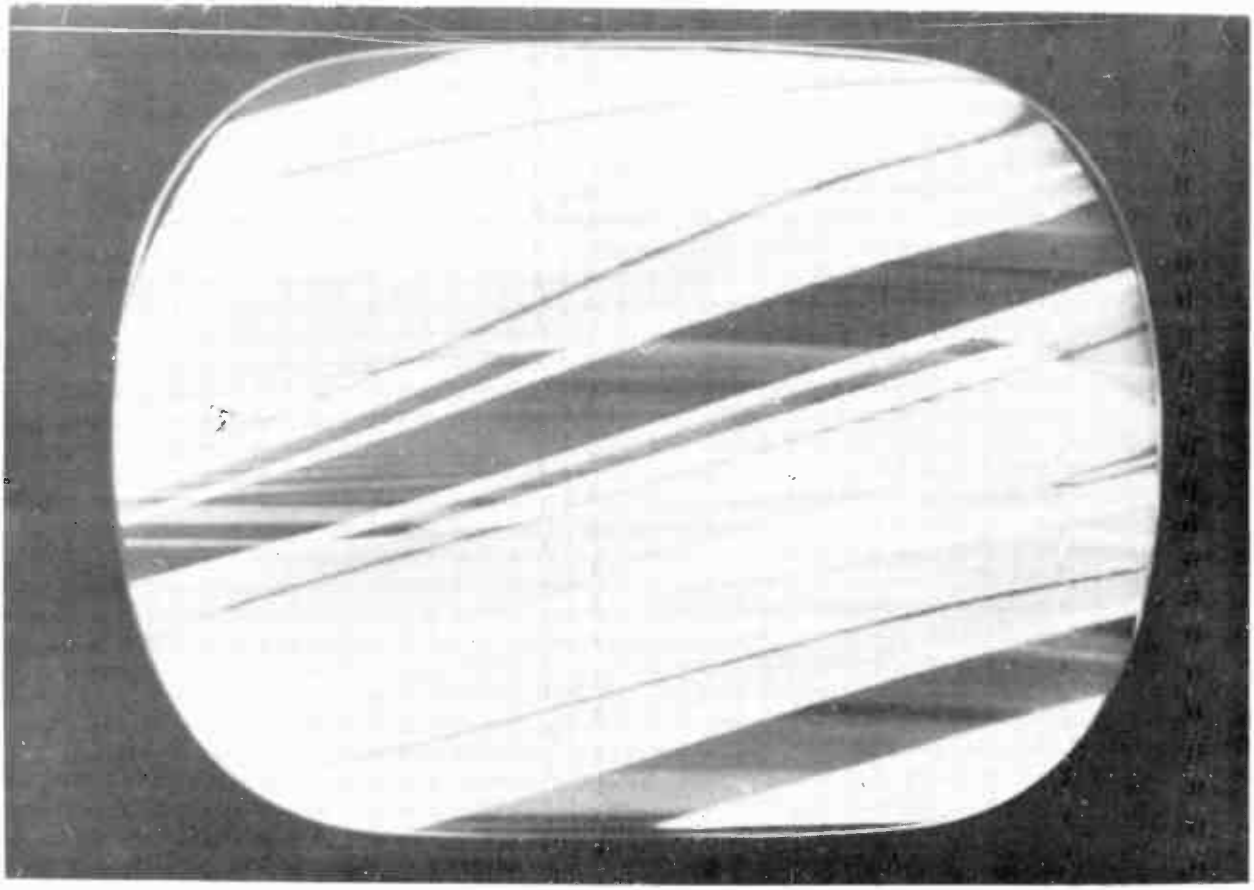


Fig. 50-2. Appearance of a pattern when horizontal synchronization is lost.

It would be desirable to use a much shorter time constant, which would allow the agc system to act on very short noise pulses. This is done in what are called keyed or gated agc systems. Before describing these systems we may discuss one other difficulty that may arise with our original methods.

This other difficulty is one that may upset synchronization of the picture, in this manner. Assume that the agc system contains one or more limiters or clippers that permit noise pulse voltages to rise no higher than sync pulse voltages. In spite of this, a short-time noise pulse can occur in between two regular sync pulses or on the leading or trailing edges of sync pulses as at A in Fig. 50-1. Longer noise pulses could obliterate several regular sync pulses as at B. The timing action of the deformed or obliterated sync pulses would be altered or destroyed.

If horizontal synchronization is upset in this way the reproduced pattern or picture may take on a momentary appearance such as pictured by Fig. 50-2. When vertical synchronization is affected the pattern or picture may roll, which means rapid movement either up or down on the screen.

The electrical action of a keyed agc system is to make the agc rectifier conductive only during instants

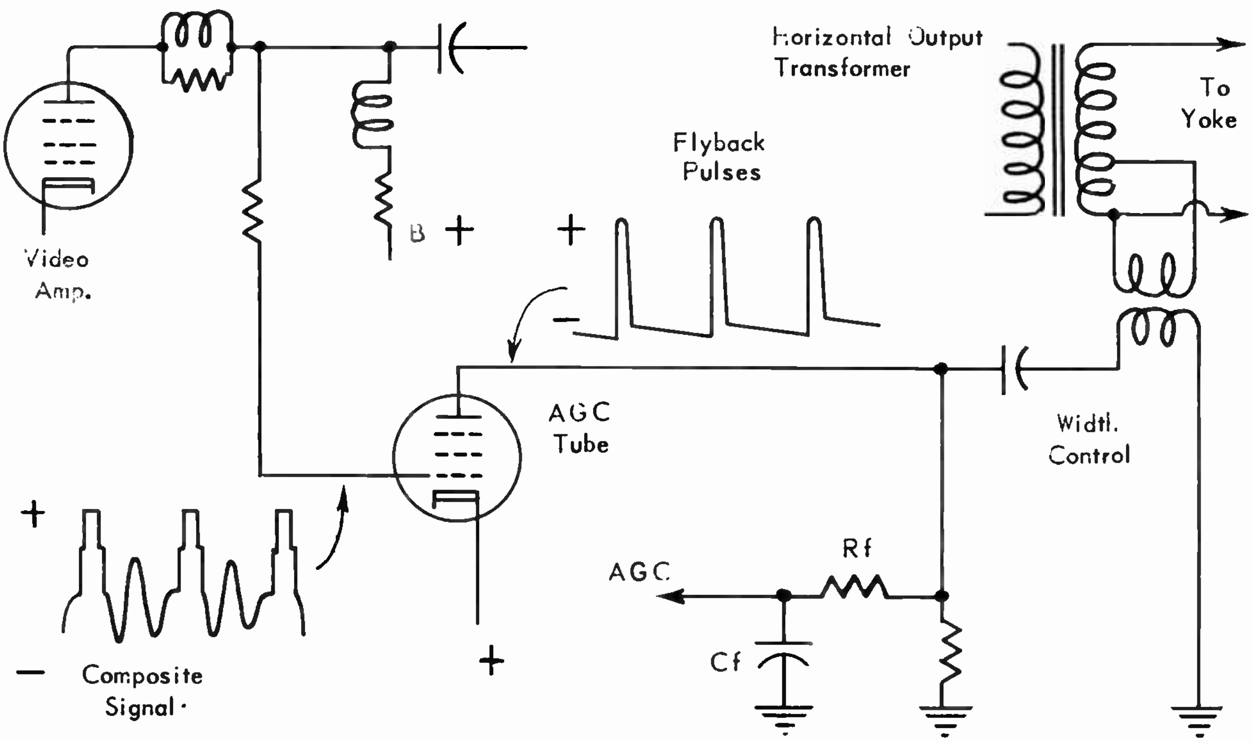


Fig. 50-3. Circuits and voltage pulses for one kind of keyed agc system.

in which occur the regular horizontal sync pulses, and to keep this tube non-conductive during all other periods in which noise pulses may occur. Then the agc voltage for controlled tubes can be developed only according to peak voltage of the regular horizontal sync pulses. The time constant of the agc capacitor and resistors may be made proportional to the frequency of horizontal sync pulses, which is 15,750 cycles per second. The short time constant allows agc action to follow rapid changes of signal strength, and compensate for these changes with corresponding variations of agc biasing voltage.

Circuits for one type of keyed agc system are shown by Fig. 50-3. To the grid of the agc tube is applied the composite signal voltage as taken from the output of a video amplifier tube. This signal voltage is taken from a point at which the sync pulses are positive. The plate of the agc tube is connected through a coupling capacitor to a small coil inductively coupled to the width control inductor which is connected to the secondary of the horizontal output transformer. In the secondary circuit is the picture tube deflection yoke. In this circuit appear strongly positive "flyback pulses" of voltage. These flyback pulses are in time with horizontal sync pulses, because horizontal deflection action is controlled by these sync pulses. The flyback pulses are the only source of plate voltage for the agc tube.

The cathode of the agc tube is connected to a B+ voltage of about 135 to 140 volts. The grid of this tube is connected to the plate voltage of the video amplifier which is somewhat lower than the B+ voltage on the plate circuit, due to drop in the plate load. Thus the grid of the agc tube is negative with reference

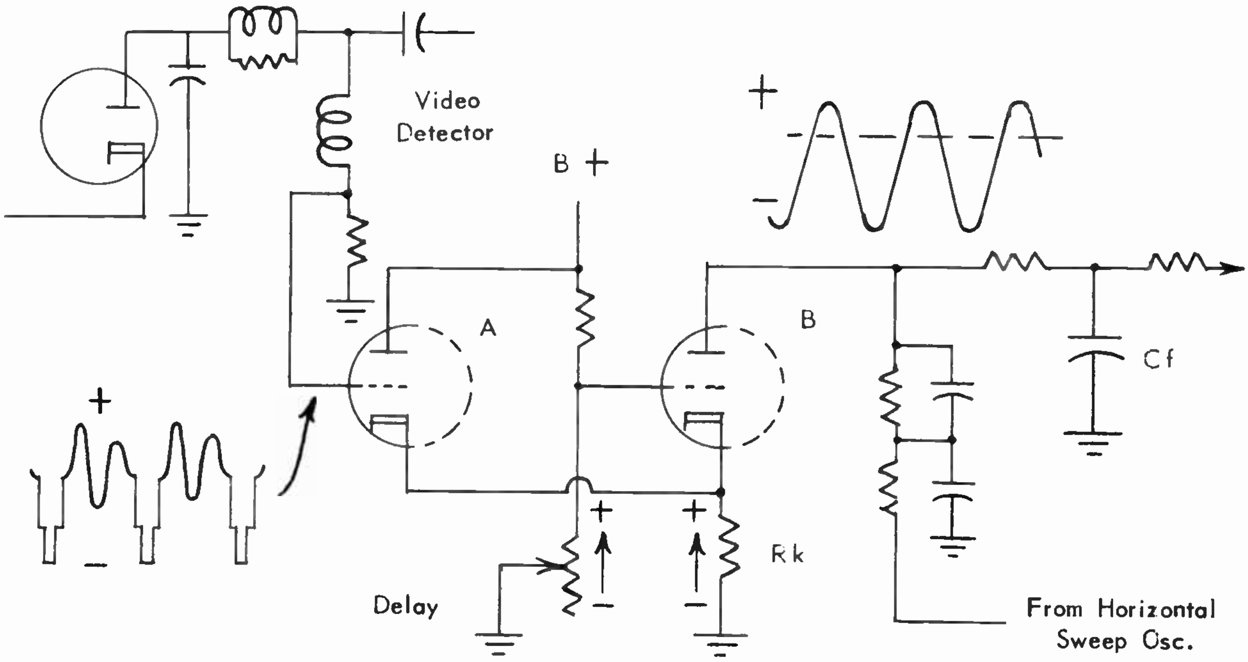


Fig. 50-4. Circuits and voltages in a gated agc system.

to its cathode. The plate of the agc tube is connected through one or more resistors to ground or B-minus, and is at zero voltage except during the application of positive flyback pulses.

When positive flyback pulses at the plate occur during the same instant as positive sync pulses at the grid the tube conducts and charges agc capacitor Cf in the usual way. There is conduction because the positive sync pulsesovercome part or all of the normal negative bias on the grid, while flyback pulses furnish the necessary positive plate voltage. If positive voltage pulses due to noise reach the grid of the agc tube in between the times of regular sync pulses the plate of this tube will be negative, since there will be no flyback pulse, and the tube will not conduct to alter the agc voltage or charge on capacitor Cf.

Fig. 50-4 is a simplified circuit diagram of a gated agc system which operates with two triodes, A and B, or with a twin triode. The plate of triode A is connected to a B+ line. To the grid of this triode is applied a composite signal voltage taken from the top of the load resistor for the video detector. Sync pulses are negative in this composite signal, so the stronger the signal the more negative becomes the grid of triode A and the less is its plate current.

Plate current of triode A flows in resistor Rk, which is connected in a way that makes it a bias resistor on triode B, although the current through this resistor and the biasing voltage across it are chiefly dependent on plate current of triode A. To the plate of triode B is applied a sine wave voltage obtained from the circuit of the horizontal sweep oscillator. The frequency of this voltage is the same as that of horizontal sync pulses on the grid of triode A, it is 15,750 cycles per second. It is only during tips of positive alternations of this plate voltage that triode B can conduct and charge agc capacitor Cf.

A stronger signal from the i-f amplifier and video detector makes the grid of triode A more negative, reducing the plate current and reducing the current and voltage at resistor Rk. This reduced voltage across Rk means a less negative bias for triode B. When sync pulses and positive alternations of plate current on triode B occur together this triode conducts, and maintains or increases the agc voltage. If noise pulses reach the grid of triode A at times in between the positive alternations of plate current on triode B this second triode is non-conductive, because its plate is negative, and the noise pulses can cause no charging of the agc capacitor.

BRIGHTNESS AND CONTRAST. Contrast and gain controls of a television receiver control the amplification in video amplifiers, or video and i-f amplifiers, to regulate the strength of the signal reaching the picture tube. The grid of the picture tube must be negatively biased to a degree that will accomodate the strength of this signal. The amount of bias voltage, in relation to signal strength, determines whether or not the reproduced picture will contain a full range of shading and will have average or overall brightness equivalent to the original televised scene. If signal strength is changed by adjusting the contrast control, the picture tube grid bias must be changed to suit. The reasons will become evident in following paragraphs.

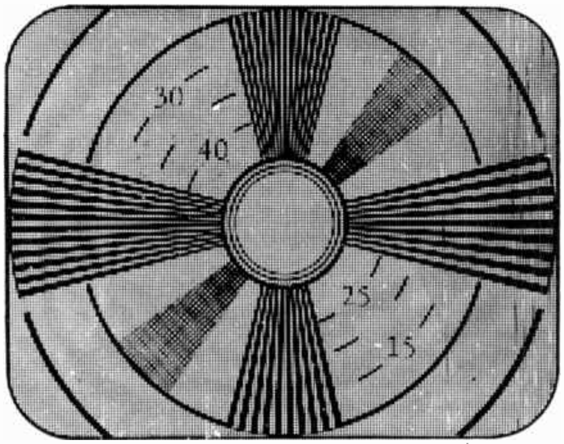
1 The adjustable control that varies the average grid bias on the picture tube most often is called the brightness control, but sometimes is spoken of as the brilliancy control, intensity control, or background control. The bias voltage on the grid of a picture tube controls average strength of current in the electron beam, just as bias on an amplifier tube controls average strength of plate current. When picture tube bias is made less negative it allows more beam current, and greater all over brightness in the picture. If the bias is made more negative it reduces beam current, and makes the picture darker or less bright.

When you "increase" the brightness control of a television receiver, or turn the brightness "up", you are making the bias on the picture tube grid less negative. When the setting of the brightness control is reduced or turned down, the picture tube grid bias is being made more negative.

2 If the brightness control is turned too far down, making picture tube grid bias too negative for the strength of signal being applied, the reproduced pattern or picture becomes dark all over, about as shown at the left in Fig. 50-5. If the brightness control is turned too high, making picture tube grid bias insufficiently negative for strength of applied signal, the effect on patterns or pictures will be somewhat as shown at the right. The picture will be light gray all over, and usually there will be white lines sloping upward toward the right. These are called vertican retrace lines. Such lines may appear also when the contrast control is turned too low for the existing setting of the brightness control.

Relations between settings of contrast and brightness controls are best illustrated as in Fig. 50-6. Signal voltages are applied to the transfer characteristic curve of a picture tube, and resulting changes of beam current are shown as changes of brightness. It is helpful to keep in mind that signal strength is determined by the contrast control, and that the amount of negative grid bias is determined by the brightness control.

Diagram A represents correct relations between signal strength and grid bias. The applied signal is an alternating voltage whose average or zero value is at the bias voltage and whose alternations go positive and negative with reference to the bias voltage - exactly as do variations of signal voltage applied to the



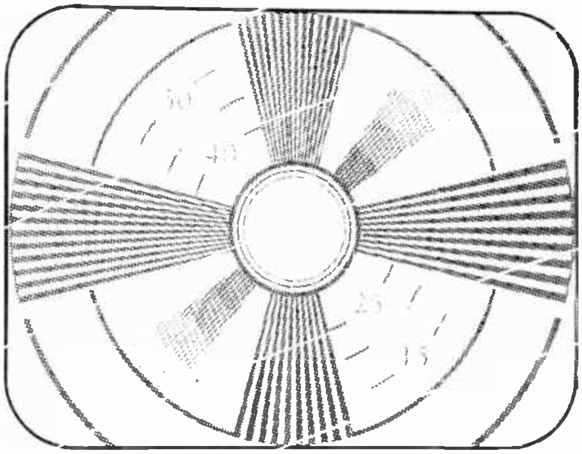


Fig. 50-5. Brightness control too low, at left, and too high, at right.

grid of a negatively biased amplifier tube. The grid bias for correct reproduction is of a value that brings the black level of the signal to the grid voltage that causes cutoff of the electron beam. When the beam is cut off, the picture tube screen is without illumination, or is black.

All signal voltages which are more negative than the voltage for cutoff leave the screen black, since there can be no beam current at any voltage below cutoff. These parts of the signal in diagram A include the sync pulses. Consequently, the sync pulses are cut off and have no effect whatever on the reproduced picture.

All signal voltages which are less negative than the voltage for cutoff, and all signal voltages which are positive, cause beam current to flow and cause a degree of illumination or brightness on the screen that corresponds to signal voltage. These parts of the signal in diagram A include all the picture variations. This means that all changes of tone or shading represented by the picture signal will appear on the screen of the picture tube.

Unless grid bias on the picture tube is such as brings the black level of the applied signal voltage to the grid voltage for beam cutoff there cannot be correct reproduction of dark and light tones in the reproduced picture.

Now let's go to diagram B of Fig. 50-6. Picture tube grid bias has been made less negative than before, but we still have the original strength of applied signal voltage. Less grid bias means that the brightness control has been turned up or increased. This should not have been done when there was no change of signal strength. The result, as you can see at the right of the characteristic curve, is to bring the reproduced pattern or picture too high in brightness. Average electron beam current is too great. Voltages of the applied signal have been moved too high on the characteristic curve.

Two bad effects show up in diagram B. First, the black level in the picture no longer is at the point of beam cutoff. Parts of the image that should be black now are raised to a dark gray. All other parts of the image above the black level are too high in brightness, and all will appear lighter than in the original

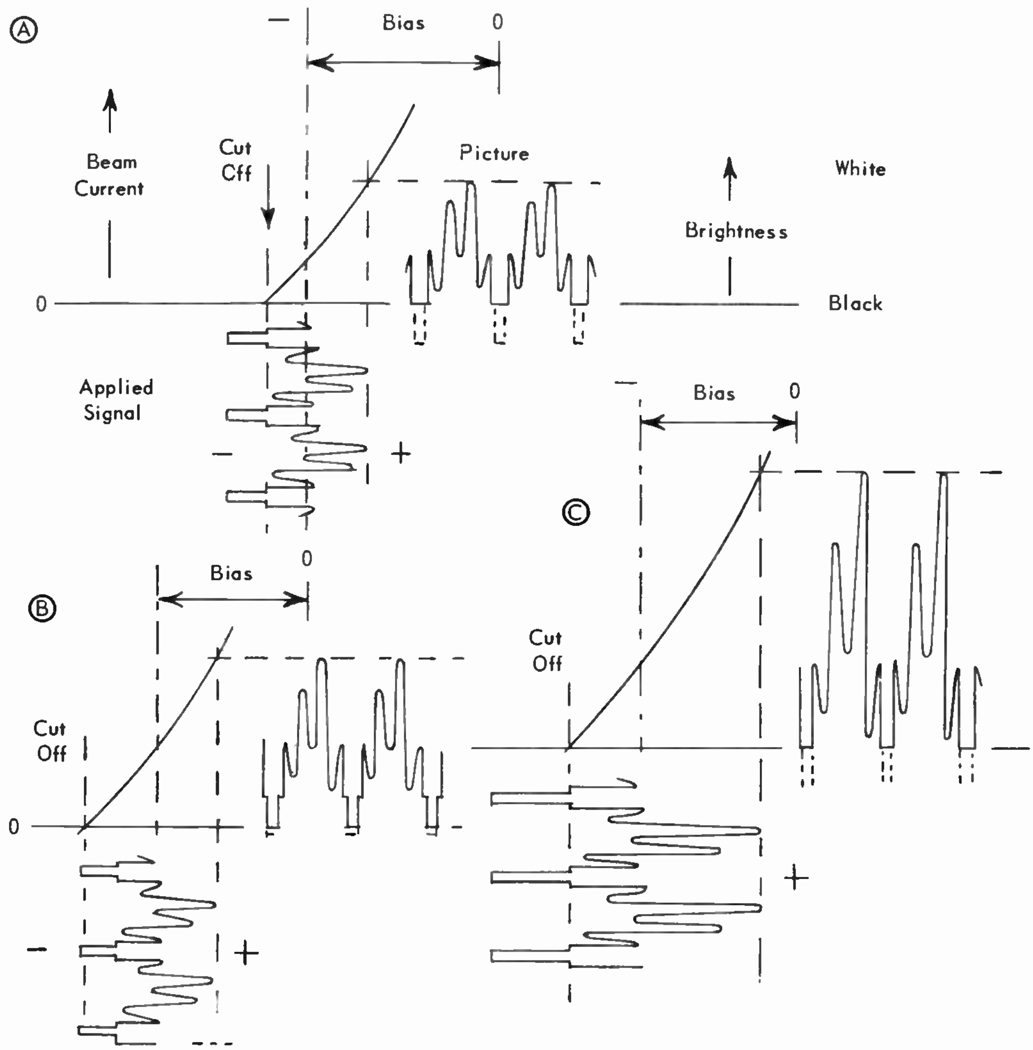


Fig. 50-6. What happens when brightness is turned too high, and how it is corrected.

scene. The second bad effect is to bring parts of the sync pulses, or their effects, into the picture. We are using sync pulse voltage as part of the picture signal, whereas this voltage should be cut off.

Next look at diagram C of Fig. 50-6. Picture tube grid bias is the same as in diagram B, meaning that setting of the brightness control is the same in both cases. But now the contrast control has been turned up to strengthen the voltage of the applied signal. The average voltage of the applied signal remains at the grid bias voltage, as always. Increasing the signal amplitude moves the positive peaks of signal voltage still more positive, and moves the negative peaks (sync pulse tips) still more negative. This expansion of signal voltage on the negative side has brought the black level of the signal down to the picture tube grid voltage for beam cutoff. This means that we have correct reproduction of the pattern or picture, with sync pulses cut off and all picture variations of the signal causing proportionate changes of beam current and brightness.

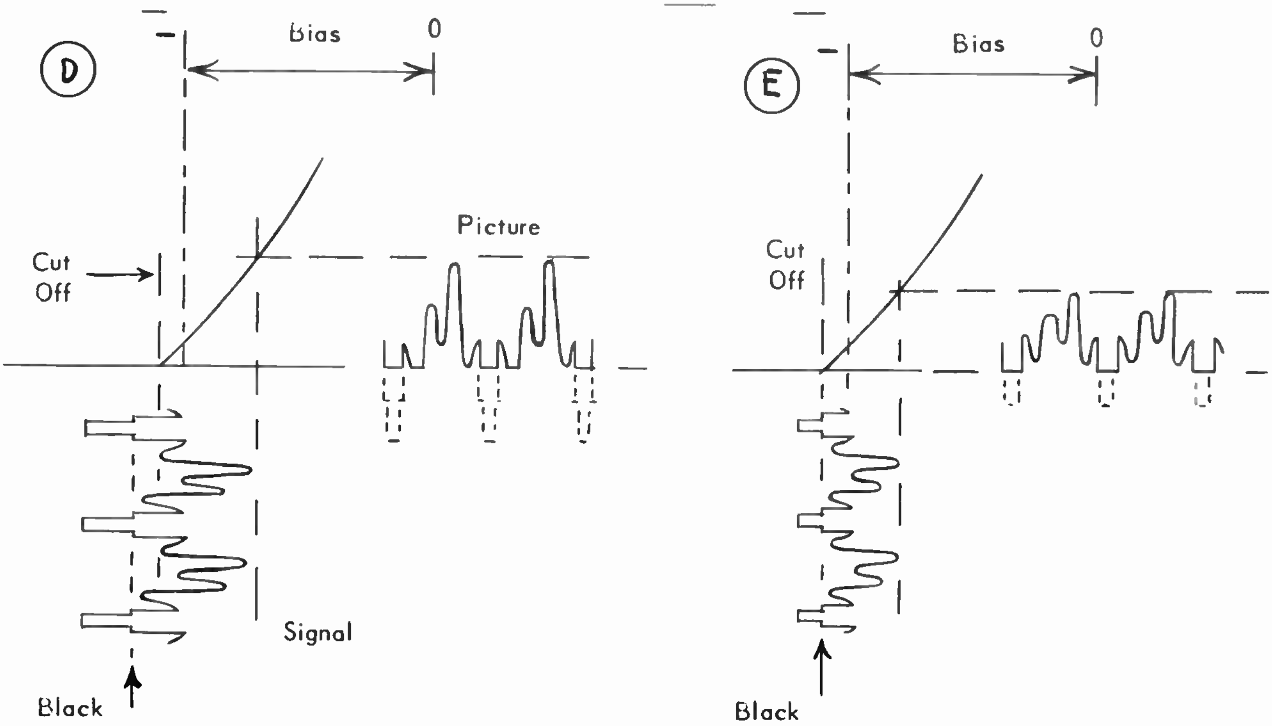


Fig. 50-7. Brightness turned too low, and compensating adjustment of contrast control.

What might have happened during the sequence shown by Fig. 50-6 is this. At A the operator has a picture with all shadings correctly related to one another, and with correct contrast, but he feels that the whole picture is too dark. At B he has increased the brightness control to lighten the picture, but now there is lack of contrast, with everything appearing in too much of the same gray tone. At C the operator has left the brightness control alone, but has increased the contrast control. He has restored the correct range of shading from darkest to lightest, and also has the brighter picture that he started out to obtain. The picture at C is brighter than the one at A because all the variations have been moved up, and the average comes higher on the brightness scale.

In Fig. 50-7 are illustrations of what happens when brightness is turned too low, when picture tube grid bias is made too negative. Diagrams are marked D and E because they really are a continuation of the series begun in Fig. 50-6. We commence once more with contrast and brightness relations as at A in Fig. 50-6, then decrease the brightness or make picture tube grid bias more negative as at D in Fig. 50-7.

We still have the original strength of applied signal. Making the bias more negative has moved the whole signal voltage down on the characteristic curve, and has brought the black level of the signal well beyond the grid voltage for beam cutoff. Along with the black level have gone some of the darker variations of the signal, and in the reproduction at the right of the characteristic curve these shadings which should have been dark gray are completely cut off and are changed to black. Only those portions of signal voltage that correspond to light tones have any effect on increasing the beam current.

Reducing the brightness or making the bias more negative, with no change of signal voltage, has reduced all the darker grays of the original image to black in the picture. The added blacks make the picture appear too dark. Actually there has been also a loss of contrast, because there is less difference between darkest and lightest remaining tones than between darkest and lightest tones in the original reproduction.

At E in Fig. 50-7 the picture tube grid bias is the same as at D, the brightness control still is turned too low. But the contrast control has been turned down to reduce the signal voltage. The average or zero voltage of the signal stays at the bias voltage, but signal voltage amplitude decreases on both sides of its zero. This contraction of signal voltage has drawn the black level over to the point for beam current cutoff. Then, of course, there is cutoff in the reproduction of only the sync pulses. All the picture variations of the signal voltage cause corresponding changes of beam current and brightness.

In diagram E we have correct contrast, because all changes of shading represented in the signal are proportionately reproduced in changes of brightness. Readjustment of the contrast control has corrected the contrast at the low brightness level caused by turning down the brightness control. But the picture at E doubtless would be too dark. If an operator were to attempt lightening the picture by turning up the brightness control, making grid bias less negative, the result would be like that shown at B in Fig. 50-6. The whole picture would be raised too high on the brightness scale, and would be an all over light gray. Then it would be necessary to increase the contrast control, and applied signal voltage, in order to bring the black level down to the point of beam cutoff.

BRIGHTNESS CONTROLS. The electrical purpose of the adjustable brightness control is to make the grid of the picture tube more or less negative than the cathode in this tube, or to vary the grid bias. The brightness control must be designed so that the grid never may become positive with reference to the cathode, and very seldom is the grid voltage allowed to become zero with reference to the cathode.

Fig. 50-8 illustrates the principles of several brightness control circuits for receivers in which the signal output of the last video amplifier is fed to the picture tube through a capacitor at Cc. In diagrams A and B the video signal is applied to the grid of the picture tube. In diagram C the video signal is applied to the cathode of the picture tube.

In diagram A the picture tube grid is connected to ground through resistor Rg. With the grid negative there is no current and no voltage drop in this resistor, so the grid is maintained at ground potential. The cathode of the picture tube is connected to the slider of the brightness control potentiometer, which is in a line between a B+ voltage and ground. Resistances and B voltage for the control circuit are of such values that the left-hand end of the control potentiometer is at +50 volts and the right-hand end is at +5 volts.

The cathode in diagram A always is positive with reference to ground. Since the grid is at ground potential, the cathode always is positive with reference to the grid. This is equivalent to always having the grid negative with reference to the cathode, or to always having negative grid bias. When the slider of the control potentiometer is moved to the +5-volt position there is a 5-volt negative bias on the grid. With the slider at the +50-volt position there is a 50-volt negative bias on the grid. There is maximum beam current and maximum brightness of the picture when the bias is minimum, with the brightness control slider at the low-voltage end of its travel. At the other end of the slider travel there is maximum negative

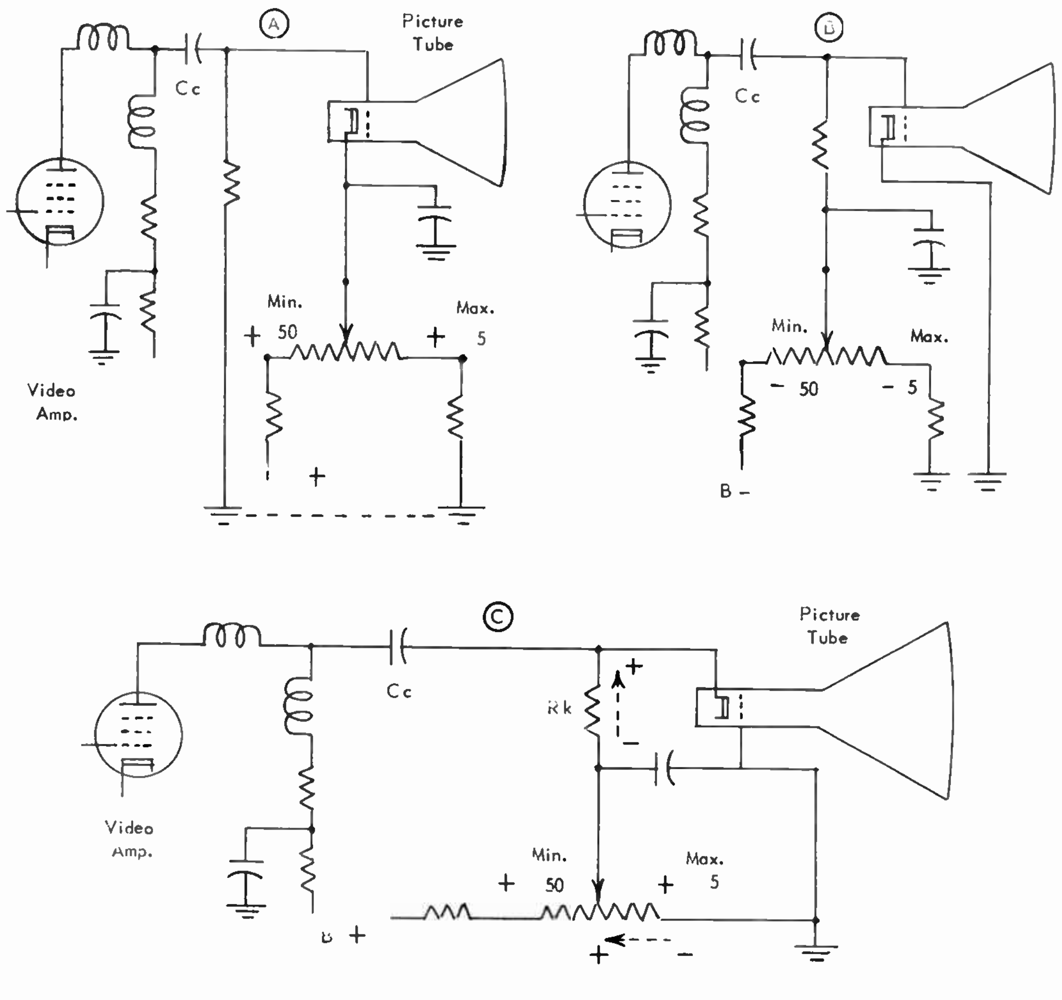


Fig. 50-8. Brightness control circuits where there is capacitor coupling from video amplifier to picture tube.

grid bias, minimum beam current, and minimum brightness of the picture. The abbreviations Min and Max on the diagram, for minimum and maximum, refer to picture brightness or to settings of the brightness control.

In diagram B the cathode of the picture tube is connected to ground, and is at ground potential, whereas in the first diagram the grid is at ground potential. Now it is necessary to maintain the grid negative with reference to ground (cathode) potential by connecting the brightness control potentiometer into a line that is in the negative side of the B power supply, or into a line between B-minus and ground. The grid of the picture tube is connected through resistor Rg to the slider of the potentiometer. Moving the slider varies the negative grid bias between 5 and 50 volts. The 5-volt negative bias causes maximum beam current and brightness, while the 50-volt negative bias causes minimum beam current and brightness.

In diagram C of Fig. 50-8 the signal output of the video amplifier is fed through capacitor Cc to the cathode of the picture tube. The grid of the picture tube is connected to ground. The cathode connects through resistor Rk to the slider of the brightness control potentiometer. This potentiometer is in a B+ line with other resistors of values that cause voltages at the ends of the potentiometer to be +50 and +5. These positive voltages on the cathode are equivalent to negative bias voltages on the grid, and we have the same results as in diagram A so far as control action is concerned.

The cathode of a picture tube must carry the current for all the anodes and also the beam current. Consequently, in circuit C there is electron flow in directions indicated by broken-line arrows through resistor Rk and through whatever portion of the potentiometer resistance is between the slider and ground. Voltage drops due to cathode current in these resistances are of such polarity as to add to the positive voltage on the cathode, with reference to ground and the grid. These voltage drops provide the minimum bias, shown as 5 volts at the right-hand end of the control potentiometer in this diagram. There is some certain minimum cathode-bias in this type of control. The cathode-bias is added to by B voltage acting in the control potentiometer.

Fig. 50-9 illustrates some brightness control circuits used in receivers where the output signal from the video amplifier is fed directly to the grid of the picture tube (diagram A) or directly to the cathode (diagram B) without having any coupling and blocking capacitor between video amplifier and picture tube. With systems such as these the grid or cathode of the picture tube is at the same voltage as the plate of the video amplifier, a voltage which must be sufficiently positive with reference to ground to operate the video amplifier tube.

The grid return circuit or the conductive circuit between grid and cathode of the picture tube includes resistors Ro and Rd in the plate circuit of the video amplifier, also the portion of the brightness control potentiometer resistance to the left of the slider, and any other resistors which may be in the brightness control circuit for dropping of voltage. In this grid return circuit must be potential differences that maintain the grid negative with reference to the cathode, or that maintain a negative bias on the grid.

In diagram A the amplifier plate circuit and brightness control circuit are fed from a point which is 200 volts positive in the B supply system. In resistors Ro and Rd there is a drop of 50 volts, leaving the amplifier plate and picture tube grid 150 volts positive. The left-hand end of the control potentiometer is connected to the 200-volt B+ voltage, so is at a potential of 200 volts with reference to ground. There is a drop of 50 volts or less in the control resistance, leaving the right-hand end with a potential of 150 volts or slightly more. The remaining voltage drop to ground occurs in resistor Ra.

Moving the slider of the control potentiometer all the way to the right brings the slider and the connected picture tube cathode to a positive potential of 150 volts or slightly more. This is either the same potential as on the picture tube grid, or a slightly greater potential. Then, with reference to the cathode, the grid bias is zero or slightly negative and there is maximum beam current and also maximum brightness. Moving the control slider all the way to the left places the slider and picture tube at a positive potential of 200 volts, while the grid potential is 150 volts positive. Then the grid is effectively 50 volts more negative than the cathode of the picture tube. This maximum negative bias reduces beam current and brightness to minimum.

Having followed the action of diagram A in some detail it should be easy to see what must happen in diagram B of Fig. 50-9, where signal input is to the picture tube cathode. The video amplifier plate and

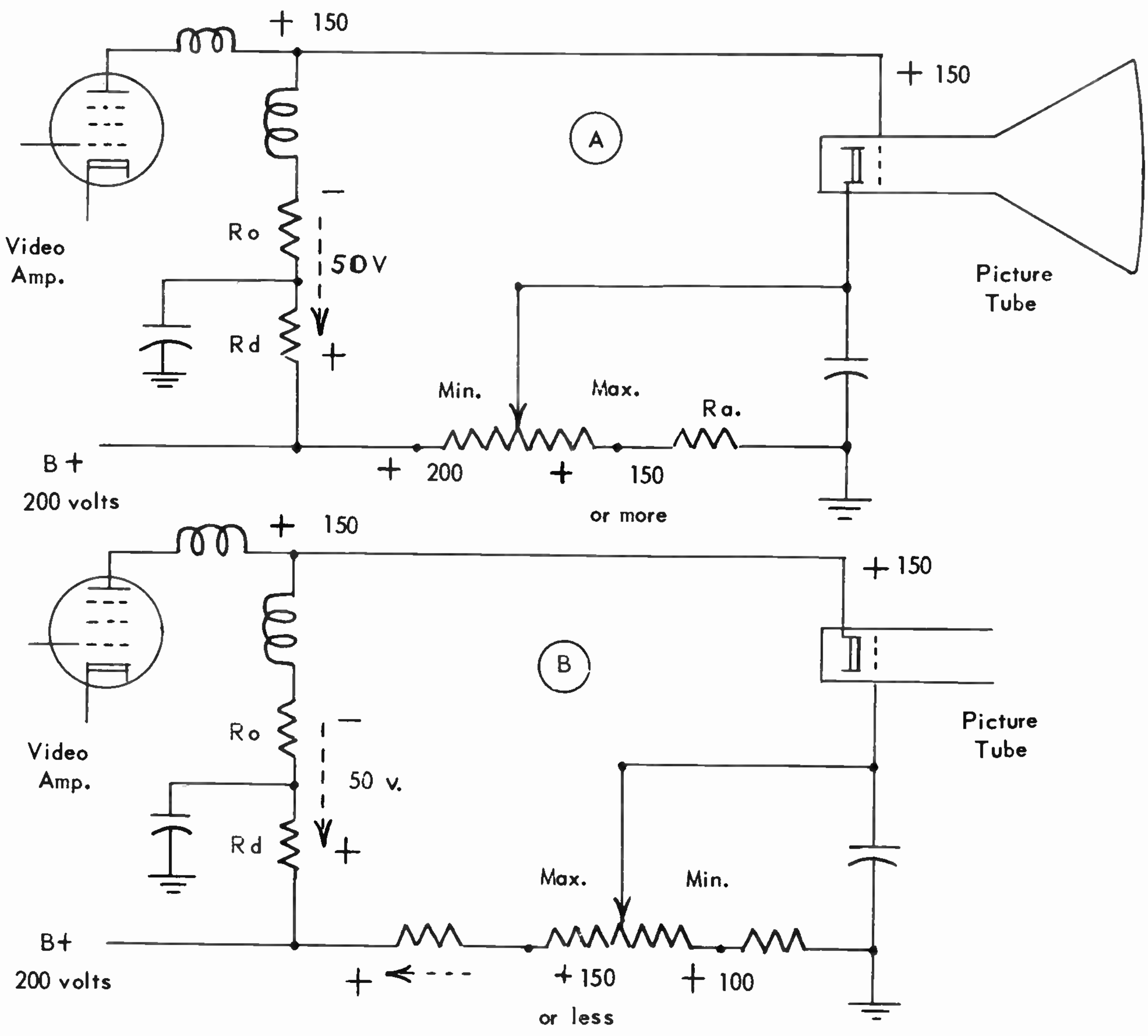


Fig. 50-9 Brightness control circuits where there is direct conductive coupling from video amplifier to picture tube.

the picture tube cathode are connected through resistors Ro and Rd to B + 200 volts. There is a drop of 50 volts in the two resistors, leaving amplifier plate and picture tube cathodes at 150 volts.

The picture tube grid is connected to the slider of the control potentiometer. The grid always must be less positive, effectively more negative, than the cathode. Then no part of the control potentiometer may be more positive than the 150 volts on the picture tube grid. Also, the positive voltage at one end of the potentiometer must be less positive by whatever amount is desired for maximum negative grid bias. Resistances and voltage drops in the control line are such as to place the left-hand end of the potentiometer at 150 volts positive or slightly less (to insure a minimum grid bias) and to place the right-hand end at 100 volts positive. Then we may have negative grid bias, variable from 50 volts negative (150 minus 100 volts) to zero or a few volts negative.

Values of voltages shown in Figs. 50-8 and 50-9 have been chosen solely for purposes of explanation, to have some definite figures with which to work. While these voltages are fairly typical of those actually used, they might be varied widely without any change in operating principles of the several circuits.

D-C RESTORATION

We have seen what happens to the picture when the contrast control is changed without making a corresponding change of brightness control, or have seen what happens when signal voltage is altered without changing the picture tube grid bias to suit the new value of signal voltage. It is easy to change the brightness control to suit any setting of the contrast control, but there is something else that alters signal strength too rapidly for anyone to follow with adjustments of a manually operated brightness control.

This other thing is rapid variation of illumination at the original scene being televised. The sun may shine brightly for a minute and then be obscured. Someone may turn a lamp on, or off. The scene may change from outdoors to indoors. There is continual and unpredictable variation of average illumination or picture tone.

What happens to the video signal when average illumination changes from dark to light and back again is illustrated by Fig. 50-10. At A is an i-f signal in which picture voltage variations or picture modulation stays rather close to the black level. This is an i-f signal for a dark toned picture.

When the i-f signal is rectified by the video detector it comes from the detector as shown at B. This is a direct voltage, it is a pulsating direct voltage in which pulsations are formed by the video signal. The maximum value of direct voltage is at the tips of the sync pulses, and the minimum value is at points where the picture is nearest to white. The average value of this direct voltage is about midway between, as marked on the diagram.

(3) When the d-c video signal, such as exists at the detector output, passes through a coupling and blocking capacitor on its way to the picture tube the signal becomes an alternating current or voltage, as at C. There is no remaining direct voltage. The average value or zero value of the a-c signal is about midway between the positive sync pulse tips and the negative amplitudes corresponding to lightest parts of the picture.

It is common practice to speak of the signal at B as having a d-c component. It is thought of as an a-c signal of which part of the voltage is direct. Then the signal shown at C has lost its d-c component. The d-c component always is lost when a signal at the detector output passes through any capacitor.

At D in Fig. 50-10 is represented an i-f signal for a picture of generally light tone. Many of the picture voltages come almost to the white level. When this signal is rectified by the video detector it comes out as shown at E. Again the signal is a direct voltage. The average value of this direct voltage is considerably less than in the d-c signal for the dark toned picture, as you can see by comparing diagrams E and B.

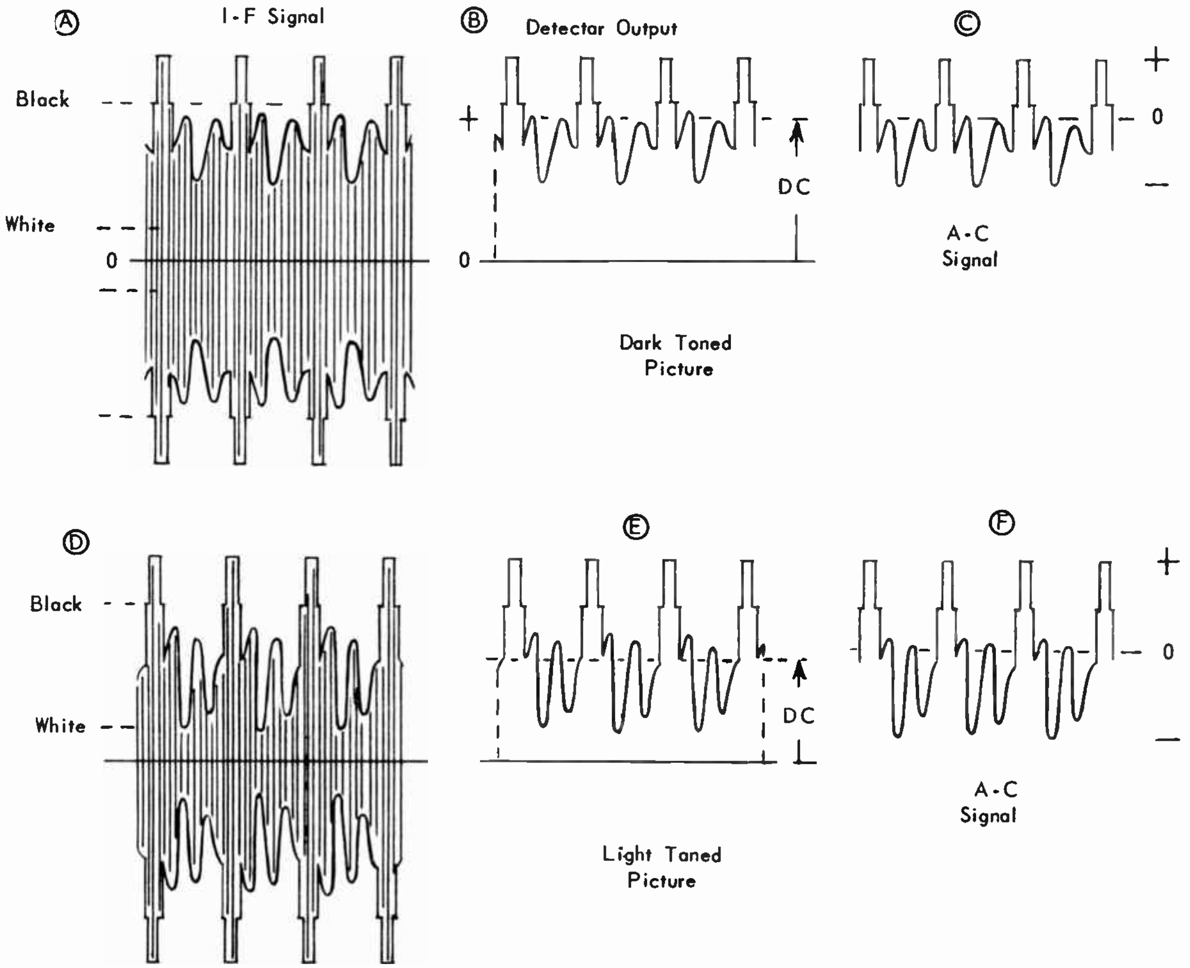


Fig. 50-10. D-c and a-c video signals for dark and light pictures as obtained from a carrier of constant strength.

When the d-c signal for the light toned picture goes through a coupling capacitor anywhere between the detector and the picture tube we have on the far side of the capacitor a signal which is an alternating current or voltage, as at F. The peak-to-peak voltage or the amplitude of this a-c signal is much greater than peak-to-peak voltage of the signal for the dark toned picture. This is apparent upon comparing diagrams F and C.

The signal for the dark toned picture has greater d-c component but less a-c amplitude than the signal for the light toned picture. Remember these differences, especially that signals for light toned pictures have relatively great amplitude.

Let's assume now that grid bias of the picture tube is suited to the signal for the dark toned picture at C in Fig. 50-10. The result of applying this signal to the picture tube grid is shown at the left in Fig. 50-11. Reproduction on the brightness scale is satisfactory. The whole picture is rather dark or is down rather close to zero beam current, as it should be. Contrast is correct, because all variations of light and dark are present in their correct relations to the original signal.

Supposing now the picture suddenly becomes bright, or, at least, that the signal changes to one representing a lighter tone, as at F of Fig. 50-10. This signal for a light toned picture is applied to the picture tube grid at the right in Fig. 50-11. There has been no change of picture tube grid bias. Reproduction is bad. The parts of the signal that still vary the beam current are down too close to zero beam current, they are made too dark. Some of the picture variations have disappeared, they would come beyond beam current cutoff.

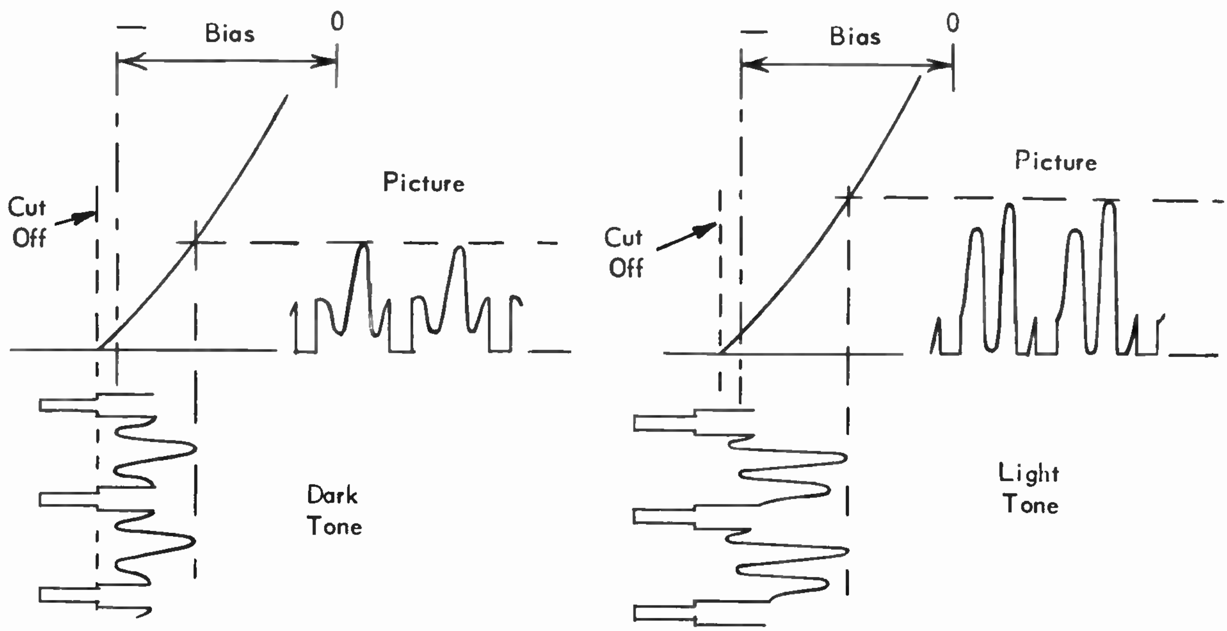


Fig. 50-11. What happens at the picture tube when the picture changes from dark to light.

What we need at the right in Fig. 50-11 is a less negative grid bias, one that will bring the entire signal higher on the characteristic curve. If picture tube grid bias could be made to follow changes of the d-c component, as shown at B and C of Fig. 50-10, everything would be all right. Then we should have a less negative bias for the light toned picture, just as there is a lesser d-c component for this kind of picture. There would be relatively great negative bias for dark toned pictures, just as there is a relatively great d-c component for such pictures.

(6) The process of varying the picture tube grid bias with every change of picture illumination, or average brightness, is called d-c restoration or it may be called d-c reinsertion. This process must be automatic, it must be dependent on the signal itself. The changes of bias never could be made rapidly enough or with sufficient accuracy with any manually operated control. D-c restoration must vary the picture tube grid bias in a manner that will hold the black level of the signal always at the grid voltage for electron beam cutoff. Then the sync pulses will be cut off, but the entire picture will remain on the screen of the picture tube.

The d-c component is lost, and the signal becomes an alternating current or voltage, when passing through a capacitor. Consequently, there must be automatic control of picture tube grid bias or there must be d-c restoration following the last capacitor which precedes the grid-cathode circuit of the picture tube. It is in this grid-cathode circuit that the d-c component must appear.

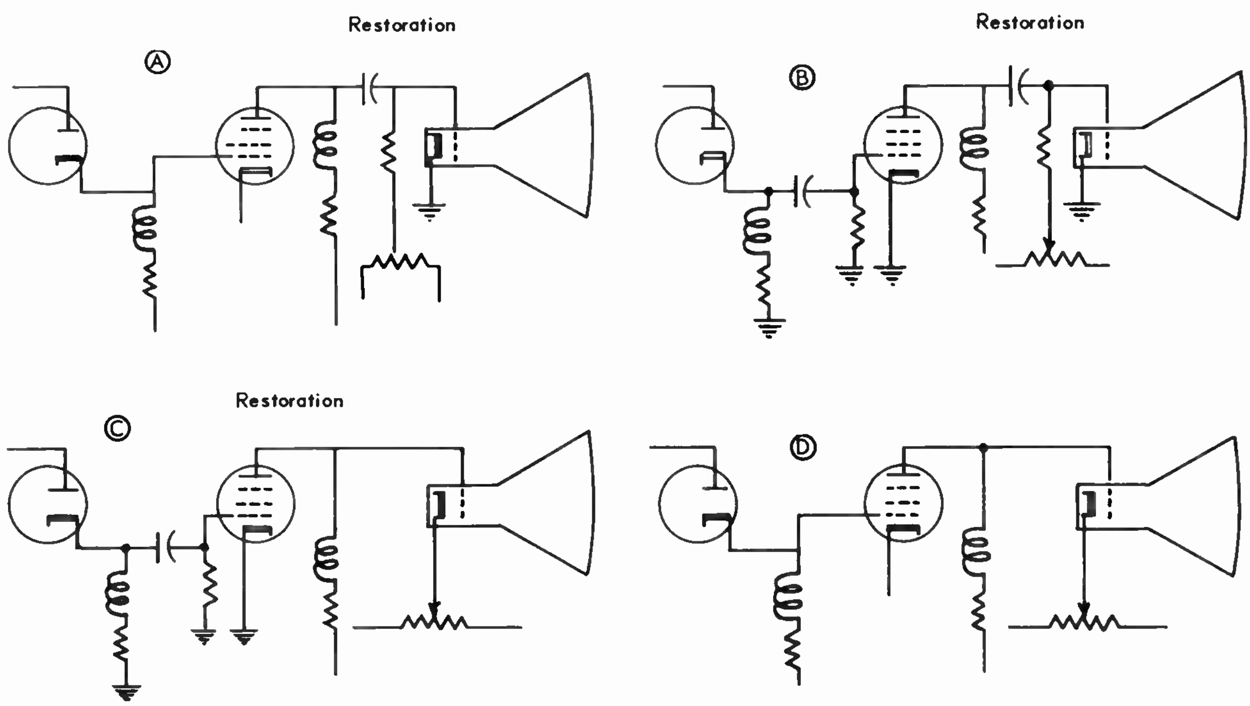


Fig. 50-12. Points at which d-c restoration may be applied when the d-c component is lost in a coupling capacitor.

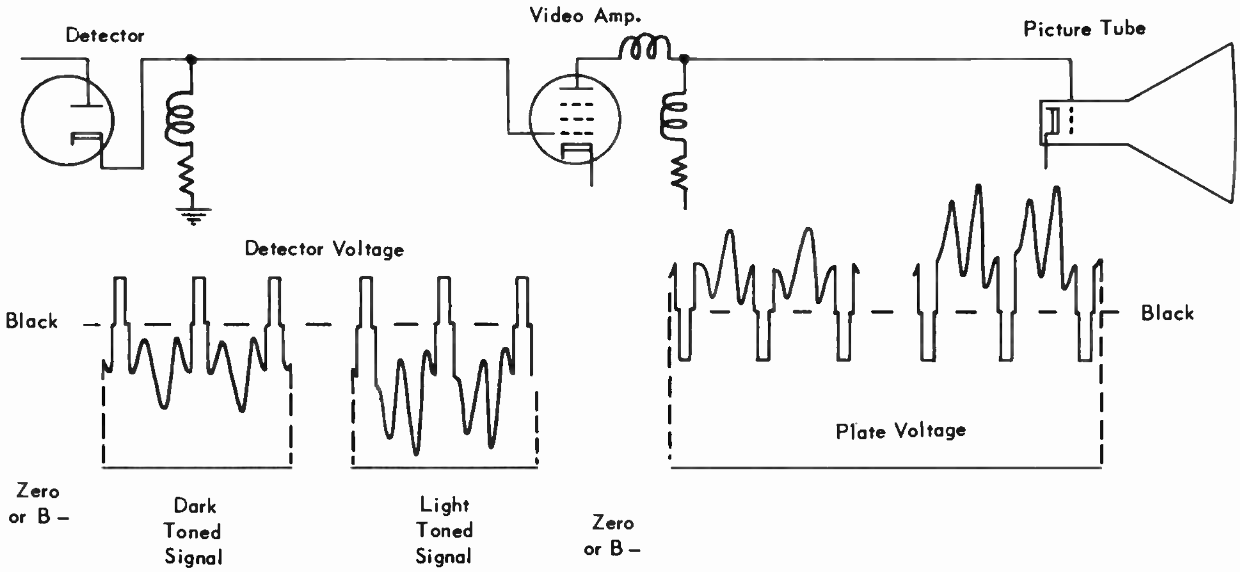


Fig. 50-13. How the d-c component is maintained when there are direct conductive couplings.

In diagram A of Fig. 50-12 there is only one capacitor in the signal line between detector and picture tube. The d-c component will be present at the detector output, and it will be carried through the video amplifier, only to be lost in the capacitor. Then d-c restoration must be carried out in the grid-cathode circuit of the picture tube.

In diagram B there are two capacitors in the signal path, one between detector and video amplifier, and the other between amplifier and picture tube. The d-c component will be lost in the capacitor that follows the detector. Were the component replaced here it would be lost in the capacitor that follows the amplifier. So the d-c component must be restored in the grid-cathode circuit of the picture tube.

In diagram C there is a capacitor between detector and video amplifier, but none between amplifier and picture tube. The d-c component might be restored in the grid circuit of the video amplifier or in the grid-cathode circuit of the picture tube. If the component is restored in the grid circuit of the video amplifier it will not be again lost between amplifier and picture tube, for between these two there is a direct conductive connection.

In diagram D of Fig. 50-12 there are no capacitors anywhere in the signal path from detector to picture tube. The d-c component will be present in the detector output, will pass through the direct conductive coupling to the video amplifier grid, and will pass through the similar coupling from amplifier plate to picture tube cathode. With circuits such as these there is no need for d-c restoration.

There is no need for d-c restoration where there are direct conductive couplings because the d-c component is not lost. The signal remains a pulsating direct voltage. How this comes about is illustrated by Fig. 50-13 where we have signal voltages for a dark toned picture and for one of light tone. At the detector output and at the input to the video amplifier these two signals are the same as at B and E of Fig. 50-10. They are pulsating direct voltages.

When the two signals go through the video amplifier they are inverted and made stronger. The added strength is not indicated in the diagram, it merely would make all the curves higher than shown. The variations which are the video signal now are variations of plate voltage. Plate voltage is a direct voltage which is pulsating. It is pulsating or varying with reference to zero or B-minus potential, which is a fixed potential. Consequently, the sync pulse tips and the black levels of both signals are at the same positive plate voltage. If the grid of the picture tube is biased to allow correct reproduction of either signal there will be correct reproduction of the other signal. If the black level of one signal is brought to the voltage for beam current cutoff the other black level also will be at this cutoff voltage.

In the following lesson we shall examine a variety of methods which are used to restore the d-c component after it has been lost in a coupling capacitor.

51→55

COMMERCIAL TRADES INSTITUTE

KNOWLEDGE + AMBITION = SUCCESS

LEARN . . .

Training Your Mind
C. T. I.
Perfecting Your Skill

ACT . . .

"The heights by great men reached and kept
Were not attained in sudden flight,
But they, while their companions slept,
Were climbing upward in the night."

ENJOY . . .

COMMERCIAL TRADES INSTITUTE
CHICAGO, ILLINOIS

TELEVISION

LESSON NO. 51

D-C RESTORATION

Training Your Mind

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

COMMERCIAL TRADES INSTITUTE

Training Your Mind

C.T.I.

Perfecting Your Skill

Chicago, Ill.

LESSON NO. 51

D-C RESTORATION

Now we are ready to restore to the video signal the d-c component that was left behind when the signal came through a coupling capacitor. On the output side of the capacitor we have only the video signal in a-c form. To this a-c signal must be added a new d-c component whose value will change in the same manner that the original d-c component is changing with every variation of picture tone. Since we have nothing to work with except the a-c video signal it is quite obvious that the value of the new d-c component must be changed by some characteristic of the a-c signal that corresponds to changes of the original d-c component, or to changes of tone and shading in the picture.

The one characteristic of the a-c video signal that changes with variations of picture tone or shading is amplitude. Amplitude of the a-c video signal increases for light toned pictures and decreases for pictures of dark tone. The amplitude always is very nearly inversely proportional to the lost d-c component.

If we can add a new direct voltage to the a-c video signal, and let amplitude of this signal vary the strength of added direct voltage, we shall have restored the d-c component. This is the basic principle of d-c restoration systems – to let amplitude or peak-to-peak voltage of the a-c video signal control the strength of a direct voltage that is added after the last coupling capacitor that precedes the picture tube.

The video signal with its new d-c component can be applied to the grid-cathode circuit of the picture tube. The added d-c portion or component of the signal can be made to vary the picture tube grid bias proportionately to picture illumination or tone, and the black level of the signal will be held at the grid voltage required for beam cutoff.

One of the simplest ways of restoring the d-c component is shown by Fig. 51-1. There is grid leak bias on the video amplifier tube that precedes the picture tube. The amplifier signal output is directly connected to the picture tube grid. Part of the d-c voltage in the amplifier plate circuit becomes the d-c component added to the video signal.

In our very elementary diagram the grid return circuit for the picture tube includes resistor *Ro* and the portion of the brightness control resistance which is to the left of the slider. If we change the average current and average voltage drop in resistor *Ro* it will change the picture tube grid bias, for average voltage drop across *Ro* is part of the potential difference or part of the biasing voltage between grid and cathode of the picture tube. Current in resistor *Ro* is plate current for the video amplifier, so changing the average plate current will alter the grid bias on the picture tube.

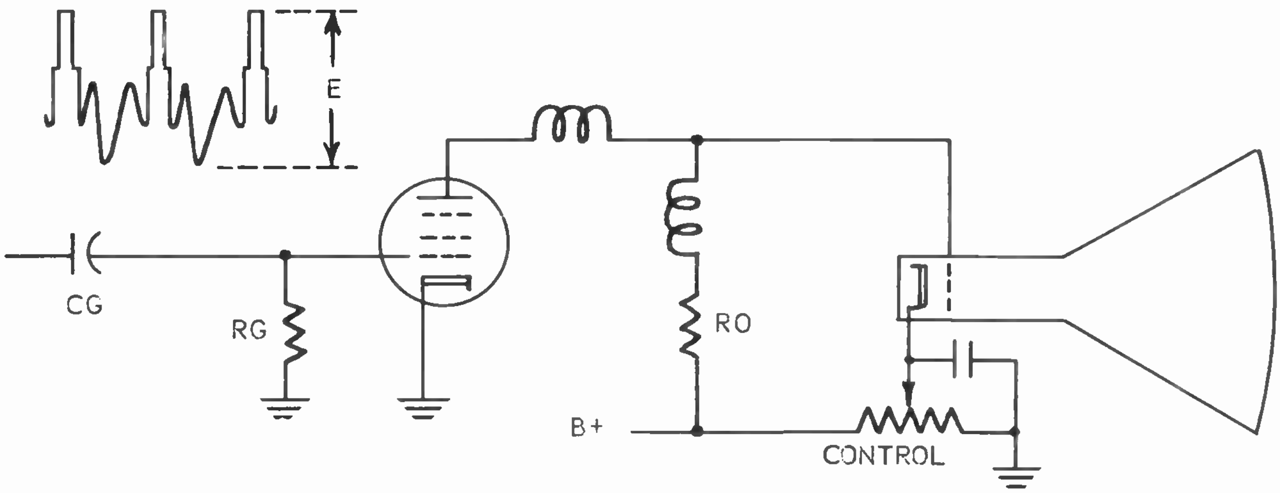


Fig. 51-1. D-c restoration from grid-leak bias on video amplifier tube.

In the grid circuit of the video amplifier tube are the familiar grid resistor *Rg* and grid capacitor *Cg* as always used for obtaining grid-leak bias. It is in capacitor *Cg* that the original d-c component is lost.

Let's assume that the a-c signal coming through capacitor *Cg* changes from that corresponding to a dark picture to one corresponding to a lighter tone. Amplitude or peak-to-peak voltage, *E*, of the signal will increase. Whenever there is grid-leak bias an increased voltage of the incoming signal makes the bias more negative. We learned all about this action long ago.

When amplifier grid bias becomes more negative there is less amplifier plate current, less current in resistor *Ro*, and less voltage drop across this resistor. This lets picture tube grid voltage come closer to cathode voltage, we have reduced the voltage between these two elements. Then the grid is less negative with reference to the picture tube cathode, which is just what we need for a picture of lighter tone. If the signal changes to that for a dark toned picture, the amplifier grid bias will decrease, there will be more amplifier plate current, more drop in resistor *Ro*, and the picture tube grid will be made more negative with reference to the cathode – as required for a picture of dark tone.

Fig. 51-2 is a circuit diagram for a complete video amplifier system in which grid-leak restoration of the d-c component occurs in capacitor *Cc* and resistor *Rg* on the second amplifier tube. In addition to grid-leak bias there is cathode-bias on this tube. There are no coupling capacitors in any parts of the signal-carrying circuits beyond the point of d-c restoration.

When the a-c video signal passes from video amplifier plate to picture tube grid or cathode through a capacitor, the d-c component from the plate load resistor cannot go through the capacitor. There are, however, several ways in which changes of d-c voltage on the plate load resistor may be transferred into the grid-cathode circuit of the picture tube to provide d-c restoration.

Some of these methods are illustrated by diagrams in Fig. 51-3. The capacitor through which the a-c

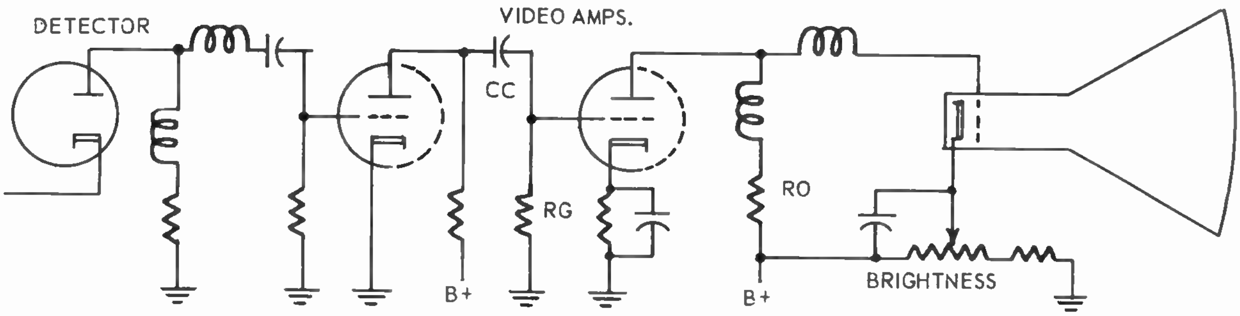


Fig. 51-2. Amplifier system with d-c restoration in grid circuit of second video amplifier tube.

video signal goes to the picture tube grid or cathode is marked *Cc*. The load resistor in the plate circuit of the video amplifier is marked *Ro* in all three diagrams. In diagram *C* there is an additional resistor, *Rn*, in series with the load resistor.

Resistors *Ra* and *Rb* are in series with each other and the two are in parallel with the load resistor or resistors. Resistance of the regular plate load units, *Ro* or *Ro* and *Rn*, will be on the order of 5,000 to 15,000 ohms. Resistances at *Ra* and *Rb* are equal to each other, and each one has a value of one to two megohms or even more. Practically no plate current will flow in these very high resistances, but across them will be the same d-c voltage drop that is across the low-resistance units which they parallel. Half of this d-c voltage, the half that appears across *Rb*, is in the grid-cathode circuit of the picture tube. This d-c voltage, which varies with d-c voltage across the plate load resistance, varies the grid bias of the picture tube to restore the d-c component.

RESTORATION WITH DIODES. There are a number of systems of d-c restoration with which a diode in the grid cathode circuit of the picture tube provides a picture tube grid bias voltage that varies with signal strength. The principle is much the same as employed for grid-leak biasing of an amplifier tube, where the grid and cathode of the amplifier, acting as a diode, provide a bias voltage that varies with strength of the applied a-c signal.

You will recall that grid-leak biasing for any tube containing a grid and cathode depends on the gridand cathode acting like the plate and cathode of a diode. The diode action rectifies part of the a-c signal voltage coming to the grid through a capacitor. Rectified positive alternations of the signal charge this capacitor. Capacitor voltage, corresponding to its charge, becomes a negative bias voltage for the grid. The charge leaks slowly away through the grid resistor. Since voltage across this resistor is the same as capacitor voltage, it too is the biasing voltage for the tube.

One general type of diode restoration circuit is shown by Fig. 51-4. The a-c video signal passes from video amplifier plate to picture tube grid through capacitor *Cc*. The a-c or high-frequency circuit is completed from picture tube cathode through capacitor *Ca* to ground, and from ground through bypass capacitor *Cb* back to the plate load of the amplifier.

The conductive grid return circuit or d-c grid circuit is completed from the picture tube grid through re-

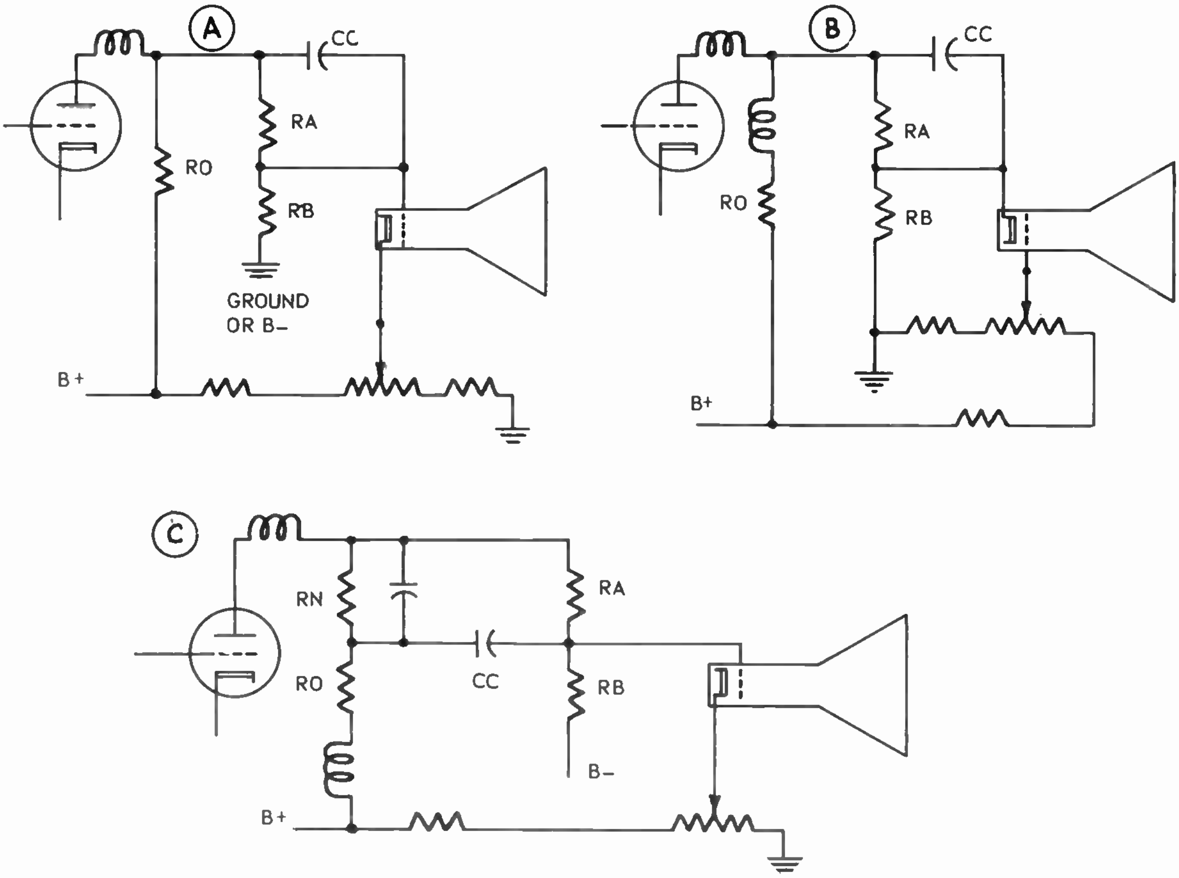


Fig. 51-3. Restoration from parallel resistors across the plate load of the video amplifier.

sistors *Ra* and *Rb* to ground, through ground to one end of the brightness control resistances, and from the control slider to the picture tube cathode. Grid bias voltage will be the potential difference across resistors *Ra*, *Rb*, the resistor in series with the brightness control, and the portion of the control resistance to the left of the slider. Voltage drop across part of the brightness control element and its series resistor tends to keep the picture tube grid negative with reference to its cathode. Now we shall see what happens to voltage across resistor *Rb*.

Resistance at *Rb* is high, usually a megohm or more, while resistance at *Ra* is only a few thousand ohms. The a-c video signal is applied through *Cc* and *Ra* to the diode. Negative peaks of the signal make the diode cathode negative with reference to its plate, and there is electron flow through the low internal resistance of the diode, as shown by broken-line arrows. This electron flow charges capacitor *Cc* in the polarity as marked, which is positive on the side toward the picture tube grid.

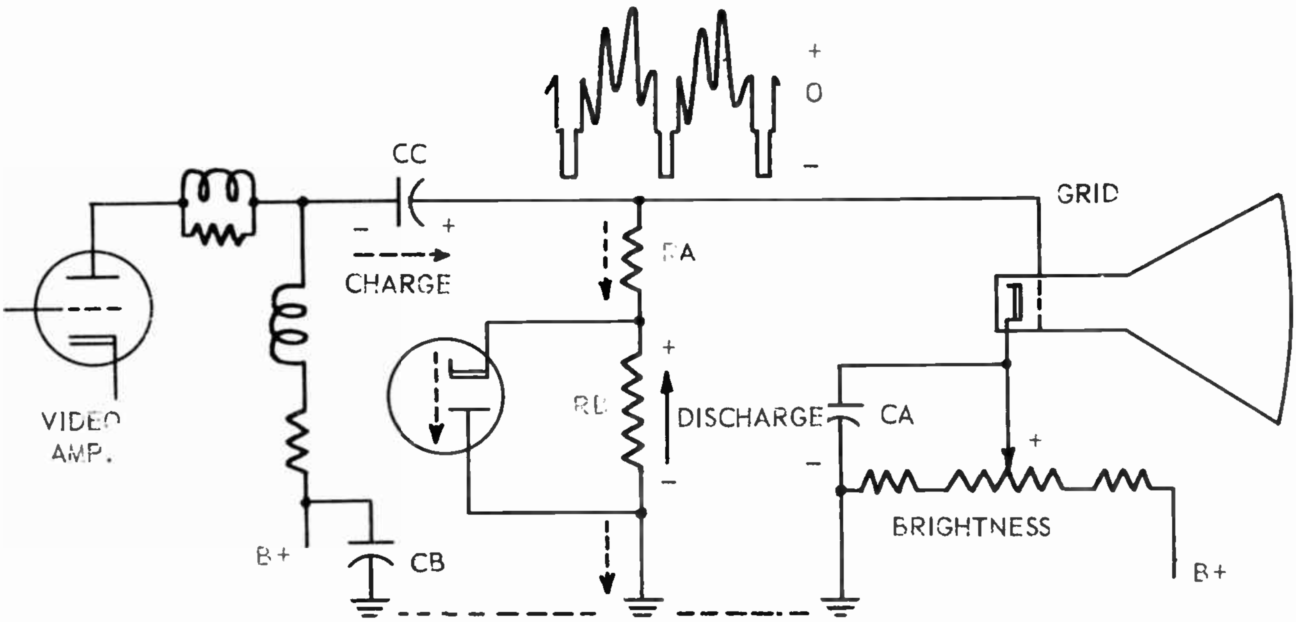


Fig. 51-4. Diode restoration with signal input to picture tube grid.

During opposite alternations of the video signal the diode cathode is made positive with reference to its plate. Then the diode is non-conductive, and charge on capacitor *Cc* can escape only slowly through the high resistance of *Rb*, which is across the diode. The charge and voltage on *Cc* are thus built up to values corresponding to amplitude or peak-to-peak voltage of the video signal. The voltage on the capacitor exists also across resistors *Ra* and *Rb*, which are between the grid and cathode of the picture tube. This voltage, like that on the grid side of capacitor *Cc*, positive at the picture tube grid.

Positive voltage applied to the picture tube grid from resistors *Ra* and *Rb* opposes the negative grid voltage from the brightness control. Since this opposing positive voltage is proportional to amplitude of the video signal the picture tube grid is made more positive, or actually less negative, by strong signals for light toned pictures. For dark toned pictures there is less signal amplitude, less positive voltage across resistors *Ra* and *Rb*, and the picture tube grid becomes less positive, or really more negative. These changes of picture tube grid voltage are just what is needed for d-c restoration.

We have not interfered in the least with passage of the a-c video signal to the picture tube grid. The combined action of the brightness control and restoration system merely holds the picture tube grid at an average voltage suited to the tone of the received picture. The a-c signal forms pulsations or variations of this average d-c voltage, exactly as in the cases where we had direct conductive connections for the signal.

When video signal input is to the cathode of the picture tube, rather than to the grid, it becomes necessary to reverse the connections of diode plate and cathode on the restoration resistor string, as has been done in Fig. 51-5. The diode is made conductive, for charging of capacitor *Cc*, when the diode plate becomes positive with reference to its cathode during positive alternations of the applied a-c signal. Electron flow through the low internal resistance of the diode then is as shown by broken-line arrows.

This electron flow in capacitor *Cc* is in such direction as to make the side toward the picture tube cathode of negative polarity, as marked. This negative voltage exists also across resistors *Ra* and *Rb*, which are in the grid return circuit of the picture tube. The stronger the a-c video signal the greater is the charge built up on capacitor *Cc* and the greater is the negative voltage across resistor *Ra* and the high resistance at *Rb*. This negative voltage, which is proportional to signal strength, tends to make the picture tube cathode more negative, or less positive, with reference to the picture tube grid. This, of course, is equivalent to making the grid less negative on stronger signals.

Voltage drop in the brightness control element and its series resistor is in such polarity as to maintain the picture tube grid negative with reference to its cathode at all times. The restoration voltage developed across resistors *Ra* and *Rb* merely alters this negative grid bias in accordance with whether video signals are for pictures of light or dark tone.

In a number of receivers the restoration diode is replaced by a triode, connected as in Fig. 51-6. If you consider the cathode and grid of this triode as equivalent to the cathode and plate of the Diode in Fig. 51-4 it becomes plain that the restoration system is identical in the two systems. The cathodes of both diode and triode are connected between resistors *Ra* and *Rb*. The plate of the diode goes to ground, and so does the grid of the triode.

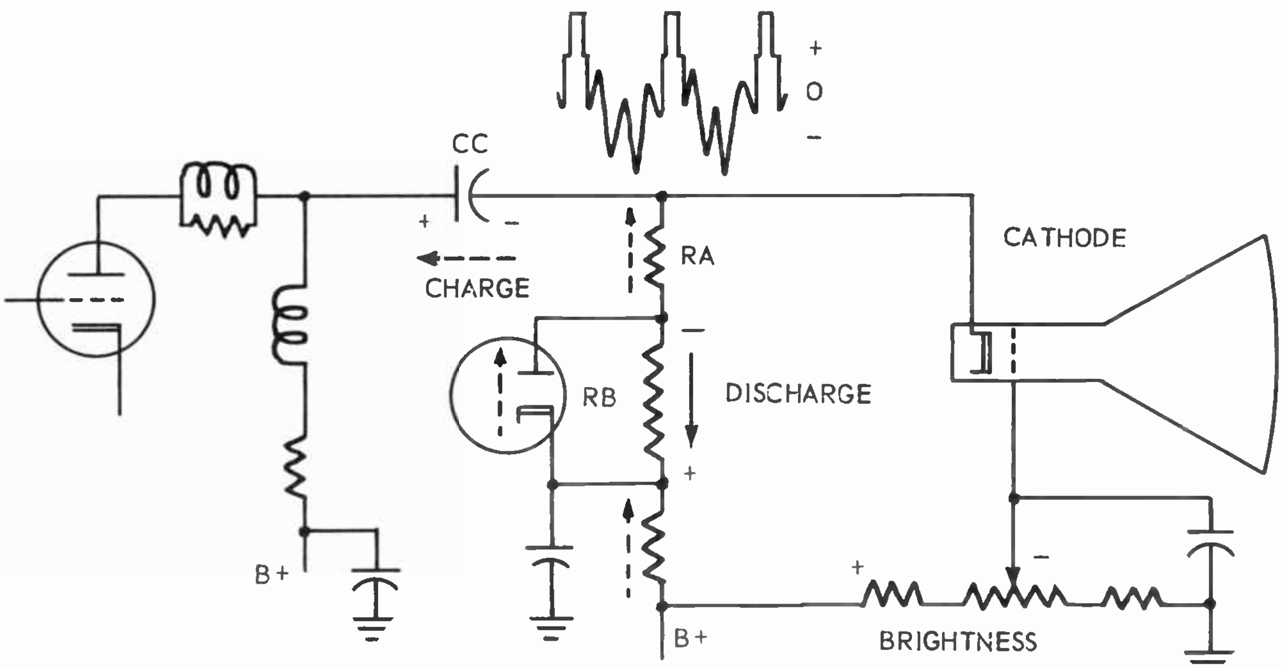


Fig. 51-5. Diode restoration with signal input to picture tube cathode.

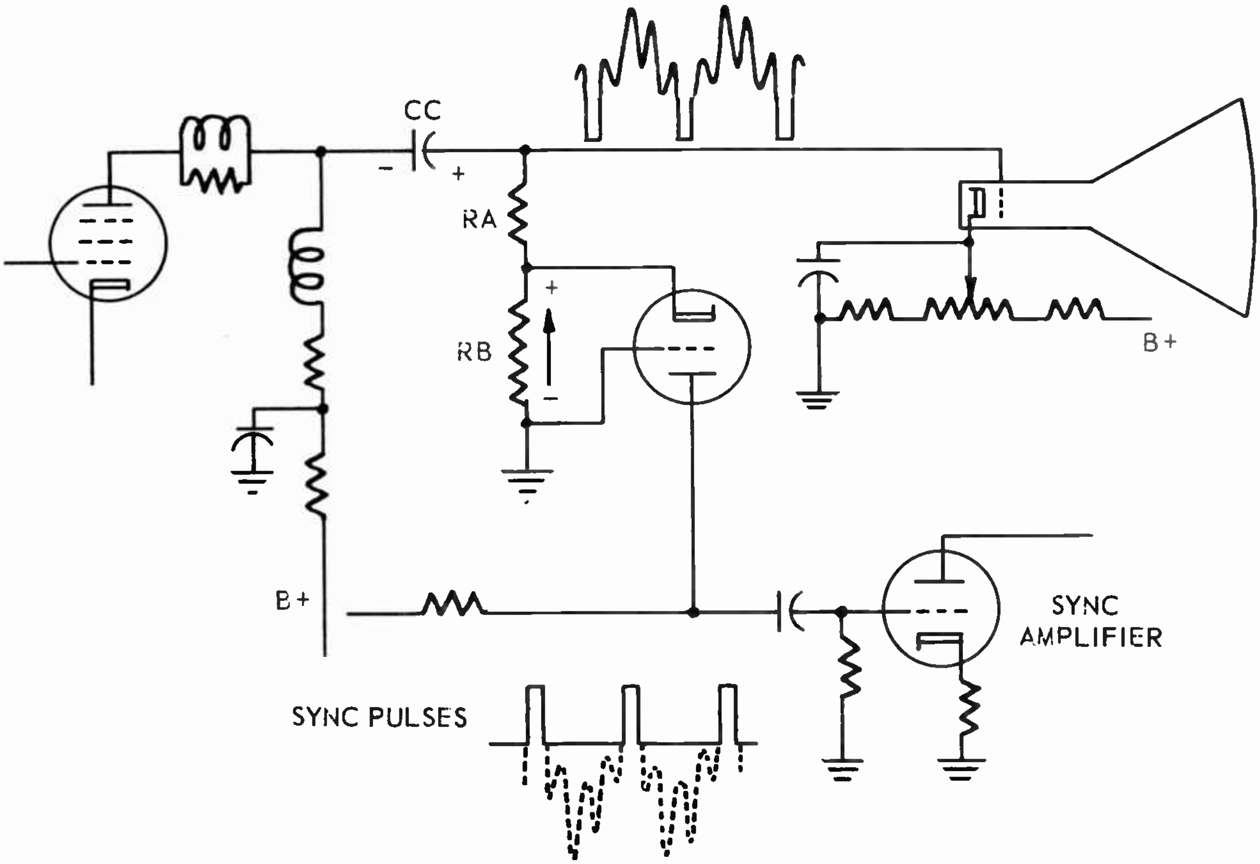


Fig. 51-6. D-c restoration by means of a triode acting also for sync takeoff.

As capacitor *Cc* is continually discharging through resistor *Rb* the electron flow is in such direction that the cathode of the triode in Fig. 51-6 is positive with reference to its grid, or the grid of this triode is negatively biased. This bias is so far negative, due to high resistance in *Rb*, that only the sync pulses cause conduction from cathode to plate in the triode. The sync pulses, which are negative in the a-c video signal, are inverted in the triode and are taken from the triode plate to the grid of an amplifier tube in the sync section of the receiver.

The triode in the grid-cathode circuit of Fig. 51-6 is acting as a diode for d-c restoration and also as a triode for takeoff of the sync pulses which eventually will control deflection in the picture tube. A little later we shall learn more about sync sections in general. This particular circuit has been mentioned here because it is so closely associated with d-c restoration.

Fig. 51-7 shows circuit connections for a system of diode restoration which is quite generally employed, often with minor modifications which do not affect the fundamental operating principles. The a-c video signal is taken from the plate of the amplifier to the picture tube grid through capacitor *Cc*. The same signal, in lesser strength, is taken from the top of plate load resistor *Ro* through another resistor *Rc* and through capacitor *Cd* to the cathode of the restoration diode. Other than this different manner of getting the a-c video signal to the restoration diode, this system is much the same in action and performance as the one illustrated by Fig. 51-4.

In Fig. 51-7 the a-c video signal applied across the diode and resistor *Rb* is rectified by the diode when signal alternations are negative, to make the cathode of the diode negative with reference to its plate. Pulses of rectified current charge capacitor *Cd* in the polarity marked. The voltage corresponding to capacitor charge exists also across resistor *Rb*, which is part of the grid return circuit of the picture tube. This positive restoration voltage is applied through resistor *Ra* to the picture tube grid, in opposition to the negative grid voltage maintained by the brightness control. Thus the picture tube grid is made less negative for light toned signals, which cause higher charge voltage on *Cd*, and is made more negative for dark toned signals, during which part of the charge and its voltage escape through resistor *Rb* and the ground connections.

4 In all of the d-c restoration systems which have been examined the voltage for restoration is developed by charging a capacitor, whose charge escapes slowly through one or more resistors. The time constant of this capacitor-resistor combination must be long enough to hold the charge and the restoration voltage practically constant during several horizontal line periods, so that average bias of the picture tube grid or amplifier grid will not be varied by changes of illumination in different parts of the same scene. On the other hand, the time constant must be short enough to change whenever there is a change of overall brightness in the scene.

In various receivers the time constant of d-c restoration may be almost anything between 0.025 and 0.100 second. A constant of 0.050 second often is used, which is somewhat longer than the time period for one frame, but less than that for two frames.

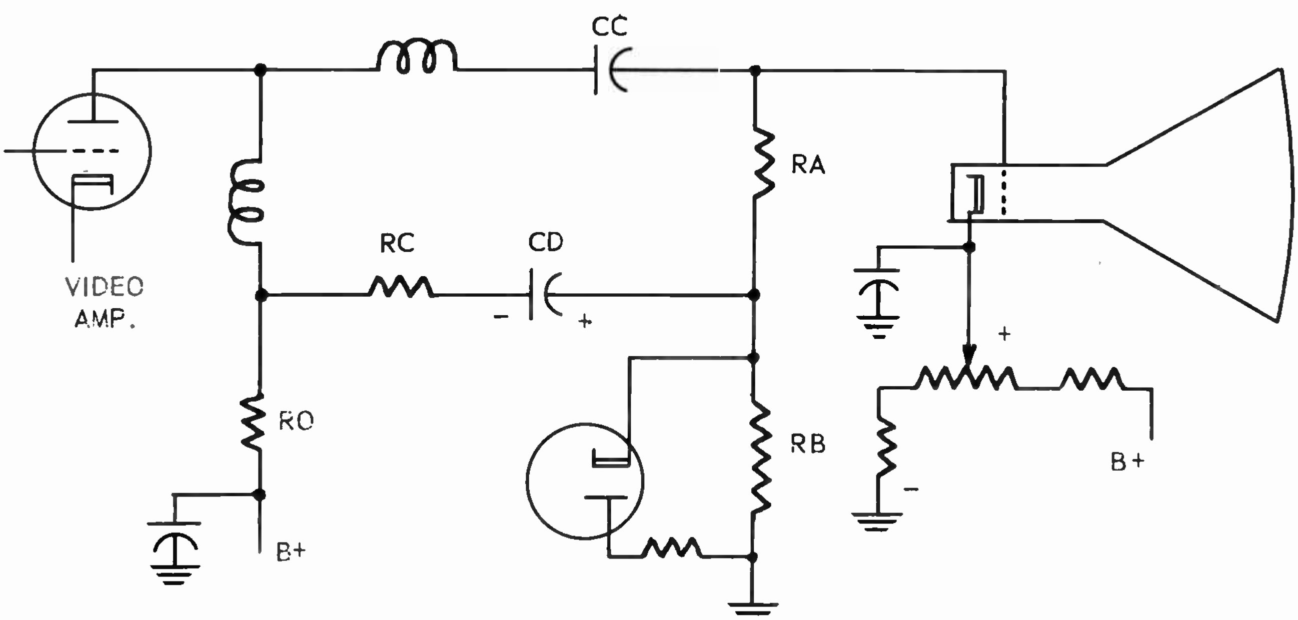


Fig. 51-7. D-c restoration with separate capacitor connection to the rectifying diode.

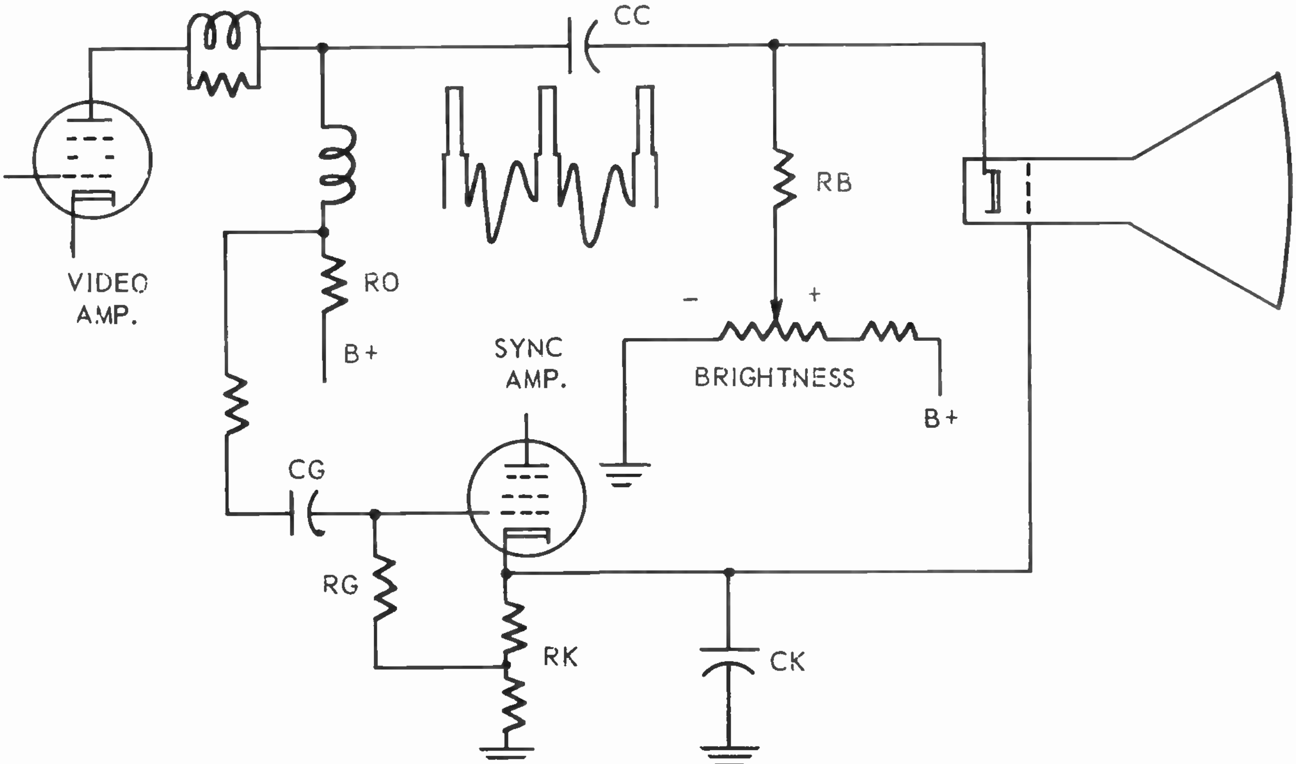


Fig. 51-8. Restoration voltage taken from cathode resistor in sync amplifier tube.

In Fig. 51-6 we looked at a system in which a triode used primarily for d-c restoration acts also as a sync takeoff tube. Now, in Fig. 51-8, we shall look at a system in which a pentode which is primarily a sync takeoff or amplifier tube acts also for d-c restoration. The a-c video signal from the plate of the video amplifier is applied through capacitor *Cc* to the cathode of the picture tube. The cathode is connected also to the usual type of brightness control. Resistance of something like 100,000 ohms at *Rb* prevents losing the signal through the brightness control connections, and keeps it on the cathode.

The same a-c video signal is taken from above load resistor *Ro* in the amplifier plate circuit and is applied through capacitor *Cg* to the grid of the sync amplifier. This sync tube is negatively biased by grid-leak action of capacitor *Cg* with resistor *Rg*, also by cathode bias voltage developed in cathode resistor *Rk*. The bias is so strongly negative that only the positive sync pulses cause conduction in the sync tube.

Pulses of current due to sync tip voltage on the grid of the sync amplifier flow through resistor *Rk*. The potential difference across *Rk* causes charging of capacitor *Ck*. There is large capacitance at *Ck*, usually between 2 and 10 mf, so that even with resistance of only a few thousand ohms at *Rk* the time constant of this combination is on the order of 0.100 second.

Were there no conduction in the sync amplifier, and neither current nor voltage drop across resistor *Rk*, the grid of the picture tube would be at ground potential for the reason that the grid is connected to the top of *Rk*. Under such a condition the picture tube grid would have the full negative bias voltage which is between the slider of the brightness control and ground.

When there is conduction in the sync amplifier, and voltage across *Rk* and *Ck*, this voltage is of such polarity as to make the picture tube grid positive, or, at least, to make it less negative. This becomes evident when you consider that electron flow must be upward in *Rk*, toward the tube cathode, which means that the top of *Rk* is positive with reference to the grounded lower end.

The stronger the a-c video signal the greater will be the positive voltage of the sync tips on the grid of the sync amplifier, and the greater will be current and voltage drop in resistor *Rk*. Thus the effect of the stronger signal is to make the picture tube grid less negative, as is necessary for d-c restoration. A weaker video signal, for a picture of darker tone, will allow escape of some charge voltage from capacitor *Ck*. Then the picture tube grid is allowed to remain more negative for the dark toned signal.

BLACK LEVEL CONTROLS. It was explained earlier that the black level of the video signal is held at the picture tube grid voltage necessary for beam cutoff by readjustment of the brightness control when there is a change in the setting of the contrast control. This is due to the fact, illustrated at *A* in Fig. 51-9, that the black level of an a-c video signal changes its position with reference to the average or zero voltage when there is a change of signal strength, such as might be caused by varying the contrast control. The black level comes closer to zero voltage for a dark toned picture than for one of light tone.

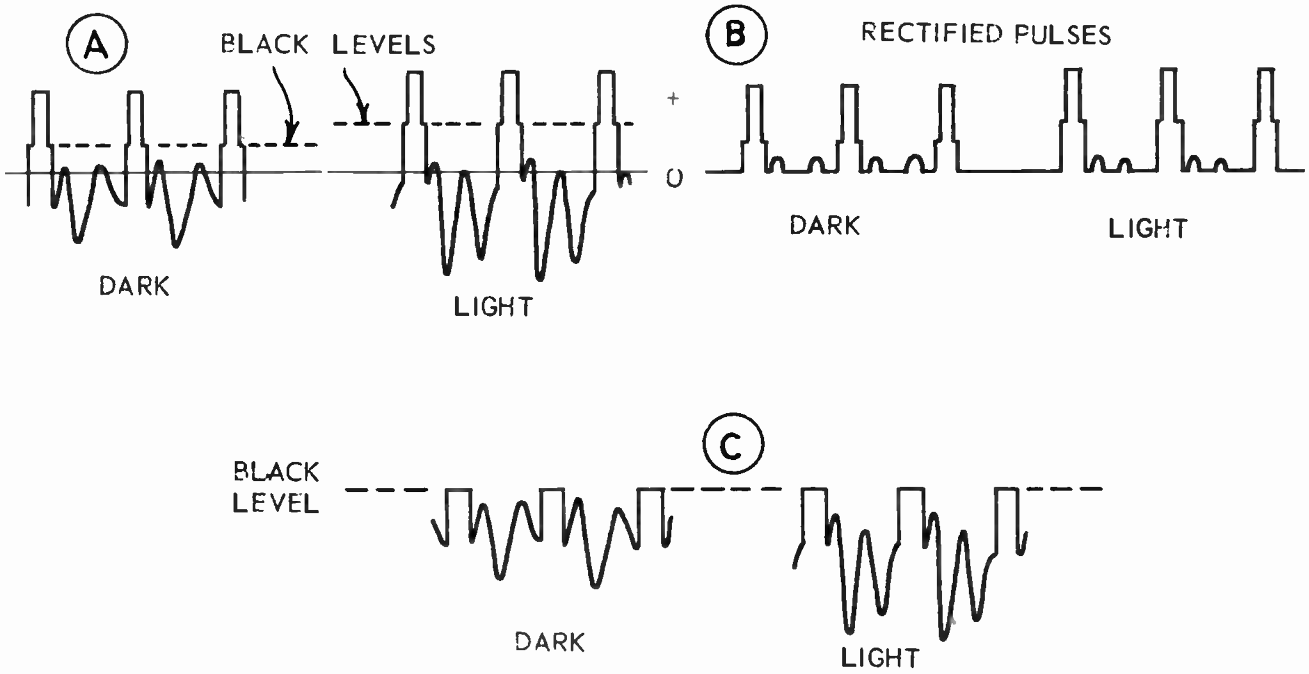


Fig. 51-9. How the black level voltage is affected by picture tone.

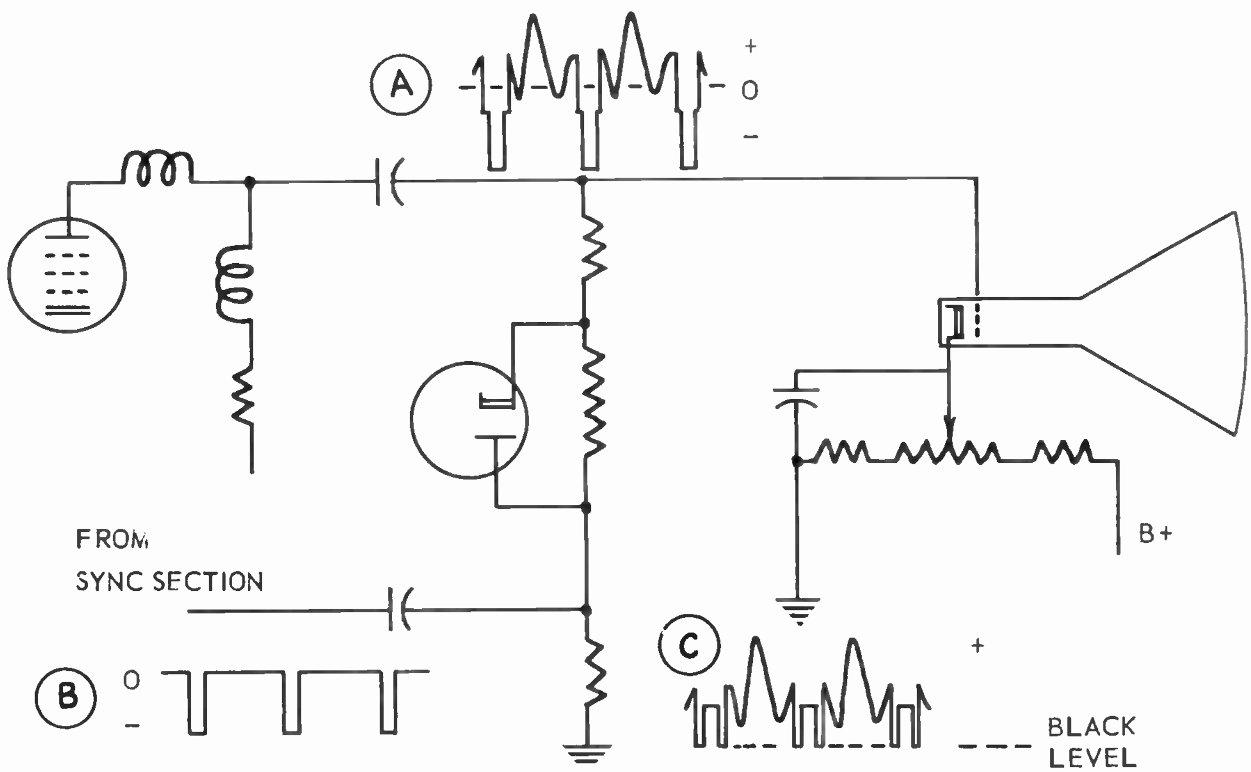


Fig. 51-10. Black level control of d-c restoration voltage.

In systems of d-c restoration employing diodes, triodes, or pentodes wherein rectification or conduction charges a capacitor to the restoration voltage, the charge is proportional to voltage or amplitude at the tips of the sync pulses. You will see, by looking back through preceding circuit diagrams in this lesson, that the sync pulse tips always act on the rectifier cathode when these tips are negative, and on the rectifier plate when the tips are positive. At *B* are shown the rectified portions of the signal after it has passed through the restorer tube. These are the voltage pulses that charge the capacitor on which is built up and maintained the restoration voltage.

Since d-c restoration is intended to compensate for changes of picture brightness rather than for changes in overall strength, it would be desirable to base the restoration voltage on the portion of the signal lying between the black level and the peaks representing lightest tone or greatest illumination. This would be the portion of the signal shown at C in Fig. 51-9. Systems have been devised by which the sync pulses are removed from the portion of the signal coming through the restorer tube. This leaves only the picture signal or the black level to determine the value of restoration voltage. Such systems may be called black level controls.

Principles utilized for one type of black level control are illustrated by circuit of Fig. 51-10. At *A* is the a-c video signal which is applied to the picture tube grid and to the cathode of the restorer diode. The negative sync pulses on the cathode would make the diode conductive. To the plate of the diode are ap-

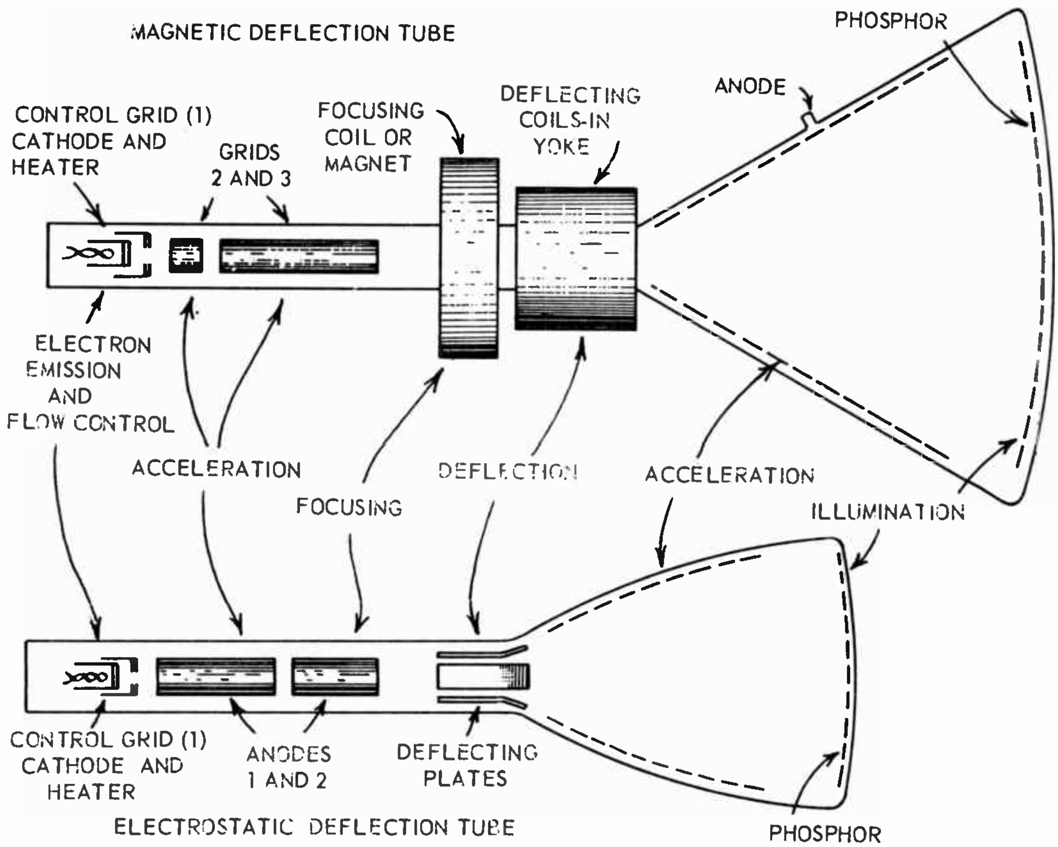


Fig. 51-11. What happens to the electrons and the electron beam in picture tubes.

plied at the same time a series of negative pulses shown at *B*. These negative-pulses are taken from the sync section of the receiver. At several points in the sync section are negative pulses which originated from the sync pulses of the video signal, hence are precisely in time with the sync pulses being applied to the cathode of the restorer diode.

The negative voltage pulses applied to the restorer plate along with negative pulses on its cathode leave plate and cathode at practically the same voltage during pulse intervals. The result, shown at *C*, is to remove the sync pulse voltages and leave only the front and back porches, which are at the black level, as the most negative portions of the voltage acting for restoration.

PICTURE TUBE PERFORMANCE. In all our work with gain, contrast, brightness, and d-c restoration we have been talking about video signal voltages which ultimately are applied to the grid or cathode of the picture tube as a means for controlling the rate of flow in the electron beam. Now it is time to see just what happens in the picture tube when video signals and other voltages are applied to the elements.

We may consider the cathode as the primary element in the picture tube, it is the one from which electrons are emitted. The grid is the first of the control elements, the one that controls the rate at which emitted electrons leave the vicinity of the cathode.

The grid and cathode of the picture tube have much the same relation to each other as the grid and cathode in amplifier tubes, a relation which has made it easy to understand the effects of grid voltage, cathode voltage, grid biases, and alternating signal voltages. Beyond the grid in the picture tube are other elements that act like the plate or the plate and screen of amplifier tubes in pulling electrons from the cathode through the grid, but which are of quite different physical form than plates or screens and which carry out their functions in different ways.

All the functions carried out in picture tubes are illustrated in Fig. 51-11. First, in tubes having either magnetic or electrostatic deflection, the electrons are emitted from the cathode because of heat energy. Then the rate at which electrons flow away from the cathode is regulated by the control grid or grid number *1*.

The next step is to accelerate the electrons to much greater velocity than that with which they leave the cathode and pass through the control grid. Then the electron stream is focused into a narrow beam which will converge to a very small spot on the screen. In the magnetic tube there is acceleration first by grid number *2*, then there is focusing by means of an electromagnetic coil or a permanent magnet or by a combination of both, placed around the outside of the tube neck. In the electrostatic tube there is acceleration and focusing by means of suitable voltages applied to anodes *1* and *2*.

After acceleration and focusing, the electron beam in the magnetic tube is deflected or swept horizontally and vertically by electromagnetic coils in the deflecting yoke, which is around the outside of the tube neck just ahead of the focusing coil or magnet. In the electrostatic tube the focused beam is deflected or swept horizontally and vertically by voltages applied to internal deflecting plates between which the electrons are traveling.

The conductive coating inside the flare of a glass picture tube, or the inner surface of the cone on a metal-cone tube, is made of or covered with dull black material, such as various preparations of finely divided graphite. A dull black surface provides minimum reflection of light from the inside of the tube back to the screen. Reflection of this nature would lighten the entire screen and would reduce contrast.

The internal conductive coating of magnetic deflection glass tubes is the anode, to which a high-voltage connection is made through a recessed terminal formed as a ball or cap. The anode of metal tubes is the cone, to which a high-voltage connection is made at the rim. The internal conductive coating of glass tubes employing electrostatic deflection is electrically a part of the second anode, being connected inside the tube to the cylindrical second anode of the electron gun which, in turn, is connected to a base pin taking the high-voltage lead.

When electrons strike the screen or phosphor material at high velocity they knock other electrons out of the screen. This action is called secondary emission. There actually is emission of electrons from the screen, not because of heat but because of the force of electrons in the beam. The emitted secondary electrons, which are negative, are attracted to the highly positive internal coating or metal cone, and through it are returned to the high-voltage circuit that includes the anode. Otherwise the emitted electrons would collect inside the non-conductive flare of glass tubes or would fall back onto the screen and accumulate there.

In a magnetic deflection picture tube there is additional acceleration of beam electrons by the anode voltage. In electrostatic deflection tubes there is additional acceleration by the part of the second anode which is the internal conductive coating.

The process of focusing, and the process of centering the beam or picture on the screen, will be studied in a later lesson. Deflection, either magnetic or electrostatic, is the final result of what goes on in the sync section and the sweep section of the receiver, so will wait for consideration until after these sections

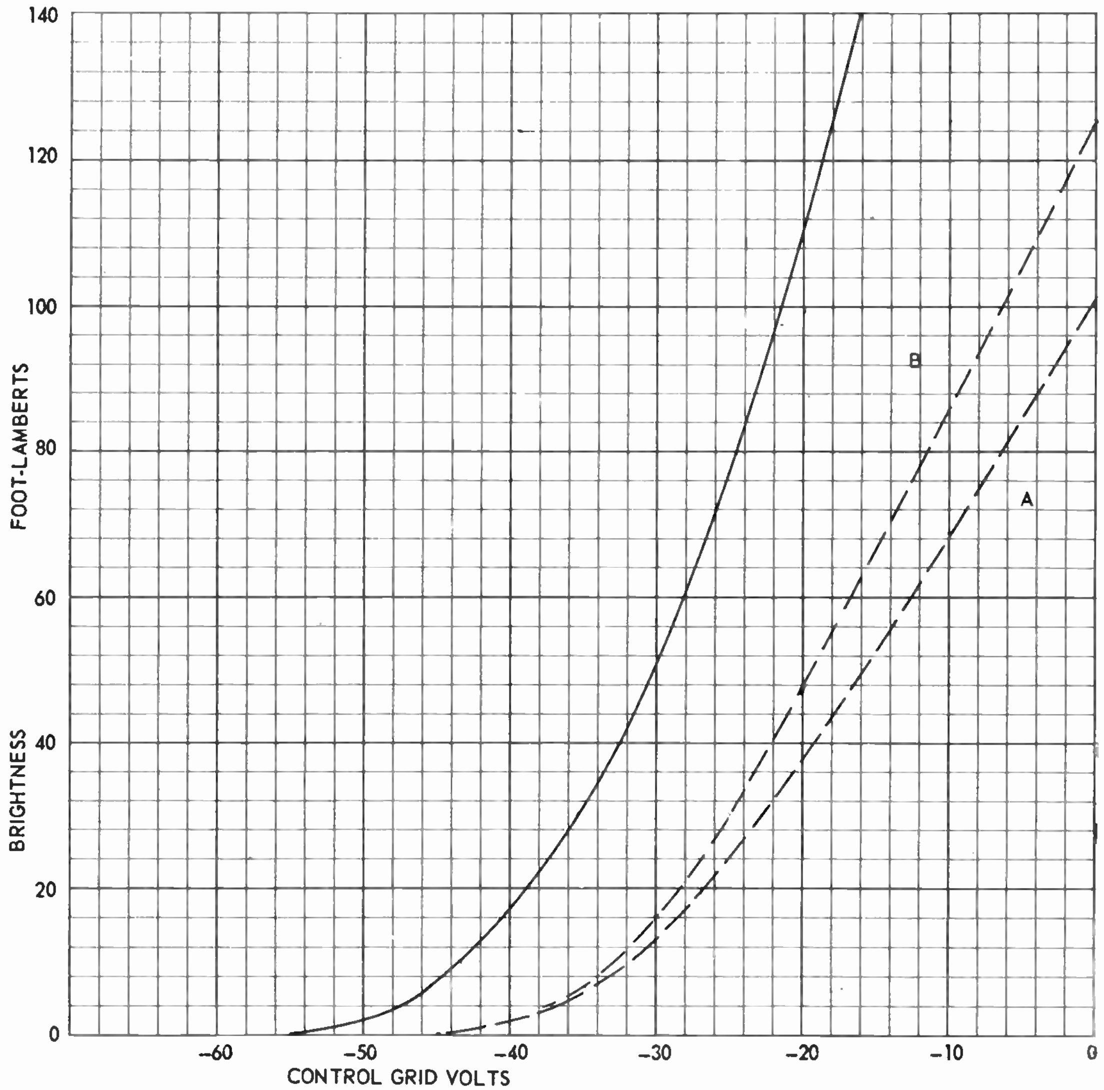


Fig. 51-12. Relations between screen brightness and control grid voltage for two magnetic deflection picture tubes.

have been examined. Just now we shall look into some of the relations between voltages that accelerate the electron beam and which determine to a great extent the brightness of the spot appearing on the screen.

The curves of Fig. 51-12 show relations between control grid voltage and screen brightness of two different magnetic deflection tubes. These curves have been selected merely as typical examples of what happens in a similar way with all types of picture tubes. The full-line curve shows performance of a 17-inch rectangular picture tube operated with 14,000 volts on its anode, which is midway between the highest permissible voltage and the minimum which allows satisfactory performance. Circuit design and operation are such that there is complete cutoff of the electron beam when the control grid is made 55 volts negative with reference to the cathode.

The broken-line curve *A* of Fig. 51-12 applies to a 12-inch round tube operated at 9,000 anode volts, the minimum which gives satisfactory performance with this particular tube. Operation is such as to cause complete cutoff of the electron beam when the control grid is 45 volts negative. Broken-line curve *B* shows performance of this 12-inch tube when its anode is operated at 11,000 volts, which is within 1,000 volts of the maximum permissible value for this tube.

The graph shows brightness in foot-lamberts. A foot-lambert is a unit commonly used for measuring average brightness of a surface that is emitting or reflecting light. This unit is equivalent to the light which would be thrown on a surface by one "international candle" placed one foot from the surface. The international candle is a sperm candle 7/8 inch in diameter burning its wax at a certain rate. You may gain some idea of the meaning of brightness measured in foot-lamberts from the fact that illumination equivalent to between 10 and 20 foot-lamberts is ample for reading ordinary printed matter, such as books or magazines.

From curves *A* and *B* of Fig. 51-12 it is evident that increasing the anode voltage increases the average brightness of a picture when control grid voltage remains unchanged, or when there is the same variation of grid voltage due to the picture portion of the video signal. Decrease of anode voltage lessens the average brightness. Excessive decrease causes poor definition, or loss of detail in pictures.

Maximum permissible anode voltage for 10-inch and 12-inch magnetic deflection tubes usually is specified as 12,000 volts. In order that there may be satisfactory brightness and definition these tubes should not be operated with less than 8,000 volts on their anodes. All anode voltages are, of course, measured with reference to cathode potential, as is the case with all tubes containing cathodes. For most 16-inch tubes the maximum anode voltage is 14,000, with a low limit of about 9,000 volts for satisfactory picture reproduction. For many 17-inch tubes the maximum is 15,000 to 16,000 volts, with a practical minimum of 12,000. For 19- to 22-inch tubes the maximum is 19,000 to 20,000 anode volts, with a practical minimum as low as 12,000 volts.

Curve *A* of Fig. 51-13 shows the relation between anode current and brilliance of the spot formed by the electron beam on the picture tube screen. We may take this current as roughly proportional to the rate of electron flow reaching the screen. As we should expect, brilliance of the spot and resulting brightness of the picture is almost directly proportional to the rate of electron flow. The greater the number of electrons reaching any given small area of the screen during some brief fraction of a second, the more intense is bombardment of the phosphor material by the electrons and the greater is the atomic activity and resulting emission of light. It is voltage on the control grid, as varied by the video signal, that regulates the rate of electron flow to the screen and thus varies the brilliance of the spot traveling across the screen.

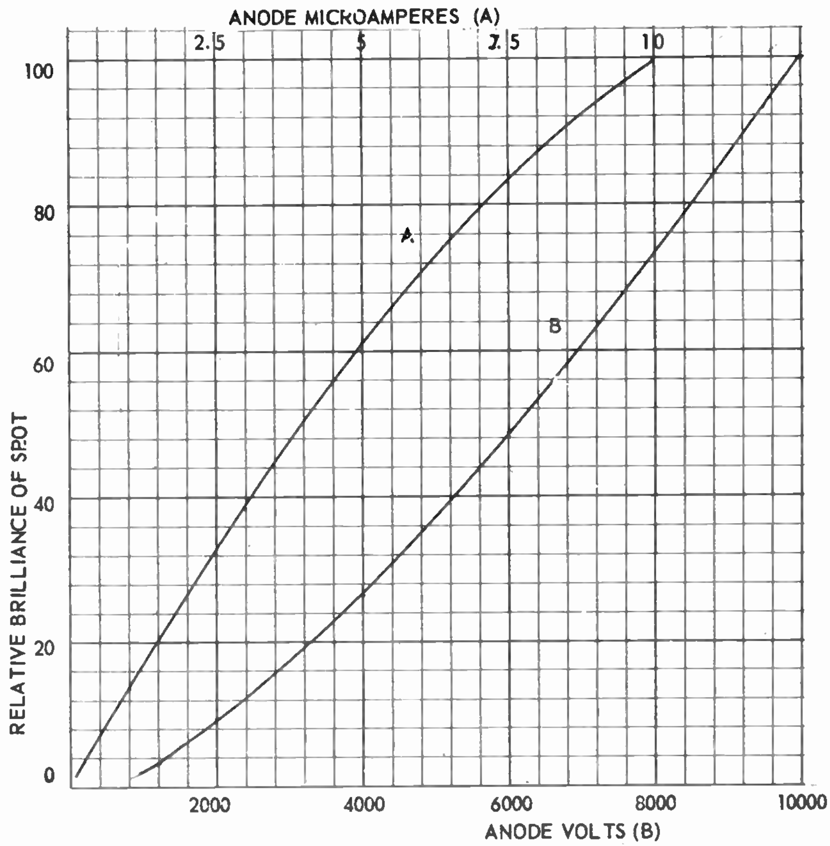


Fig. 51-13. Relations between spot brilliance and anode current (curve *A*) and between brilliance and anode voltage (curve *B*).

Curve *A* of Fig. 51-13 shows the effect on brilliance of varying the rate of electron flow in the beam when voltage on the picture tube anode remains constant. Curve *B* shows the effect on brilliance of varying the anode voltage when the rate of electron flow remains constant. Increase of anode voltage causes a nearly proportional increase of spot brilliance. This is because the greater voltage imparts greater velocity to every electron. The electrons then strike the phosphor material with greater force. Each electron then causes greater activity of particles making up the atoms of phosphor material, and there is increased emission of light. The curves in this figure do not apply to any particular picture tube, rather they apply to any tube containing the type of phosphor used in television picture tubes.

The chief purpose of grid number *2* in magnetic deflection picture tubes is to improve the focusing, or to maintain good focus when there are rather large variations of anode current and variations of spot brilliance.

Maximum values of brightness as shown on the graph of Fig. 51-12 is mnch greater than maximum average brightness utilized for picture reproduction. For illustration, assume that you have tuned a receiver to

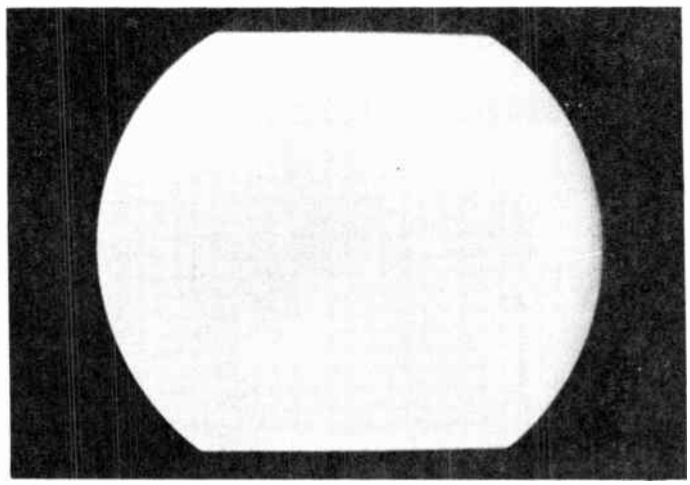


Fig. 51-14. The brightness of a raster seldom exceeds 25 foot-lamberts.

a channel in which there is no transmission at the moment, and have turned the brightness control nearly to maximum. The effect is as pictured in Fig. 51-14. Average brightness of the screen seldom would be much in excess of 25 foot-lamberts. Brilliance along the horizontal trace lines would be greater, for line

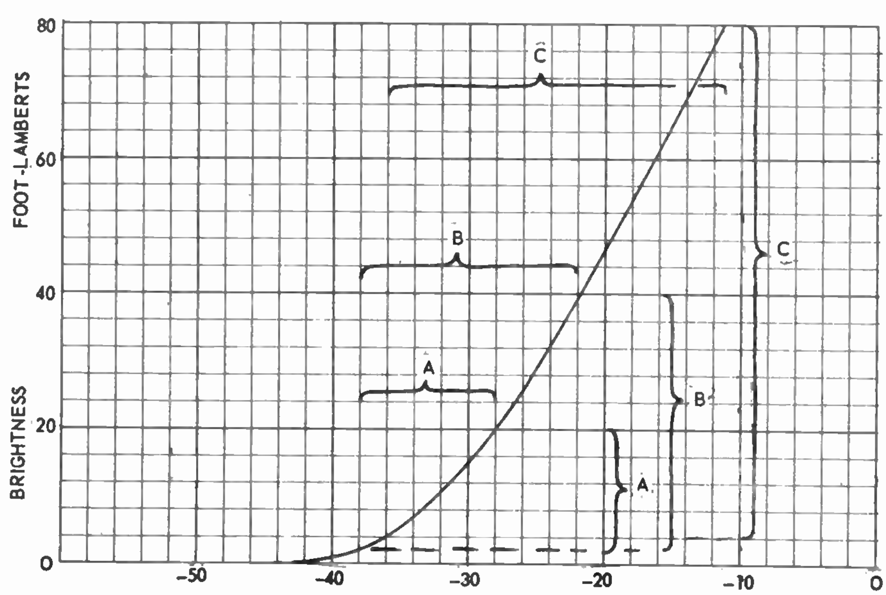


Fig. 51-15. Determination of contrast ratios and accompanying grid swings.

brilliance depends on intensity of light in the spot itself while average brightness would be reduced by the relatively dark horizontal spaces between successive lines being traced by the spot.

If you tune in a picture and make adjustments which allow a good range of shading, good contrast and definition, and enough brightness for viewing in a well lighted room, the average screen brightness may be on the order of 12 to 15 foot-lamberts. There is a reduction of average brightness because the picture contains many gray tones and nearly black tones in addition to "highlights" of maximum brightness.

Those with minimum brightness are merely a very dark gray. They are gray because of light dispersed from the electron spot to adjacent areas of the screen, because of some internal reflection, and because room light is reaching the screen material and is being reflected back to your eyes.

The degree of contrast, as it appears to viewers, is the ratio of maximum to minimum apparent brightness in various parts of the screen. The true contrast resulting from changes of intensity or rate of flow in the electron beam is the ratio of maximum to minimum brightness as shown along the curves of graphs which are being examined. This true contrast always is somewhat greater than visual or apparent contrast.

The contrast or brightness ratio may be as low as 15 or 20 to 1 and still give acceptable reproduction. Ratios increasing up to as much as 50 to 1 gives pictures more pleasing to most people. Fig. 51-15 illustrates some contrast ratios and the "swings" or picture variations of video signal voltage which produce the contrasts. Between points on the transfer characteristic curve enclosed by bracket *A* the maximum brightness is 20 foot-lamberts and the minimum is 2. The contrast ratio is 10 to 1. The minimum brightness results from –38 volts on the control grid, and maximum from –28 volts. This is a 10-volt grid swing, or the range of picture variations in the video signal is 10 volts from least to greatest.

Within bracket *B* of Fig. 51-15 the maximum brightness is 40 foot-lamberts and the minimum again is 2. The contrast ratio has been increased to 20 to 1. The voltage on the control grid now is changing between –38 and –22 volts, which is a grid swing of 16 volts. These contrast ratios and the grid swings that produce them illustrate something well known to experienced technicians – you always can have a brighter picture and better contrast with a strong video signal than with a weak one. True, it is possible to increase brightness and maintain correct contrast by readjustment of controls, but still you are working on the original video signal and the stronger this signal the easier it will be to get desired results.

At *C* in Fig. 51-15 the maximum brightness is 80 foot-lamberts and the minimum is 4. The contrast ratio is 20 to 1, the same as at *B*. Now the control grid voltage is –36 for minimum brightness and is –11 for maximum brightness. The grid swing is 25 volts, as caused by picture variations in the video signal. Although we have not changed the contrast ratio, the picture is of much greater average brightness. This has been made possible by the greater grid swing or by the still stronger video signal.

A grid swing of 25 volts may seem rather small for a bright picture of full contrast. But we must remember that picture variations can occupy no more than 60 percent of the signal voltage range from zero to the tips of sync pulses. So a picture voltage of 25 requires an overall rectified voltage at the video amplifier output of almost 42 volts.

TELEVISION

TEAR — CUT ALONG THIS L

LESSON NO. 52

THE SYNC SECTION

Training Your Mind

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON NO. 52

THE SYNC SECTION

The signal which enters the television receiver at the tuner has been followed all the way through to the grid-cathode circuit of the picture tube. In all the parts and circuits which have been examined only three things have been done to the original signal. It has been amplified at many points, all along the line. The modulated carrier frequency has been changed to a modulated intermediate frequency. The intermediate frequency has been demodulated or "detected" to obtain the composite television signal. This composite signal is applied to the picture tube.

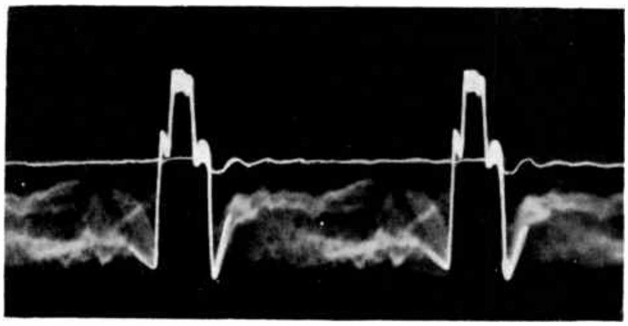
The composite signal at the picture tube is exactly like the modulation envelope of the original carrier. The carrier frequency and intermediate frequency have merely been means for transferring the composite signal through space and through the i-f amplifiers of the receiver, until this signal is recovered at the video detector. The "waveform" of the composite signal at the picture tube is unchanged from that existing at the antenna and tuner.

Now we are ready to follow the television signal through the sync section of the receiver. This section, as you know, extends from a takeoff point at the video detector or amplifier through to the sweep oscillators. In this portion of the receiver the entire character of the signal will be altered. The only resemblance between voltages or signals at the input and output of the sync section is that the same horizontal and vertical sync frequencies are present at both places.

We shall observe how the waveform of the signal is altered at various points in the sync section with the help of an oscilloscope. The luminous trace on the screen of an oscilloscope tube shows variations of a voltage, just as the trace on the screen of a picture tube shows the outlines of a picture. There is no essential difference between the cathode-ray tube of the oscilloscope and an electrostatic deflection picture tube in a television receiver. We shall use the oscilloscope in the same way that it would be used by a television technician during service operations.

Many times you have looked at drawings of television signal waveforms similar to the one at the left in Fig. 52-1. Here are represented two horizontal blanking intervals, the pedestals, the black level, the sync pulses, and the picture signal traces between the blanking intervals. On the screen of the oscilloscope the same parts of a signal appear as at the right. The irregular faint tracings between the blanking intervals are caused by varying signal voltages that are forming lights and shadows on the picture tube of the receiver. Most of these tracings are near the black level, indicating that a picture of generally dark tone was being reproduced while making this photograph.

When the signal is for a picture in which there is movement the picture tracings shown by the oscillo-



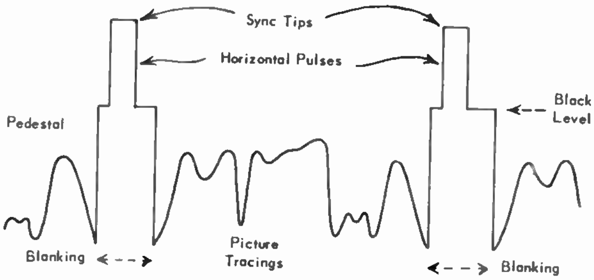


Fig. 52-1. Horizontal sync pulses and blanking as they appear in a drawing and on the oscilloscope screen.

scope between blanking intervals will weave and shift with every movement of the image. This accounts for the cloudy effect in the photograph. If the signal is for a stationary test pattern the tracings in the picture space remain stationary.

You will recall, from descriptions of sync sections in earlier lessons, that tubes in these sections are used for a number of purposes which are quite different from anything found in other parts of the television receiver or in sound receivers. The functions served by tubes and circuits in the sync section may be summarized as follows.

1. Remove from the composite television signal the picture voltages, leaving only the horizontal and vertical sync pulses. This is done by sync separator tubes.

2. Make these sync pulses of uniform strength, and remove irregularities which might result from noise voltages or other interference. Tubes called clippers or limiters do this part of the work.

3. Separate the horizontal from the vertical pulses, and produce from each kind of pulse a voltage at the corresponding horizontal or vertical frequency. This is accomplished by suitable filter circuits, not by tubes.

4. Make the strength of these final output voltages independent of ordinary variations in strength of received signals and of changes in setting of the contrast control. The filter circuits perform most of this function.

5. Amplify the signal or the sync pulses wherever it may be necessary for correctly carrying out the other functions. This is the work of sync amplifiers.

6. Invert the signal or the sync pulses in case this is necessary for producing final output voltages of whichever polarity, positive or negative, may be required for application to the sweep oscillators. Any tube used especially for this purpose is called a sync inverter.

There is no standardization of names of tubes used for the several purposes. Names mentioned in the

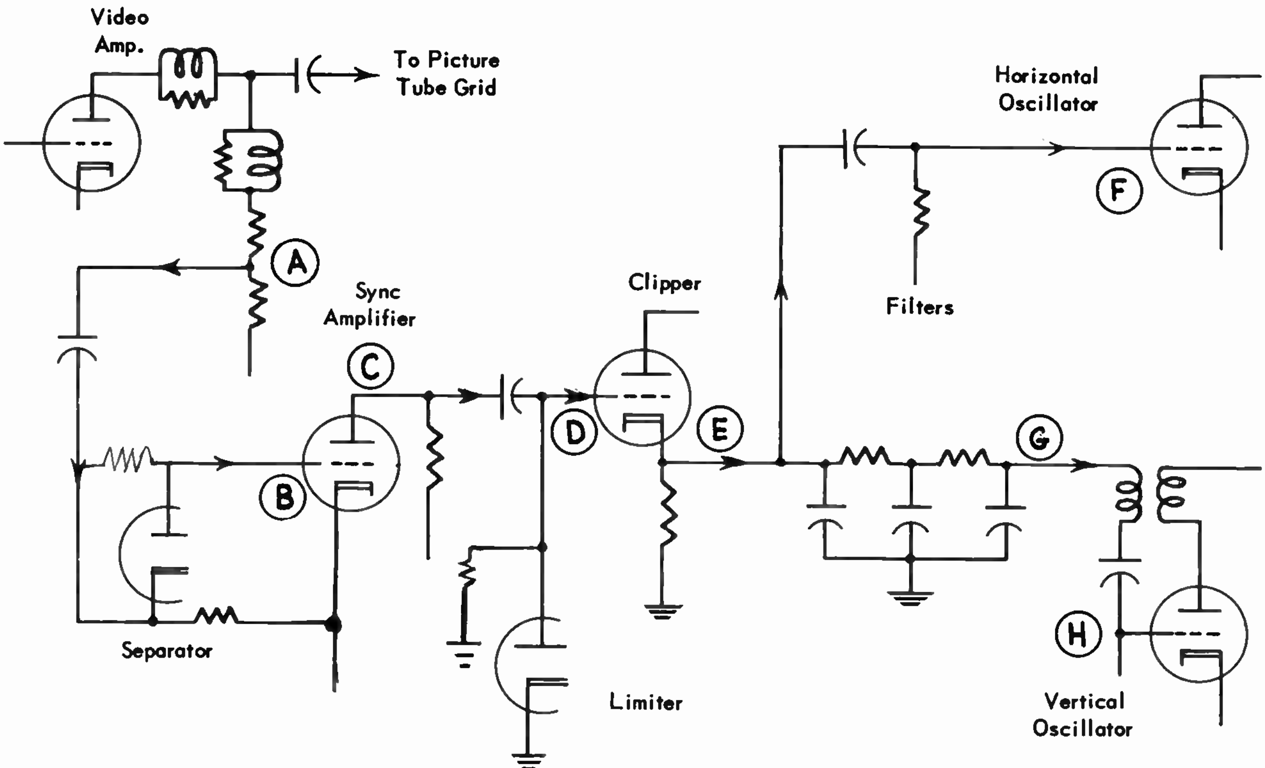


Fig. 52-2. Principal circuit connections of the sync section whose performance is to be observed.

preceding summary are in general use, but some manufacturers deviate from this practice. For example, what we are calling a separator may be called a stripper. Although separate tubes sometimes are used for each of the functions, it is more common practice in modern receivers to operate tubes in a way that allows handling more than one of the jobs. For instance, a single tube may both separate and amplify, or it may amplify and clip, or it may invert while doing almost anything else. Twin diodes and twin triodes are in common use.

To illustrate the changes undergone by signal voltages in passing through a typical sync section we shall follow the path that is "mapped" on the diagram of Fig. 52-2. Here are shown the signal-carrying connections in one particular receiver which uses a magnetic deflection picture tube. The first tube in the sync section is a diode used as a separator. It is followed by a sync amplifier, then a limiter, and then a clipper tube. Between the clipper and the sweep oscillators, horizontal and vertical, are filter circuits.

To begin with we shall see what happens to the horizontal sync pulses. Later we shall follow the vertical pulses in similar fashion. The entire composite signal is taken from point A between two resistors that form the plate load for the video amplifier. Since the output of this amplifier is shown as

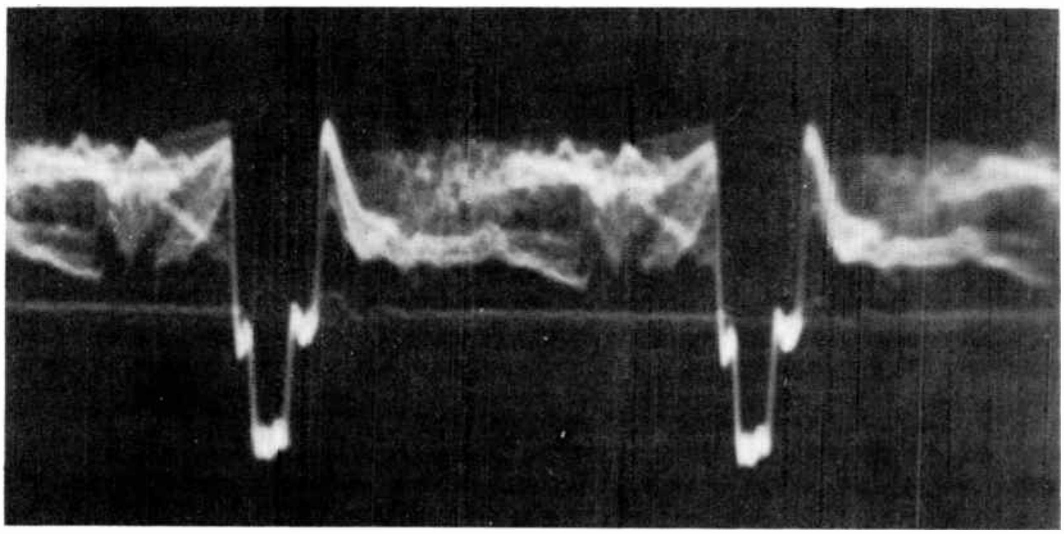


Fig. 52-3. Positive picture voltages and negative sync pulses in the composite signal.

going to the grid of the picture tube, picture voltages must be positive and sync pulses negative. The oscilloscope connected to takeoff point A shows this polarity of the composite signal as in Fig. 52-3. Except for inversion of polarity this signal is like the one at the right in Fig. 52-1.

The composite signal taken from the video amplifier outpnt goes through a capacitor to the cathode of the separator diode. This diode is conductive when negative alternations of the composite signal act on the cathode. The sync pulses are part of the negative alternations. Then the output from the plate of the separator consists of negative voltage pulses. The oscilloscope shows these negative pulses as in Fig. 52-4 when the instrument is connected to point B of the diagram, or to the separator plate and sync amplifier grid. One-way conductivity in the separator diode has removed most of the positive picture voltages but has preserved the negative voltage pulses whose frequency is the same as that of the horizontal sync pulses. These negative pulses form the input voltage for the sync amplifier grid.

Pulse voltage at point B of Fig. 52-2 is much less than peak-to-peak voltage of the composite signal at A. This is because we have lost much of the picture portion of the composite signal, and because there are losses in the separator diode and in its circuit connections not shown by the diagram. Peak voltage at B, the sync amplifier grid, is only about one-fourth of the peak-to-peak voltage at A.

Our next step is to increase the pulse voltage by means of the sync amplifier tube. The oscilloscope shows pulse voltages at the plate of the sync amplifier, point C on the diagram, as in Fig. 52-5. These voltage pulses are actually about six times as strong as at the amplifier grid. To bring the stronger voltages down to a height suitable for viewing and photographing the gain control of the oscilloscope has been turned down.

There is no material change in the waveform of voltage pulses at the amplifier plate and grid. There

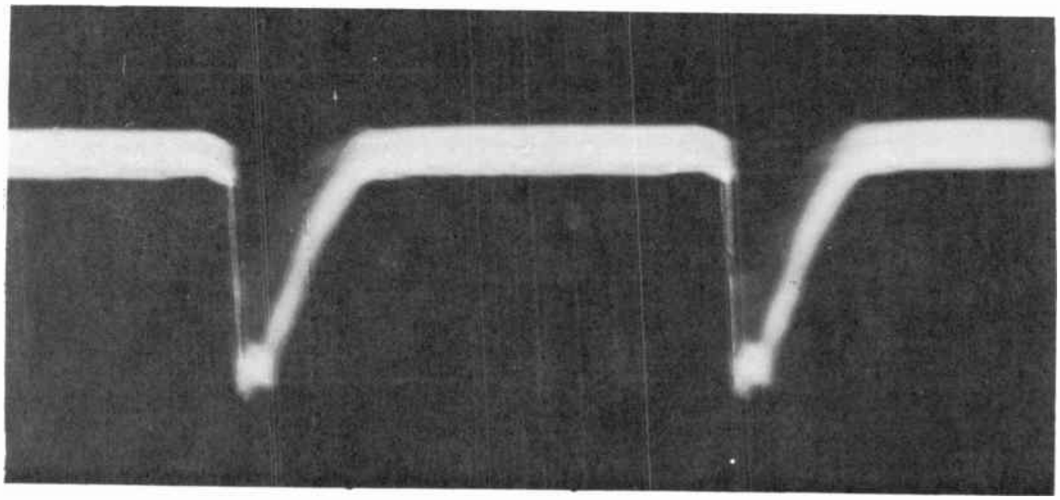


Fig. 52-4. Horizontal pulses at the separator plate and sync amplifier grid.

has, however, been an inversion of polarity. Now we have positive pulses at the plate (Fig. 52-5) whereas the pulses are negative at the grid (Fig. 52-4).

The positive voltage pulses from the plate of the sync amplifier pass through a coupling and blocking capacitor to the grid of the clipper, point D in Fig. 52-2. To the grid of the clipper is connected the plate of the diode limiter. Consequently, the plate of this diode is at the negative voltage which is the bias voltage for the grid of the clipper tube. The diode limiter will not conduct unless its plate is made more positive than its cathode. Then to have conduction in this limiter the peaks of the positive voltage pulses (Fig. 52-5) must exceed the negative voltage which is on the limiter plate.

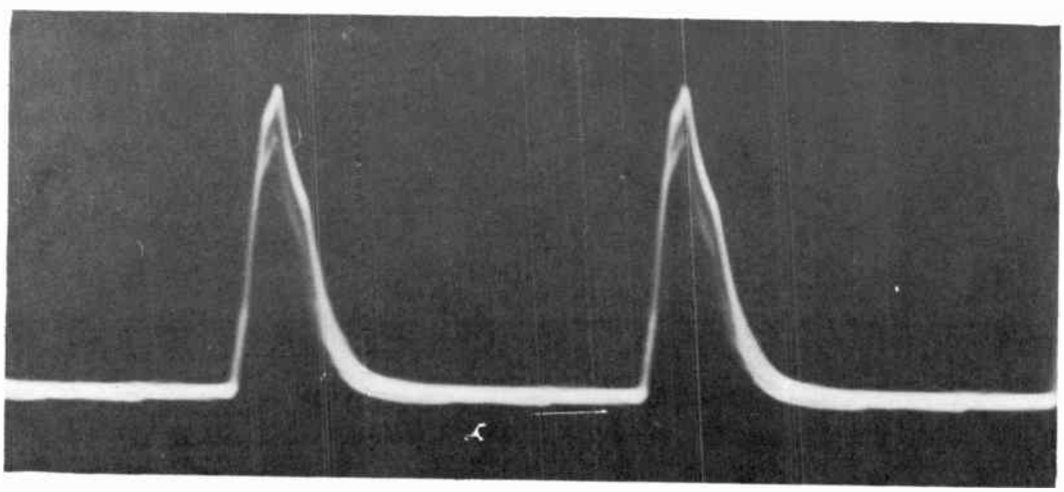


Fig. 52-5. Horizontal pulses at the sync amplifier plate.

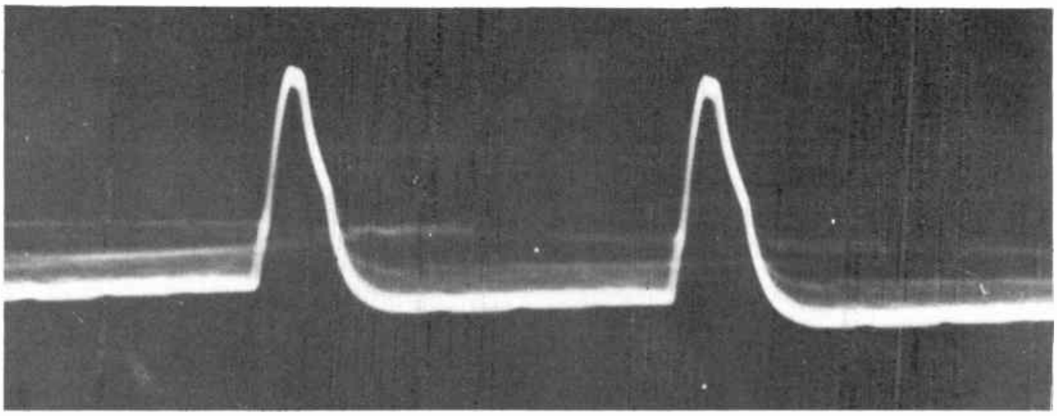


Fig. 52-6. Horizontal pulses at the clipper grid or limiter plate.

If the peaks of the positive voltage pulses do exceed the negative plate voltage the resulting conduction through the limiter to ground will carry off the excess pulse voltage. Thus the limiter diode acts to limit the maximum positive voltage of the pulses to a value which is equal to the negative bias on the clipper grid. Any sudden pulses of high voltage, such as might result from noise or other interference, are absorbed by the limiter action. A tube used in this general manner sometimes is called a noise limiter.

When the oscilloscope is connected to point D of our circuit diagram, at the grid of the clipper, we have the voltage trace pictured by Fig. 52-6. The waveform is much the same as at point C (Fig. 52-5) except for some flattening of the pulse tips, this being due to action of the limiter. The actual peak voltage in Fig. 52-6 is about one-third of the peak voltage at the plate of the preceding amplifier tube.

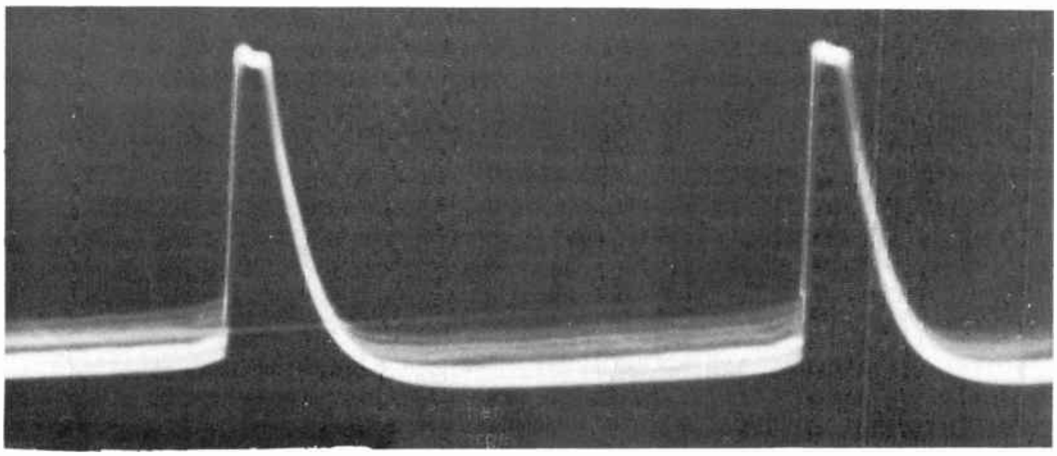


Fig. 52-7. Horizontal pulses applied to the filters from the clipper cathode.

If you look back at the diagram of Fig. 52-2 you will see that the clipper tube is operated as a cathode follower, with output voltage taken from the cathode rather than from the plate. The connection from the clipper to the two filters is taken from the high side of the cathode resistor, at point E. Fig. 52-7 shows the voltage waveform at point E, as observed on the oscilloscope. Since there is no inversion of polarity between grid and cathode the voltage pulses at the cathode are positive, just as at the grid. There is little change of waveform between grid and cathode. A cathode follower provides little if any gain, so peak voltage at the cathode output is about the same as at the grid input to this tube.

The voltage pulses shown by Fig. 52-7 are applied to the two filters, one of which leads to the horizontal sweep oscillator and the other to the vertical sweep oscillator. The design and operation of the filters are much the same in nearly all receivers, but there is great variety in the sync circuits leading up to the filters, or in the circuits between the sync takeoff and the filters. Therefore, before discussing the action of the filters we shall go all the way back to the video amplifier system and examine some designs which differ from that of Fig. 52-2.

SYNC SEPARATORS AND TAKEOFFS. The composite television signal which is delivered to the sync section may be taken not only from the output of a video amplifier, but also from any other point at which this signal exists. Fig. 52-8 shows one way of taking this signal from the output of the video detector. Sync pulses are positive and picture voltages are negative at this takeoff point. The composite signal is applied to the plate of a separator diode. The diode plate is negatively biased by the charge built up on

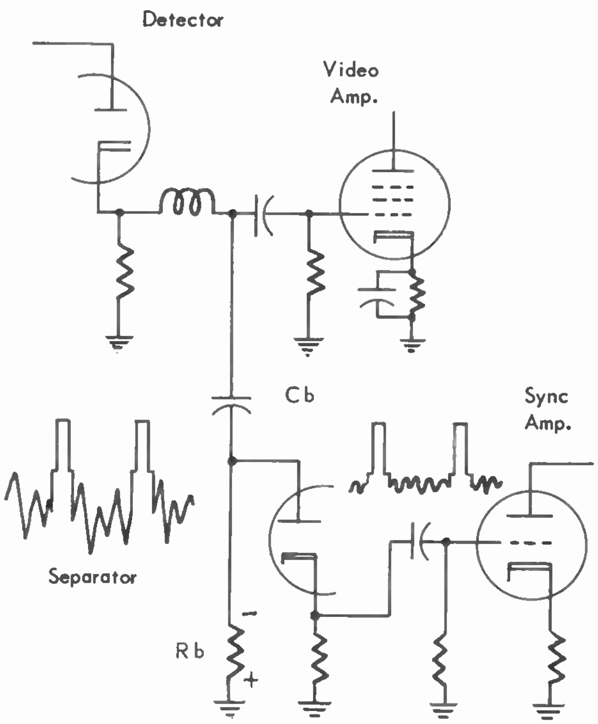


Fig. 52-8. Composite signal taken from video detector output.

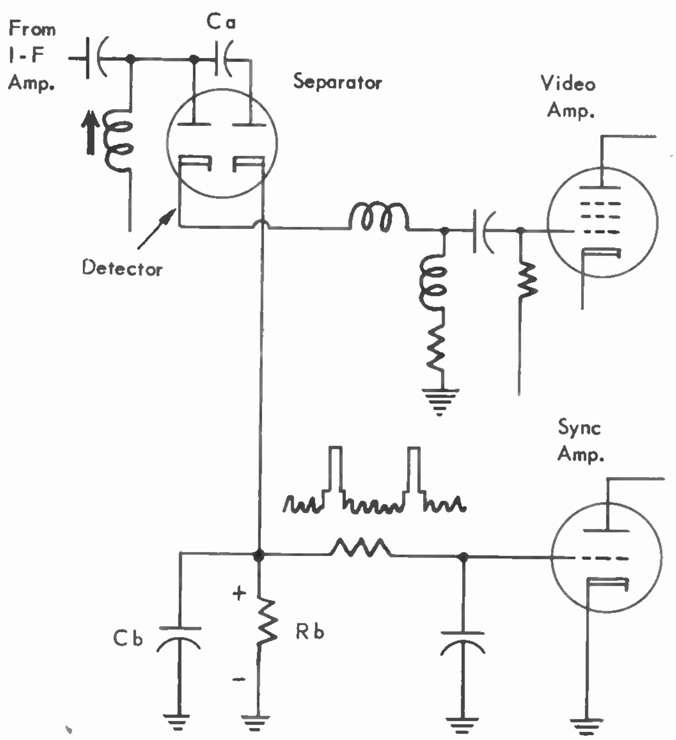


Fig. 52-9. Composite signal obtained from i-f amplifier.

capacitor Cb, from which there is slow leakage through resistor Rb. The separator conducts on only the more positive portions of signal alternations, and in the output from the separator cathode the sync pulses are much stronger than remaining picture voltages. This output is applied to the grid of the following sync amplifier triode.

In Fig. 52-9 the composite signal is secured from the output of the last i-f amplifier. The output of this i-f amplifier is applied to the plate of the detector diode directly, and is applied through capacitor Ca to the plate of the separator diode. The detector and separator are the two sections of a twin diode. The cathode of the separator diode is connected to the grid of the sync amplifier shown down below.

The cathode of the separator diode in Fig. 52-9 is positively biased by action of capacitor Cb and resistor Rb. You always can determine the polarity of a capacitor-resistor bias system by noting the direction in which electrons must flow through a resistor to the cathode of the diode or from the plate of the diode. This direction always is from the negative end to the positive end of the resistor. This electron flow through resistor Rb in Fig. 52-8 is from the diode plate downward, which means that the plate is connected to the negative end of Rb, and is negatively biased. In resistor Rb of Fig. 52-9 the electron flow is upward to the cathode of the separator diode, which means that the cathode is connected to the positive end of Rb, and is positively biased.

1 If the plate of a diode is negatively biased the positive alternations of a signal applied to the plate must be strong enough to exceed the bias voltage before the plate is made positive with reference to the cathode and before the diode can conduct. Of course, the same thing could be accomplished by applying to the cathode a signal whose negative alternations are stronger than the bias voltage on the plate. This would make the cathode so strongly negative that it would be negative with reference to the plate in spite of the bias voltage on the plate, and there would be conduction.

If the cathode of a diode is positively biased the negative alternations of a signal applied to the cathode must be strong enough to overcome the bias before there can be conduction. Or, if positive alternations of a signal applied to the plate are stronger than the positive bias on the cathode the plate will be made positive with reference to the cathode, and there will be conduction. Whenever we have a biased diode the applied signal must be strong enough to make the plate positive with reference to the cathode, or to make the cathode negative with reference to the plate, before there can be conduction through the diode.

Were the biasing voltage on a diode to be almost as great as peak voltage of the applied signal the diode would be made conductive only by the tips of the sync pulses; all the rest of the applied signal would be cut off. If this biasing voltage were maintained at constant strength while signal strength decreased, not even the tips of the sync pulses would cause conduction. The entire signal would be stopped and nothing would be passed into the sync section. Were the applied signal to increase in strength, with constant biasing voltage, the diode would pass all of the sync pulses and too much of the picture voltages for effective separator action. Naturally, both of these effects would be undesirable.

5 In order that voltages passed into the sync section always may be the same portion of the composite signal (essentially the entire strength of the sync pulses) the bias on the separator diode must vary in accordance with signal strength. The bias must increase for stronger signals, and decrease for weaker ones. This is accomplished automatically by using capacitor-resistor biasing, which acts just like grid-

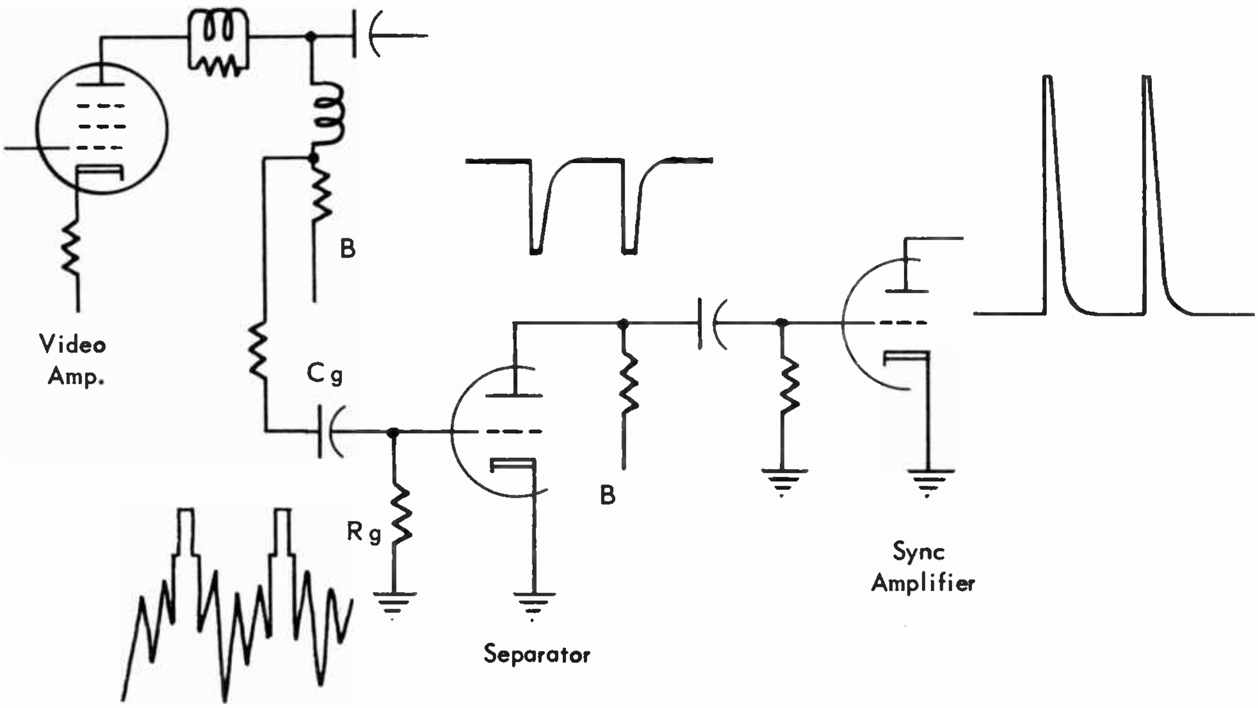


Fig. 52-10. Using a triode as the sync separator.

leak biasing. The biasing capacitor, Cb in Figs. 52-8 and 52-9, is charged to a voltage proportional to strength of applied signal. This voltage is maintained by slow leakage of the charge through resistor Rb in these diagrams.

Now we shall leave the diode type of sync separator and turn our attention to the manner in which triodes and pentodes are used in taking sync voltages from a composite signal in the output of a video amplifier or video detector.

Fig. 52-10 shows how the composite television signal taken from the top of the load resistor for a video amplifier may be applied to the grid of a triode used as a sync separator. The separator is operated with grid-leak bias provided by capacitor Cg and resistor Rg. The time constant of this capacitor-resistor combination is such that the biasing voltage on the separator grid will cause plate current cutoff at or slightly above the black level of the applied signal.

Such biasing and its effect are illustrated by Fig. 52-11. Conduction results only from the sync pulses, which are more positive than the black level of the applied signal, and only these pulses appear in the output of the separator. In order to have cutoff at the valve of negative biasing voltage obtained from the composite signal the separator is operated with moderate plate voltage. The plate voltage may be something like 30 to 50 volts, or even less, for a tube which could take 150 or more plate volts when used as an amplifier. A pentode separator tube may be used in a generally similar circuit. The pentode is operated with low voltages on both plate and screen. In addition to cutting off the picture voltages of the composite signal there may be some amplification of the sync pulses.

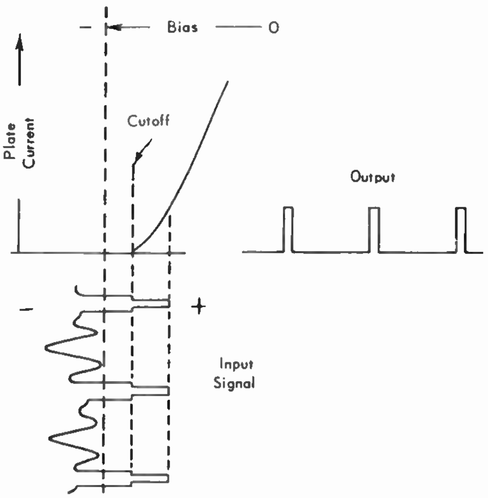


Fig. 52-11. The triode is biased for plate current cutoff at the black level.

Quite obviously, the composite signal must be taken from a point where sync pulses are positive and picture voltages negative, since the sync pulses must make the separator grid relatively more positive, or less negative, than the average bias. It would be desirable to have cutoff right at the black level, but to avoid rectifying some of the picture voltage the average cutoff may be somewhat above the black level. The more negative the average bias is maintained the smaller will be the fraction of the sync pulses appearing in the separator output, and the more amplification will be needed farther along in the sync section. The output of the separator in Fig. 52-10 goes to the grid of a following sync amplifier.

Grid-leak bias is used on the separator in order that the bias voltage and the plate cutoff may remain proportional to strength of the composite signal. This type of bias voltage varies directly with changes of signal strength, and will maintain cutoff at the same point in relation to the black level, or very nearly at the same point. An additional fixed minimum bias sometimes is used, in addition to grid-leak bias. The added negative bias may insure cutoff of all picture voltages and may lessen the effects on synchronization of noise voltages which get through to the sync section. Separator tubes nearly always are medium-mu triode voltage amplifiers or sharp cutoff pentodes, although semi-remote cutoff pentodes sometimes are found.

Instead of applying the composite signal to a separator, and applying the output of the separator to an amplifier, as in Fig. 52-10, the first sync tube may be an amplifier and the next one a separator. Then the composite signal is strengthened by the amplifier, much as in a video amplifier, and the picture voltages then are cut off by a following separator tube. The difference between a separator and an amplifier usually is only in the operating voltages. A separator works with low voltage on the plate or on both plate and screen, and with a bias highly negative in proportion to these voltages. An amplifier works with normally high plate and screen voltages, and with a bias which does not cause plate current cutoff anywhere on the composite signal waveform when this is the signal applied to the amplifier grid, or does not cause cutoff of a sync pulse signal which may be applied to the amplifier grid.

Sync amplifiers may serve several purposes. First, they provide amplification which is needed to bring

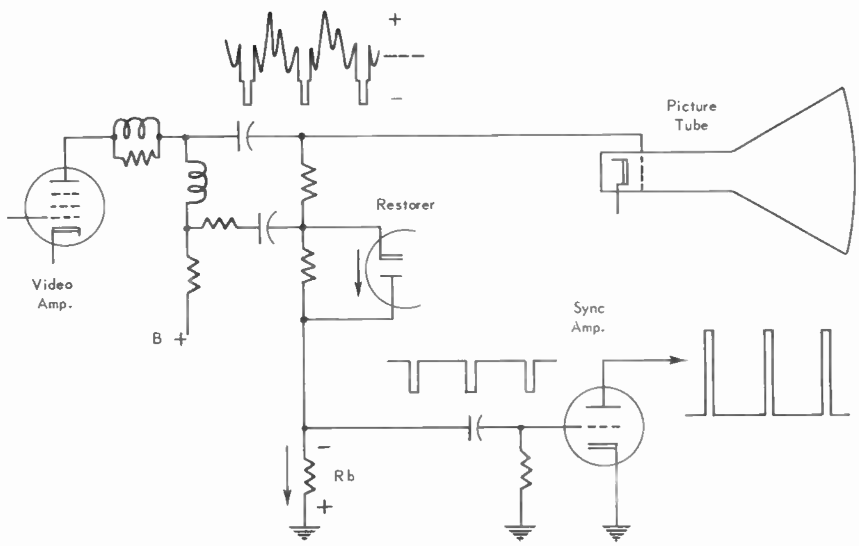


Fig. 52-12. Sync pulse voltages taken from a diode restorer tube.

signals up to necessary strength after the voltages have been reduced by separators or limiters. How much total amplification is required depends largely on the point from which the original composite signal is taken. If takeoff is from the final video amplifier the signal is strong, and needs little added amplification. If from a preceding video amplifier, or the output of the video detector, or the output of the last i-f stage, the composite signal is relatively weak and more total sync amplification will be needed.

Since the sync amplifier must have good gain at frequencies as low as those of the vertical sync pulses it ordinarily is designed with large grid coupling capacitances and with B-plus plate feed circuits which favor the low frequencies. These things were discussed in connection with video amplifiers. There is, however, no problem of high-frequency amplification, as with video amplifiers, for we do not wish to amplify the picture voltages in the sync section of the receiver. Sync amplifiers are designed and constructed much like resistance coupled audio voltage amplifiers.

So far as is practicable the output from a sync amplifier should remain fairly constant in voltage when there are considerable changes of strength in the composite signal. This will help to maintain steady synchronization of the sweep oscillators which follow the sweep section. Grid-leak bias automatically increases, and lowers the gain, on strong signals, while decreasing and raising the gain on weak signals. Consequently, grid-leak bias is much favored for sync amplifiers.

Now we shall examine still another way of obtaining voltage pulses for use in the sync section. This way is through the d-c restorer tube which is used in many receivers. One circuit in quite general use is shown by Fig. 52-12. The restorer diode conducts only when its cathode is made negative by negative alternations of the composite signal. The negative alternations include the sync pulses. The pulses of diode current flow as shown by arrows, downward through resistor Rb. This resistor is in the grid circuit of the sync amplifier, so negative pulse voltages form the input for this amplifier.

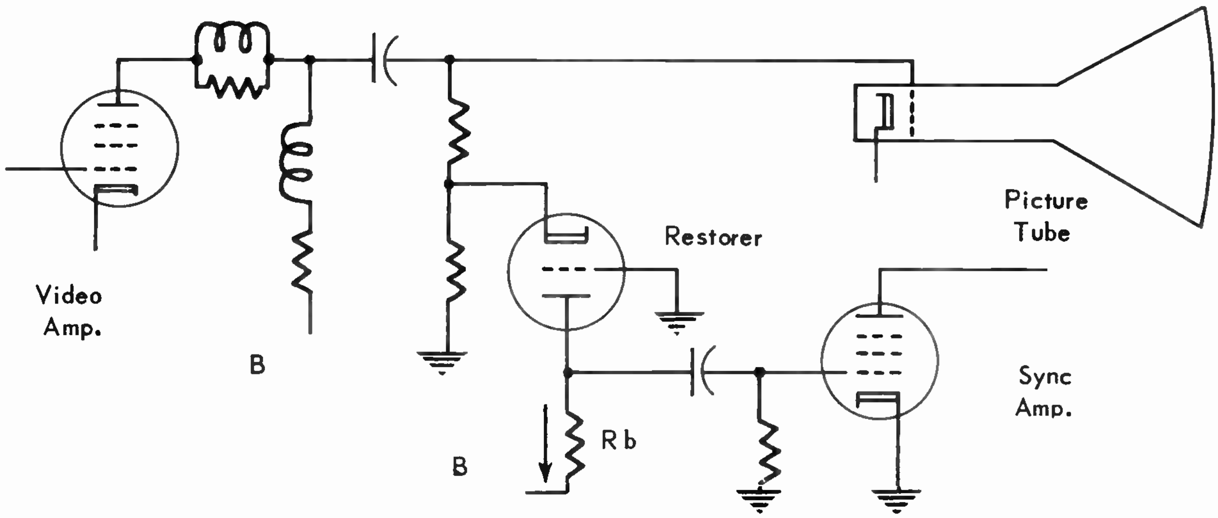


Fig. 52-13. Sync pulse voltages taken from the plate of a triode restorer.

Sync takeoff may be from the plate of a restorer used in any of the usual circuits, and regardless of whether the restorer is a diode, a triode, or a pentode. Fig. 52-13 shows the sync takeoff connection from the plate of a triode restorer. As you will remember, the cathode and grounded grid of the triode are used like a diode type of restorer. The triode plate is used only for sync takeoff. The plate is connected to the B-supply through resistor Rb, across which are produced sync voltage pulses each time the restorer tube conducts. These negative sync pulses are applied to the grid of the sync amplifier.

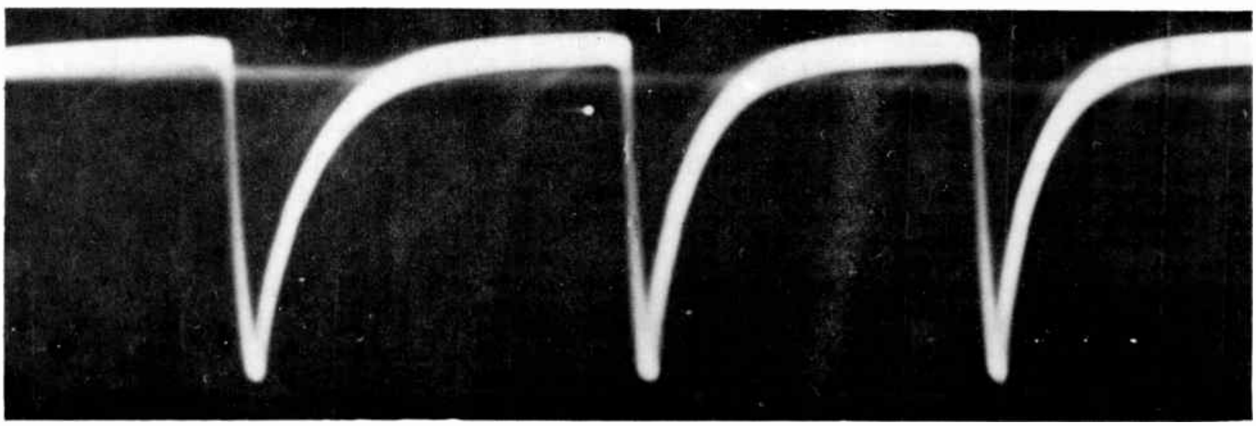


Fig. 52-14. Horizontal voltage wavetorm at the plate of a triode restorer.

Fig. 52-14 is a photograph of the horizontal sync pulse voltages as they appear on the screen of the oscilloscope with the instrument connected to the plate of a triode restorer. The voltage goes suddenly negative during charging of whatever capacitors are in the plate circuit of the restorer and the grid circuit of the sync amplifier. Then the voltage returns more slowly to zero as the capacitors discharge through resistors which are in these circuits. The waveform at the grid of the sync amplifier is the same as at the plate of the restorer. There will be inversion of polarity in the sync amplifier, making the voltage pulses positive in the output or plate circuit of this amplifier.

CLIPPERS AND LIMITERS. In Fig. 52-11 we examined the effect of biasing a triode or pentode in such a way as to cut off all the picture voltages from a composite signal, and leave only the sync pulses. This is the action which we have called sync separation. It also may be called clipping, since the sync pulse voltages are clipped from the composite signal.

When first discussing the functions of tubes shown in Fig. 52-2 it was explained how the limiter diode removes excessive positive voltage from sync pulses, or how this diode acts to bring all pulses down to approximately a uniform strength. We may think of this limiter diode as leveling the positive tips of the voltage pulses, while the clipper triode cuts off the pulses uniformly on the negative sides.

It is possible to operate a triode in a manner which clips the sync pulse voltages from the composite signal and at the same time limits the maximum voltage of pulses at the triode output. This is done by using plate current cutoff for the clipping action, and plate current saturation for limiting. Plate current cutoff resulting from a strongly negative bias is illustrated by Fig. 52-11. The steady rise of the transfer characteristic curve in that figure shows that plate voltage is high enough to cause continued rise of plate current with a grid voltage that is made less and less negative.

If we use a very low plate voltage the transfer characteristic curve will be changed to a form such as shown in Fig. 52-15. As grid voltage is made less and less negative there is first an accompanying increase of plate current. There is a corresponding increase of voltage drop or voltage loss in the plate load resistance, which leaves less and less voltage at the plate of the tube. Soon the voltage at the plate becomes so small that no more current will flow even though the grid voltage comes almost to zero. This is the action called plate current saturation or plate current limiting.

When the triode tube is operated with a grid-leak bias or a combination of fixed and grid-leak biasing which causes cutoff at or near the black level, and is operated with a very low plate voltage at the same time, we may have both clipping and limiting in the one tube. In Fig. 52-15 the three sync pulses shown as input are of unequal strength. Plate current saturation or limiting brings all three to the same strength in the output. The action is not quite so perfect as shown by the diagram, but we do have satisfactory clipping on the negative side and have a great deal of limiting action for the positive side of the voltage pulses.

To have limiting action, the plate voltage on a triode, or plate screen voltages on a pentode, may be 10 volts or even less. Such low voltages cause only small plate currents even when no signals are coming to the tube, and when the grid-leak bias is zero or nearly zero. Pulse currents in the output of a combination clipper-limiter are rather small. If, however, there is high resistance in the plate load the pulse voltages produced across this load may have peak values of 20 to 30 volts or even more. In some receivers using a pentode clipper-limiter tube the pulse voltage in the plate circuit approaches 100 volts peak-to-peak.

VERTICAL SYNC PULSES. In Figs. 52-3 to 52-7 we used the oscilloscope to observe horizontal sync pulses and resulting voltages as they appeared in circuits all the way from the video amplifier through to the input for the filters of a particular sync section. Now we shall use the oscilloscope to look at the vertical sync pulses and resulting voltages in the same circuits.

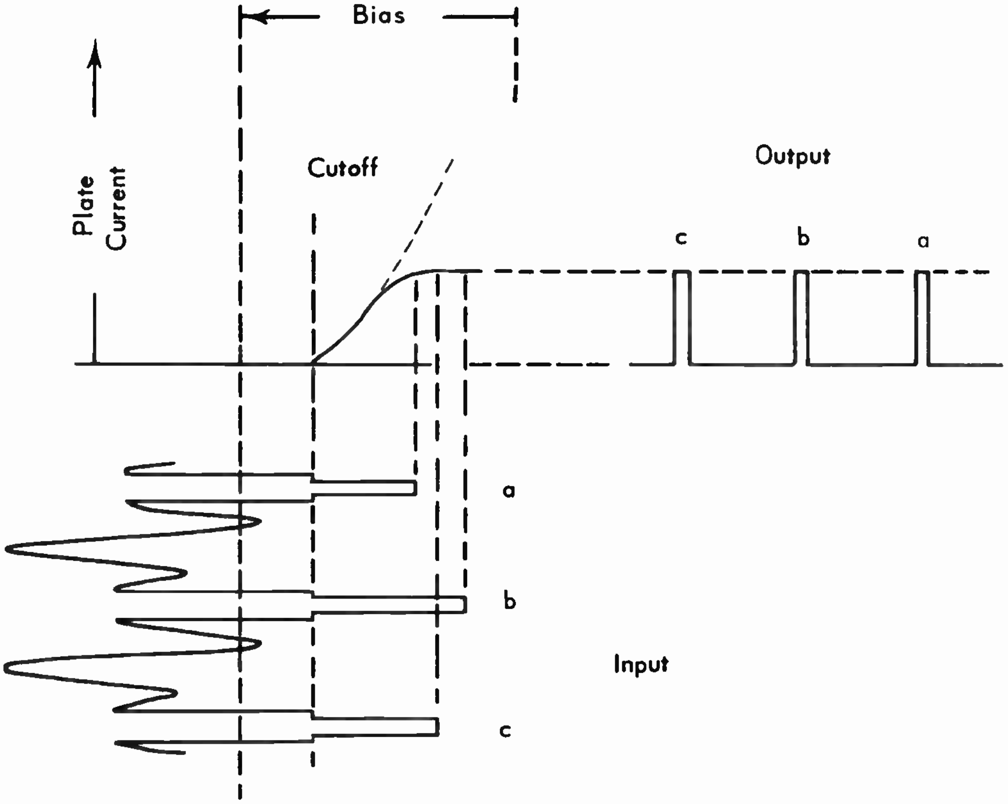


Fig. 52-15. Plate current saturation or limiting and plate current cutoff in the same tube.

First it may be interesting to know how the oscilloscope may be made to show either horioz ntal or vertical pulses and their effects. In the case of horizontal pulses, we wish to look at something which is occuring at the rate of 15,750 times per second. For such an observation the electron beam in the cathode-ray tube of the oscilloscope may be swept across its screen from left to right at the same rate, 15,750 times per second. At the same time the beam is being deflected vertically by changes of the voltage which we wish to observe. Then, while the beam is moving across the screen, the luminous trace will rise and fall with increases and decreases of observed voltages, as the changes occur during this period of time.

If the oscilloscope beam is swept from left to right only half as fast, or at the rate of 7,875 times per second, each crosswise travel will take twice as long as before. In twice the time we shall see two horizontal pulses or their effects. To look at three pulses at one time, crosswise of the oscilloscope screen, we would cause the beam to travel only one-third as fast, or to take three times as long in moving from left to right.

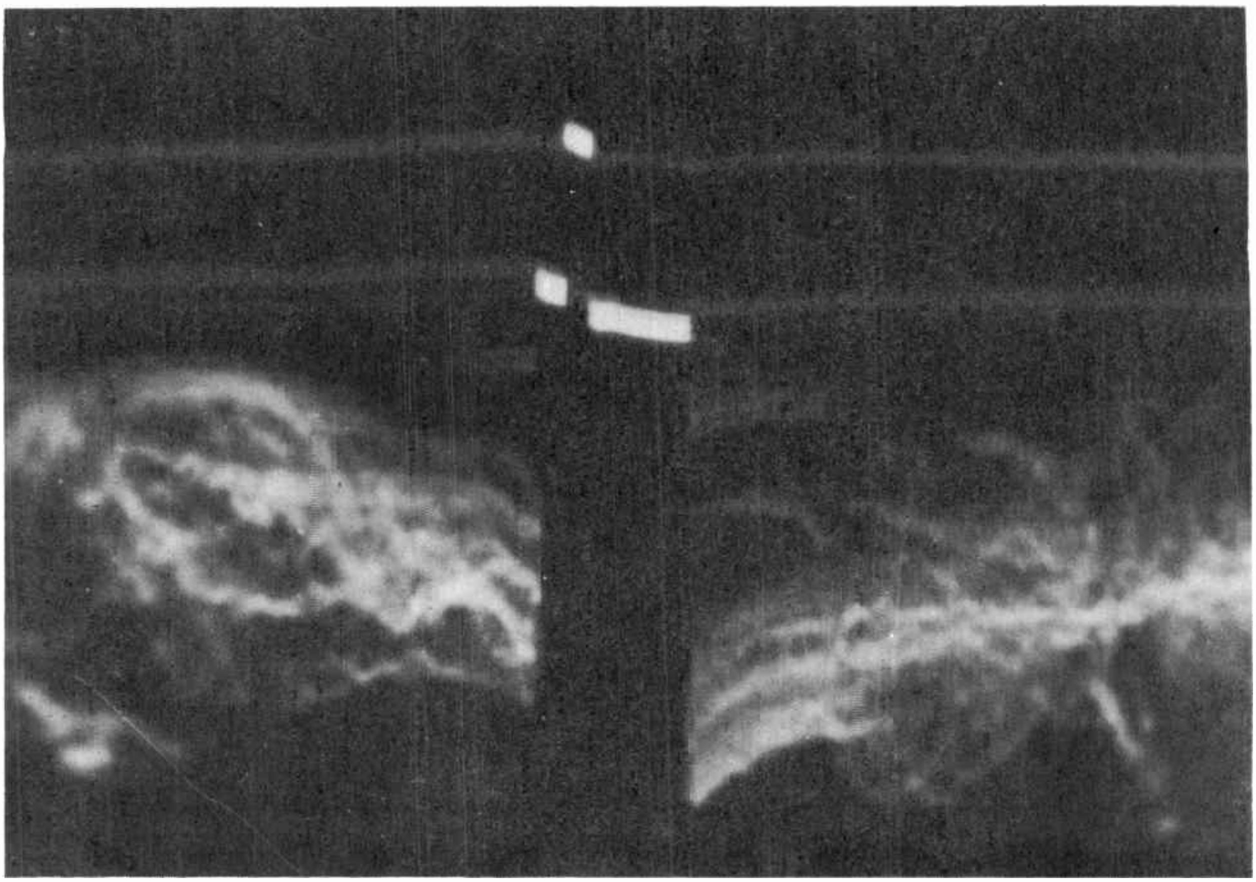


Fig. 52-16. Sync pulse and equalizing pulse voltages during a vertical blanking interval.

To observe only one vertical sync pulse or any resulting voltage we must sweep the oscilloscope beam from left to right only 60 times per second, because this is the rate at which vertical pulses recur. To look at two vertical pulses or their effect, at one time, we would sweep the beam only half as fast, making it cross the screen only 30 times per second. In each 1/30 second there are two vertical pulses. Now that you know how oscilloscope photographs are made we shall go back to the output of the video amplifier of Fig. 52-2 and see how the vertical pulses appear.

In Fig. 52-16 we are looking at part of a composite signal in the output of a video amplifier. Sync pulses are positive and picture voltages negative, this polarity being chosen because it corresponds to that of diagrams with which we have become familiar in many earlier lessons.

The cloudy effect at the left is caused by picture voltages in approximately the last 70 lines of a field. Then the picture space becomes dark during the vertical blanking interval. Toward the right the cloudy effect reappears. It is caused by the first 70 odd lines of the next field.

During the first part of the blanking interval the voltage rises as shown by the short bright space. This

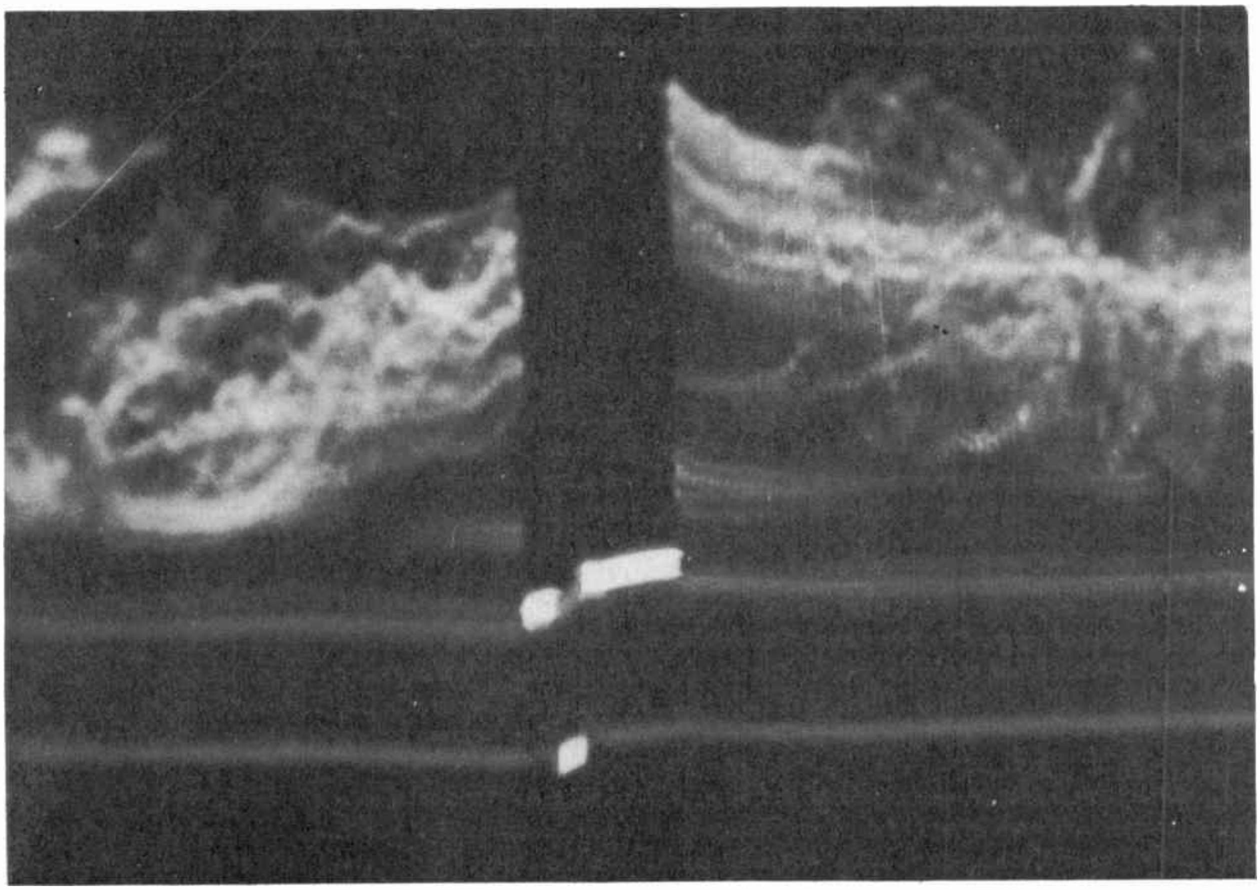


Fig. 52-17. Positive picture voltages and a negative vertical sync pulse.

is the voltage of the horizontal sync pulses and the equalizing pulses that occur ahead of the vertical sync pulses. Then comes a much higher voltage, shown by the highest bright space. This is the voltage resulting from the serrated vertical sync pulses. Next the voltage drops down and continues along the bright space during the remainder of the blanking interval, as a result of the equalizing pulses and horizontal sync pulses which continue until the start of the next field. Here we are looking, in actual fact, at things which previously have been seen only as diagrams or drawings.

You will remember that, in the output of the video amplifier of Fig. 52-2, point "A" the sync pulses actually are negative and picture voltages positive. This output polarity is shown by Fig. 52-17. This figure, which shows vertical synchronization voltage, is taken from the same point as Fig. 52-3, which shows horizontal synchronization voltage. It is evident that the two photographs were not taken with the same picture being received, for in one case the picture voltages are close to the black level(a dark toned picture) while in the other case the picture voltages extend far from the black level (a light toned picture). The time for the vertical blanking interval of Fig. 52-17 is about 1,250 millionths of one second. The time for each of the horizontal blanking intervals of Fig. 52-3 is about 11 millionths of one second.

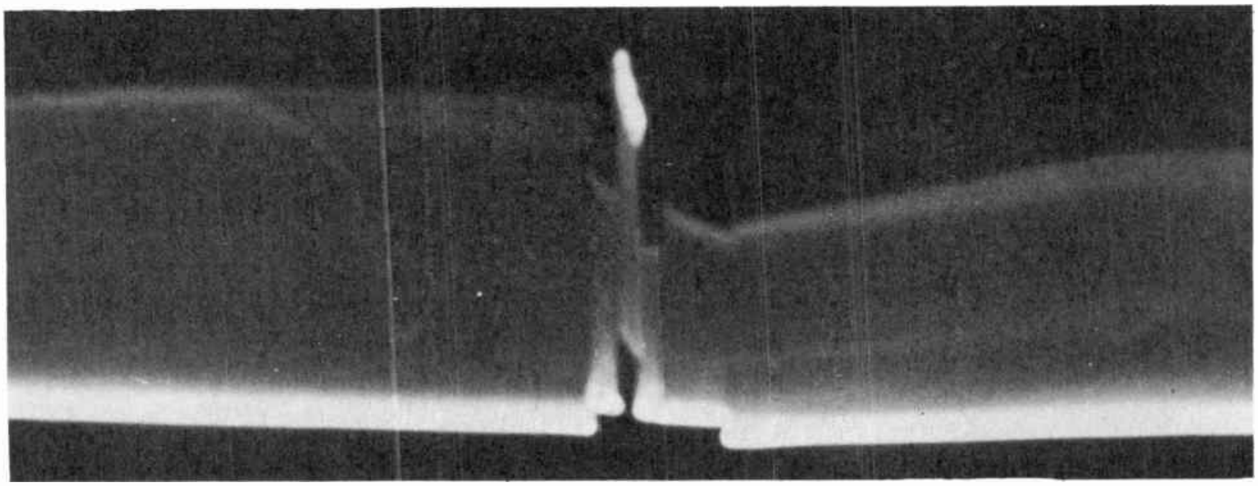


Fig. 52-18. A vertical pulse at the sync amplifier plate.

When watching oscilloscope traces for horizontal synchronization we went next to the grid of the sync amplifier, point B of Fig. 52-2. But since input and output waveforms at this amplifier are practically alike, except for polarity inversion, we shall skip this point for vertical observations and go to the amplifier output at point C on the circuit diagram. Here we obtain the vertical waveform shown by Fig. 52-18. The voltage peak produced by the serrated vertical pulses now is much higher than the voltages produced by horizontal and equalizing pulses during other parts of the vertical blanking interval. This vertical blanking interval still shows quite distinctly along the lower edge of the waveform. Keep in mind that this vertical waveform is taken from the same point as the horizontal waveforms shown by Fig. 52-5. The gray tone which shows before and after the vertical blanking interval results from great numbers of the horizontal voltage pulses pictured by Fig. 52-5.

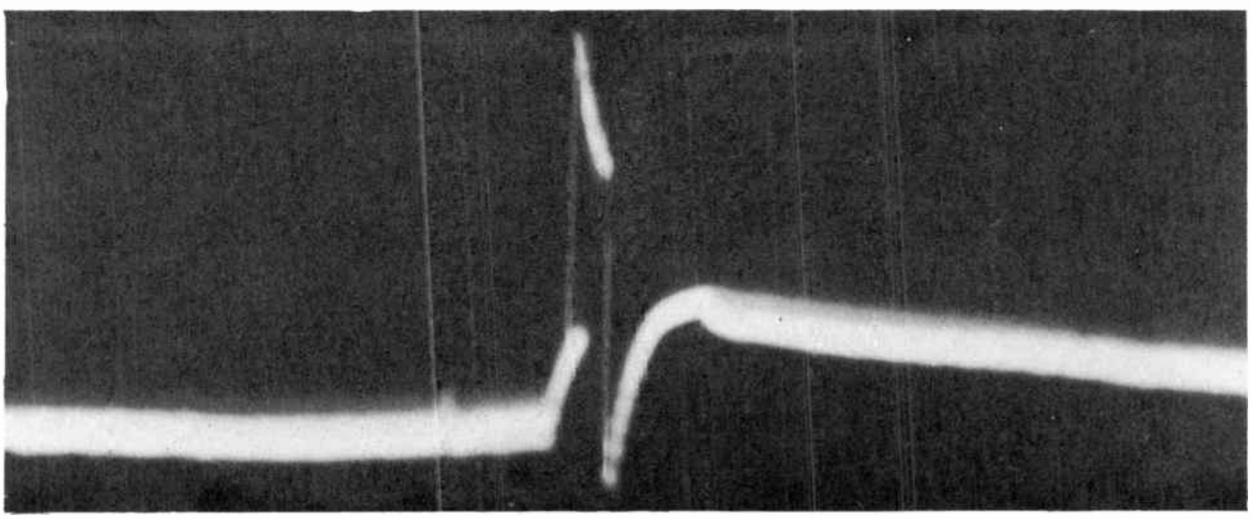


Fig. 52-19. A vertical pulse at the clipper grid or limiter plate.

At the clipper grid or limiter plate, point D of Fig. 52-2, we obtain the vertical waveform shown by Fig. 52-19. Now we are getting a strong voltage pulse from the serrated vertical pulses of the signal, and have relatively little voltage variation due to sync pulses on either side of the vertical pulse. This vertical pulse voltage is taken from the same point as the horizontal pulse voltages of Fig. 52-6.

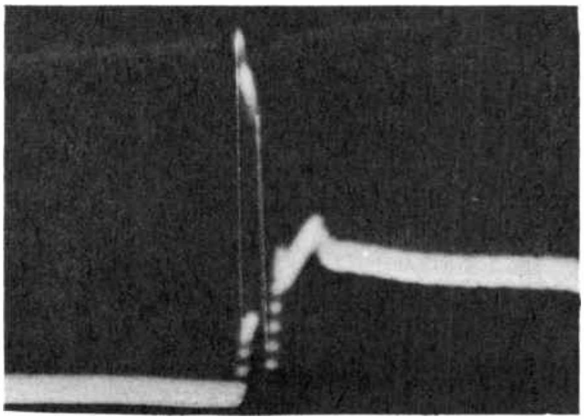


Fig. 52-20. A vertical pulse applied to the filters from the clipper cathode.

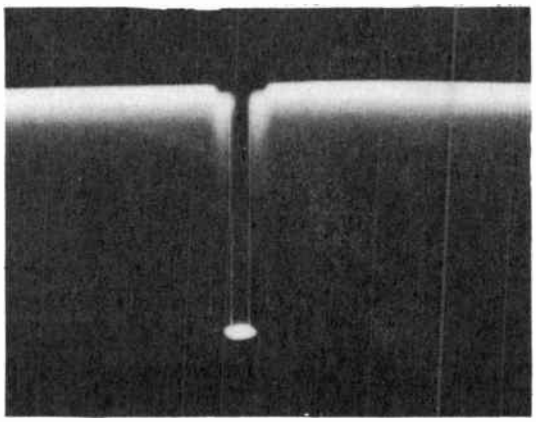


Fig. 52-21. A vertical pulse at the plate of a triode restorer.

When looking at horizontal pulse voltages we proceeded from the clipper input at its grid to the output at its cathode. There was no inversion of polarity and but little change of waveform between these two points, as you will see by looking back at Figs. 52-6 and 52-7. When we observe the vertical sync voltages at the same two points there again will be no inversion of polarity and little change of waveform. The vertical voltage at the clipper cathode is pictured by Fig. 52-20. It is almost the same as at the clipper grid (Fig. 52-19).

Back in Fig. 52-14 we looked at the voltage waveform appearing at the plate of a triode restorer tube used as the point of sync takeoff. The vertical synchronization voltage at the same point is shown by the oscilloscope as in Fig. 52-21. This voltage is applied to the grid of the sync amplifier in the diagram of Fig. 52-13. The serrated vertical sync pulses of the signal are causing a strong negative voltage about half way through the vertical blanking interval.

We have followed the horizontal and vertical sync pulses and resulting sync voltages by means of oscilloscope traces because you will find this method generally used in the service instruction manuals of receiver manufacturers. As a rule there will be a series of waveform photographs commencing at the point of sync takeoff and continuing all the way through the sync section and the following sweep section of the particular receiver covered by the instructions. Usually the horizontal and vertical waveforms taken at any one point are shown together, since it is convenient to leave the oscilloscope connection in place while changing the sweep rate to show these two waveforms.

After you become familiar with the meanings of waveforms they form a rapid means for locating troubles in the sync and sweep sections of television receivers. Of course, you do not have to use an oscilloscope for such trouble shooting. Voltage measurements, resistance measurements, and the appearances of pictures or patterns on the picture tube can be used for the same class of work. But, inasmuch as oscilloscope traces illustrate in an easily understood way what really is happening, we shall continue to look at them as we work on toward the deflection systems for the picture tube.

TELEVISION

LESSON NO. 53

THE SWEEP OSCILLATORS

Training Your Mind ...
C. T. I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind
C.T.I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON NO. 53

THE SWEEP OSCILLATORS

In the preceding lesson we took a composite television signal from the video section of the receiver, put this signal into the sync section, and at the output of the last tube in the sync section obtained voltage pulses at the same frequencies as those of the original sync pulses. We ended our observations at the input to the filters. Although these filters often are considered to be parts of the sync section, their purpose is to change the rather long pulses of sync voltage into sharp peaks of short duration, such as are required for timing or synchronizing the sweep oscillators. It is for this reason that we have reserved an examination of the filters until we are ready to study the horizontal and vertical sweep oscillators.

The filter whose output times, or synchronizes, the horizontal sweep oscillator may be called a differentiating filter or differentiating network, because it differentiates or distinguishes between the horizontal and vertical pulses. The filter whose output times the vertical sweep oscillator may be called an integrating filter or integrating network. To integrate means to unite many parts into a whole. The vertical filter unites the effects of all the serrated vertical sync pulses into a single voltage pulse. We shall use the names horizontal filter and vertical filter.

THE HORIZONTAL FILTER. The horizontal filter or differentiating filter consists only of a capacitor Cf and resistor Rf connected as shown in Fig. 53-1, between the output of the sync section and circuits leading more or less directly to the grid of the horizontal sweep oscillator. The input voltage applied to the filter capacitor from the sync section is of approximate square wave form. The output, which appears across the filter resistor, consists of sharply peaked positive and negative pips or spikes.

The output pips from the horizontal filter may be seen by themselves on the oscilloscope only when working on one of the relatively few receivers in which the filter connects directly to the oscillator grid. In the great majority of sets the oscillator grid is connected also to other circuits whose effects obscure the simple waveform which is delivered by the filter.

The fact that the horizontal filter really does change square waves into positive and negative spikes of voltage is illustrated by Fig. 53-2. At the left is an oscilloscope trace of a square wave produced by a square-wave signal generator. The frequency was adjusted to 15,750 cycles per second, which is the horizontal sync frequency. This square wave was applied to a filter circuit such as shown by Fig. 53-1, using a capacitor of 150 mmf and a resistor of 2,200 ohms. These are values quite often used in television receiver circuits. The oscilloscope shows the filter output, across the resistor, as pictured at the right.

The voltage which is forming the square waves first goes suddenly positive, to form the rising leading

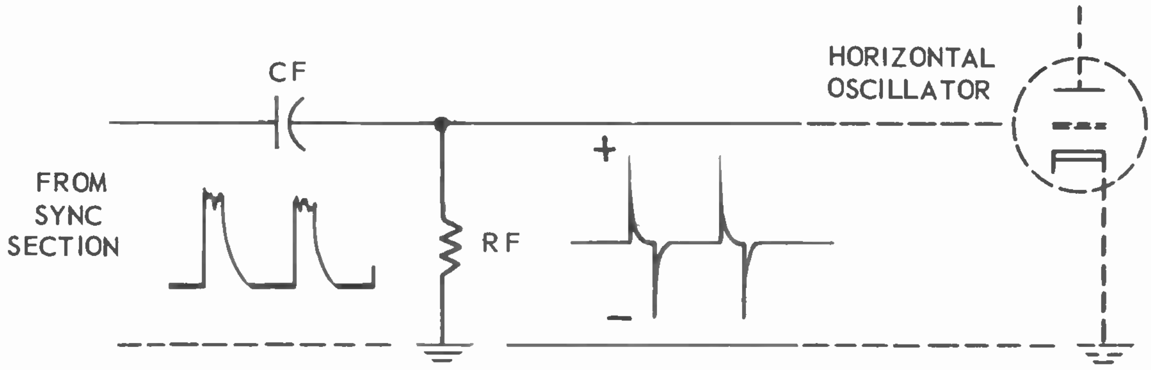


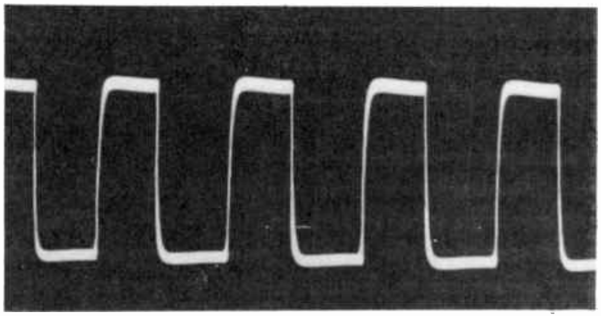
Fig. 53-1. A horizontal filter and its input and output waveforms.

edge of these waves. This voltage charges the filter capacitor, as shown by the corresponding positive spike of the output voltage. The capacitor charges almost instantaneously through the resistor, as shown by the quick drop of the spike voltage.

This charge of the filter capacitor occurs while the square wave voltage is at its maximum positive value, so we have maximum positive input voltage and a charged capacitor. Then the square wave goes suddenly back to zero, which is equivalent to a negative change of its voltage. This negative change of square wave voltage discharges the capacitor and recharges it in a negative polarity. This is shown by the negative spike of output voltage. Again the capacitor charges almost instantaneously, and is ready to be discharged and charged all over again by the next positive edge of the square wave voltage.

During normal operation of the television receiver the input to the horizontal filter consists of horizontal, equalizing, and serrated vertical sync pulses as represented at the top of Fig. 53-3. Resulting voltage spikes in the filter output are shown down below. Each time the input pulse voltage rises, at a leading edge, there is produced a positive voltage spike in the output. When the input voltage goes back to zero, effectively going negative, there is produced a negative voltage spike in the output.

The horizontal sweep oscillator operates in such manner that it can be timed or "triggered" only by



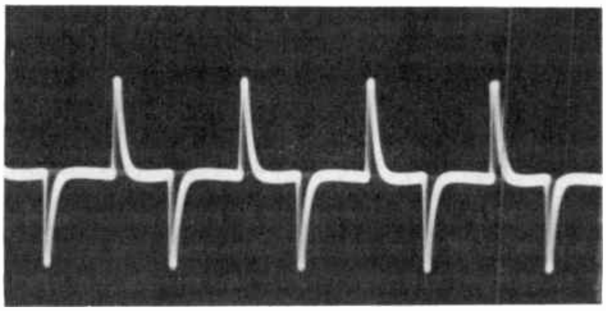


Fig. 53-2. A sawtooth wave applied to the input of a horizontal filter causes positive and negative voltage spikes at the output.

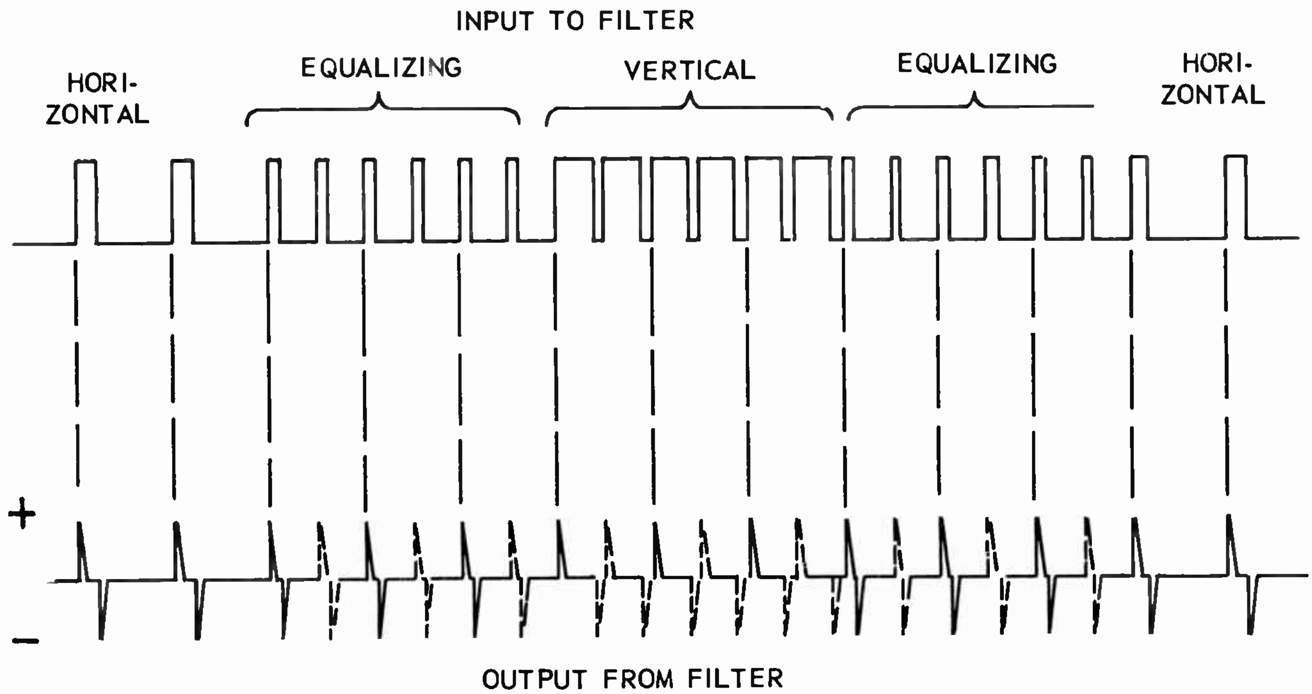


Fig. 53-3. Sync pulses and their effects in the horizontal filter.

positive voltage spikes from the filter output. Consequently, we may disregard all the negative spikes. Furthermore, the horizontal oscillator can be triggered by a positive spike only during certain portions of its oscillating cycle, only during periods when it is almost ready to oscillate even without the added positive voltage from a spike. Then only the positive voltage spikes that are shown by full lines can be effective in timing the oscillator. Shown by broken lines are intervening spikes caused by alternate equalizing pulses and by alternate serrations in the vertical pulse, but these have no effect on the horizontal oscillator.

Note that the filter capacitor discharges very quickly because its time constant, in connection with the filter resistor, is extremely short. The time constant of the capacitor and resistor used in making the traces of Fig. 53-2 is 0.33 microsecond. Time constants for this filter usually are less than one microsecond.

THE VERTICAL FILTER. A vertical filter most often is constructed with the parts represented by Fig. 53-4. The voltage pulse from the sync section appears first across resistor Ra. This resistor ordinarily would be in the cathode circuit or plate circuit of the last tube in the sync section. The voltage pulse acts through resistor Rb to charge capacitor Cb. The resulting voltage wave across capacitor Cb acts through resistor Rc to charge capacitor Cc. Finally, the voltage which appears across Cc acts through Rd to charge capacitor Cd, and the voltage developed across capacitor Cd is applied directly or through intervening circuits to the grid-cathode circuit of the vertical sweep oscillator.

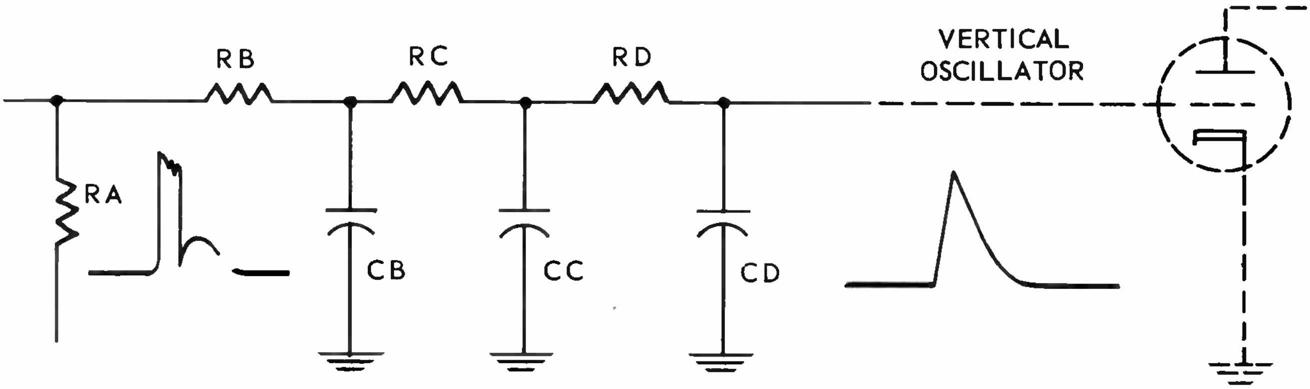


Fig. 53-4. The parts of a vertical filter, and its effect on waveform.

Fig. 53-5 shows how the output waveform from the vertical filter is produced. To this filter are applied the horizontal, equalizing, and serrated vertical sync pulses, just as they are applied to the horizontal filter. But because the filter capacitors and resistors are arranged in a different manner, and because they have far longer time constants, the action in the vertical filter is decidedly different than in the horizontal filter. At the top of the diagram are shown the various sync pulses that occur before, during, and after a vertical blanking interval. Down below is shown the rise and fall of charge voltage on the first of the filter capacitors.

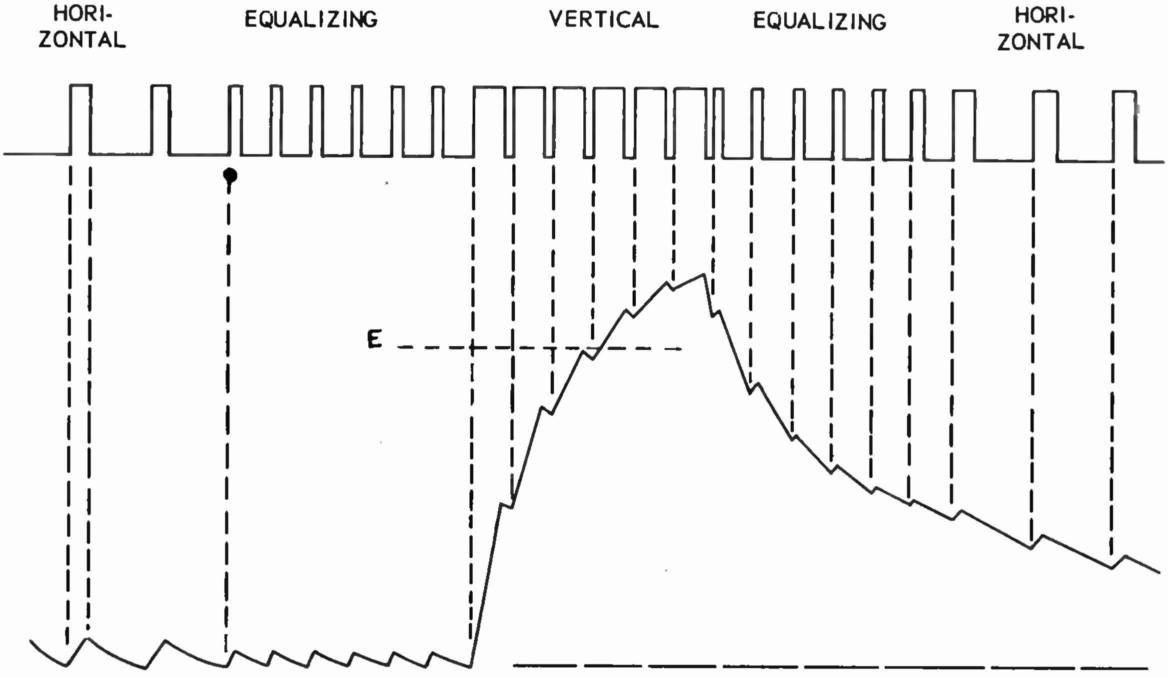


Fig. 53-5. How the synchronizing voltage is built up by the vertical filter.

Starting from the left-hand side of the diagram we see that each horizontal pulse of voltage imparts a small charge to the capacitor. This charge leaks away through the filter resistors before the next sync pulse and the next charge. Each of the equalizing pulses causes a similar action. There is complete discharge of the capacitor because the intervals between pulses are long in comparison with the duration of the pulses.

But when we come to the first of the serrated vertical pulses there is a relatively long charging time, followed by a very brief discharge. The capacitor discharges only a little between successive charges. As a result, the capacitor voltage builds up higher and higher. The vertical sweep oscillator is operated in such manner that it is triggered when the capacitor voltage rises to some value such as indicated by line E on the diagram. The charge voltage is allowed to rise still higher in order that the oscillator surely may be triggered.

After the last vertical pulse there is a brief period for discharge. Because of the high voltage on the capacitor this discharge is rapid. Then follow short charges from each of the following equalizing pulses, with relatively long periods for discharge between these pulses. These alternate short charges and long discharges bring capacitor voltage to zero during these equalizing and the following horizontal pulses. Then, with all remaining horizontal sync pulses, we have the same action as in the beginning, at the left-hand side of the diagram. The capacitor charges and voltages developed by the horizontal and equalizing pulses are nowhere near great enough to trigger the sweep oscillator, so it remains inactive until the next series of vertical pulses come along at the end of 1/60 second.

The added sections of the vertical filter get rid of the irregular changes of voltage and deliver at the filter output a smoothly rising and falling voltage which is ideally suited for triggering the sweep oscillator. What is happening may be seen with the help of the oscilloscope. At A in Fig. 53-6 is pictured the oscilloscope trace taken from the filter input, at the top of capacitor Cb in Fig. 53-4. You can plainly see the successive charges being put into the first capacitor by current flowing through resistor Rb. This is an actual picture of what is shown by the lower part of the diagram in Fig. 53-5.

Photograph B of Fig. 53-6 is of the oscilloscope trace taken from the middle of the vertical filter, at the top of capacitor Cc in Fig. 53-4. Although it still is possible to distinguish the successive charges and discharges, the change of voltage is much smoother. The rapid charges and discharges from horizontal and equalizing pulses, which appear along the lower line of the pictures, now have all but disappeared. Effects of any high-frequency noise voltages or similar interference would likewise be reduced in going through the filter.

Photograph C of Fig. 53-6 shows the oscilloscope trace taken from the filter output, at the top of capacitor Cd in Fig. 53-4. This is the waveform of voltage that will be used for triggering the vertical oscillator. The sharp rise at the beginning is due to the six serrated vertical pulses, whose effects here have been integrated into a single steady increase of voltage. The following gradual discharge takes place during the time of the equalizing pulses and horizontal sync pulses which follow the vertical pulses.

The time constant of each filter resistor and the following capacitor usually is on the order of 30 to 50 microseconds. Resistors usually are of some value between 7,000 and 20,000 ohms, with capacitances ranging from 0.002 to 0.006 mf.

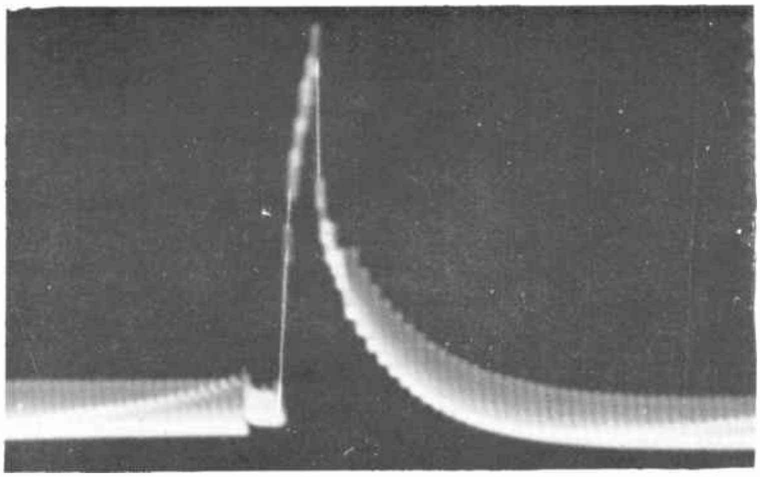
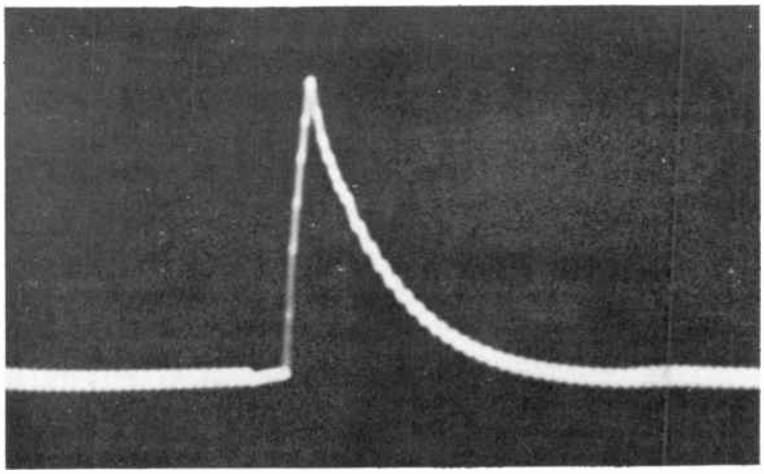
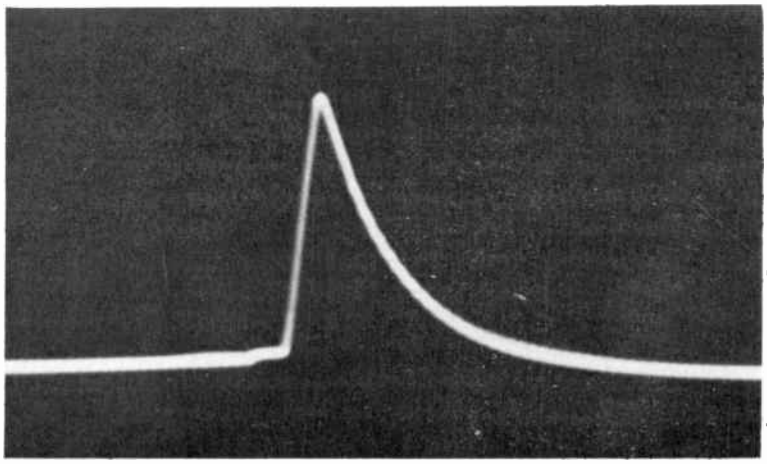


Fig. 53-6. Voltage waveforms at the input

(A), the center



(B), and the output



(C) of a vertical filter.

(A) *BLOCKING OSCILLATORS.* Television sweep oscillators may be either of two principal types, one called the blocking oscillator and the other called the multivibrator. We shall first examine the action of a blocking oscillator, whose basic circuit is shown by Fig. 53-7. This is a familiar circuit in which oscillation is maintained by feedback of energy from plate to grid through a transformer having one winding in series with the plate and the other winding in series with the grid. The oscillating frequency would depend on values of inductance and capacitance in plate and grid circuit. It would be the frequency of resonance for the inductance and capacitance.

It will be well to briefly review the action of this oscillator. To begin with, we shall assume that plate current is increasing in the tube and through the plate winding of the feedback transformer. This constitutes a change of current in the transformer winding. A changing current induces an emf in a coupled winding, which here is the winding in the grid circuit. This second winding is so connected into the grid circuit that the induced emf is making the grid more positive.

As the grid is made more positive it causes further increase of plate current. The continued change of plate current in the original direction continues to induce emf in the grid winding, making the grid still more positive and causing still more plate current to flow. This action requires an appreciable length of time, because the rise of plate current is opposed by counter-emf in the plate winding, just as any change of current is opposed by counter-emf in any inductance. As we learned in an early lesson, the rate of current increase is at first very rapid, and then slows down. The emf being induced in the grid winding rises rapidly at first, and then slows down along with the increase of the inducing plate current.

Finally, the decreasing rate of rise in induced emf no longer can force enough energy back into the grid circuit to overcome losses of energy which are occuring in that circuit. Then the grid goes no further positive, and plate current is increased no more. When plate current ceases its change there is no further induction, since induction results only from a changing current. Then there is no induced emf for the grid, and the grid voltage goes back to zero.

As grid voltage goes from highly positive to zero it is effectively changing in a negative direction.

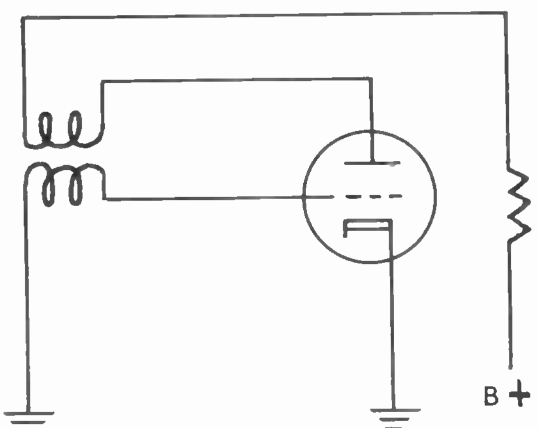


Fig. 53-7. Basic circuit for a blocking oscillator.

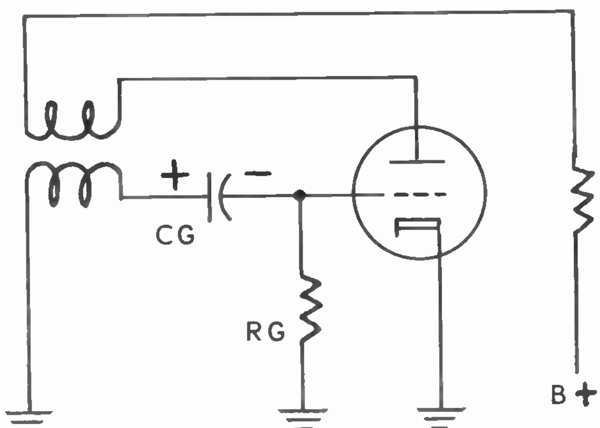


Fig. 53-8. Blocking results from a grid capacitor and resistor.

This, of course, reduces the plate current. With plate current decreasing it is changing in a direction or polarity which is opposite to the original increase. This changing current again induces emf in the grid winding, but now this emf is of reversed polarity - it is making the grid more and more negative.

The grid quickly reaches the negative voltage that causes plate current cutoff. When the plate current thus is stopped from flowing there is no further change of this current. With no change of current there is no induction, and no emf is applied to the grid. This lets the grid voltage return to zero from the negative value that caused cutoff of plate current.

In returning from negative to zero the grid voltage is effectively changing in a positive direction. As the grid goes less negative or more positive it causes plate current to resume, and to increase. Now we have an increasing plate current, and have arrived back at the point where we came in. We have followed the action during one cycle of current and voltage changes, or during one cycle of oscillation. These cycles now repeat at the natural or resonant frequency of the circuits.

Our next step will be to introduce the blocking action by means of capacitor Cg and resistor Rg which have been added to the oscillator grid circuit in Fig. 53-8. As we have just learned, the oscillator grid is driven positive during the first oscillating cycle by feedback from the plate. With the grid positive there is electron flow from cathode to grid in the tube, and through capacitor Cg and the grid winding back through ground to the cathode. Electrons thus added to the side of Cg which is toward the grid make this side negative, as marked. The capacitor is charged to a voltage approximately equal to maximum positive grid voltage.

The point is soon reached where the plate current starts leveling off, and less positive voltage is fed back to the grid. This change of grid voltage causes a reduction of plate current. We know that decreasing plate current induces through the transformer a negative emf at the grid. This feedback action makes the grid go rapidly more negative. The combined effects of negative feedback and negative charge on capacitor Cg drive the grid voltage far more negative than the value for plate current cutoff. The feedback emf still further increases the capacitor charge, in a polarity which is negative toward the grid. The oscillator now is "blocked".

Now the strong negative charge on grid capacitor Cg holds the grid negative and keeps the plate current cut off. The charge leaks slowly away through grid resistor Rg. This allows the grid to become less and less negative, until its voltage finally reaches the value at which plate current can resume. Once more we have come back to the point of increasing plate current, and the whole performance repeats.

The oscillator no longer can operate at its natural or resonant frequency. There can be only the first part of a single cycle, during which the grid goes momentarily positive and then far negative. The next cycle cannot begin until enough charge leaks off capacitor Cg, through the grid resistor, to allow resumption of plate current flow. The rate or the frequency at which these single separated cycles may occur depends on the time constant of capacitor Cg and resistor Rg, because the time constant determines how long it will take for the capacitor to discharge to the point where plate current can resume its flow.

Obviously, if resistance at Rg is adjustable, this adjustment may be used to vary the time constant and thus change the rate or frequency of oscillation. Such an adjustable grid resistor forms a hold control, about which we shall learn much more a little later on.

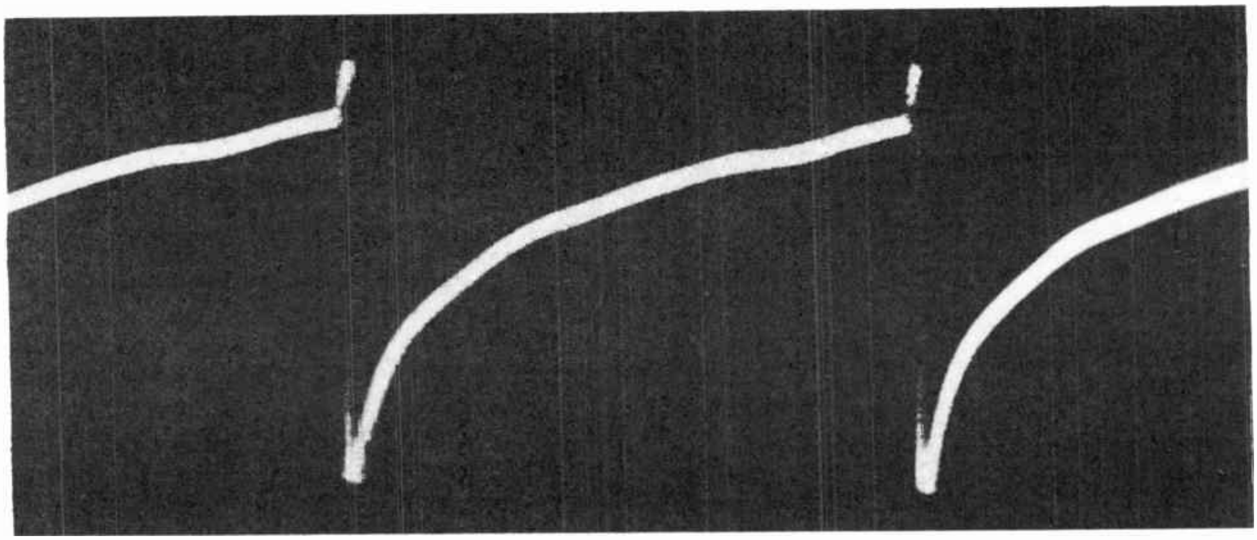


Fig. 53-9. Changes of grid voltage in the blocking oscillator.

The oscilloscope shows grid voltage of the blocking oscillator as in Fig. 53-9. Plainly evident are the sudden changes of voltage (downward) to a value which is strongly negative. Then the grid voltage rises slowly toward zero along the curved part of the trace as the grid capacitor discharges through the grid resistor. At the instant in which the oscillator tube becomes conductive, and when plate current starts in again, there is a short but sharp upward pulse of positive voltage. This positive voltage is followed instantly by the next drop to maximum negative. This type of oscillator is called a blocking type because its action is blocked while the capacitor is discharging, or while the grid voltage is held more negative than required for plate current cutoff.

FORMING A SAWTOOTH VOLTAGE. To have horizontal deflection of the electron beam in the picture tube it is necessary that the beam be moved at a uniform rate of speed from left to right during formation of each luminous line, then returned very quickly to the left-hand side of the screen in readiness for starting the next line. In a picture tube employing electrostatic deflection this movement of the beam is secured by a deflection voltage that increases uniformly in one polarity to shift the beam from left to right, and that then decreases or changes in the opposite polarity very rapidly to bring the beam back to its starting point. This is a sawtooth voltage. For magnetic deflection we require a sawtooth current, which is derived from a sawtooth voltage.

Vertical deflection is caused similarly by a sawtooth voltage for an electrostatic picture tube or by a sawtooth current for a magnetic picture tube. In any case we require, to begin with, a sawtooth voltage whose rise and fall is controlled by the horizontal sweep oscillator or the vertical sweep oscillator. The sawtooth voltage becomes available when we add to our oscillator circuit the sawtooth capacitor marked Cs in Fig. 53-10.

In diagram A the oscillator grid is strongly negative and the tube is non-conductive. Electron flow is

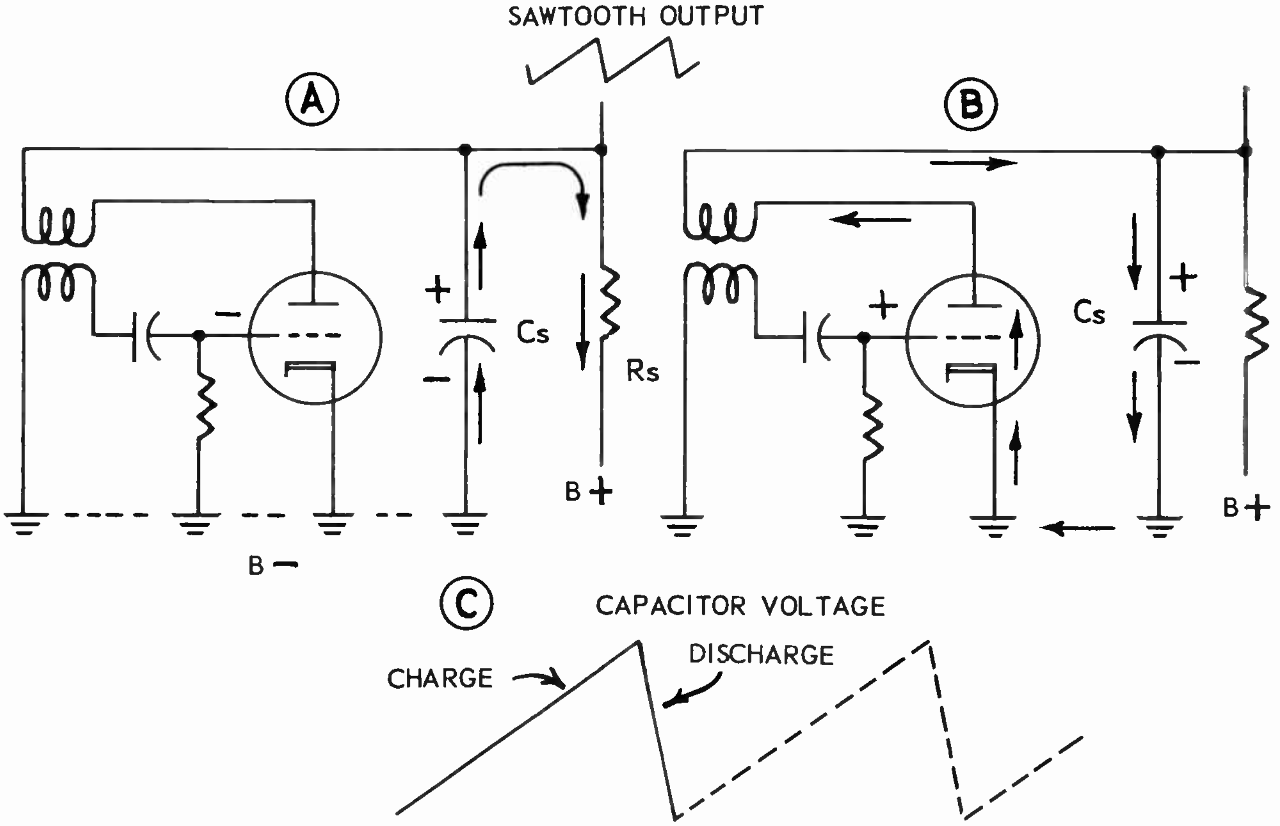


Fig. 53-10. The sawtooth wave results from charge and discharge of a capacitor.

from B-minus, or ground, through capacitor Cs and resistor Rs to B-plus. This electron flow charges the capacitor in the marked polarity. In diagram B the grid of the oscillator tube is momentarily positive, and the tube is highly conductive for electron flow from cathode to plate. The negative side of capacitor Cs is connected to the oscillator cathode through ground, and the positive side of this capacitor is directly connected to the oscillator plate. Consequently, the capacitor discharges very rapidly through the tube.

Diagram C shows the changes of capacitor charge and voltage. The charging and accompanying rise of voltage takes place rather slowly. This is because the capacitor is being charged through resistor Rs, and the rate of charge is determined by the time constant of this capacitor-resistor combination. Discharge is very rapid, because there is nothing to hinder it except the negligible internal resistance of the oscillator tube, whose grid is positive, and the small resistances of the circuit connections and transformer plate winding.

If we are dealing with a horizontal sweep oscillator the relatively slow rise of capacitor voltage will be used to control electron beam deflection from left to right on the picture tube screen, while a luminous line is being formed. The rapid fall of capacitor voltage during discharge will control the horizontal retrace time. In the case of a vertical sweep oscillator the relatively slow rise of voltage will control downward travel of the electron beam during formation of one field. Then the rapid fall of voltage during capacitor discharge will control the vertical retrace time. Fig. 53-11 is a picture of a sawtooth voltage wave as shown by the oscilloscope connected to the plate of a sweep oscillator.

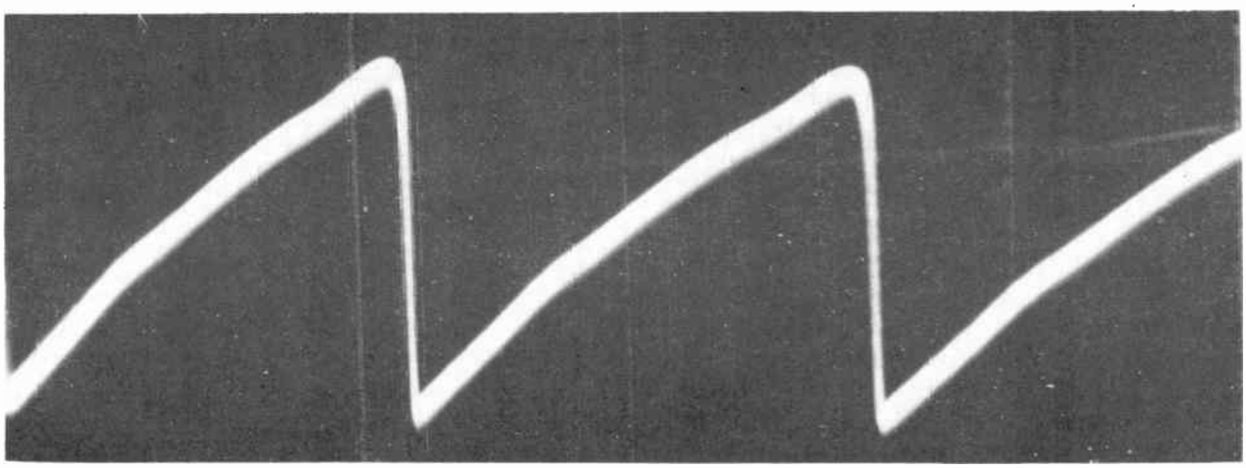


Fig. 53-11. A sawtooth wave as seen on the oscilloscope.

If charging of the capacitor were not interrupted by the oscillator grid becoming positive, and allowing discharge, the voltage on the capacitor would increase to the full voltage of the B-power line. The time consumed in reaching this voltage, or any fraction of it before discharge commences, depends on the time constant of capacitor Cs and resistor Rs.

How far the picture tube electron beam is swept vertically or horizontally depends on how high the sawtooth voltage rises before there is discharge of the capacitor and retrace in the picture tube. This is the the same as saying that the height or width of the picture is proportional to the maximum voltage during charge of the sawtooth capacitor.

If resistor Rs is adjustable in value it may be used to change the rate of electron flow and the maximum voltage reached during charge of the sawtooth capacitor. Since this action will alter the height or width of the picture, an adjustable resistor at Rs forms one kind of size control, or one kind of height or width control.

SYNCHRONIZING THE SAWTOOTH. We have developed a sweep oscillator which will produce a sawtooth deflecting voltage at a frequency controlled by the time constant of a capacitor and resistor in the oscillator grid circuit. But what we actually are aiming for is a sawtooth voltage whose frequency is controlled by sync pulses of the received signal. These sync pulses, which are part of the original composite signal, have been changed by the sync section of the receiver into sharp spikes or pips of voltage occuring at the same frequency as that of the original sync pulses. This voltage now will be used to time the action of the sweep oscillator, and thereby to control the frequency of the sawtooth voltage at the oscillator output.

In the circuit diagram of Fig. 53-12 we have opened the grid circuit of the sweep oscillator, on the side between the transformer grid winding and ground. This opened side of the oscillator grid circuit is connected to the output of the sync section, so that voltage spikes from the sync section are applied to the oscillator grid.

The effect of synchronizing voltage on the oscillator grid is shown at the right in Fig. 53-12. Up above is represented the voltage wave at the oscillator grid when no synchronizing voltage is being applied. The frequency here is that determined by the time constant of the capacitor and resistor in the grid circuit. It is called the free running frequency of the sweep oscillator. The grid resistor Rg is adjusted to make the free running frequency just a little slower than the frequency of the sync voltages, which would be the horizontal line frequency of 15,750 cycles per second for a horizontal sweep oscillator, or the vertical field frequency of 60 cycles per second for a vertical sweep oscillator.

Each spike of synchronizing voltage now occurs just a little before the oscillator grid would go positive at the free running frequency. The positive sync voltage drives the grid positive at the sync frequency, and the oscillator then goes through a cycle just as though the positive grid had occured at the free running frequency. On the next cycle the next sync voltage drives the oscillator grid positive just before the cycle would have ended with the free running frequency. Thus the oscillating cycles are forced by the sync voltages to take place at the sync frequency rather than at the free running frequency.

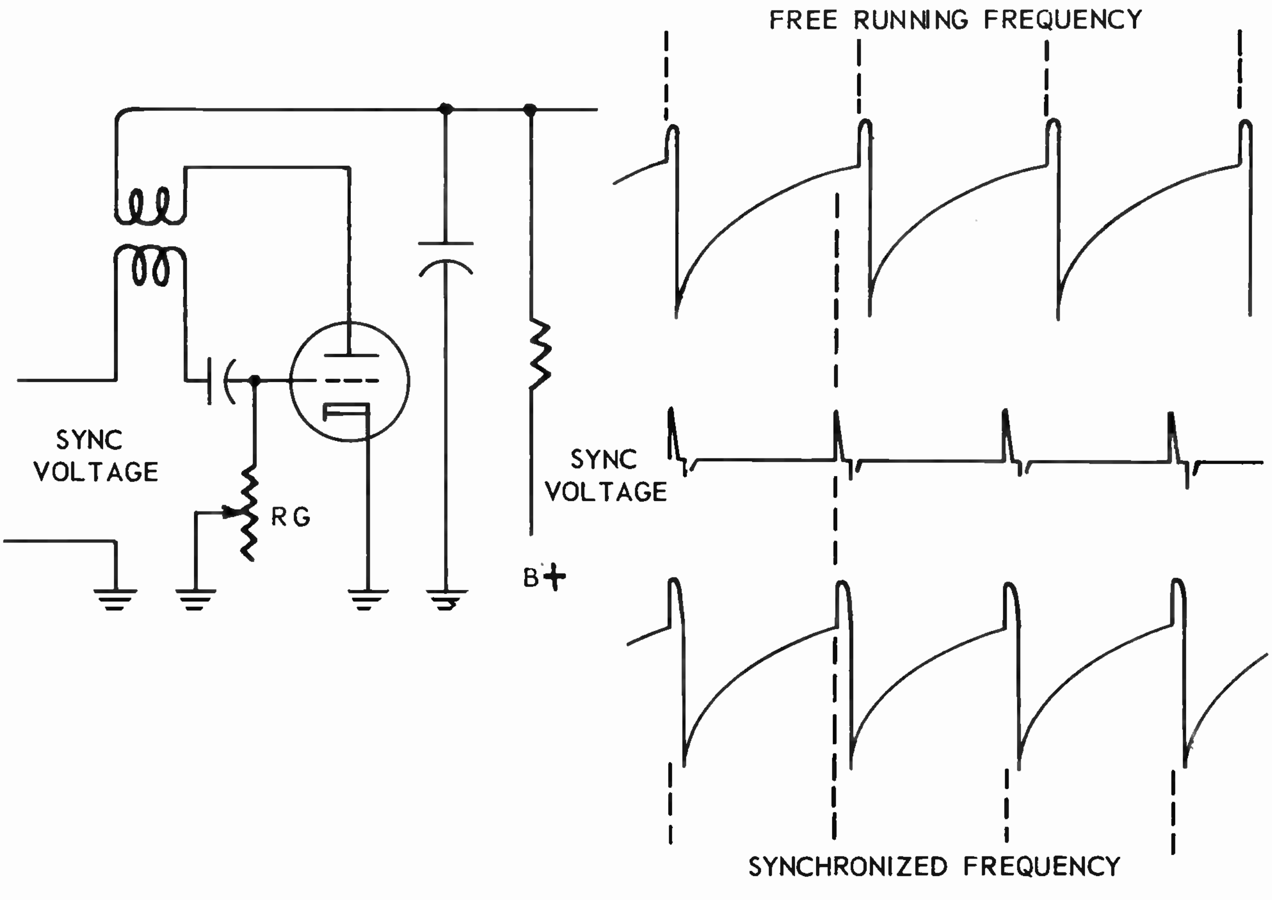


Fig. 53-12. How pulses of synchronizing voltage act to time the oscillator.

The adjustable resistor Rg in the oscillator grid circuit is a hold control. As you adjust this hold control to bring the free running frequency closer and closer to the sync frequency, the picture or pattern on the picture tube will lock into synchronization when control of oscillator frequency is taken over by the sync voltages. If the free running frequency is made too greatly different from the sync frequency, the synchronizing voltages will occur so far down on the curve of grid voltage that they cannot overcome the remaining negative voltage on the grid, and there will be no synchronization.

If no transmitted signal is being received there will be no sync pulses, and no sync voltages will come to the grid of the sweep oscillator. Then the oscillator will continue operating at its free running frequency. Any fault in the receiver that prevents received signals from causing sync voltages at the sweep oscillator will prevent synchronization of pictures or patterns, but the oscillator will continue to operate at its free running frequency. This fact is important in service work, for continued operation of the sweep oscillator allows making tests with the oscilloscope and other instruments at and beyond the oscillator even when no signals are being received.

Continued operation of the sweep oscillators when no signal is being received is effective also in protecting the screen or phosphor in the picture tube. Were there to be no oscillations there would be no sawtooth deflecting voltages, and the electron beam in the picture tube would remain stationary. The stationary bright spot on the screen soon would burn or destroy the phosphor at this point. So long as the sweep oscillators operate there is deflection of the electron beam, and illumination is distributed over the picture tube screen, even though there is no synchronization of pictures or patterns.

DISCHARGE TUBE. Blocking oscillators in some receivers are operated in connection with a discharge tube as shown by Fig. 53-13. The purpose is to provide a separate tube, or separate section of a twin tube, for discharge of the sawtooth capacitor Cs, while allowing the oscillator to control the timing as determined by sync pulse frequency.

In the oscillator and discharge tubes the grids are connected directly together, and so are the cathodes. Consequently, changes of grid voltage in the discharge tube are exactly the same as in the oscillator. Both grids become positive or negative at the same instants of time, and to the same degree. The discharge tube is conductive while the oscillator is conductive, and is non-conductive while the oscillator is non-conductive.

The sawtooth capacitor Cs charges through resistor Rs in the usual way. But instead of discharging through the cathode and plate of the oscillator, this capacitor discharges through the cathode and plate of the discharge tube. The charge and discharge, and the corresponding sawtooth voltage at the output, have the same waveforms as when using the blocking oscillator alone.

Some of the reasons for using a separate discharge tube are as follows: It is possible to use a higher B-plus voltage on the discharge tube than on the oscillator. A high B-plus voltage on the discharge tube allows obtaining a greater change of voltage in the sawtooth wave, or allows a stronger sawtooth deflecting voltage. With a high B-voltage on the oscillator there is a large plate current and large grid current while the grid is positive. The large plate current or great changes of plate current must go through the plate winding of the feedback transformer, where the current changes induce strong counter-emf's which tend to extend the period of positive grid voltage and to extend the time for retrace. With capacitor discharge through the oscillator, the discharge current must go through the plate winding of the feedback

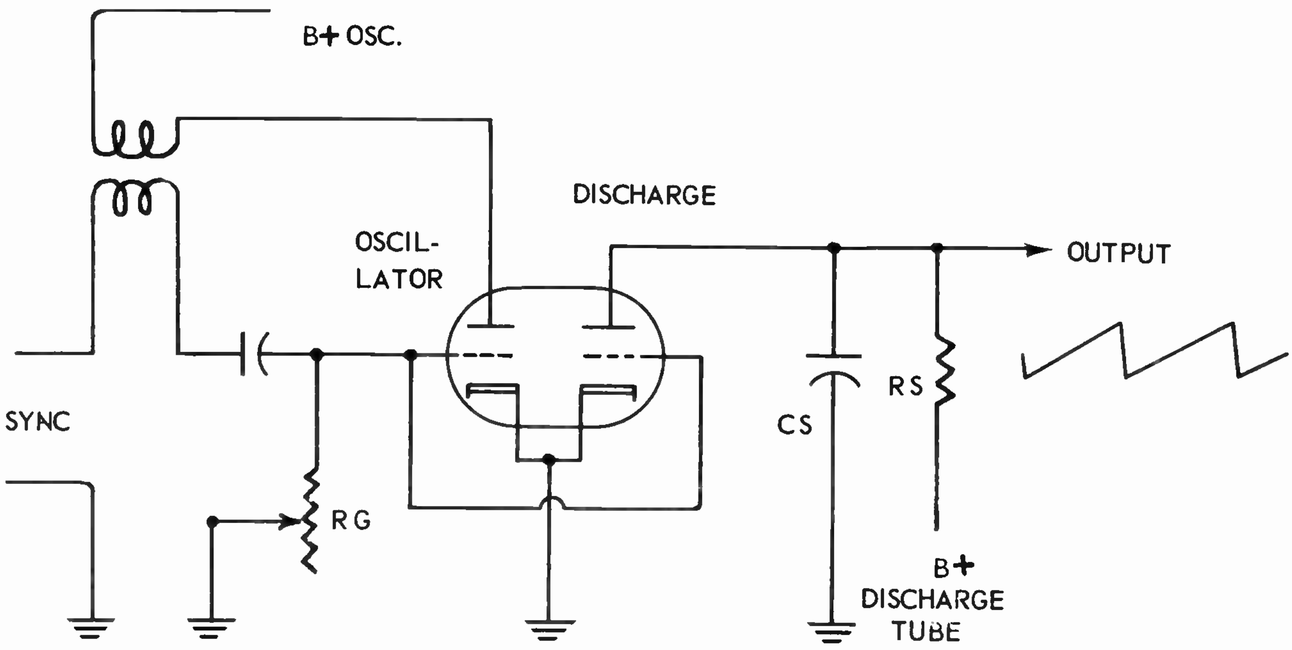


Fig. 53-13. A discharge tube used with a blocking oscillator.

transformer. The large and sudden discharge current induces strong counter-emf in the transformer, and further slows down the discharge to extend the time of retrace.

In spite of all the apparently good reasons for using a separate discharge tube it seldom is employed. Retrace time and other operating features are entirely satisfactory with a well designed and constructed blocking oscillator used alone. Counter-emf's are reduced by using no more inductance than necessary for voltage feedback through the transformer. Sawtooth voltages of moderate strength are easily brought up in a following sweep amplifier tube.

OSCILLATOR VOLTAGES AND CURRENTS. In Fig. 53-9 we looked at an oscilloscope trace showing changes of voltage at the grid of a blocking oscillator. In explanation of the action there shown it was stated that oscillator plate current rises to a high value when the grid goes positive, and drops to zero as grid voltage goes down through plate current cutoff. Now we shall look at a trace showing changes of plate current.

The oscilloscope is not designed to give direct indications of changes in current, its traces are produced only by varying voltage. This is because the input terminals of the oscilloscope are connected to or coupled to the grid-cathode circuits of amplifier tubes within the instrument. The outputs of these amplifiers are used to deflect the electron beam in the cathode-ray tube of the oscilloscope. The amplifiers are actuated only by voltages or changes of voltage applied to the grids, as is the case with all amplifiers.

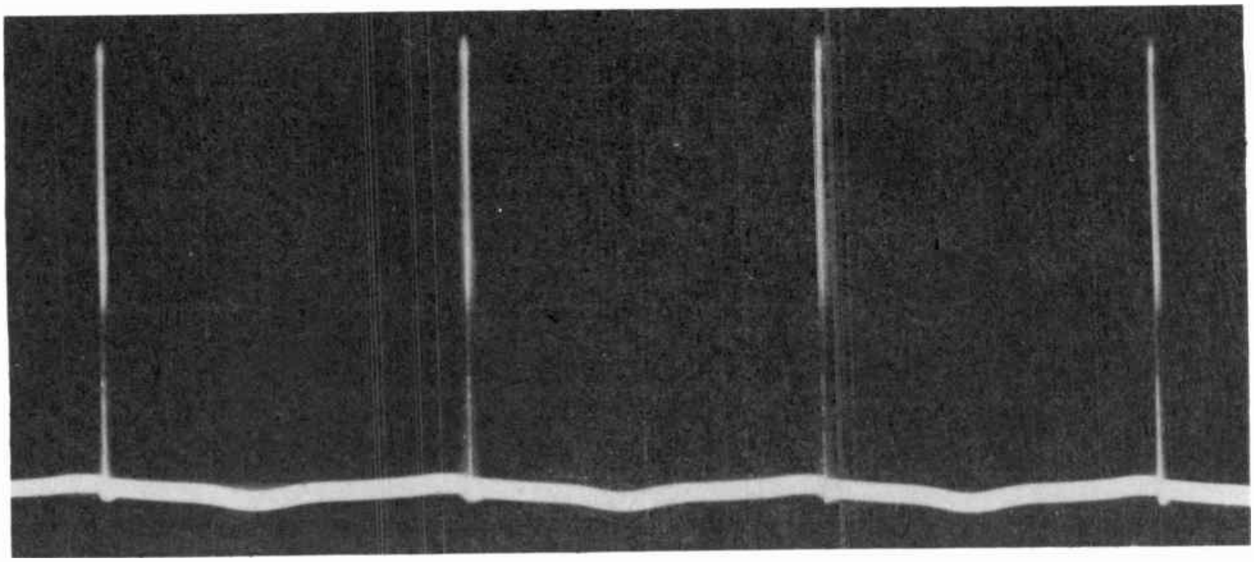


Fig. 53-14. Pulses of plate current in the blocking oscillator.

In spite of this limitation it is possible to observe changes of current by means of the oscilloscope. All we need do is connect in series with the circuit whose current is to be observed a non-inductive resistor, then observe the changes of voltage across the resistor. In a pure resistance the voltage and current are in phase, and voltage changes occur at the same instants and in the same polarities as current changes.

This expedient of employing a series resistor was used in obtaining the trace of Fig. 53-14. The plate of a blocking oscillator was disconnected from the plate winding of the feedback transformer, and was reconnected through a 100-ohm resistor. This amount of resistance is too small to alter the action in the oscillator plate circuit. Then the input terminals of the oscilloscope were connected across the series resistor, and the trace photographed. It is clearly evident that, every time the oscillator grid goes instantaneously positive, there is an accompanying very brief pulse of plate current. Then the plate current falls back to zero until the next positive voltage on the grid. Plate current remains at zero during the entire time for discharge of the grid capacitor.

With the series resistor removed, and the oscillator plate again connected directly to the feedback transformer, the oscilloscope was connected to the plate. The trace is pictured by Fig. 53-15. It shows changes of voltage at the oscillator plate, not changes of plate current. The voltage here represented is a combination of the oscillator plate voltage and the charge-discharge voltage on the sawtooth capacitor. Looking back at preceding circuit diagrams you will see that the oscillator plate is directly connected to the sawtooth capacitor <u>Cs</u>, so voltage observed at the plate is being taken from both the tube and the capacitor. Incidentally, we would obtain the same traces at the plate and at the grid of a discharge tube operated from a blocking oscillator.

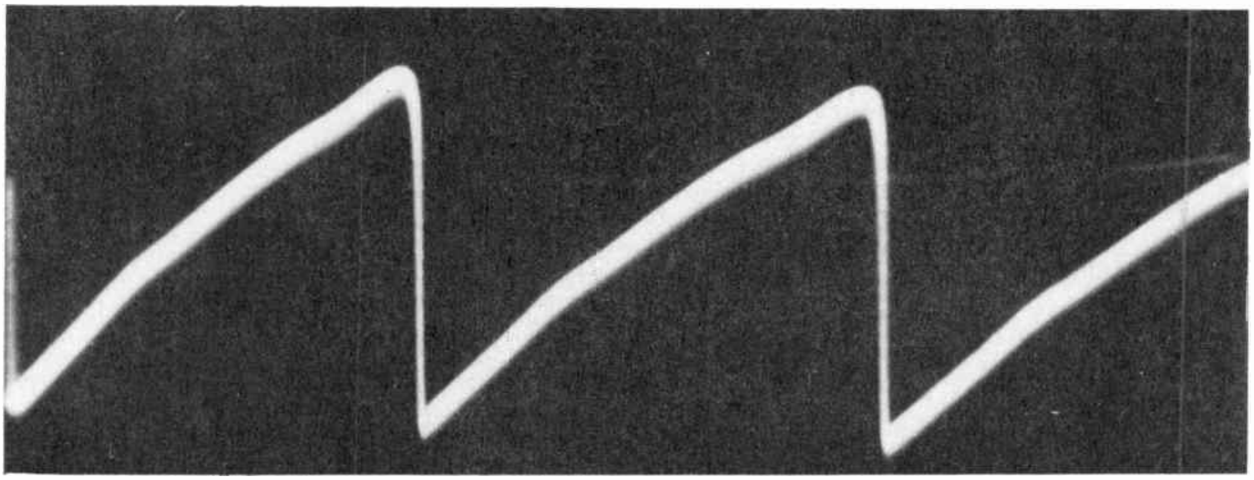


Fig. 53-15. Voltage wave at the plate of the blocking oscillator and at the sawtooth capacitor.

The steep vertical lines on the voltage trace of Fig. 53-15 show sudden drops of plate and capacitor voltage. Plate voltage drops because of the very large momentary plate current that accompanies the positive grid. As you know, plate voltage drops when current rises, because the current causes an increased drop in plate load resistance. Capacitor voltage drops because the capacitor is discharging through the oscillator or discharge tube.

The longer upward traces show plate voltage rising as plate current decreases with the grid coming back through zero and going to cutoff voltage. These traces show also that the capacitor has discharged and is ready for recharging. The long traces sloping gradually upward toward the right show that the sawtooth capacitor is being charged and that its voltage is increasing. This capacitor charging voltage also exists at the tube plate, because capacitor and plate are direct connected.

We have made a rather detailed analysis of what happens in the grid and plate circuits of the blocking oscillator, and have shown how these things are pictured by the oscilloscope. The purpose has been not only to learn how this important type of oscillator acts, but also to show how any apparently intricate electrical actions can be explained by the elementary principles of electricity and electronics.

You, as a television technician, will use the oscilloscope and other service instruments in diagnosing and locating receiver troubles, much as a physician uses X-rays and the indications of many instruments in diagnosing human ailments. The doctor is able to draw useful conclusions from instrument indications only because he is thoroughly acquainted with the human body and the laws that govern its performance. You will be able to draw useful information from service instruments only when you correctly interpret their indications. This you can do only by relying on the fundamental laws relating to resistance, inductance, and capacitance, and by understanding how television and radio apparatus is supposed to perform.

THE MULTIVIBRATOR. Sweep oscillators employing the multivibrator, which will be studied in the following lesson, have many features in common with blocking oscillators. At least, the sweep circuits utilizing the two kinds of oscillator get their final results in much the same way. In both cases the saw-tooth output voltage is formed by charging a capacitor through a resistance and discharging it through a tube. Voltage pulses from the sync section control the timing of the multivibrator when applied to its grid circuit, just as when applied to the grid circuit of the blocking oscillator.

The most apparent difference between a multivibrator and a blocking oscillator are, first, that two tubes or two sections of a twin tube nearly always work together to form a multivibrator. Second, feedback in the multivibrator is secured through resistance or capacitance, rather than through the inductive feedback used for the blocking oscillator.

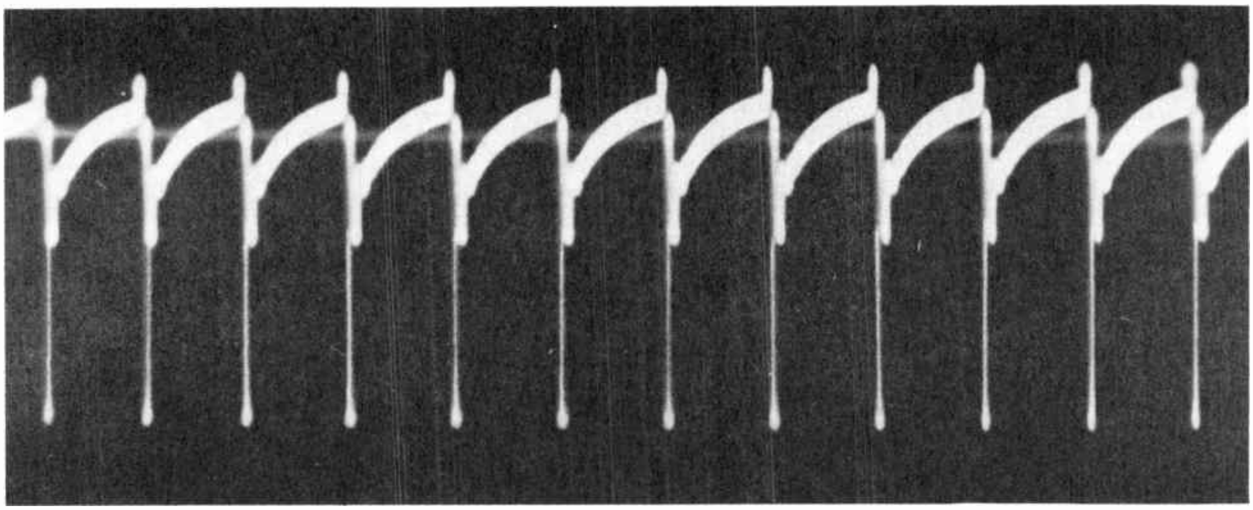


Fig. 53-16. You should be able to identify the voltage shown by this oscilloscope trace.

TELEVISION

LESSON NO. 54

MULTIVIBRATOR SWEEP OSCILLATORS

Training Your Mind...
C. T. I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind

C.T.I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON NO. 54

MULTIVIBRATOR SWEEP OSCILLATORS

The blocking oscillator which we studied in the preceding lesson is used more than all other types combined in the vertical sweep sections of television receivers. Now we shall examine the multivibrator oscillator, which is found most often in horizontal sweep circuits, although frequently used also in vertical sections. There are many varieties or modifications of the multivibrator, but the one most generally employed in television receivers is the cathode coupled type, which may be called the Potter sweep circuit. The action of this multivibrator may be explained by considering the complete oscillator circuit to include two triodes and their couplings as shown by Fig. 54-1.

Voltage pulses from the sync section of the receiver are applied to the grid of tube A. In the plate circuit of tube B is the sawtooth capacitor Cs and resistor Rs through which this capacitor is charged. The charge escapes from Cs through tube B when its grid is made positive and the tube becomes conductive. Note that in the plate circuit of tube B we have the same parts as in the plate circuit of a blocking oscillator or the plate circuit of a discharge tube used with a blocking oscillator.

For the time being we shall disregard the effect of sync pulses and learn how the multivibrator becomes a free running oscillator. When B-voltage is applied through resistor Ro a moderate plate current for tube A will flow to its cathode through resistor Rk. Resulting voltage drop across Rk becomes a small negative bias voltage for tube A, whose grid return resistor goes to ground and through ground to the bottom of resistor Rk. Tube A now is normally conductive.

At the same time there is an electron flow that charges capacitor Cg through the path marked by arrows in diagram 1 of Fig. 54-2. This flow is upward through resistors Rk and Rg. Because Rg is of many megohms resistance it carries only a small current, but the effect is to make the top of Rg and the grid of tube B go positive with reference to the cathode of this tube.

With the grid of tube B thus made positive, the internal resistance between cathode and grid becomes very small. This low-resistance cathode-to-grid path acts practically as a short circuit around the high resistance of Rg, and from cathode to grid flows a large current that charges capacitor Cg almost instantly.

The heavy pulse of current that charges capacitor Cg flows through resistor Rk. Since the top of Rk is connected to the cathode of A, while its lower end is connected through ground to the grid of this tube, the voltage drop across Rk acts as a strongly negative pulse of voltage on the grid of tube A.

The reduction of plate current that accompanies the negative voltage pulse on the grid of A normally would allow a considerable increase of voltage at the plate of this tube, because of less voltage drop in resistor Ro.

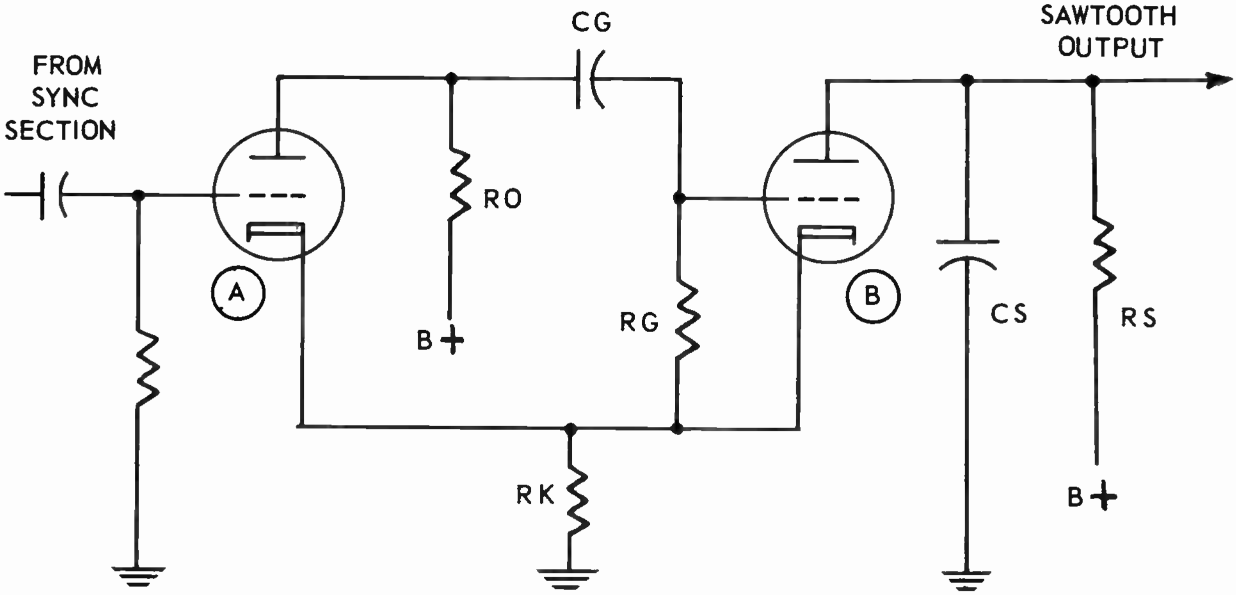


Fig. 54-1. Circuit connections for a cathode coupled multivibrator.

During the brief time of the charging pulse for capacitor Cg, we have on tube A a strongly negative pulse of grid voltage. Naturally, the result is to drive A to plate current cutoff. The total effect of charging of capacitor Cg has been to make tube A non conductive and tube B conductive. These conditions last only until completion of the charge into capacitor Cg.

As soon as capacitor Cg is fully charged, or is charged to the value of applied voltage, we no longer have the electron flow shown at 1 in Fig. 54-2. There is no longer any upward flow of electrons through resistor Rg, and there is no potential difference across this resistor. The grid of B goes momentarily through zero voltage with reference to its cathode, and the plate current is reduced considerably. This decrease of plate current reduces the voltage drop across resistor Rk and thus the grid bias of tube A. Tube A now is again subjected to normal cathode bias voltage from resistor Rk, and consequently, returns to a state of normal conduction. The conduction of tube A immediately commences a discharge of capacitor Cg through the path shown by diagram 2 of Fig. 54-2. Electron flow for discharge goes downward through resistor Rg to the cathode of tube A, thence from cathode to plate in this tube (which is conductive as explained) and back to the positive side of the capacitor.

The electron flow for discharge of capacitor Cg immediately makes the top of Rg and the grid of B negative with reference to the cathode of this tube. Then there can be no electron flow through the tube and all the remaining discharge from the capacitor must go through the high resistance of Rg. The period during which this discharge lasts depends almost entirely on the time constant of Cg and Rg, since internal resistance of tube A in the discharge path is much less than the several megohms of resistance in Rg.

The negative voltage maintained at the grid of tube B during the relatively slow discharge of Cg holds this tube at plate current cutoff. It is during this period of non-conduction in B that the sawtooth capacitor takes its charge.

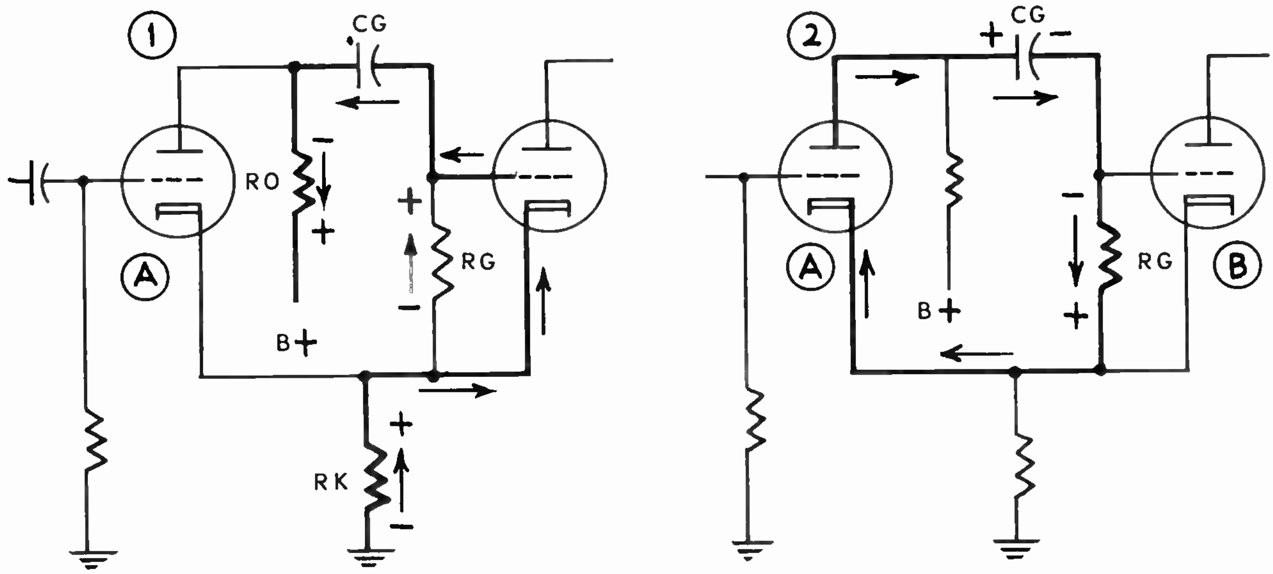


Fig. 54-2. Electrons flows during charge and discharge of the coupling capacitor.

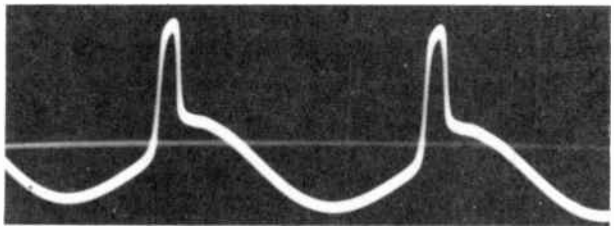
When capacitor Cg completes its discharge there is no longer any electron flow downward through resistor Rg, and grid voltage at B returns to zero. This change of grid voltage allows B to once more become conductive and there is immediate repetition of the action shown by diagram 1. While B is conductive, as indicated in that diagram, there is discharge of the sawtooth capacitor through the cathode-to-plate path within this tube.

Charge and discharge of capacitor Cg have made tube B briefly conductive, then non-conductive for a much longer time. The sawtooth capacitor connected to the plate circuit of this tube discharges during the period of conduction, and the rapidly falling sawtooth voltage controls retrace of the electron beam in the picture tube. During the longer non-conductive interval for tube B the sawtooth capacitor gradually takes a new charge, during which the sawtooth voltage increases to control movement of the picture tube beam during either a horizontal luminous line or the downward travel of the beam during a field, depending on whether the sweep oscillator is in a horizontal or a vertical deflection system.

Changes of voltage at the plate of tube A are shown by the oscilloscope trace at the left in Fig. 54-3. Each positive (upward) peak occurs while capacitor Cg is charging. During this charge there is plate current cutoff in A, which accounts for the highly positive plate voltage. When the charge is complete the plate voltage gradually drops, while plate current increases, and then the voltage rises upon approaching the next peak.

At the right of Fig. 54-3 is a trace showing changes of plate current in tube A. Here we see current dropping to the cutoff point during charge of capacitor Cg, then a sharp increase followed by a more gradual rise upon approaching the next cutoff.

Because the grid of tube B and the plate of tube A are connected together through capacitor Cg, the



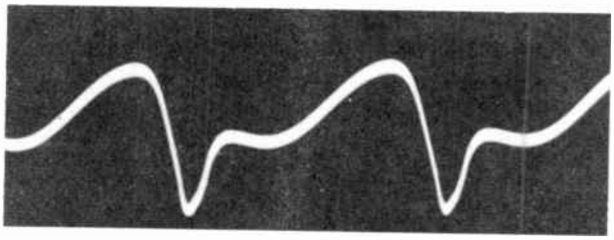


Fig. 54-3. Plate voltage (left) and plate current (right) of the first tube in the cathode coupled multivibrator

capacitor acts like a coupling capacitor in any resistance-capacitance coupled amplifier stage, transferring a voltage or signal from the plate of one tube to the grid of another.

Fig. 54-4 is an oscilloscope trace showing changes of plate current in tube B. There is a sudden increase of current as the sawtooth capacitor discharges through this tube. Then plate current falls to zero and remains there while capacitor Cg is discharging. During this period of zero plate current or plate current cutoff the sawtooth capacitor is being recharged from the B-supply.

With the oscilloscope connected across resistor Rk, or connected between the tube cathodes and ground, we obtain the trace of Fig. 54-5. This trace is almost the same as that for plate current in tube B, since by far the greatest of all currents in the cathode resistor is that from discharge of the sawtooth capacitor. This trace taken from the cathode resistor shows changes of both current and voltage, because in a resistance the voltage and current are in phase.

At the top of resistor Rk the voltage of the pulse is positive, and at the lower end it is negative. Since the grid of tube A connects through ground to the lower end of Rk, we have strong pulses of negative polarity at the grid of this tube. These negative voltage pulses which result from discharge of the saw-

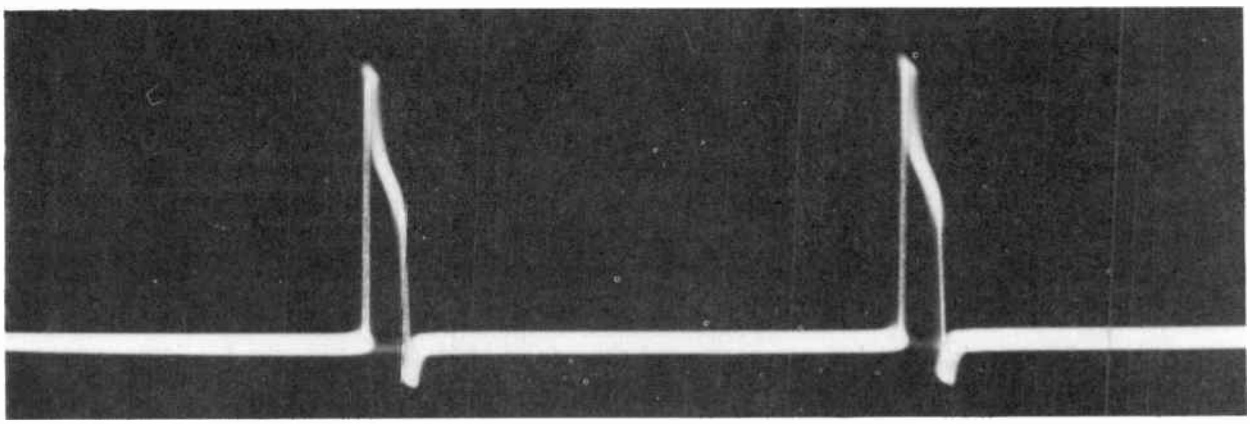


Fig. 54-4. Plate current in the second or discharge tube of the cathode coupled multivibrator.

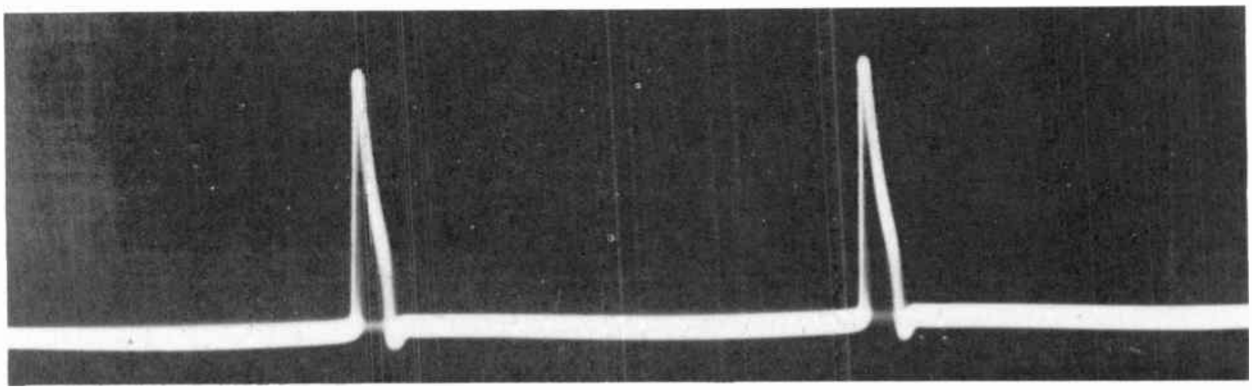


Fig. 54-5. Waveform of voltage and current in the cathode resistor of the multivibrator.

tooth capacitor add to the strength of negative pulses resulting from charge of capacitor Cg, and the grid of A is driven still farther beyond plate current cutoff.

Now we shall look at changes of voltage (not current) that occur at the plate of tube B, as shown by the oscilloscope trace of Fig. 54-6. This voltage is entirely unlike the change of plate current shown by Fig. 54-4. Here we have not only the voltage change that accompanies change of plate current, but also the voltage resulting from charge and discharge of the sawtooth capacitor which is connected directly to the tube plate.

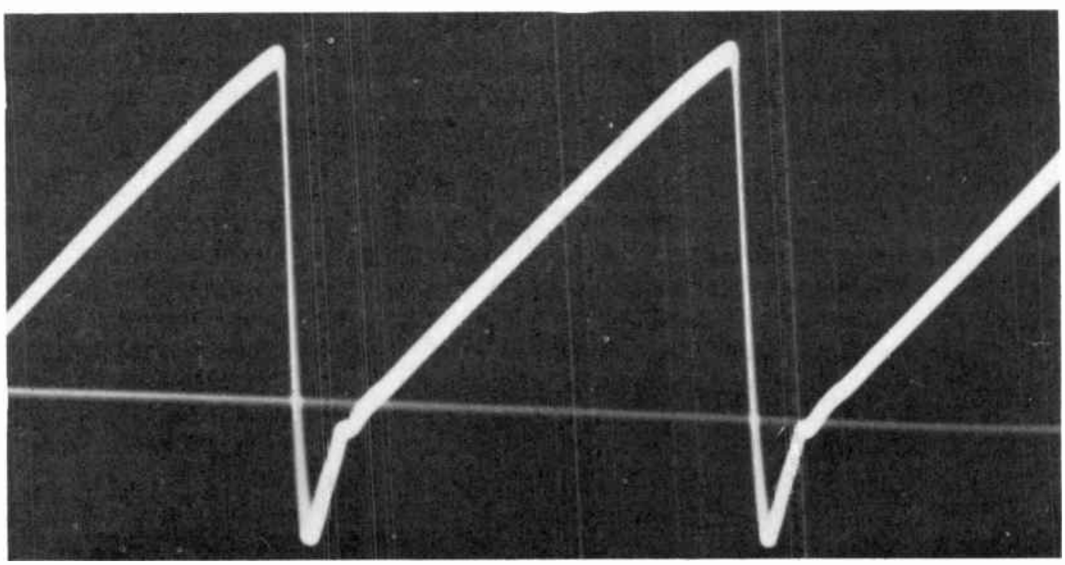


Fig. 54-6. Voltage at the plate of the discharge tube and at the plate side of the sawtooth capacitor in the multivibrator circuit.

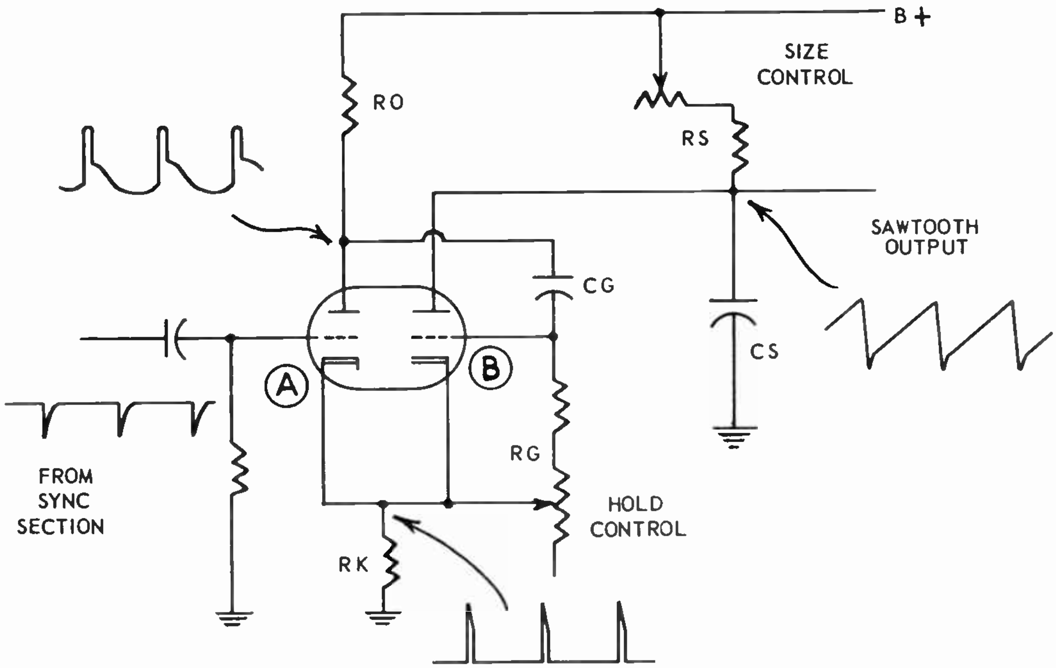


Fig. 54-7. Circuits of the cathode coupled multivibrator often are shown like this in service diagrams.

The sudden drops of voltage at plate and sawtooth capacitor occur while this capacitor is discharging through tube B while its grid is positive. The entire voltage wave is practically the same as the one observed at the tube plate and sawtooth capacitor for the blocking oscillator. The relatively slow rise of voltage following the sudden drop indicates that the sawtooth capacitor is being charged through resistor Rs of Fig. 54-1, and is getting ready for the next discharge and sudden drop of voltage.

The circuit connections for a cathode coupled multivibrator sweep oscillator often are shown by diagrams generally similar to that of Fig. 54-7. Diagrams in manufacturers' service manuals often carry drawings of correct forms of oscilloscope traces obtained from various points, just as they are shown on this circuit diagram.

1 Electrical connections in Fig. 54-7 are the same as in Figs. 54-1 and 54-2, and parts are identified by the same letters. The two tubes usually are the two sections of a twin triode. The maximum voltage to which the sawtooth capacitor may charge between instants of discharge is varied by making resistor Rs adjustable. Since this maximum voltage determines the extent of horizontal or vertical deflections of the picture tube beam the adjustable resistor is called a size control. For a horizontal sweep oscillator it might be called a width control, and for a vertical sweep oscillator it might be called a height control.

The length of periods during which tube B is held non-conductive by negative voltage at the top of resistor Rg depends on the time constant of Rg and capacitor Cg. Varying the resistance at Rg alters this time constant, changes the rate at which tube B allows charge and discharge of the sawtooth capacitor.

and accordingly varies the free running frequency of the oscillator. Resistance at Rg is made adjustable, and is used as a hold control. The hold control is set to make the free running frequency of the oscillator just a little slower than the frequency of sync pulses which are to time the operation of this oscillator. Then the sync pulses bring the oscillator into synchronization with the received signal, just as when using a blocking oscillator.

The start of each cycle of oscillation in the multivibrator occurs when the grid of tube B is made positive and when the sawtooth capacitor discharges through this tube. The grid of tube B goes positive at the same instant in which the grid of A goes negative, as we learned when examining the action of the multivibrator. In order that voltage pulses from the sync section may act to time the oscillations these pulses must be negative when applied to the grid of tube A. Then these sync pulses have exactly the same effect as negative pulses produced at the grid of A by charging of capacitor Cg, and they cause the grid of B to go positive because of polarity inversion between grid and plate of A. The negative sync pulses arrive at the grid of A just before there would be cutoff in this tube at the free running frequency.

Sync pulse voltages from the sync section sometimes are applied to the top or at some point between top and bottom of cathode resistor Rk instead of to the grid of tube A. Then the synchronizing voltages must be positive instead of negative. This is because, as shown by Fig. 54-5, the pulses of cathode current and voltage that help in driving tube A to cutoff are positive at the cathode resistor. In order that voltage pulses from the sync section when applied to the cathode, may have the same effect as feedback voltage pulses at the cathode, these sync voltages must be positive. Positive sync voltages at the cathode, of tube A have exactly the same effect as negative pulses at the grid.

In the multivibrator which we have examined the sudden changes of voltage during the retrace portion of the sawtooth wave cause pulses of current and voltage in the cathode resistor. From this resistor the voltage pulses go to the grid of the first tube. Thus there is an effective feedback by way of the cathode circuit, and we have a cathode coupled multivibrator.

CAPACITOR COUPLED MULTIVIBRATOR. The first modification of the multivibrator to be studied is shown by Fig. 54-8. The feedback from output to the grid of the first tube is through a capacitor rather than through a cathode resistor. In explaining this circuit we shall assume, to begin with, that negative voltage pulses from the sync section are being applied to the grid of tube A. Later we shall see how the multivibrator would be free running in the absence of sync pulses.

The negative sync pulses at the grid of A cause positive voltage pulses at the plate, because of the ever present inversion of polarity between a control grid and a plate. These positive pulses act through capacitor Cg to make the grid of tube B momentarily positive. While the grid is positive there is a large electron flow from ground or B-minus to the cathode of B, thence to the positive grid, into the side of capacitor Cg that is connected to the grid, and out of the other side of this capacitor to the positive or plate side of the circuit. Capacitor Cg thus receives a strong charge in the polarity marked on the diagram, negative toward the grid of tube B.

Part of the charge current for capacitor Cg must flow upward through resistor Rg placing a positive voltage on the grid of tube B, increasing plate current and decreasing the plate voltage. This decrease of voltage discharges capacitor Ca through resistor Ra, biasing and holding tube A at plate current cut-off for the "Rc" time of capacitor Ca and resistor Ra. The point is soon reached when capacitor Ca is dis-

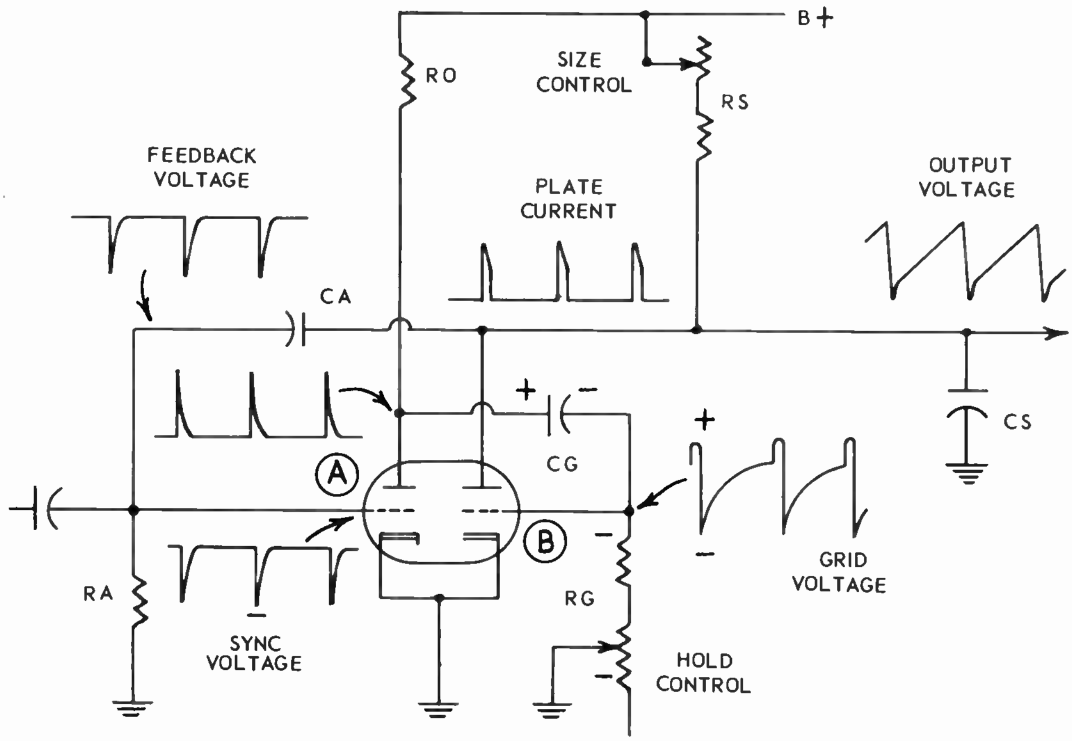


Fig. 54-8. Circuit connections and waveforms of the capacitor coupled multivibrator.

charged, and no voltage developed across resistor Ra. Tube A control grid voltage momentarily goes through zero, and begins conduction. Conduction of tube A begins a discharge of capacitor Cg down through resistor Rg and driving the grid of tube B far negative. Capacitor Cg discharging through the high resistance of the hold control Rg places a negative voltage on the control grid of tube B. As discharge decreases, there is a slow return of grid voltage to its zero value.

Sawtooth capacitor Cs charges through resistance at Rs or the size control while the negative voltage on the grid of B holds this tube non-conductive. The sawtooth capacitor discharges through B in the usual manner, while the grid of this tube is momentarily positive.

There is plate current in B only during the positive voltage pulses on its grid. This current is shown by the oscilloscope trace of Fig. 54-4. These sudden increases of plate current flow through resistor Rs, or the size control resistance, and cause equally sudden and brief increases of voltage drops across this resistance. Each increase of voltage drop across resistance Rs must be accompanied by an approximately equal decrease of voltage at the plate of B. So we have negative pulses of voltage at the plate of B.

The negative pulses of voltage at the plate of B act through feedback capacitor Ca to make the grid of A momentarily negative. That is, the negative pulses of voltage at the plate of B cause negative pulses on the far side of capacitor Ca in the same manner that positive pulses at the plate of A are causing positive pulses on the far side of capacitor Cg.

The negative voltage pulses appear across resistor Ra and thus are applied to the grid of Tube A. The time constant of Ca and Ra is only about 1/50 as long as that of Cg and Rg. Therefore, the negative feedback pulses at the grid of A last for only very brief periods. In fact, they are almost equivalent in duration to the negative pulses that come from the sync section.

Now we have at the grid of A a continuing series of negative feedback pulses which are just as effective as the negative sync pulses in initiating each cycle of oscillation. Were no signal being received, or if sync pulses were to cease for any other reason, the oscillator would continue to operate at a free running frequency because of the continuing feedback pulses.

The free running frequency is determined by the discharge time constant of capacitor Cg and the hold control resistance at Rg. This frequency is adjusted by means of the hold control to be just a little slower than the sync frequency. Then each sync pulse comes to the grid of A just before a feedback pulse would have arrived to start a cycle, and the oscillator is pulled into time with the sync pulses.

MULTIVIBRATOR WITH OUTPUT AMPLIFIER. In the next modification of the multivibrator to be examined the second tube is used not only as a portion of the oscillating system but also as an amplifier for the sawtooth voltage wave. The output of the second tube then is applied through an impedance matching transformer to the deflection coils for a magnetic-deflection type of picture tube.

A circuit of this general type is shown by Fig. 54-9. Note that the sawtooth capacitor Cs and the size control resistance Rs through which this capacitor is charged are in the plate circuit of tube A, rather than in the plate circuit of tube B as in the other multivibrators. Plate voltage of A is only about 15 volts positive, due to the large drop from B-plus through several megohms of resistance at Rs.

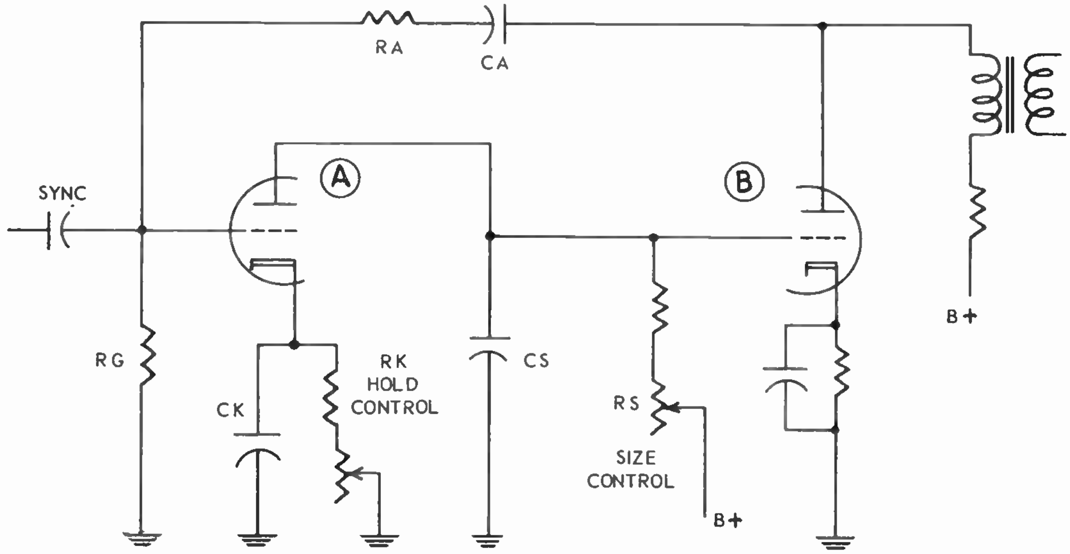


Fig. 54-9. Multivibrator circuit in which the second tube is an output amplifier.

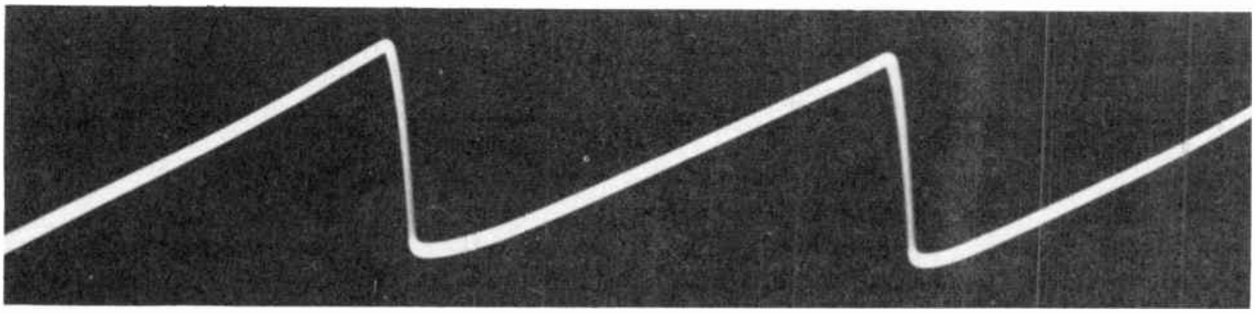


Fig. 54-10. Voltage at the first plate and at the sawtooth capacitor.

The grid of tube B is directly connected to the plate of A, making the grid of B about 15 volts positive with reference to ground. There is, however, a negative bias on the grid of B due to action of its cathode resistor. In the cathode resistor the voltage drop normally is 20 to 25 volts. The difference between this drop and the plate voltage from A becomes negative bias for B. In some receivers there is a blocking and coupling capacitor between the plate of A and the grid of B, with the return circuit for the grid of B through a resistor to ground.

There is feedback from the plate of B to the grid of A through capacitor Ca and resistor Ra in series with each other. The plate of B connects also to the primary winding of the output transformer, whose secondary goes to the deflection coils. The hold control resistance Rk, which varies the free running frequency, here is an adjustable cathode resistance on tube A, rather than a grid resistor as in the other multivibrators. The hold control resistance is paralleled by capacitor Ck.

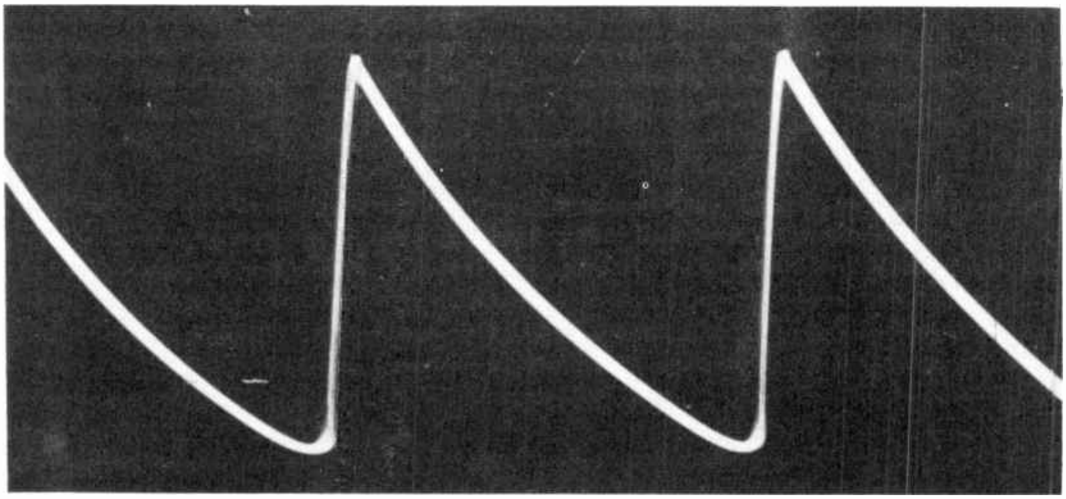


Fig. 54-11. Voltage at the plate of the output amplifier section of the multivibrator.

In explaining the action of this multivibrator we shall assume that capacitor Cs has been charged through resistance at Rs and that a positive sync pulse is applied to the grid of A. This tube is thus made conductive, and capacitor Cs discharges through the tube almost instantaneously. The pulse of discharge current through the cathode-plate path in A is like the current pulse shown by Fig. 54-4.

Fig. 54-10 is an oscilloscope trace taken at the plate of tube A. Here we have both the plate voltage and the voltage of charge and discharge for the sawtooth capacitor. As we should expect, this waveform for capacitor Cs is a sawtooth voltage suitable for control of beam deflection in a picture tube. The sawtooth voltage is applied to the grid of tube B, which amplifies and inverts the wave. The output from B is fed to the primary of the output transformer.

Voltage at the plate of B is shown by the oscilloscope trace of Fig. 54-11. This voltage will result in a sawtooth current in the transformer secondary and the deflection coils. It appears in the trace that the sawtooth voltage is of the wrong polarity for deflection, with sudden rises and gradual drops rather than the reverse. This causes no difficulty, because the current wave (or a voltage wave) from the secondary of the transformer can be applied to the deflection coils in either polarity. Reversal of polarity requires only a reversal of the connections of the two ends of the transformer secondary into its circuit.

The voltage wave at the plate of B is applied to the feedback capacitor and resistor Ca and Ra. After passing through these two units the voltage goes to resistor Rg and thus the grid of A. The time constant of capacitance and resistance in this path is only about 1/5000 second. Capacitor Ca is suddenly charged by the part of the wave in Fig. 54-10 that rises steeply to maximum positive. Then, due to the short time constant mentioned, capacitor Ca discharges very quickly.

The positive voltage pulse at the grid of A, resulting from charge of capacitor Ca, is shown by Fig. 54-12. Here we have the positive voltage which, when applied to the grid of A, was assumed to start the performance by allowing discharge of the sawtooth capacitor through A. The events explained will repeat over and over again, once for each oscillation cycle.

Now to examine the manner in which the hold control varies the intervals between discharges of the sawtooth capacitor, and varies the free running frequency of this multivibrator.

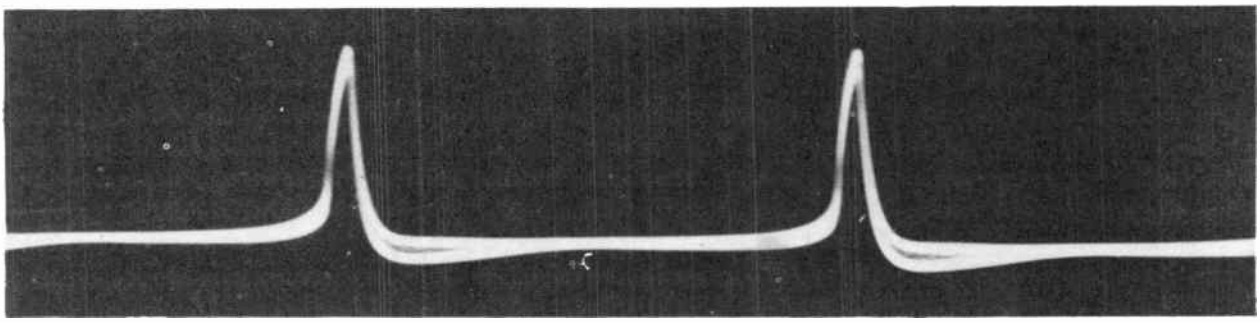


Fig. 54-12. Pulses of positive voltage at the grid of the first section in the combination multivibrator and output amplifier.

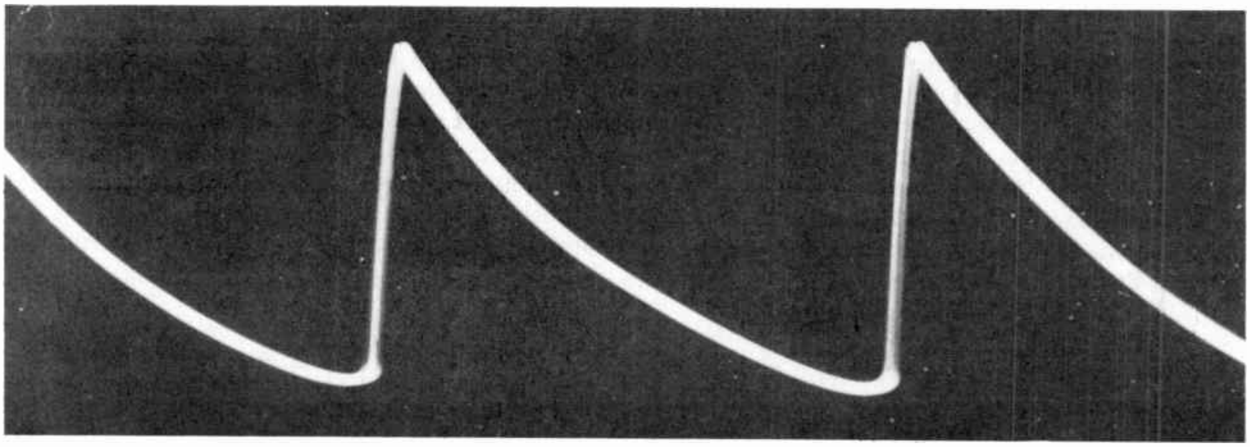


Fig. 54-13. Voltage at the cathode of the first section of the combination multivibrator.

Fig. 54-13 is a trace of the voltage at the cathode of A, at the top of resistor Rk and capacitor Ck. The sharp rises of voltage in positive polarity result from pulses of plate current, which is current from discharge of the sawtooth capacitor. The voltage developed across Rk is applied across capacitor Ck, and charges Ck. When a current pulse ends, capacitor Ck discharges at a relatively slow rate through the resistance of Rk. These decreases of capacitor and resistor voltages are shown by the long downward slopes on the oscilloscope trace. The greater the resistance in hold control resistor Rk the longer will be the time constant of Ck and Rk, and the longer it will take Ck to discharge.

Resistor Rk acts like a cathode bias resistor, because the grid of tube A is connected through resistor Rg and ground to the bottom of Rk. Every positive change of voltage at the top of Rk, and the tube cathode, is equivalent to a negative change at the bottom of Rk and the grid. These negative changes of grid voltage drive A to cutoff and hold it there, temporarily preventing discharge of the sawtooth capacitor through this tube.

5 The longer it takes capacitor Ck to discharge through the cathode resistor, the longer will be the intervals between returns of tube A to a condition near enough conduction that a moderate positive voltage pulse at the grid will allow discharge of the sawtooth capacitor. These longer intervals mean a lower free running frequency. Therefore, increasing the hold control resistance to increase the time constant of Ck and Rk will lower the free running frequency. Conversely, less hold control resistance will increase the free running frequency, because then capacitor Ck can discharge more quickly.

The sawtooth capacitor can discharge only when the grid of tube A is made positive and the tube conducts. Voltage pulses from the sync section applied to the grid of A may be of only positive polarity or they may consist of both positive and negative spikes. In a double polarity sync pulse only the positive spikes or pips will be effective in synchronizing this oscillator.

If it is desired to synchronize the oscillator by means of negative sync voltages these voltages are applied not at the grid of tube A but at or near the top of the cathode resistor or hold control. As you know, negative pulses at a cathode have the same effect as positive pulses at a grid of the same tube.

As with all sweep oscillators, the hold control is used to adjust the free running frequency to a rate just a little slower or lower than the sync frequency of the received signal. Then the pulses of voltage from the sync section trigger the oscillator just before it would operate at the free running frequency. In this way the discharges of the sawtooth capacitor are forced to occur in time with the sync pulses of the received signal.

Fig. 54-14 illustrates the action of positive voltage pulses coming from the sync section to the grid of tube A in the multivibrator. While examining these diagrams we must keep in mind that changes of voltage at the grid of the tube result from a combination of three voltages. First, we have the grid voltage or the varying bias due to the cathode resistor or hold control Rk. Second, there are the feedback pulses coming through capacitor Ca and resistor Ra from the plate of tube B. Third, there are the sync pulse voltages from the sync section.

Diagram 1 shows the combination of hold control voltage, practically a sawtooth wave, and a feedback pulse. Here the timing is that of the free running frequency as determined by the hold control adjustment. Diagram 2 represents a voltage pulse coming from the sync section just a little before there would be discharge of the sawtooth capacitor with conditions as in the first diagram. As shown by diagram 3 the positive sync pulse more than overcomes the remaining negative voltage of the hold control, the grid is driven positive, and there is discharge of the sawtooth capacitor. The time of the cycle at 3 is less than at 1, showing that oscillation frequency has been speeded up by the sync pulse, and brought into synchronization with the received signal.

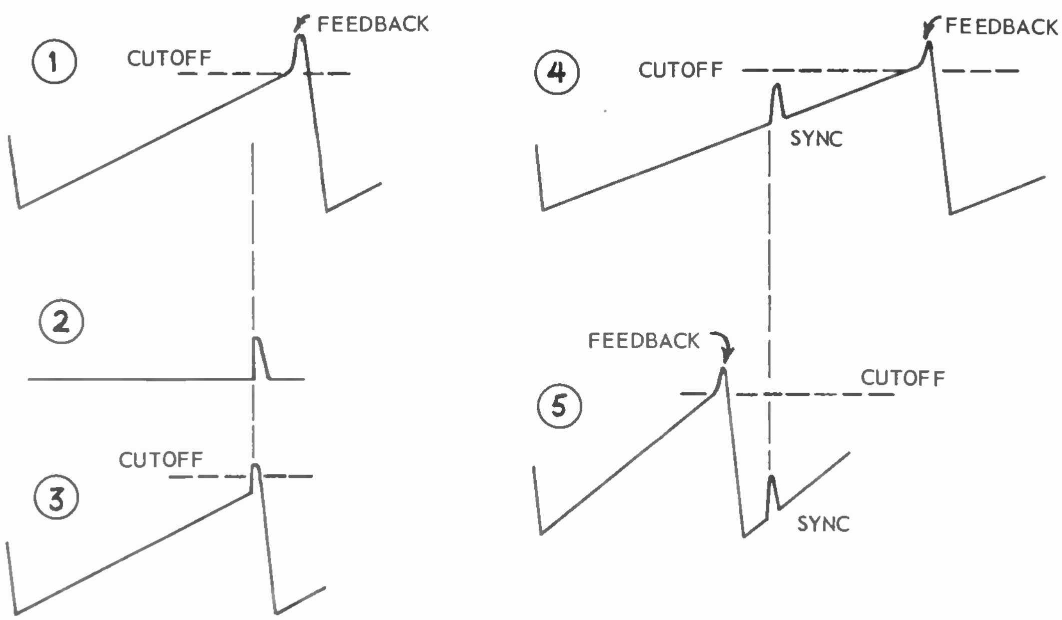


Fig. 54-14. How pulses of positive sync voltage act on the grid of the first section of the multivibrator.

Diagram 4 of Fig. 54-14 shows what happens when the hold control is adjusted for an excessively long free running frequency. The positive sync pulse arrives at the same time as before, but now the negative grid voltage due to the hold control is still so far below that of cutoff that the sync pulse cannot overcome it. The sync pulse cannot bring the grid out of cutoff and, of course, cannot make the grid positive. Then there is no synchronization with the received signal, and the oscillator continues to operate at the free running frequency.

Diagram 5 shows what happens when the hold control is adjusted for an excessively short free running period. Again the sync pulses arrive at the same time as in other diagrams. But now the oscillator already has tripped from its own feedback pulse, and grid voltage is far below cutoff. Naturally, the sync pulse cannot overcome this highly negative grid voltage, and there is no synchronization with the received signal.

This relation between free running frequency and sync pulses is generally true with any type of sweep oscillator. The diagrams of Fig. 54-14 might represent grid voltage from a grid resistor type of hold control, as used on other multivibrators and on blocking oscillators. By means of the hold control, the free running frequency must be brought reasonably close to and a little slower than the sync frequency in order that pulses of sync voltage may lock the oscillator into time with the received signal.

From Fig. 54-14 it is apparent that negative sync pulses occuring at any time in the cycle would have no effect. They merely would make the grid of the tube still more negative, and cutoff would be maintained. This is why sync voltages which go both positive and negative are entirely satisfactory for timing.

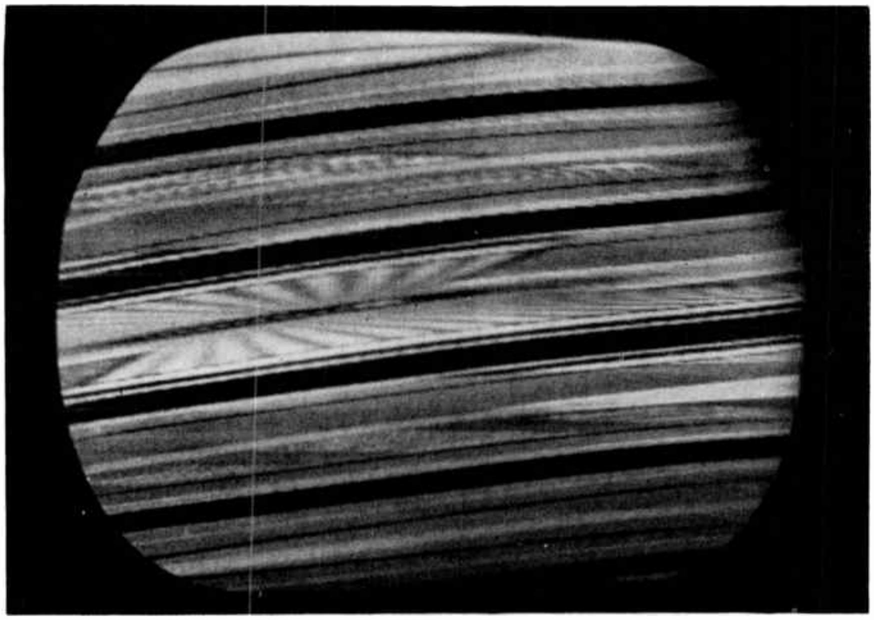


Fig. 54-15. Appearance of a pattern on the picture tube when the horizontal hold control is slightly out of adjustment.

If there are positive sync pulses midway between beginnings or ends of regularly synchronized cycles these intermediate pulses would occur far down on the curve of grid voltage, as at 4 in Fig. 54-14, and would have no effect. Such intermediate pulses would be produced by alternate equalizing pulses and by alternate serrations of the vertical pulse in a composite signal. When any such pulses arrive, the oscillator is so far from tripping that there is no effect on frequency.

HOLD CONTROL ADJUSTMENT. When a horizontal hold control is out of adjustment to a moderate degree, the pattern on the picture tube screen will appear about as shown by Fig. 54-15. There will be a series of dark horizontal bars sloping upward or downward toward the right, depending on whether the free running frequency is being made too fast or too slow. There is a complete pattern or picture between each two adjacent bars, but it is compressed by incorrect horizontal frequency and sloped by vertical synchronization, which here is assumed to be correct.

As the hold control is brought more nearly to a correct setting the bars will become fewer in number and will become more nearly vertical. Finally, the picture will snap into synchronization. If the hold control is turned still farther in the same direction until synchronization again is lost, usually the bars will reappear and will be sloped in the opposite direction.

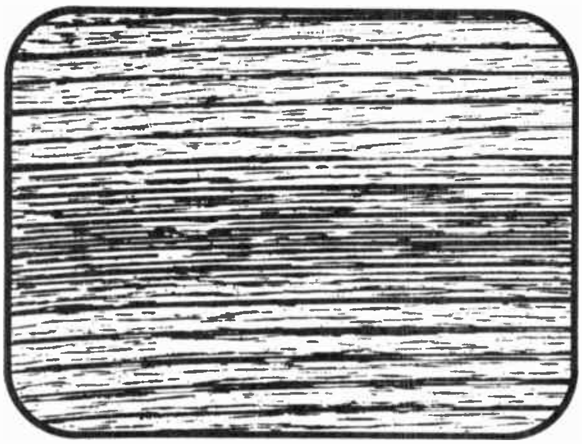


Fig. 54-16. When horizontal synchronization is far out of the way the picture tube shows many horizontal lines.

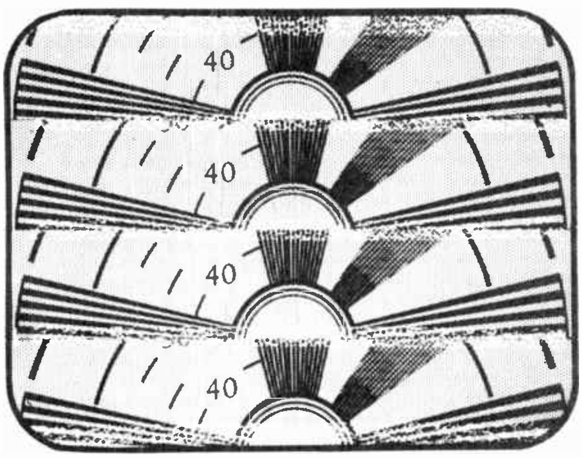


Fig. 54-17. When vertical synchronization is lacking the pattern on the picture tube "rolls" vertically.

If the horizontal hold control is far out of adjustment, or is inoperative, the screen of the picture tube will be covered with a network of horizontal streaks and narrow lines, about as shown by Fig. 54-16.

When a vertical hold control is out of adjustment the pattern on the picture tube screen will move either up or down, either slowly or rapidly. To reproduce the effect would require a motion picture, but it is somewhat as shown by Fig. 54-17. The pattern or picture is said to be rolling or slipping vertically. As the hold control is turned more nearly to a correct adjustment the vertical movement will slow down, and eventually the picture will lock into its correct position.

If either a horizontal or vertical hold control will throw the picture out of synchronization at both extremes of the adjustment range you should note the positions of the knob or pointer at which the picture falls out of sync when the adjustment is turned first clockwise and then counter-clockwise. The best setting usually is midway between the two positions where synchronization is lost.

After making an approximate adjustment as just described, turn the receiver off and immediately back on again, or switch to another channel and back again to the first one. If the picture does not quickly pull into synchronization it may be possible to make a slight readjustment to improve the pull-in.

When a picture is synchronized, with the hold control, at a satisfactory setting, it will be necessary to turn the control some certain amount in either direction to lose synchronization. This is the point of drop-out. As the control is slowly brought back toward its original setting the picture will not return to syn-

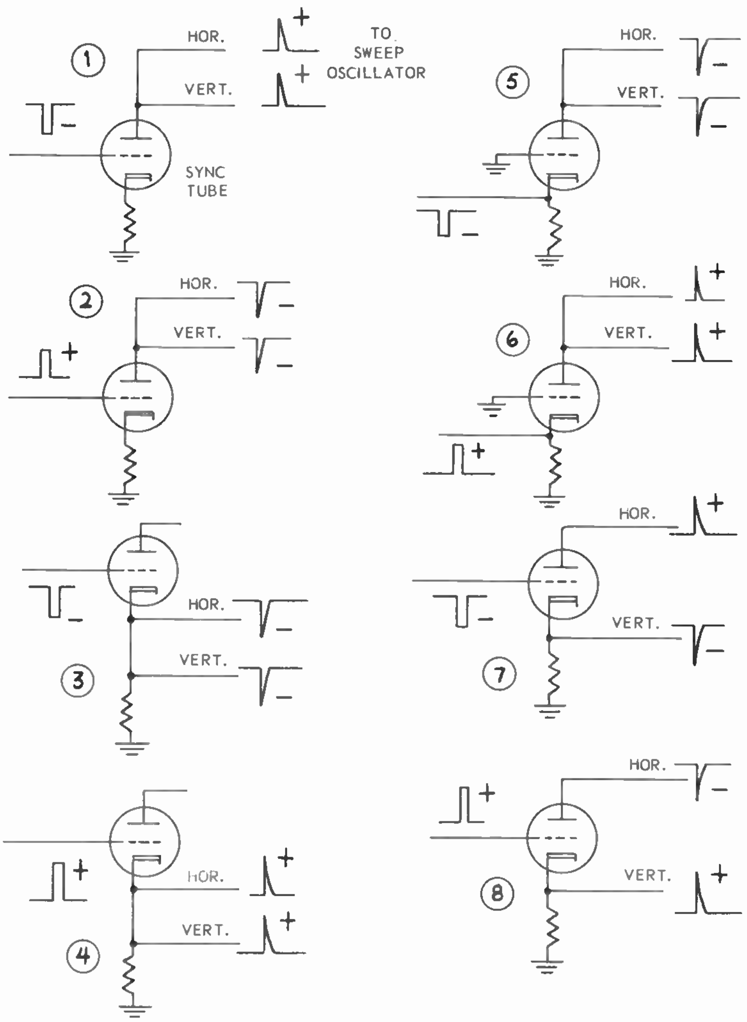


Fig. 54-18. Connections from the final tube in the sync section, as determined by required polarity for oscillator synchronization.

chronization at the point where it dropped out, but the control will have to be moved still farther toward its original correct setting. The position at which the picture again synchronizes is the pull-in point. There may be large differences between the points of drop-out and pull-in.

Once a hold control is correctly adjusted it should not require readjustment except at long intervals, as when tubes gradually change their characteristics after considerable use. If a hold control requires frequent adjustment it is probable that resistors or capacitors in the control circuit are changing their values by an excessive amount as the set warms up.

Hold control knobs or pointers may be readily accessible, in plain view of the operator, on the front panel. Otherwise these controls may be concealed by a small cover on the front panel, which is easily opened or removed for control adjustment. In still other receivers the hold control adjustments are on the back or top of the chassis, rather inaccessible except for service adjustments.

OSCILLATOR SYNC POLARITIES. As we have seen in this lesson and the one preceeding, sweep oscillators may require synchronization by either positive or negative pulse voltages. The blocking oscillator used alone or with a discharge tube requires positive pulses. Cathode coupled multivibrators, and the capacitor coupled type first examined, require negative pulses at their grids or positive pulses at their cathodes. The multivibrator in which one section is an output amplifier requires positive pulses at its grid or negative at its cathode. Different types of oscillators often are used for horizontal and vertical sweeps, so the polarity problem sometimes becomes rather involved.

Fig. 54-18 illustrates a few of the ways in which sync polarities required for sweep oscillators are obtained from the final tube in the sync section. Conventional symbols indicate polarities of sync pulses to grid or cathode of the sync tube and of output polarities to the two sweep oscillators. By remembering that there is inversion between grid and plate, but not between grid and cathode nor between cathode and plate, it should be easy to follow the diagrams with the help of the accompanying table.

POLARITIES TO SWEEP OSCILLATORS (Fig. 54-18)

Diagram No.	Input Pulse Polarity	Element to Which Applied	Element From Which Output Taken	Polarity To Sweep Oscillators Horizontal	Vertical
1	neg	grid	plate	pos	pos
2	pos	grid	plate	neg	neg
3	neg	grid	cathode	neg	neg
4	pos	grid	cathode	pos	pos
5	neg	cathode	plate	neg	neg
6	pos	cathode	plate	pos	pos
7	neg	grid	plate and cathode	pos	neg
8	pos	grid	plate and cathode	neg	pos

Fig. 54-19 illustrates a few combinations which involve two or more tubes at or near the output of the sync section. Input to the grid of the first tube in every diagram has been shown as negative. Were the

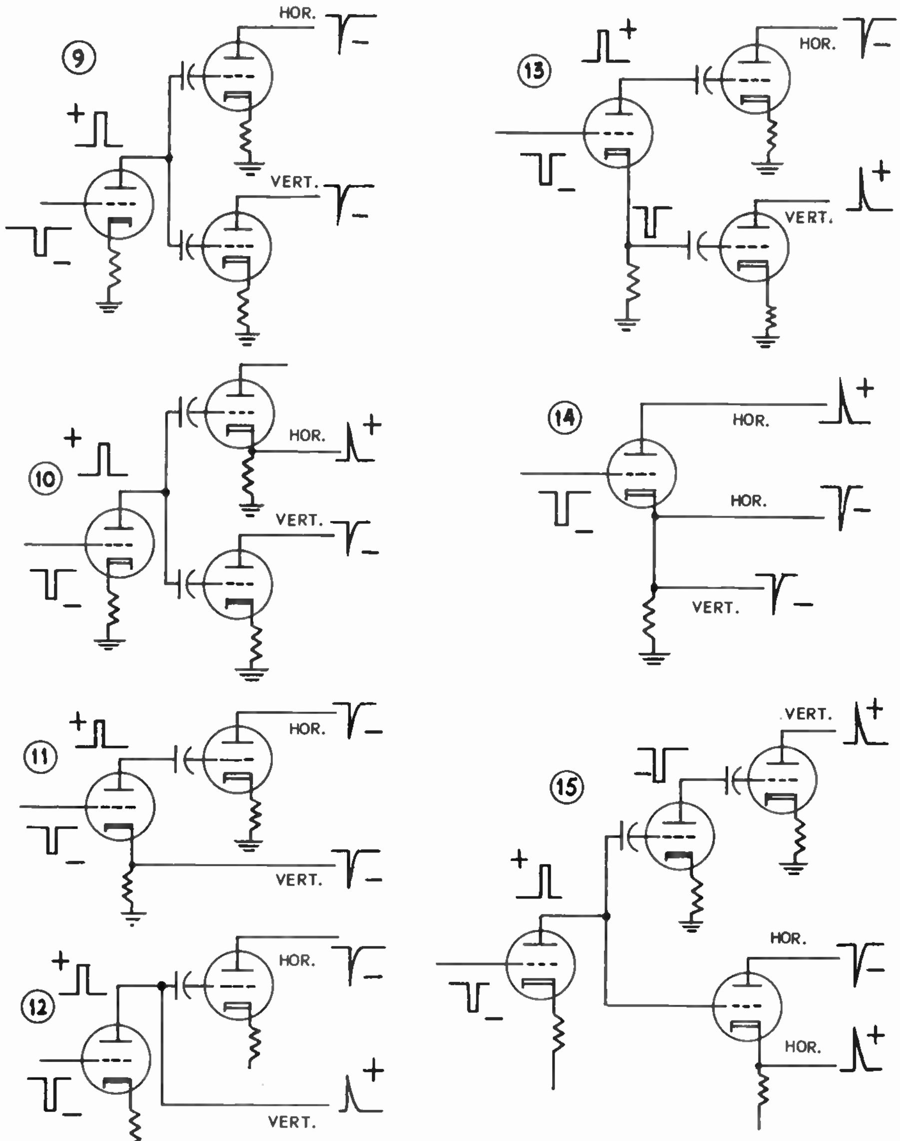


Fig. 54-19. Sync section output in which two or more tubes feed voltages to the following sweep oscillator circuits.

input for this first tube to be positive all the output polarities to the oscillators would be inverted. This became evident in Fig. 54-18, where in successive diagrams from 1 through 8 the grid inputs were shown alternately negative and positive, with accompanying reversals of outputs to the oscillators. The numbering of diagrams commenced in Fig. 54-18 is continued in Fig. 54-19.

In diagrams 14 and 15 there is a single vertical output, but there are two horizontal outputs of opposite polarities. These horizontal outputs are required for certain types of oscillator frequency control which we shall study in the following lesson.

TELEVISION

LESSON NO. 55

AUTOMATIC CONTROL OF SWEEP FREQUENCY

Training Your Mind
C. T. I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind

C.T.I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON NO. 55

AUTOMATIC CONTROL OF SWEEP FREQUENCY

In all oscillators of ordinary commercial construction there are temporary slight variations of operating frequency, due to equally slight and temporary changes of resistance, capacitance, and inductance that occur with heating, cooling, atmospheric changes, shifting of electric or magnetic fields, and other similar causes. Hold control resistors and capacitances are themselves subject to these disturbing influences, and cannot maintain sweep frequency at the oscillator output in perfect synchronization with pulses of the received signal.

These natural variations of frequency in vertical sweep systems are minimized because at a frequency so low as 60 cycles per second the oscillation circuits are quite stable in operation, and pulses from the sync section are able to hold the oscillator in time.

But in horizontal sweep systems the frequency is $262\frac{1}{2}$ times as high as in the vertical systems. Every change of capacitance or inductance, or of resistance which affects a time constant, will have proportionately greater effect in the horizontal sweep. The result, in spite of correct setting of the hold control, would be an irregular sidewise shifting of the picture decidely noticeable on screens of 10-inch and larger tubes.

This horizontal shifting of the picture could be compensated for by continual manipulation of the hold control, which manifestly is impracticable. What we need is a sort of automatic hold control that will instantly bring the sweep frequency back into synchronization when it tends to vary. Practically all receivers except those using 7-inch or smaller electrostatic deflection picture tubes do have automatic frequency controls for the horizontal sweep oscillator. A few have additional automatic frequency control for the vertical sweep oscillator.

There is great variety in design and in the exact methods of accomplishing desired results with the various afc (automatic frequency control) systems for sweep oscillators. Fortunately, however, our study of these systems is made relatively easy by the fact that all of them employ the same basic principle, which is illustrated by Fig. 55-1.

Between the sync section and the sweep oscillator is inserted a frequency control circuit. Voltage pulses from the sync section, which are at the correct sweep frequency, go to this control circuit. To the control circuit are applied also voltage pulses at the actual oscillator frequency, which are brought back from some point in or following the sweep oscillator output. The frequency of these feedback pulses will be incorrect if the oscillator is not in exact synchronization with the received signal.

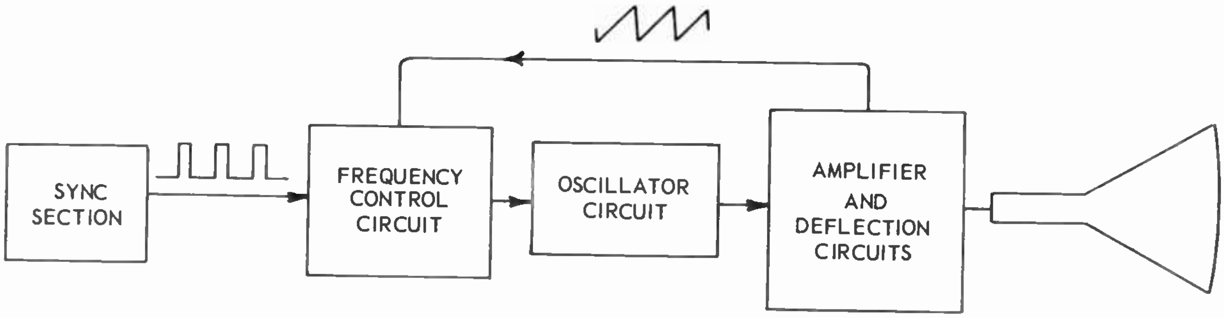


Fig. 55-1. Where the frequency control circuit is used.

The sync pulses and the actual sweep pulses act together in the control circuit to produce a combination voltage called the correction voltage. The correction voltage is applied to the input circuit of the sweep oscillator. If actual sweep frequency and received sync frequency are exactly the same, the oscillator frequency is unaffected. But if actual sweep frequency gets out of time with sync pulses the resulting correction voltage from the control circuit acts to lower or raise the oscillator frequency as may be required to bring it back into synchronization with the received signal.

The majority of afc systems utilize the correction voltage to alter the grid bias of the sweep oscillator. Making the bias of a blocking oscillator more negative lowers its frequency, because a longer time then is needed for the oscillator to return to a conductive condition after being blocked. Making the bias less negative increases the oscillator frequency, which is the actual sweep frequency.

If the oscillator is a cathode coupled or a capacitor coupled multivibrator, making the grid of the first section more negative will raise the frequency, since this is equivalent to making the second grid less negative or to lessening the resistance of the hold control. With the multivibrator which is combined with an output amplifier, making the oscillator grid more negative will lower the frequency, since this is the same effect as produced by increasing the hold control resistance.

Afc systems which do not act on oscillator grid bias employ what is called a reactance tube. The correction voltage then is used to alter the bias or grid voltage of the reactance tube. Thereupon the reactance tube changes the effective inductive reactance in the grid-plate circuit of the oscillator. This change of reactance has the same effect as a change of inductance in the grid-plate circuit, and oscillator frequency is altered accordingly.

All afc systems for sweep sections help to hold sweep frequency in time with sync pulses even when there is moderate interference of the noise type. Noise voltages, though they may be weak when reaching the sweep section, are likely to temporarily upset synchronization unless there is automatic frequency control.

We shall examine typical circuits of afc systems which are in general use. All of the commonly employed principles will be explained. Thereafter it should not be difficult for you to understand the effects of modifications or of new methods which later may be introduced.

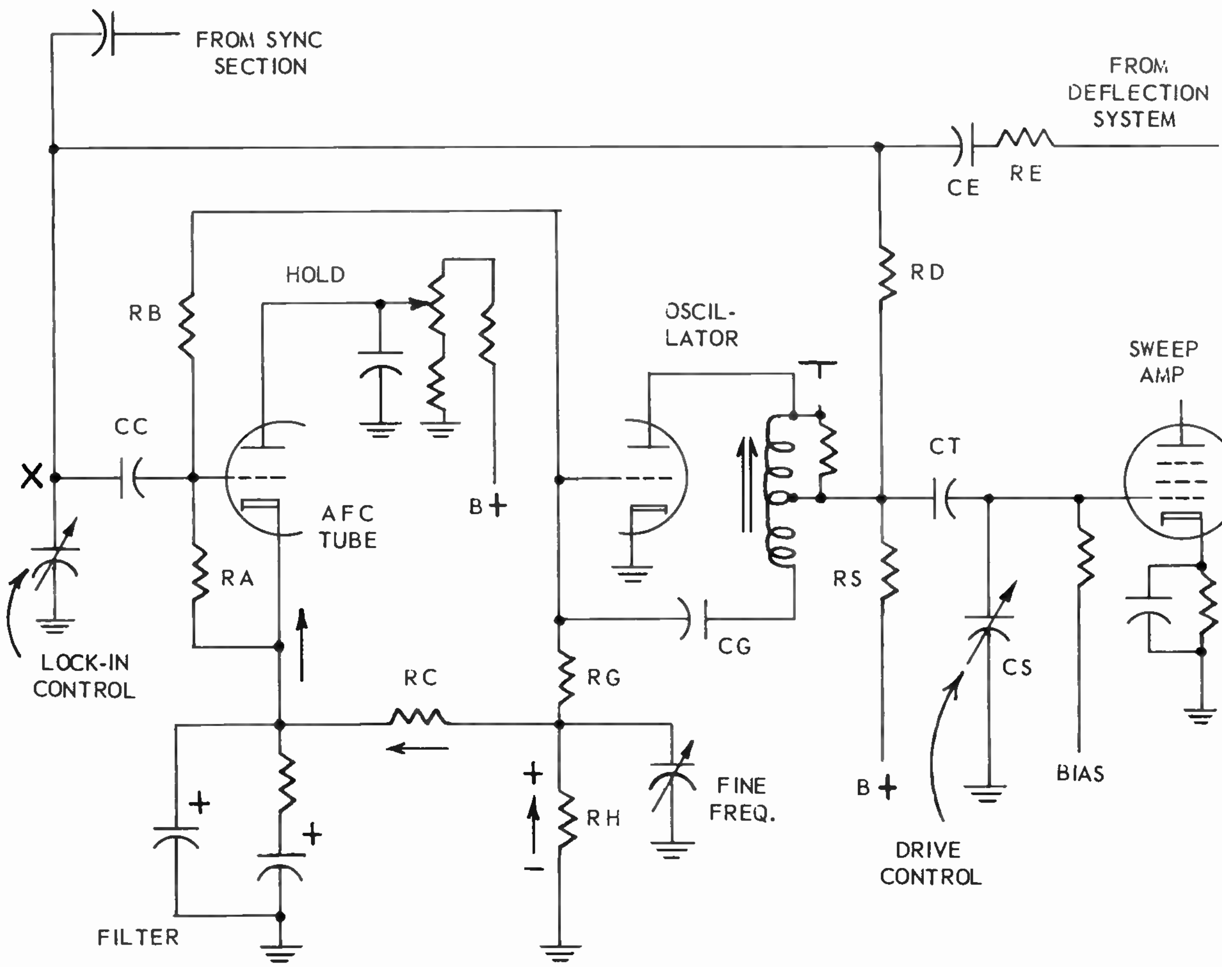


Fig. 55-2. An afc system using variable conduction in a triode to regulate sweep frequency.

AFC WITH TRIODE CONTROL TUBE. Fig. 55-2 shows circuit connections for a widely used afc system in which a triode tube controls the grid bias of a blocking oscillator.

Let's look first at the parts of this circuit with which we already are familiar. The oscillator feedback transformer, T, is an autotransformer type with continuous winding. One end is connected to the oscillator plate, and the other end is connected through capacitor Cg to the oscillator grid. A sawtooth output voltage appearing at the transformer tap is applied through capacitor Ct to the grid of the sweep amplifier tube.

The capacitance across which is developed the sawtooth voltage, by slow charge and rapid discharge, here consists of capacitors Cs and Ct in series with each other. These capacitors are charged by electron flow through them from ground (B-minus) and through resistor Rs to B-plus. The capacitors discharge through ground to the grounded cathode of the oscillator tube, through this tube to its plate, and from there through the upper part of the oscillator transformer and a paralleled resistor back to the capacitors. Discharge occurs when the oscillator grid is made positive.

The purpose of the resistor on the oscillator transformer is to provide a load that "damps" out oscillations which would tend to continue at the natural or resonant frequency of the oscillator feedback circuit. The portion of the sawtooth capacitance marked Cs is adjustable. It is called the drive control. The purpose of this control will be explained later.

The oscillator grid is connected to ground, and thus back to the cathode, through resistors Rg and Rh. With blocking oscillators having no afc one of these resistors would be adjustable, and would be the hold control for varying the free running frequency. Here we shall apply a correction voltage to the top of resistor Rh. This correction voltage will come through resistor Rc from the afc system, and will vary the oscillator grid voltage or grid bias to alter the frequency.

Note this: Although the parts of our oscillator circuit appear quite different from those on blocking oscillators previously examined, there really is nothing new except omission of the adjustable hold control and substitution of an afc voltage.

Voltage at the grid of the oscillator is shown at the left in Fig. 55-3. The grid goes far negative on the long downward traces, as the grid capacitor begins to discharge. While the capacitor continues discharging through resistors Rg and Rh the voltage becomes less negative along the gradual upward slope. Along this slope we find first a few cycles of oscillation at the natural frequency. These rapid oscillations are subdued or damped out by resistance in the grid circuit, and the capacitor discharge continues at a uniform rate. Where the oscillator grid voltage reaches the value allowing conduction we see brief upward peaks of positive voltage. It is during these positive peaks that the sawtooth capacitors Cs and Ct discharge through the oscillator tube.

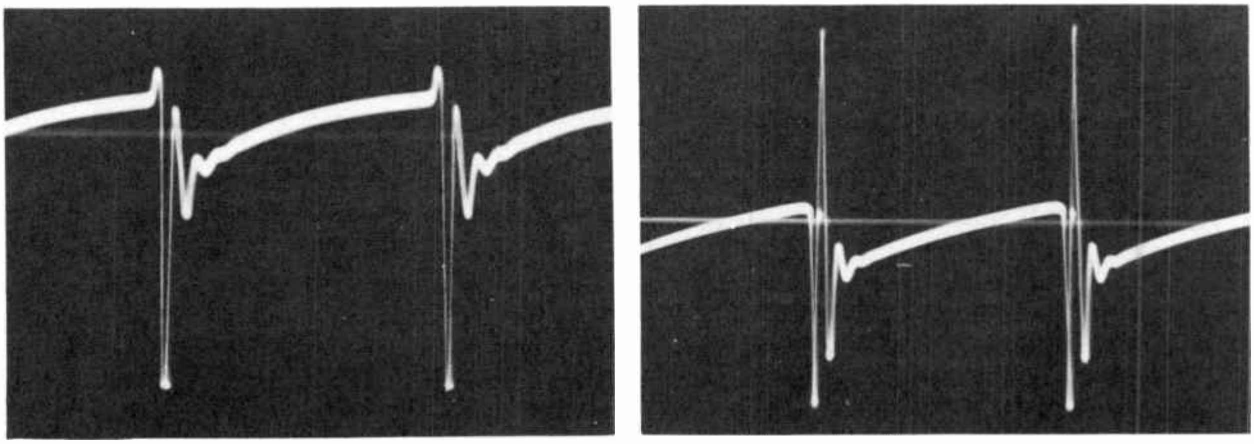


Fig. 55-3. Voltages at the grid (left) and at the plate (right) of the blocking oscillator.

Voltage at the oscillator plate is shown at the right in Fig. 55-3. Here again we have rapid fluctuations due to changes of grid voltage at the natural frequency, then a straightening of the upward slope of the curve.

Fig. 55-4 shows voltage at the tap or the output point of the oscillator feedback transformer, T. All the natural frequency oscillations which existed at the oscillator plate now have been removed by choking or filtering action of the transformer inductance. There is a smooth sawtooth wave, except for rather sharp negative downward peaks just before the curve commences its gradual upward slope to the right.

The negative peaks at the beginning of the upward slope in the sawtooth wave are due to resistance in the discharge path of the sawtooth capacitors. This resistance prevents complete discharge of the sawtooth capacitors during the period in which the oscillator is conductive. The remaining charge of the capacitors brings the voltage upward by a small amount from its extreme negative value as soon as the oscillator is cut off. Then follow the smooth upward slope indicating recharging of the sawtooth capacitors. Later we shall have much more to say about this negative peaking, because it is of great importance in securing correct deflection of the picture tube beam.

Having determined with the help of the oscilloscope that our oscillator really is operating like any other blocking oscillator, we shall consider what is happening in the afc portion of the circuits.

Note first in Fig. 55-2 that the grid of the afc tube is connected to the oscillator grid through resistor Rb, whose value is about 3 megohms. Through this resistance there is applied to the afc grid, part of the negative biasing voltage which exists at the oscillator grid. The grid return path to the cathode is through

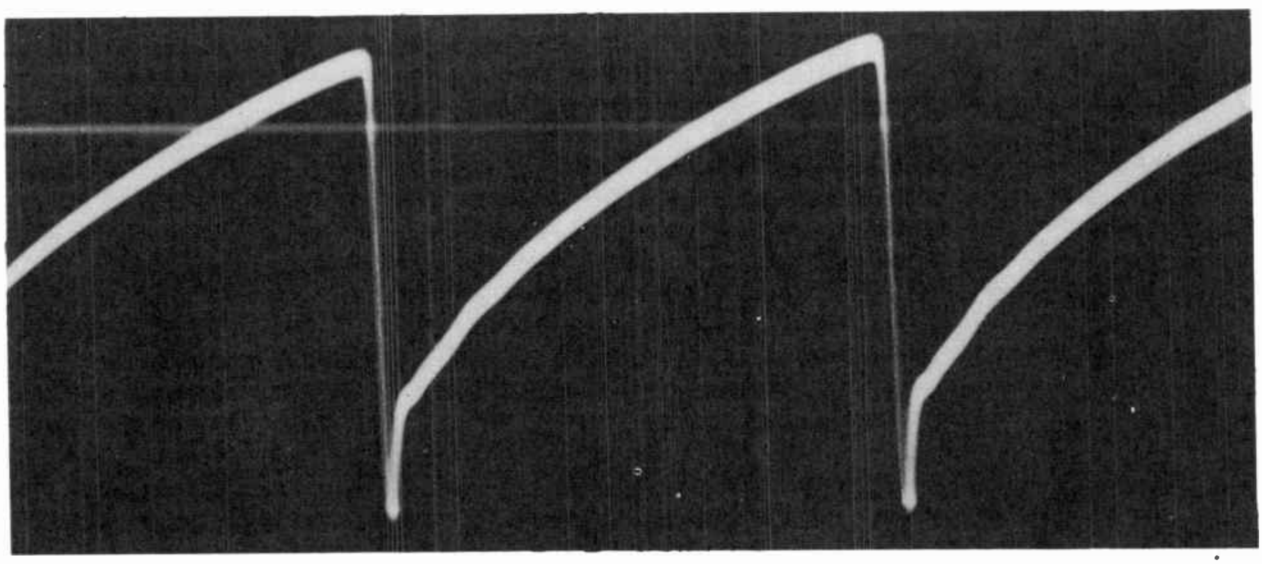


Fig. 55-4. Voltage at the transformer output and the sawtooth capacitors.

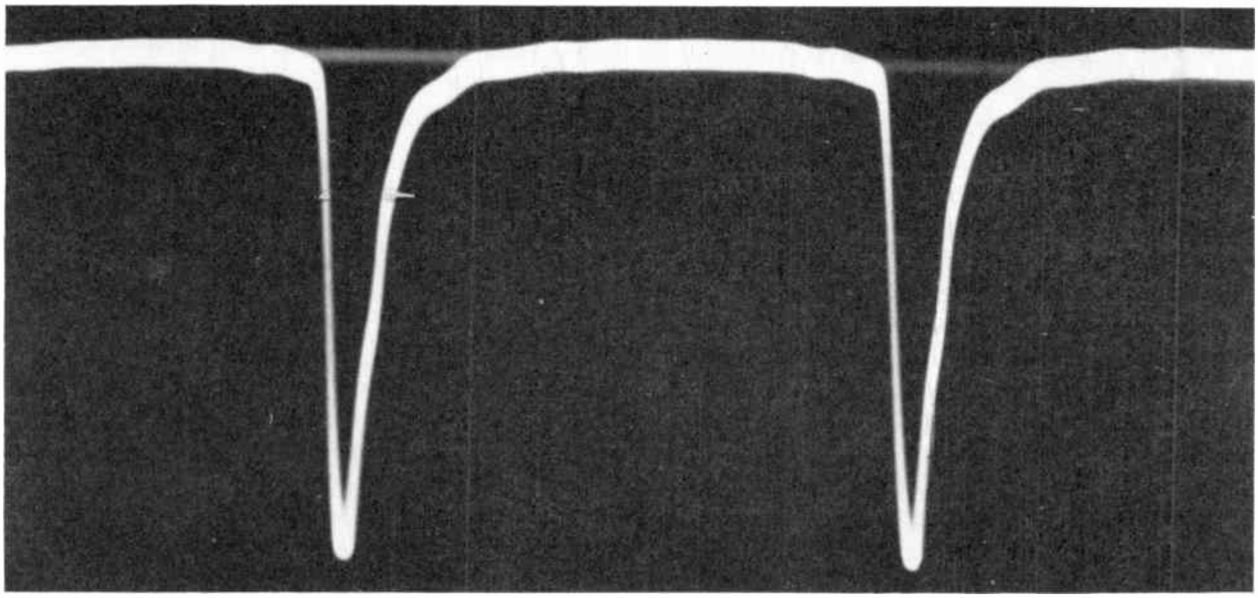


Fig. 55-5. Negative pulses from the deflection circuit, used in forming the combination waveform at the afc grid.

resistor Ra, whose value is considerably less than one megohm. Thus the afc tube is negatively biased to an extent which allows conduction only during the more positive portions of the other voltages which come to the grid.

Coming to the grid of the afc tube through capacitor Cc are three voltages, which combine to produce the frequency correction effect. One of these voltages is the sawtooth wave at the oscillator output, shown by Fig. 55-4. This voltage comes through resistor Rd to point X and through capacitor Cc to the grid.

Another of the three voltages is taken from the deflection system, from some point in the circuit for the horizontal deflection coils for the picture tube. This voltage, as it exists in the deflection circuit, consists of sharp negative pulses. These voltages are shown by the oscilloscope as in Fig. 55-5. These deflection pulses come to point X and the afc grid through a resistor and capacitor at Re and Ce. The resistor is of about a half-megohm and the capacitor of less than 10 mmf, to greatly reduce the high voltage of the deflection pulses and make them somewhat narrower.

The sawtooth voltage from the oscillator transformer and the negative pulses from the deflection circuit combine at the grid of the afc tube to give the waveform shown by Fig. 55-6. The upward slopes of this wave are from the sawtooth output of the oscillator and the negative dips are from the deflection circuit. Since everything that contributes to formation of this wave comes from the output side of the oscillator, the frequency must be the actual sweep frequency at which the oscillator is operating. If actual sweep frequency increases there will be an increase of the frequency in the wave of Fig. 55-6, and a decrease of sweep frequency will be reflected in decreased frequency of the combination wave.

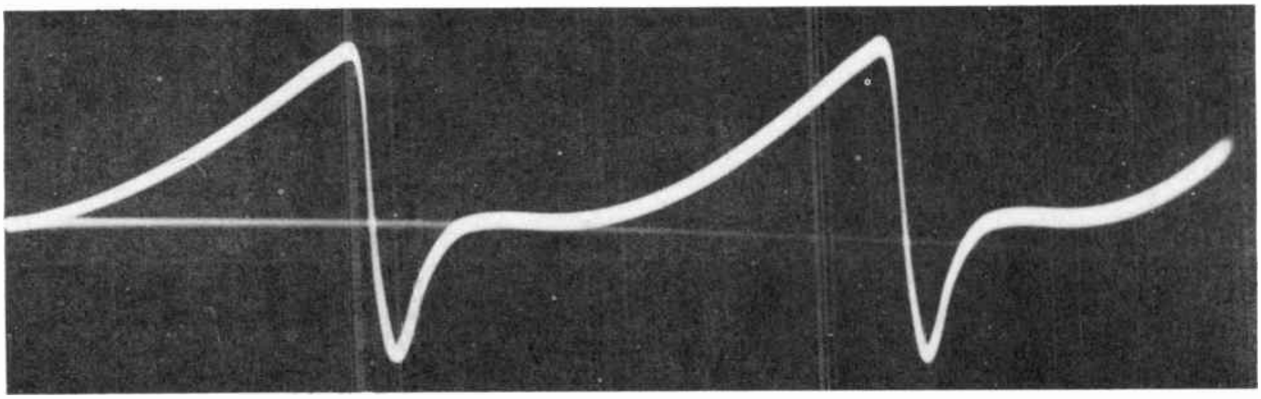


Fig. 55-6. The combination voltage at the afc grid before sync voltage pulses are added.

The third voltage to be applied at the afc grid is taken from the last tube in the sync section. This sync voltage is shown by Fig. 55-7, as it appears at the output of the last sync tube. Here we have positive pulses whose frequency is the horizontal sync frequency of the received signal, and which will remain at this correct frequency no matter what happens to the frequency of the combination wave derived from the sweep output.

When the sync voltages add themselves to the combination wave at the afc grid we have the effect shown by Fig. 55-8. The positive sync pulses add small positive peaks at the end of each upward slope in the combination wave. It is only these positive peaks that cause conduction in the afc tube, because because of the negative bias mentioned previously. You should compare this trace with that of Fig. 55-6.

5 With any given plate voltage applied to the afc tube the length of conduction time and the resulting total quantity of electrons passing through this tube will depend on the width or the time duration of the positive peaks that appear at the tops of the combination wave.

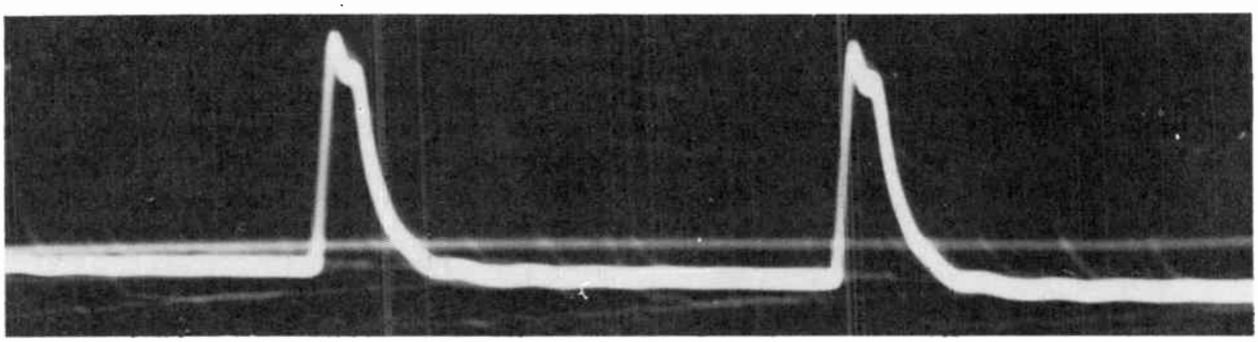


Fig. 55-7. Voltage pulses from the sync section, which are added to the combination voltage at the afc grid.

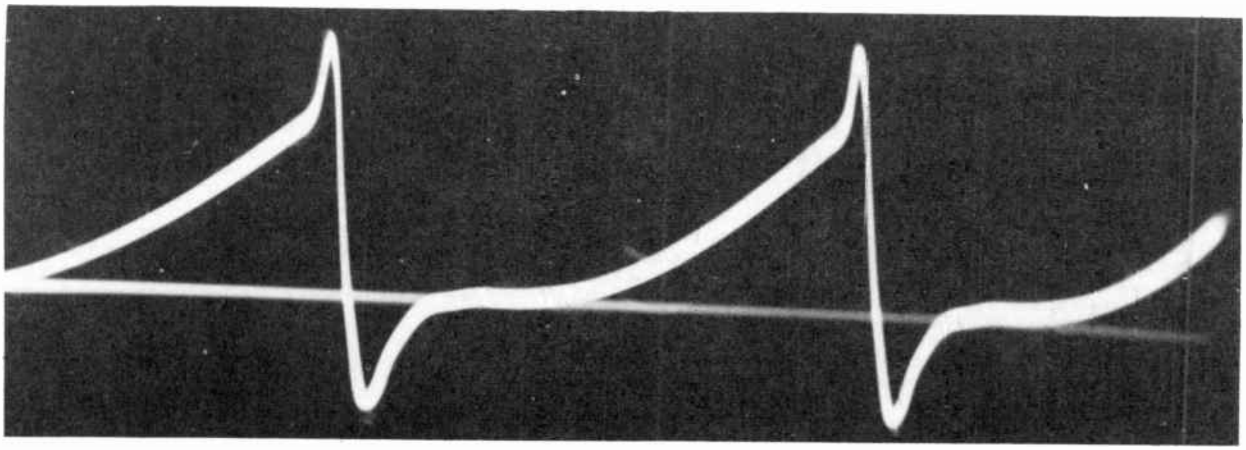


Fig. 55-8. With correct synchronization the combination voltage has positive peaks of moderate width.

What happens to the positive peaks when there is a change of actual sweep frequency is illustrated by diagrams in Fig. 55-9. At the top we have relations between sync voltage and combination voltage for correct synchronization of oscillator frequency. About half of the total length of time for the sync voltage is before the sharp negative drop of combination voltage. Half the time duration of sync voltage remains on top, to cause conduction in the afc tube. The remainder of the time of the sync voltage occurs later than the peak, and this portion of the sync voltage adds itself to the combination wave away down in the negative dip where it can cause no conduction in the afc tube. This is the condition shown by the oscilloscope in Fig. 55-8.

In the middle diagram of Fig. 55-9 the oscillator has slowed down, its frequency has become lower. Then each sharp peak of the combination voltage occurs a little later than before. The timing of sync pulses has not varied, and as a result the entire time of sync voltage occurs before the negative dip. The afc tube is made conductive for a longer time in each cycle by the decrease of oscillator frequency. This condition is shown by the oscilloscope as at the left in Fig. 55-10, where it is apparent that the positive peaks added to the combination wave are lasting for a much longer time than before.

In the lower diagram of Fig. 55-9 the oscillator is running faster, its frequency has increased. Now the sharp peaks on the combination wave occur earlier. Only a brief fraction of time duration of the sync voltage occurs before the combination voltage goes negative, and only a very little of the sync voltage remains to cause conduction in the afc tube. The afc tube has been made less conductive, or conductive during briefer intervals in each cycle by increase of oscillator frequency. The condition represented by this lower diagram is shown by the oscilloscope as at the right in Fig. 55-10. The positive peaks have all but disappeared.

Plate current or cathode current for the afc tube flows from ground (B-minus) through resistors Rh and Rc of Fig. 55-2, thence through the tube and the hold control resistances to B-plus. Voltages which accompany the current in Rh and Rc charge the capacitors in the filter system connected to the afc cathode. The time constant of the filter capacitors and resistors Rh and Rc is such as to remove rapid fluctuations

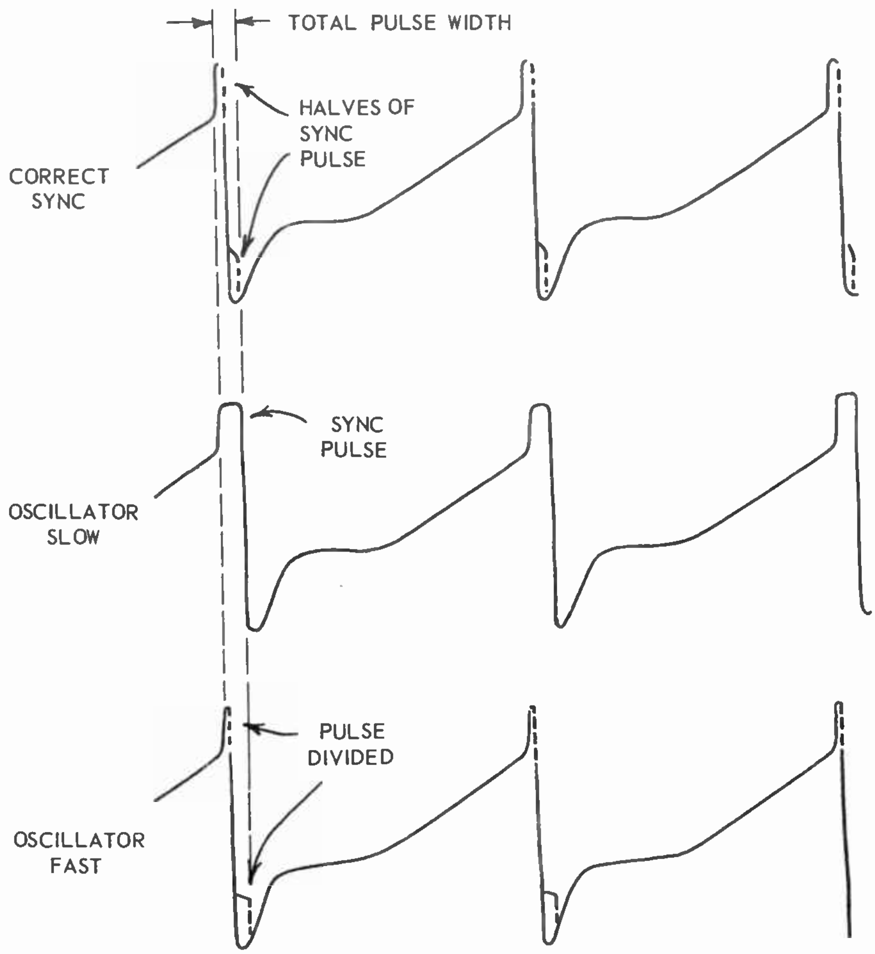


Fig. 55-9. How the positive sync voltage pulses combine with the sawtooth wave which has a negative peak.

of voltage. The charges built up on the filter capacitors are positive on the sides connected through resistor Rc to the oscillator grid circuit. Then the greater the charge put onto the filter capacitors the greater will be the positive correction voltage applied to the oscillator grid circuit, and the less negative will be the bias on the oscillator grid.

Capacitor charge and positive correction voltage is increased by longer conduction periods of the afc tube with a slow running oscillator, as shown by the middle diagram of Fig. 55-9. Then the less negative grid voltage for the oscillator increases the oscillation frequency. When capacitor charge and voltage decrease with shorter conduction periods of the afc tube with a fast running oscillator (bottom diagram of Fig. 55-9) the oscillator grid remains more negative. This decreases or slows down the oscillation frequency. Thus we have automatic frequency control which holds oscillation and sweep frequency in time with the received sync pulses.

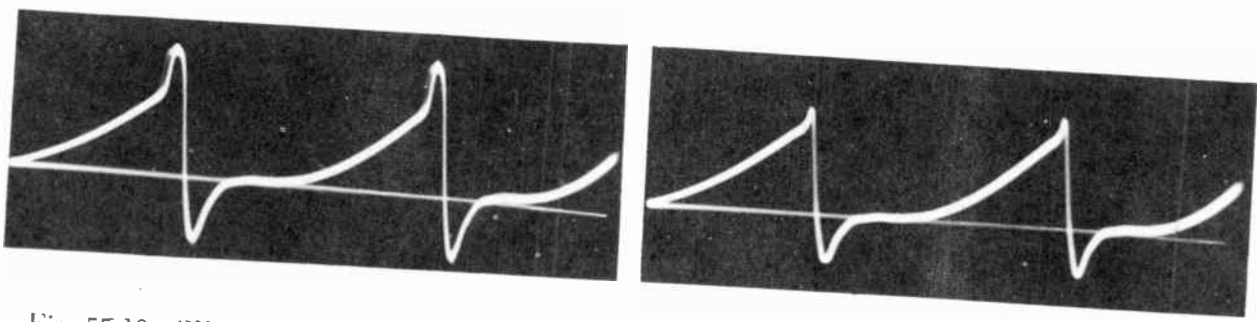


Fig. 55-10. Effect on positive voltage peaks of a slow oscillator (left) and of a fast oscillator (right).

The hold control of this afc system varies the plate voltage on the afc tube. Higher plate voltage increases the current during each period of conduction. This puts a greater charge on the filter capacitors, makes the oscillator grid less negative, and increases the oscillation and sweep frequency. If the hold control is adjusted for less plate voltage there is less current during each conduction period, a smaller charge on the filter capacitors, a more negative voltage on the oscillator grid, and there is decrease of oscillation and sweep frequency.

There are four adjustments or controls on this afc system. The hold control is accessible to the operator. The other three are service adjustments. One of these service adjustments is the movable core in the oscillator feedback transformer. This core varies the free running frequency to make it enough slower than sync frequency that the hold control may be effective for synchronization.

The fine frequency control is an adjustable trimmer capacitor on the oscillator grid circuit. It allows bringing the free running frequency to a more nearly correct value when the transformer core is not accurately adjusted. This control capacitor is not always included in afc systems of this general type.

The lock-in control on the grid circuit of the afc tube is another trimmer type adjustable capacitor. It acts as a partial bypass to ground for the voltages reaching the afc grid. Adjustment for less capacitance increases the reactance of this capacitor, reduces the bypassing effect, and there is increase of amplitude of the wave at the afc grid. More capacitance decreases the amplitude.

Adjustment of the lock-in determines the effectiveness of the hold control in synchronizing the picture. If the lock-in capacitor is correctly adjusted and the hold control is at the approximate center of its adjustment range, a picture will almost instantly pull into synchronization when the set is switched from channel to channel, or when it is turned off and on again. After a synchronized picture appears, the hold control may be turned from end to end of its adjustment range or very nearly through the entire range without losing the synchronization that has been established. If no adjustment of the lock-in capacitor allows this behavior of the hold control it indicates incorrect adjustment of the core in the oscillator transformer or of the fine frequency control when this latter control is used.

TRIODE AFC WITH OSCILLATOR-AMPLIFIER. Another method of using a triode for producing the correction voltage in an afc system is illustrated by the diagram and waveforms of Fig. 55-11. Here the sweep oscillator acts also as a sweep amplifier, feeding its output through an impedance matching transformer to the deflection coils on the picture tube.

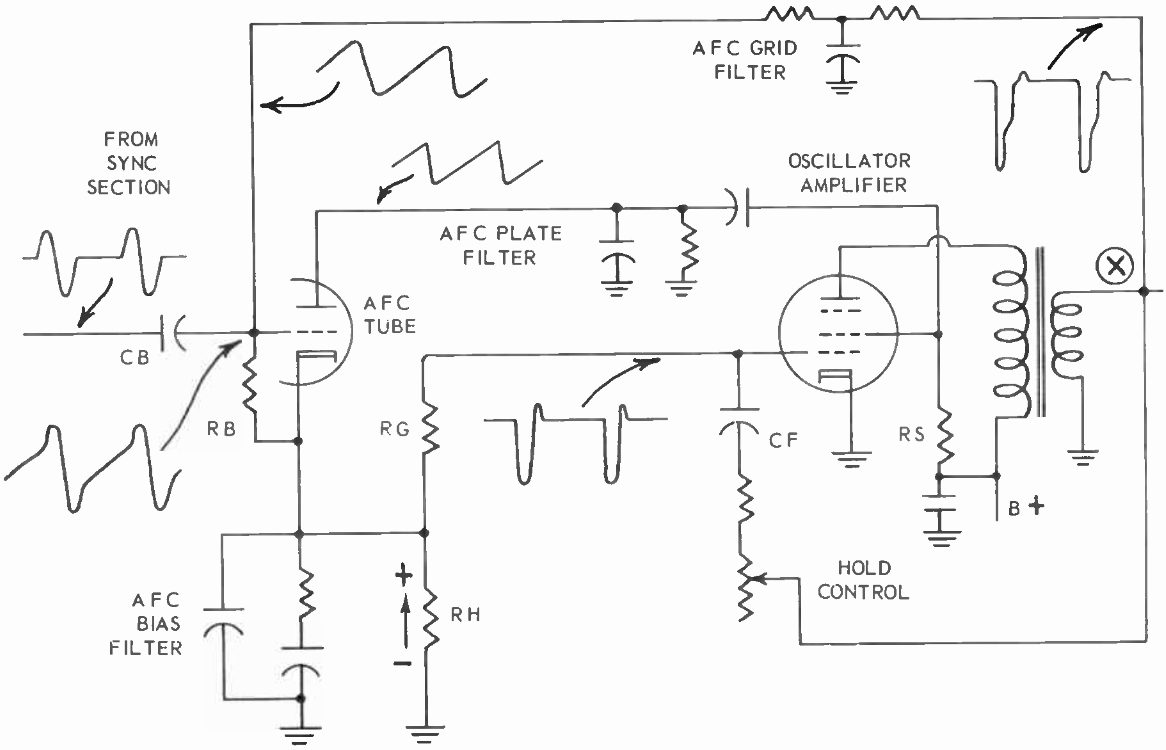


Fig. 55-11. An afc system controlling a combined oscillator-amplifier.

The oscillator is essentially a blocking type. The primary of the output transformer acts much like the plate winding of the feedback transformer in blocking oscillator circuits studied earlier. The secondary acts much like the grid winding of a feedback transformer.

Assume to begin with, that the oscillator-amplifier becomes conductive during its operating cycle. There is a momentary large rush of current from cathode to plate and through the transformer primary to B-plus. The quick change from zero current to relatively large current induces a strong emf in the secondary winding of the transformer. The upper end of the secondary, marked X, becomes strongly positive while the lower grounded end becomes negative.

There is a connection from point X through the hold control resistance and capacitor Cf to the grid of the oscillator-amplifier, so the grid of this tube is also driven positive. The time constant of the circuit including capacitor Cf is short, allowing this capacitor to discharge very quickly after being charged by the voltage pulse which we originally assumed as having made the oscillator-amplifier conductive to start the action.

Although plate current through the transformer primary began with a sudden rush, it did not rise to a high value, because of the strong counter-emf induced in the transformer primary. Thereafter the plate current increased at a relatively slow rate in accordance with the time constant of inductance and winding resistor in the transformer primary. This gradual rise of plate current is the rising portion of the sawtooth wave which is required for deflection. Note that here we are not forming the sawtooth by charge and discharge of a capacitor, but by the build-up and decay of a magnetic field in the output transformer.

When plate current in the oscillator amplifier approaches its maximum value, as determined by applied B-voltage, the rate of increase slows down. This reduces the induction of emf in the secondary winding, and soon this emf becomes too small to hold the grid of the oscillator amplifier positive. Grid voltage then drops to the cutoff value and goes negative. At the instant of plate current cutoff there is sudden stoppage of current in the transformer primary. The change from increase to very fast decrease is the steep downward part of the sawtooth wave.

Note that this much of the performance of the oscillator-amplifier is very similar to that earlier described in connection with more common types of blocking oscillators.

Now we shall go to the afc section of Fig. 55-11. The only voltage applied to the plate of the afc tube is a sawtooth wave obtained from the screen circuit of the oscillator-amplifier and brought through the afc plate filter elements. As oscillator plate current is increasing, the d-c voltage at the plate is decreasing, as always happens. The screen voltage will also change, at the same rate, due to the unbypassed resistor Rs. This is the change of screen voltage that is taken to the plate of the afc tube. It is important to note that the frequency of this sawtooth plate voltage for the afc tube is the actual sweep frequency being applied to the deflection system. It is the frequency which may vary in relation to sync frequency, and which must be brought back into synchronization.

Two voltages are applied to the grid of the afc tube. One of these is a sawtooth voltage derived from point X in the deflection circuit and brought back through the two resistors and capacitor of the afc grid filter. The capacitor in this filter is rapidly charged by the strong negative pulses from X. This negative change of capacitor voltage forms the steep downward slope of the sawtooth applied to the afc grid. Then the filter capacitor discharges rather slowly through the resistance paths, forming the gradual upward slope of the sawtooth wave.

The other voltage applied to the afc grid comes from the sync section of the receiver. This sync voltage consists of positive and negative peaks occuring at the beginning of each cycle, with zero voltage between times.

The combination of the two voltages at the grid results in a sawtooth wave with positive peaks at the beginning of each downward slope, and negative peaks at the end of this slope. We shall make use of the positive peaks of grid voltage. They are occuring at the sync frequency of the received signal. The first use of the positive peaks is to bias the grid of the afc tube by grid-leak biasing action of capacitor Cb and resistor Rb.

If the sweep oscillator is correctly in time with the received signal the positive peaks of afc grid voltage will occur at the instant wherein the sawtooth plate of the afc tube voltage has dropped half way from maximum to zero. This relation of sync pulses to plate voltage is shown at the top of Fig. 55-12.

Supposing now that oscillation frequency decreases, the oscillator commences to run too slow. The downward slope of plate voltage will occur later than it should. The result is shown at the middle of

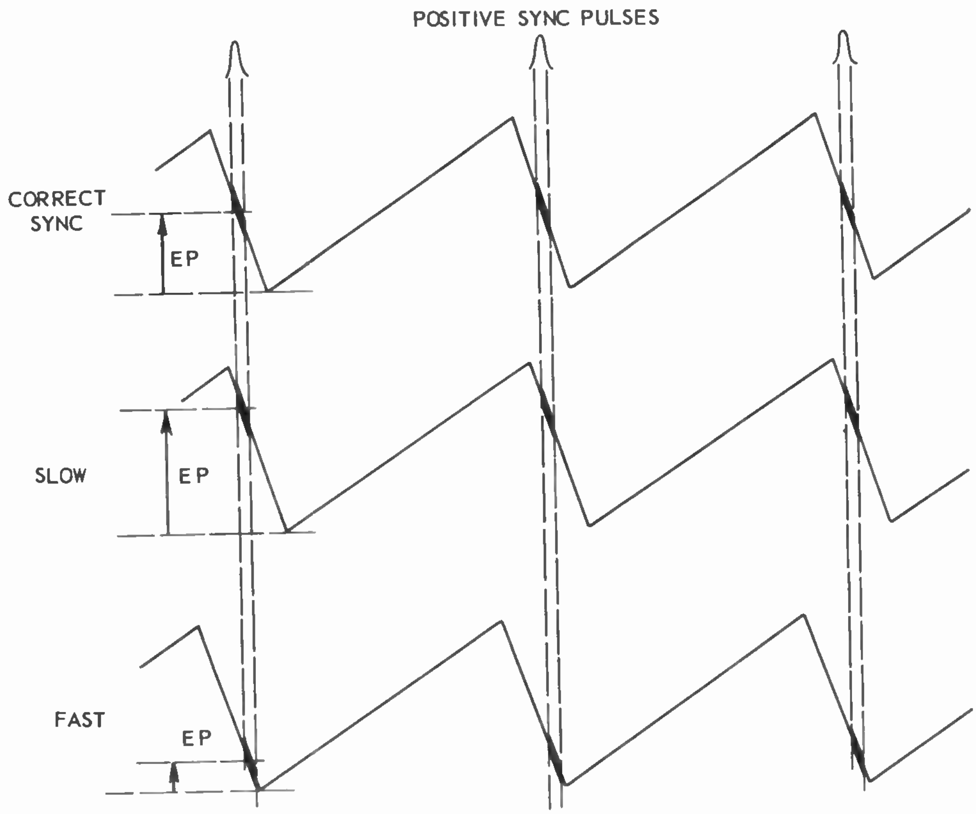


Fig. 55-12. How oscillator frequency varies the point on the sweep-frequency wave at which there are sync pulse voltages.

Fig. 55-12. Since the sawtooth wave occurs later in time, the sync pulses are relatively earlier and they occur while plate voltage is higher than in the first condition. This higher plate voltage causes increased plate and cathode current through the afc tube during each of the sync pulses.

Should oscillation frequency increase, or become too fast, we have the condition shown at the bottom of Fig. 55-12. Now the positive sync pulses occur when there is much lower plate voltage on the afc tube. There is less plate and cathode current during each sync pulse.

By looking at the cathode circuit of the afc tube in Fig. 55-11, and comparing it with the cathode circuit of the afc tube in Fig. 55-2, you could write the remainder of this explanation for yourself. Greater plate-cathode current with a slow running oscillator increases the charge and the voltage on capacitors in the bias filter. This makes for a more positive voltage at the top of resistor Rh. This resistor is in the oscillator grid circuit, so oscillator grid bias is made less negative (the same as more positive) and oscillator frequency is increased. Less plate-cathode current in the afc tube has exactly opposite effects. The oscillator is slowed down to overcome the fast action that reduced the plate-cathode current.

AFC CORRECTION VOLTAGE FROM DISCRIMINATOR. The discriminator circuit which we used for a detector or demodulator in f-m sound receivers quite often is used for developing a correction voltage in afc systems for sweep oscillators. Doubtless you recall that the discriminator produces an output voltage that goes positive and negative when the frequency of the f-m signal becomes greater or less. In the sound receiver the output voltage of the discriminator becomes alternately positive and negative at an audio frequency rate, thus forming the audio signal.

In the sound demodulator we obtain a reference frequency from the transformer primary or from the plate of the last i-f amplifier. In an afc system we may use the sync frequency as our reference. In the sound system we obtain from the ends of the transformer secondary two voltages whose phase relation to the reference voltage changes with changes of signal frequency. In an afc system the voltage or voltages which shift their phase (or their frequency) may be secured from the output of the sweep oscillator, and will be the actual sweep frequency. Discriminator output then will become positive when actual sweep frequency varies in one direction, and become negative when sweep frequency varies in the other direction. This positive or negative output voltage is used in various ways to bring the oscillator back into synchronization.

The discriminator circuit shown by diagram 1 of Fig. 55-13 looks much like some of the sound demodulator circuits. Sync pulses here are applied to the center of the transformer secondary. The diode circuits

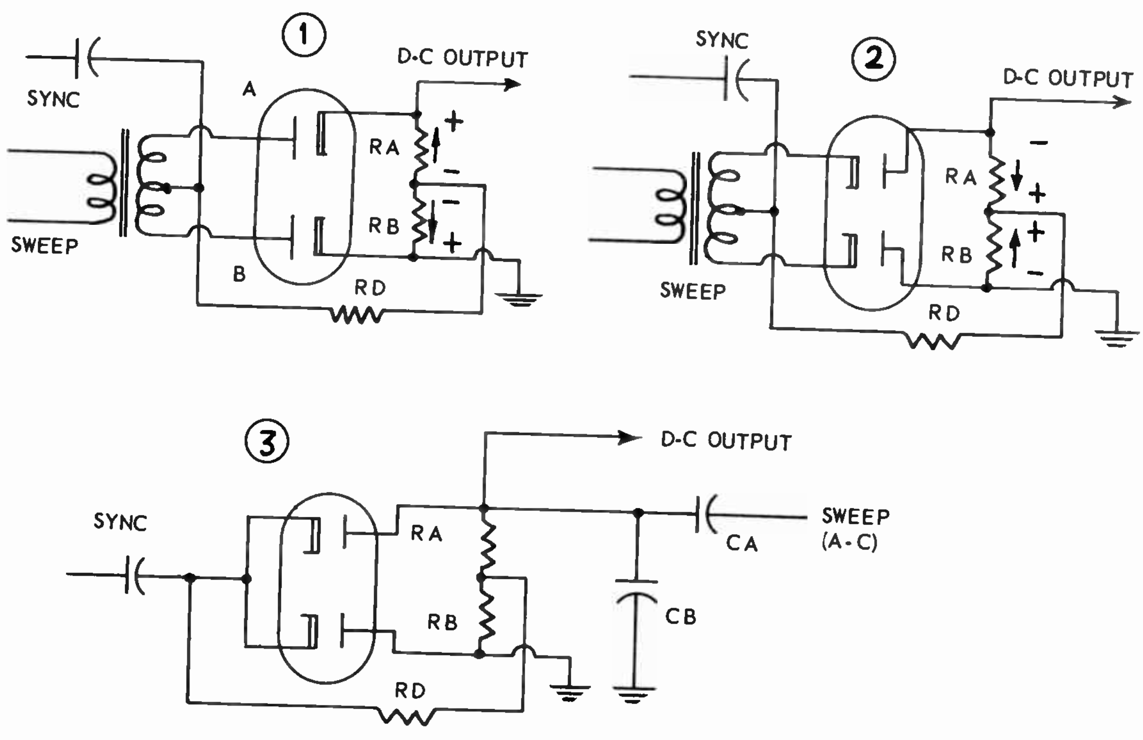


Fig. 55-13. Some of the discriminator circuits employed in afc systems.

are completed through resistor Rd. Sweep voltage is applied to the transformer primary, and goes in opposite phase from the secondary to the plates of the two diodes. The diodes become alternately conductive when alternations of sweep voltage make their plates positive. Conduction currents flow in opposite directions through load resistors Ra and Rb. Output is taken from the top of resistor Ra. Anything which causes diode A to conduct more than B makes the output positive, while anything that increases conduction in diode B over that in A makes the output negative.

Except for using sync pulses and sweep voltage, the explanation in the preceding paragraph would apply to a sound discriminator. If you do not remember details of discriminator performance it would be well to read about them in the lesson dealing with demodulation in f-m receivers.

If we reverse the connections to plate and cathode of each diode, the discriminator circuit appears as in diagram 2 of Fig. 55-13. Conduction currents have been reversed in the load resistors. Now an increase of conduction in diode A makes the output negative, while increase of conduction in diode B makes the output positive.

Generally similar results may be had with the circuit modification of diagram 3. Only the sync pulses, which must be of negative polarity, are applied to the cathodes of the two diodes. The alternating sweep voltage is brought through capacitor Ca and appears across capacitor Cb. The top of Cb is directly connected to the top of the load resistors, and the lower end is connected through ground to the bottom of the resistors. Thus the sweep voltage appears across the resistors and is applied to the plates of both diodes.

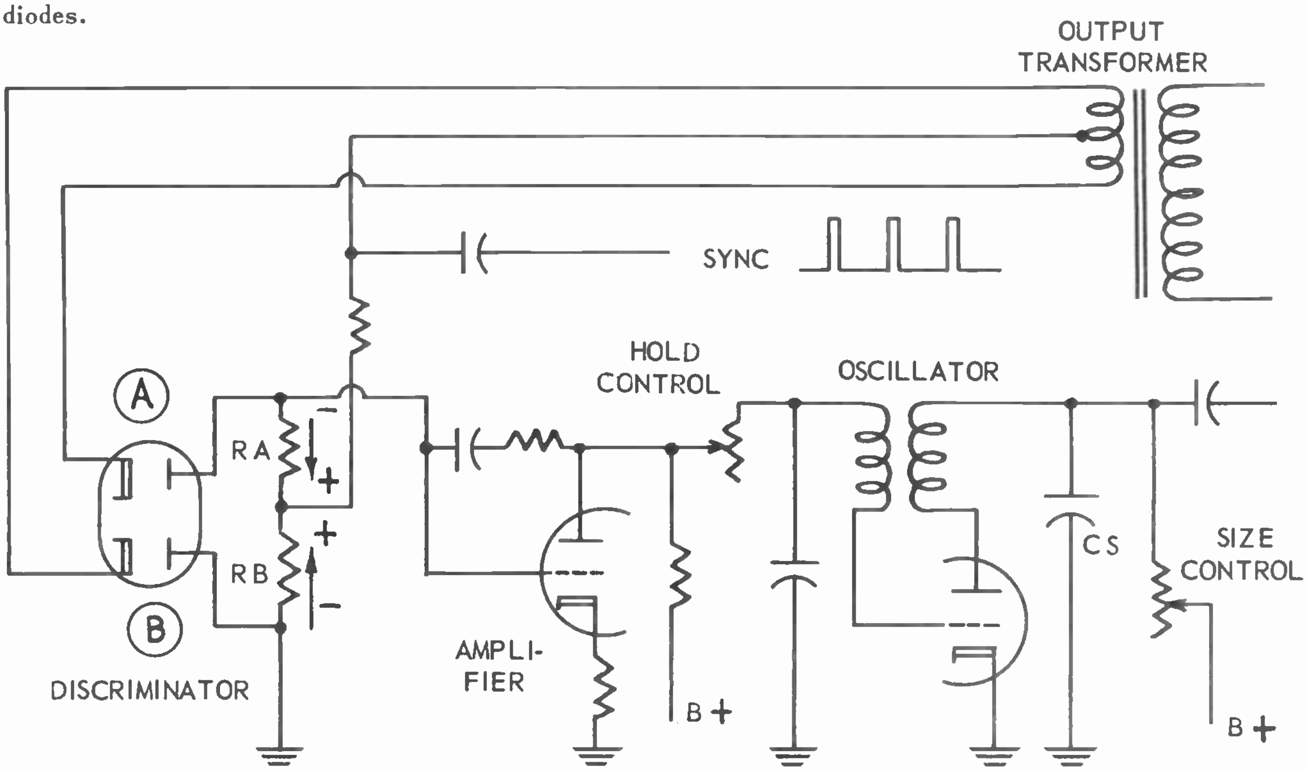


Fig. 55-14. A discriminator varying the plate resistance of an amplifier as a means for controlling oscillator frequency.

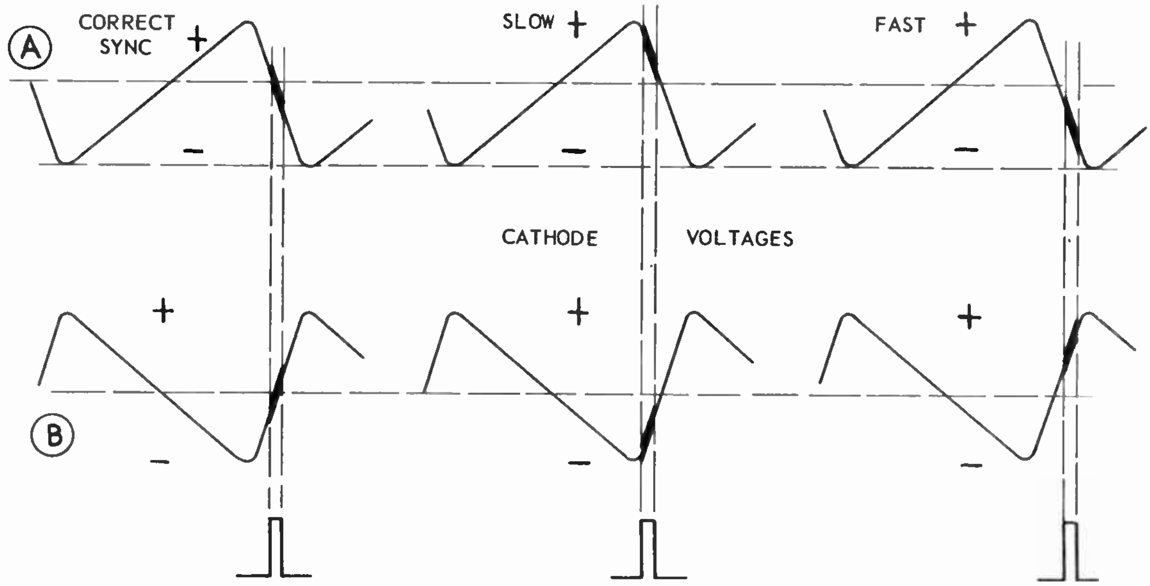


Fig. 55-15. How discriminator diodes are made more or less conductive by shifting of the sweep frequency.

Fig. 55-14 shows one of the many applications of discriminators in afc systems for sweep oscillators. Here we have the discriminator acting as the afc tube, then an amplifier to strengthen the d-c output of the discriminator, then a blocking oscillator in whose grid circuit we use the amplifier plate resistance for varying the oscillator bias and its frequency. There is nothing about the oscillator itself with which we are not already familiar.

The cathodes of the discriminator diodes are connected to the two ends of a winding which is on the core of the sweep output transformer. The voltage coming from this winding to the discriminator is of sawtooth waveform. Since alternating potentials from opposite ends of any transformer winding are of opposite phase, the voltages on the discriminator cathodes are of opposite phase. Fig. 55-15 shows these opposite simultaneous voltages as they are applied to the cathode of diode A and to the cathode of diode B. One cathode is made negative while the other is made positive. Each diode conducts only while its cathode is negative. These polarities are, of course, with reference to the diode plates.

To the center tap of the load resistors Ra and Rb on the discriminator are applied positive voltage pulses from the sync section of the receiver. These positive voltage pulses pass through each of the load resistors to the respective plates of the discriminator. Thus the two plates are made positive at the same time. Now we must realize that, even though the diode plate is made moderately positive by a sync pulse, the diode cannot conduct unless its cathode is negative with reference to the plate.

If sawtooth waves from the deflection system are in correct time relation to sync pulses from the received signal we have the condition shown at the left in Fig. 55-15. The pulses act just when the diode cathodes are going through zero voltage. If there is any small conduction it will be equal in both diodes. Currents in the load resistors will balance, since they are in opposite directions, and there will be no d-c output from the discriminator.

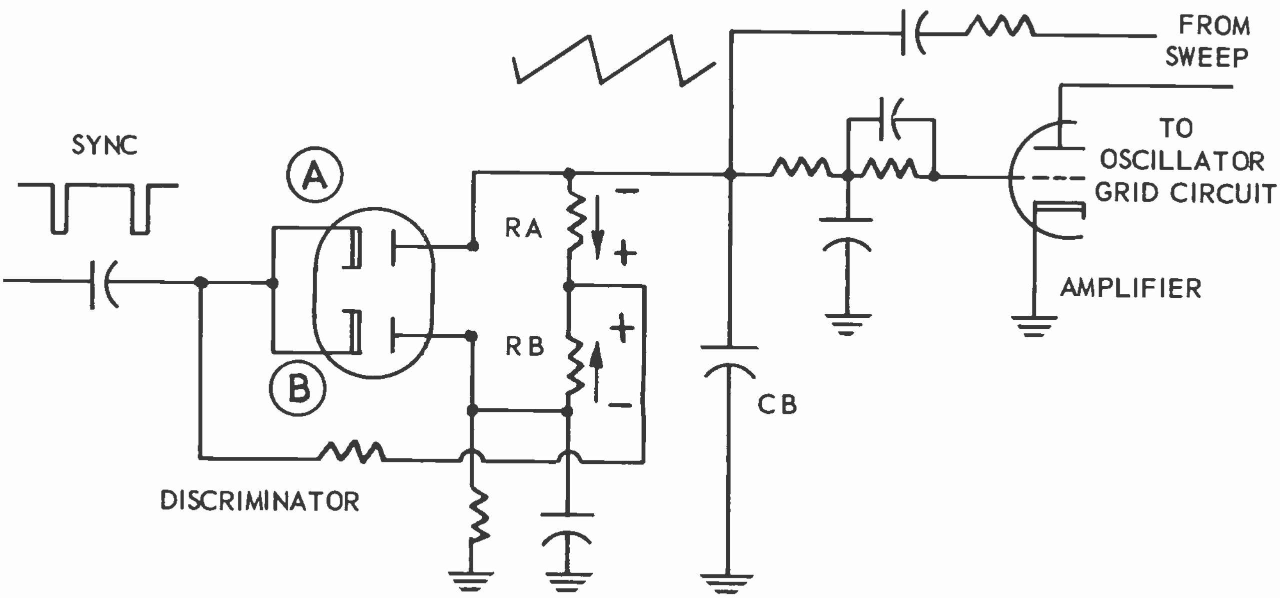


Fig. 55-16. A discriminator output connected through an amplifier to the grid circuit of a blocking oscillator.

Should the oscillator commence to run slow, or at a frequency lower than that of the sync pulses, we have the condition shown at the center of Fig. 55-15. The sawtooth waves occur later, which is equivalent to having the sync pulses come farther ahead on the waves. The positive sync pulse acts on diode A while the cathode is positive and the diode is non-conductive. But the positive pulse acts on diode B while its cathode is negative and while this diode is conductive. There will be a large current through diode B and load resistor Rb. This makes the discriminator output a positive d-c voltage, as you can see from Fig. 55-14.

If the oscillator commences to run fast with reference to sync frequency we have the condition shown at the right in Fig. 55-15. Now the positive sync pulses act on diode A while it is conductive, because of its negative cathode. Diode B now is non-conductive, so the sync pulse has no effect. There is a large current through diode A, and the discriminator output becomes a negative d-c voltage due to current in resistor Ra.

These variations of discriminator output polarity are applied to the grid of the amplifier. With a slow oscillator the discriminator output is positive and the amplifier grid is made positive with reference to its grounded cathode and the grounded lower end of the discriminator load resistors. Positive grid voltage has the effect of reducing plate resistance of the amplifier tube. This plate resistance is part of the grid circuit resistance for the oscillator. The reduction of grid circuit resistance allows the oscillator grid bias to become less negative, and there is a resulting increase of oscillator frequency.

With a fast oscillator all the actions are reversed. Discriminator output and amplifier grid bias go negative. Amplifier plate resistance increases. Oscillator grid circuit resistance is increased, the oscillator bias becomes more negative, and the oscillator is slowed down.

Another method of using a discriminator in an afc system for sweep oscillators is shown by Fig. 55-16.

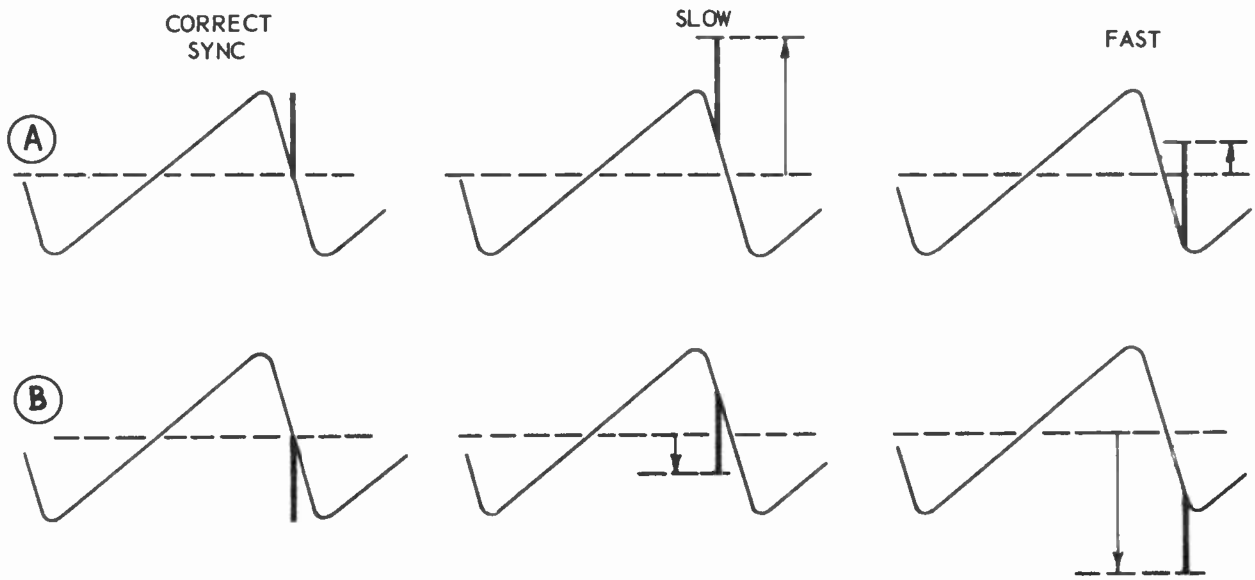


Fig. 55-17. How changes of oscillator frequency vary the correction voltage from a discriminator.

Negative voltage pulses from the sync section are applied to the cathodes of the two diodes, as in diagram 3 of Fig. 55-13. From the sweep output is brought back a sawtooth voltage which is applied to capacitor Cb and to the diode plates. This sawtooth voltage is in phase at the two diode plates, as shown by Fig. 55-17.

Since diode conduction currents flow opposite directions in the two load resistors, and thus have opposite effects on polarity of the discriminator output, these effects may be shown on the sawtooth plate voltages as in Fig. 55-17. When sweep voltage is correctly in time with sync pulses there are equal increases of conduction in both diodes, and the opposite effects cancel to leave no d-c output.

If the sweep oscillator runs slow the sawtooth waves occur later. Then there is more conduction in diode A than in B, and we have a negative d-c output. A fast oscillator, or an oscillator frequency higher than sync frequency, causes more conduction in diode B than in A. The result is a positive d-c output.

Correction voltage from the discriminator is applied to an amplifier, and from the amplifier is applied to the grid circuit of the first section in a blocking type sweep oscillator. The polarity is reversed in the amplifier. A negative discriminator output with a slow oscillator results in a positive voltage at the grid circuit of the oscillator, and its frequency is increased. A positive discriminator output with a fast oscillator has the opposite effect, and oscillator frequency is decreased.

There are other applications of the discriminator which involve the action of a reactance tube in the afc system. We shall examine these other applications after studying the reactance tube and some of its other uses in the following lesson.

58 → 60
56

COMMERCIAL TRADES INSTITUTE

KNOWLEDGE + AMBITION = SUCCESS

LEARN . . .

ACT . . .

ENJOY . . .

*"The heights by great men reached and kept
Were not attained in sudden flight,
But they, while their companions slept,
Were climbing upward in the night."*

COMMERCIAL TRADES INSTITUTE

CHICAGO, ILLINOIS

TELEVISION

LESSON NO. 56

PHASE DETECTORS AND REACTANCE TUBES

Training Your Mind

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON NO. 56

PHASE DETECTORS AND REACTANCE TUBES

In many television receivers the horizontal sweep frequency is automatically controlled by a twin-diode tube used as a phase detector. A phase detector produces a d-c voltage whose polarity and strength are proportional to differences in phase or in times of sync pulses and the actual output frequency of the sweep oscillator. One method of using a phase detector is shown by Fig. 56-1, which is a circuit diagram for an afc system in which the sweep oscillator is a cathode coupled multivibrator.

The cathode of one section and the plate of the other section of the phase detector are connected together. To them is applied a sawtooth voltage at the actual oscillator frequency, as obtained from some point in the horizontal deflection circuit. The voltage in the deflection circuit itself consists of very strong positive peaks, shown by the oscilloscope as at the left in Fig. 56-2. These voltage pulses charge the capacitors which are along the line that goes to the phase detector. The capacitors then discharge rather slowly through their associated resistors, in accordance with the capacitance-resistance time constant. The result is an approximate sawtooth wave at the joined cathode and plate of the detector. This wave is pictured at the right in Fig. 56-2. The frequency of this sawtooth voltage is, of course, that of the oscillator output. It may or may not be in time with the sync pulses.

Voltage pulses at the horizontal frequency of the received signal come to the phase detector from a sync inverter tube, sometimes called a splitter when used in this general type of circuit. Sync voltage pulses at the grid of the inverter tube are negative, as shown at A in Fig. 56-3. Since there is no inversion of polarity between grid and cathode, we have at the cathode of the inverter tube a series of negative voltage pulses at the sync frequency. These pulses are shown at B. The negative sync pulses are applied through capacitor Cb to the cathode of diode B in the phase detector, also to the lower end of load resistor Rb.

From the plate of the inverter tube we obtain positive sync pulses as shown at C in Fig. 56-3. These pulses are positive at the plate because there is inversion of polarity between the grid and the plate of the inverter. There is considerable amplification between the grid and the plate of the inverter tube, which makes the voltage of the plate pulses much greater than the voltage of pulses coming to the grid. Then the plate pulses will be much stronger than those from the cathode, because there is little or no difference between output voltage at the cathode and input voltage at the grid. To make the plate pulses and cathode pulses of approximately equal strength we take the plate pulses from a point in the plate circuit that follows dropping resistor Rd.

The positive sync pulses from the inverter plate circuit are taken through capacitor Ca to the plate of diode A in the phase detector, and also to the top of load resistor Ra.

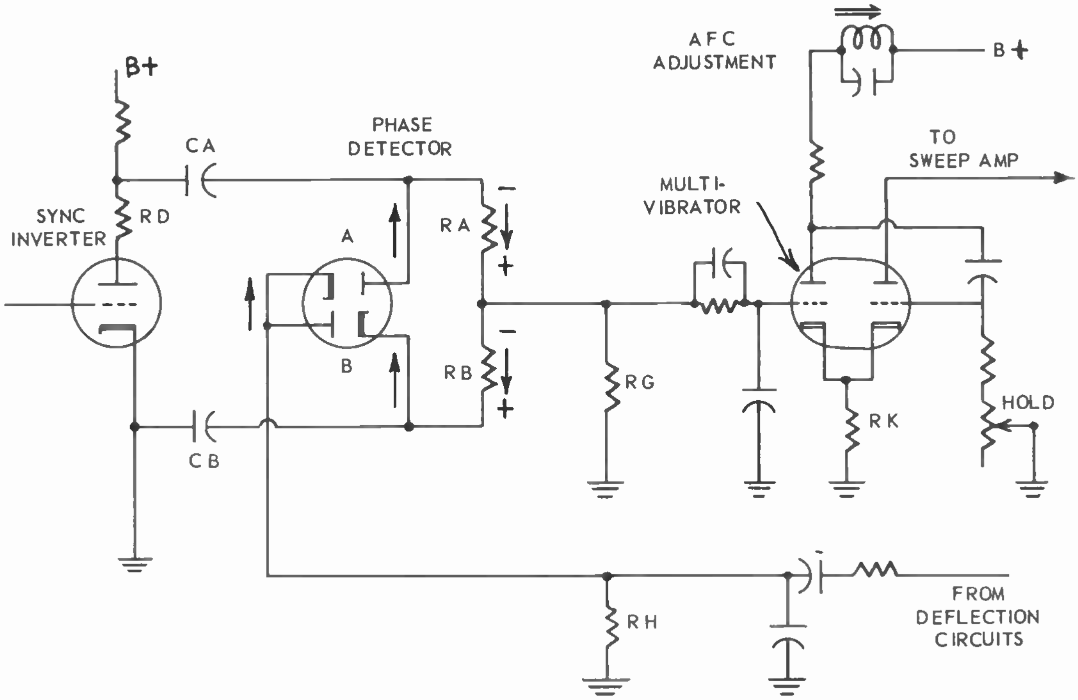


Fig. 56-1. Afc system with phase detector and multivibrator.

At the outer ends of the load resistors on the phase detector we now have a combination of two voltages. One is the sawtooth voltage whose frequency is that at which the oscillator actually is operating. The other voltage is that due to the sync pulses, whose frequency is that for correct timing of deflection in accordance with the received signal.

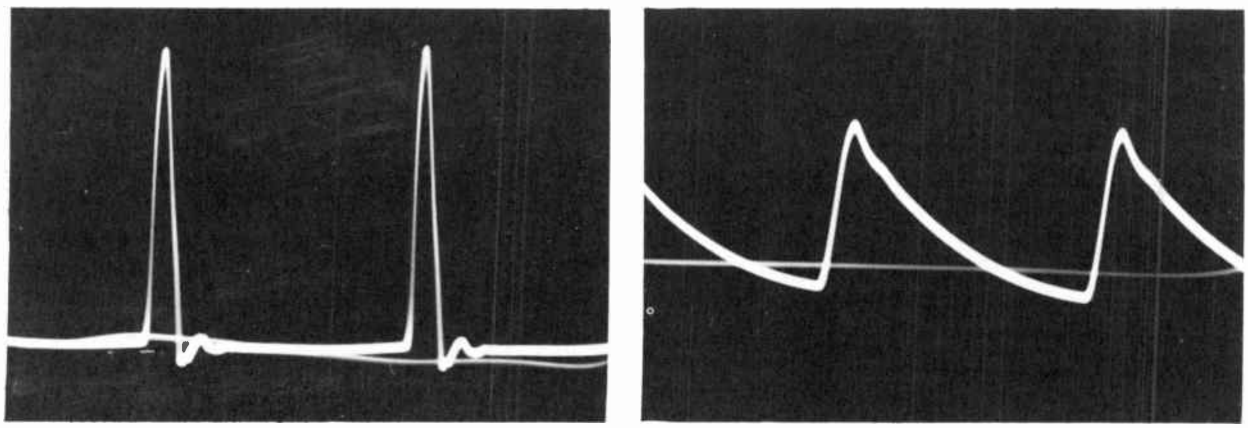


Fig. 56-2. Positive voltage pulses from deflection circuit (left) and sawtooth voltage fed to phase detector input (right).

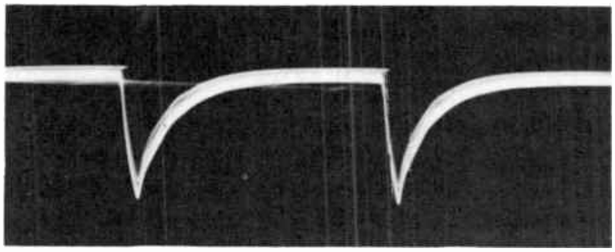
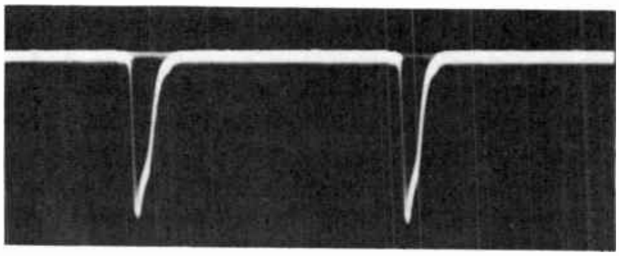
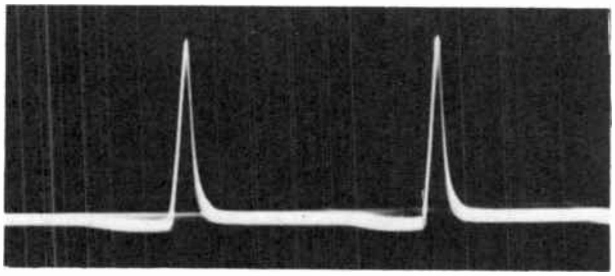


Fig. 56-3. A. negative pulses at inverter grid.



B, negative pulses at inverter cathode.



C, positive pulses from inverter plate.

The combination voltages are as shown by Fig. 56-4. Along the top row are voltages acting on diode A, between its cathode and its plate. The bottom row shows voltages acting in diode B, as measured from plate to cathode. Since the sawtooth wave is applied to the cathode of A and to the plate of B, this wave appears of the same polarity in all diagrams.

Sync pulse voltages ride on the steep slopes of the sawtooths. These pulse voltages are positive at the plate of diode A and are negative at the cathode of diode B. Thus the pulse voltages are of such polarities as will cause conduction in the diodes.

The two diagrams at the left in Fig. 56-4 show the sync pulses riding about half way down the sawtooth slopes. This is the condition when the sawtooth wave is correctly synchronized with the signal pulses. Peak-to-peak voltages are equal, and the result is equal conduction on both diodes. This means equal currents in the two load resistors Ra and Rb of Fig. 56-1, and equal voltages across these resistors.

Next toward the right are shown conditions when the oscillator runs too slowly, or at a frequency lower than that of the sync pulses. The sawtooth now occurs later in time, which would move it toward the right in these diagrams. Since the timing of the sync pulses does not change, or, at least, remains at the points which we wish to attain in deflection, the pulses effectively move to the left. The pulse on diode A rides higher. This increases the peak-to-peak voltage acting on diode A, increases conduction in this diode, and increases the current and voltage in load resistor Ra. At the same time the pulse on diode B has moved into such a time position as to decrease the peak-to-peak voltage on this diode. There is less conduction in diode B, and less current and voltage in load resistor Rb.

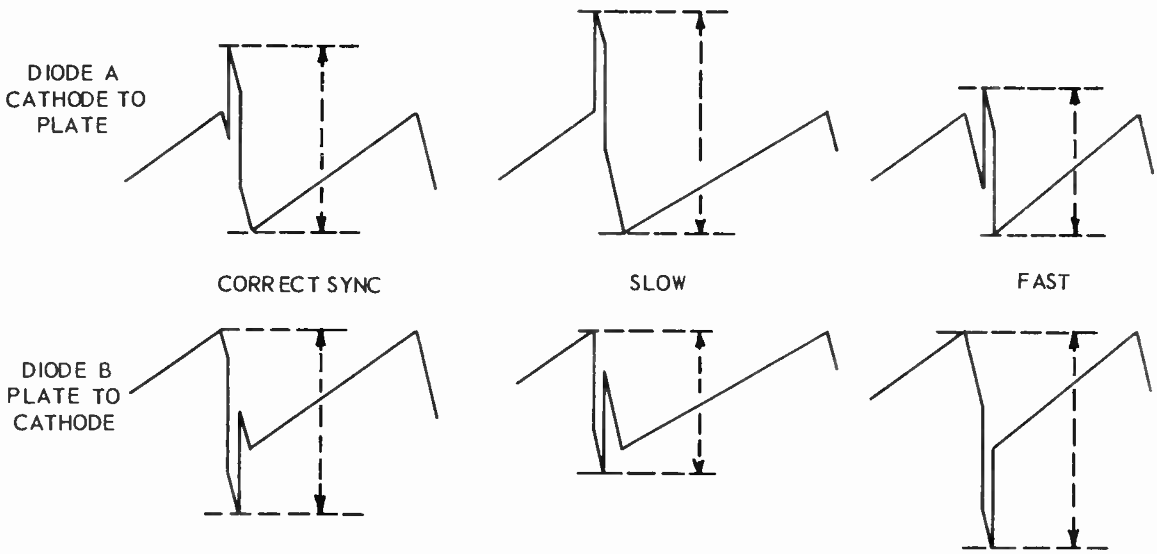


Fig. 56-4. How sync pulses combine with the sawtooth wave at various oscillator frequencies.

If the oscillator runs too fast, at a frequency higher than sync frequency, we have conditions shown by the two diagrams at the right in Fig. 56-4. The sawtooth wave occurs earlier in time, or moves toward the left in the diagrams. This brings the sync pulse on diode A down into the trough of the sawtooth, to reduce peak-to-peak voltage on this diode. There is less conduction in A, and less current and voltage in load resistor Ra. Simultaneously the sync pulse moves higher on the sawtooth for diode B, increasing the peak-to-peak voltage, the conduction, and increasing the current and voltage in load resistor Rb.

When currents in the two diodes and their respective load resistors are equal, the electron flow is as shown by arrows in Fig. 56-1. Flow is from the plate of diode A through resistor Ra, resistor Rb, and to the cathode of diode B, thence from the plate of this diode to the cathode of A and to the plate of A.

If there is more conduction through diode A than through B the excess electron flow must go through the path shown by heavy lines at the left in Fig. 56-5, since this is the only path between plate and cathode of A which is completed through conductors. Part of this path consists of resistor Rg which, as you can see from Fig. 56-1, is in the grid circuit of the oscillator and which provides part of the grid biasing voltage for the oscillator. Electron flow is downward in this grid resistor, making the oscillator grid more negative with reference to ground. This is the result of a slow oscillator, or of an oscillator frequency lower than sync pulse frequency.

If the oscillator runs fast there is more conduction in diode B than in A. Then the excess electron flow through B must follow the conductive path shown by heavy lines at the right in Fig. 56-5. This flow is upward in grid resistor Rg, making the oscillator grid more positive with reference to ground.

The oscillator grid always has an average negative bias due to the voltage drop across cathode resistor

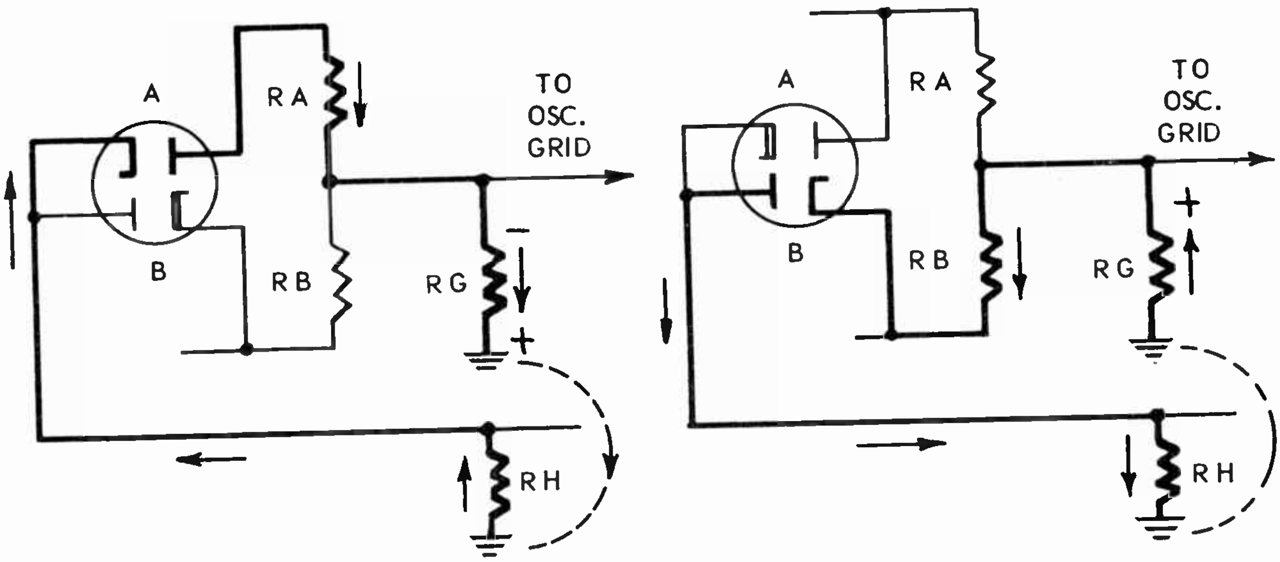


Fig. 56-5. Electron flow when the oscillator is slow (left) and when the oscillator is fast (right).

Rk of Fig. 56-1. The phase detector voltages in resistor Rg merely make the oscillator grid more negative for a slow oscillator and less negative for a fast oscillator. These changes of grid voltage on the first section of the cathode coupled multivibrator are equivalent to changes of opposite polarity at the grid of the second section. Then we have the effect of a lower oscillator frequency making the grid of the second section less negative, to increase the frequency. Conversely, higher oscillator frequency makes the grid of the second section more negative, to lower the frequency. This is automatic frequency control.

The changes of phase detector output voltage which occur with the continual slight shifting of oscillator and sweep frequency may be checked with a sensitive electronic voltmeter. The voltmeter may be connected between ground and either the top of resistor Ra or the bottom of resistor Rb. Otherwise the meter may be connected between ground and the top of resistor Rg. By closely watching the meter pointer you will see almost continual small changes of d-c voltage as the sweep oscillator attempts to vary its frequency and is brought back into time by the correction voltage.

The changes of correction voltages may be seen by connecting an oscilloscope from ground to the top of resistor Ra or to the bottom of resistor Rb. Traces observed with these connections are shown by Fig. 56-6. When the oscillator is slow the correction voltage from diode A is pictured by trace number 1, and from diode B by trace number 2. It is plain that diode A is furnishing the greater voltage, just as at the left in Fig. 56-5. Trace number 3 shows the voltage from diode A when the oscillator is fast, and trace number 4 shows the voltage from B with a fast oscillator. Here it is apparent that diode B is furnishing the greater voltage, as at the right in Fig. 56-5.

The oscilloscope traces show the sawtooth voltages with opposite slopes in the two diode circuits, which indicate opposite polarity. This is because both voltages here are being observed with reference

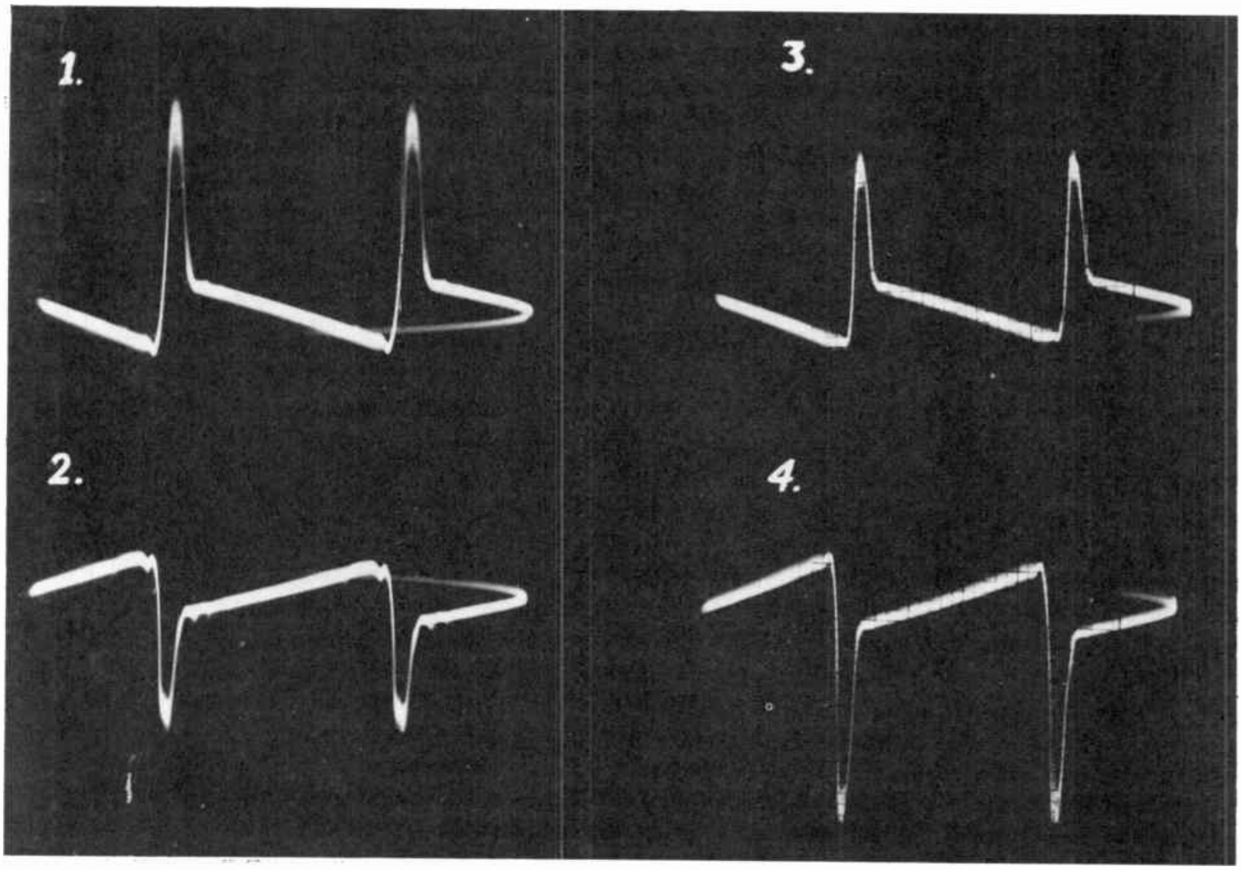


Fig. 56-6. Voltages observed at the diodes with the oscillator running too slowly or too rapidly.

to ground, whereas in Fig. 56-4 the two voltages are with reference to the cathode of one diode and the plate of the other diode.

The only service adjustment in this control system is marked AFC Adjustment in the diagram of Fig. 56-1. This is a parallel resonant circuit which is tuned to the horizontal sync frequency by means of a movable core in the inductor or coil. With correct adjustment there is maximum impedance in the oscillator plate circuit at the horizontal frequency, and the oscillator gives its maximum voltage output at this frequency.

Adjustment is made as follows:

1. Set the manually operated hold control at the center of its range.

2. Tune the channel selector and set the contrast control as for normal reception in any channel on which a program is being transmitted.

3. Move the core of the afc adjustment until the picture or pattern locks into horizontal synchronization.

4. Turn the hold control adjustment fully clockwise, then slowly back counter-clockwise until the picture just locks in. Three or four sloping black bars should appear just before lock-in.

5. Repeat the check by turning the hold control fully counter-clockwise, then back just far enough for lock-in. The bars should slope the opposite direction just before lock-in.

6. Carefully adjust the core of the afc coil to obtain lock-in with the hold control turned about the same part of its travel from either extreme position. The picture or pattern then should remain locked when the hold control is rotated through one-fourth to one-third of its range.

This general method of frequency control may be used when sync pulses to the grid of the inverter tube are positive rather than negative. Then the plate of the inverter is connected through a blocking capacitor to the cathode of the phase detector, and the cathode of the inverter is connected through another blocking capacitor to the plate of the detector. Correction voltage polarities then will be the same as described for the system with negative pulses to the inverter grid. Fig. 56-7 shows this method of connection, also the pulse polarities and the direction of electron flow. Negative pulses are applied to the cathode of the phase detector and positive pulses are applied to its plate, just as with the system previously described.

There are numerous other afc systems in which a phase detector is used for combining sync pulse voltages with sawtooth voltages to obtain a d-c output that varies with frequency differences. In Fig. 56-8 the

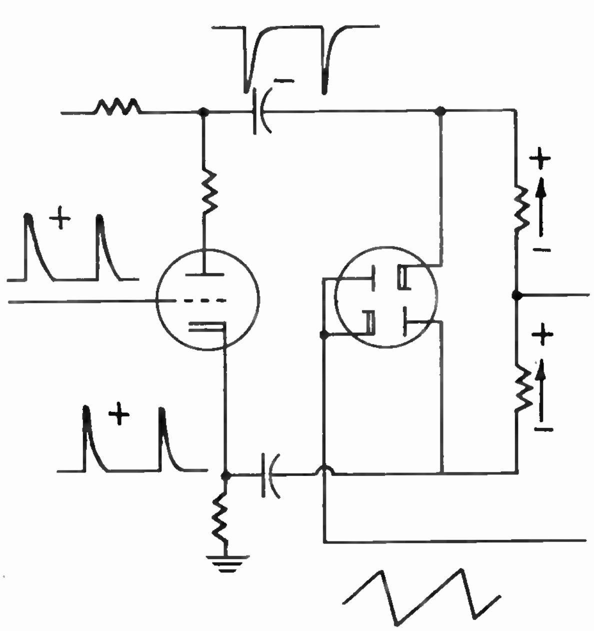


Fig. 56-7. Connections for positive sync pulses at inverter grid.

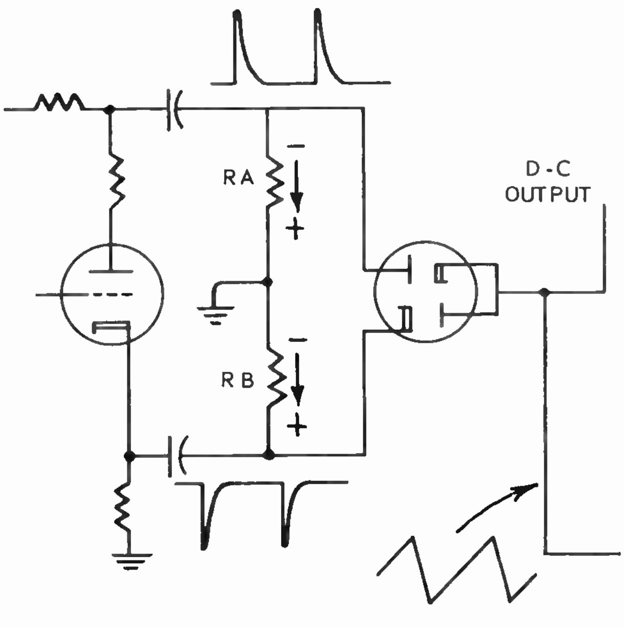


Fig. 56-8. D-c output taken from a cathode and plate connected together.

sync pulse voltages of opposite polarity are applied to the outer ends of resistors Ra and Rb and to one plate and one cathode of the detector. The remaining cathode and plate are connected together, and from them is taken the d-c correction voltage for the sweep oscillator grid circuit.

The sawtooth voltage derived from some point in the deflection circuit is applied to the joined cathode and plate of the phase detector. This wave then will act in the same polarity through both diodes to give a correction effect similar to that illustrated by Fig. 56-4. Correction voltage is proportional to the difference between voltage across resistor Ra and the voltage across resistor Rb.

PHASE DETECTORS WITH AMPLIFIERS. In Fig. 56-9 we have a sync inverter and phase detector similar to these units in Fig. 56-1, and there will be similar changes of output voltage from the detector when the actual sweep frequency varies from the correct sync frequency. But now we have the output of the phase detector connected to the grid of a triode used as an afc amplifier.

At the upper right is a blocking oscillator with the usual hold control resistance connected to its grid. But instead of the far end of the hold control resistance going to ground it is connected to the plate of the afc amplifier tube. Then voltage at the oscillator grid will be altered by every change of voltage at the plate of the amplifier, or at point X. If plate voltage becomes less positive, or effectively more negative,

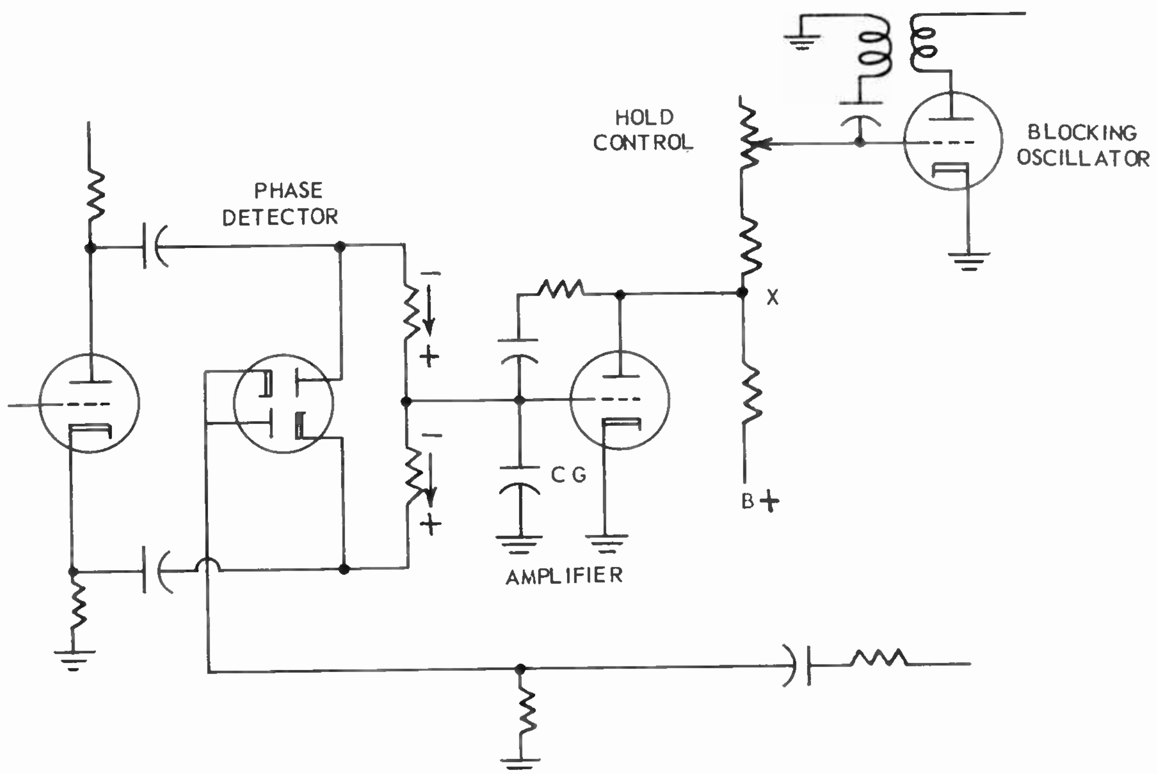


Fig. 56-9. Afc stetem with phase detector, amplifier, and blocking oscillator.

the oscillator grid voltage will be made more negative and oscillation frequency will be lowered. Amplifier plate voltage more positive will make oscillator grid voltage less negative, and oscillation frequency increased.

Let's assume that oscillator frequency tends to increase above sync frequency. Correction voltage from the phase detector will go more positive, just as it did when applied to the grid of the first section of a multivibrator oscillator. This will place a positive charge on the grid end of capacitor Cg and on the amplifier grid. Now the amplifier draws increased current. There is more voltage drop in the resistor from the plate to B-plus, and voltage at the plate itself becomes less positive. The result is to make the oscillator grid voltage more negative, and the oscillator slows down.

Should oscillator frequency tend to drop below sync frequency the correction voltage from the phase detector will go negative. All the other changes will be reversed, and oscillator frequency will be increased. This is automatic frequency control. The series resistor and capacitor between plate and grid of the amplifier are needed to prevent sudden changes of voltage in the oscillator grid circuit from acting on the phase detector through the amplifier. When a system of this general type is connected to the grid of the first section in a multivibrator the series resistor and capacitor are omitted.

The afc amplifier may be used also on a phase detector output of the kind shown by Fig. 57-8, with the amplifier grid connected to the joined plate and cathode of the detector. The amplifier plate circuit and oscillator grid circuit then will be as described in connection with Fig. 56-9, or the oscillator may be a multivibrator instead of the blocking type shown in that figure.

REACTANCE TUBE. In all the afc systems which previously have been examined the frequency of the sweep oscillator is controlled by varying its grid voltage in one way or another. Now we shall look at some systems in which the oscillator frequency is determined by a tuned resonant circuit consisting of inductance and capacitance. The oscillation frequency will be varied to hold it in synchronization by varying the effective tuning inductance. Strange as it may seem, the effective tuning inductance for the sweep oscillator will be varied without using an adjustable inductor. Instead we shall use a reactance tube, which may be made to behave like an inductance whose value is varied simply by changing the voltage applied to the grid of this tube.

The elementary principle of reactance tube action may be explained with the help of Fig. 56-10. The oscillator tuning inductor or coil is marked La. The top of this coil is connected to the oscillator grid through capacitor Cg. The oscillator grid is biased in the usual way by this capacitor and resistor Rg. The lower end of the tuning coil is connected through ground and capacitor Cb to the oscillator plate. The oscillator cathode is connected to a tap on the tuning coil. You will recognize that this much of the circuit is nothing more than a common type of Hartley oscillator.

Across coil La are capacitor Ca and resistor Ra, in series with each other. La and Ca are the tuning inductance and capacitance. Resistance at Ra is only 10 ohms or thereabouts, so has little effect on oscillation. Inductance and capacitance in this tuned oscillator circuit are of such values as to be resonantat the horizontal sync frequency of 15,750 cycles per second.

In parallel with capacitor Ca and effectively in parallel with the inductance of La (because of small resistance at Ra) we have the reactance tube. The plate of this tube is connected to the top of the tuned

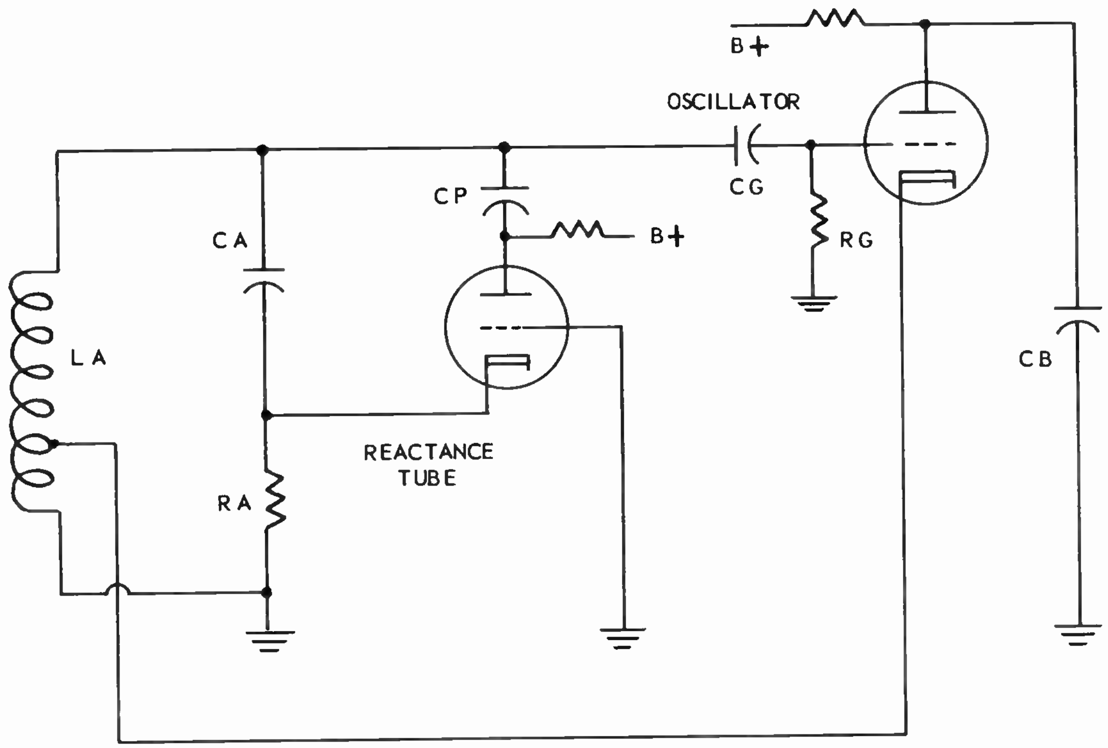


Fig. 56-10. Reactance tube connected to act as a variable inductance.

circuit through capacitor Cp. This is merely a blocking capacitor to keep d-c voltage for the plate of the reactance tube from affecting the oscillator grid circuit, while furnishing a path for alternating current and voltage. The cathode of the reactance tube is connected to the bottom of tuning capacitor Ca, and effectively to the bottom of the tuned circuit through the small resistance at Ra.

We wish the reactance tube to have the same effect as would a variable inductance in parallel with the tuned circuit. Let's first inquire what this effect should be. We know that in any pure inductance the alternating current will lag the alternating voltage by 90 degrees. Therefore, if alternating plate current in the reactance tube can be made to lag the alternating plate voltage by 90 degrees this tube will act like an inductance. Now to see how this is accomplished.

Since capacitor Ca is part of the oscillatory circuit there is oscillating or alternating voltage across this capacitor. Ca is connected across the plate and cathode of the reactance tube. Therefore, the alternating voltage which is across Ca is applied as alternating plate voltage to the reactance tube. This alternating plate voltage and the capacitor voltage are represented by curve 1 of Fig. 56-11.

Because capacitor Ca and resistor Ra are in series with each other in the oscillatory circuit the current is the same in these two elements. That is to say, current in Ra is in phase with current in Ca. But, as

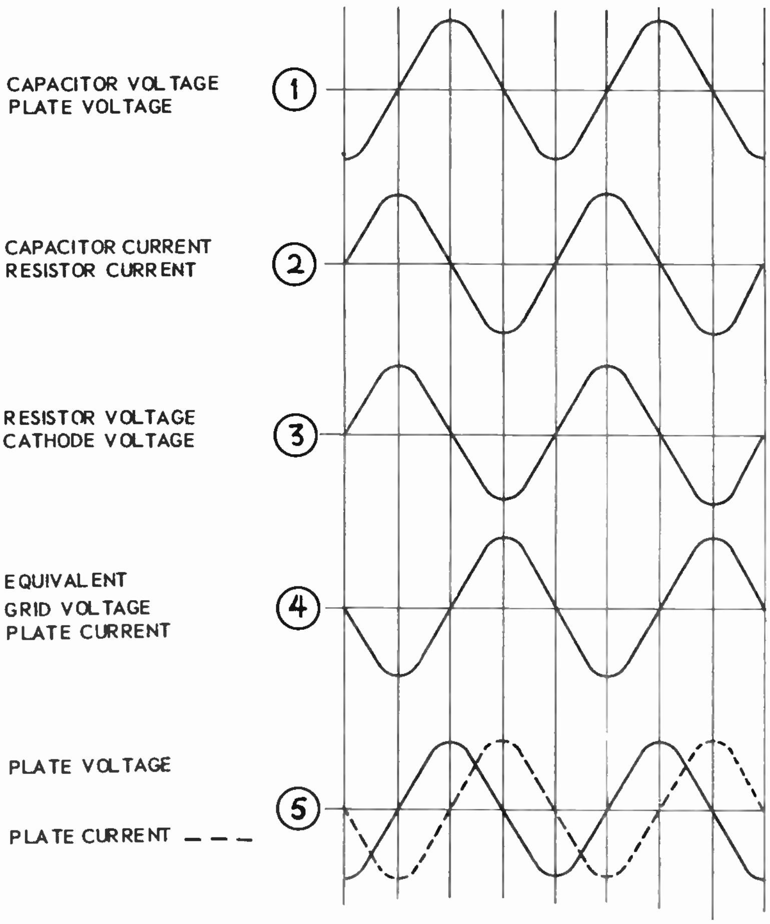


Fig. 56-11. Phase relations of voltages and currents in the reactance tube circuit.

in any capacitance, the alternating current in Ca leads the alternating voltage by 90 degrees. Consequently, the current in Ca and the same current in Ra is leading the capacitor voltage and reactance tube plate voltage by 90 degrees. This leading current is represented by diagram 2.

The alternating voltage across any resistance is in phase with the alternating current in that resistance. Then the voltage across resistor Ra must be in phase with its alternating current. This voltage is represented by curve 3. Voltage from resistor Ra acts on the cathode of the reactance tube, so curve 3 also represents this alternating cathode voltage.

Any voltage at the cathode of a tube has the same effect on plate current as a voltage of opposite phase at the grid of the same tube. This effect is observed every time we consider a cathode biasing system. Making the cathode more positive is equivalent to making the grid more negative. It follows that a reactance tube grid voltage equivalent to the alternating voltage on the cathode would be as shown by curve 4.

COMMERCIAL TRADES INSTITUTE

In any tube having a grid, plate, and cathode the alternating plate current is in phase with alternating grid voltage. A more positive grid voltage increases the plate current, and a more negative grid voltage decreases the plate current. Then reactance tube plate current is also represented by curve 4.

At 5 in Fig. 56-11 the reactance tube plate voltage (from curve 1) is shown by the full-line curve. The plate current (from curve 4) is shown by the broken-line curve. It is apparent that plate current is lagging the plate voltage by 90 degrees, and that the reactance tube is acting like an inductance connected in parallel with the tuned circuit or in parallel with the inductance of tuning coil La.

An increase of plate current in the reactance tube will increase its inductive effect. We may increase the plate current by making the grid of this tube more positive or less negative. Such a change of d-c grid voltage will be provided by opening the grid circuit, which goes directly to ground in Fig. 56-10, and inserting a source of variable d-c voltage. A decrease of reactance tube plate current will lessen the inductive affect. The plate current will be decreased by applying a less positive or more negative d-c voltage to the grid of this tube.

Now we must consider the effect on oscillator frequency of increasing or decreasing the inductive effect of the reactance tube, which is acting like one inductance in parallel with another inductance in coil La. Adding a paralleled inductance reduces the effective combined inductance of the two inductors, just as adding one resistance in parallel with another reduces the combined resistance. Less total inductance in the oscillator tuned circuit increases the oscillation frequency. All this adds up as follows:

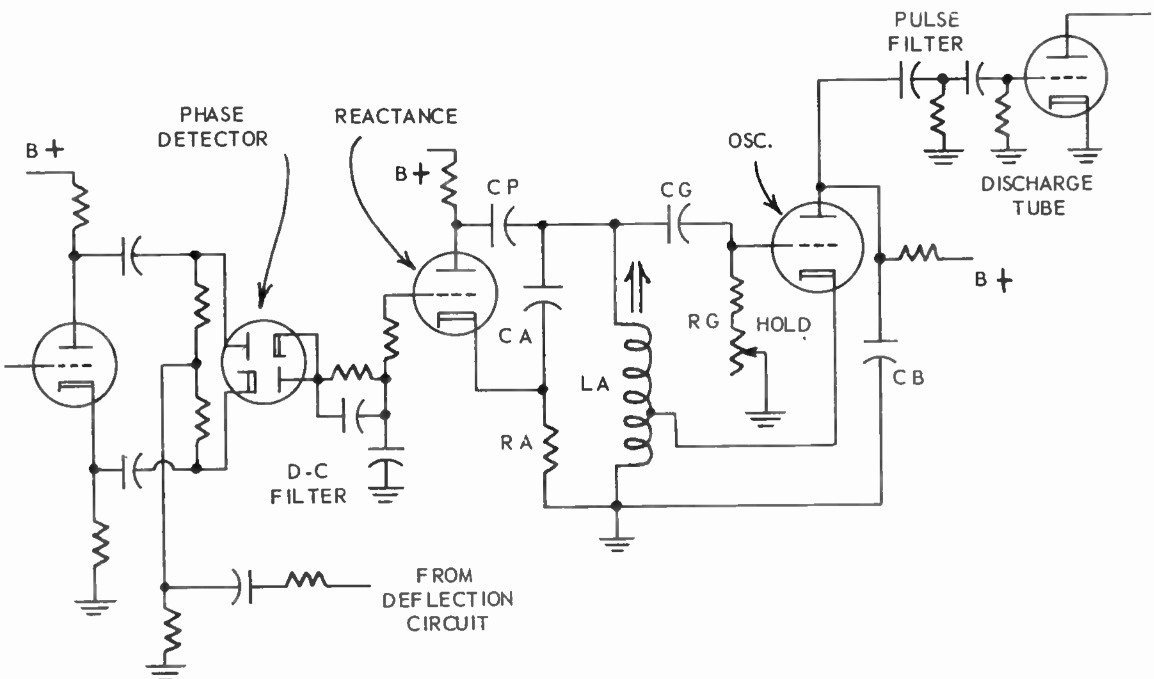


Fig. 56-12. Afc system with phase detector, reactance tube, and Hartley oscillator.

Reactance tube grid more positive. More inductive effect in tube. Less total or combined inductance in oscillator tuned circuit. Higher oscillation frequency. If the grid of the reactance tube already is negatively biased, as normally is the case, making the grid less negative will have exactly the same effect - it will raise the oscillation frequency.

If the grid of the reactance tube is made more negative, everything will be reversed and oscillation frequency will be lowered. All this is quite easy to remember, because it so happens that a given change of voltage at the grid of the reactance tube has the same effect on oscillation frequency as the same change of voltage at the grid of a blocking oscillator or at the second grid of a cathode coupled multivibrator.

The d-c voltage at the grid of the reactance tube may be varied by connecting this grid to the d-c output of either a phase detector or a discriminator. Both methods are used for automatic control of oscillator frequency. We shall look first at a system employing a phase detector.

PHASE DETECTOR AND REACTANCE TUBE. Fig. 56-12 shows a practical application of the reactance tube in a circuit wherein control voltage for the grid of this tube is furnished by a phase detector. Between the output of the phase detector and the grid of the reactance tube is a d-c filter system composed of resistors and capacitors. This filter removes from the detector output any remaining sync pulse voltages and other pulse or alternating frequencies, to leave a practically smooth d-c control voltage for the grid of the reactance tube. Another name sometimes used is noise filter.

The various parts of the phase detector circuit and the oscillator grid circuit are identified by the same reference letters used in Fig. 56-10. Two adjustments have been added. One is a movable core in the oscillator tuning coil La, called a horizontal lock control or a horizontal frequency control. The other is a hold control, which is an adjustable resistor from the oscillator grid to ground at Rg.

The hold control is a front panel adjustment for use by the set operator. The lock or frequency control is a service adjustment. With the hold control at the approximate center of its range the lock control is adjusted until the pattern or picture falls into synchronization, if not already there. It should be possible to move the hold control some distance each way from center while holding synchronization, and to bring the pattern or picture back into sync after the hold control has been turned to the limit of its range in either direction. If this is not possible, the lock control is carefully readjusted until the hold control operates as described.

In the circuit illustrated the sawtooth capacitor does not discharge through the oscillator, but through a separate discharge tube that follows the oscillator. This discharge tube is needed because oscillating voltage at the grid of this particular oscillator does not have sharp positive peaks suitable for allowing discharge through it of a sawtooth capacitor connected to its plate. You will recall that such positive peaks are produced at the grids of blocking oscillators and of multivibrators which act also as discharge tubes.

The hold control on the oscillator grid of Fig. 56-12, in connection with the pulse filter leading from the oscillator plate to discharge tube grid, act to form sharp positive pulses at this latter grid. Then the discharge tube becomes conductive and non-conductive to allow discharge and recharge of a sawtooth capacitor in the plate circuit of this tube. We shall examine the action a little later in this lesson.

There are many reactance tube circuits other than the one that we have been studying. Any connection which allows a tube to act like a variable inductance or a variable capacitance in parallel or in series with a tuned circuit provides what may be called a reactance tube. As an example, the positions of capacitor Ca and resistor Ra might be interchanged in Fig. 56-10. Then plate voltage on the reactance tube would be in phase with resistor voltage, and cathode voltage would be in phase with capacitor voltage. Plate current in the reactance tube then would lead the plate voltage, and we would have the effect of a variable capacitance across the tuned circuit for the oscillator.

Fig. 56-13 shows another way of using a reactance tube to vary the effective capacitance in the tuned circuit for the oscillator grid. Correction voltage for the grid of the reactance tube is furnished by a discriminator which follows the last sync amplifier tube. Discriminator output, from the top of the two load resistors, is applied to the reactance tube grid through a d-c filter or noise filter.

Inductance coil La is in the tuned grid circuit of the oscillator. In parallel with this inductance are three capacitive branches of the tuned circuit. One is capacitor Ca. The second branch consists of capacitor Cp in series with the plate-cathode path through the reactance tube.

When the hold control resistance is increased it increases the resistance as compared with capacitive reactance of capacitor Ch in this branch of the circuit. This is equivalent to having less capacitance, and oscillation frequency is raised. Reducing the hold control resistance lowers the oscillation frequency.

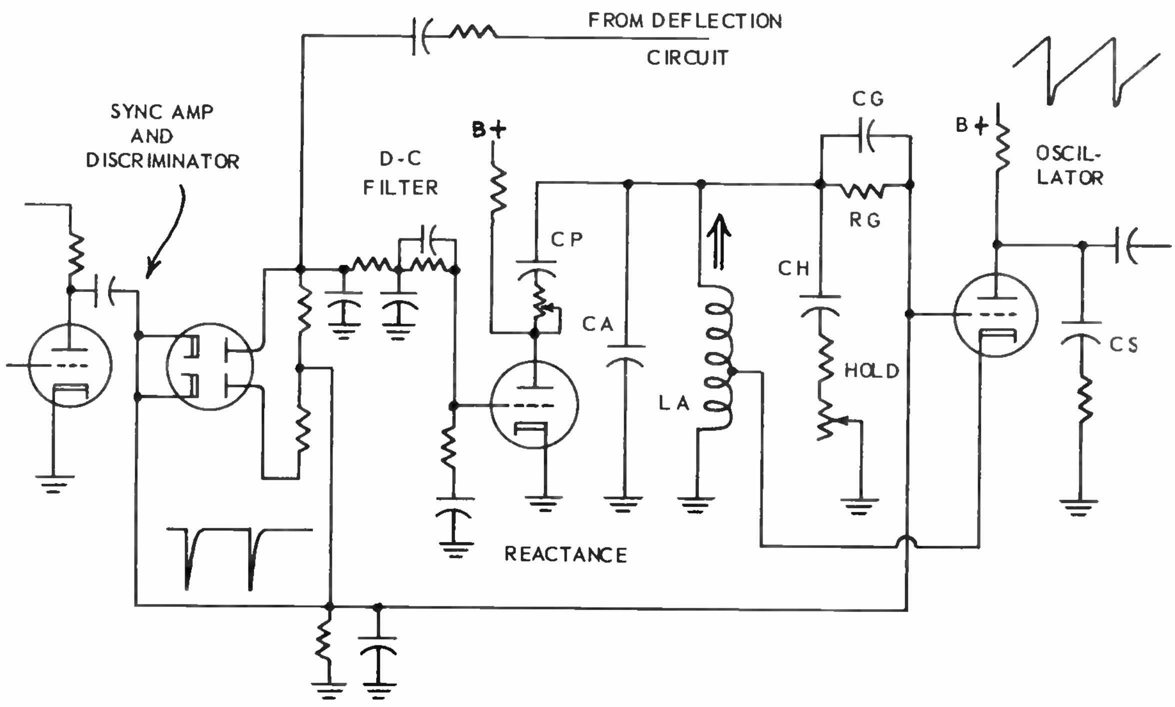


Fig. 56-13. Afc system in which a reactance tube operates as a variable capacitance.

The reactance tube plate resistance is a variable resistance in series with capacitor Cp. Increasing the plate resistance of the reactance tube has the same effect as increasing the hold control resistance, it allows capacitor Cp to have less effect in the tuned oscillator circuit or is equivalent to reducing the capacitance. This, of course, increases the oscillation frequency. Plate resistance of the reactance tube is varied by change of its grid voltage as determined by the d-c output of the discriminator. If the discriminator makes the reactance tube grid more negative, there is greater plate resistance in the reactance tube, the tuning effect of capacitor Cp is reduced, and oscillation frequency is raised. A less negative grid voltage on the reactance tube has the opposite effect, and oscillation frequency is lowered.

Inductance of La is altered by a movable core in this coil. This is a service adjustment. The hold control is a front panel adjustment accessible to the operator. The movable core in La is adjusted to allow correct action of the hold control by following the same method as explained in connection with these two adjustments in Fig. 56-12.

No discharge tube follows the oscillator in the system of Fig. 56-13. The oscillator itself is made to act as the discharge tube for sawtooth capacitor Cs in its plate circuit. This is accomplished by having grid capacitor Cg and grid resistor Rg of such values as provide grid-leak bias so strongly negative that the oscillator is held non-conductive except during the most positive tips of oscillation voltage on the oscillator grid. These positive tips of grid voltage allow discharge of sawtooth capacitor Cs through the oscillator tube. There is additional shaping of the grid voltage waveform and a sharpening of the change from negative to positive by means of sync pulse voltages brought to the oscillator grid through a connection from the sync amplifier and discriminator.

DISCRIMINATOR AND REACTANCE TUBE. Fig. 56-14 is a circuit diagram for an afc system in which a discriminator supplies d-c correction voltage to the grid of a reactance tube. The reactance tube varies the effective inductance in the tuned grid circuit of a Hartley oscillator. The output of the oscillator is fed to a discharge tube. Each of these circuit elements has been examined in other combinations. Here we have them in the afc system which was first to be generally used, and which still is found in many of the newest receivers.

The feature which chiefly distinguishes the appearance of this circuit from others containing a reactance tube is the two-winding transformer between oscillator and discriminator. The transformer primary, La, is in the oscillator grid circuit. It takes the place of oscillator tuning coil La of Fig. 56-10. This winding, and all parts of the reactance tube circuit, operate as explained in connection with that figure, and all are similarly lettered so that you may identify them.

The secondary of the transformer Ld is connected to the discriminator plates, which are on the input side of the discriminator. The secondary is tuned to resonance at the horizontal sync frequency by its inductance and the capacitor Cd. The frequency is adjustable by means of a movable core in the secondary winding. Primary resonant frequency is adjustable by a separate movable core in the primary.

Positive sync pulses from the last tube in the sync section are applied to a center tap on the transformer secondary. Because these pulses are put into the center tap they appear at both ends of the secondary and at both discriminator plates in the same positive polarity. In cases where the sync pulses consist of positive and negative spikes only the positive spikes are effective, since they go to the plates of the discriminator and only positive voltages at the plates will make the diodes conductive.

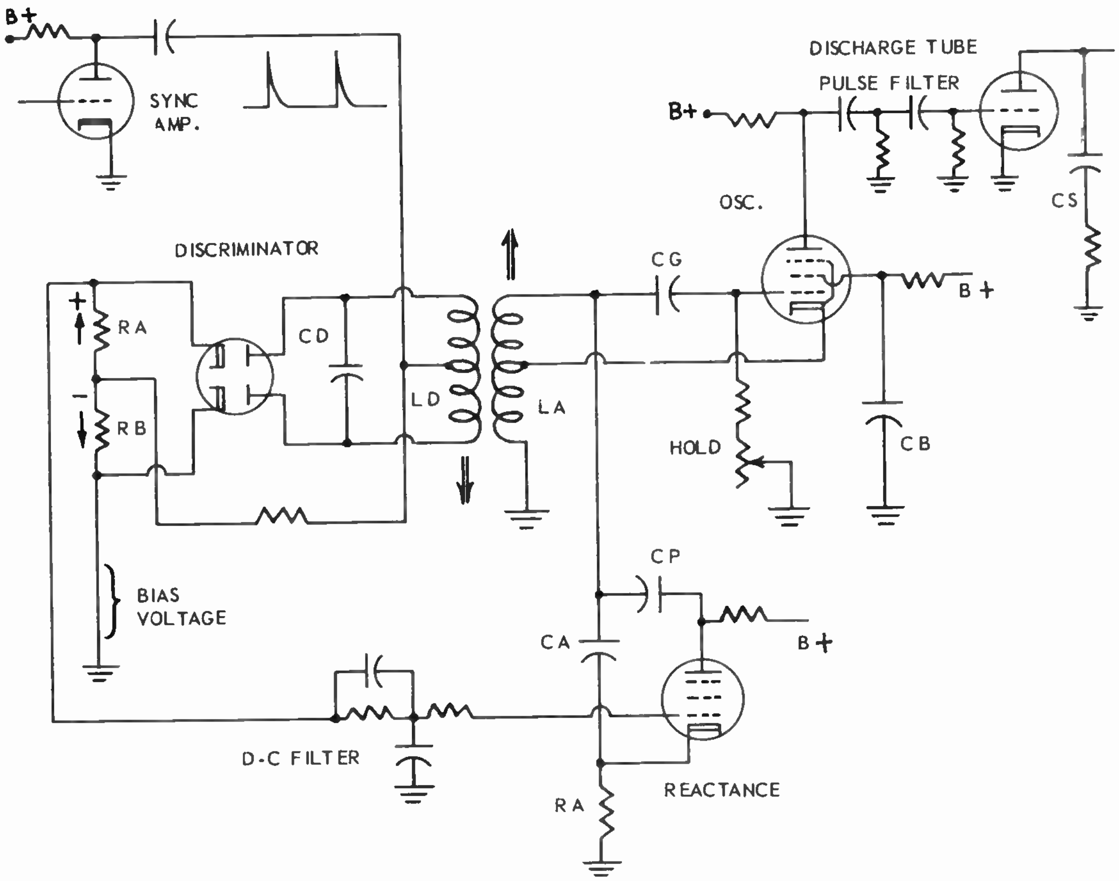


Fig. 56-14. Afc system with discriminator, reactance tube, and Hartley oscillator.

Although the sync pulses make the plates of both discriminator diodes positive at the same time, the resulting diode currents flow opposite directions in load resistors Ra and Rb. If conductions are equal in the two diodes there are equal opposite voltages across the load resistors and the d-c output of the discriminator is zero. Then if the only voltage applied to the discriminator were that of the sync pulses the d-c output always would be zero.

To the discriminator plates is applied also a voltage induced in the transformer secondary because of its coupling to the primary. This secondary voltage, like the primary voltage, is at the actual operating frequency of the oscillator and at the actual sweep frequency, which may or may not be correctly in time with the received sync pulses. The transformer secondary voltage is of opposite polarity at the outer ends of this winding, and is applied in opposite polarity to the plates of the two discriminator diodes. This voltage has the form of a sine wave, because there is sine wave current in the transformer primary.

The sync pulse voltages add themselves to the sine wave voltages at the plates of the discriminator diodes, with the results shown by Fig. 56-15. The diodes are conductive during positive alternations of sine wave voltages, and are non-conductive during negative alternations. On the upper line are represented voltages on the plates of diodes A and B when the sine wave oscillator voltage is correctly synchronized with received sync pulses. Sync voltages occur at zero points on the sine waves in both diodes. The pulses act just as though there were no sine wave voltage, and they cause balanced or zero d-c output voltage from the discriminator.

Conditions with a slow oscillator are shown by the middle line of Fig. 56-15. The sine wave voltage occurs later in time, it has moved to the right. Now the sync pulse voltage rides high on the sine wave for diode A, and there is increased conduction in this diode. The result is a positive d-c voltage from the discriminator output, as you can see by referring to Fig. 56-14. This makes the grid of the reactance tube more positive or less negative, and oscillator frequency is increased.

If the oscillator tends to run fast the voltages on the discriminator plates change as shown at the bottom of Fig. 56-15. Pulse voltage on diode A has dropped into the non-conductive region, but on diode B this voltage is riding high on the sine wave. This increases conduction in diode B over that in diode A, and there is negative d-c output voltage from the discriminator. This output makes the reactance tube grid more negative, and oscillator frequency is lowered to bring it back into synchronization.

The fixed negative bias for the reactance tube is provided at the bottom of load resistor Rb in Fig. 56-14 by making a connection to any point which is between 1½ and 3 volts more negative than ground. The grid return for the reactance tube is through the d-c filter, thence through load resistors Ra and Rb to the biasing point. The d-c output of the discriminator adds to or subtracts from the fixed bias voltage.

The d-c filter or noise filter acts similarly to such filters shown previously. The filter time constant is as long as several horizontal lines. Thus we have on the reactance tube side of the filter an average of momentary variations of discriminator output voltage, practically unaffected by any instantaneous noise voltages coming from the sync section to the discriminator plates. Were it not for the averaging effect or the time delay effect of this filter the picture could "jitter" when reception is noisy.

The action of the hold control in the afc system being examined is so different from anything with which we are acquainted as to be well worth some discussion. What happens between the grid of the oscillator and the grid of the discharge tube is illustrated by Fig. 56-16. This applies also to the hold control of Fig. 56-12.

Curve 1 represents the sine wave voltage coming from oscillator coil La to grid capacitor Cg of Fig. 56-16. The positive alternations make the grid so positive as to cause plate current saturation, due to the large current and large voltage drop in the resistor from the oscillator plate to B plus. Then the top of the plate current wave (curve 2) is flattened off as at a. As grid voltage becomes less positive the plate current decreases to point b. But the grid capacitor and grid resistor of the hold control are applying a grid-leak bias so negative as to cause plate current cutoff when grid voltage falls to point b. Plate current goes to zero at c. Current remains at cutoff or zero until the next positive alternation of grid voltage brings the grid back to cutoff at d. Then the performance repeats on every cycle of sine wave voltage at the grid.

Changes of plate current are accompanied by the changes of plate voltage shown by curve 3. As you know, plate voltage is in opposite phase to plate current. This is the voltage at the oscillator plate, it is the voltage applied to the pulse filter. The filter is a so-called differentiating type, and the applied plate voltage is an approximate square wave. Then every sudden rise of plate voltage causes a positive pulse at the filter output, and every sudden drop of plate voltage causes a negative pulse. These pulses are applied to the grid of the discharge tube. The positive pulses trigger this tube and allow discharge through it of the sawtooth capacitor. Curve 4 shows positive and negative pulses at the filter output.

Note that the leading edges of positive triggering pulses occur when sine wave voltage at the oscillator grid drops to the cutoff value. We are not interested in the negative pulses, for they have no effect on the discharge tube action.

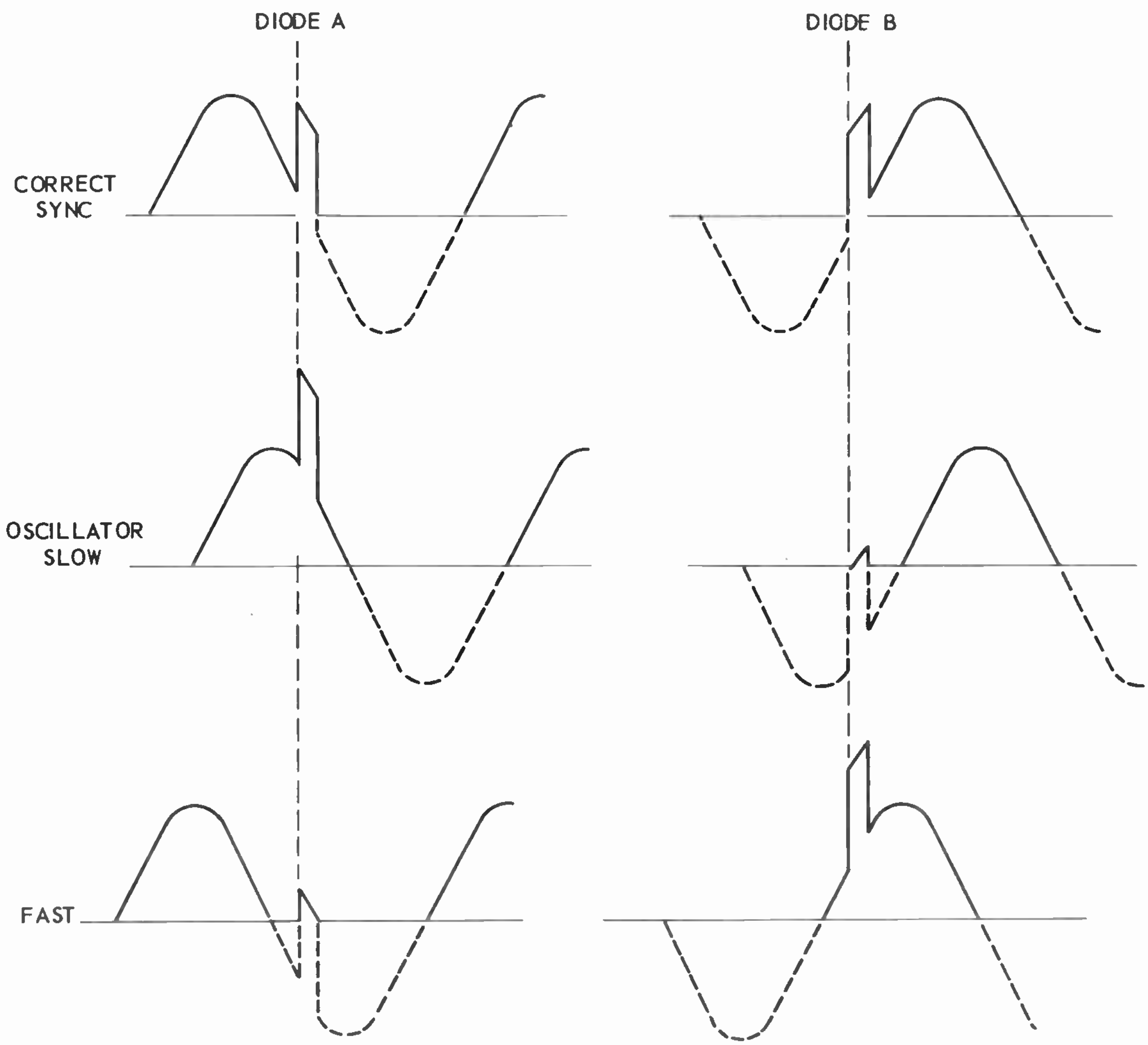


Fig. 56-15. How sync pulses combine with the sine wave voltage from the oscillator circuit.

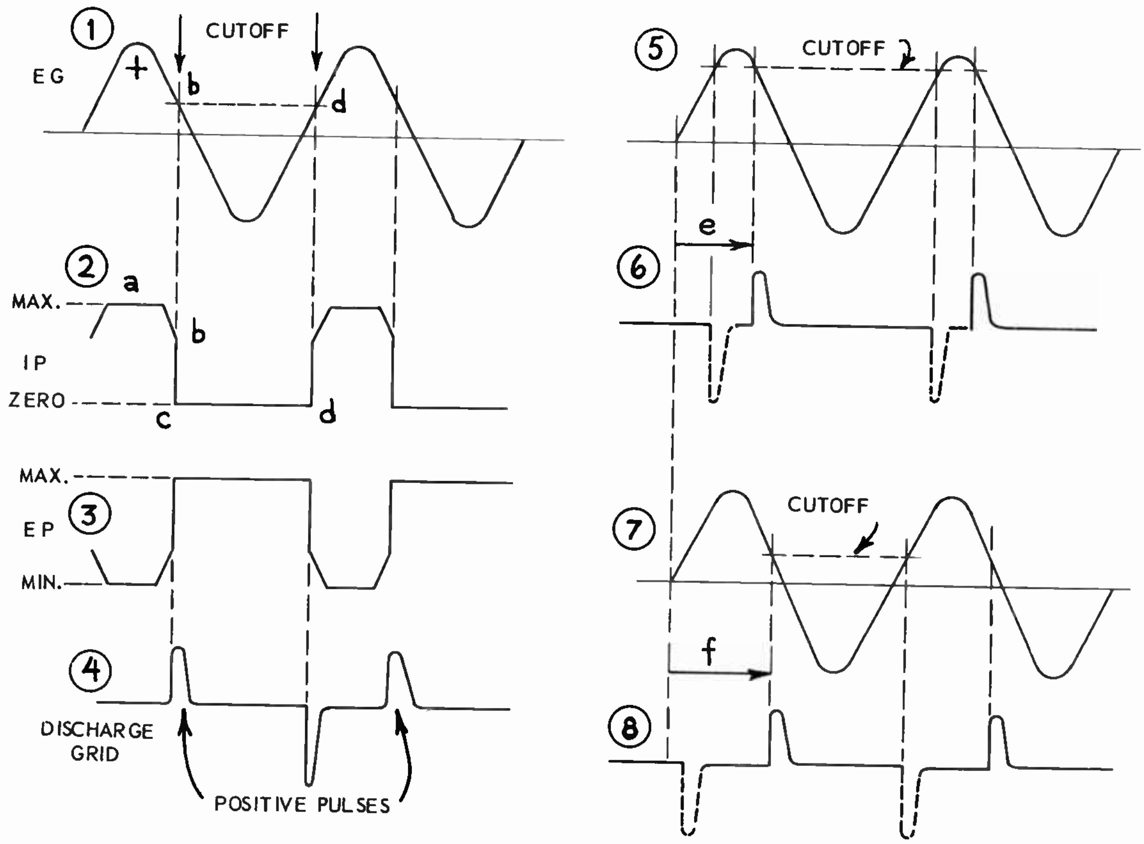


Fig. 56-16. How the hold control varies the timing of positive pulses to the grid of the discharge tube.

Now look at curves 5 and 6 of Fig. 56-16. The hold control resistance has been readjusted to change the negative bias applied to the oscillator grid by voltage across this resistance. Cutoff now occurs while grid voltage is more positive or occurs earlier during the sine wave. Arrow e indicates the time between zero voltage of the oscillation sine wave and the leading edge of a positive triggering pulse.

Next look at curves 7 and 8. Hold control adjustment has been changed to make bias voltage at the oscillator grid less negative. Now sine wave voltage must drop to a lesser positive value before there is plate current cutoff and a positive triggering pulse. Arrow f indicates the time between zero sine wave voltage and the triggering pulse. This time is longer than that indicated by arrow e. Thus we find that adjustment of the hold control varies the time of the triggering pulses in relation to the time of the oscillator frequency, as represented by the instants at which the sine wave goes through its various values.

There are three adjustments for the afc system of Fig. 56-14. One is the hold control, which is accessible to the set operator. There are two service adjustments, the movable core in the transformer primary, which usually is called the frequency control, and the movable core in the secondary, which usually is called the phasing control. The frequency control most often is a screw or nut on top of the transformer, while the phasing control is reached from the bottom of the transformer.

Adjust the frequency control as follows.

1. Tune in a signal, preferably a test pattern.

2. Turn the hold control slowly back and forth through its full range. The picture or pattern should hold throughout the entire adjustment range, or at least through the center half or two-thirds of the range.

3. If the hold control does not act thus, carefully adjust the frequency control to obtain the desired action.

The phasing control requires adjustment only if a vertical black bar appears on the screen or if the picture or pattern appears as though "folded" back on itself at either the right or left. Make the adjustment as follows.

1. Turn the hold control to the center of its range.

2. Adjust the contrast control to make the picture or pattern a little darker than usual, then adjust the brightness control so that sloping white lines just become visible.

3. Adjust the phasing control to move a vertical black bar just off the screen to the right. The edge of the picture or pattern should then be close to the left side of the screen area. Make the adjustment so that there is no folded appearance at either side, and no vertical white streak at the left side.

Check your adjustments as follows.

1. Rotate the hold control one way and the other. The picture or pattern should move to the right and left, and if folding appears it should be about equal at one side, then at the other.

2. With the hold control anywhere in the central half or two-thirds of its range the picture or pattern should pull back into sync when tuning is changed from channel to channel and when the set is turned off and then on again.

TELEVISION

LESSON NO. 57

DEFLECTION SYSTEMS

Training Your Mind...
C.T.I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

DO NOT TEAR — CUT ALONG THIS LINE AND SEND IN FOR GRADING.

Training Your Mind

C.T.I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON NO. 57

DEFLECTION SYSTEMS

The synchronizing portions of the television signal have been followed through the sync section and through the vertical and horizontal sweep oscillators. We have succeeded in obtaining sawtooth voltages that are held in synchronism with the received signal. Now we must use these sawtooth voltages for controlling vertical and horizontal deflection of the electron beam in the picture tube.

What we do with the sawtooth voltages after they leave the sweep oscillators or discharge tubes depends first on how the electron beam is to be deflected in the picture tube. If there is to be electrostatic deflection, sawtooth voltages must be applied to the vertical and horizontal deflecting plates which are inside the tube. These deflecting voltages will be of practically the same waveform that comes from the sweep oscillators, and may need no special treatment other than possible amplification.

For magnetic deflection we must have sawtooth currents, not sawtooth voltages, in the deflection coils which are placed around the neck of the picture tube. The original or amplified sawtooth voltages then must be applied to the primary of a transformer whose induced secondary current goes through the deflection coils. Unfortunately, a sawtooth voltage does not induce a sawtooth current of similar form when applied to a circuit having large inductance, such as one containing a transformer primary. Therefore, the sawtooth voltages from the sweep oscillators will require special treatment before they can cause sawtooth currents.

It is not too difficult to obtain satisfactory vertical magnetic deflection, because the deflection frequency is only 60 cycles per second. But when we come to the horizontal deflection frequency of 15,750 cycles per second there are many difficulties, and the circuits between horizontal sweep oscillators and horizontal deflection coils are much more complicated than those for the vertical system.

Because of their relative simplicity we shall first examine electrostatic deflection systems. Then we shall proceed to magnetic vertical deflection systems, and finally to the methods of securing horizontal magnetic deflection.

ELECTROSTATIC DEFLECTION. As we learned in earlier lessons, the electrons issuing from the cathode of a picture tube are focussed into a beam of very small diameter and accelerated to high velocity as they travel toward the screen. In an electrostatic picture tube this electron beam passes through spaces between pairs of vertical and horizontal deflecting plates. The velocity of the electrons is not changed as they pass between the plates, but their direction may be changed.

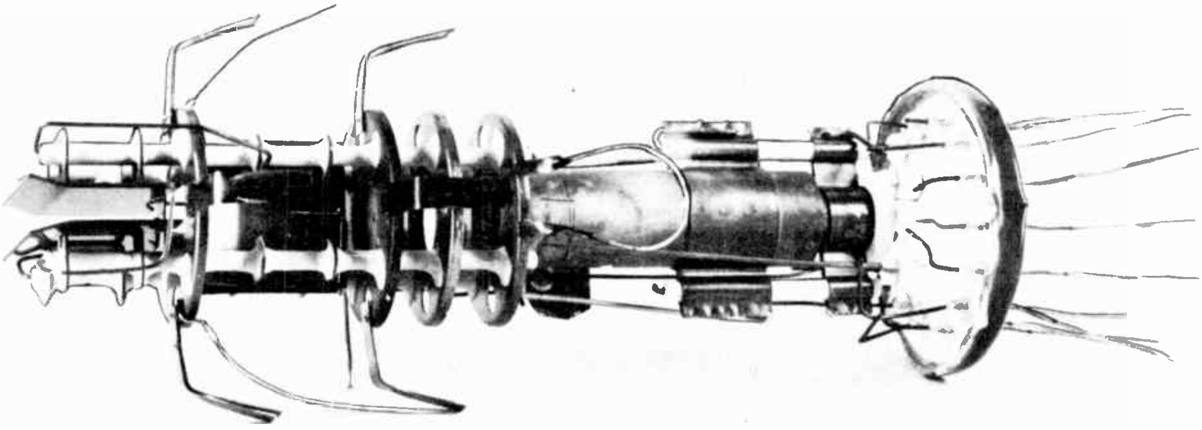


Fig. 57-1. The elements of an electrostatic deflection picture tube. The two pairs of plates are toward the left. Then comes accelerating and focusing electrodes. Grid and cathode are toward the right, just inside the glass press through which pass the wire leads.

The electrons in the beam are negative. If the two deflecting plates of a pair are at zero potential, or if both are at the same potential and have equal charges, the electrons are not influenced in any way. But if one plate is more positive than the other it attracts the negative electrons and changes the direction in which they are traveling. If one plate is more negative than the other it repels the negative electrons and changes their direction of travel.

In actual practice the potentials or voltages of the two plates in a pair are made to vary as shown by Fig. 57-2. There is an alternating sawtooth voltage on each plate. As one plate is driven positive the opposite plate is driven negative. Then the first plate is made negative and the opposite plate positive. At any one instant the electron beam is attracted toward whichever plate is positive, and at the same time is repelled by the opposite negative plate. As the plate voltages alternate, the beam is bent or deflected one way and the other. After the electrons leave the space between deflecting plates they travel straight to the screen in the direction of their deflection.

There are various ways of utilizing the single sawtooth voltage from the sweep oscillator to form the opposite-phase sawtooth voltages of Fig. 57-2. One method is illustrated by the circuit diagram of Fig. 57-3. At the left is represented any type of sweep oscillator or discharge tube in whose plate circuit is formed a sawtooth voltage by charge and discharge of capacitor <u>Cs</u>. This sawtooth voltage is applied to the grid of sweep amplifier tube <u>A</u>. The amplified sawtooth voltage, which also is inverted by the amplifier, goes from the plate of <u>A</u> to one of the deflecting plates of either the vertical or horizontal pair in the picture tube at the right.

The plate circuit load on amplifier tube <u>A</u> consists of resistors <u>Ra</u> and <u>Rb</u>. The strength or amplitude of the sawtooth voltage at the top of resistor <u>Ra</u> is the same as at the amplifier plate and the same as applied to the picture tube. At point <u>X</u> between resistors <u>Ra</u> and <u>Rb</u> the sawtooth amplitude is much less,

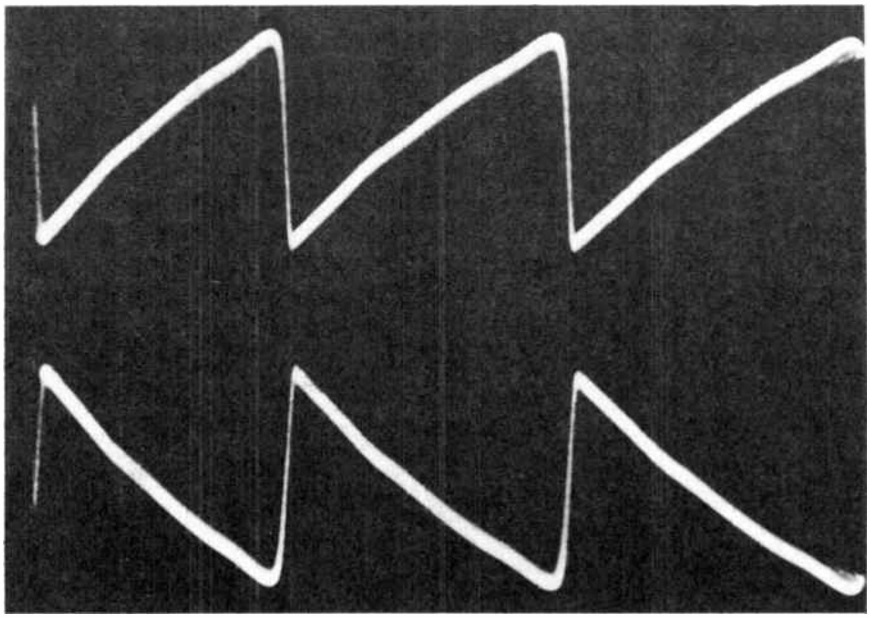


Fig. 57-2. Sawtooth voltages of opposite polarity fed to opposite plates of a pair in an electrostatic deflection tube.

and at the bottom of Rb the amplitude of this alternating voltage is practically zero. Resistors Ra and Rb form a voltage divider. The fraction of the original sawtooth amplitude remaining at point X is applied to the grid of tube B, which is the phase inverter. The phase of the sawtooth voltage at the grid of tube B is the same as at the plate of tube A. This voltage is amplified and inverted by tube B, and from the plate of this inverter tube the sawtooth voltage is applied to the other deflecting plate in the picture tube.

The inversion of polarity which occurs in tube B causes the sawtooth voltages to be in opposite phase at the two deflecting plates of the pair.

6 The sawtooth voltages on opposite plates in the picture tube must be of equal amplitudes. The equal amplitudes result from the voltage divider action of resistors Ra and Rb. This divider reduces the sawtooth amplitude at the grid of tube B, as compared with that at the plate of A. When this reduced amplitude is amplified by B it will be brought up to the same strength as at the plate of A.

The sweep amplifier and inverter ordinarily are the two sections of a twin tube. Both sections have the the same amplification when operated with equal plate voltages and grid biases. Sawtooth voltage taken from the plate of A to the grid of B must be reduced proportionately to the amplification occuring in B. For example, if amplification in B is 10 times then sawtooth grid voltage at B must be 1/10 of the sawtooth voltage at the plate of A. If B amplifies 20 times, the sawtooth voltage at its grid would be reduced to 1/20 of the sawtooth voltage or amplitude at the plate of A. Then sawtooth output from the plate of B will equal that from the plate of A.

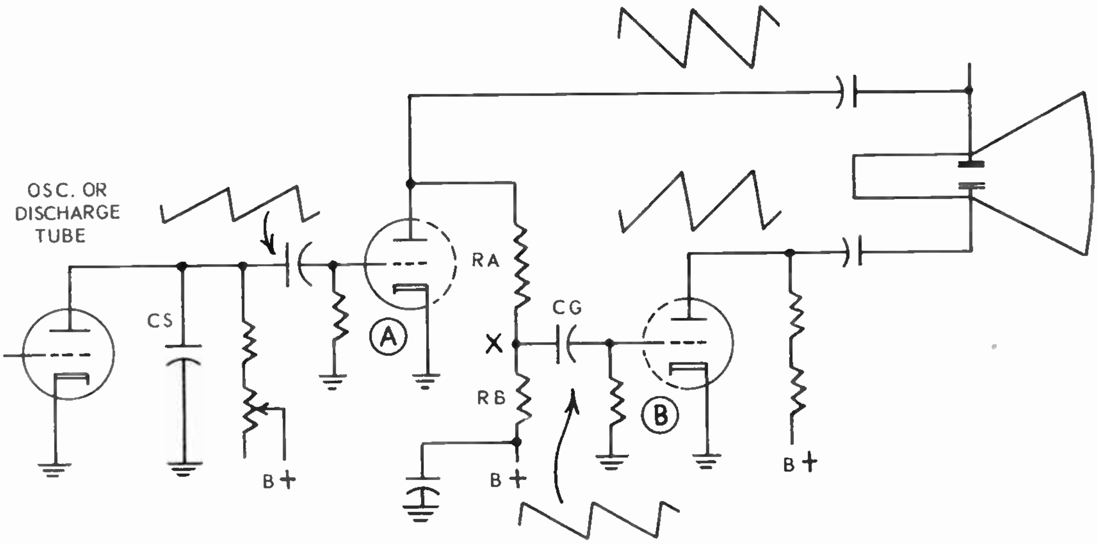


Fig. 57-3. Resistance voltage divider in phase inversion circuit for electrostatic deflection.

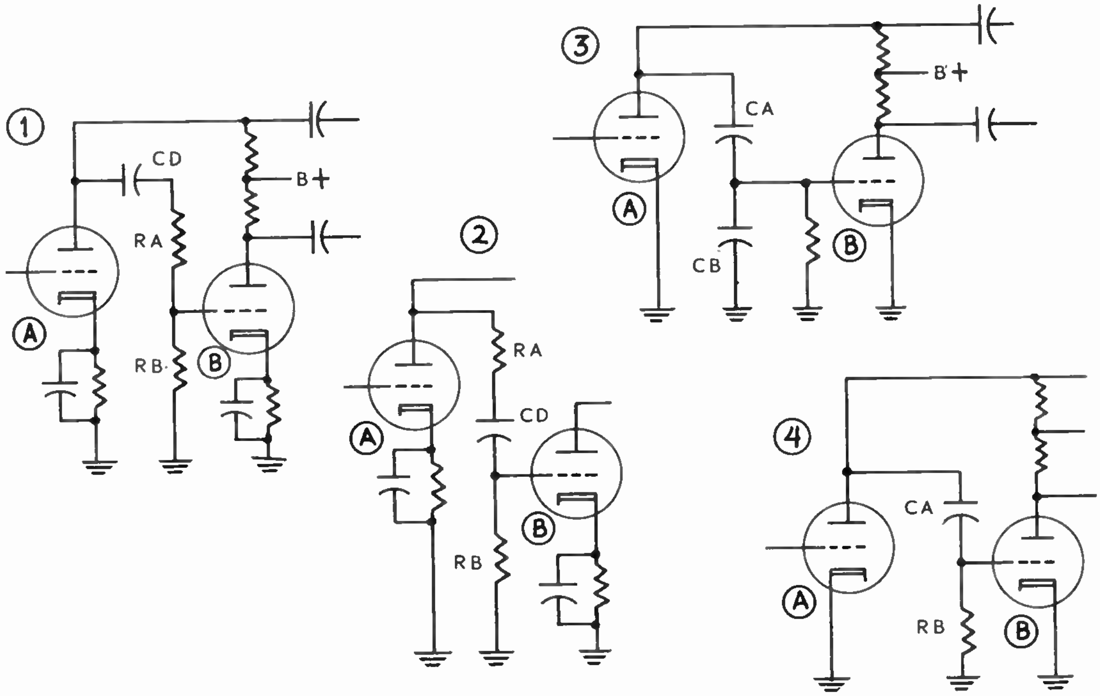


Fig. 57-4. Various voltage divider arrangements used for electrostatic deflection.

Fig. 57-4 shows some other types of voltage dividers used between the plate of a sweep amplifier and the grid of an inverter tube. In diagram 1 the lower end of divider resistor Rb goes to ground rather than to B plus. Sawtooth voltage from the amplifier plate is taken to the divider through blocking capacitor Cd. Always there must be a blocking capacitor to isolate the high d-c voltage on the plate of A from the grid circuit of B. In diagram 2 the blocking capacitor is between the two divider resistors.

In diagram 3 of Fig. 57-4 we have a capacitance voltage divider between the plate of amplifier A and the grid of inverter B. Voltage divider action results from the relative capacitive reactances in capacitors Ca and Cb. In order to have small sawtooth amplitude at the inverter grid the reactance at Ca must be much greater than at Cb, which means that capacitance at Ca will be small and at Cb will be relatively large. In diagram 4 the voltage divider action results from capacitive reactance in capacitor Ca and from resistance in resistor Rb. In both these latter diagrams capacitor Ca acts as a blocking capacitor.

An entirely different method of obtaining inversion of one sawtooth voltage is illustrated by the circuit of Fig. 57-5. One voltage is taken from the plate of the oscillator and applied to the grid of amplifier A. At the oscillator cathode is a similar voltage, but of opposite polarity. This inverted sawtooth is applied to the grid of amplifier B. We might consider that the inversion is due to taking the two sawtooths from opposite sides of sawtooth capacitor Cs, since one side of this capacitor becomes more negative as the other becomes more positive, and vice versa.

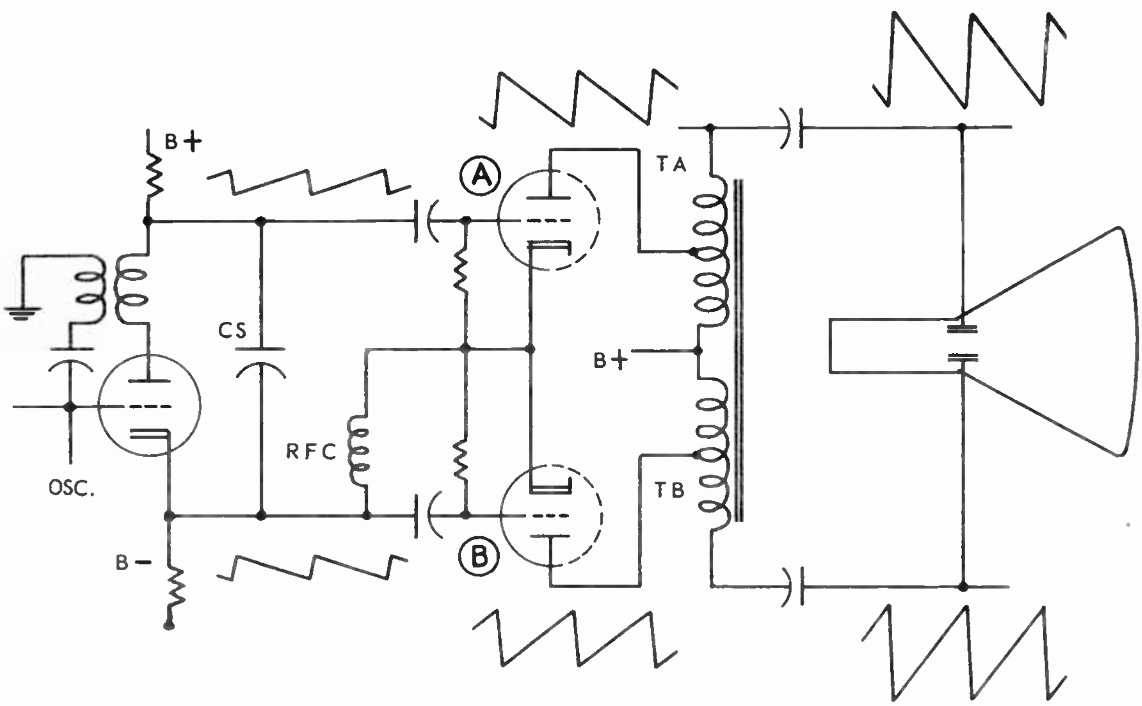


Fig. 57-5. Electrostatic deflection with two amplifiers and two auto-transformers.

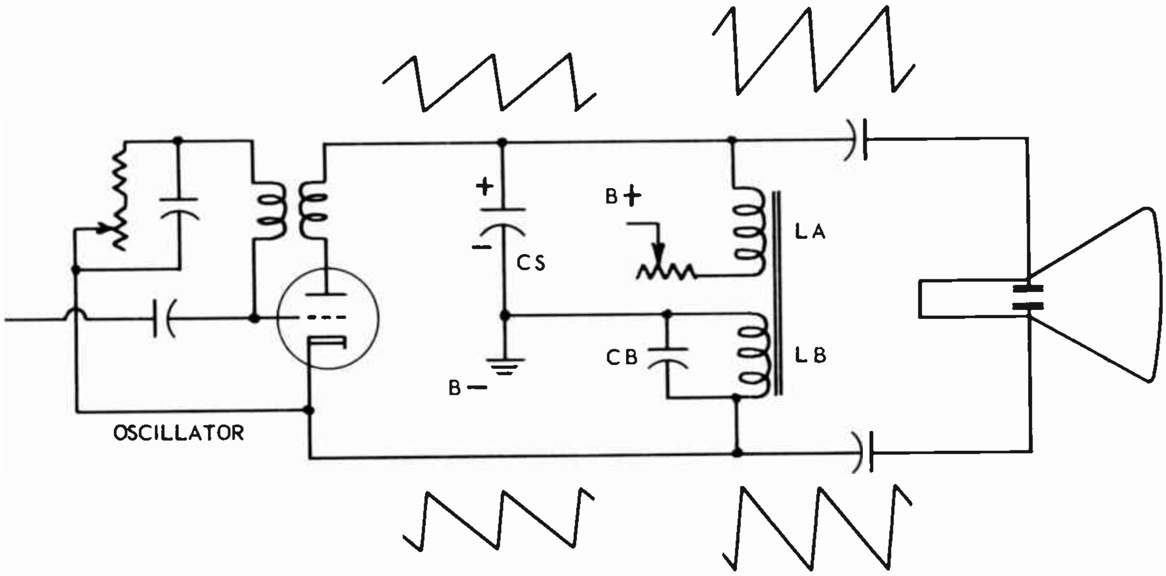


Fig. 57-6. Connections from sweep oscillator to picture tube plates, with no amplifier.

Sawtooth voltages from the plates of the two amplifiers are applied at the taps on auto-transformers Ta and Tb, which are wound on the same core. There is no polarity inversion between the tap and the high-voltage end of an auto-transformer, but there is a step-up of voltage. The stepped-up voltages of opposite polarity are applied to the two deflecting plates in the picture tube.

Fig. 57-6 shows how the plate and cathode of a sweep oscillator may be connected to the two plates of a deflecting pair without the use of an amplifier tube or tubes. While the oscillator is blocked and non-conductive the sawtooth capacitor Cs is charged in the marked polarity by electron flow from B minus or ground through this capacitor, through the winding of choke coil La, and through the size control resistance to B plus.

When the oscillator grid is made momentarily positive, and this tube becomes conductive, the sawtooth capacitor discharges through capacitor Cb to the oscillator cathode, through the tube to its plate, through the plate winding of the oscillation feedback transformer, and to the positive side of capacitor Cs. The very sudden change (increase) of current during capacitor discharge attempts to go through chokes La and Lb, but it induces such high counter-emf's in the chokes as to force practically the whole discharge current to flow through the conductive oscillator tube. The counter-emf or the "inductive kick" voltage induced in the chokes is greater than the applied voltage, with the result that sawtooth voltages going to the deflecting plates are considerably greater than those from the oscillator.

The sawtooth voltages applied to the deflecting plates are in opposite phase because one sawtooth comes from the plate side of the oscillator while the other comes from the cathode side, or one sawtooth comes from the positive side of capacitor Cs while the other comes from the negative side. The two choke windings are on the same core, and are so connected or so wound in relation to each other that their inductive voltages act together - each acts to increase the other. This effect contributes to the relatively high sawtooth voltages going to the deflecting plates.

MAGNETIC DEFLECTION. The coils for magnetic deflection of the electron beam are carried in a structure called a yoke which is placed around the neck of the picture tube just back of the cone or the flare. Fig. 57-7 is a picture of a set of deflecting coils with their protective coverings and the supports removed to expose the windings. The tube neck goes through the opening which you can see through the center of the coils.

There are four deflecting coils, two connected in series with each other for vertical deflection, and the other two in series for horizontal deflection. The two coils for vertical deflection are opposite each other and are mounted at the left-hand and right-hand sides of the tube neck, as shown by the upper row of sketches in Fig. 57-8. The magnetic field of these two coils then extends horizontally through the tube neck and through the space traversed by electrons of the beam in going from the electron gun to the screen of the tube. We must have horizontal field lines for vertical deflection because a magnetic field deflects electrons at right angles to the direction of the field lines.

The two coils for horizontal deflection are placed above and below the neck of the picture tube, as shown by the lower row of sketches in Fig. 57-8. The lines of the magnetic field between these two coils then are vertical, and when they deflect the electron beam at right angles to their own direction this deflection will be horizontal.

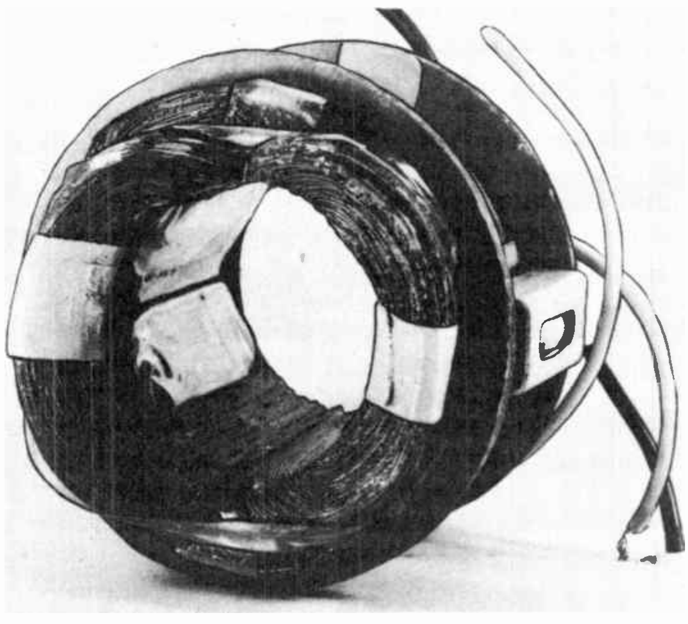


Fig. 57-7. The vertical and horizontal pairs of coils in a magnetic deflection yoke.

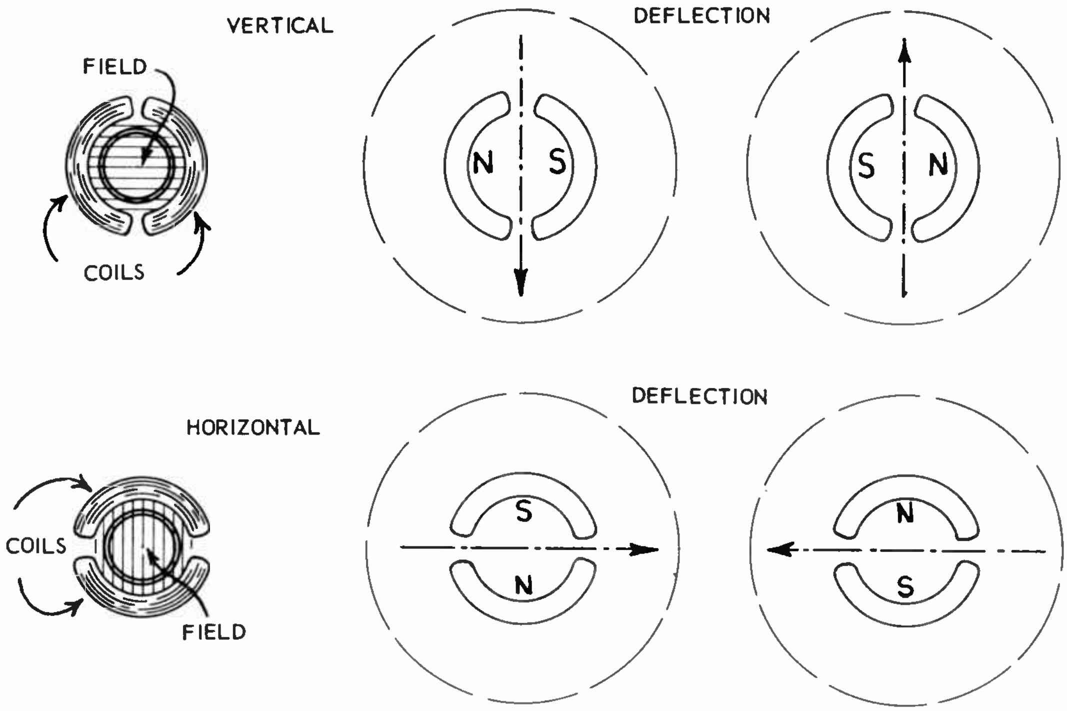


Fig. 57-8. Relation between magnetic field directions, coil positions, and directions of beam deflection when looking at the screen end of a picture tube.

The four diagrams showing directions of beam deflection are drawn with the assumption that you are looking toward the face or screen end of the picture tube, looking back through the cone or flare and through the center of the neck toward the base. When the direction of electron flow through the two vertical deflecting coils is such as to produce a north magnetic pole on your left and a south magnetic pole on your right the electron beam will be deflected downward. When coil current reverses there is reversal of the magnetic poles, and the beam is deflected upward. Of course, we have to remember that the beam is blanked during vertical retrace, but coil currents and magnetic fields are such as would cause upward deflection were there any beam to be deflected.

When electron flow through the two horizontal deflecting coils is in the direction that produces a south magnetic pole from the top coil and a north magnetic pole from the bottom coil the beam is deflected from left to right. When coil current and magnetic poles reverse, the beam would be deflected to the left were there any beam to be deflected during this retrace action.

The action of the magnetic fields that deflect the electron beam is entirely independent of the action of the electrostatic fields that accelerate the electrons. The two kinds of fields each act as though the other field were not present. A magnetic field will deflect electrons only when the electrons are in motion, and only when the motion is in a direction not parallel to the lines of the magnetic field. The magnetic deflecting fields have no effect on velocity of the electrons.

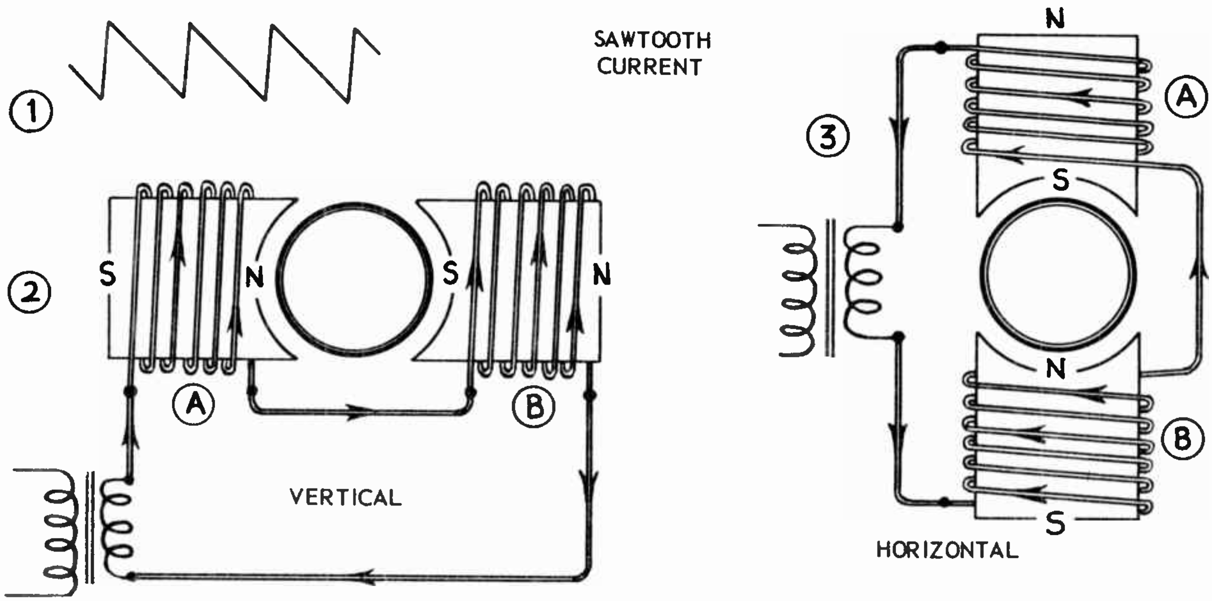


Fig. 57-9. Directions of electron flows in deflecting coils for establishing sawtooth currents and magnetic fields of opposite phase.

For deflection of the beam in the picture tube we require magnetic fields whose polarities are opposite at each instant of time, just as for electrostatic deflection we require electric fields produced by voltages which are opposite or in opposite phase at each instant of time. The magnetic fields must change their strengths in sawtooth fashion in order to have fast retrace times and relatively slow trace times. These magnetic fields are produced by sawtooth currents in the deflection coils, as shown at 1 in Fig. 57-9.

The two coils for vertical deflection are connected in series with each other and with the secondary winding of the sweep output transformer, whose primary is connected to the plate circuit of the vertical sweep output amplifier tube. This is shown by diagram 2. The two coils for horizontal deflection are similarly in series with each other, and are in series with the secondary of the horizontal sweep output transformer whose primary is in the plate circuit of the horizontal sweep output amplifier tube. This is shown by diagram 3.

The magnetic fields which are in opposite phase at the ends of the coils that are toward the neck of the picture tube are secured by suitable directions of current in the coils. These current directions are shown by diagrams 2 and 3 of Fig. 57-9. Arrowheads indicate directions of electron flow in the coil circuits during one alternation. Opposite magnetic polarities are produced on opposite sides of the tube neck, resulting in a magnetic field that deflects the beam in one direction. This deflection is shown in Fig. 57-8. When electron flow reverses, the magnetic field polarities reverse, and the beam is deflected in the other direction. Because electron flows or currents in the coils are of sawtooth waveform we have the required fast retrace times and the relatively slow trace times.

Fig. 57-10 shows complete circuit connections for a vertical sweep system that you will find in a number of receivers. At the left is the vertical sweep oscillator, shown here as a blocking type. Then comes the vertical output amplifier. At the right are the two vertical deflection coils. Waveforms of currents as observed on an oscilloscope are marked I. Waveforms of voltage are marked E.

In the deflecting coils we have a current whose waveform is a nearly perfect sawtooth. But the voltage across these coils is a weak sawtooth with very strong and exceedingly brief positive pulses at every retrace period. By making tests or measurements at the primary of the output transformer or at the plate of the output amplifier we find similar conditions. The current is a well formed sawtooth, with only a slight curvature in the long slopes. But again the voltage across the primary, from amplifier plate to the B plus connection, is a very weak sawtooth with high positive pulses at the instants of retrace.

Continuing this investigation we go next to the grid of the amplifier tube. Here we have the same voltage waveform as at the plate, but the polarity is inverted because there always is such inversion between grid and plate. At the amplifier grid there is a sawtooth voltage with strong negative peaks at the instant of every retrace period. This same voltage waveform appears at the oscillator output, or at the upper connection to sawtooth capacitor Cs. Voltage at the cathode of the amplifier tube remains very nearly constant because capacitance of bypass Ck is something from 50 to more than 200 microfarads, and it absorbs and smooths out the changes of cathode current to leave a nearly steady bias voltage for the amplifier grid.

There are many makes and types of receivers in which the grid-cathode circuit of the vertical output amplifier are connected as in the small diagram at the bottom of Fig. 57-10 rather than as up above. In the small diagram the lower end of the resistor Rs in series with sawtooth capacitor Cs is connected to the amplifier cathode rather than to ground. Charge and discharge of the sawtooth capacitor then goes through cathode resistors Rk on the amplifier. The waveforms at all points in this circuit, through to the deflection coils, are practically the same as shown on the main diagram.

NEGATIVE PEAKING. In the grid circuit of Fig. 57-10 we have observed the effect called negative peaking, with which there is a negative dip or peak at the instant of each retrace period, preceding the trace portion of the waveform. This negative peaking, which changes to positive peaking on the plate side of the sweep amplifier, is required because there is a great deal of inductance in the output transformer and in the deflection coils.

The portions of the sweep circuit from the amplifier plate to and including the deflecting coils contain both inductance and resistance. If a steady voltage is applied to such an inductive-resistive circuit the current will not instantly rise to a value proportional to applied voltage, but due to counter-emf induced in the inductance will rise rather slowly to that value. The magnetic field resulting from the current will increase its strength in the same manner and at the same rate as the current. If applied voltage then is suddenly cut off, the current and the magnetic field will not instantly drop to zero. Rather they will decrease somewhat slowly, at a rate determined by the inductance-resistance time constant of the circuit. All this is simply a re-statement of principles we learned long ago.

Instead of working with an applied voltage that changes between maximum and zero we may try an alternating square wave voltage as in diagram 1 of Fig. 57-11. Resulting current in the inductive-resistive circuit is shown below the voltage wave. The current falls to zero and goes into the opposite polarity

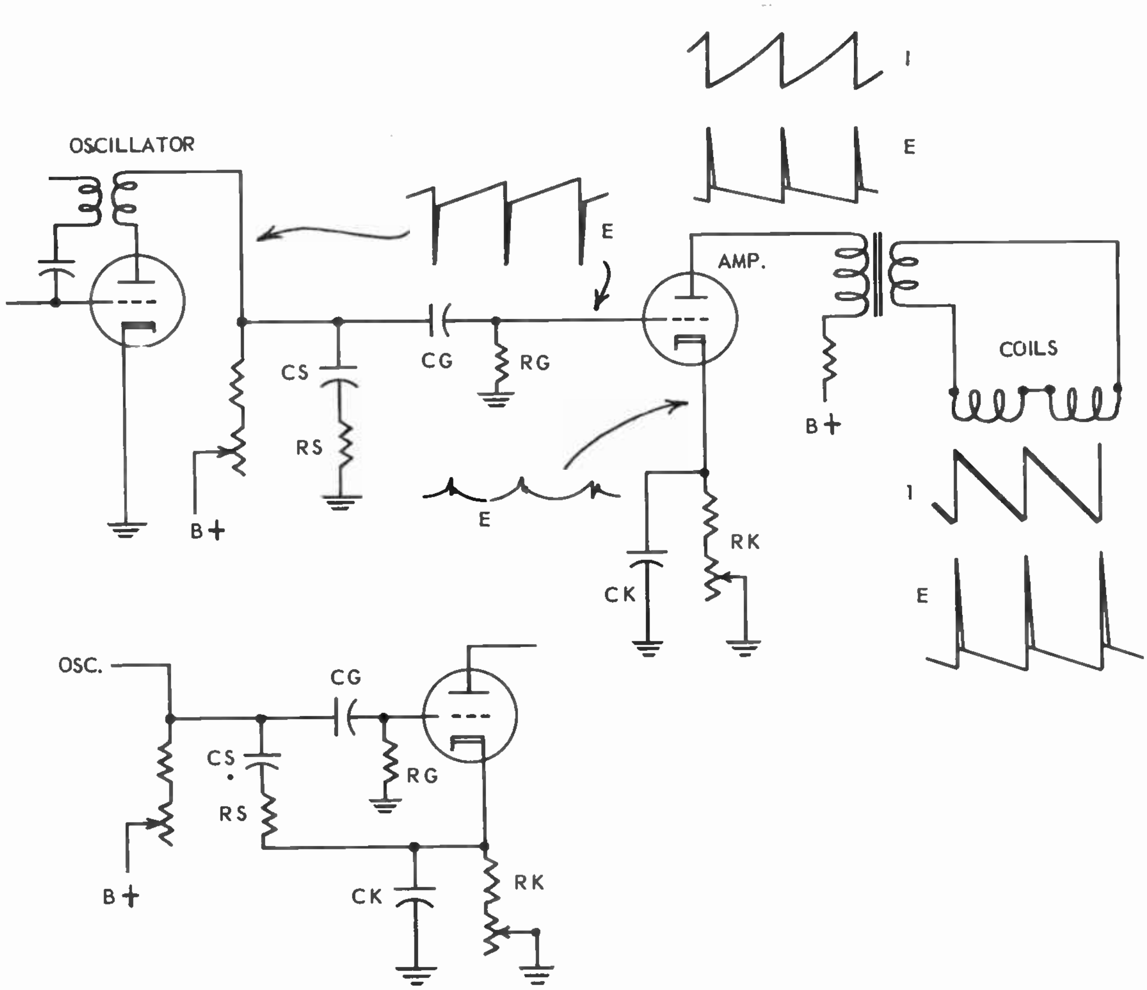


Fig. 57-10. Connections and waveforms of currents and voltages in a vertical magnetic deflection system.

some time after the reversal of voltage polarity, because of delay proportional to the time constant. This is somewhat on the order of a sawtooth, but times for retrace and trace would be equal. The retrace time must be greatly shortened.

In diagram 2 strong peaks of voltage have been added at each reversal of polarity. These voltage peaks force the current to reverse its direction much more suddenly than before, but after reversing so quickly the current continues to increase only according to the inductance-resistance time constant of the circuit. Still we have no satisfactory sawtooth current wave.

An improvement is made by retaining only the positive peaks of voltage and removing the negative

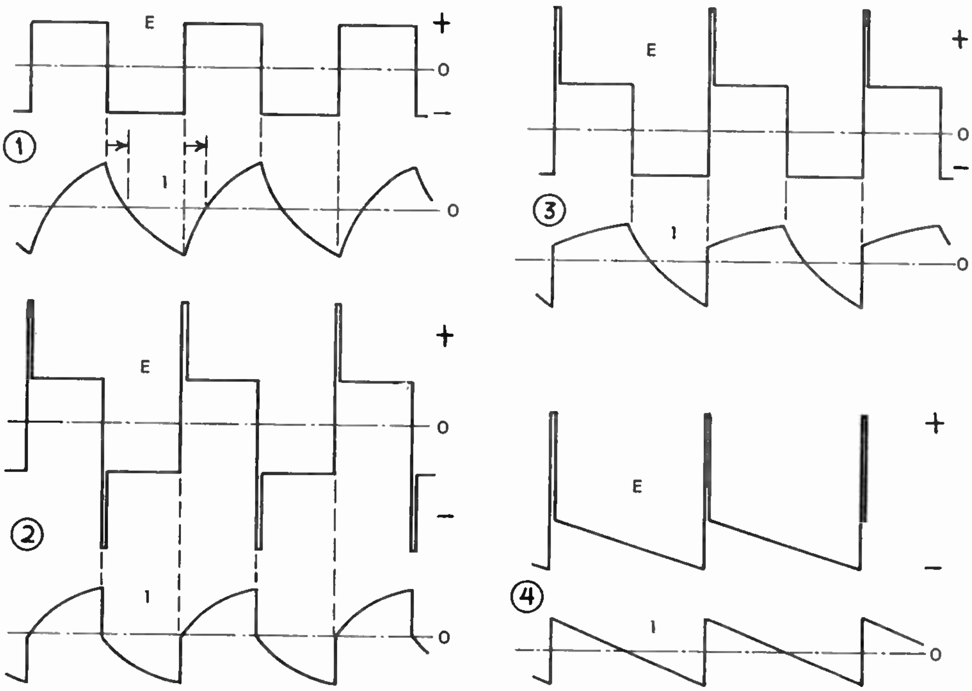


Fig. 57-11. Why voltage peaking is needed in order to have a sawtooth current.

peaks, as in diagram 3. The positive peaks cause the desired sudden reversals of current. However, the abrupt reversal of voltage in the middle of each cycle puts a hump in the current curve.

In diagram 4 the positive peaks of voltage are retained, but voltage changes at a uniform rate between the peaks. The necessary sudden reversals of current are present, but the hump in the current wave is removed by the steadily changing voltage. The current changes smoothly as it decreases in one polarity, goes through zero, and increases in the other polarity.

The strong positive peaks of voltage applied to the deflection coils are needed in order that current in the coils may reverse its direction within the very brief time allowed for retrace. Otherwise the retrace periods would last as long as the trace periods. The positive peaks in the voltage output of the sweep amplifier are caused by the negative peaking on the grid side.

How much peaking is needed depends largely on the relative values of resistance and of inductive reactance in the circuits. There is resistance in all the conductors and in the plate resistance of the sweep amplifier. There is inductance in primary and secondary of the output transformer, and in the deflection coils. The greater the proportion of inductance the stronger must be the negative peaks. Were the circuit to contain only resistance, with no inductance, a sawtooth current would result from a plain sawtooth voltage wave, and current waveform would be the same as voltage waveform.

Negative peaking in the circuits of Fig. 57-10 is produced by resistance at Rs or any other resistance which is in series with the sawtooth capacitor. Electron flow through the sawtooth capacitor and series resistor during charging of the capacitor is shown by arrows in the left-hand diagram of Fig. 57-12. Charging current flows to B plus through the size control and associated resistors, whose total resistance ordinarily is on the order of several megohms. Consequently, there is relatively slow charging, as from a to b on the voltage wave at the center of the figure. Resistance at Rs usually is something between 5,000 and 10,000 ohms, and is so small in comparison with resistance in the size control connection as to have no appreciable effect on either the time of charge or the voltage to which the capacitor is charged.

Electron flow during discharge of the sawtooth capacitor is shown by arrows in the right-hand diagram of Fig. 57-12. The only resistances in the discharge circuit are the small ones at Rs and in the cathode-to-plate path inside the oscillator or discharge tube. Practically all the electrons which have been accumulated in the sawtooth capacitor pass through the resistance at Rs within the very brief time allowed for retrace. Now we must remember that current is defined as a rate of electron flow, coulombs per second. Even though capacitor Cs does not hold a great quantity of electrons, the discharge is so very fast that the rate in coulombs per second is high, and there is thus a large current through resistor Rs.

We must remember also that the potential difference across a resistance is directly proportional to the product of current times resistance. Consequently, the large discharge current through only the moderately great resistance at Rs causes a large potential difference across this resistance. The direction of elec-

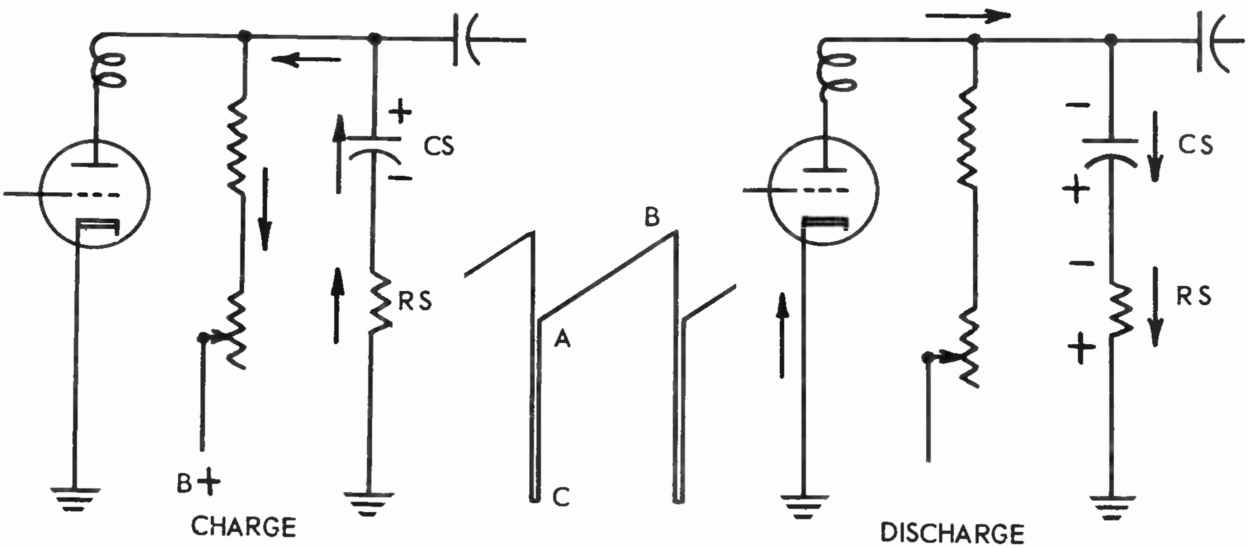


Fig. 57-12. Electron flows through series resistor during charge and discharge of the sawtooth capacitor.

tron flow is such as to make the upper end of Rs negative with reference to ground. Thus there is produced the strong negative peak shown between a and c on the voltage wave diagram. This peak exists only during the fast discharge of the sawtooth capacitor, while the oscillator or discharge tube is conductive. As soon as the tube becomes non-conductive, which is almost instantly, the sawtooth capacitor again is charged as before.

The strength of depth of the negative voltage peak is increased by increasing the resistance of resistor Rs, which is directly in series with the sawtooth capacitor and which is not bypassed by any other capacitor. Decreasing this resistance lessens the negative peak. When, as in the small diagram of Fig. 57-10, the sawtooth capacitor discharges through the cathode resistor Rk on the output amplifier, a change of resistance at Rk has practically no effect on the degree of peaking when Rk is bypassed by a very large capacitance at Ck. There are a number of receivers in which the sawtooth series resistance at Rs is made adjustable as a means for securing a correctly shaped sawtooth current in the deflection coils. Such an adjustable resistance sometimes is called a linearity control, because it helps in making the trace portions of the sawtooth appear as straight slanting lines, or linear. Sometimes this adjustable resistance is called a drive control.

HORIZONTAL MAGNETIC DEFLECTION. The circuits which have been described and shown on the input or grid side of the sweep amplifier tube are found in both vertical and horizontal sweep systems for magnetic deflection tubes. In vertical sweep systems the parts and connections on the plate or output side of the amplifier are simple, consisting chiefly of the vertical output transformer with its primary connected to the amplifier plate and its secondary to the deflecting coils.

But between the plate of the horizontal sweep amplifier and the horizontal deflection coils the circuits are not simple. There are a number of important parts not found in vertical systems, and many new principles must be investigated. Furthermore, the circuits on the grid side of horizontal output amplifiers may be quite different from anything that we have studied.

Many of the added features in horizontal output systems are required because of the rather high operating frequency, 15,750 cycles per second instead of the 60 cycles per second in vertical systems. Other added features allow the horizontal output circuits to perform such extra work as supplying voltage higher than any B-plus voltage that comes from the regular power supply. This higher voltage, used for plates and screens of various tubes, may be about 100 volts greater than taken from the B-power supply, and for the anode of the picture tube it may be between 8,000 and 12,000 volts or more.

The actions which occur in these horizontal output systems are so interrelated and so dependent on one another that they cannot be considered separately until we have a general understanding of all the things that happen simultaneously. To gain such an understanding we shall examine the circuit diagram of Fig. 57-13, which applied to a fairly typical television receiver using magnetic deflection.

The sawtooth voltage supplied to this system comes from the sweep oscillator and is developed across sawtooth capacitor Cs in the usual way. Sawtooth voltage for the grid of the amplifier tube is not taken directly from the top of the sawtooth capacitor through grid capacitor Cg, but passes first through capacitor Ca.

Capacitors Ca and Cb are in series with each other, and the two are in parallel with the sawtooth capacitor. Sawtooth voltage from Cs is applied across Ca and Cb. These latter two capacitors form a capacitance voltage divider. The total voltage from Cs divides proportionately to capacitive reactances in the two elements of the divider. For example, if the reactance of Ca is twice that of Cb then twice as much of the sawtooth voltage will be across Ca as across Cb. You might consider the total applied voltage as divided into three equal parts, with two-thirds across the greater reactance at Ca and one-third across the lesser reactance at Cb. With such a division, only one-third of the total sawtooth voltage would be applied to the amplifier grid, because the grid-to-ground circuit is across capacitor Cb.

Capacitor Cb is adjustable. It is called a drive control. When this capacitance is reduced the reactance is increased, and a stronger sawtooth voltage is applied to the amplifier grid. If capacitance at Cb is increased there is a decrease of capacitive reactance, and a weaker sawtooth voltage is applied to the amplifier grid. When a stronger sawtooth voltage is applied to the amplifier grid the effects of the increase carry all through following circuits, and there is greater horizontal deflection and a wider picture because of greater amplitude of sawtooth current in the deflecting coils. Adjustment of drive control capacitance has other effects even more important than change of picture width, as we shall learn when working with methods of setting this control.

In Fig. 57-13 there is no resistor in series with the sawtooth capacitor to provide negative peaking. There is, however, a moderate amount of peaking due to discharge of grid capacitor Cg through grid re-

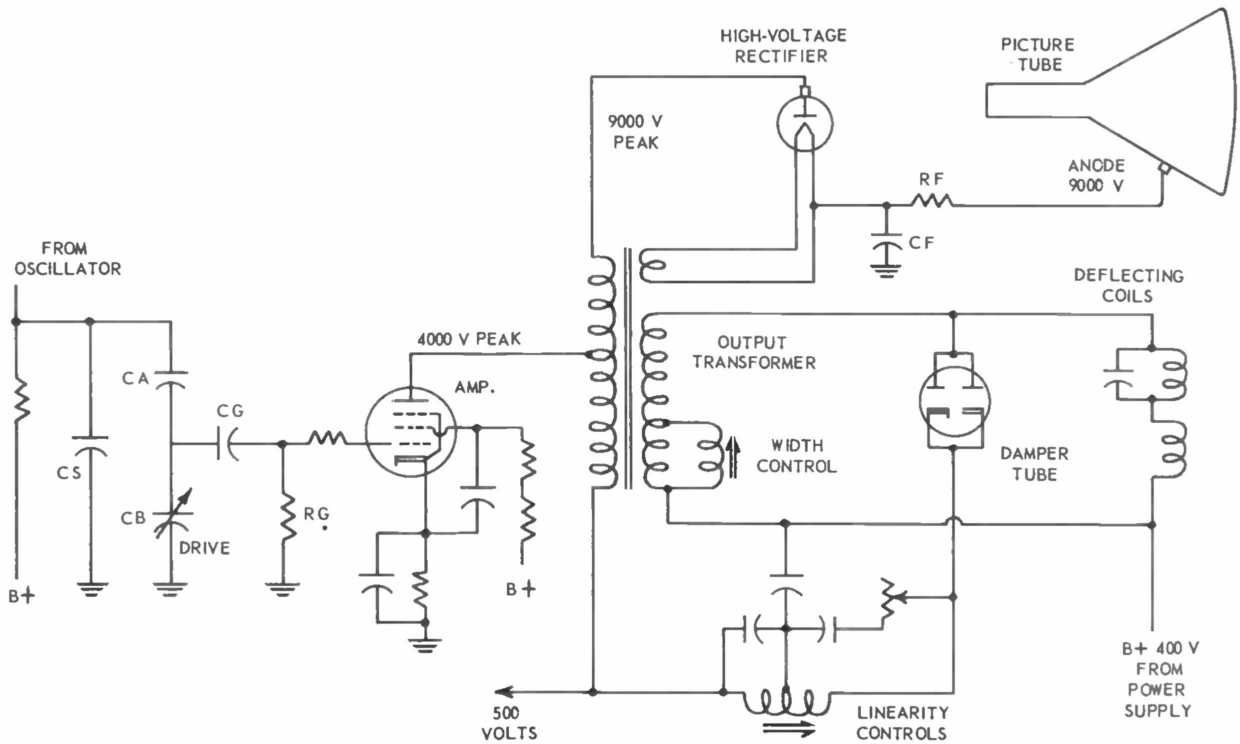


Fig. 57-13. Connections for one of the more common types of horizontal magnetic deflection systems.

sistor Rg at the same time that the sawtooth capacitor discharges through the oscillator or discharge tube. Capacitors Ca, Cb and Cg all are connected between the top of the sawtooth capacitor and ground, so that all charge and discharge in the same manner. Discharge of Cg through the grid resistor is in such direction as to make the grid end of this resistor negative during the brief instant in which peaking is needed.

Only a moderate amount of negative peaking is needed in the horizontal circuits. One reason is that inductances may be and are smaller than in vertical circuits, due to the higher operating frequency. There is more resistance in the horizontal circuits because the amplifiers are pentode types having plate resistances around 30,000 to 40,000 ohms, while vertical sweep amplifiers usually are medium-mu power triodes with plate resistances around 6,000 to 7,000 ohms.

The plate of the horizontal sweep amplifier is connected to a tap on the primary winding of the output transformer. The lower section of this primary acts in the usual way to induce a secondary emf and a sawtooth current for the horizontal deflecting coils, to which the secondary is connected.

When the sawtooth grid voltage on the amplifier tube goes suddenly negative and to a negative peak, there is equally sudden cutoff of plate current, which is the current in the transformer primary. Then the magnetic fields which have been built up in the transformer and the deflecting coils begin to collapse. The rate of collapse is determined by the inductance-resistance time constant of the circuits.

The abrupt change of current in the transformer secondary and the deflecting coils starts a cycle of oscillation in this circuit. The frequency of this oscillation is that corresponding to the circuit inductance and all the stray and distributed capacitances, which are small enough to have a natural resonant frequency much higher than the horizontal deflection frequency. Were the oscillation allowed to continue it would badly distort the trace portion of the sawtooth current wave.

When the oscillating current and voltage gets half way through its first cycle and goes into a positive alternation the positive voltage acts on the plates of the damper tube to cause heavy conduction through this tube. The tube acts almost like a short circuit at this time, and loads the transformer and coil circuit so heavily as to stop the oscillation. The purpose of the damper tube is to prevent continued oscillation at the natural frequency of the deflecting coil circuit.

When the magnetic field around the transformer collapses at the instant of plate current cutoff, the contracting lines of force cut through the primary as well as the secondary. The high rate of cutting, due to sudden change of magnetic field strength, induces in the primary winding a pulse of emf having a peak value of several thousand volts. The value is arbitrarily marked as 4,000 volts on Fig. 57-13. In any case this voltage pulse is so strong that touching the amplifier plate terminal or any conductors connected to it will produce painful results if not more serious personal injury.

The two sections of the primary make this winding act as an auto-transformer, stepping up the pulse peaks to something like 9,000 volts. This very high pulsating voltage goes to the plate of a half-wave rectifier tube. Positive voltage pulses from the rectifier filament are smoothed by the filter consisting of capacitor Cf and resistor Rf. The filter time constant is enough longer than the horizontal pulse frequency to provide a d-c voltage nearly equal to the peak pulse voltage. This d-c voltage is used for the anode of the picture tube.

The high voltage system just described usually is called a flyback type because it depends on voltage pulses produced during the retrace or flyback periods. Fig. 57-14 is a picture of a horizontal output transformer designed for use in a flyback system. In this particular unit the core extends all the way around the outside of the windings and through their centers. Between insulating discs on the side of this transformer which is toward you are two turns of wire acting as a secondary to provide heating current for the high-voltage rectifier. The outer edge of one of the other windings is visible. These other windings are of the duolateral type, in the form of discs and rings around the central part of the core. The terminal plate, of insulating material, is on the far side of the windings.

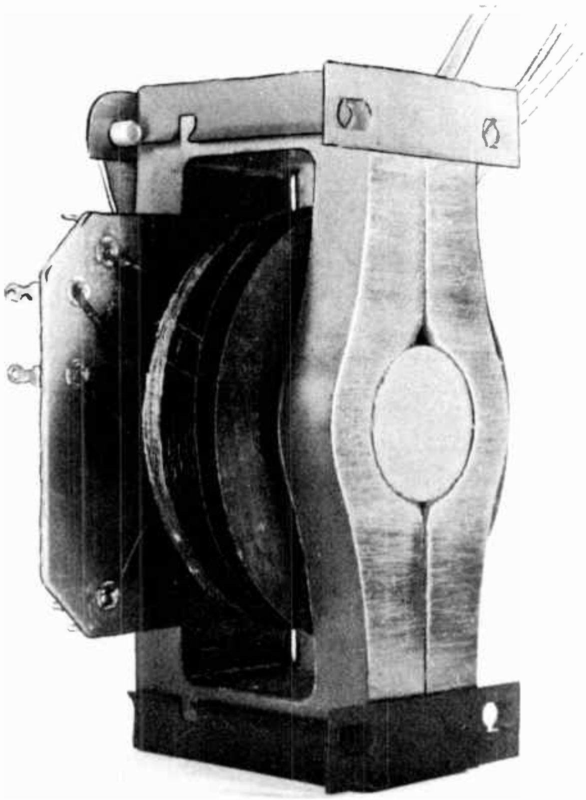


Fig. 57-14. A flyback type of output transformer used in horizontal magnetic deflection systems.

2 Connected across part of the turns in the main secondary winding of the output transformer in Fig. 57-13 is a width control. This control usually has the form of a coil an inch or so long and about 3/8 inch in diameter. Its inductance is adjustable by a movable core.

Increasing the inductance of the width control inductor, by turning the core farther into the coil, will make the picture wider and wider until the core is centered lengthwise to give maximum inductance. It would be rather natural to assume that disconnecting this added inductor from the transformer secondary would make the picture very narrow. Yet if you do disconnect one end of the width control coil from the transformer winding the picture will be wider than with any setting of the core in the control unit. The reason is something for you to figure out. It will be explained later, when we consider this and all other types of width controls and their adjustment. All that you need do to solve the problem is remember what happens to inductance and to inductive reactance when inductances are in parallel.

Returning once more to Fig. 57-13, notice the coil-capacitor combination marked Linearity Controls. Then follow the path of direct electron flow through the output amplifier and to the connection for B-plus from the power supply at the lower right-hand corner of the diagram. From ground (B-minus) the flow is to the amplifier cathode, through this tube to its plate, through the lower section of the primary winding of the output transformer, thence through the coil of the linearity controls to the cathode of the damper tube, through this tube to its plate, and then through the transformer secondary and deflecting coils to B-plus.

We are assuming that 400 volts positive potential is available from the B-supply of the receiver. This potential goes to the damper tube. The damper is conductive during the greater portion of every horizontal deflection cycle. The alternating deflection voltage is rectified by the damper tube, and the rectified voltage is added to the B-plus voltage from the regular power supply. The combination is a pulsating direct voltage. The coil of the linearity control and the capacitors between this coil and the deflecting coil circuit on the B-plus side act much like the choke and capacitors in the filter of an ordinary power supply, they partially smooth the pulsations coming from the damper tube. Then we have on the end of the linearity control coil that is connected to the output transformer primary a voltage about 100 volts higher than that from the regular B-plus line.

The energy removed from the deflecting circuit for the primary purpose of stopping oscillations is converted into additional positive voltage applied through the transformer primary to the plate of the output amplifier. This is called the booster action of the damper circuit. The boosted positive voltage is used also for plates and screens of other tubes in the receiver.

The primary purpose of the linearity controls is not to smooth the boosted positive voltage but to help obtain linear horizontal deflecting currents. A linear sawtooth current is one that changes at a uniform rate from the end of one retrace to the beginning of the next retrace, it is a current for which this portion of the waveform would be a straight line. A linear sawtooth current deflects the electron beam at uniform velocity all the way from left to right across the screen, and objects and outlines of the pictures are distributed in precisely correct relative positions or just as they are distributed in the original televised scene.

If there is curvature or wavering along the portion of the sawtooth wave that should be linear there will be any of a wide variety of distortions in reproduced pictures and patterns. The subject of linearity, how it is caused and how corrected, is much too extensive to get into right now, but it will come later.

In the horizontal sweep section which has been examined all the many functions have been carried out with particular parts and circuits. Nearly all these things may be done also in one or more other ways. In lessons immediately following we shall look at the other ways, and for all the methods to be studied we shall learn how service adjustments are to be made.

Do Not Tear — Cut Along This Line and Send In For Grading.

TELEVISION

LESSON NO. 58

FOCUSING, CENTERING, AND SIZE CONTROL

Training Your Mind

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

COMMERCIAL
TRADES
INSTITUTE

Chicago, Ill.

LESSON NO. 58

FOCUSING, CENTERING, AND SIZE CONTROL

In order that television receivers may sell at prices within the reach of millions of people there must be mass production, using parts of usual commercial tolerances. When such receivers come off the factory line they require alignment and other adjustments to compensate for normal variations in their parts and circuits. This is the first reason for providing so many adjustable capacitors, inductors, and resistors.

Even though a receiver were to be in perfect operating condition when leaving the factory it would not continue so. Parts and their connections are sure to be affected by jolts and jars and other "accidents". There will be inevitable changes in the characteristics of all the tubes and of many other circuit components, due to continued use and the natural processes of aging. This is the second reason for providing so many adjustments, so that you may bring back the original perfection of performance when trouble occurs.

So far in our work we have looked at many of the adjustments provided in tuners, i-f amplifiers, and video detector circuits, but as yet have not learned how these adjustments are to be handled when servicing a receiver. We have examined contrast and brightness controls, and have learned how each affects the other when their adjustment is changed. We have looked at hold controls and at frequency controls for afc systems in sweep oscillator systems, and have learned how most of them should be adjusted.

But still to be considered are all the various operating principles and methods of making service adjustments for focus, for vertical and horizontal centering, for vertical and horizontal size (height and width), for peaking, drive, and for linearity. We shall commence working on this list by considering the matter of focusing.

If there is to be good definition and clarity of detail in reproduced pictures, the electron beam must be made of such small diameter as to produce the smallest possible spot or narrowest line of light on the picture tube screen. This is accomplished by focusing the beam. Focusing is needed to counteract the natural tendency of the beam to spread out and form a large irregular spot of light on the screen. The beam tends to spread because all its electrons are negative. Negative charges repel one another and try to get as far apart as possible.

There are two basic methods of focusing now in use. One method utilizes differences of potentials on elements inside the picture tube. The different potentials cause electrostatic fields to appear between them, and the fields act to focus the electron beam. This is electrostatic focusing.

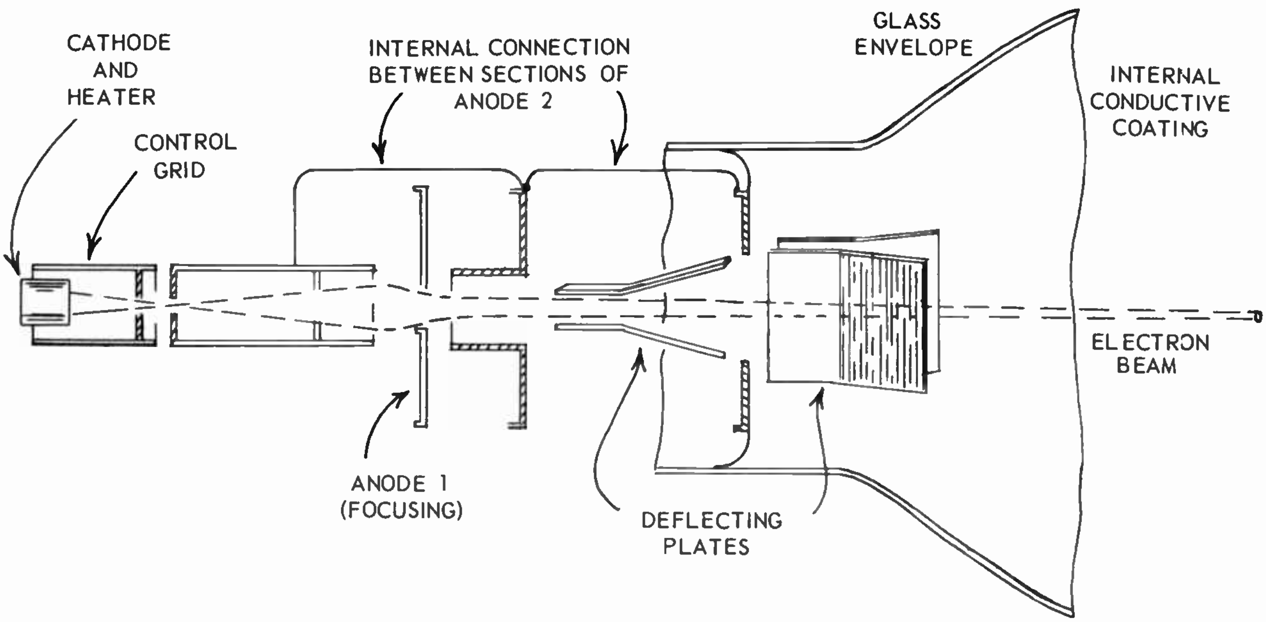


Fig. 58-1. The tube elements that take part in electrostatic focusing.

The other method utilizes direct current in an electromagnetic coil, or uses a permanent magnet, or a combination of electromagnet and permanent magnet to produce a magnetic field that focuses the beam. This is electromagnetic focusing or magnetic focusing.

Picture tubes designed for electrostatic deflection are constructed with the internal elements required for electrostatic focusing. This method of focusing may be used also in tubes operated with magnetic deflection. However, the great majority of picture tubes operated with magnetic deflection are designed also for magnetic focusing.

ELECTROSTATIC FOCUSING. Fig. 58-1 is a sectional drawing showing the arrangement of elements inside an electrostatic deflection tube commonly used in television receivers. At the left is the cathode shell which supports the electron emitting material. Inside the cathode shell is the heater. Almost completely surrounding the cathode is the control grid. Just to the right of the control grid is a cylinder that is the first section of a multi-section element usually called anode 2, or sometimes grid 2.

To the right of the first section of anode 2 is a disc-shaped element called anode 1. This is the focusing electrode. Then comes a short cylinder mounted on another disc to form the second section of anode 2. This is followed by two pairs of deflecting plates, used for horizontal and vertical deflection. A disc between the plates is the third section of anode 2. On the outer edge of this disc are thin flat springs that make contact with a conductive coating on the inside of the glass of the tube envelope. This conductive coating acts as still another section of anode 2.

Voltage on anode 2, with reference to the cathode, is the highest applied to the tube, usually something between 4,000 and 6,000 volts. This high voltage acts on the electrons all along their path from cathode to screen, and accelerates them to very great velocity. Voltage on anode 1, the focusing electrode, usually is somewhere between 1,000 and 2,400 volts. The electron beam, and how its diameter is altered by changes of element voltages, is shown by broken lines in Fig. 58-1.

Between the first cylindrical section of anode 2 and the opening through anode 1 the electrons are moving through an electrostatic field between high and relatively low potentials. In such a field the electrons are forced to converge toward the center or toward the lengthwise axis of the tube. This converging action is so great that were the electrons to continue toward the axis they would all come to a point of focus only a little beyond the opening through anode 1.

In traveling from anode 1 to the second section of anode 2, the electrons move through a field between low and high potentials. Such a field causes the electrons to diverge from the axis, or to spread out. The divergence is enough to prevent the electrons from coming to a focus point within the elements, and to bring them to a sharp focus on the screen of the tube.

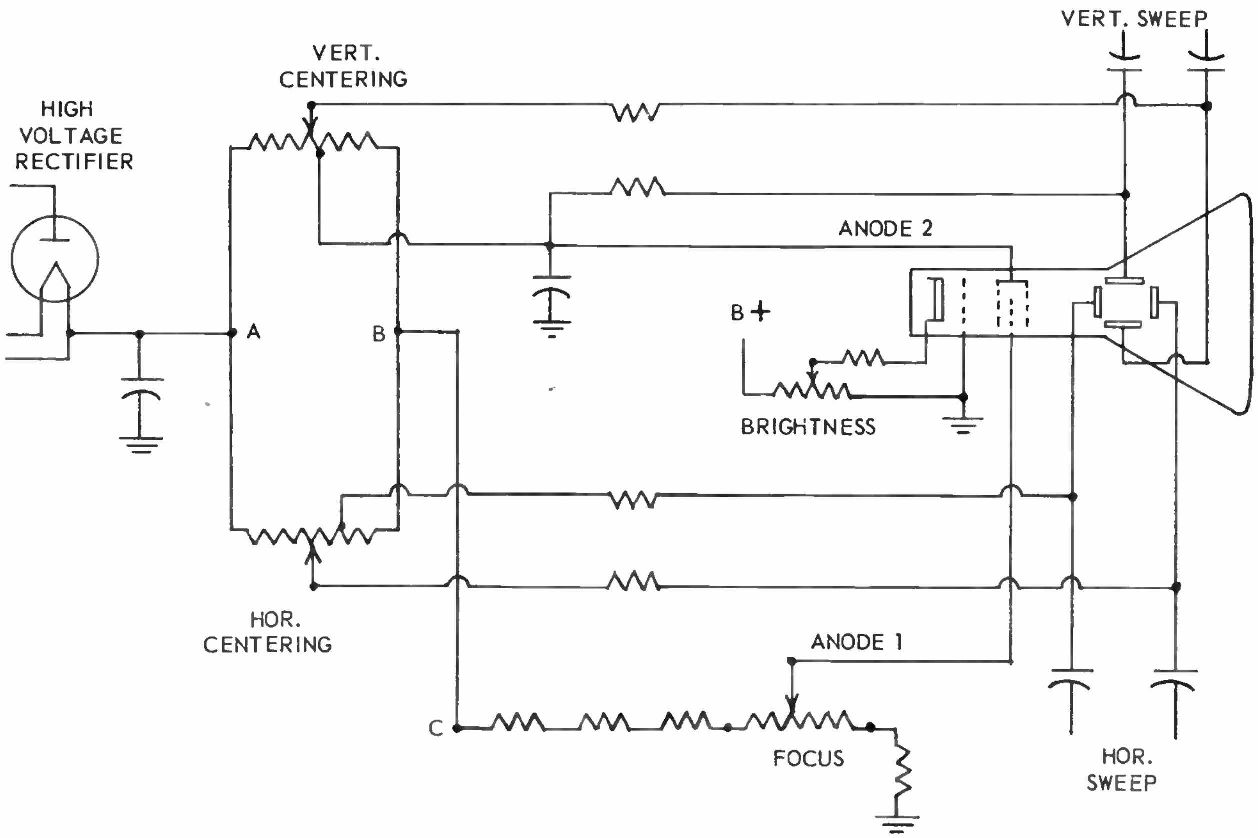


Fig. 58-2. Circuits that control focusing and centering for an electrostatic picture tube.

COMMERCIAL TRADES INSTITUTE

The beam is brought to a sharp focus on the screen by varying the ratio of voltages on anodes 1 and 2. In practice, the voltage of anode 2 remains constant while voltage on anode 1 is varied by means of the focusing control. Fig. 58-2 shows the usual method of obtaining adjustable voltage for focus control of an electrostatic picture tube. This diagram shows not only the focus adjustment, but also all other connections between the high-voltage rectifier of the power supply and elements in the picture tube. The complete diagram is shown because it is difficult to follow the drops of voltage without seeing all the connections. The centering controls which are in this power supply system will be explained later.

The highest positive voltage is at the filament of the high-voltage rectifier. The two centering control resistors are in parallel with each other, and there is a drop of voltage through them, from a to b. From a point midway along the upper (vertical) centering control resistance a connection goes to anode 2 of the picture tube. Consequently, the potential on anode 2 is very nearly as high as at the rectifier.

From the low-voltage side of the centering control resistances, at b, a line goes to point c, which is followed by a string of several resistors totaling nine or ten megohms. Then follows the focus control potentiometer whose slider connects to anode 1 in the picture tube. Voltage drop from point c to the control slider is the difference between voltages or potentials on anodes 2 and 1. All voltages are with

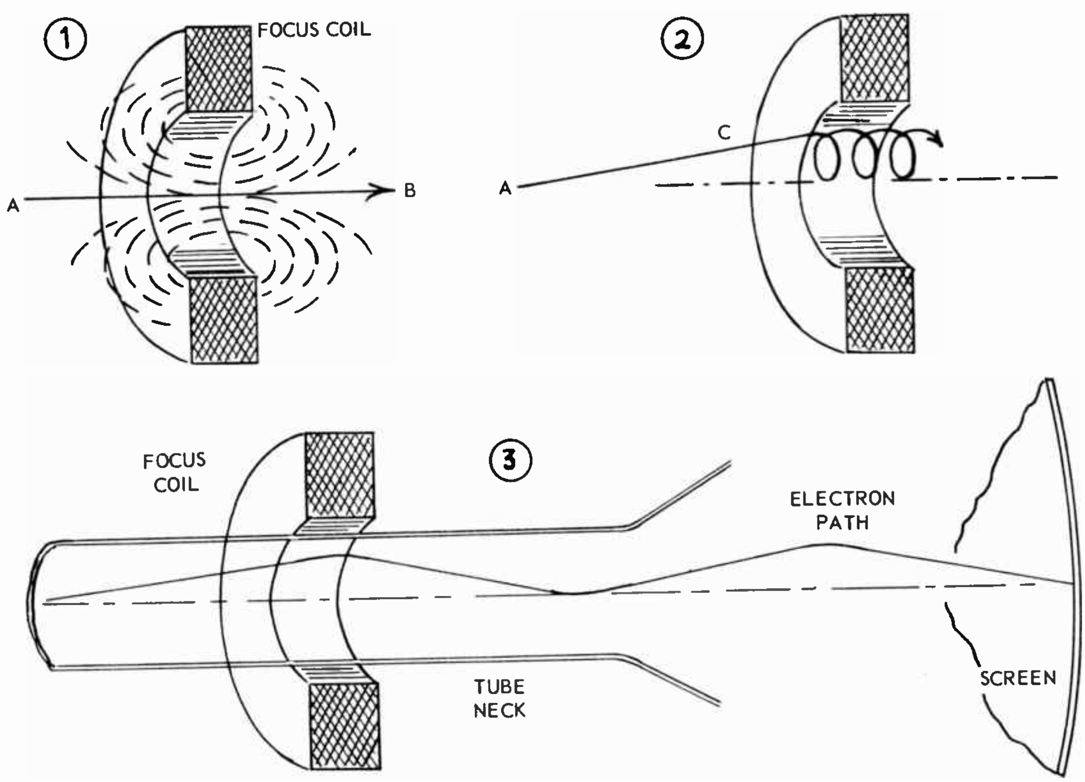


Fig. 58-3. The principle of magnetic focusing.

reference to the picture tube cathode because the cathode is connected through the brightness control to ground and because the low end of the high-voltage divider system is connected to ground beyond the focus control. Circuits essentially like those shown by Fig. 58-2 are used in nearly all receivers which have electrostatic picture tubes.

MAGNETIC FOCUSING. The principle of magnetic focusing is illustrated by Fig. 58-3. The focusing coil or magnet has the form of a short hollow cylinder with thick walls. This coil or magnet is placed around the neck of the picture tube just ahead of the end of the electron gun that is toward the tube screen and just back of the deflecting yoke. The axis through the center of the focusing coil or magnet is in line with the axis of the tube neck. The magnetic poles are at the front and rear of the focusing coil or magnet, so that lines of the magnetic field which pass through the tube neck run lengthwise of the tube axis, as in diagram 1.

If an electron traveling exactly along the tube and magnet axis were to enter the focusing magnetic field this electron would continue along the axis, as from a to b, because an electron moving parallel to the lines of a magnetic field is unaffected by the magnetic force.

But were the electron to enter the field along a path which diverges from the axis, as from a to c in diagram 2, the electron would be deflected at right angles to the field lines, just as with vertical and horizontal magnetic deflection. This right-angle deflection would force the electron to go around and around inside the magnet while continuing to travel toward the screen. The path followed by the electron would be a helix, which is the same as the path of the thread on a straight screw or bolt. The helical path would be in such position that once during every revolution the electron would pass through the axis of the focusing magnet and of the picture tube.

Because the diverging electron is traveling at very great velocity as it goes through the focusing field the turns of the helical path are not so close together as in diagram 2. The electron continues flying toward the screen at a great rate, with the result that the turns of the helical path are made quite long, as in diagram 3. But once during every turn the electron passes through the axis of the magnet and tube. This is true whether the electron goes around one turn or around several turns of its helical path between the focusing coil and the screen of the picture tube.

Every other electron that enters the magnetic field is forced to follow a helical path that goes through the magnet axis once in every turn. No matter how great or how little is the divergence of the electron path from the axis as the electron enters the magnetic field, the time required for the electron to get around one turn of its helical path remains the same. Then every electron that has arrived at any one point along its travel gets around to the axis at the same instant of time, because all electrons arriving at one point must have left the cathode at the same instant of time. This means that all electrons moving along together must go through the axis at the same time.

The length of one turn of the helical path followed by an electron gets longer as the strength of the magnetic field is decreased, for then the right-angle deflecting force is weakened while the lengthwise velocity remains unchanged. If the field strength is made exactly correct, all the electrons will come to a small spot on the axis of the tube just as they reach the screen. Thus we have focusing of the beam.

When the electrons of the beam leave the focusing magnetic field all of them are traveling in such directions as would bring them to the center of the screen. The electrons in this focused beam then are deflected vertically and horizontally as they go through the deflecting yoke. But because vertical and horizontal deflecting forces act on all the electrons alike, all still will strike the screen at one point. The position of this point is determined by the vertical and horizontal deflecting forces which are acting on the focused beam.

Focusing coils of the plain electromagnetic type, without any added permanent magnet, are mounted within a housing of iron or soft steel which helps concentrate the field lines within the tube neck while preventing the focusing magnetic field from reaching the deflecting yoke in such strength as to cause interaction. When focusing is by means of a combination permanent magnet and electromagnetic coil the ring-shaped permanent magnet is mounted around the outside of the coil.

Focus is adjusted to obtain the desired small spot or thin trace line by varying the coil current and consequent strength of the electromagnetic field. Normal currents for plain electromagnetic focusing coils

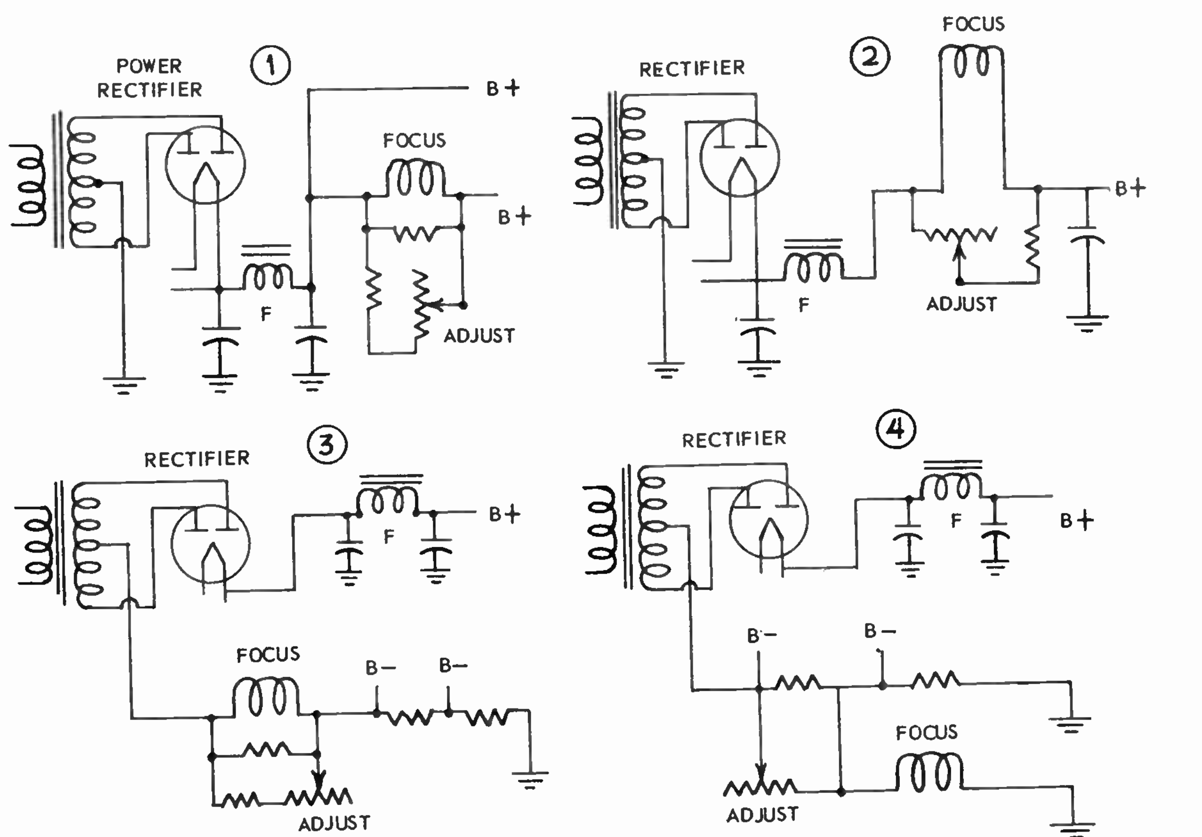


Fig. 58-4. How focusing coils are connected into low-voltage B power circuits.

range from about 75 to 150 milliamperes. The strength of the magnetic field is roughly proportional to the number of ampere-turns. The required magnetic strength varies with different types of tubes. It is not proportional to the diameter or size of the tube face. Current for the coil in a combination focusing magnet is on the order of 30 milliamperes. Additional field strength is provided by the permanent magnet.

Current for the focusing coil most often is all or part of the current which flows in the low-voltage B supply for plates and screens of various tubes in the receiver. A few of the many circuit arrangements are shown by Fig. 58-4. These diagrams show only how the focusing coil is connected into the power supply system, and must not be taken to indicate that this coil is located close to the power rectifier. The power supply filter choke and capacitors are marked F in the diagram.

Diagrams 1 and 2 of Fig. 58-4 show focusing coils connected into the positive side of the B supply. How much or how little of the total plate and screen current goes through the coil is determined by adjustable resistors shunted across the coil or connected in parallel with the coil. Increasing this parallel resistance forces more of the total current to pass through the coil. The total of current through the coil and the adjusting resistance is affected very little by change of adjustment. This is because the combined resistance of the focusing coil and paralleled resistance is only a small fraction of the plate and screen resistances of tubes which are in the B plus line. Any change of focusing resistance makes but slight difference in the total resistance and in the total current.

Diagrams 3 and 4 of Fig. 58-4 show two of the many ways in which the focusing coil and its paralleled adjusting resistance may be connected into the negative side of the low-voltage power supply. At 3 the entire power supply current goes through the coil and its resistance to ground. At 4 the focusing coil is on a separate line carrying only part of the total current.

FOCUS ADJUSTMENT. When focusing is incorrect the pattern or picture will be blurred and indistinct, as in Fig. 58-5. Correct adjustment for either magnetic or electrostatic focusing is carried out as follows.

1. Turn the channel selector to a position where no transmitted signal or program is received. Only the raster is used for focus adjustment.

2. Place the contrast control in its usual position for normal reception, not too high. Turn up the brightness control until the entire raster area is somewhat whiter than for normal reception. The bright diagonal vertical retrace lines will become visible, as shown by the photograph of a raster at the left in Fig. 58-6.

3. Adjust the focusing control so that it becomes possible to see separated narrow horizontal trace lines covering more or less of the raster area. Such lines are shown at the right in Fig. 58-6, which is enlargement of part of the raster at the left.

4. Try different settings of the focusing control. Usually the area of sharpest focus or most distinct horizontal lines can thus be shifted toward center or the edges of the raster area. The adjustment should be finally set so that there is reasonably sharp focus at the center and as far out toward the edges as possible without losing distinctness at the center.

5. Reduce the brightness control somewhat and try the effect of again moving the focusing control to make sure of having the best compromise between distinctness at the center and out toward the edges.

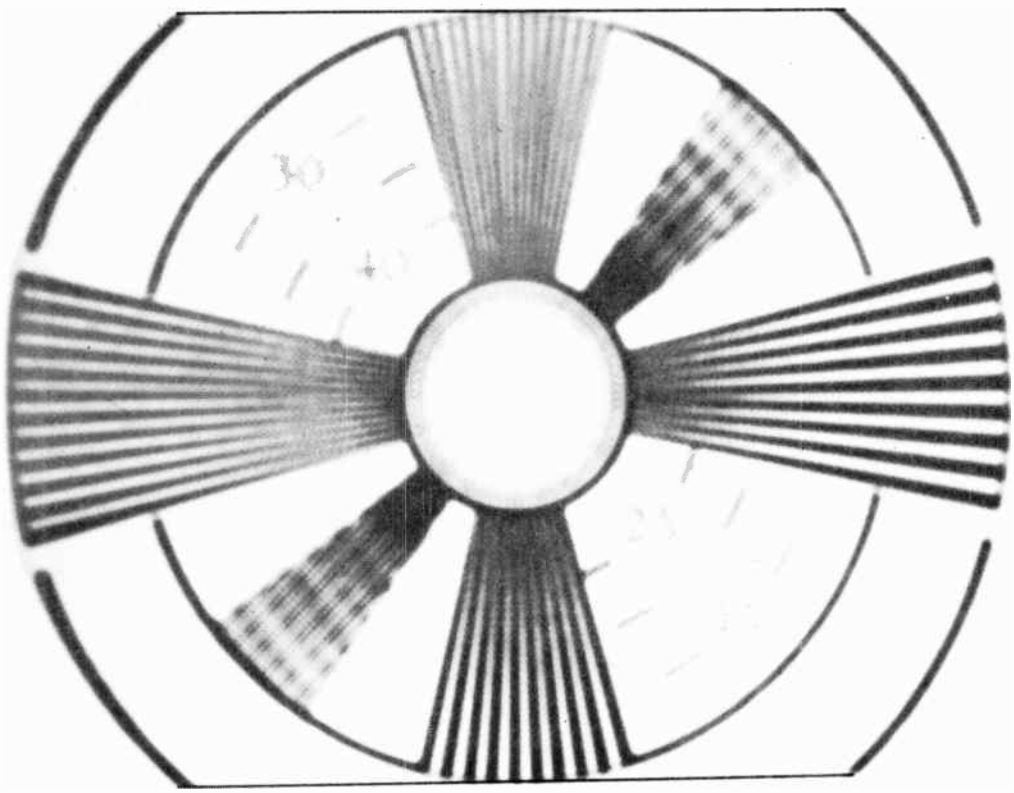


Fig. 58-5. A pattern that is out of focus.

If you encounter difficulty with a magnetic focusing system check the position of the focus coil in relation to the deflection yoke. Most often the space between the front (screen end of the focus coil)and the rear side of the yoke should be about 3/8 inch or possibly a little less. If possible, move the focus coil forward and back, while making readjustments of the control, to get the best compromise of sharp lines at the center and edges. With too much separation between focus and deflecting coils it will be impossible to obtain distinct horizontal lines. If changing the focusing control tends to rotate a pattern or picture the coils probably are too close together.

In moving or attempting to move the focus coil use extreme care not to strike the neck of the picture tube with anything hard, and under no circumstances place any pressure on the neck. It is safest to wear heavy goggles. Breakage of the tube will release several tons of atmospheric pressure which is acting on the outside of the glass, and the resulting implosion and flying of broken glass can cause serious injury.

CENTERING. Centering means to bring the center of the reproduced picture or pattern to the center of the mask opening in front of the picture tube. When centering is to be secured by changing the centering adjustments it is essential first to see that horizontal and vertical hold controls are adjusted about mid-

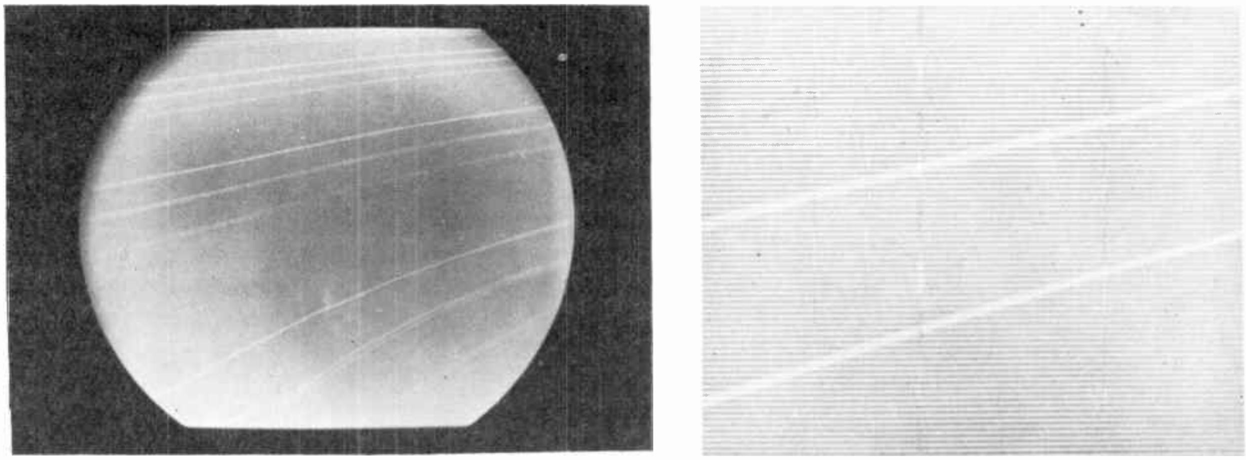


Fig. 58-6. The raster used for focusing (left) and how the horizontal lines appear when focusing is sharp (right).

way between their positions at which the picture or pattern drops out of synchronism. Otherwise the movement of the picture or pattern with correct adjustment of the hold controls may throw it too far off center.

The picture or pattern may be too low in the mask, or it may be too high as shown by diagram 1 in Fig. 58-7. The position may be too far to the right, or to the left as in diagram 2. The picture or pattern may be off center both horizontally and vertically at the same time, as in diagrams 3 and 4.

ELECTROSTATIC CENTERING. In Fig. 58-2 were shown centering controls together with other parts of the high-voltage circuits for an electrostatic picture tube. The parts of that diagram which enter into the control of centering are shown by themselves in Fig. 58-8.

The deflecting plates numbers 1 and 2 in the picture tube are used for vertical deflection. To them are applied the opposite-phase sawtooth voltages from the vertical sweep amplifiers or the vertical sweep oscillator. Deflecting plates numbers 3 and 4 are used for horizontal deflection, with these two plates receiving the sawtooth voltages from the horizontal sweep amplifiers or oscillator. The four deflecting plates are farthest from the electron gun and closest to the screen of all the picture tube elements. Consequently, to maintain high velocity of electrons in the beam, these deflecting plates must be at high voltage. They are connected through the centering controls to the positive output of the high-voltage rectifier.

Vertical deflecting plate number 2 is connected to a center tap on the vertical centering control potentiometer. Potentials on the left of the tap are relatively positive, and those on the right are relatively negative with respect to the tap and to deflecting plate 2. The other vertical deflecting plate, number 1, is connected to the slider of the vertical centering control. If the slider is moved to the left of the tap, plate number 1 is made more positive than plate number 2, and the electron beam is drawn over toward number 1. This shifting of the beam merely establishes a new average position. The beam still is deflected up and down from this average position by the sawtooth voltages. If the slider is moved to the right, plate number 1 is made negative with reference to plate 2, and the average position of the beam is

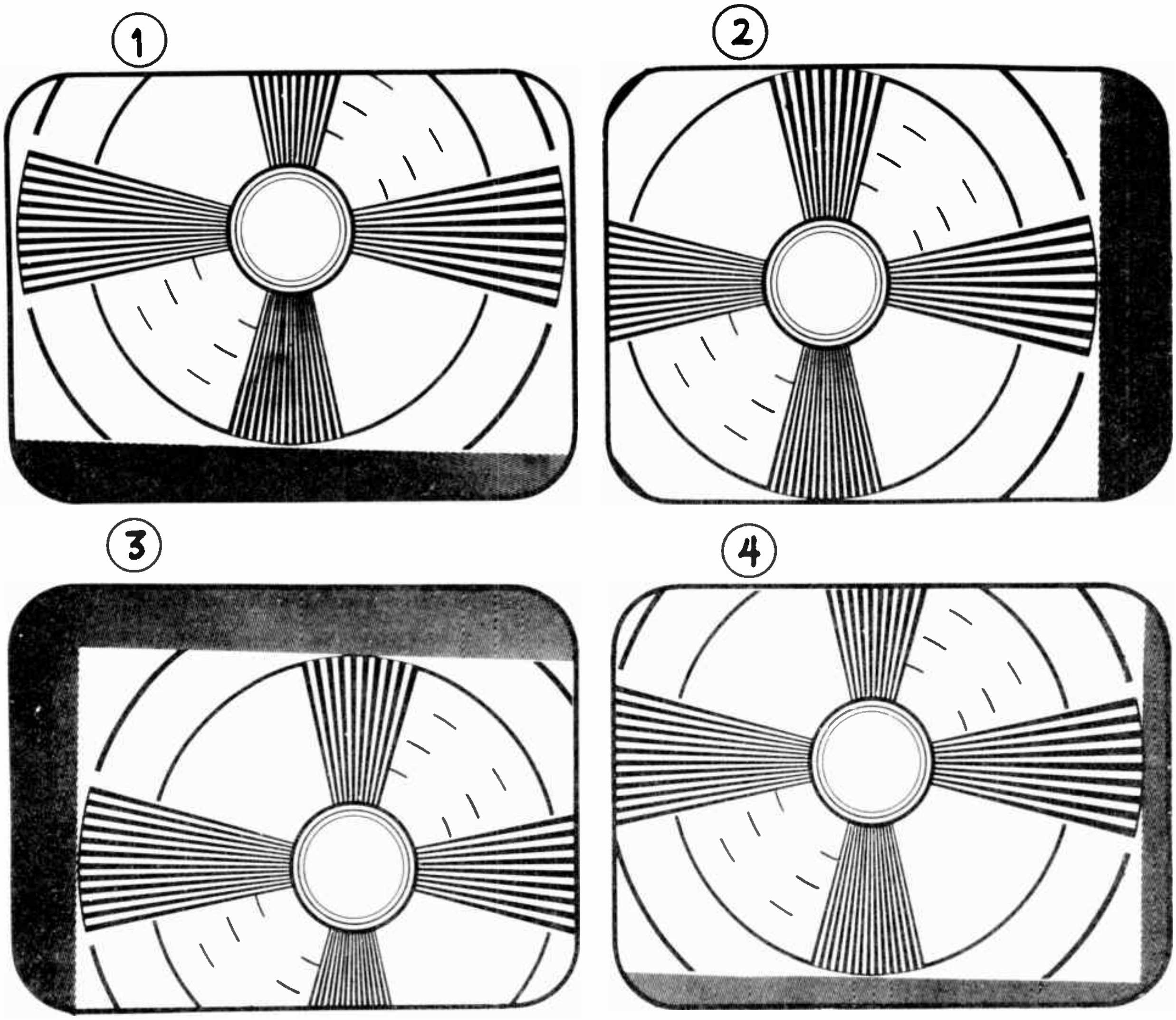


Fig. 58-7. Incorrect centering of patterns.

shifted toward plate 2 or away from plate 1. Since deflections are equal above and below the average position of the beam the entire picture is moved up or down on the screen by moving the slider of the centering control.

The effect of the horizontal centering control on horizontal deflecting plates numbers 3 and 4 is similar to that described in the preceding paragraph, except that the horizontal control moves the average position of the electron beam to the left or right, closer to plate 3 or else closer to plate 4. Thus the entire picture may be shifted sideways on the screen of the picture tube.

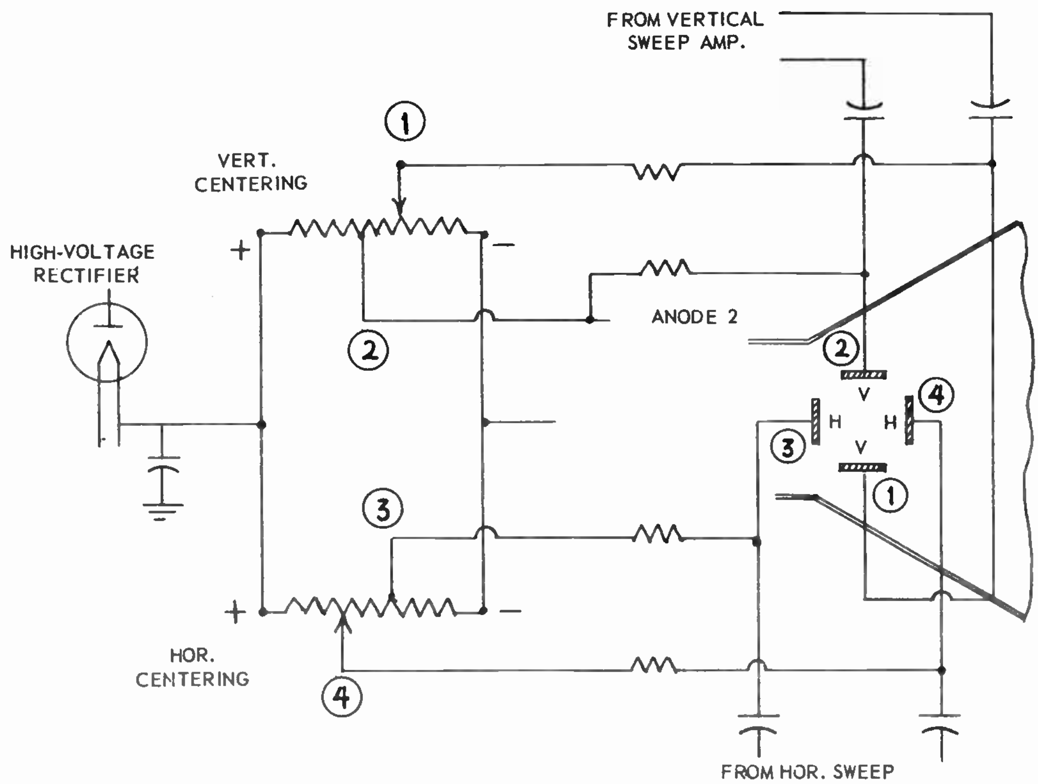


Fig. 58-8. Connections for centering controls on an electrostatic picture tube.

The tap connection on the horizontal centering control is closer to one end than to the other. As a result there is a greater voltage drop from the positive end to the tap than from the tap to the negative end. The two voltage drops then match the drops on opposite sides of the tap in the vertical centering control. In that control there is more voltage drop on the positive side because this side is carrying the additional current for anode number 2 in the picture tube. It is desirable to have equal voltage drops in corresponding sides of both controls so that both will have about equal effects on picture position as the sliders are adjusted.

MAGNETIC CENTERING. Centering with picture tubes having magnetic focusing most often is accomplished by tilting the focus coil or magnet up, down, or sideways as may be required. The action is much as though the focus coil were aiming the electron beam or the picture at the screen of the picture tube, and as though tilting the axis of the coil were causing the picture to fall at different positions on the screen.

The focus coil is mounted or supported in some manner which allows tilting of its axis with reference to the axis of the picture tube. There are many variations in mechanical details of the mountings. Unless you already are familiar with the construction of a receiver being worked on it is necessary to make a preliminary examination of the screws, nuts, and levers, paying especial attention to screws or studs which pass through elongated slots instead of through round holes only large enough to take the screw or stud. Such slots are intended to permit movement of the coil or the coil support one way and another along the length of the slot.

In Fig. 58-9 the deflecting yoke is carried in a bracket extending upward from the chassis, with the forward end of the yoke against the flare or cone of the picture tube. The focus coil is carried by a rearward extension of the yoke bracket. On opposite sides of the shell of the focus coil are attached threaded studs which extend outward through lengthwise slots in the bracket. Wing nuts and washers hold the coil housing in position.

With the wing nuts loosened on the studs the top and bottom of the focus coil may be moved forward or back to tilt the axis up or down, with the studs acting as pivots. It is possible also to move either or both sides of the coil housing forward or back, to shift the coil axis toward the right or left. The entire coil may be moved forward or back to place it the correct distance from the yoke.

With the construction of Fig. 58-10 the yoke and focus coil again are carried by a bracket mounted on

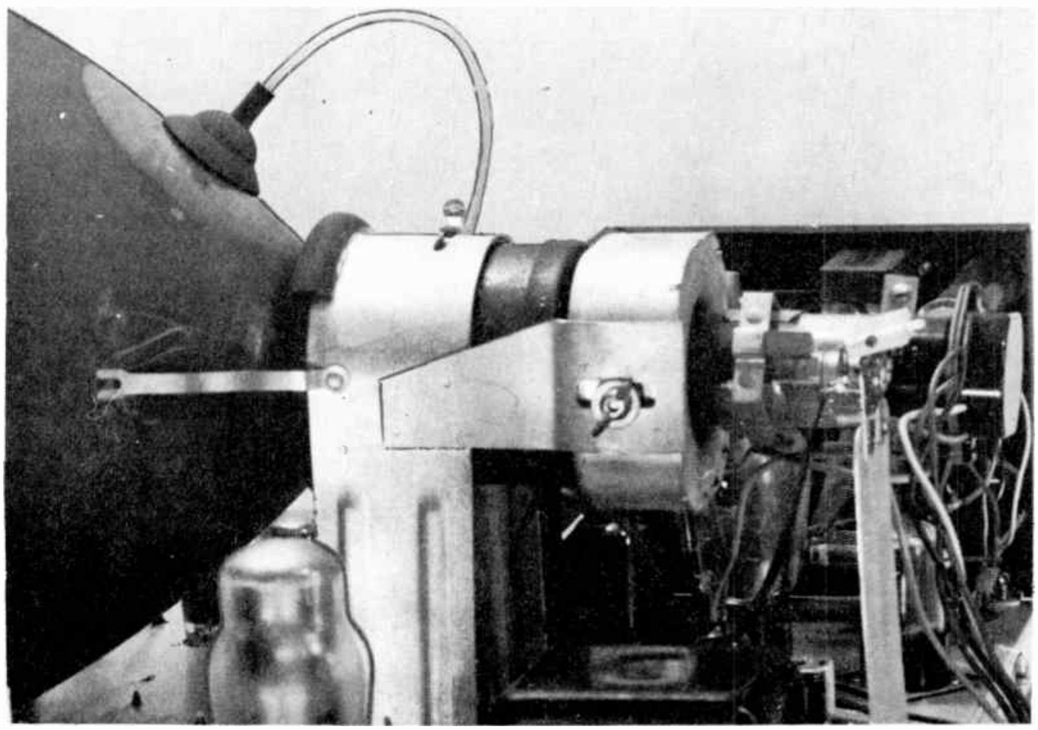


Fig. 58-9. The deflection yoke is immediately back of the picture tube flare. The focus coil is carried by the bracket that supports the yoke.

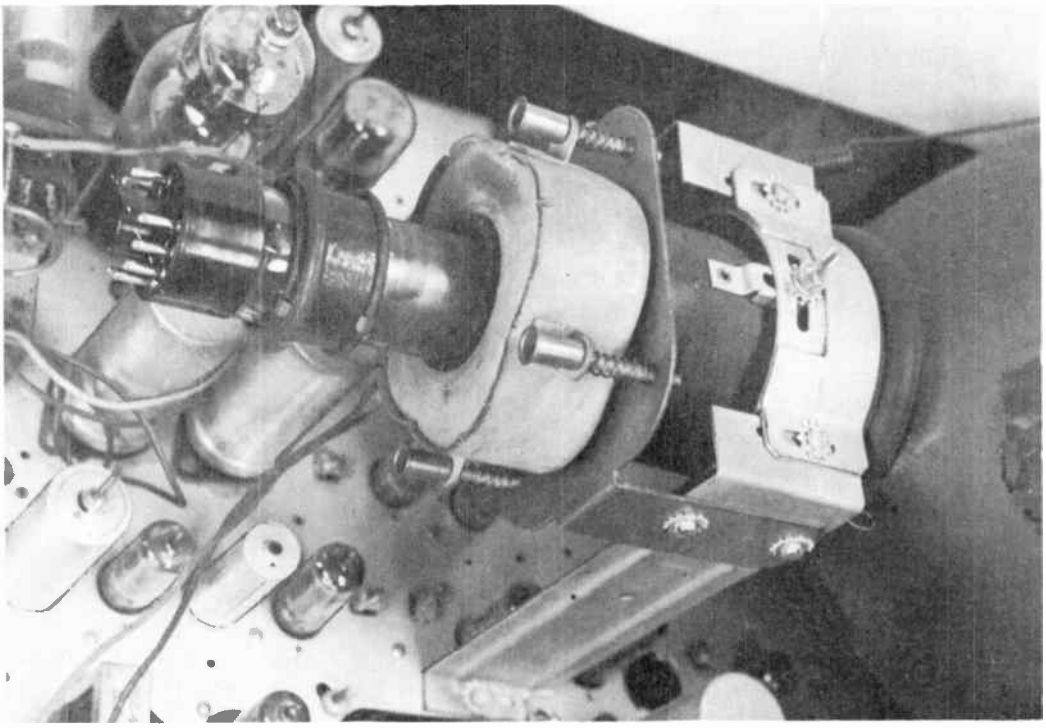


Fig. 58-10. A focus coil supported by adjusting screws and springs that allow centering of the picture.

the chassis. On the housing of the focus coil are ears or lugs through which pass long screws that thread into the stationary part of the supporting bracket. Between the lugs and the bracket are small coiled springs placed around the screws. These springs push the focus coil away from the stationary bracket to a distance determined by how far the screws are turned into their threaded holes in the bracket.

There are only the three adjusting screws which you can see in the picture. Turning the middle screw, the one closest to you in the picture, will move its side of the focus coil forward or back and will tilt the beam accordingly. When you tighten one of the two screws that are opposite each other, and loosen the second one of this pair, the beam is tilted in a direction at right angles to the direction of tilt secured with the single screw on one side of the housing. Turning all three screws equally forward or backward will move the focus coil closer to or farther from the deflecting yoke.

Other mountings are of the gimbal type, such as used on marine compasses to keep them level no matter how a ship may tilt. The focus coil is held within two rings. There are pivots on opposite sides of both rings, and the pivots in one ring are at right angles to those in the other ring. The focus coil may be tilted into any desired position, where it is held by tightening wing nuts on the pivot studs. Sometimes the focus coil is in supports that allow movement, but which have considerable frictional resistance to such movement. Then the coil may be moved by a long lever or handle attached to its housing.

The picture on the screen will not move in the same direction that the axis of the focus coil is tilted. When the coil is moved to tilt its axis straight upward or downward with reference to the axis through the neck of the tube the picture will shift diagonally at the same time it shifts up or down. Tilting the coil axis to the left or right will move the picture diagonally as well as sideways. By noting the direction of picture movement as the coil is tilted one way and the other it is easily possible to bring the picture to any desired position on the screen.

Although the focus coil axis is intentionally tilted to center the picture, the center of the opening through the coil must be and must remain on the axis of the picture tube. If the picture tube is temporarily removed, and you look lengthwise through the openings of the focus coil and the deflecting yoke these openings should be in line. The center of the focus coil must not be higher or lower nor to one side or the other from the center of the tube. If the focus coil is not so positioned with reference to the axis of the picture tube it may be impossible to secure correct centering without having black shadows at one or two corners of the picture. These shadows are caused by the electron beam striking the neck of the tube where it joins the flare or cone, rather than having free passage to all parts of the screen. In rare cases it may be necessary to slightly rotate the picture tube around its own axis or the axis of the neck in order to get rid of corner shadows.

5 In order that there may be freedom from corner shadows, also to allow good focusing, it is essential that the deflection yoke be as far forward as it can be moved, or right up against the flare or cone of the picture tube. Forward and back movement of the yoke usually is allowed by lengthwise slots through which pass screws or studs that hold the yoke on its bracket. These adjustments are visible on the yoke mounting in Fig. 58-10. In some receivers the large bracket that carries the yoke and focus coil has lengthwise slots for the screws holding this bracket on the chassis or other support. Then the bracket and all coils may be moved together while the screws are loosened.

Centering in some receivers having magnetic deflection picture tubes is accomplished by a magnetized steel ring supported between the back of the deflecting yoke and the front of the focus coil, surrounding the neck of the picture tube. This centering ring is mounted so that it may be raised or lowered, or moved sideways. Moving the ring up or down shifts the picture horizontally, to one side or the other, while moving the ring sideways shifts the picture vertically, up or down. Rotating the magnetized ring has very little effect on picture position or centering.

Another magnetic centering device makes use of a magnetized steel ring having a small air gap at one point. This ring is supported on the rear end of the deflecting yoke. The ring is carried by a mounting that allows rotation through a large part of a turn, and that allows limited movement vertically or horizontally. If the gap in the centering ring is on one side or the other of the tube neck, rotation of the ring will shift the picture sideways. If the ring gap is above or below the tube neck, rotation will shift the picture vertically. When the ring is moved bodily up or down, or from side to side, it has a limited effect on picture position.

Fig. 58-11 illustrates a method of centering which makes use of controlled direct currents in the deflecting coils. The horizontal deflecting coils are connected through the secondary of the sweep output transformer to a B plus point in the voltage divider system that is connected to the low-voltage rectifier furnishing plate and screen currents for various tubes in the receiver. The other end of the horizontal deflecting coils is connected to the slider of a centering control potentiometer in the voltage divider system.

More or less direct current is caused to flow in the deflection coils by the difference of potential across the two ends of the coil circuit. This steady direct current tends to give the electron beam a constant deflection to one side or the other. With the control slider at the center of its range the focus coil or magnet is adjusted for correct centering. Thereafter, should the picture move off the horizontal center position, adjustment of the control potentiometer will vary the steady direct current and shift the picture in a direction to center it. The steady direct current for centering is in addition to the sawtooth current for deflection.

Vertical centering is similarly adjusted. The vertical centering potentiometer has a tap that is connected to one end of the deflecting coil circuit, while the other end of the coil circuit is connected to the control slider. Movement of the slider to one side or the other of the tap varies the amount and the polarity of steady direct current sent through the deflecting coils. Thus the picture may be shifted either up or down on the screen. The centering current is, of course, in addition to the sawtooth deflecting current in the coils.

SKEW OF PICTURES. If the entire picture or pattern is twisted obliquely, so that the sides are not vertical and so that the top and bottom are not horizontal, the fault may be described as skewing. There

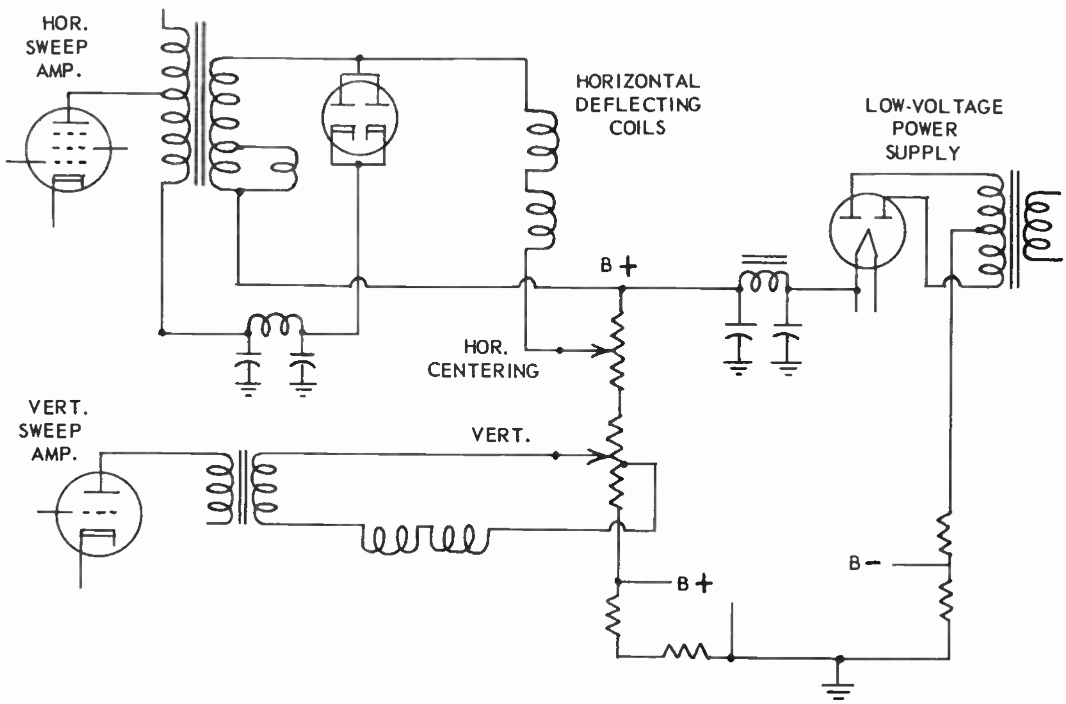


Fig. 58-11. Connections for centering by means of variable direct current in the deflecting coils.

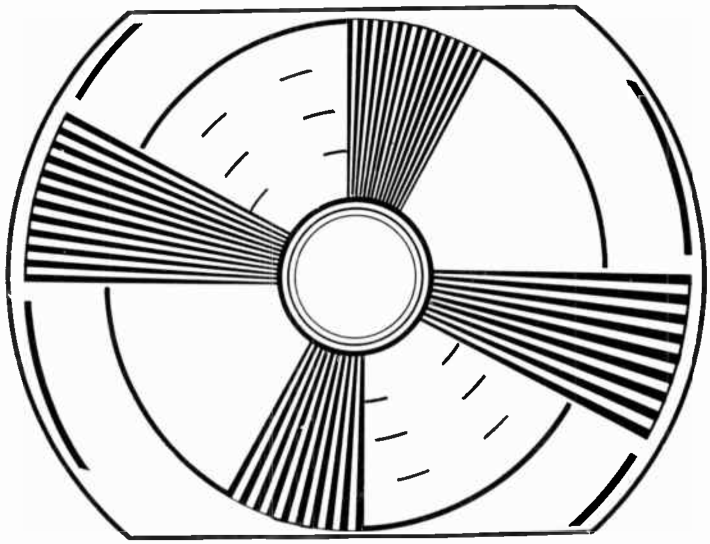


Fig. 58-12. Skewing of a pattern.

is skewing in Fig. 58-12, as is evident from the fact that lines which should be vertical are oblique and those which should be horizontal are oblique.

Skewing is corrected with an electrostatic picture tube by loosening the mountings and rotating the tube bodily to square the picture or pattern with the opening in the mask which is on the cabinet in front of the screen.

With a magnetic deflection tube the skewing is corrected by loosening the nut or nuts which hold the yoke in its position in the supporting bracket, then rotating the yoke to make the picture or pattern square with the mask. Heavy goggles and gloves must be worn when handling the picture tubes, as personal protection in case of breakage. An electrostatic deflection tube may be rotated while a picture or pattern is being reproduced. A magnetic deflection tube should be rotated only while the receiver is turned off. Otherwise there is too much danger of some part of your body coming in contact with one of the high-voltage leads in the anode circuit. Shocks from this circuit are quite likely to have serious consequences.

SIZE CONTROLS. Vertical size or height of the picture is increased by anything which increases the amplitude of the sawtooth voltages on vertical deflecting plates of an electrostatic picture tube, or which increases the amplitude of sawtooth current in vertical deflecting coils for a magnetic deflection tube. Horizontal size or width is similarly increased by anything which increases amplitudes of sawtooth voltages on horizontal deflecting plates or sawtooth current in horizontal deflecting coils. This you would expect, because the distance the electron beam is deflected in any direction is proportional to maximum amplitude of deflecting voltages or currents. There are many ways in which these amplitudes may be varied to increase or decrease the picture size.

At the left in Fig. 58-13 is shown a method of size control used with either electrostatic or magnetic

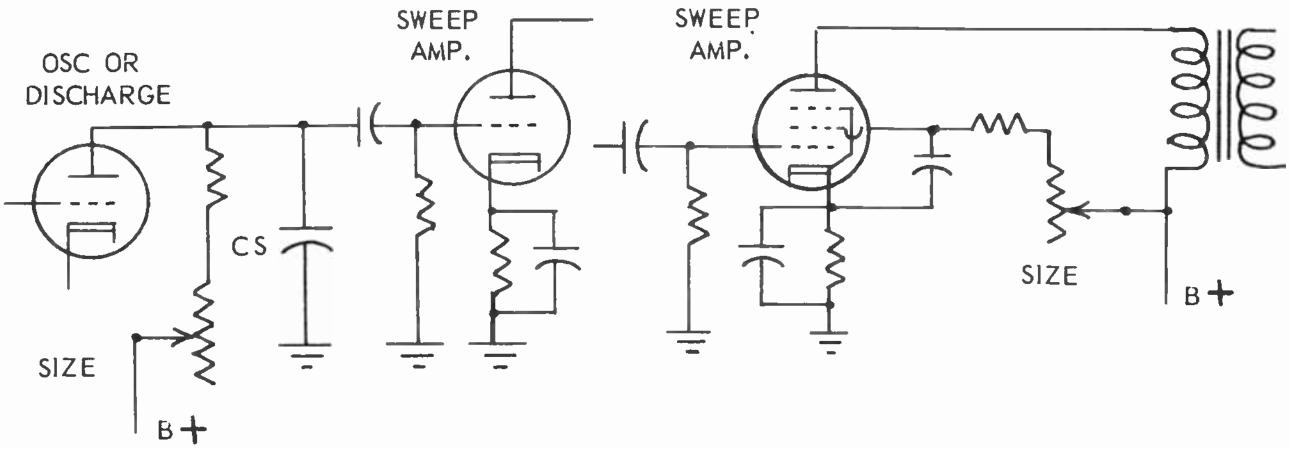


Fig. 58-13. A size control that varies maximum charge of the sawtooth capacitor (left) and one that varies screen voltage on the sweep amplifier (right).

deflection systems and with either vertical or horizontal deflection. Sawtooth capacitor Cs charges through the adjustable size control resistance during the time in which the oscillator or discharge tube is non-conductive. The quantity of charge and resulting maximum voltage or amplitude of the sawtooth wave is increased with less resistance and decreased with more resistance. The effect of the changed sawtooth amplitude is carried on through to the deflecting plates or coils.

At the right in Fig. 58-13 the size is controlled by varying the voltage on the screen of the sweep amplifier. Greater screen voltage increases the amplification less screen voltage decreases it. The amplitude of sawtooth voltage and current in the amplifier output thus is varied. In some receivers the size is varied by an adjustment for plate voltage on the sweep amplifier. In a few cases the amplification is varied by an adjustable cathode-bias on the sweep amplifier.

Fig. 58-14 shows circuits which include a type of horizontal size control widely used in receivers having a flyback type high-voltage supply system for the picture tube anode. This width control consists of an adjustable inductor connected in parallel with or shunted across part of the turns in the secondary winding of the horizontal output transformer.

The operating principle of this width control may be explained as follows. First, the greater the total inductance in the transformer secondary winding the greater will be the induced emf and the greater will be induced current in this winding and in the deflecting coils. There is greater width of picture with more inductance, less width with less inductance.

When part of the secondary winding is shunted with another inductance, the width control, the combined inductance of the two is less than the inductance of either part alone. Therefore, any shunted inductance is going to reduce the total inductance in comparison with inductance without any shunting, and picture width is reduced.

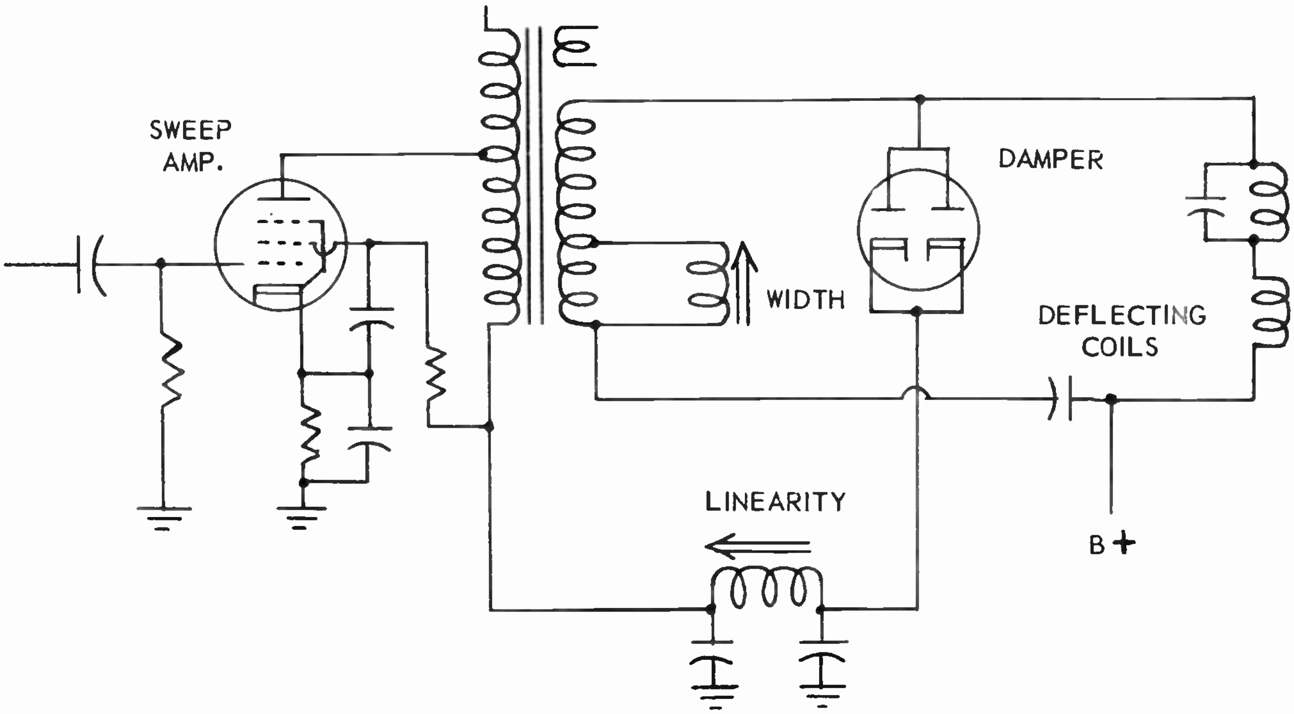


Fig. 58-14. A width control that varies the effective inductance of the output transformer secondary.

If the inductance of the shunt (width control) is increased, there is an increase of combined or parallel inductance in that portion of the secondary circuit consisting of the shunt and the shunted turns of the main winding. Then there is an increase of total secondary inductance as compared with the inductance before the shunt is increased. Thus an increase of width control inductance increases the picture width. Conversely, a decrease of width control inductance decreases the picture width.

Since secondary inductance is decreased by any value of shunted inductance, disconnecting the width control inductor from the secondary winding will allow maximum inductance and will give the widest possible picture. This is the action that you were going to figure out for yourself when this method of width control was first shown in a preceding lesson.

There are numerous variations or modifications of the width control that employs an adjustable inductor on the horizontal output transformer. Some of them are shown by Fig. 58-15. In diagram 1 the width control is connected across the damper end of the transformer secondary. The deflecting coils are not connected across the part of the secondary which carries the width control inductor. In diagram 2 the width control is tapped, with one side of the deflecting coils connected to the tap. In diagram 3 the width control inductor is connected across a separate winding on the core of the output transformer. With any of these circuits an increase of width control inductance increases the deflecting coil current and the picture size.

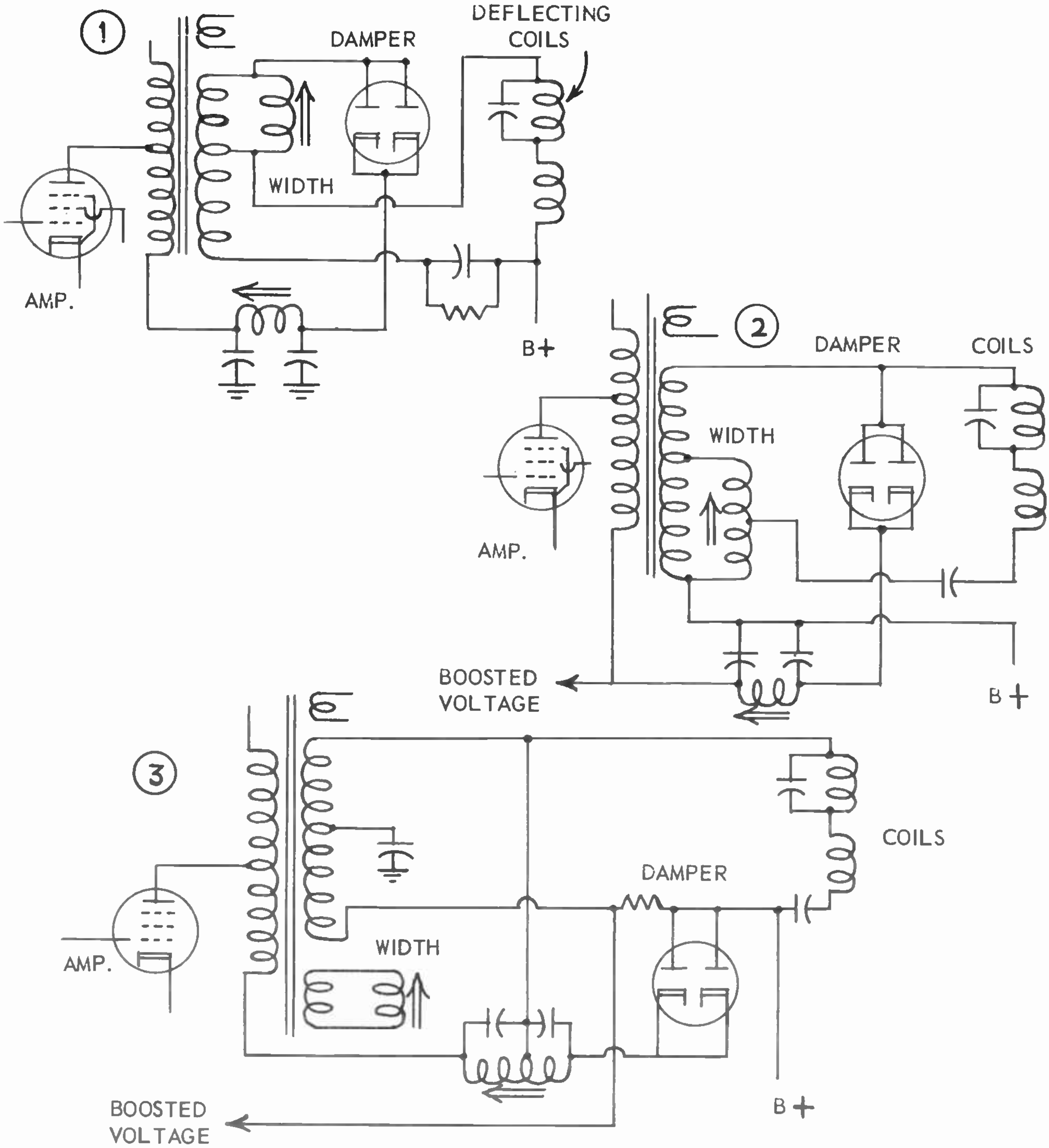


Fig. 58-15. Various ways of connecting width control inductors to the sweep output transformer.

The vertical size control or height control should be adjusted so that the top and bottom of the reproduced picture or pattern just disappear behind the top and bottom of the mask that frames the face of the picture tube. The horizontal size control or width control is to be adjusted so that the opposite sides of the picture or pattern just disappear back of the sides of the mask. In other words, the picture or pattern is to be made just a little greater in both dimensions than the opening of the mask. Of course, the picture must be correctly centered, or the centering must be adjusted, while making settings of the size controls.

Picture size is varied not only by settings of the size controls, but also by changes of other controls. Readjustment of linearity controls nearly always will change the picture size. Variation of drive controls will change the size. Anything that alters the high voltage applied to the anode of a magnetic picture tube or to the deflecting plates and high-voltage anode of an electrostatic tube will change the picture size in both dimensions at the same time.

The higher the anode voltage or electron accelerating voltage the smaller the picture will become when no other changes are made. This is because the accelerating voltages try to keep the electrons traveling along a straight path from electron gun to screen in the picture tube. The greater the accelerating voltage the harder the deflecting sawtooth voltages or currents have to work to bend the beam in any direction, and the less is the resulting deflection both vertically and horizontally. Anything which reduces the accelerating voltage on anodes or plates allows the deflecting voltages or currents to have proportionately greater effect on the electron beam, and the picture will become both higher and wider.

DO NOT TEAR — CUT ALONG THIS LINE AND SEND IN FOR GRADING.

TELEVISION

LESSON NO. 59

LINEARITY, DRIVE AND DAMPING

Training Your Mind!!

C.T.I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

COMMERCIAL TRADES INSTITUTE

Training Your Mind ... Perfecting Your Skill — C.T.I.

Chicago, Ill.

LESSON NO. 59

LINEARITY, DRIVE AND DAMPING

When a television picture has irregular twists and curves where there should be straight lines, and has curves that turn in the wrong directions, there is highly unsatisfactory reproduction. There are many possible causes for such pictures. One cause is a misshapen sawtooth current or voltage wave at the deflecting coils or plates. Obviously, if current or voltage in the trace portions of the sawtooth does not change at a constant rate, the electron beam will be forced to travel horizontally or vertically at speeds which vary at different points on the screen. Then parts of the picture which should be in certain positions will be advanced or retarded, and we have distortion.

There are controls whose special purpose is to insure that horizontal and vertical deflections are at constant rates, so that the sawtooth waveforms shall consist of lines that are practically straight, or are linear. These are the vertical and horizontal linearity controls. In this lesson we shall learn about faults that are compensated for by these controls, and about the action of the controls and how they should be adjusted during service work. But first we are going to learn about the action of the damper tube in preventing a form of distortion that would distort the left-hand side of pictures on magnetic deflection tubes were the damper not provided.

You know that the retrace portion of the sawtooth wave must be completed within a few microseconds. This fast retrace is obtained by utilizing the high resonant frequency of distributed and stray capacitances in the deflecting coil circuit, together with its inductances. The retrace is formed by the first part of a cycle of oscillation at the self-resonant frequency of these capacitances and inductances. This frequency is many times greater than the horizontal deflection frequency.

Oscillation is started in this manner. When plate current in the horizontal sweep amplifier is suddenly cut off by negative grid voltage, there is sudden collapse of the magnetic fields in the output transformer and deflection coils. The very fast change of magnetic field strength induces a pulse of high voltage in the transformer secondary winding. This pulse shocks the self-resonant circuit into oscillation. Were nothing done to stop it, the oscillation would continue until enough energy was absorbed in circuit resistance to leave a steady current. There would be several oscillating cycles, about as shown by Fig. 59-1, and the picture would be distorted.

Damping action is illustrated by Fig. 59-2. Plate current in the sweep amplifier, and induced current in the output transformer secondary and deflecting coils, increase to the peak marked _a_ in diagram _1._ Then

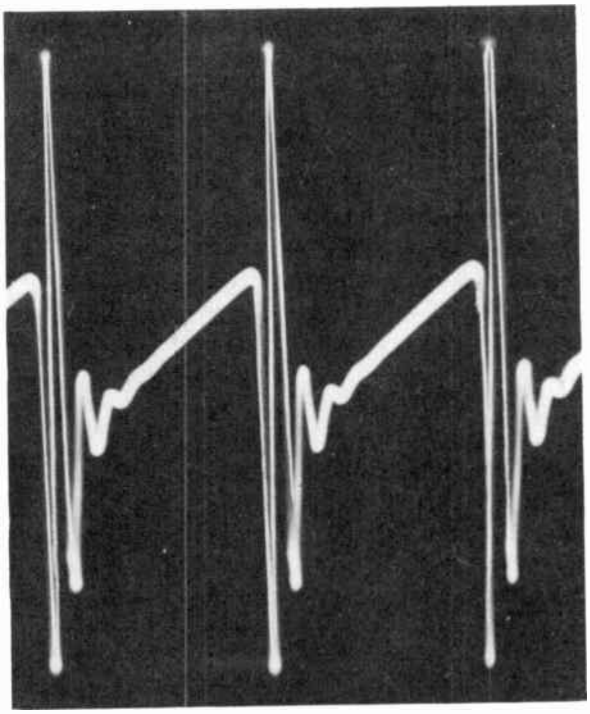


Fig. 59-1. Oscillation at the beginning of a sawtooth wave.

there is plate current cutoff, which causes the currents to decrease to point b. The currents then go in opposite polarity to point c during the first part of the first oscillation cycle. The change of current from a to c is the retrace portion of the sawtooth wave.

Since the oscillatory circuit consists chiefly of the deflecting coils and the transformer secondary it is essentially an inductive circuit. In such a circuit the changes of voltage lead the changes of current by approximately 90 degrees, which is simply another way of saying that current lags the voltage in an inductive circuit.

The plate of the damper tube is connected to the oscillatory circuit or deflecting coil circuit. When current reaches point a in the diagram, the leading voltage has reached a value corresponding to point b. That is, voltage on the damper plate then is zero and is going negative. When current falls to point b the leading voltage has reached a value corresponding to point c. The voltage on the damper plate then is of maximum negative value. The negative voltage keeps the damper non-conductive while current is changing from a to c, during the entire retrace. The non-conductive damper has no effect on the retrace.

But as current commences to change from c toward x, the leading voltage would be changing from x toward y. That is, the voltage commences to go positive on the damper plate. This makes the damper conductive, and it acts like a low-resistance load on the deflecting coil circuit. So much energy is taken from

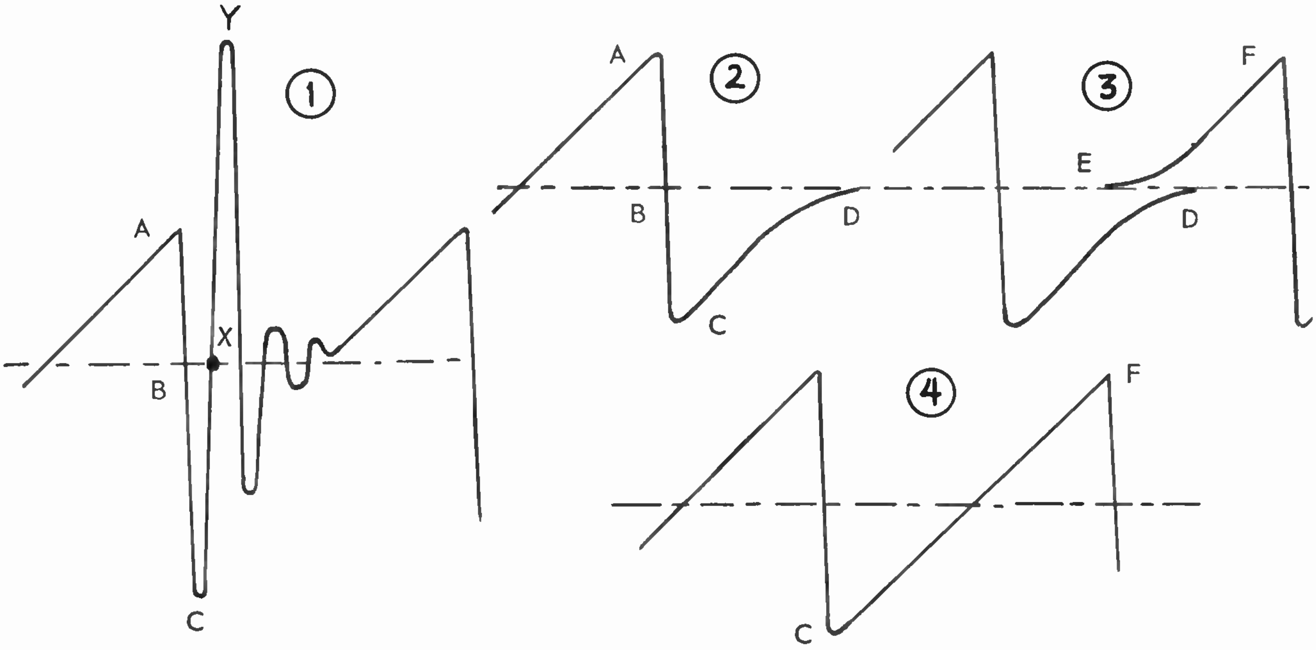


Fig. 59-2. How the oscillating current is damped.

the coil circuit that not enough remains to support continued oscillation. Consequently, the oscillatory current never does get all the way from c to x, but it dies out along the line from c to d in diagram 2. Remember, we are talking now about current that is due to oscillation. The deflecting coil current that is due to induction, or due to changes of sweep amplifier plate current and its cutoff, stops at point b.

Before the oscillatory current dies out at point d, grid voltage on the sweep amplifier has returned from its strongly negative value to the value to which plate current resumes. Plate current commences to flow, and a new induced current appears in the transformer secondary and deflecting coils. This new induced current starts from e in diagram 3 and increases to f, where again there is cutoff. Point f represents the same instant of time in the sawtooth cycle that is represented by point a. While oscillatory current is dying out as it approaches point d, the new induced current is increasing from point e.

Of course, there are not two separate currents at the same time in the deflecting coil circuit. The dying oscillatory current combines with the increasing induced current and, as in diagram 4, there is a continuous and smooth change of deflecting current from c to f. Here we have the fast retraces due to plate current cutoff and the start of oscillation, and have also the smooth intervening changes which are due to damping of the oscillation and resumption of plate current.

The less the effective damping resistance the faster the oscillatory current will die out, and the greater this resistance the longer it will take. Part of this resistance is the internal plate-cathode resistance of the damper tube. Opposition to flow of damper current is varied also by all resistors, inductors, and capacitors that are connected into the plate or cathode circuits of the damper. All such circuit elements affect linearity, because they help determine the rate at which oscillatory current dies out and they help determine whether the oscillatory and induced currents combine to produce a smooth change at and near the center of the trace portion of the sawtooth wave.

Most damper tubes are types originally designed as power rectifiers. Quite commonly used are the 6W4-GT half-wave rectifier and the 5V4-G full- wave rectifier. Two-element rectifier type dampers, which may be either of these types, have been shown in the various diagrams of deflection circuits.

The damper tube sometimes is a twin triode (6AS7-G) especially designed for this purpose. Fig. 59-3 shows one method of using this twin triode. Plates, cathodes, and grids are connected in parallel with each other. Small resistors in series with the grids help prevent sustained oscillation. Voltage applied to the grids is secured from the plate-cathode circuit. The time constant of the grid circuit may be varied by the adjustable linearity control resistor, thus altering the time of damper conduction and the internal resistance of the tube. In some receivers the cathode resistor Rk also is adjustable to provide an additional linearity control. The triode damper is not widely used at the present time.

A Damper tubes such as have been described are used in horizontal sweep circuits. In vertical sweep circuits sufficient damping to prevent self-resonant oscillation in the deflecting coils is provided by connecting resistors of a few hundred ohms across each of these coils. The resistor currents absorb enough energy to prevent oscillation. Damping resistors have been shown in some of the earlier diagrams.

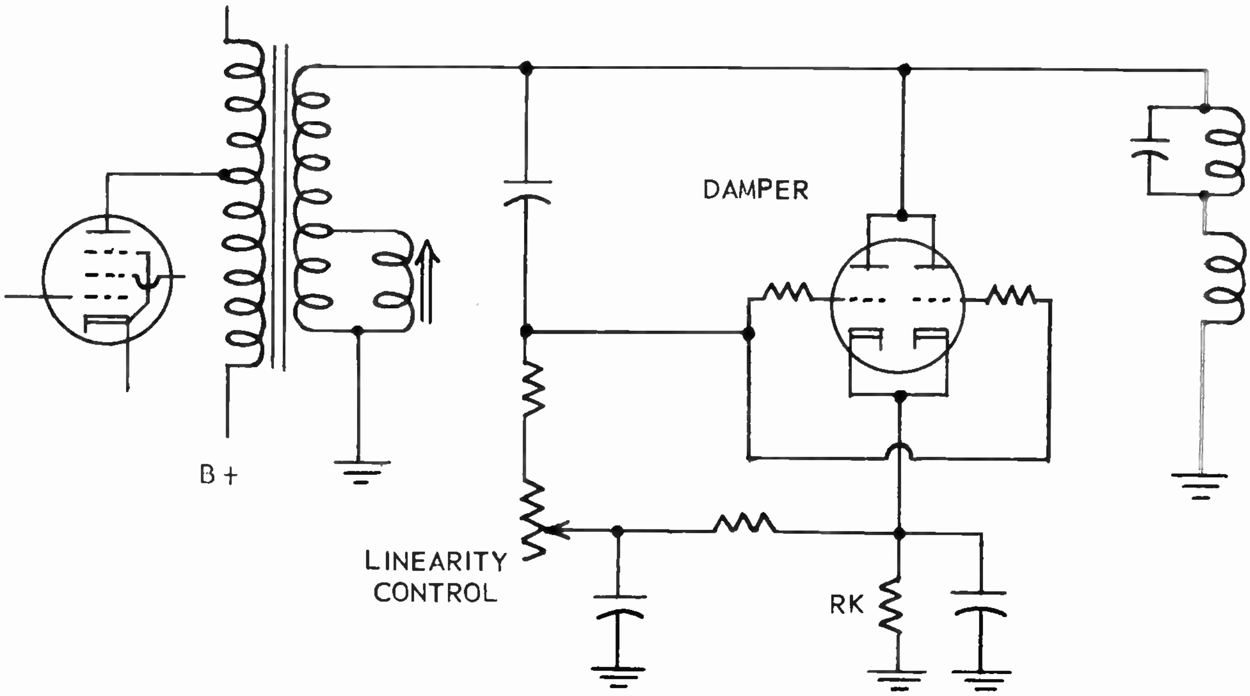


Fig. 59-3. Connections for a twin-triode damper tube.

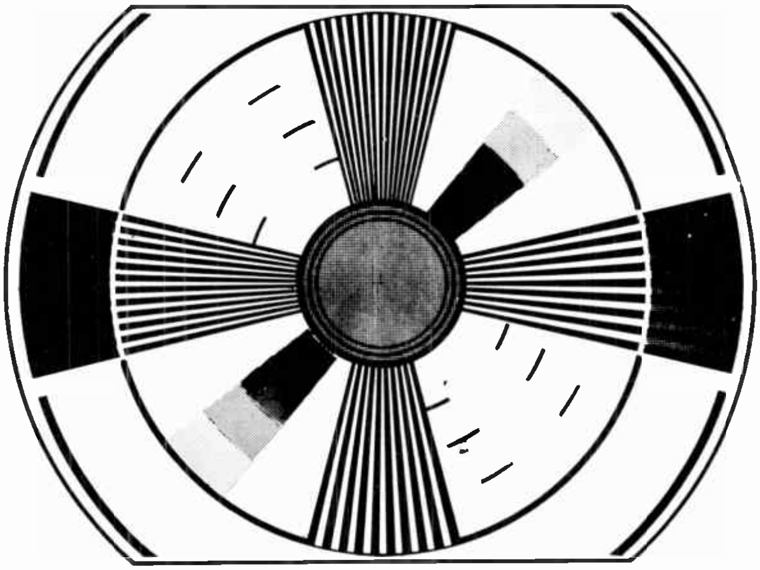


Fig. 59-4. Circles, lines, and wedges such as found in test patterns.

LINEARITY. Linearity or lack of it is best checked while receiving a test pattern transmitted by a television broadcasting station. Most stations transmit their own designs of test patterns which carry the call letters and channel number. There is a rather wide variety in patterns from different stations, but most of them consist of lines, circles, and shadings on the general order of those shown by Fig. 59-4. The station pattern usually is transmitted for a greater or less length of time preceding regularly scheduled programs, and sometimes between programs.

The designs on test patterns allow identification of many faults in receiver operation. So far as checks of linearity are concerned the circles should be truly circular, not egg-shaped, not partially squared off, nor otherwise of irregular form. The straight lines should run symmetrically, not in odd directions. The center of the test pattern should be centered between top, bottom, and sides of the illuminated raster area. The pattern should be symmetrical in all parts, with equal spacings between similar details above and below the center, and to the left and right of the center. When these requirements are satisfied there is linear reproduction. Even though they are not perfectly satisfied, the actual moving pictures reproduced during normal reception will be entirely satisfactory to nearly everyone.

Controls or adjustments for linearity will correct for faults mentioned in the preceding paragraph, but there are other distortions which must be corrected with other controls. For example, an effect such as shown at 1 in Fig. 59-5 is not due to poor linearity, it is due to too much vertical size or height. Circles have become ovals, but they are well shaped and symmetrical ovals. At 2 is shown one effect of poor vertical linearity. Circles have become egg-shaped, pointed toward the top. The center of the pattern now is below the center of the mask opening or below the center of the raster area.

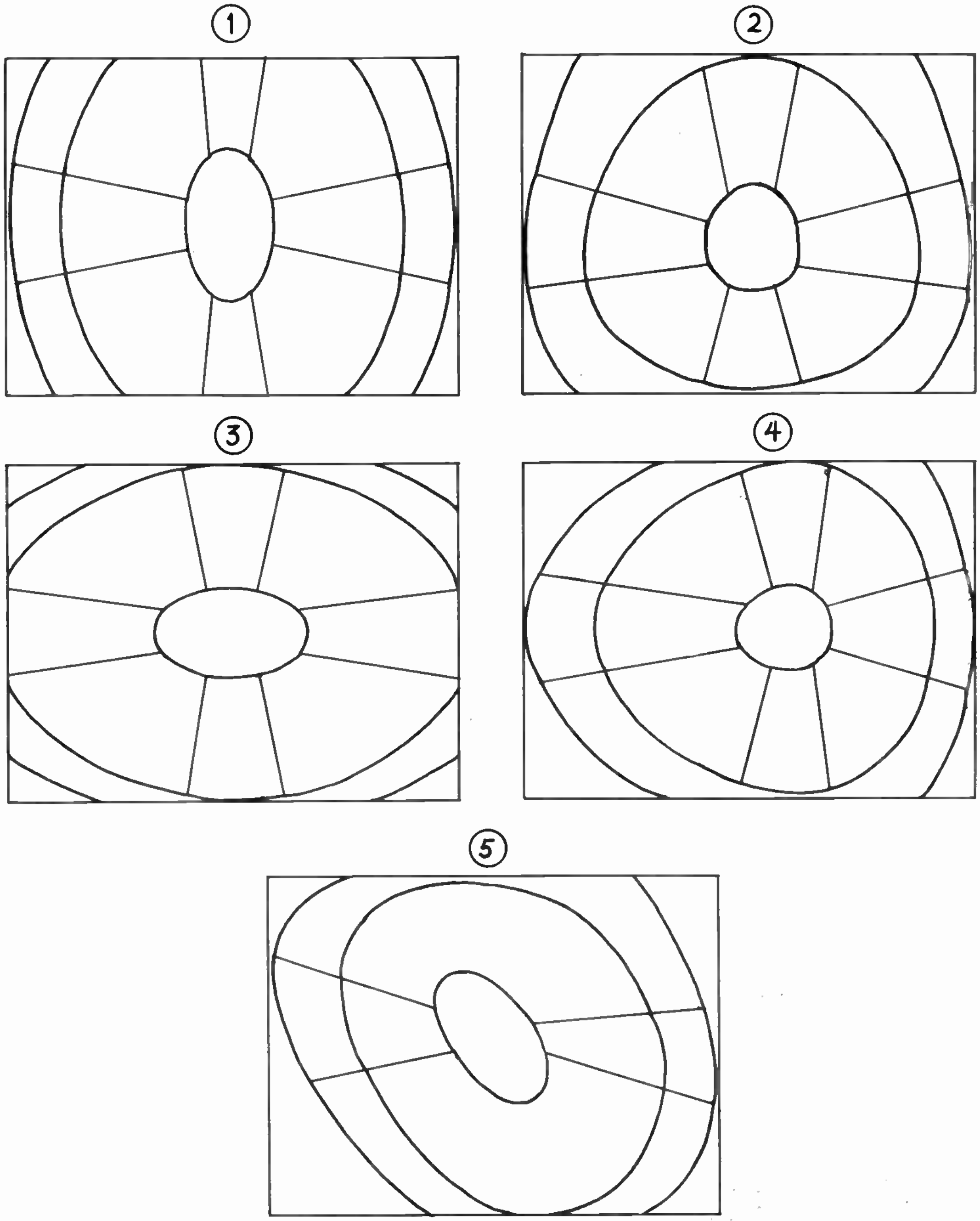


Fig. 59-5. Defects which may appear in reproduction of test patterns.

At 3 in Fig. 59-5 is shown the effect of too much horizontal size or width. Again the circles have become symmetrical ovals. One kind of poor horizontal linearity is illustrated at 4. The circles have become egg-shaped, pointed toward the left. The center of the pattern is displaced toward the right in the raster. Stretching of the pattern which is shown on the left-hand side in diagram 4 might be on the right, with compression on the left. Stretching shown at the top of the pattern in diagram 2 might be at the bottom.

An effect such as shown by diagram 5 seldom if ever results from incorrect adjustment of linearity controls. More often this distortion is due to too little gain, as from setting the contrast control too low, or it may be due to a defective video amplifier tube or to other defects in the video amplifying system.

GRID BIAS FOR LINEARITY. Our difficulties in maintaining linearity commence with charging of the sawtooth capacitor. When voltage is applied to a capacitor and resistor in series with each other the capacitor does not charge at a uniform rate. The charge and the capacitor voltage increase in relation to time as shown by the curve at the left in Fig. 59-6. The amount of series resistance affects only the length of time required to reach any certain percentage of full charge. The curve showing charge versus time never can be made a straight line.

The non-uniform rate of increase in sawtooth capacitor voltage would cause a similarly non-uniform shape of the trace portion of the sawtooth voltage wave. This distortion of sawtooth waveform is held within reasonable limits by using only the first part of the charging curve, as from a to b in the diagram. That is, discharge and retrace begin when capacitor charge has increased only to point b, and before the charging curve commences to bend over more sharply.

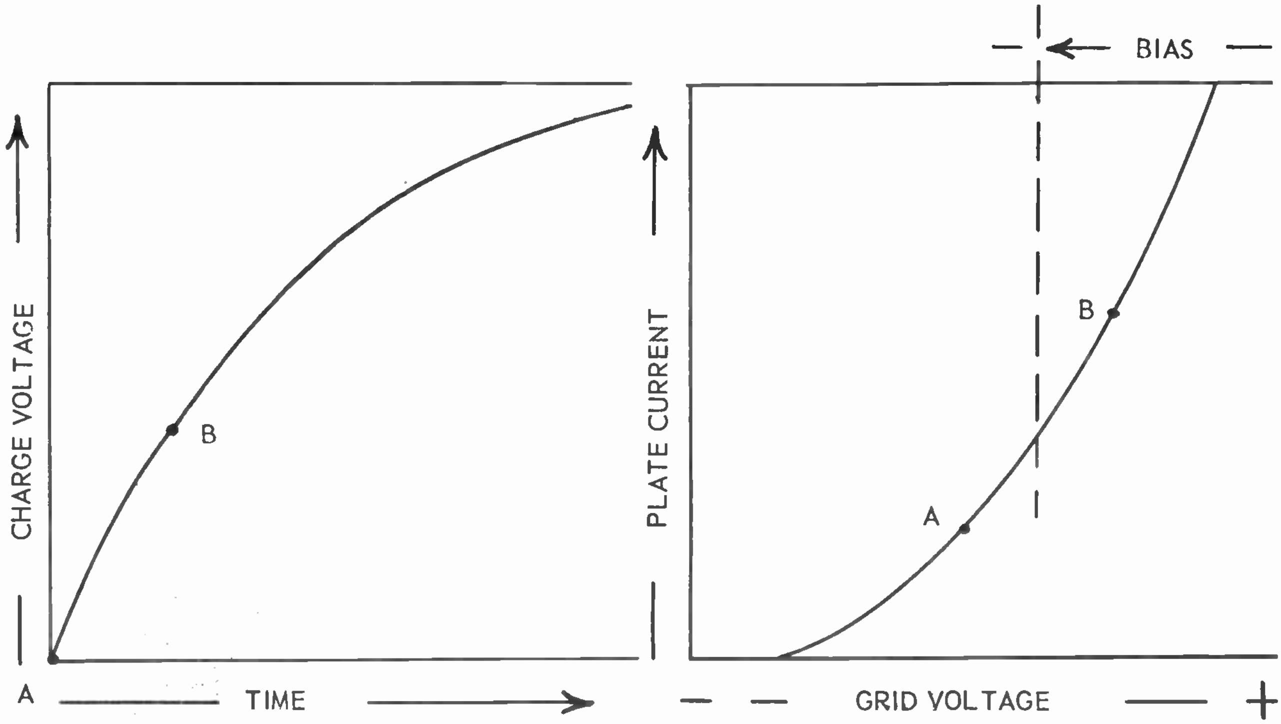


Fig. 59-6. Rate of change of capacitor charging voltage (left) and compensating curvature of amplifier transfer characteristic (right).

The sawtooth capacitor voltage is applied to the grid of the sweep amplifier. If the amplifier is operated on a straight portion of its grid-voltage plate-current transfer characteristic the curvature of sawtooth voltage will produce a corresponding non-uniform plate current. But if the amplifier is so biased that the sawtooth voltage changes fall on a curved part of the transfer characteristic, as between a and b at the right in Fig. 59-6, the opposite curvature of the characteristic will correct for all or most of the non-uniformity of sawtooth voltage. Then amplifier plate current will vary at a practically uniform rate between minimum and maximum.

Since the sawtooth voltage ordinarily comes to the amplifier grid through a capacitor this voltage is alternating at the amplifier grid. The zero value of an alternating grid voltage always falls at the bias voltage of an amplifier. By varying the negative bias voltage on the amplifier grid the portion of the characteristic curve acted upon by the sawtooth voltage may be moved up or down. The bias may be adjusted to a value that allows compensation of sawtooth voltage curvature, or to a value which very nearly attains this ideal.

When an adjustable amplifier bias is used for linearity control the biasing element usually is a variable

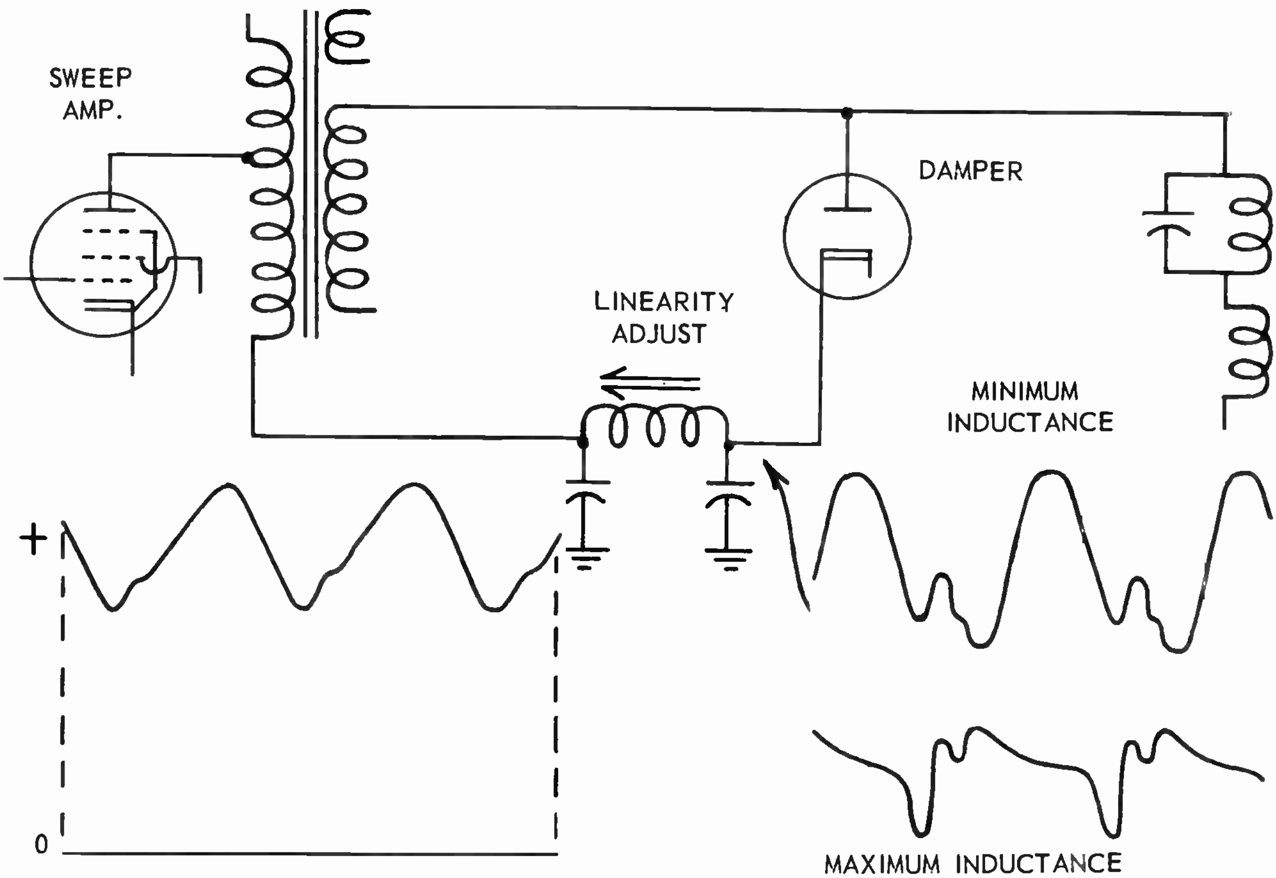


Fig. 59-7. Waveforms at input and output of linearity adjusting choke in damper cathode circuit.

cathode resistor bypassed by a large capacitance. This type of linearity control will also compensate to a greater or less extent for non-linear effects resulting from time constants in the inductive-resistive circuits of the output transformer and deflecting coils. As you know, magnetic fields around inductive circuits do not instantly increase to their maximum strength when voltage is applied. The current and the field strength increase at a decreasing rate which might be shown by the same curve used at the left in Fig. 59-6 to illustrate increase of charge on a capacitor.

DAMPER ADJUSTMENTS FOR LINEARITY. Fig. 59-7 shows connections for a linearity adjusting inductor or filter in series between the damper cathode and the primary winding of the output transformer. This circuit, and others generally similar, have appeared in several earlier diagrams. Typical oscilloscope traces taken at the damper plate or at the input to the linearity filter are shown for minimum inductance and for maximum inductance of the filter choke. The inductance is varied by a movable core in the choke coil. The voltage waveform at the filter output and at the transformer primary carries a series of pulses rather than being a smooth direct current. This is the B-voltage for the plate circuit of the sweep amplifier.

As inductance of the linearity filter is adjusted one way or the other there is a shifting of the phase of the plate voltage pulses with reference to sawtooth grid voltage being applied to the sweep amplifier. This phase shift alters the plate characteristic of the amplifier and alters the waveform of the sawtooth currents in the plate circuit and in the deflecting coil circuit which is connected to the transformer secondary.

Adjustment of this particular type of linearity control usually has greatest effect at the horizontal center of the pattern. The central circle will move a short distance toward the left or right, with accompanying slight crowding of the side toward which the central circle is moved and slight stretching of the opposite side. If a sawtooth wave of deflecting current is observed while the adjustment is altered, the middle of the trace portion may be seen to hump slightly upward or downward.

In addition to the adjustable damper inductor for linearity control there may be either or both of the adjustments shown by Fig. 59-8. There are several taps near one end of the secondary winding in the output transformer. The principal reason for these taps is to allow matching the impedance of the secondary to the impedance of the deflecting coils in order that there may be satisfactory power transfer to the coils. Changing the connection from one tap to another also varies the voltage applied to the damper plate. This, in turn, varies the conduction through the damper and may have considerable effect on horizontal linearity. Too much damper plate voltage will compress the left side of the pattern or picture. Too little voltage may cause stretching or rippling on the left side.

The second linearity adjustment shown in Fig. 59-8 is a tapped resistor, R, connected between the damper plate and cathode. In some receivers this resistor is continuously adjustable by means of a slider instead of solder lugs on the taps. This resistance damps the deflecting coil circuit at all times, but has its maximum effect during that portion of the sawtooth cycle in which the oscillatory current is dying away and before induced current has resumed. This is the part of the trace that forms the left-hand side of the pattern or picture.

Increasing the resistance at R means less damping or less loading of the deflecting coil circuit. This will stretch the left-hand side of the pattern or picture while having little or no effect on parts to the right of the center. Less resistance increases the damping, and will contract the portion of the pattern or picture to the left of center.

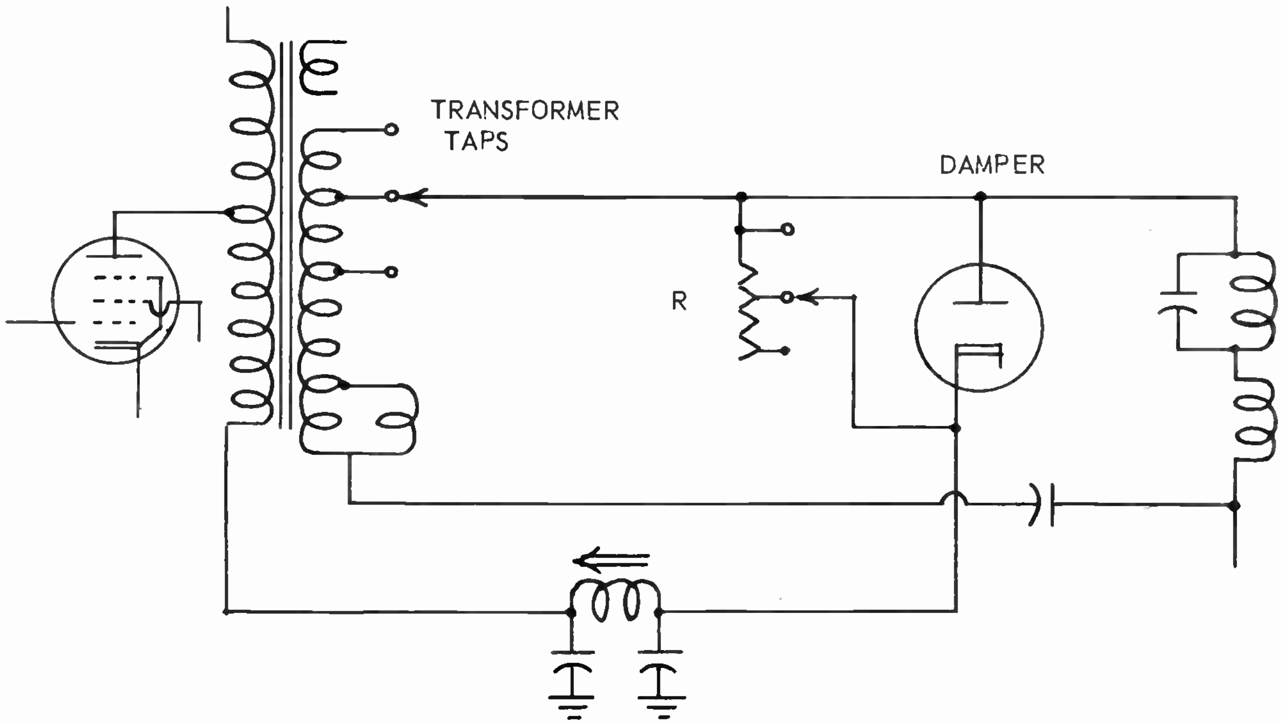


Fig. 59-8. Linearity control by altering damper plate voltage and resistance connected across the damper.

DRIVE CONTROLS. Fig. 59-9 shows connections and effects of adjusting the type of drive control that consists of a capacitance voltage divider across the sawtooth capacitor, with the center of the divider connected to the grid of the horizontal sweep amplifier.

The approximate sawtooth voltage waveforms applied to the amplifier grid are illustrated at the upper left. These are the voltages which appear across the drive control capacitor. When this capacitor is adjusted to have less capacitance it has greater capacitive reactance. More of the total sawtooth voltage then is across the drive capacitor, and there is increase of amplitude in the voltage at the amplifier grid.

Changing the amplitude of sawtooth voltage applied to the amplifier grid causes corresponding changes in amplitude of sawtooth current in the horizontal deflecting coils, as shown by the current waveforms at the upper right in Fig. 59-8. Less drive capacitance increases the amplitude of the deflecting current, while more capacitance, and less capacitive reactance, decreases the amplitude of the deflecting current.

Increasing the amplitude of sawtooth deflecting current, by using less drive capacitance, will increase the width of the reproduced pictures. Although there is some increase of overall width, nearly all of the change occurs in stretching the portion of the picture or pattern which is to the left of center. Too much drive capacitance, and too little capacitive reactance, will cause crowding or cramping to the left of center.

Increasing the amplitude of sawtooth current in the deflecting coils and secondary of the output trans-

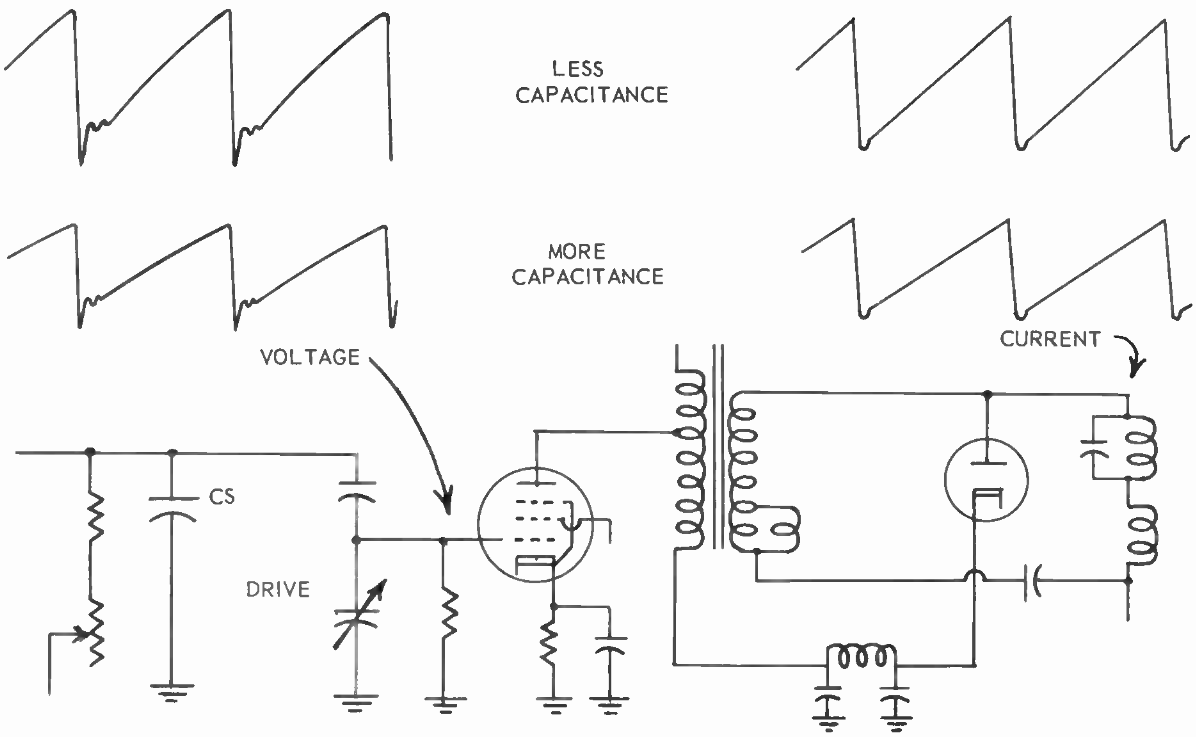


Fig. 59-9. The effects of adjusting a capacitor type drive control.

former increases the amount and rate of current change during retrace periods. With a flyback high-voltage power supply for the picture tube anode, the strength of the voltage pulse induced in the transformer primary, and fed to the high-voltage rectifier, is proportional to this rate of current change. Then the greater the sawtooth current amplitude the higher will be the voltage on the picture tube anode. Thus it comes about that a decrease of drive control capacitance causes an increase of anode voltage at the same time it changes the linearity on the left-hand side of the picture.

Fig. 59-10 illustrates another adjustment often called a drive control, although sometimes it is called a linearity control or a peaking control. This circuit was examined earlier, in connection with the subject of negative peaking. The adjustable unit is a resistor, usually of about 25,000 ohms, which is in series with another resistor of 3,000 to 10,000 ohms and with sawtooth capacitor Cs. Waveforms at the right in the figure show how adjustment of the drive resistance alters the peaking. There is little change in amplitude of the sawtooth portion of the wave, exclusive of the negative peak.

The change of negative peaking at the grid of the sweep amplifier varies the point during a cycle at which this tube resumes conduction after being cut off. There are resulting changes in plate current waveform and in waveform of current induced in the transformer secondary and deflecting coils. The principal effect of misadjusting this drive control is to cause crowding on the right and stretching on the left of a pattern or picture.

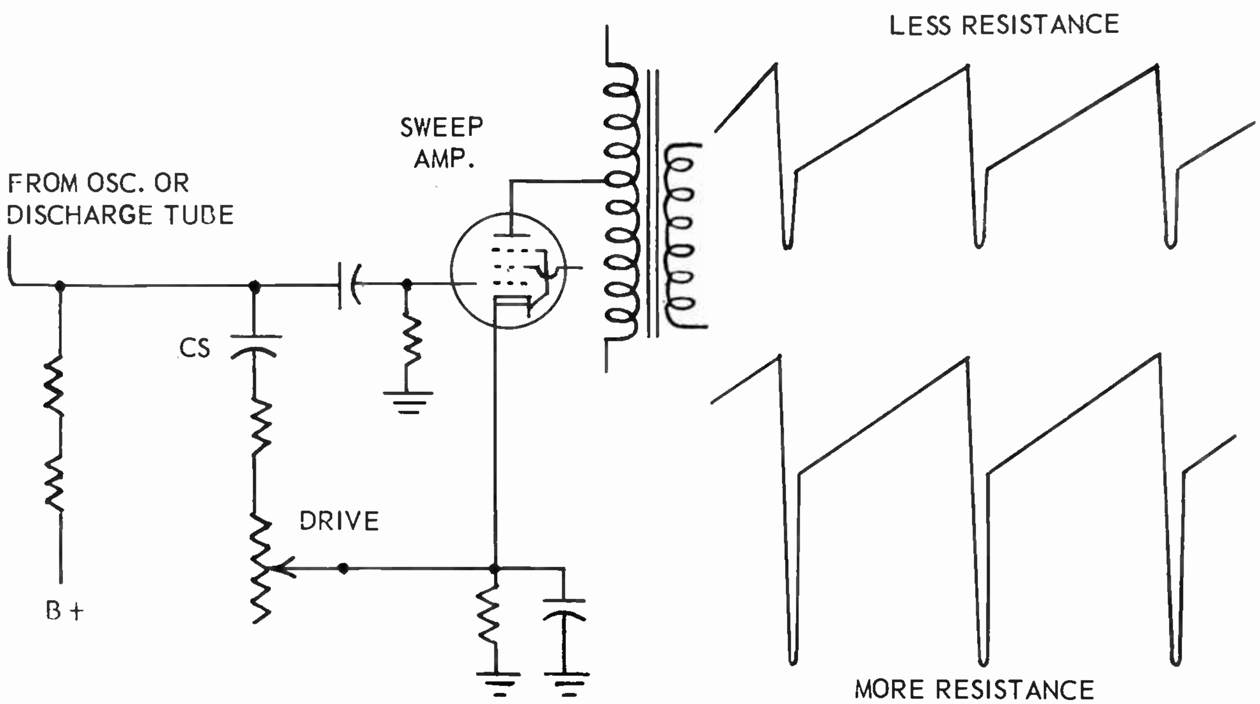


Fig. 59-10. Effects of adjusting a resistance type drive control.

LINEARITY WITH ELECTROSTATIC DEFLECTION. In electrostatic deflection systems which employ an amplifier and an inverter there is a certain amount of natural compensation for non-linearity. As an example, should the trace portion of the sawtooth voltage from the amplifier show an upward curvature, this curvature will be inverted in the output from the inverter. If, however, there is very much distortion of the waveform the point in the cycle at which both voltages go through zero will come too early or too late, causing crowding in one direction and stretching in the opposite direction.

Non-linearity may be due to unequal sawtooth voltage outputs from the amplifier and the inverter. This tendency is compensated for in some receivers by using an adjustable voltage divider between amplifier and inverter. One method is shown by Fig. 59-11, where the linearity control is an adjustable capacitor. Changing the capacitance and its reactance will alter the percentage of amplifier sawtooth voltage applied to the inverter grid, and thus will alter the output of the inverter to match that of the amplifier. An adjustable resistor would be used for linearity control in a resistance type voltage divider.

Linearity often is improved by a voltage feedback from the plate of the inverter to the grid circuit of the of the amplifier. The connection may be made as in Fig. 59-12, through a filter circuit consisting of capacitors and a resistor. In other cases the feedback may omit the series capacitor, using only the series resistor and the capacitor to ground, or there may be only a series resistor of several megohms. If the strength of feedback voltage is adjustable we have a service control for linearity. One such control is shown by the circuit diagram of Fig. 59-13.

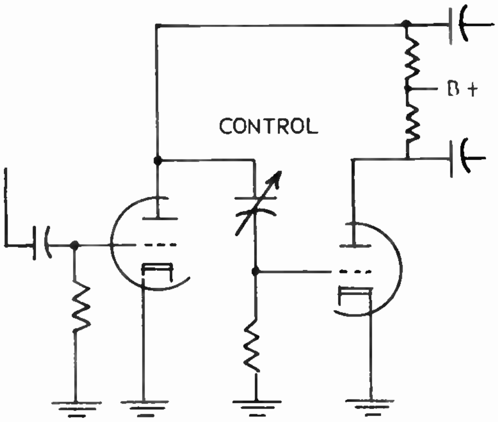


Fig. 59-11. Adjustable voltage divider for linearity control.

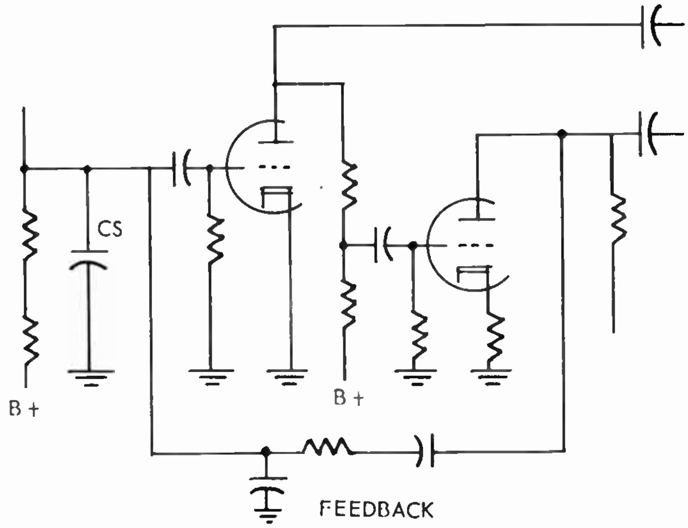


Fig. 59-12. Voltage feedback for linearity control.

ADJUSTMENTS FOR LINEARITY. Should you examine the reproduction of transmitted test patterns on receivers which are delivering pictures of quality pleasing to the average observer, you would find that many of the patterns are not linear. Perfect linearity is difficult to attain, and in most cases it would call for a servicing charge greater than most set owners are willing to pay.

When you undertake the adjustment or control of linearity you should begin with checking certain other controls. First, operate the contrast control and leave it at a setting not so high as to cause picture distortion. Try turning the contrast higher and lower, and leave it somewhat below the point at which distortion appears.

Next, operate both the vertical and the horizontal hold controls. Leave these controls set at points about midway between positions at which the picture or pattern drops out of synchronism, or at which it pulls back after synchronism is lost. Leave the hold controls where slight rotation in either direction does not cause dropout. Make certain that the horizontal hold control is not set at a position where it causes twisting or curving of vertical lines which should be straight.

Do not attempt to make linearity adjustments while receiving pictures in which there is movement. Work only while receiving a regularly transmitted test pattern or while using one of the special signal generators that furnish line patterns suitable for checking linearity. If there is more than one adjustment marked for horizontal linearity or for vertical linearity, try the effects of moving each one a little ways back and forth. Note the parts of the pattern that are most affected by each adjustment.

Linearity controls and drive controls always must be adjusted during the same operation. Both of them affect linearity, and what you accomplish with one may be nullified or at least altered to a greater or less extent by adjustment of the other. Both linearity controls and drive controls affect picture width or picture

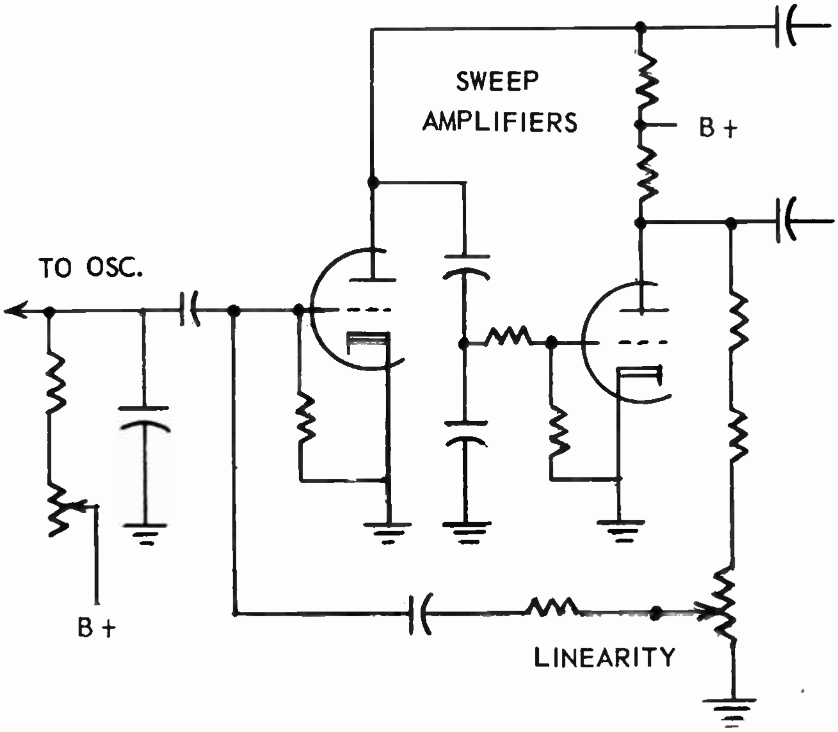


Fig. 59-13. Adjustable linearity control used with electrostatic picture tubes.

height, depending on whether you are working with the horizontal or vertical controls. Therefore, these size controls always will require readjustment when you make changes of linearity or drive settings. It is also quite possible that centering controls will require readjustment after you obtain satisfactory linearity.

Linearity adjustments such as shown by Fig. 59-8 should be changed only in case it is impossible to obtain a **satisfactory** pattern by continued adjustment and re-adjustment of other linearity controls and drive controls. Any change of resistance across the damper tube, or a change of damper plate voltage, will require complete readjustment of the other linearity controls.

You may be able to obtain excellent linearity with a test pattern from one station or on one channel, and much less satisfactory from other stations or channels. Do not conclude that non-linear test patterns are being transmitted. The fault most probably is due to unequal strengths of received signals, or to lack of equal receiver sensitivity for different channel settings, or to misadjustments of tuner circuits. The best that can be done is to make linearity adjustments giving the best compromise on all channels which it is possible to tune.

ADJUSTMENTS FOR DRIVE. Capacitor type drive controls in some receivers having flyback high-voltage systems may be adjusted to cause excessively high voltage at the picture tube anode and excessive plate current in the sweep amplifier. That is, the drive may be adjusted for so little capacitance and such high capacitive reactance that voltage and current ratings of the picture tube and output amplifier will be exceeded. Manufacturers of receivers in which this is possible usually recommend that the drive control be adjusted only while measuring the resulting voltage at the picture tube anode. Because of the very high anode voltage and the danger of shock you should not attempt such measurements until suitable methods and precautions have been discussed in a later lesson.

Drive control capacitors usually are of the trimmer type with which capacitance is increased by turning the adjusting screw clockwise or to the right, which forces the capacitor plates closer together. Turning the screw to the left, or loosening it, reduces the capacitance and increases the amplitude of sawtooth voltage applied to the grid of the sweep amplifier. With these compression type trimmer capacitors, by far the greater part of the possible change of capacitance and possible adjustment range is obtained during the first outward turn of the screw from the position at which the capacitor plates are forced closest together. Therefore, you should change the drive control adjustment very slowly and only a very little at a time.

With the drive control capacitor screw turned all the way in, for maximum capacitance, the picture or pattern will be much too narrow and there will be crowding on the left-hand side. The pattern probably will be fuzzy, and brightness will be far below normal. As the adjustment is changed for less and less capacitance the brightness will increase, the picture will become wider, and the left-hand side will expand. The correct position for drive adjustment is where you obtain the best balance between opposite sides of the picture and have satisfactory brightness with the brightness control set about midway in its adjustment range or slightly above this point.

Too little capacitance in the drive control may cause some crowding of the pattern on the right-hand side and there may be excessive brightness. Vertical bright lines or bars one-fourth to one-half inch wide may appear just to the left of the pattern center, or possibly the pattern may appear to be folded on itself along a vertical line. The bright bars or the foldover effect may appear when you make readjustments of linearity and width controls after having made a preliminary setting of the drive control. In this case the bars or lines should be removed by adjustment of the drive control, even though they seemed to result from changes of linearity or width controls.

The drive control of Fig. 59-10, which consists of an adjustable resistor in series with the sawtooth capacitor, is adjusted in much the same general manner as the capacitor drive control. More resistance causes more peaking, and higher voltage at the picture tube anode when the high-voltage power supply is of the fly-back type. In receivers using this type of drive control there is little if any danger of causing excessive anode voltage.

The resistance type drive control should be adjusted so that the pattern is well balanced on both sides, not crowded on either the left or right. As a general rule the control is set for maximum resistance with which there still is good balance or good linearity in a horizontal direction. This will allow picture tube anode voltage high enough for satisfactory brightness.

PICTURE ENLARGEMENT. Some receivers allow the picture to be enlarged, when desired, by operation of a switch that is accessible to the user. This switch may be manually operated or it may be operated electrically by means of a relay connected to a remote control push button at the end of an extension cord. When the picture is enlarged the outer parts disappear as they move beyond the limits of the screen, but the central part becomes both wider and higher.

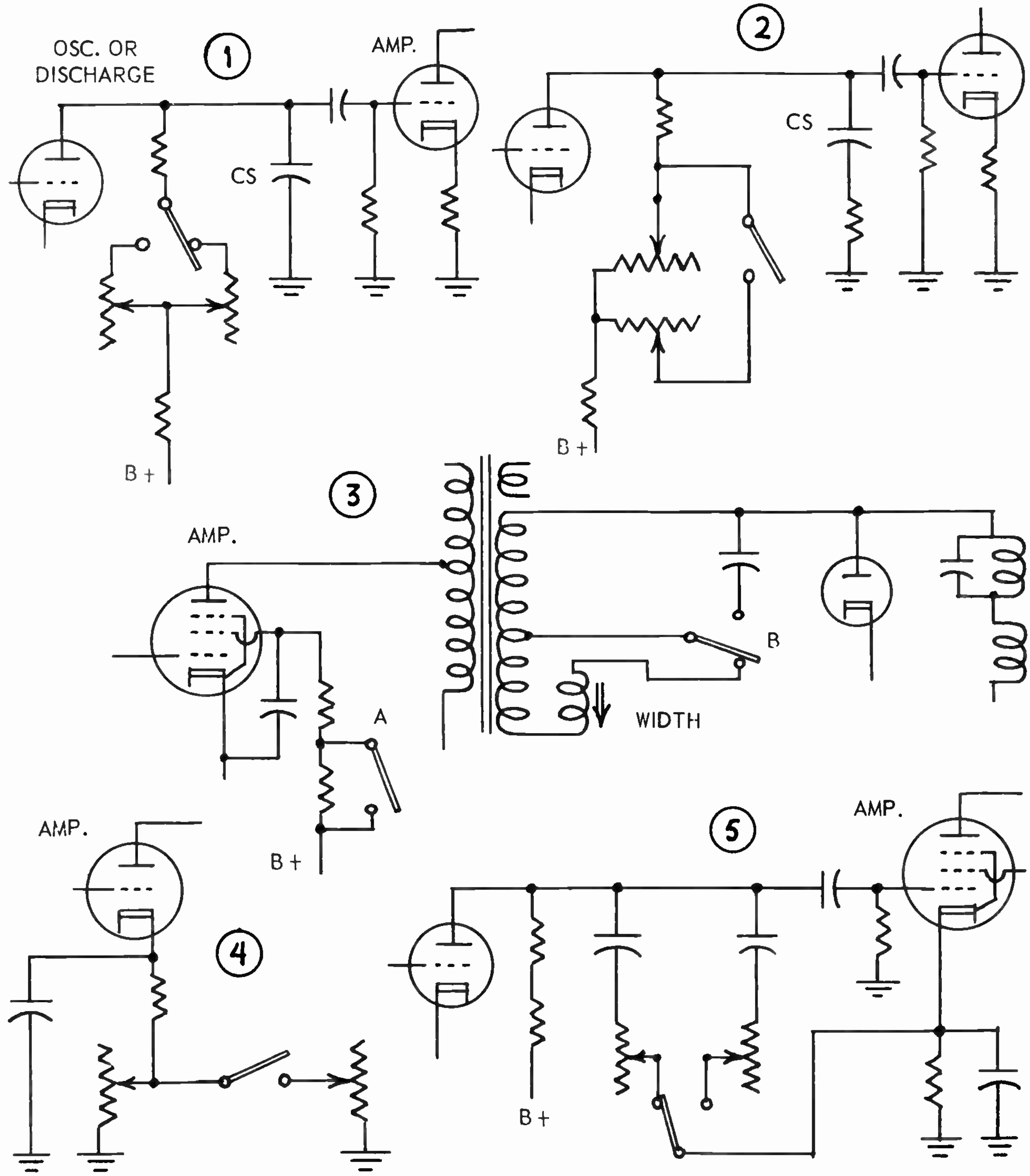


Fig. 59-14. Switching methods used for temporary enlargement of the picture.

At first thought it might seem necessary to provide switching only for the width and height controls. But, as we have just learned, these controls interact with many others. Increase of width and height may require simultaneous changes for any or all of the controls for linearity, drive, hold, brightness, focus, and centering. The changes which provide enlargement must not alter any of the settings or adjustments that control the picture when of normal size. Consequently, dual control elements must be provided and there must be separate adjustments for each of them.

The exact methods used for enlagerment vary with the different electrical designs of different receivers, or with the types of controls that are employed for normal picture size. Fig. 59-14 shows some of the dual controls that are employed. Either height or width may be changed with a two-position switch such as used in diagram 1. The sawtooth capacitor is charged through one of the adjustable resistances for normal picture size, and for enlargement is charged through the other resistor which is adjusted for less resistance to allow increased sawtooth amplitude. Diagram 2 shows another method of changing the charging resistance in series with the sawtooth capacitor. Size is increased by closing the switch to place the two charging resistances in parallel and thus reduce their combined resistance.

Diagram 3 shows several methods for increasing picture width. Closing switch a in the line to the sweep amplifier screen shorts out part of the series voltage dropping resistance. This increases amplification in the tube. When switch b is moved upward from the normal size position in which it is shown one end of the width control inductor is disconnected from the secondary of the output transformer to allow maximum width. This switch also closes a connection to a capacitor, which thus is connected across part of the secondary winding. The effect of the capacitor is to decrease the flyback voltage applied to the picture tube anode. Less anode voltage allows the deflecting magnetic fields to have greater bending effect on the electron beam and to cause greater deflection both horizontally and vertically. This reduction of anode voltage also helps compensate for the increase that would be brought about by raising the amplifier screen voltage.

Diagram 4 of Fig. 59-14 shows a dual control for vertical linearity. One of the adjustable cathode bias resistors is used for normal picture tube size. Closing the switch places the two adjustable resistors in parallel to reduce their combined resistance. Diagram 5 shows connections of dual resistors for a resistance type drive control. Two sawtooth capacitors of different values are each in series with one of the adjustable drive resistors. The switch places either one or the other of the capacitor-resistor combinations in circuit.

If hold controls are to be varied, a switch is arranged to short circuit part of the hold control resistance used for normal picture size. If brightness is to be varied, there will be a switch that short circuits part of the brightness control resistance when the picture is to be enlarged. This reduces the negative bias on the picture tube grid. A change of focus would be compensated for by a switch that varies the current in the focus coil. All switch elements required for changing the picture size in any given application are parts of a single unit, and all are operated together by the control knob or relay.

With the control switch unit in its position for normal picture size all adjustments are made in the usual way. Then the switch is changed to its position for enlargement and all the auxiliary controls are adjusted.

CONTROL ADJUSTMENTS. It is time to take stock of all the control adjustments that have been examined in recent lessons. They are listed in the accompanying table. In most receivers three types of controls are accessible to the operator, for use during reception. The other seven are service adjustments. Some of the service adjustments ordinarily are found only in horizontal sweep systems. Others are used in both the horizontal and the vertical sweep sections. Some service adjustments are needed only with magnetic deflection picture tubes. Others are needed whether the picture tube is of the magnetic or the electrostatic type, although the adjustment or control methods will be different for the two kinds of tubes.

If you count the X's on the lines for service controls you will find that twenty varieties have been examined up to this point in your studies. How much can you recall about each one?

For instance, when thinking about size controls do you remember that size may be changed by varying the sawtooth amplitude, by varying gain of the sweep amplifier tube, by using an adjustable inductor on the secondary of the output transformer, and so on? Unless you have a rather clear idea of how each of all these listed controls is intended to act, and how it should be adjusted, it is in order to check back through the lessons to refresh your memory.

CONTROLS AND ADJUSTMENTS

NAME OF CONTROL	ADJUSTED DURING	SWEEP SYSTEM		PICTURE TUBE	
		Hor.	Vert.	Magnetic	Electrostatic
1. Contrast	Operation				
2. Brightness	Operation				
3. Hold	Operation	X	X		
4. Frequency, afc systems	Service	X		X	
5. Size	Service	X	X	X	X
6. Focus	Service			X	X
7. Center	Service	X	X	X	X
8. Skew	Service			X	X
9. Linearity	Service	X	X	X	X
10. Drive	Service	X		X	

TELEVISION

LESSON No. 60

TELEVISION PICTURE TUBES

"Training Your Mind"

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind
C.T.I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON No. 60

TELEVISION PICTURE TUBES

We have brought the picture and sync signals, or voltages resulting from these signals, all the way from the antenna to the picture tube. Various important features relating to performance of picture tubes have been discussed. Now we shall examine the construction, the electrical and light emitting characteristics, and various accessory parts which relate to these tubes.

Television picture tubes may be classified in various ways. Considering first the materials used for envelopes we have either all-glass construction or else have metal cones or shells combined with glass face-plates and glass necks. Second, from the standpoint of screen shape there are round tubes and rectangular tubes. These physical characteristics combine into four general styles; all-glass round tubes, all-glass rectangular tubes, round tubes with metal cones, and rectangular tubes with metal shells or flares.

The proportions of one style of all-glass round picture tube are shown by the upper drawings of Fig. 60-1 and those of a comparable all-glass rectangular tube are shown by the lower drawings. Dimensions on all drawings are to the same scale. Both tubes here represented are of nominal 16-inch size.

At the top of Fig. 60-2 are shown the proportions of a 16-inch round tube with metal cone. Below are shown the proportions of a 17-inch rectangular tube which has a metal shell.

Picture tubes may be classified also in accordance with methods of focusing and deflecting for which they are designed. During all the recent years in which television became widely used, the majority of receivers have been equipped with tubes designed for magnetic deflection by means of the usual yoke and for magnetic focusing by means of an electromagnetic coil. More lately we have seen many magnetic deflection picture tubes focused by means of permanent magnets applied in a rather wide variety of mechanical designs.

Picture tubes constructed for electrostatic deflection and electrostatic focusing use a focusing anode as part of the electron gun, and have deflecting plates ahead of the gun. This type of tube has been made in sizes from three-inch to twenty-inch diameter.

Some years ago quite a number of picture tubes were designed for magnetic deflection and electrostatic focusing. Most of these types became obsolete before the great popularity of television, hence never came into general use. Tubes of this general style are again being used, because they need only a focusing anode instead of a wire-wound focus coil containing much copper or else permanent magnets made with some of the rare alloy metals.

People not concerned with technical details most often classify picture tubes according to nominal size in inches, which is the outside diameter of a round tube or the diagonal distance across the front of a rec-

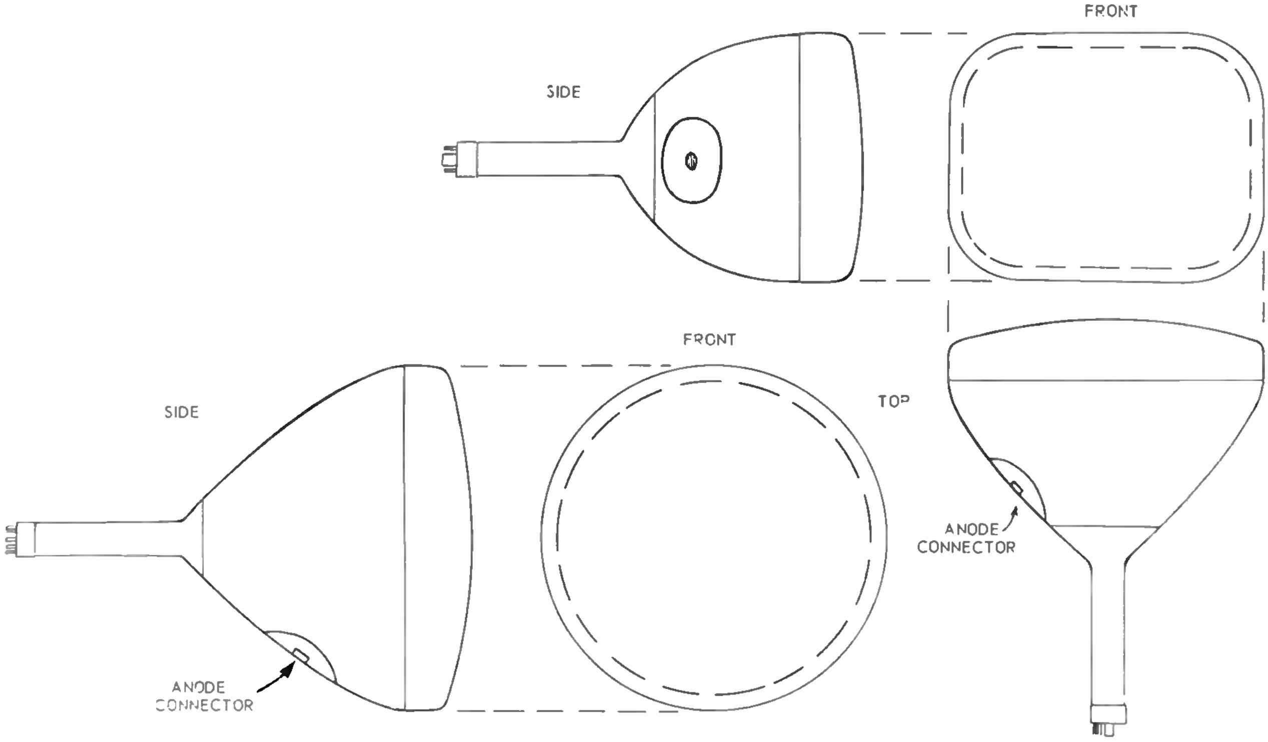


Fig. 60-1. Proportions of all-glass round and rectangular picture tubes.

tangular tube.

TYPE DESIGNATIONS. The numeral or numerals which appear first in the type designations of picture tubes indicate the approximate greatest outside diameter of round types or the approximate distance across diagonally opposite outside corners of rectangular types. For example, a 24-inch round tube is about 24 inches in diameter at the face or screen end, while a 24-inch rectangular tube measures about 24 inches across the diagonal at the face or screen end.

The first number of the type designation is followed by a letter indicating merely the order in which that particular design was first registered with the RTMA. To illustrate, we have 24-inch round tubes with type numbers such as 24AP4 and 24BP4. Among the 24-inch rectangular tubes are types 24CP4, 24DP4, 24TP4, and others.

Following the registration letter of all direct view television picture tubes are the letter P and the numeral 4. These refer to the type of phosphor material used on the screen. The P4 phosphor gives white lines or traces and is the only type phosphor used in television picture tubes.

Following the P4 combination may be another letter, usually A or B. This indicates that the tube is generally similar to an earlier type or is an improvement over an earlier type whose designation is the same except for having no final letter. Improvements may be in the screens or faceplates, or possibly in some featu-

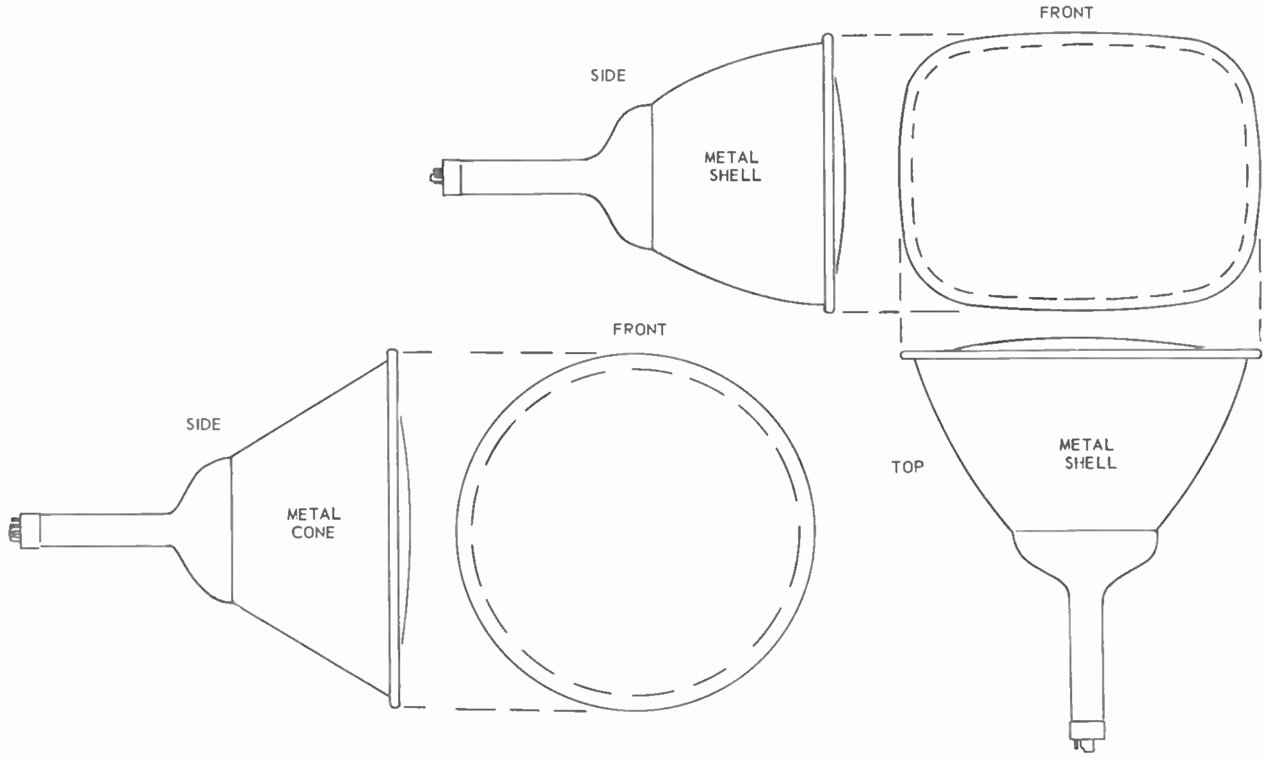


Fig. 60-2. Proportions of picture tubes having metal cones or shells.

res of performance, but tubes with and without the final letter are near enough alike to be interchangeable without making any circuit alterations.

The name picture tube is used for all makes and styles. There also are various trade names, among the best known of which are Kinescope for RCA picture tubes and Teletron for those made by DuMont.

BASING AND CONNECTORS. All present picture tubes designed for magnetic deflection and magnetic focus use the same base, whose pin positions and internal connections as seen from the outside of the base are shown at "A" in Fig. 60-3. This base is called a small shell duodecal type. The spacing would allow for a total of 12 pins were all positions filled. Pin numbering begins at the left of the locating key with 1 and 2, then proceeds as though all positions were filled until coming to the last three pins numbered 10, 11, and 12.

Tubes using electromagnetic deflection with electrostatic focusing use the same duodecal type base with the exception that an additional pin connection is added in position No. 6 for the focusing anode. This is shown at "B" in Fig. 60-3. Certain electrostatic type focus tubes which, by the way, are in the minority, do not have an additional pin connection for this purpose, as voltage for the focusing anode is obtained internally from the cathode of the tube. This is illustrated at "C" in Fig. 60-3.

7 The high voltage anode of all-glass tubes attaches internally to a connector on the flare, as shown in Fig. 60-1. On nearly all such tubes this connector is a recessed metal cavity about 5/16 inch in diameter inserted in the glass. Into this cavity pushes and snaps a spring terminal which is attached to the end of the anode lead. A cup-shaped insulator of soft rubber encloses the lead terminal and covers the cavity connector. On a few tubes the high voltage anode connector is a small ball over which slips a spring terminal on the end of the high-voltage cable. The ball is recessed in a metal insert.

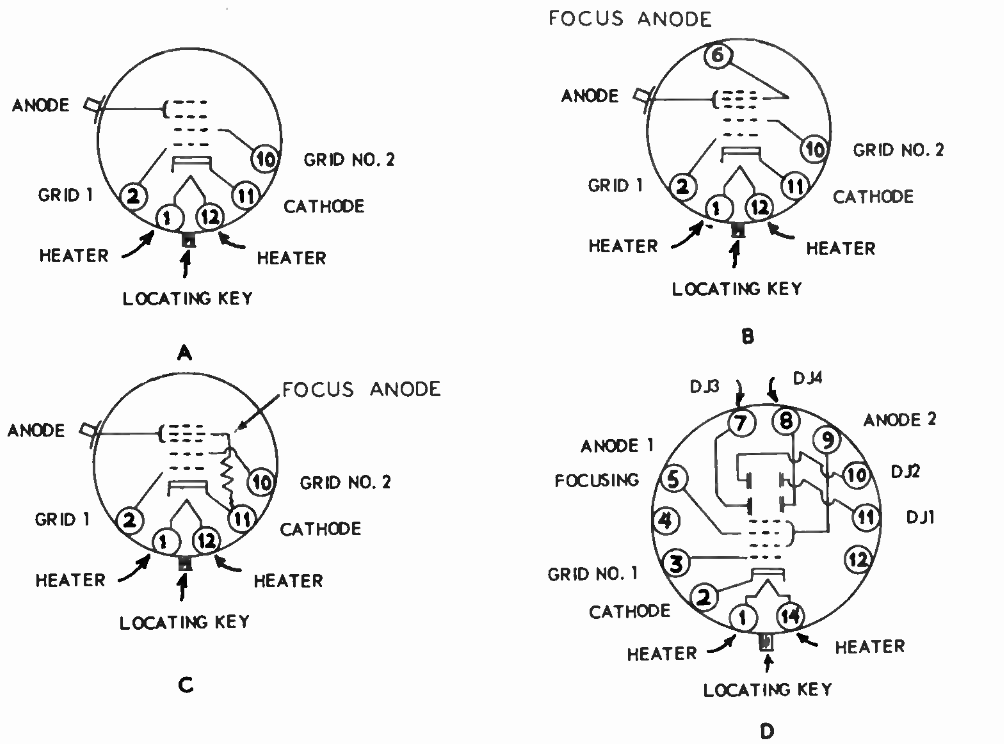


Fig. 60-3. Base pin positions and connections for magnetic and electrostatic tubes.

With metal cone or metal shell tubes the high voltage anode is attached externally to the metal. The anode lead or cable is fitted with a spring clip that fastens onto the outer rim or edge of the metal cone or shell. The rim usually is protected with a rubber or plastic insulating ring which is put in place after the clip is attached.

At "D" in Fig. 60-3 are shown base pin positions and internal connections for the seven-inch 7JP4 picture tube which operates with electrostatic deflection and electrostatic focus. Pin positions are shown as viewed from the exposed end of the base. This base is called a medium shell diheptal 12-pin type. The pin spacing would allow for 14 pins were all positions filled, but pins 6 and 13 are missing. The pair of deflecting plates toward the screen are designated as DJ1 and DJ2. They are connected to pins 11 and 10, and ordinarily are used for vertical deflection. Plates DJ3 and DJ4, connected to pins 7 and 8, are toward the base and ordinarily are used for horizontal deflection.

PHOSPHORS. The inner surface of the glass face of picture tubes is coated with a substance called the phosphor. The phosphor contains chemical salts such as zinc sulphide, zinc silicate, and others which have the ability to emit light when struck by electrons in the picture tube beam. The color of emitted light depends on the chemical composition of the phosphor; it may appear white, green, blue, yellow, orange, red, or of almost any desired hue. Phosphors used in television picture tubes emit practically white light. White light may come from mixtures of salts which, by themselves would emit blue and red-orange, blue and yellow, or other combinations.

The emission of light from the phosphor is called luminescence. Luminescence consists of two effects. First there is light emission that occurs only while the electrons are striking the phosphor, and which ceases instantly when the beam travels onward or is cut off. This part of the emission is called fluorescence.

The phosphors used in picture tubes continue to emit light for a greater or less length of time after they have been excited by the electrons and after the beam has moved on or has been cut off. This continuing emission is called phosphorescence. The fraction of a second during which there is continued emission, or phosphorescence, is called persistence. The time of persistence is determined by the kinds and proportions of salts used in the phosphor. Persistences are roughly classified as short, medium, or long. All television picture tubes are of the medium persistence type.

Except with long persistence types the persistence of the phosphor itself lasts for a time which is shorter than the visual persistence of the human eye. Your eyes retain some impression of the light after both fluorescence and phosphorescence have ceased, or after all luminescence has ceased. With a medium persistence phosphor the brightness drops to about 1/5 of maximum within 6/1000 second after the electron beam passes on, and drops to about 1/20 of the maximum within 1/40 second. During the period in which some persistence of vision lasts in the average human eye the brightness has dropped to less than one per cent of its original maximum.

If phosphor persistence were made too long it would cause the picture reproduced during one frame to hang over into the next frame, or it might cause fast moving light colored objects to be followed by a tail of light, like a comet. If persistence were too short it could cause noticeable flickering with presently used field and frame frequencies.

Each type of phosphor used in picture tubes and other cathode-ray tubes is identified by a number preceded by the capital letter P. As mentioned before, the white phosphor P4 is used in all present day picture tubes. Early picture tubes used a yellowish-green P3 phosphor. Others used the P6 type. Both of these latter phosphors now are obsolete. Cathode-ray tubes for oscilloscopes use phosphors P1, P2, P5, or P11, which have characteristics suited to different purposes. Phosphors P7, P12, and P14 are used chiefly in radar tubes.

The full-line curve of Fig. 60-4 shows relative values of energy emitted by the P4 phosphor at various wavelengths of light. The broken-line curve shows average relative sensitivities of the human eye at the various wavelengths. While radio wavelengths ordinarily are measured in meters or centimeters, the wavelengths of light and related radiations are too short to be expressed conveniently in these units. For instance, the wavelength of green light would be 0.0000522 centimeter. The corresponding frequency is about 575 millions of megacycles. For light wavelengths we most often employ the angstrom or angstrom unit, which is a length equal to one hundred-millionth of a centimeter or about four billionths of an inch.

The wavelengths of Fig. 60-4 are in angstroms. At the top of this graph are indicated the colors which we see in light of various wavelengths. The shortest visible wavelengths are those giving violet light. Still shorter wavelengths are called ultra-violet. They do not excite our eyes to vision, but give the impression of black. The longest visible wavelengths are those producing red light. Still longer radiant wavelengths are called infra-red, and cause only a visual impression of black. We see things as black also when they do not radiate or reflect any visible wavelengths or colors. We see them as white when they radiate or reflect all colors or all visible wavelengths at the same time, and also when they give off certain complementary colors which combine to give the impression of white.

Human eyes have average maximum sensitivity for a wavelength of about 5,500 angstroms, which is a yellow-green hue. Phosphor P4 has one peak of radiant energy at very nearly this same wavelength, and another peak in the blue-violet region where eye sensitivity is low. The combined effects of radiant energies and eye sensitivities at all the various wavelengths give the kind of white light seen on the screen of picture tubes.

IMPROVING CONTRAST. A satisfactory television picture should have a nearly full range of tones from black through grays to white, even when viewed in a normally lighted room. Then there is good contrast. Although contrast may be improved by correct operation of the contrast and brightness controls there are

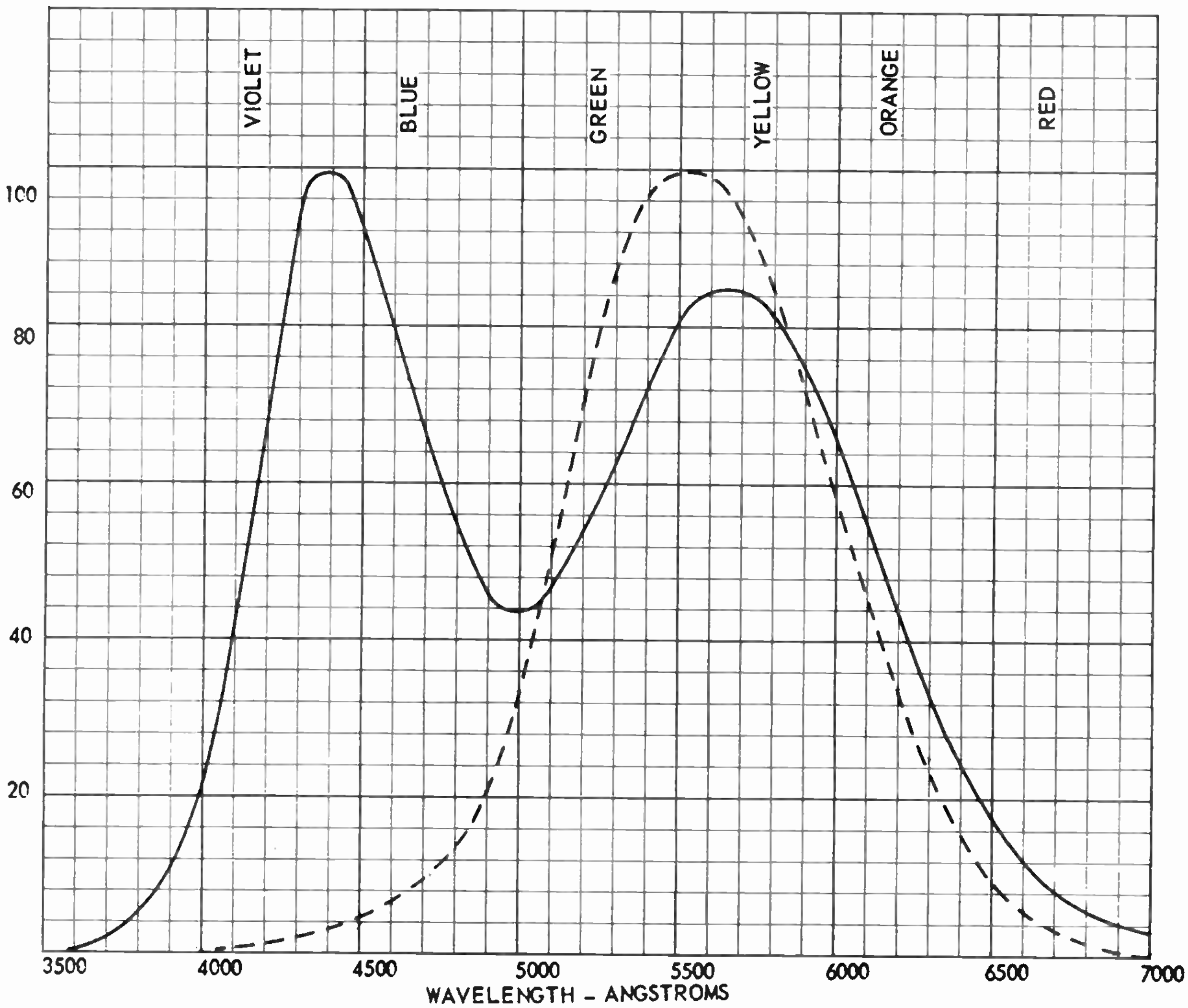


Fig. 60-4. Relative radiant energies of phosphor P4 (full-line curve) and relative sensitivities of average human eye (broken-line curve) at various wavelengths of light.

certain features of the glass faceplates and the phosphors of picture tubes which lighten the blacks and grays to reduce the apparent contrast regardless of control adjustments.

The glass face or faceplate of a picture tube is from 3/8 to 1/2 inch thick. The phosphor and screen coating on the inside of the faceplate is thin but practically opaque, you cannot see through it. The smooth outer and inner glass surfaces and the opaque inner coating allow various reflection effects that reduce apparent contrast. Reflection from the outer glass surface causes images of lighted lamps, of windows opening to daylight, and other bright objects to be mirrored on this surface and to obscure parts of the picture.

Part of the ordinary diffused room illumination passes through the glass and is reflected outward again because the opaque coating on the smooth inner surface makes this surface act much like a mirror. The effect is to lighten the entire picture area. Portions of the picture which are intended to be white are made no whiter, but all the darker areas are made lighter than they should be.

Light emitted from the point on the phosphor being struck by the electron beam goes through the front of the faceplate, but part of this light is reflected from the front surface back onto the screen. The resulting scattering of light all over the screen tends to lighten the grays and blacks of pictures, and thus lessens the contrast between these tones and the areas that already are white. This particular effect may be called halation.

(8) Many picture tubes, especially the newer types, improve the apparent contrast under ordinary room lighting conditions by means of a "gray filter faceplate". The glass in such a faceplate is of a kind which transmits quite freely the light emitted outward from the phosphor, but which absorbs in the glass itself a much greater proportion of the reflections which occur within the faceplate and from the inner coating. Black and grey tones remain of more nearly correct values in relation to the highlights or nearly white areas.

Other "black face" picture tubes provide improvement of apparent contrast by mixing with the originally white screen coating some substance that materially darkens this entire coating which is on the inside of the faceplate. The phosphor still becomes normally bright where excited by the electron beam, but there is less reflection of external light from the darkened inner surface.

The light output of picture tubes has been considerably increased by incorporation of a reflector within the picture tube. This reflector consists of a thin coating of aluminum over the phosphor which acts like a mirror. It is thin enough to permit the electron beam to strike the phosphor in a normal fashion, and thick enough to reflect the light from the phosphor toward the front of the picture tube. It acts much the same as the reflector behind the bulb in a flashlight which concentrates the light into a single direction, giving us greater light output. Picture tubes of this type are referred to as being aluminized. Some manufacturers call them DAYLIGHT, SILVER VISION, SILVER SCREEN, etc.

Reflection images of lighted or luminious object in the room may be reduced in various ways. The faceplate of many of the newer picture tubes has a lightly frosted outer surface which destroys the mirror effect. Special lacquers and other anti-glare coatings are applied to the faceplates of other tubes. There still may be reflections from the protective glass or plastic sheet mounted in the cabinet opening. In some receivers this sheet is tilted slightly forward at its top, so that reflections of bright objects are directed toward the floor. Two or more of the methods for reducing reflection effects may be combined in the same picture tube or in the same receiver.

(9) *MASK AND SCREEN SIZES.* In Figs. 60-1 and 60-2 the front views show in full lines the outside dimensions, and broken lines show the useful screen areas of the picture tubes. These useful areas include the portions of the faceplate coated with screen material or the portions which are sufficiently flat for picture reproduction. The useful areas are from one, to one and one-half inches less in diameter or in width and height than the extreme outside dimensions.

Transmitted pictures have almost square corners and an aspect ratio or width-to-height ratio of 4 to 3. The width is 4/3 or 1-1/3 times the height, and the height is 3/4 of the width. If the entire picture were to be made visible on a round tube the width would be very close to 0.8 times the useful diameter, while the height would be approximately 0.6 times the useful diameter. This proportion is illustrated at A in Fig. 60-5. As an example, on a 16-inch tube with 15-inch useful diameter the reproduced picture would be 12 inches wide (0.8 times 15) and 9 inches high (0.6 times 15).

(10) The visible portion of a picture is that seen through the mask, which is the frame or the opening on the front of the cabinet. A square cornered and straight sided mask opening is not of most pleasing appearance, so the corners always are rounded off to a greater or less extent. Rounding the corners while retaining straight edges is represented by diagram B. Since little of importance ever appears in the extreme corners

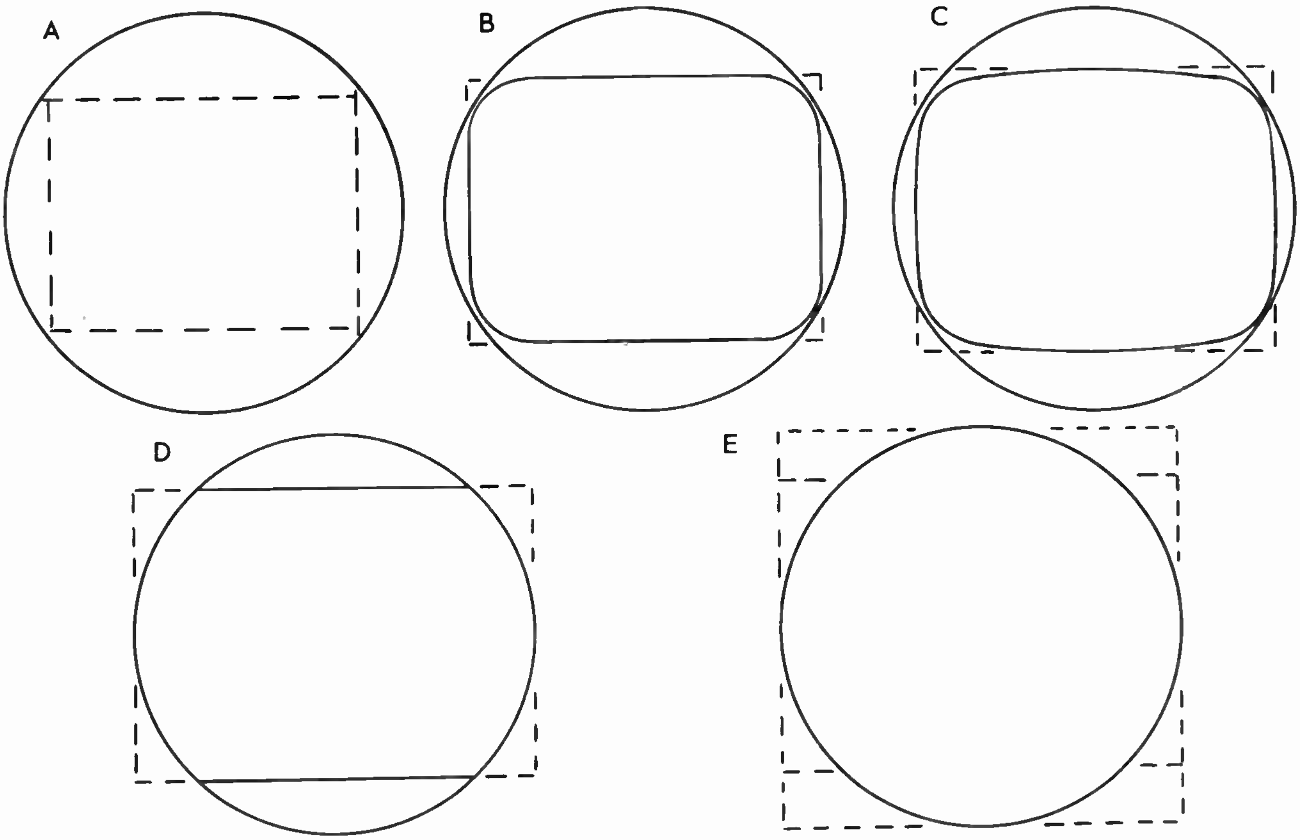


Fig. 60-5. Mask openings in relation to raster dimensions.

of a picture only a negligible portion disappears behind the mask. But the raster and the remaining visible portion of the picture are desirably enlarged.

In addition to rounding the corners of the mask it is sometimes the practice to slightly round the four edges, as in diagram C. By sacrificing still more of the corners the width of the picture may be brought out to the useful screen diameter, as at D. Here we have the largest obtainable raster or picture of correct aspect ratio on a given tube diameter. In some receivers the picture height is made equal to the width, and the entire useful screen area is used as shown by diagram E. Such a picture is not of the original or transmitted aspect ratio. All objects will be somewhat higher in proportion to width than they should be.

Fig. 60-6 shows front outside dimensions and useful areas of typical 16-inch picture tubes of the rectangular type at the left and the round type at the right. These diagrams represent the proportions which have been used in the majority of tubes in this particular size up to the present time. The useful area enclosed by the broken line for the rectangular tube has the standard aspect ratio of 4/3, with the corners rounded.

Within the circular useful area for the round tube has been drawn with a solid line the outside limits for the largest possible picture having the same rounding of corners (the same radius) as used for the rectan-

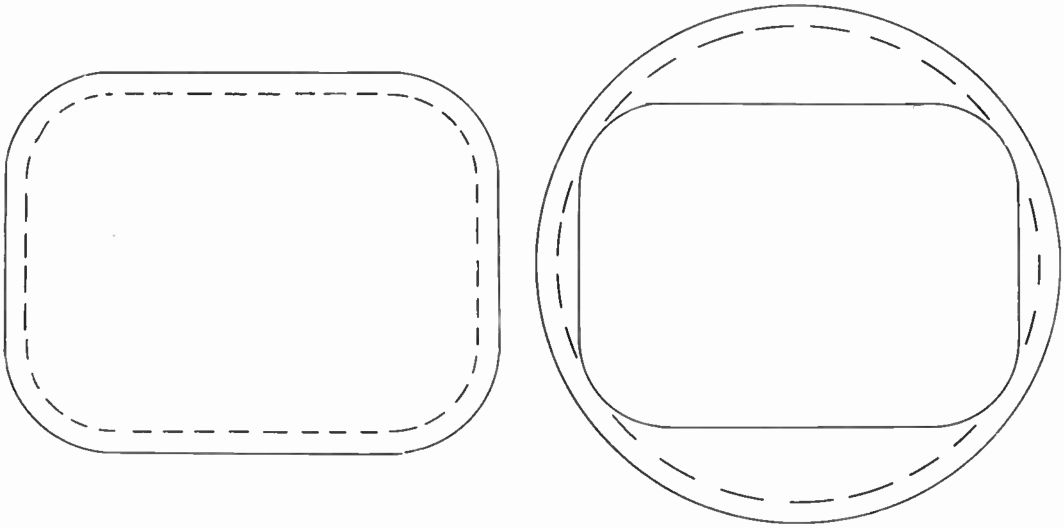


Fig. 60-6. Relative front dimensions of typical 16-inch rectangular and round picture tubes.

gular tube. Picture areas for the two tubes are very nearly equal, which explains why the rectangular tube is called a 16-inch size although it measures only 14-3/4 inches from side to side and 11-1/2 inches from top to bottom.

For any given picture area a rectangular tube will fit into a cabinet that is both narrower and lower than a cabinet accomodating a round tube. This is probably the chief advantage of the rectangular tube over the round type. If a cabinet is wide enough to take a 12-inch round tube with a fraction of an inch to spare it will take a 14-inch rectangular tube so far as width is concerned. If a cabinet is high enough to take a 10-inch round tube, and has two inches or slightly more to spare in width, it will take the 14-inch rectangular tube. These statements refer only to cabinet size, they do not mean that you always can make a quick change from a round tube to a larger rectangular type. There may be mechanical problems related to the chassis parts and to the supports for the tube and its accessories.

ELECTRON GUN AND ANODE. The structure of an electron gun used in an electrostatic tube was examined in an earlier lesson. The parts of a gun used in a tube designed for magnetic deflection and magnetic focus are shown by Fig. 60-7. Visible from the outside of the tube neck are three metal cylinders, each about a half-inch in diameter. The section nearest the tube base is grid 1 or the control grid. This grid almost completely encloses the cathode material and its support. Inside the cathode support is the heater. Just ahead of the control grid is grid 2, which is the first electron accelerating and beam forming element. Connections from two of the five base pins go to the heater. The other three pins are connected to the cathode, the control grid, and grid 2.

Ahead of grid 2 is another longer cylinder which is one part of the anode, the part that is included in the gun structure. On a disc at the forward end of this anode cylinder are carried thin contact springs whose outer free ends bear firmly against an internal conductive coating that extends back into the neck of the tube from the glass flare or from the metal cone or shell, and which is electrically continuous with the flare coating or the metal. The internal conductive coating in the flare of a glass tube, or the cone or shell of metal tubes, forms the other portion of the anode. Voltage reaches the anode cylinder of the electron gun through its connection to the coating or metal in the forward part of the tube, to which voltage is applied through its connector or lip from the high-voltage lead or cable.

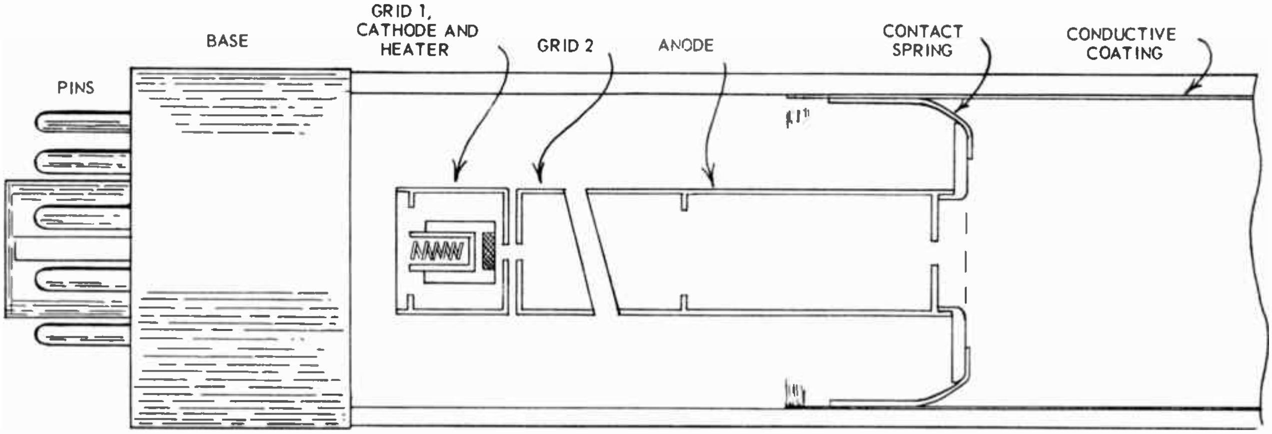


Fig. 60-7. The parts of an electron gun in a magnetic deflection-focus picture tube.

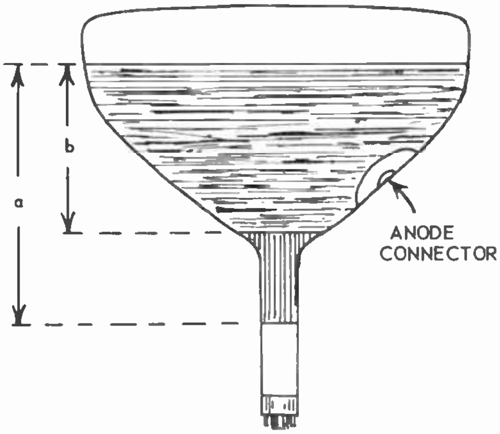
At the left in Fig. 60-8 is an outline view of an all-glass tube. The internal conductive coating extends over the distance a, from near the screen to a point well back inside the neck. On some all-glass tubes there is also an external conductive coating extending over the distance marked b, from near the screen to a point just ahead of the front end of the deflecting yoke.

The external coating is completely insulated from the internal coating. Around the anode connector is an uncoated and non-conductive circular area of the glass. This external coating and the internal coating form the two plates of a capacitor, whose dielectric is the glass between them. The capacitance is used in the d-c filter system of some types of high-voltage power supplies. With the external coating grounded, as always is the practice, it acts also as a shield against external electric fields which might affect operation of the tube.

At the right in Fig. 60-8 is the outline of one style of metal cone picture tube. The internal conductive coating is made electrically continuous with the metal of the cone, and extends back through the glass portion of the tube to point a. This extension permits electrical connection to the anode portion of the electron gun through the contact springs.

Part of the metal cone tube of Fig. 60-8 is a glass flare at b, extending from the back end of the cone proper to the neck of the tube. A similar flare may be seen on the metal shell tube of Fig. 60-2. These glass flares provide insulation and non-conductive spacing between the metal part of the tube and the front end of the deflecting yoke which is placed around the neck. This glass barrier should be kept free from finger prints and reasonably clean in general.

ELECTRICAL RATINGS AND CHARACTERISTICS. In tables of picture tube ratings issued by manufacturers you will find values of maximum permissible electrode voltages. For tubes operated with magnetic deflection and focusing there will be maximum voltages for the high-voltage anode and for grid 2. For tubes operated with electrostatic focusing there will be listing of maximum voltage for the high-voltage electrode, anode 2, and also for the focusing anode 1. For all tubes you will find listings relating to the control grid 1 which give maximum negative bias voltages. It is understood that bias on this grid never may be positive, and during actual operation the bias always will be more or less negative. With negative bias the signal may



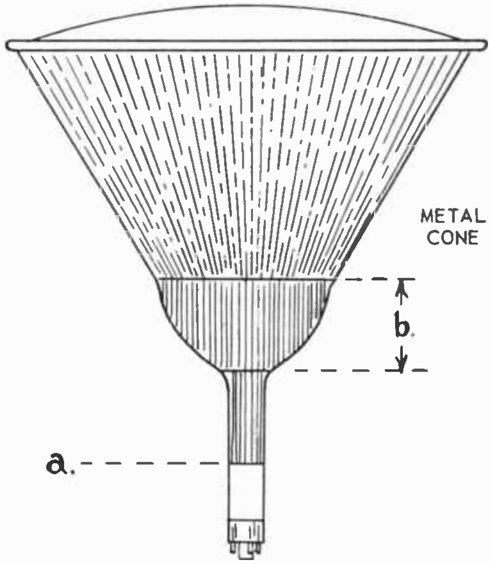


Fig. 60-8. The extent of conductive coatings on all-glass tube (left) and on a metal cone tube (right).

cause a peak positive voltage, whose maximum value will be listed. Included also will be maximum allowable d-c potential differences between the cathode and the heater.

In addition to the maximum ratings will be given typical operating conditions when using certain combinations of anode and grid voltages. Listed here will be the value of control grid negative voltage required to extinguish the light spot on the screen when the beam is focused but is not being deflected. The accompanying table applies specifically only to the 21FP4, 21FP4A, or 21FP4C tube which operates with magnetic deflection and electrostatic focus. It illustrates about what you may expect to find for picture tubes in general, but, of course, with values suitable for each particular type of tube.

In an earlier lesson we studied the effects of electrode voltages and currents on brightness and contrast, arriving at the following conclusions.

Brightness is almost directly proportional to anode voltage, also to anode current or beam current. Greater anode voltage means that electrons arrive at the screen with greater velocity, while greater current means that more electrical energy is being delivered at the screen.

(9) Good contrast and high brilliancy are more easily secured with a strong signal from the video amplifier than with a weaker signal. For first class performance with magnetic deflection tubes it is desirable to have a maximum grid swing of 25 to 30 volts. The same contrast ratio which is obtained with a strong video signal may be had from a somewhat weaker signal by adjustment of contrast and brightness controls, but the picture will not be of equal quality.

It would be an excellent idea to go back and review all the information on picture tube performance given

MAXIMUM RATINGS – Design Center Values.	21FP4-21FP4A-21FP4C	
Heater Voltage	6.3	A.C. volts
Heater Current	0.6	A.C. amp
Anode voltage (High-voltage electrode)	18,000 max.	D.C. volts
Grid No. 4 Voltage (Focusing electrode)	-500 to +1000 max.	” ”
Grid No. 2 Voltage (Accelerating electrode)	500 max.	” ”
Grid No. 1 Voltage (Control electrode)		
Negative bias value	125 max.	” ”
Positive bias value	0 max.	” ”
Positive peak value	2 max.	” ”
Peak Heater-to-cathode Voltage		
Heater positive with respect to cathode	180 max.	” ”
Heater negative with respect to cathode	180 max.	” ”
Heater negative, during warmup period not to exceed 15 seconds	410 max.	” ”

TYPICAL OPERATING CONDITIONS

Anode voltage	16,000	D.C. volts
Grid 4 voltage	-64 to +350	” ”
Grid 2 voltage	300	” ”
Grid 1 voltage, for visual extinction of undeflected focused spot	-27 to -72	” ”
Ion Trap magnet strength (approx.)	35	Gausses

in that earlier lesson. The graphs now will be more instructive because you are better acquainted with the many problems of electron beam control.

In addition to the effects of anode voltage on brightness, this voltage has also a material effect on the size of the light spot formed on the screen. The lower the anode voltage the greater is the diameter of the spot. This results in poorer definition or in poorer rendering of picture detail. The smallest spot and best definition are secured with voltages approaching the maximum or, at least, no lower than those listed for typical operating conditions.

3 *DEFLECTION SENSITIVITIES.* The deflection sensitivity of a picture tube is the distancethat the electron beam will be moved away from the center of the screen by a deflecting field of specified strength. With electrostatic deflection the strength of deflecting field is assumed to be proportional to the peak potential

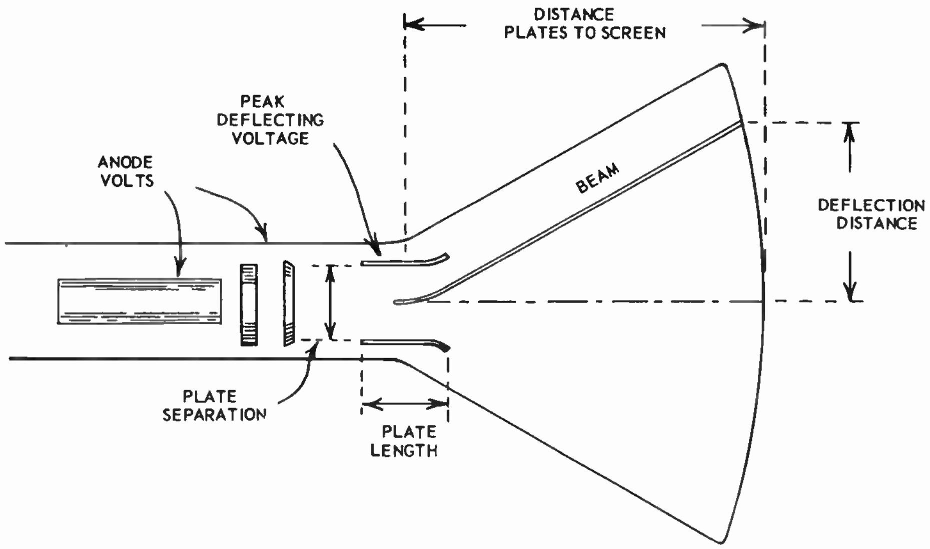


Fig. 60-9. Factors that affect electrostatic deflection distance.

difference or voltage between opposite deflecting plates of a pair. For such a tube we specify the deflection sensitivity as some certain fraction of an inch per volt.

Were a 7JP4 electrostatic deflection tube operated with 4,000 volts on anode 2 the deflection sensitivity for the pair of vertical plates might be between 0.006 and 0.008 inch per deflecting volt, and for the horizontal plates between 0.007 and 0.010 inch per deflecting volt. Within these limits the sensitivity would increase or the deflection would become greater with less voltage on the focusing anode, number 1 anode, and would decrease with higher focusing voltage.

Factors which affect the deflection distance in electrostatic tubes are illustrated by Fig. 60-9. The distance increases in direct proportion to (a) the deflecting voltage applied between the plates, (b) the distance from the plates to the center of the screen, and (c) the length of the deflecting plates. Doubling any of these three factors would, theoretically, double the distance of deflection. Halving any of them would reduce the deflection distance to half.

The deflection distance is inversely proportional to (a) the voltages on the anodes with respect to the cathode, and (b) the separation between the plates of the pair. Doubling either of these factors would, theoretically, cut the deflection distance in half. Halving either of them would double the deflection distance.

You cannot alter the distance from the deflecting plates to the screen in any existing tube, nor can you alter the plate length or separation. Therefore, deflection distance may be varied only by changes of deflecting voltage applied to the plates of a pair and by changing the voltages on the anodes. You may increase the deflection distance with more deflecting voltage from the sweep amplifiers or with lower voltages on anode 2 and anode 1, or with both these changes. Less deflection, which means less width or height of the picture, will result from smaller deflecting voltage or greater anode voltages, or both.

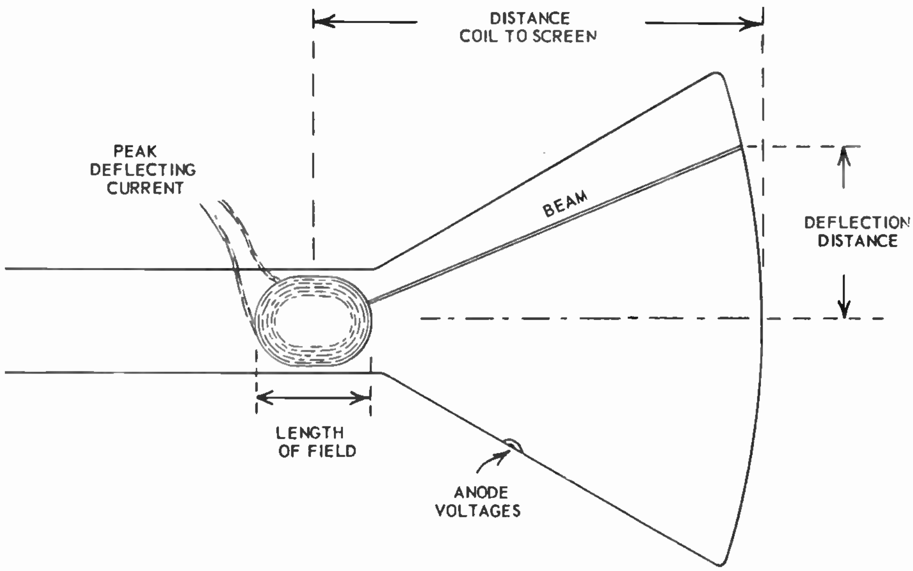


Fig. 60-10. Factors that affect magnetic deflection distance.

Magnetic deflection sensitivities usually are specified as the distance that the electron beam is moved away from the center of the screen per gauss of deflecting magnetic field. The gauss is a unit of measurement for magnetic field density. One gauss is the intensity of a field in which there are 6.45 magnetic lines of force per square inch of cross section. Field intensity of an electromagnet coil varies more or less directly with number of turns in the winding and with amperes of current in the winding. You cannot vary the number of turns in a deflecting coil which is already constructed, so the only way of changing the field strength is to change the deflecting current in the coil. With air-core coils the field intensity in gausses is almost directly proportional to coil current, but when the coil has a magnetic core the field intensity may change either more or less rapidly than current.

The factors which affect magnetic deflection distance are illustrated by Fig. 60-10. Deflection is directly proportional to the distance from the center of the deflecting coil to the center of the screen, also to the length of the deflecting magnetic field. Deflecting coils or yokes always are longer from front to back than are focusing magnets.

Deflection is directly proportional to intensity of the deflecting magnetic field, which, for any given construction, means more or less directly proportional to peak deflecting current in the coil. It would be desirable to have the field lines of force perfectly straight and uniformly distributed where the beam passes through them, but this is not practicable in commercial construction. The non-uniform field which exists in actual yokes tends to make the beam spot somewhat oblong or elliptical rather than perfectly round. There is also a tendency to distort the shape of the raster, so that its edges are not perfectly straight or so that the edges curve slightly inward or outward. Careful design and shaping of the deflecting coils reduces these distortions to where they are not objectionable for picture reception.

Deflection distance is inversely proportional to the square root of the anode voltage. What this means in practice is best illustrated by an example. Assume that with 8,000 anode volts we have some certain def-

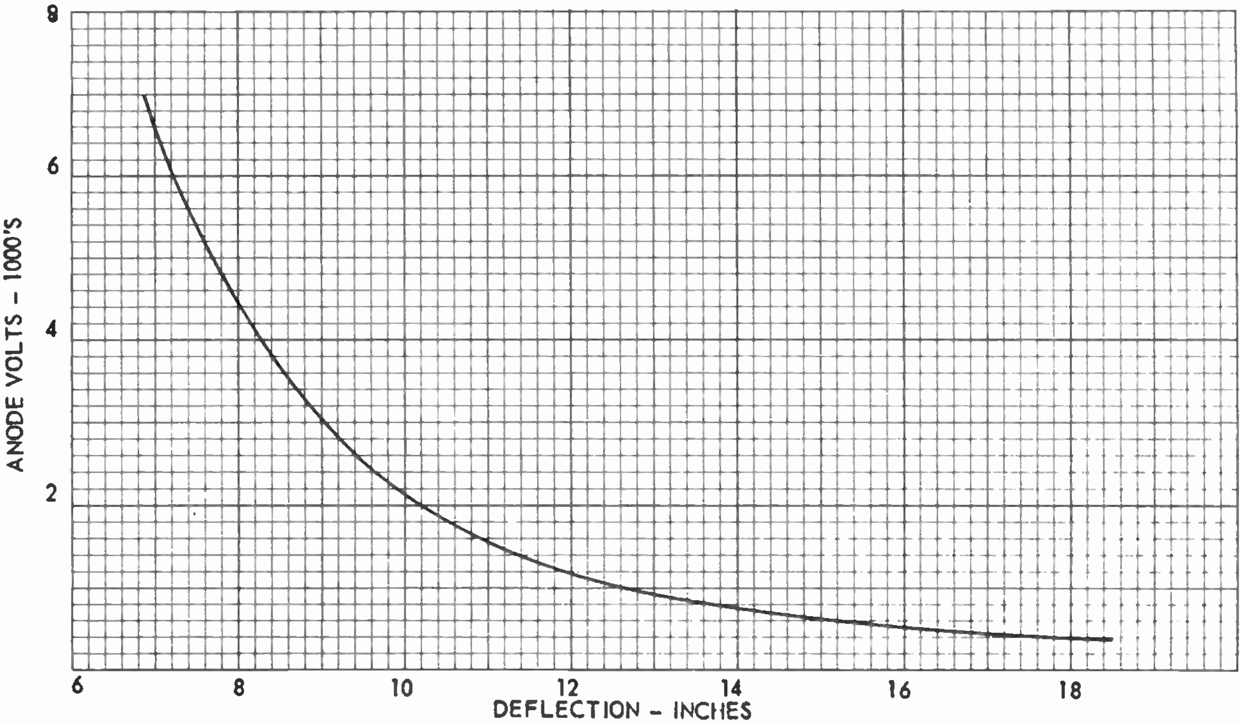


Fig. 60-11. The effect of anode voltage on magnetic deflection distance.

lection, and that the voltage is increased by 50 per cent to 12,000 volts. The square roots of 8,000 and 12,000 are respectively about 89.3 and 109.4. Because of the inverse relation we take the reciprocals of these roots, which are about 0.01120 and 0.00914. Then the deflection with 8,000 volts is proportional to 0.01120 and with 12,000 volts is proportional to 0.00914. Increasing the anode voltage by 50 per cent has reduced the deflection by only 18.4 per cent.

In the case of an electrostatic deflection tube an increase of 50 per cent in anode voltage would decrease deflection by 33.3 per cent, because deflection is inversely proportional to anode voltage itself rather than to the square root of this voltage. To maintain an original deflection in an electrostatic tube with a given percentage increase of anode voltage would call for a much greater increase of deflecting voltage than the required increase of deflecting current in a magnetic coil.

Fig. 60-11 shows the manner in which deflection distance increases with reduction of anode voltage when deflecting current is maintained of constant peak-to-peak value in a certain air-core deflecting coil. The curve illustrates only the effect of anode voltage on deflection. We could not use the lower voltages because

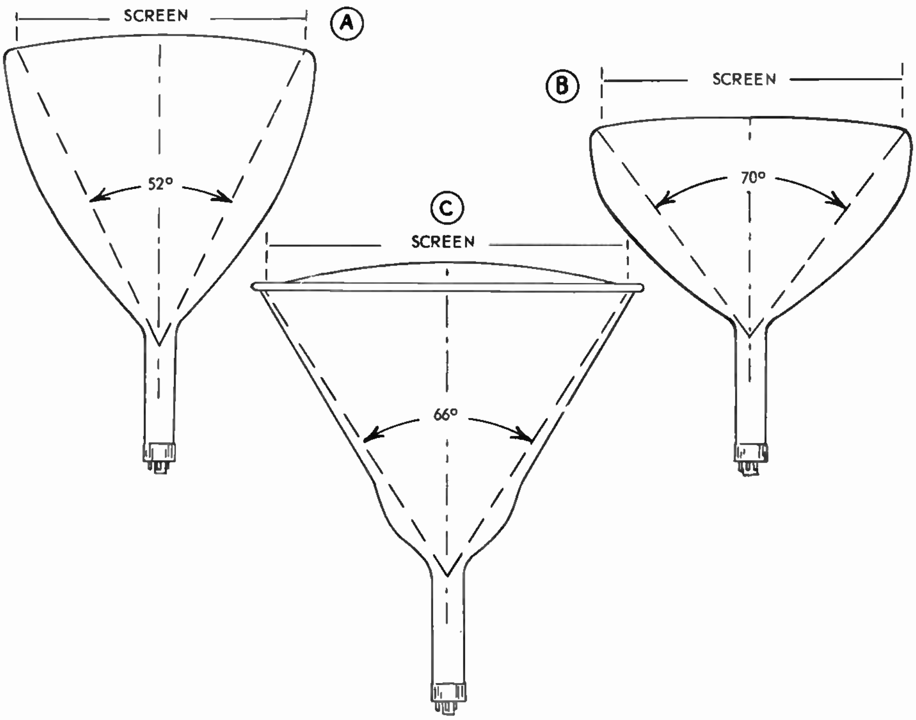


Fig. 60-12. Deflection angles for round picture tubes.

brightness would be so low and definition so poor as to produce no useful pictures.

DEFLECTION FACTORS. A deflection factor is the electrostatic deflecting voltage or the deflecting magnetic field strength required to produce deflection through a specified distance, usually one inch or one millimeter. A deflection factor is the reciprocal of a corresponding deflection sensitivity. That is, you divide 1 by the sensitivity to find the factor, or divide 1 by the factor to find the sensitivity.

Earlier in this lesson were mentioned some deflection sensitivities, one of which was 0.008 inch per volt. Dividing 1 by 0.008 gives 125, which is the corresponding deflection factor in volts per inch of deflection.

DEFLECTION ANGLES. The deflection angle is the number of degrees through which the electron beam must be swept to cover the raster area or the useful area of the screen. Angles for magnetic deflection picture tubes now in use range all the way from 50 degrees to 90 degrees. With 50-degree deflection, the beam is swept 25 degrees each way from a line through the lengthwise axis of the tube, and with 90-degree deflection, the sweep is 45 degrees each way from center.

Fig. 60-12 illustrates deflection angles for round tubes. At A are shown the proportions of a tube requiring a 52-degree deflection angle. At B is another tube in which the angle is 70 degrees. Screen diameters or useful diameters are almost exactly the same on both these tubes, but the flare on the tube at A is much longer than on the one at B and, of course, the first tube has greater overall length than the second one. The lengths of the necks back of the flares are the same for both tubes.

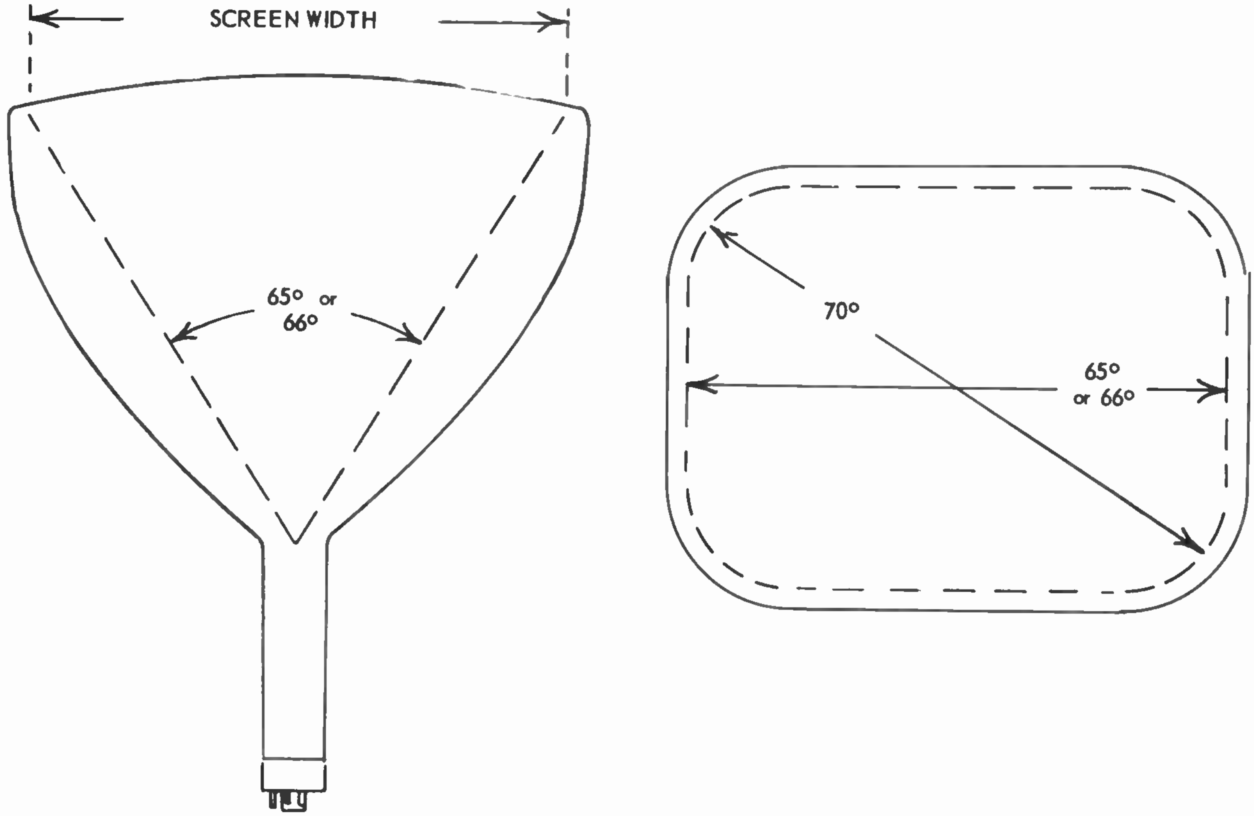


Fig. 60-13. Deflection angles for rectangular picture tubes.

For any given useful diameter or screen diameter, the less the overall length of the tube the greater will be the deflection angle. The greater the sweep of the beam or the greater the deflection angle the stronger must be the deflecting magnetic field produced by the yoke coil. It follows that relatively short tubes of any given screen diameter require stronger deflecting fields that longer tubes with the same screen diameter.

The deflection angle depends on screen size as well as on tube length. At C in Fig. 60-12 are shown the proportions of a tube of 19-inch nominal diameter which is of about the same overall length as the 16-inch tube at A. Because the screen diameter is greater at C than at A the deflection angle is 66 degrees instead of 52 degrees.

Fig. 60-13 shows one set of deflection angles for rectangular tubes, either all glass or metal and glass types. If we consider the width of the screen in our illustration, from left to right, we obtain a deflection angle of 65 or 66 degrees. This is referred to as the horizontal deflection angle. If we consider the maximum combined horizontal and vertical deflection with the electron beam at one of the corners, the deflection angle will be 70 degrees. This is referred to as the diagonal deflection angle. The deflection angle for rectangular tubes is sometimes given either way. However, the trend is towards the figure giving the diagonal angle. This is illustrated by the deflection angles listed for various tubes in Fig. 60-14. Tubes listed as having a diagonal angle of 90 degrees have a horizontal angle of 85 degrees.

TYPE	ROUND OR RECT.	METAL OR GLASS	DIAGONAL ANGLE DEGREES
21AP4	□	M	70
21ACP4	□	G	90
21ACP4A	□	G	90
21AFP4	□	G	70
21ALP4	□	G	90
21ALP4A	□	G	90
21ALP4A/B	□	G	90
21AMP4	□	G	90
21AMP4A	□	G	90
21ANP4	□	G	90
21ANP4A	□	G	90
21AQP4	□	G	90
21AQP4A	□	G	90
21ARP4	□	G	70
21ARP4A	□	G	70
21ASP4	□	G	70
21ATP4	□	G	90
21AUP4	□	G	72
21AUP4A	□	G	72
21AUP4A/B	□	G	72
21AVP4	□	G	72
21AVP4A	□	G	72
21AVP4A/B	□	G	72
21AWP4	□	G	72
21AYP4	□	G	70
21DP4	□	M	70
21EP4	□	G	70
21EP4A	□	G	70
21EP4B	□	G	70
21FP4	□	G	70
21FP4A	□	G	70
21FP4C	□	G	70
21JP4	□	G	70
21JP4A	□	G	70
21KP4	□	G	70
21KP4A	□	G	70

TYPE	ROUND OR RECT.	METAL OR GLASS	DIAGONAL ANGLE DEGREES
21MP4	□	M	70
21WP4	□	G	70
21WP4A	□	G	70
21XP4	□	G	70
21XP4A	□	G	70
21YP4	□	G	70
21YP4A	□	G	70
21ZP4	□	G	70
21ZP4A	□	G	70
21ZP4B	□	G	70
22AP4	○	M	70
22AP4A	○	M	70
24AP4	○	M	70
24AP4A	○	M	70
24AP4B	○	M	70
24BP4	○	M	70
24CP4	□	G	90
24CP4A	□	G	90
24DP4	□	G	90
24DP4A	□	G	90
24QP4	□	G	90
24TP4	□	G	90
24VP4	□	G	90
24VP4A	□	G	90
24XP4	□	G	90
24YP4	□	G	90
27AP4	□	M	90
27EP4	□	G	90
27GP4	□	G	90
27LP4	□	G	90
27MP4	□	M	90
27NP4	□	G	90
27RP4	□	G	90
27SP4	□	G	90
27UP4	□	G	90
30BP4	○	M	90

The chart in Fig. 60-14 illustrates the variations in tube construction, starting with a 21-inch and going to a 30-inch picture tube. The column after Type indicates if the tube is square by the square symbol and round by the circle. The next column indicates whether the tube is all glass "G", or if it utilizes metal "M", for the cone portion of the tube. The last column shows the diagonal angle in degrees for each of the tubes.

61 → 65

COMMERCIAL TRADES INSTITUTE

KNOWLEDGE + AMBITION = SUCCESS

LEARN . . .

Training Your Mind

C. T. I.

Perfecting Your Skill

ACT . . .

"The heights by great men reached and kept
Were not attained in sudden flight,
But they, while their companions slept,
Were climbing upward in the night."

ENJOY . . .

COMMERCIAL TRADES INSTITUTE

CHICAGO, ILLINOIS

TELEVISION

LESSON No. 61

PICTURE TUBE ACCESSORIES

Training Your Mind

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

COMMERCIAL TRADES INSTITUTE

Training Your Mind

C.T.I.

Perfecting Your Skill

Chicago, Ill.

LESSON NO. 61
PICTURE TUBE ACCESSORIES

Many parts essential to the operation of magnetic picture tubes are on the outside of the tube, as shown by Fig. 61-1. Commencing at the base end we may note first that the picture tube socket does not support the tube, rather the tube supports the socket. The tube is mounted and supported at the forward end of the flare or cone by some suitable mechanical structure and is given additional support by the bracket that carries the deflecting yoke and the focus magnet or coil. The socket merely presses onto the base pins. Flexible leads extend from the chassis circuits into the socket, where connections of these leads to the contact lugs for the pins are enclosed and insulated by a cap over the outer end of the socket.

96 Between the base and the focus coil or magnet on most magnetic deflection tubes is an ion trap magnet. This magnet, and the remainder of the trap which is part of the electron gun structure, removes ions from the electron beam. Were the ions allowed to reach the screen coating they would cause a brown spot, called an ion burn. Even though you set the trap to catch the ions, wrong adjustment can cause trouble with focusing, centering, brilliance, and easily may cause such permanent internal damage as to make the tube quite useless. The trapping of ions will come in for some rather lengthy explanation later in this lesson.

Moving forward along the picture tube neck we come next to the focus coil or magnet. The principles of focusing by means of a magnetic field already have been studied. There remain to be examined, however, quite a few focusing systems which utilize permanent magnets and magnetic shunting. This will be another subject for the present lesson.

The front end of the permanent magnet focusing structure in a few designs is mounted right up against the rear end of the deflecting yoke, and in other designs may be anywhere between 1/8 and 3/8 inch back of the yoke. The position selected is that which allows the best compromise between sharp focusing at the center and at the edges of the screen, just as with focus coils or with combined coils and permanent magnets.

Between the support for the deflecting yoke and the glass flare on all types of picture tubes you will see a cushion consisting of a rubber or plastic ring. In order to have completely satisfactory deflection and focusing it is essential that the flare of the tube be snugly seated against this cushion and that the deflecting yoke be just as far forward as the cushion will allow. The yoke supports should have enough adjustment to allow this.

7 If the picture tube is all glass and has an external conductive coating on the flare this coating must be grounded to the chassis. One or more grounding contacts usually have the form of long wire loops or leaf type springs which are fastened to the yoke bracket and rest firmly on the tube coating.

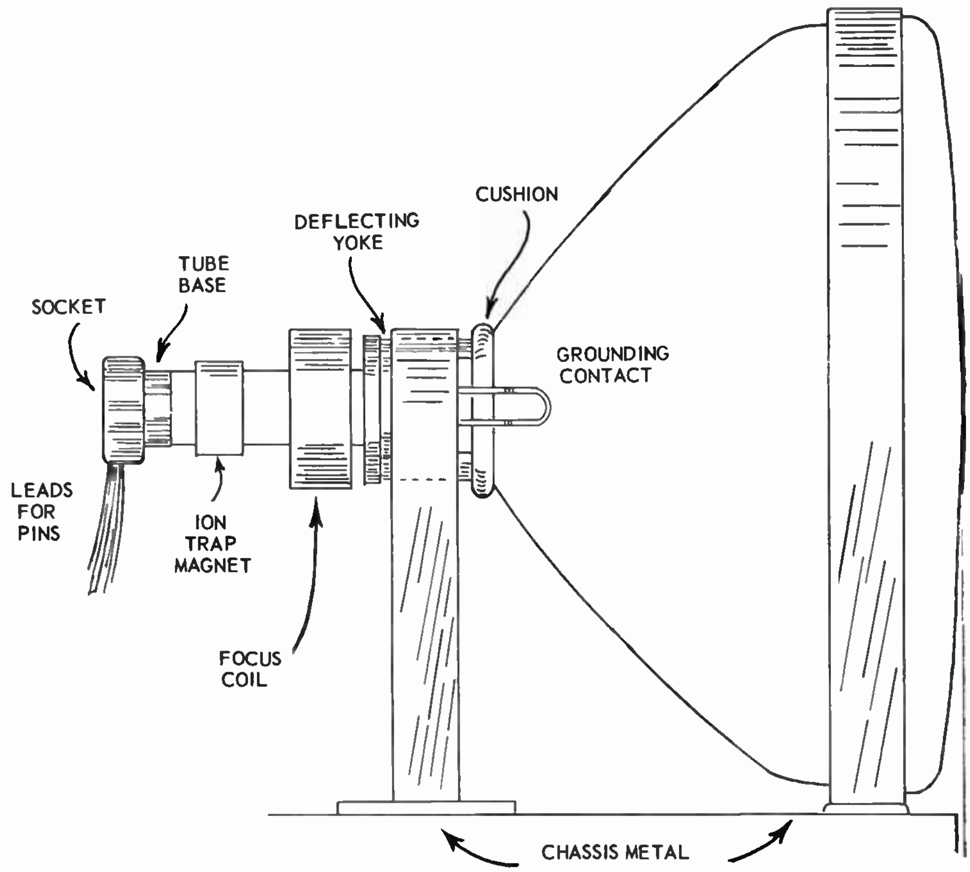


Fig. 61-1. The parts which are mounted on or around the neck of most picture tubes.

ION TRAPS. The ions which are to be trapped out of the electron beam before it reaches the screen are negatively charged molecules of gases which have remained when the picture tube was evacuated. It is possible for a free negative electron to attach itself to a neutral molecule, thereby giving a net negative charge to the molecule – which then is called a negative ion.

(a) Before talking about separation of ions from electrons we should review a few simple facts on which the action depends. First, when ions and electrons pass through an electrostatic or electric field at an angle to the direction of field lines both the ions and the electrons are equally deflected. In a tube which employs electrostatic deflection the ions thus are distributed all over the screen, just as are the electrons. There is no concentration of ion bombardment on any one area of the screen, and because there are relatively few ions compared to electrons no damage results. Consequently, electrostatic deflection tubes do not require ion traps.

The second important fact is this: When electrons and ions pass through a magnetic field at an angle to the field lines the deflection is proportional to the square root of the mass or weight of the particles. An ion contains all the parts of a complete atom plus an electron, and its mass is at least 1,870 times as great

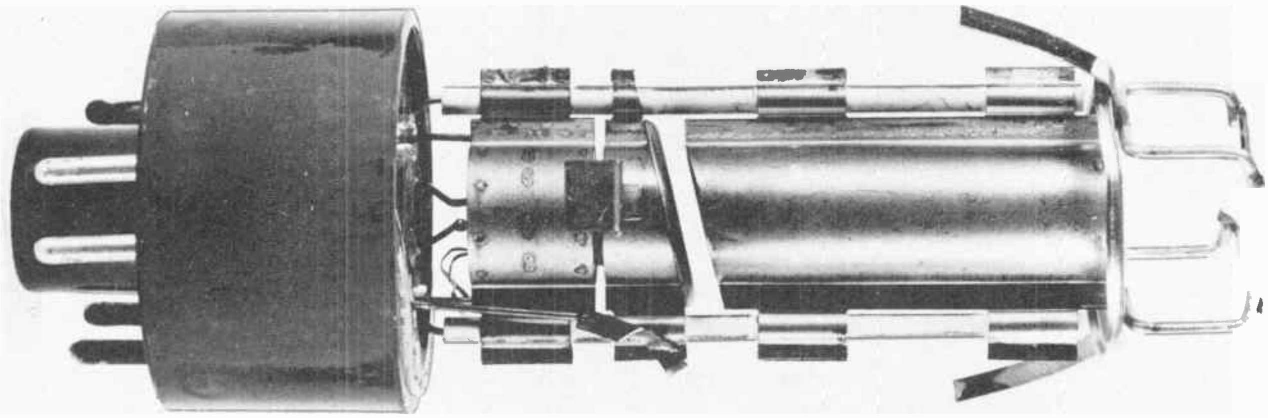


Fig. 61-2. An electron gun in which is an angular gap forming part of the ion trap.

as that of an electron. The square root of 1,870 is approximately 43, so the magnetic force deflects an ion only about 1/43 as far as it deflects an electron. By using a suitable combination of electric and magnetic fields we may form a trap which takes the ions out of the electron stream and holds them in the gun of the picture tube instead of allowing them to proceed to the screen.

The ion trap which first came into common use, and which still is generally employed, consists of an angular gap in the electron gun and of two magnets placed around the tube neck near the gap. This angular

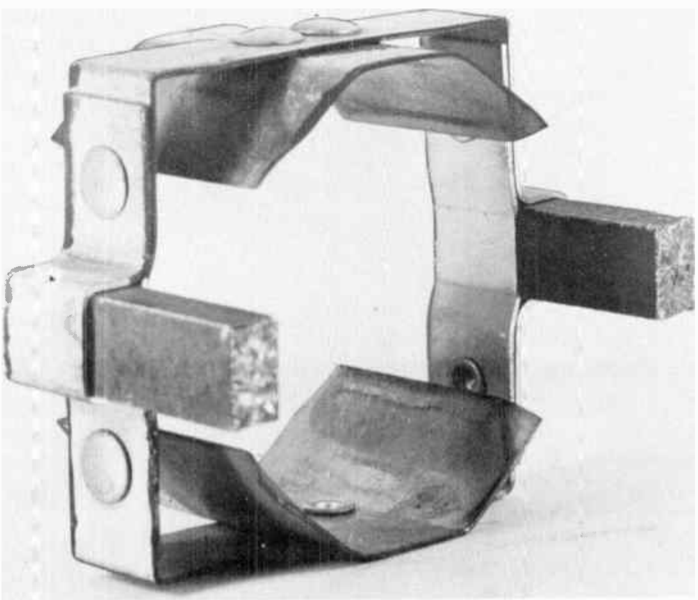


Fig. 61-3. Two ion trap magnets in a mounting that holds them on opposite sides of the picture tube neck.

gap is clearly visible in Fig. 61-2, which is a picture of the electron gun from a tube designed for magnetic deflection and focusing. The oblique gap is between the second grid and the anode. Potential on the second grid is 250 to 300 volts in most cases, and on the anode it is several thousand of volts. This great potential difference causes a strong electric field to be formed between these two electrodes, with the lines of the field running perpendicular to or square to the angular sides of the gap opening.

There are many styles of ion trap magnets which furnish the two magnetic fields required for operation with the angular gap as pictured. One of the simplest magnet structures is illustrated by Fig. 61-3. Two small but very strong bar magnets are held in a rectangular frame which slips onto the neck of the picture tube, so that the two magnets are parallel with the tube axis and are on opposite sides of the electron gun containing the gap. Spring bronze pieces attached inside the magnet frame hold it in position by friction on the glass of the neck.

Looking toward the electron gun from one side of the tube the magnet nearest you would be in the position shown at 1 in Fig. 61-4. Looking down on top of the tube the two magnets would be on opposite sides of the diagonal gap, as shown by diagram 2. North and south poles of the two magnets are opposite each

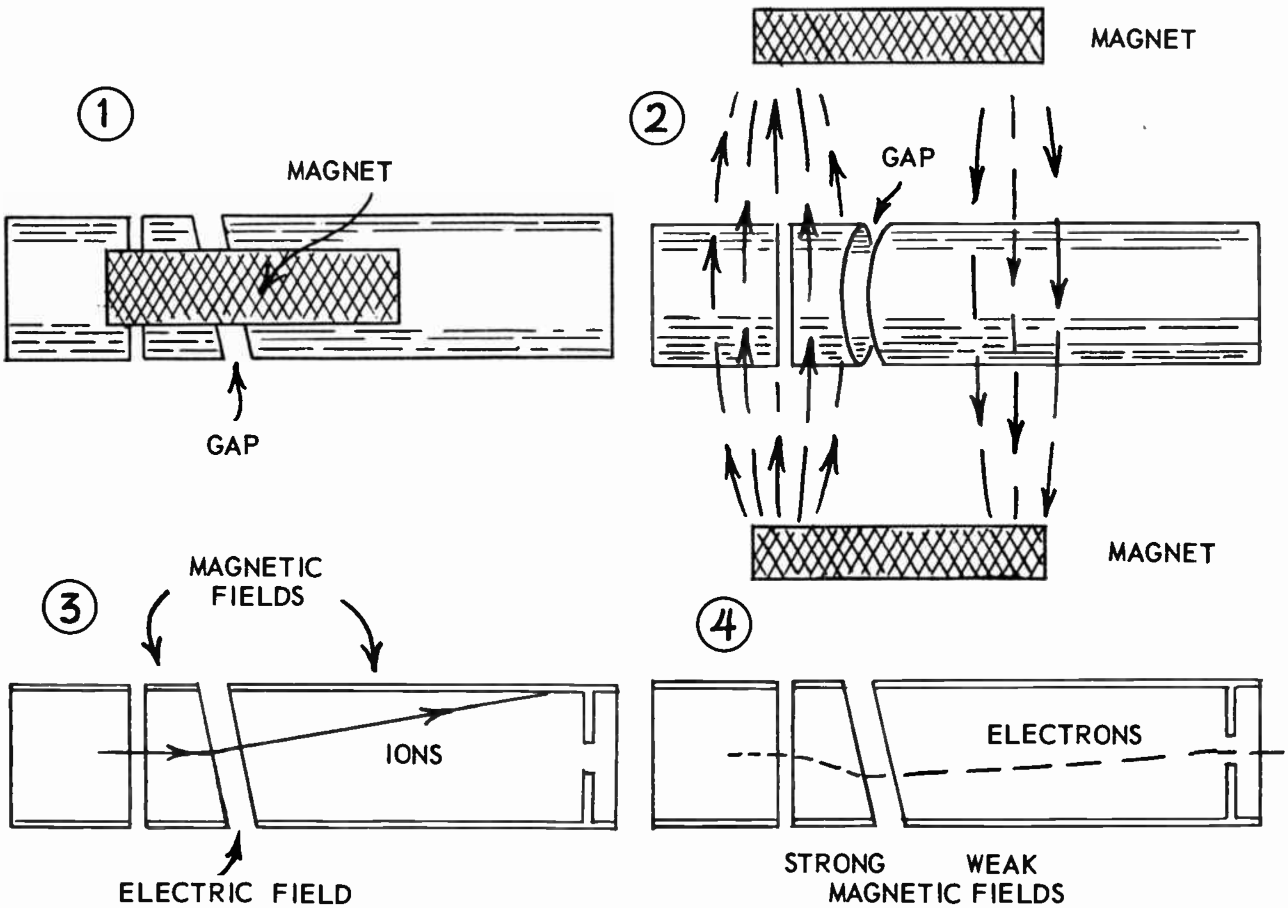


Fig. 61-4. How ions and electrons are deflected by magnets and electric fields of an ion trap.

other, so that the lines of force are in one direction through the magnetic fieldthat is toward the tube base and are in the opposite direction through the field that is toward the screen.

The magnetic field toward the tube base is much stronger than the one toward the screen. In the structure of Fig. 61-3 this difference of strength is secured by making the supporting frame that carries one end of each magnet out of iron or soft steel. Then part of the lines of force go through the frame and only the remainder go through the space in which is the electron gun of the picture tube. The free ends of the magnets, between which is the stronger field, are placed toward the base of the tube.

Ions passing from the tube cathode space into the electron gun are forced to follow the path shown at 3 in Fig. 61-4. The ions are deflected by only negligible amounts in the two magnetic fields. But in passing through the electric field of the diagonal gap the ions are deflected in line with the electric lines of force and are turned so that they strike a point inside the gun structure instead of going on to the screen.

As shown by diagram 4, the electrons are deflected downward by the relatively strong magnetic field that is toward the tube base. The extremely small mass or weight of the electrons allows their velocity to be far greater than that of the heavy ions, and the electrons go through the electric field of the diagonal gap without too much deflection. Then the forward magnetic field brings the electrons up just enough so that they pass through the small opening in the front end of the anode. If you refer to the lesson in which we studied the principles of magnetic deflection of the electron beam you can use the information given there to check the directions of deflection shown at 4 in Fig. 61-4. Keep in mind that the beam is traveling from left to right and that magnetic lines of force extend away from you in diagrams 1 and 4.

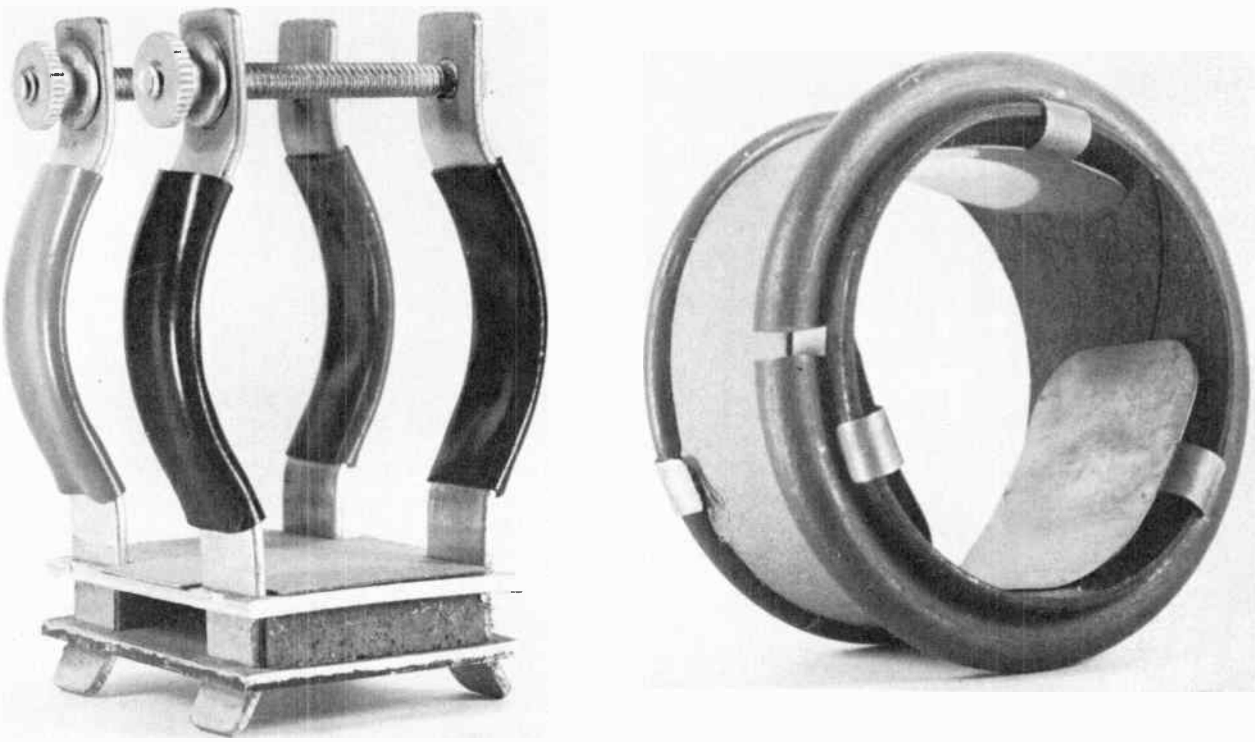


Fig. 61-5. Two styles of double magnets for ion traps.

Two of the many other types of double magnets for ion traps are pictured by Fig. 61-5. In the design at the left the two magnets are clamped between pieces of flat non-magnetic metal at the bottom of the structure. From the opposite ends of each magnet curved pole pieces extend upward and pass around opposite sides of the picture tube neck. This unit is clamped on the tube neck by tightening the nuts on two screws that pass through the upper ends of the pole pieces. These screws are of non-magnetic metal so that they do not partially short circuit or shunt the magnetic fields. One of the magnets is larger and stronger than the other. It is placed toward the base end of the picture tube. The stronger and weaker magnetic fields which extend from pole to pole pass through the electron gun.

In the design pictured at the right in Fig. 61-5 there are two ring-shaped magnets carried on a fibre sleeve that fits around the neck of the picture tube. Inside the fibre are spring bronze plates that hold the unit in position by friction. At one point around the circumference of each ring is an air gap. The two ends of the ring on both sides of the gap are of the same magnetic polarity, forming, in effect a single pole. At a point directly opposite the gap is a "consequent" pole of the opposite polarity. The field lines extend across the center opening of the ring from the gap to the opposite consequent pole, consequently pass through the space occupied by the electron gun.

The larger of the two rings, which provides the stronger magnetic field, is placed toward the base end of

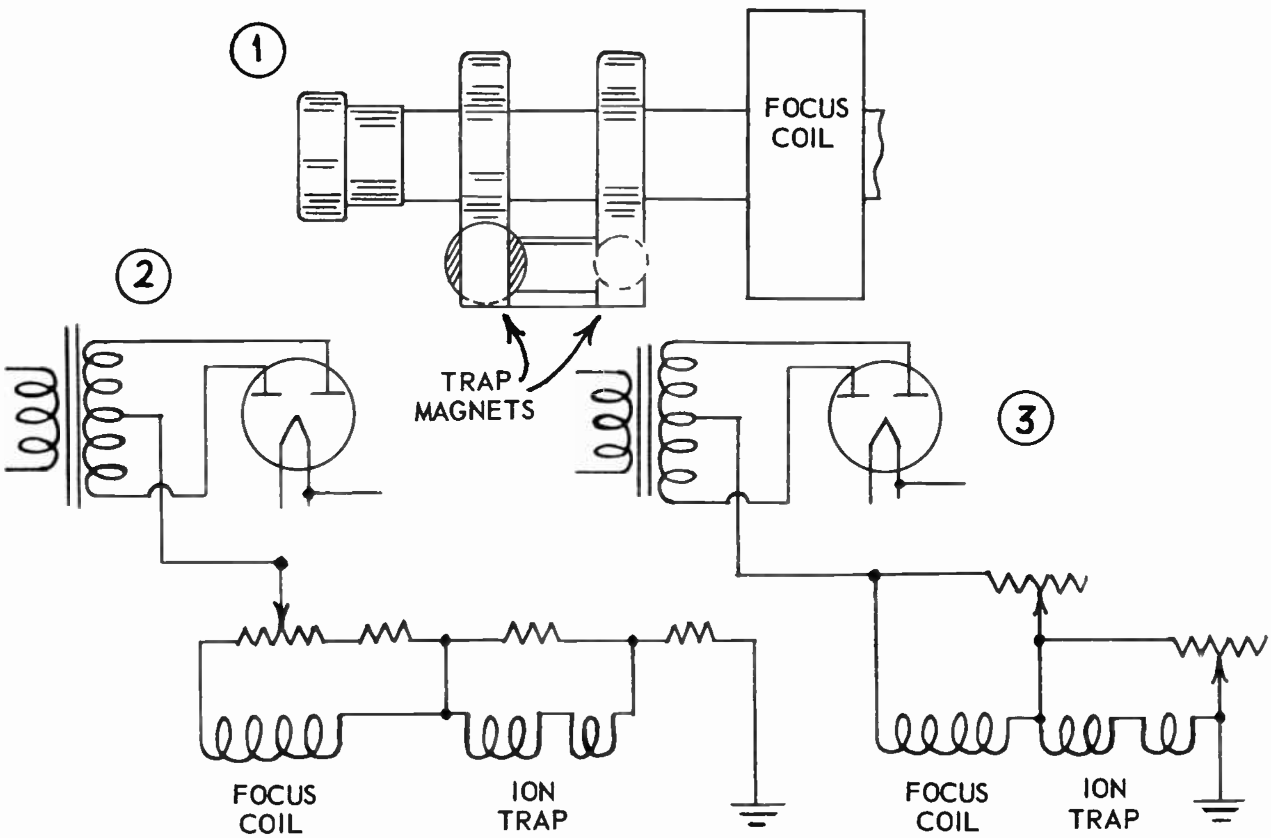


Fig. 61-6. Wound coils used as ion trap magnets, and how they may be connected into the power supply circuit.

the picture tube, with the smaller ring and weaker field toward the screen. The gap in the small ring is half way around from the gap in the large ring. The rear magnetic field and front magnetic field are of opposite polarity, as with all double magnets.

5 In many of the earlier television receivers using magnetic picture tubes the ion trap is operated with two electromagnets instead of two permanent magnets. As shown at 1 in Fig. 61-6 the larger and stronger electromagnet is placed toward the tube base, with the smaller and weaker electromagnet toward the screen. The illustration shows pole pieces extending upward around the tube neck from the ends of the two electromagnet coils in about the same manner that they extend up from the permanent magnets at the left in Fig. 61-5.

The windings for the two electromagnets are connected in series with each other, and usually are in the same part of the power supply circuit as the focus coil. Two methods of connection are shown in Fig. 61-6. In both cases the focus coil and the trap magnets are in series and both are shunted by resistors. At 2 there is no adjustment for trap magnet current, this being the usual method of operation. At 3 there are service adjustments for focus coil current and for trap magnet current.

INCLINED GUN ION TRAPS. In some picture tubes the base end of the electron gun is offset and the lengthwise axis of the gun is inclined with reference to the axis of the tube neck. This construction is illustrated by Fig. 61-7. After the electrons and ions meet the electric field lines in the diagonal gap the heavy ions proceed practically in line with this field until they strike against the inside of the anode. After leaving the gap the electrons are bent downward by a magnetic field produced by a single external ion trap magnet. Correct positioning of this magnet causes the electron beam to pass through the small opening at the forward end of the anode cylinder.

BENT GUN ION TRAP. In the bent gun design of ion trap, illustrated by Fig. 61-8, the end of the electron gun toward the tube base is inclined with reference to the neck axis while the forward part of the gun is coaxial with or is lined up with the neck axis. The angle of inclination of the grid end of the electron gun "aims" both the electrons and the ions toward a point inside the forward end of the anode. Around the outside of the tube neck, approximately in the position marked, is placed a single permanent magnet with its north pole toward you when looking at the gun from the position indicated in the drawing.

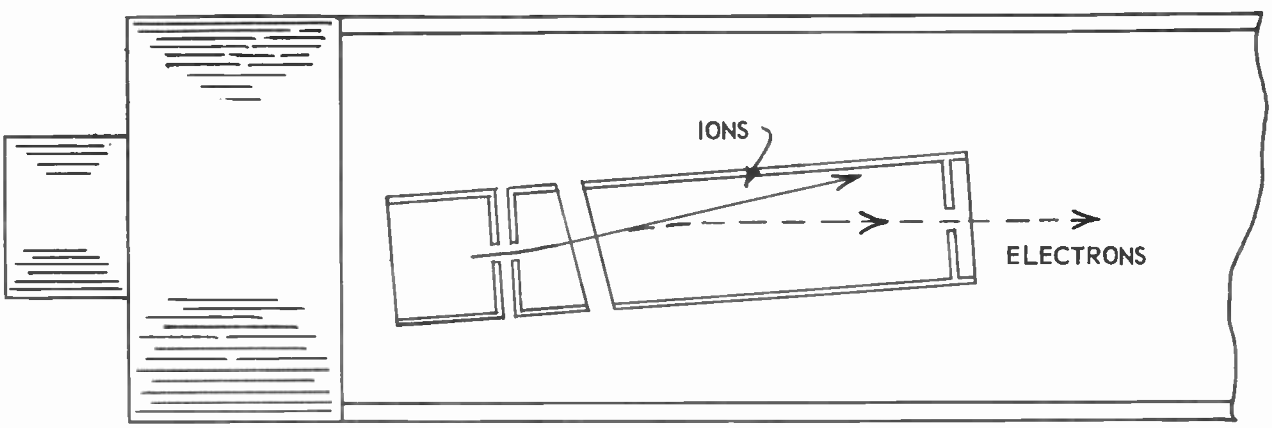


Fig. 61-7. Ion and electron paths in an inclined gun ion trap.

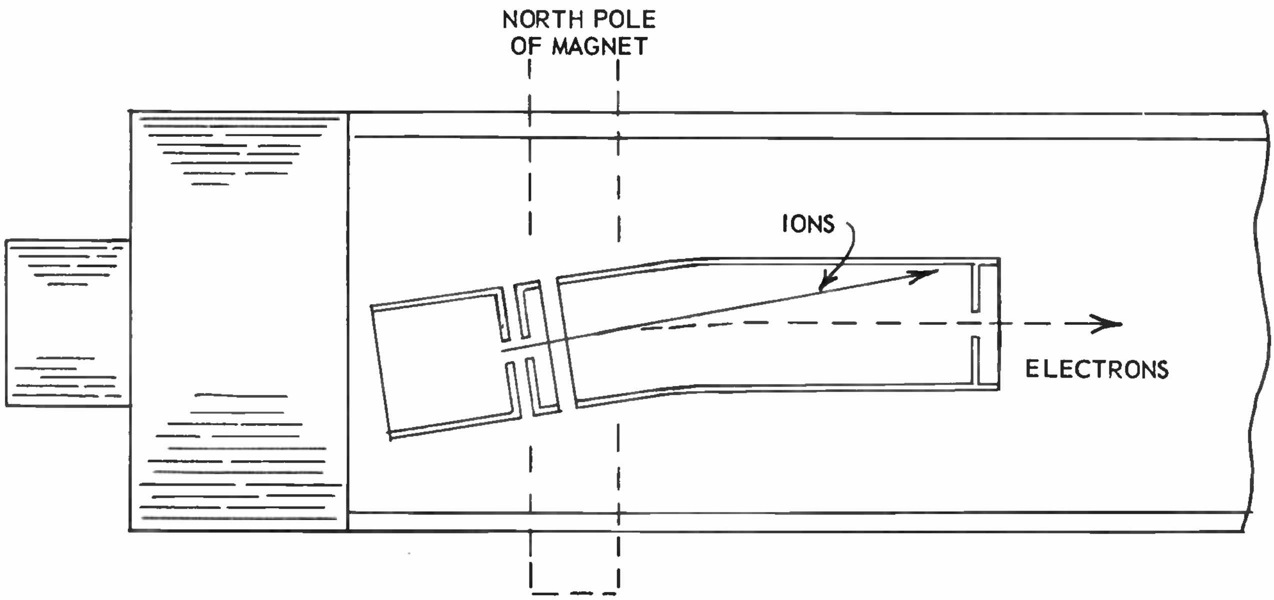


Fig. 61-8. Paths of ions and electrons in a bent gun ion trap.

The ions travel forward along the electrostatic field lines formed between the second grid and the anode and since they are deflected hardly at all by the field of the permanent magnet the ions are trapped within the gun. The magnetic field deflects the electrons after they have passed through the gap, bringing the electron stream down just far enough so that it passes through the small opening at the forward end of the anode.

Single magnets for any tubes having ion traps that require only a single element are constructed in much the same mechanical fashion as double magnets. At the left in Fig. 61-9 is a double magnet and at the right is a single magnet of the corresponding style. Single magnets are made also like half of the double type pictured at the left in Fig. 61-5 as well as in many other styles which serve to hold the magnet and its pole pieces permanently in position after they are correctly adjusted.

POSITIONS FOR ION TRAP MAGNETS. To the electron guns of many picture tubes are attached two small metal "flags" which help identify the point along the tube neck at which you should place the poles of the stronger magnet in a two-magnet unit or of a single magnet when the trap requires only one magnetic field. In Fig. 61-10 you can see the two flags attached to the second grid and extending over the space between first and second grids. The two magnet poles are to be placed opposite the two flags and approximately in line with the flags along the length of the neck. This is the preliminary settingto be made before proceeding with an exact adjustment of magnet position.

If there is an arrow marked or stamped anywhere on the magnet structure this arrow is to point forward, toward the screen and away from the base end of the tube.

(10) Many ion trap magnets have sleeves of colored fabric or plastic over the extended magnetic pole pieces. You can see such sleeves on the pole pieces in Fig. 61-5 at the left. When one pair of sleeves is black

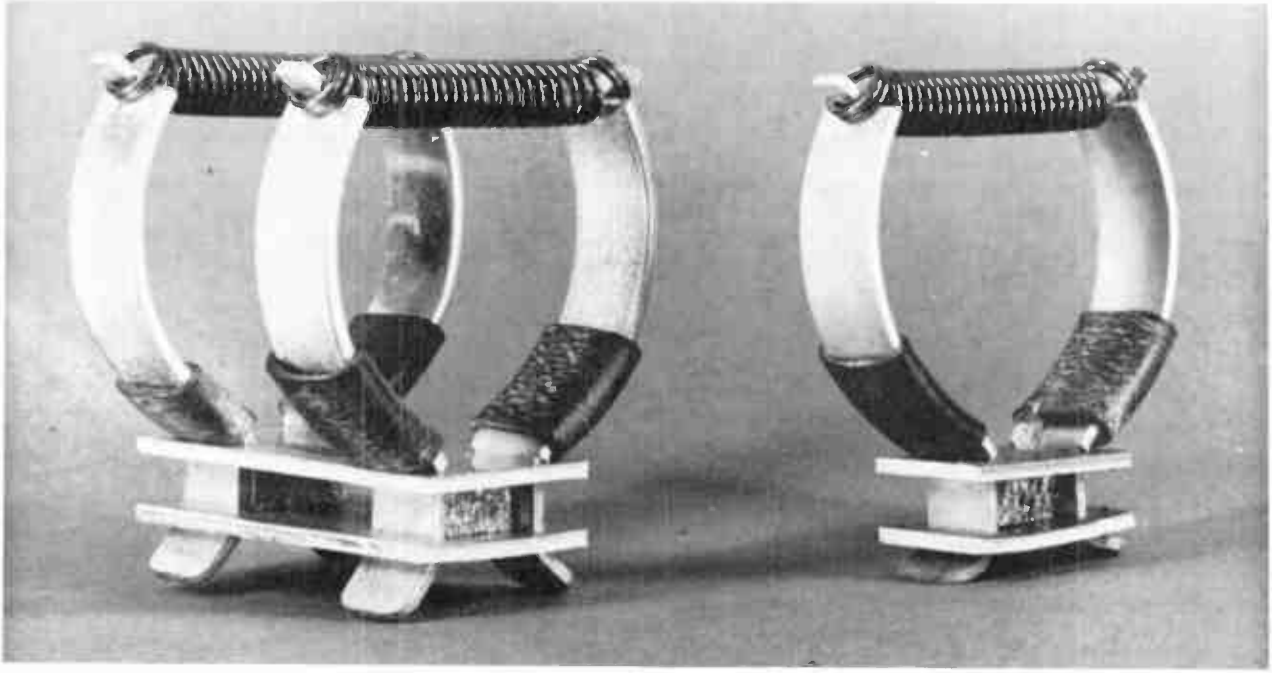


Fig. 61-9. Double and single ion trap magnets of similar mechanical design.

this end of the magnet structure goes toward the base of the picture tube. The other sleeves, which usually are blue or red, go toward the screen end of the picture tube.

When there are two magnets the stronger one always goes toward the base of the picture tube and the weaker one toward the front or screen end. This applies to electromagnets and to permanent magnets alike.

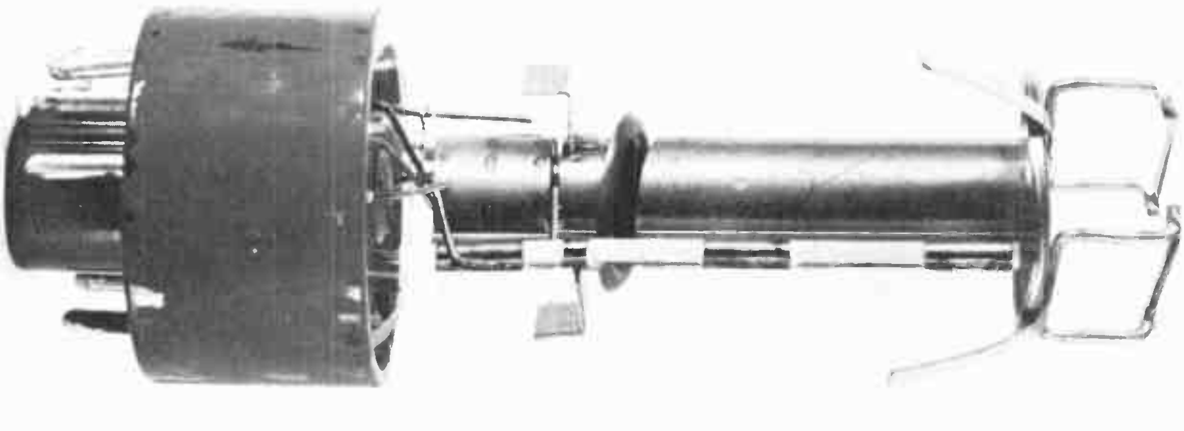


Fig. 61-10. The flags on an electron gun which help in correct placing of the ion trap magnet.

The stronger of two electromagnets is the one with the larger cylindrical form and housing. The larger and stronger ring-shaped magnet is easily identified on the unit illustrated at the right in Fig. 61-5. The free or extended ends of the magnets in Fig. 61-3 have the stronger fields, so they go toward the base end of the picture tube.

Units which contain one or two small magnets, with extended pole pieces, usually must be mounted with the magnets on the bottom or chassis side of the tube neck, this being true when the anode connector of an all-glass picture tube is on top. If the tube is turned to some other position, as often is the case, the magnets may have to be either above or at one side of the neck.

You should note that our explanations of the directions in which electrons and ions are deflected assume certain positions of the electron guns with reference to diagonal slot inclination or to the slope of tilted gun or bent gun types. If the picture tube is turned to some other position the field polarities and directions of deflections will change accordingly. This would not change any of the actions occuring in the tube, but you might have to look at a diagram upside down or from the back of the page on which it is printed.

The magnet positions which have been mentioned are merely the preliminary settings to be made before you undertake the process of adjustment. There are some single-magnet units on which there are no markings to indicate how they should be turned or positioned for an initial setting. If you cannot get any brightness on the raster after going through the steps for adjustment it is probable that the magnet must be turned around, front and back.

ION TRAP ADJUSTMENT. Following are the steps in making a correct adjustment of an ion trap magnet. After you do this kind of work a few times the whole job will take far less time than needed to read these instructions.

1. Check the initial position of the trap magnet as explained in preceding paragraphs.
2. Examine the deflecting yoke to see that it is forward against the flare cushion. Make sure that the picture tube flare is back against the cushion.
3. See that the focus coil or magnet has its center on the axis of the tube neck.
4. Turn on the receiver, set the tuning to a channel where there is no transmission, and view the raster.
5. If centering is incorrect adjust the electrical centering controls if they are used, or adjust the focus coil or magnet for correct centering.
6. Set the brightness control for the least brightness that allows a clearly visible raster. Never use a bright raster before adjusting the ion trap magnet. A high-intensity electron beam striking the edge of the front opening in the anode can permanently damage the picture tube in less than a half minute.
7. If the ion trap magnet is held with screw clamps loosen them just enough to allow sliding and rotating the magnet, but not so much that the magnet will not remain where you place it.
8. Rotate the magnet through a small angle around the tube neck, and at the same time move it slightly backward and forward along the neck to find the position at which the raster becomes brightest. You may find two positions, one forward and one back, at which the raster appears equally bright. Always place the magnet in the rearmost of these two positions, nearer the tube base. The forward position will make difficult centering and focusing, and may cause permanent damage to the picture tube.
9. Reduce the brightness to a value somewhat less than you would use for picture observation and carefully adjust the focus control for the best compromise between sharpness at the center and edges of the raster.
10. Repeat step 8 to again obtain greatest possible brightness – without changing the brightness control. It is absolutely necessary that the ion trap magnet be placed for maximum brightness with any setting of the brightness control, otherwise the electron beam is striking the edges of the anode opening rather than coming through.

11. Increase the brightness control until the distinct line focusing commences to disappear, or until the raster just commences to "bloom", which means to become both higher and wider. Again repeat step 8 to obtain maximum brightness.
12. Readjust the focusing control for the most distinct lines with the brightness control turned high, as in the preceding step.

Shadows may appear at the edges or corners of the raster when the ion trap magnet is positioned for maximum brightness. These shadows should not be removed by shifting the magnet if this reduces the brightness. The shadows will not remain if you correctly follow the first five steps of the preceding adjustment instructions, and if the trap magnet is of normal strength. If you cannot make the adjustments as described, or if the trap magnet has to be moved more than about one-quarter inch from the flags on an electron gun, the chances are that the magnet is weak and should be replaced. Some double-ring magnets, generally similar to the one in Fig. 61-5 at the right, have been made with a small front ring that may be rotated independently of the larger ring. Shadowing will result if the small ring is out of its correct position.

TUBES WITHOUT ION TRAPS. Some picture tubes have an exceedingly thin layer of metallic aluminum over the inside of the screen and do not require ion traps. The heavy slow-moving ions cannot penetrate through the aluminum to damage the screen, but the almost weightless high-velocity electrons go through this added coating and excite the phosphor as usual. Some of the tubes in this class include types 10DP4, 10FP4, 10MP4, and 12KP4.

Other tubes which do not have to have ion traps include types which are pumped to an extra high vacuum, leaving relatively few gas molecules to form ions. Some of the larger screen tubes are in this class. The greater the deflection distances the less concentrated is the ion bombardment over the central area, and the less is the likelihood of ion burns.

FOCUSING AND CENTERING WITH PERMANENT MAGNETS. Focusing and centering by means of adjustable permanent magnets have become rather common practices in many television receivers of recent design. Permanent magnets for these two functions were mentioned briefly when we studied the general principles of focusing by means of magnetic fields, but it is desirable that we become acquainted with the action, mechanical details, and adjustment of these devices.

Focusing of the electron beam is brought about by changing the strength of the magnetic focusing field. With an electromagnetic focus coil this change results from varying the direct current in the coil. When using permanent magnets for focusing we vary their field strength by means of adjustable magnetic shunts of one kind or another. A magnetic shunt consists of one or more pieces of iron or soft steel placed at or near the poles of a permanent magnet, or any other magnet, to allow a greater or less proportion of the field lines to pass through the shunt instead of through the external field space. The strength of the free field in the space between magnet poles may be varied by moving the shunt.

Fig. 61-11 illustrates some principles of magnetic shunting as employed for focusing. At 1 is shown a cylindrical permanent magnet with partially closed ends through which would extend the neck of the picture tube. The two magnetic poles are at the front and rear openings, and the magnetic field lines between poles extend as shown, so that the direction of this field is the same as that of a field formed with a focusing coil.

In diagram 2 the magnetic shunt consists of an iron cylinder that may be moved lengthwise of the magnet axis through the opening in which is the picture tube neck. Moving this shunt farther into the magnet allows more and more of the magnetic circuit to be completed through the iron of the shunt, and the remaining field in the open space and through the tube neck becomes progressively smaller and weaker to alter the focusing of the electron beam.

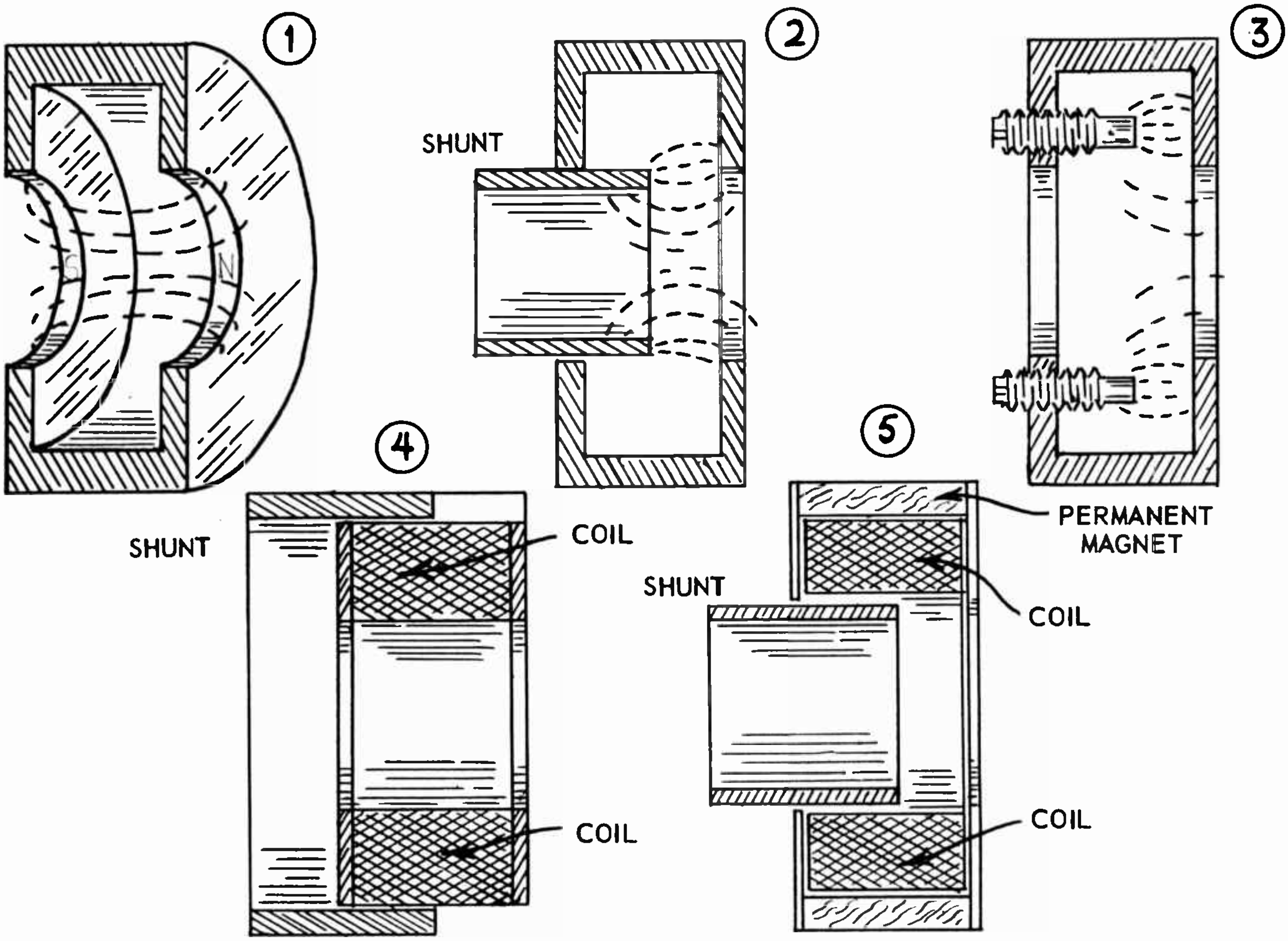


Fig. 61-11. Some of the principles employed for focusing with magnetic shunts.

At 3 the shunting effect is secured with iron or soft steel screws that may be turned farther into or out of the space between magnet poles. Turning one or both screws farther in increases the proportion of the magnetic circuit formed by iron and steel. This weakens the field remaining in the free space and passing through the neck of the picture tube.

Diagram 4 illustrates the principle of a magnetic shunt sometimes employed with an electromagnetic focus coil. The shunt consists of an iron or soft steel cylinder or ring that may be moved lengthwise of the coil axis around the outside of the coil. Moving this shunt ring farther onto the coil allows more and more of the magnetic lines of force to pass through this path, and the field that extends through the neck of the picture tube is weakened.

An internal cylindrical magnetic shunt is shown at 5. This type sometimes is found with a combination of electromagnetic coil and permanent magnet for focusing. As the shunt is moved farther into the coil it completes more of the magnetic circuit through itself, and weakens the magnetic field that passes through the tube neck.

Centering of the picture or pattern, when accomplished by tilting a focus coil, results from tilting the axis of the electromagnetic field with reference to the axis of the picture tube neck. This, in effect, aims an undeflected electron beam at one place or another near the center of the screen. While the beam is being deflected during production of a raster or picture the illuminated area on the screen is shifted up, down, or sideways by tilting of the coil and its field lines. Exactly the same effect may be had by tilting any style of permanent magnet which is used for focusing.

It is possible to keep the main body of the focusing magnet stationary while inclining its field axis by shifting a plate or ring which is part of the magnetic circuit. This principle is illustrated by Fig. 61-12. Diagram 1 shows a centering plate mounted at the front end or screen end of a focus magnet. The opening through the centering plate is somewhat smaller in diameter than the opening or openings which form the poles of the focus magnet, but is something like a quarter-inch larger than the outside diameter of the tube neck.

The centering plate is mounted or supported in some manner that allows it to be moved up, down, or to either side. If the plate is moved up, as in diagram 2, the axis of the magnetic field through the entire magnetic structure is tilted upward with reference to the tube axis. If the centering plate is moved down, as at 3, the magnetic axis is tilted downward with reference to the tube axis. If the plate is shifted toward either the right or left the magnetic axis will be correspondingly tilted sideways away from the tube axis.

It has been explained earlier that the picture or pattern will not move on the screen in the same direction that the magnetic field is tilted, whether the focusing magnet is a coil type or a permanent magnet type. The centering plate is shifted in whatever direction the desired direction of picture shift occurs.

There are many styles of focusing and centering devices which employ permanent magnets. To illustrate in a general way what you may expect to find in these units we shall examine in detail the one pictured by

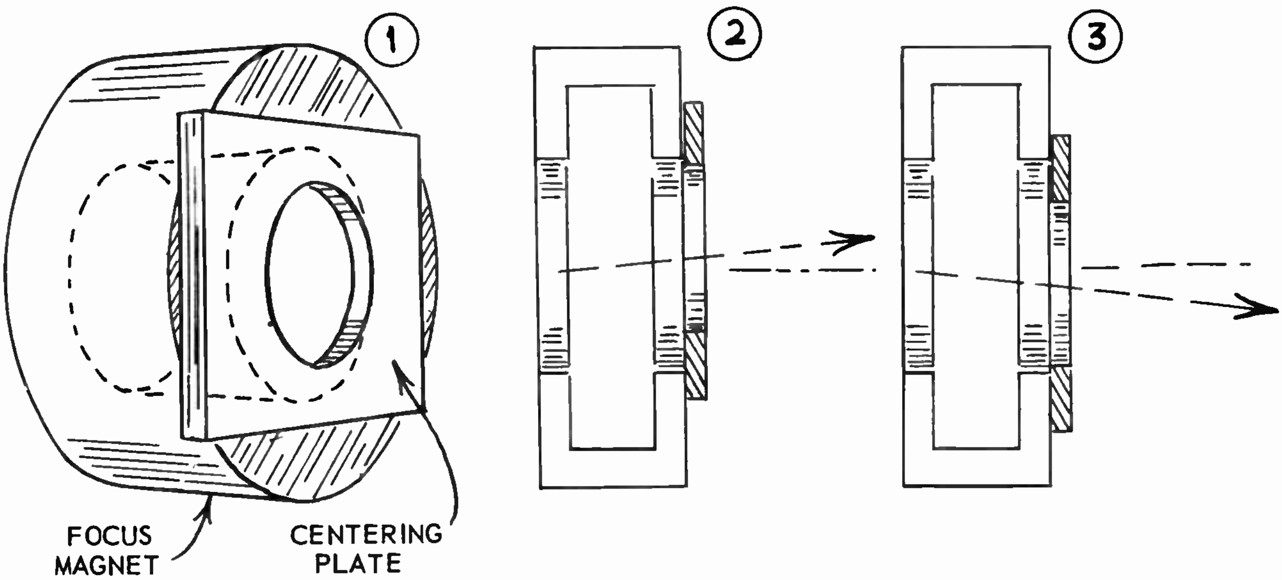


Fig. 61-12. The action of a centering plate mounted on a permanent magnet used for focusing.

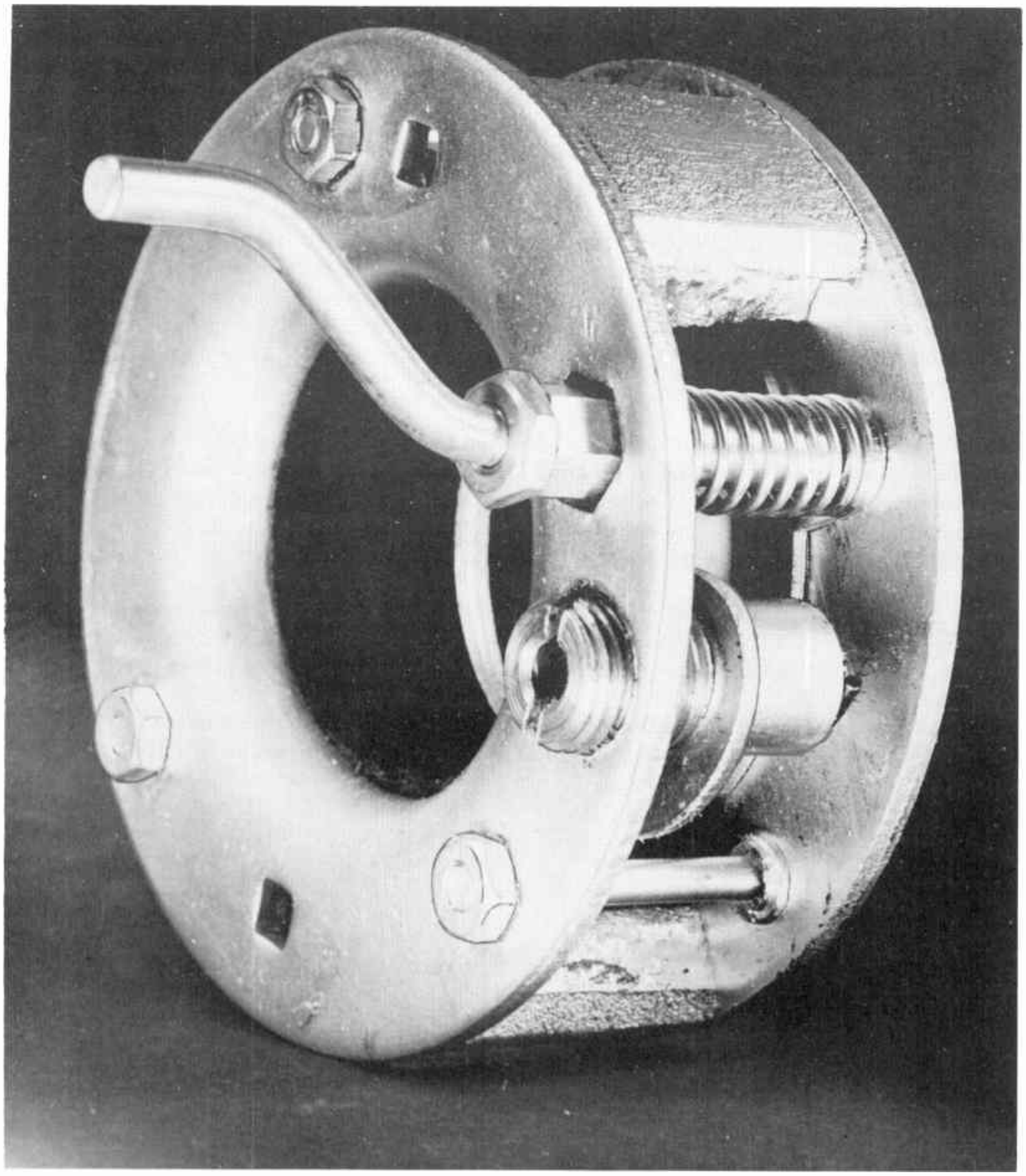


Fig. 61-13. One style of permanent-magnet focusing and centering device.

Fig. 61-13. This is a type used in a number of television receivers. The stationary magnetic structure consists of front and back circular plates between which are three equally spaced Alnico magnets. Focusing is adjusted by turning the large screw whose slotted end protrudes through the rear plate. This screw moves a shunting ring lengthwise of the internal openings. The centering adjustment is a curved handle which moves a plate to shift the axis of the focusing magnetic field.

Various structural details of this particular device are illustrated by the sectional drawings of Fig. 61-14. At the left is shown the manner in which the focusing screw moves the shunting sleeve or ring. The rear magnet plate is formed with an inward cylindrical extension that goes around the neck of the picture tube. This extension is lined with felt to protect the glass of the tube. The focusing sleeve is around the outside of the extension, allowing the metal portion of the magnetic circuit to pass from the extension through the movable sleeve. The sleeve is supported from its integral boss through the opening in which extends the inner end of the focusing screw. The screw is threaded through the rear stationary magnet plate, so that turning the screw moves the screw and the attached focusing sleeve one way or the other. Moving the sleeve to the right in the drawing, or toward the picture tube screen, shortens the air gap between the sleeve and the front magnet plate. This reduces the strength of the magnetic field that passes through the picture tube electron gun.

At the right in Fig. 61-14 is shown the centering adjustment. There is a centering shutter or aperture plate that operates on the principle illustrated in Fig. 61-12. One side of the shutter has an extended tongue in which is a slot that fits over a pivot set into the front magnet plate. This slot allows the shutter to be moved up and down when in the position shown by the drawing. On the opposite side of the shutter plate is another tongue having a small opening into which fits the tip of the focusing handle.

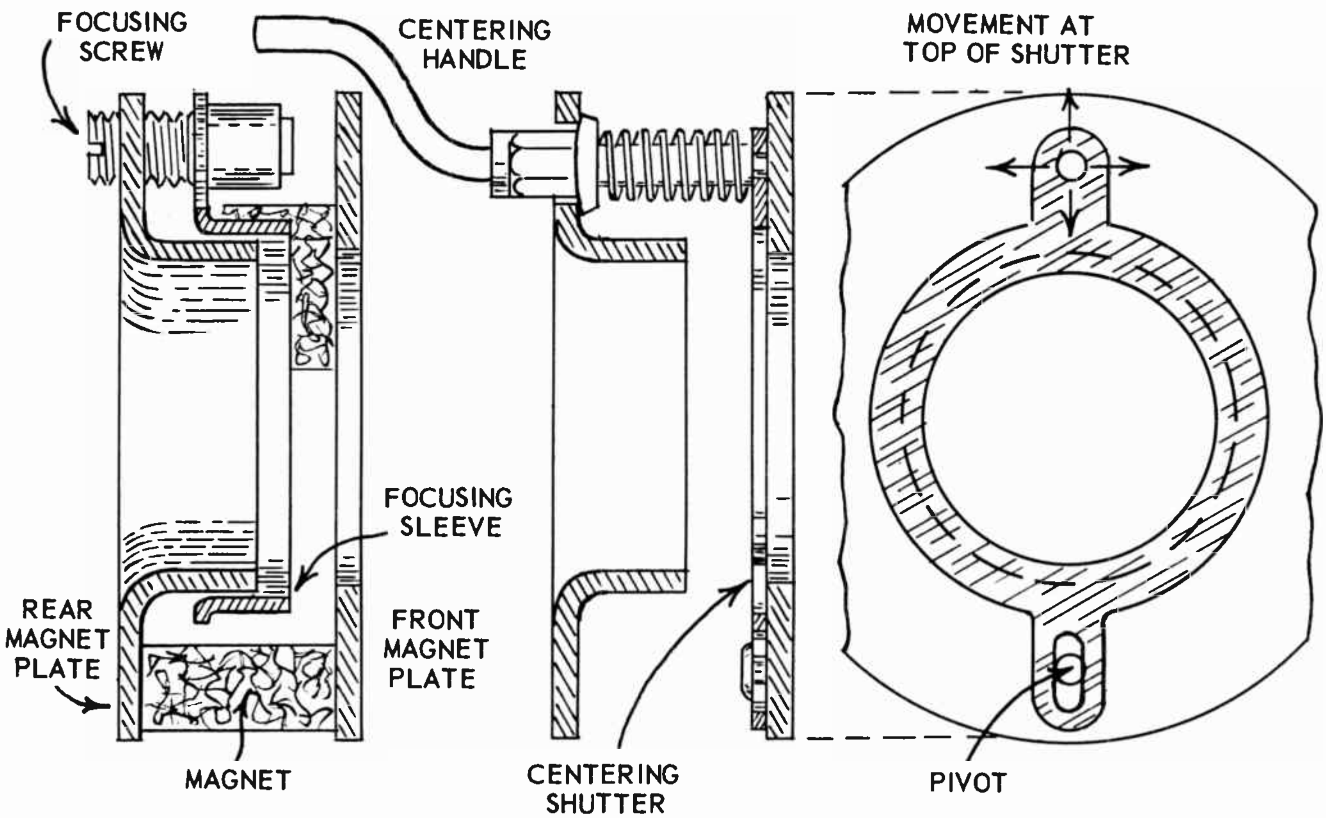


Fig. 61-14. Details of adjustments for focusing and centering with a permanent magnet structure.

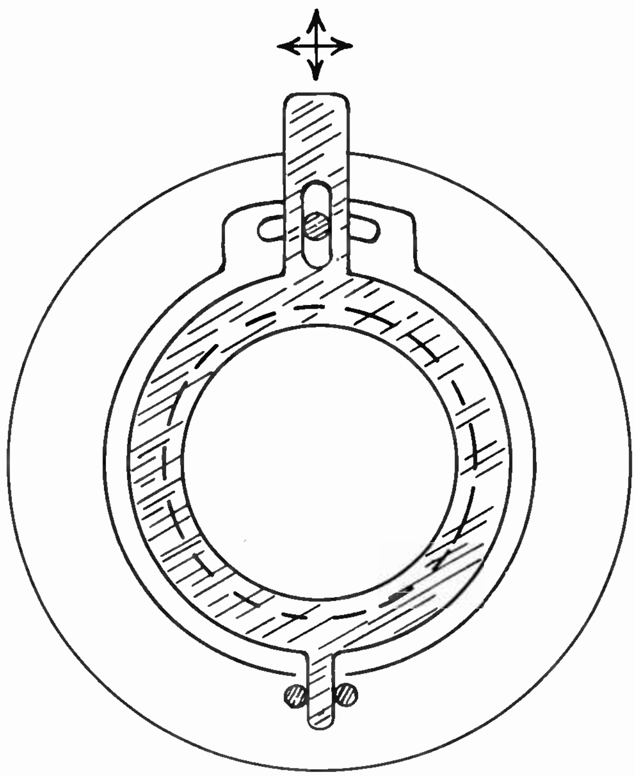


Fig. 61-15. A method for adjusting the position of a centering plate.

The focusing handle is supported in the rear magnet plate by a sort of universal joint. Moving the exposed end of the handle down will move the shutter up, moving the handle up moves the shutter down, moving the handle to the left shifts the shutter to the right as it pivots on its lower tongue opening, and moving the handle to the right shifts the shutter opening to the left. The universal mounting allows any possible combination of these shutter movements, and allows tilting the axis of the focusing magnetic field in any direction with reference to the picture tube axis. The shutter is held in its adjusted position by friction between the front face of the shutter and the inside of the front magnet plate. Frictional contact pressure is maintained by the coiled spring around the centering handle extension between the two magnet plates. The centering handle or centering "stick" may be straight rather than curved, and may extend to several inches back of the magnetic structure.

All styles of permanent magnet focusing and centering devices are supported on the bracket that carries the deflecting yoke. Holes for the supporting screws are enlarged or slotted to allow preliminary positioning of the magnet unit with its center on the center line through the picture tube. Some of these magnetic units are carried on trunnions or adjustable mountings which allow them to be tilted bodily, just as a focus coil may be tilted. This tilting of the entire magnet structure may be the sole means for centering the picture, with the only magnetic adjustment that for focusing. The focusing adjustment may be by means of magnetic shunting screws, as at 3 in Fig. 61-11, instead of with a shunting ring.

Centering plates quite often are adjusted with some such arrangement as shown by Fig. 61-15. A tongue on the bottom of the plate is held between two guide pins. On the plate at a point opposite this tongue is a slotted extension through which passes a locking screw fitted with a thumb nut. This screw passes also through a curved slot whose "tangent" is at right angles to the slot in the centering plate extension. This right-angled slot is formed in some stationary part of the device. With the thumb nut loosened, the centering plate may be shifted to whatever position centers the picture, and held in this position by tightening the thumb nut.

In still other designs the centering plate is moved into the required position by turning two eccentrics, which are discs carried by a pin or screw at a point that is offset from the center of the disc. The plate is held in its adjusted position by the tension of small coiled springs that press against the sides of the centering plate that are opposite the eccentrics.

All these focusing and centering adjustments on permanent magnet units are mechanical arrangements whose methods of operation are quite evident upon inspection, or whose effects become apparent when making careful preliminary movements of what appear to be the adjustments.

Focusing and centering adjustments almost always must be altered during setting of an ion trap magnet. As emphasized in earlier instructions, the trap magnet should be placed only for maximum brightness of the raster. Shadowing should be eliminated by the centering adjustment, or, if necessary, by moving the entire magnetic structure to bring its center on the tube axis or to tilt it with reference to the structure to bring its center on the tube axis or to tilt it with reference to the tube axis.

Adjustments for focusing and centering with permanent magnet units should be made only with a non-magnetic tool. A steel-bladed screw driver may so affect the focusing and centering magnetic field that performance will change as soon as you take away the tool. Focusing screws of the general type pictured in Fig. 61-13 are easily turned by using a silver dime or a copper cent, both of which are non-magnetic. The practice of using a non-magnetic tool is a good one to follow also when adjusting the tilt of an electromagnetic focus coil.

ELECTROSTATIC FOCUS WITH MAGNETIC DEFLECTION. To retain the advantages of magnetic deflection while eliminating external focusing devices of all kinds some tubes are designed for electrostatic focusing and for use of the usual type of magnetic deflecting yoke. This combination was used in several types of picture tubes which have become obsolete, and is found in one 7-inch tube (7DP4) which still is found in some receivers, and again has been revived in a number of comparatively new types in sizes from 14- to 20-inch face diameter or the equivalent rectangular sizes.

The electron gun structure is similar to that of the familiar electrostatic deflection and focus tubes except for having no deflecting plates. The focusing anode is operated usually at 20 to 25 per cent of the voltage applied to the second anode or high-voltage anode. Voltage on the first (focusing) anode is adjustable for focusing the electron beam. There is, of course, no focus coil or magnet which may be tilted for centering the picture. Centering is handled by adjustable direct currents in the horizontal and vertical deflecting coils, just as earlier described for tubes having magnetic deflection combined with magnetic focusing.

TELEVISION

LESSON NO. 62

HIGH-VOLTAGE POWER SUPPLIES - I

"Training Your Mind"
C. T. I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

COMMERCIAL TRADES INSTITUTE

Training Your Mind

C. T. I.

Perfecting Your Skill

Chicago, Ill.

LESSON NO. 62

HIGH-VOLTAGE POWER SUPPLIES - I

Picture tubes operated with magnetic deflection require from 8,000 to 14,000 d-c volts or more for their anodes. Tubes using electrostatic deflection and focus require 4,000 to 6,000 d-c volts for their high-voltage anode, and something like 800 to 1,200 volts for the focusing anode. Such anode voltages are furnished by a high-voltage power supply system that is separate from the low-voltage system used for plate, screen, and grid biasing voltages of all the other tubes.

Energy for the high-voltage power supply is taken from the low-voltage supply in the form of a small direct current which, in most cases, is at 125 to 350 volts. The only way of increasing the voltage with apparatus such as may be installed in television receivers is to use a step-up transformer. A transformer will operate only from alternating or pulsating current, not from direct current as taken from the low-voltage power supply. Consequently, the first step toward raising the voltage is to change the direct current from the low-voltage system into alternating or pulsating current for the step-up transformer.

There are two commonly used methods for securing an alternating or pulsating current from the original direct current. In receivers having magnetic deflection picture tubes a suitable pulsating current is produced incidentally in the sweep circuit, from energy taken into this circuit from the low-voltage supply. The sudden reversals of horizontal deflection current cause corresponding pulses of voltage in the sweep output transformer. By auto-transformer action these voltage pulses are stepped up to a value high enough for the picture tube anode circuit. Then we have a pulsating high-voltage at the horizontal frequency of 15,750 cycles per second. This high voltage is rectified, and the resulting direct current is fed to the picture tube anode. Such an arrangement is called a flyback high-voltage power supply.

The parts of one flyback high-voltage system are illustrated by Fig. 62-1. We have looked at pictures of other arrangements in earlier lessons. In the present design the rectifier tube is supported in a horizontal position. Back of the rectifier are the horizontal sweep amplifier and the damper tubes. These three tubes, amplifier, damper, and rectifier, nearly always are close together on the chassis. Down below the rectifier may be seen part of the sweep output transformer. In this design the remainder of the of the transformer extends under the chassis. At the extreme right, supported by a stud on the chassis, is the high-voltage filter capacitor, and near the rectifier socket is the filter resistor.

With the second type of high-voltage power supply we change the original direct current into alternating current by using an oscillator whose output frequency usually is between 100 and 300 kilocycles. The high-frequency oscillator current is used in a step-up transformer to secure a voltage high enough for the picture tube anode or anodes. This high voltage then is rectified and fed to the anode circuit. This

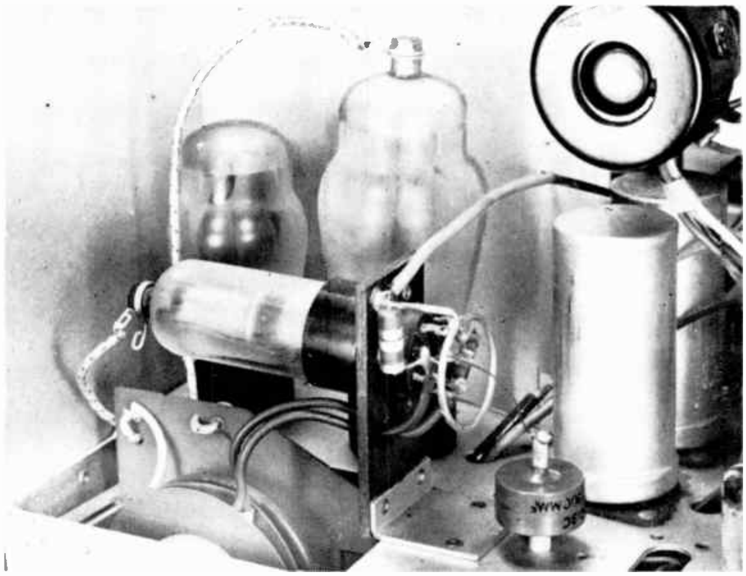


Fig. 62-1. The principal parts of a flyback type high-voltage power supply.

arrangement usually is called an r-f (radio-frequency) high-voltage power supply. It is used with electro-static-deflection picture tubes, where there are no sawtooth currents whose reversals would operate a fly-back power supply, and is used also instead of the flyback system for many magnetic-deflection picture tubes.

Fig. 62-2 is a picture of one style of r-f high-voltage power supply. We have looked at other examples of this type in earlier lessons. Here the oscillator tube is at the left, and at the right is the high-voltage rectifier tube. Between the two tubes is the air-core step-up transformer. Efficient operation is obtained with an air-core transformer rather than an iron-core type because of the high operating frequency. Other parts of this power supply are underneath the base plate. Directly below the oscillator is an adjustable capacitor used for varying the output voltage from this system.

(a) With either type of high-voltage power supply we commence with d-c energy from the low-voltage system. This energy first produces pulsating or alternating current and voltage. Then the voltage is increased by a transformer, and finally is rectified to furnish high direct voltage and current for the picture tube anodes. In order to obtain a high voltage we have to go from direct current to alternating or pulsating current, then back to direct current.

Terminals at which potentials are thousands of volts in the flyback system include the top caps of the sweep amplifier and rectifier when these tubes are of types having such caps, also the terminals or lugs on the rectifier socket. With an r-f high-voltage supply the high-potentials are at the rectifier top cap, if used, and also at the rectifier socket and the transformer terminals.

These high-voltage points always are covered in some manner that protects you and others against shock. In the majority of receivers the parts of the high-voltage supply are enclosed within a metal shield that is perforated or has other ventilating openings. You positively must not work on high-voltage elements while their cover is removed and the power cord is plugged into a line receptacle. Turning off the receiver switch is not enough to avoid danger.

Nearly all receivers are equipped with some form of safety interlock that opens the line power circuit when the cover is removed from the high-voltage supply or, in many cases, when the cabinet is opened to give access to the chassis. This interlock commonly consists of a double-pin recessed male socket mounted on the chassis and a female plug that fits into the socket. The socket pins are connected to the off-on switch and to the primary of the low-voltage power transformer in the chassis. The plug is attached to the cover of the power supply or to a removable back panel of the cabinet, and is on the receiver end of the line cord. The power supply cover or the cabinet back cannot be removed without taking the attached plug with it, which automatically cuts off the line power.

To avoid the necessity of replacing the power supply cover or the cabinet back every time you wish to apply power for observing the effect of adjustments, you can use a special line cord available from radio and television supply stores. One end of this cord is a double female plug that fits the safety socket on most receivers, and on the other end is a regular two-prong plug to fit a line power receptacle. When using a cord of this kind you always must disconnect it before touching any parts of the power supply with your fingers or any tool.

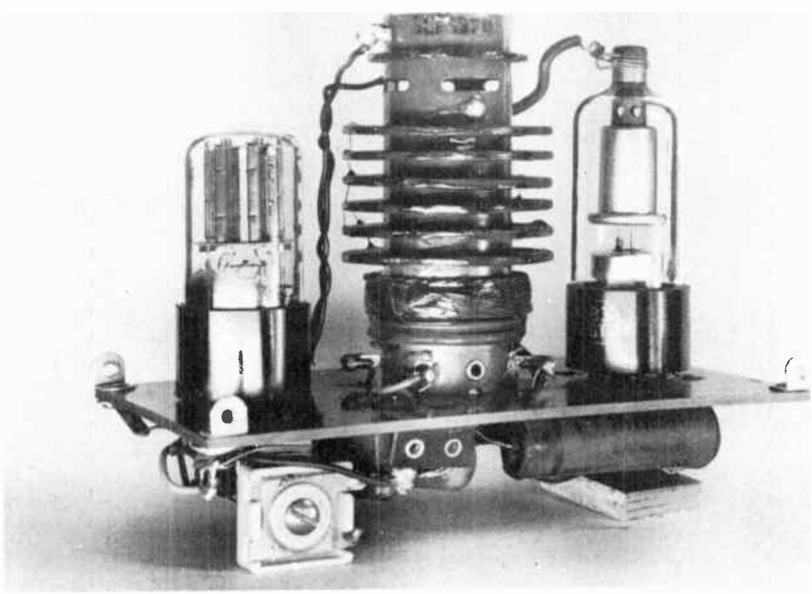


Fig. 62-2. Tubes, transformer, and voltage adjusting capacitor of an r-f high-voltage power supply.

Some receivers have in series with their regular off-on switches a safety switch of the spring plunger type. This safety switch remains closed while the plunger is depressed, and opens when the plunger is released. The switch is pressed closed by the cabinet back when put in place, thus allowing power to be applied when the regular off-on switch is turned on. Removing the cabinet back or any access panel releases the switch and opens the line power circuit. Such a switch may be pressed closed while you observe the effects of adjustments or replacements, but it always must be open or else the line cord must be disconnected while you do any work on the high-voltage units.

FLYBACK POWER SUPPLIES. Fig. 62-3 is a circuit diagram for a flyback power supply and parts associated with it. When plate current is cut off in the horizontal sweep amplifier, at the instant for retrace, there is sudden collapse of the magnetic fields which have been built up around the output transformer and deflecting coils during the relatively slow increase of sawtooth current. The rapid cutting of the field lines through the turns of the transformer winding induces proportionately high emf's in the windings. The induced emf is positive on the rectifier plate end of the primary while negative on the damper plate end of the transformer secondary.

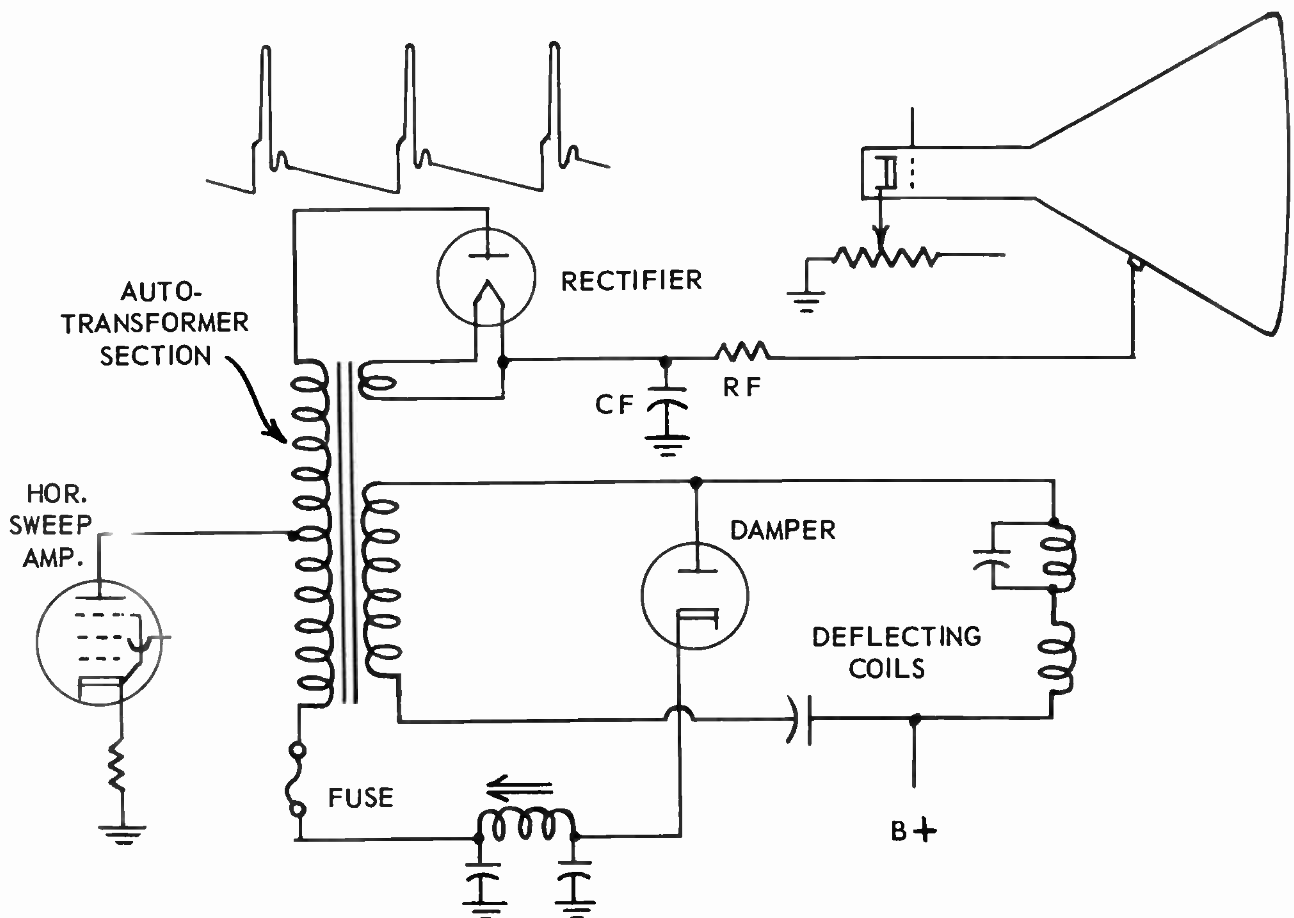


Fig. 62-3. Circuits of a typical flyback type high-voltage supply.

Due to the relatively great number of turns in the primary winding the emf induced in this winding reaches a peak value of around 4,000 to 6,000 volts. The primary winding has an extension or has additional turns beyond the connection for the sweep amplifier plate. The entire winding becomes an auto-transformer in which the originally induced emf is stepped up to between 8,000 and 12,000 volts, or to whatever voltage is required for delivering power to the picture tube anode. The power that is furnished to the anode circuit comes from energy that has been stored in the magnetic fields of the transformer and deflecting coils during the gradual rise of sawtooth current between retrace times.

The positive pulses of high voltage from the auto-transformer section are applied to the plate of the high voltage rectifier. The filament-cathode of the rectifier is affectively at ground potential so far as the high -frequency pulses are concerned, because of its connection to ground through the rather small capacitive reactance of the filter capacitor Cf. Thus the filament-cathode of the rectifier is on the negative side of the high-voltage power circuit while the plate is positive. The voltage pulses are rectified to produce across the filter capacitor a direct voltage nearly equal to the peak-to-peak value of the pulse voltage applied between plate and filament-cathode of the rectifier.

Current taken from the filter system for the picture tube anode is very small, only a few microamperes. This makes it possible for the small filter capacitor and the filter resistor Rf to remove the horizontal frequency variations from current that goes from the filter to the picture tube anode.

In some high-voltage supplies there is a 1/4-ampere fuse in the line going to the low end of the horizontal output transformer. This fuse will blow and prevent burn-out of the transformer winding should the sweep amplifier tend to draw excessive current. Excessive plate current may result from an internal short circuit between elements of the tube, or it may result from the tube becoming gassy. A gassy tube is one in which gases have been released from some of the metal elements due to overloading or to manufacturing defects. Should the fuse blow, it nearly always indicates a defective horizontal sweep amplifier, which should be replaced with a good tube before installing a new fuse.

It is important to note that pulses of induced emf, at 4,000 volts or more, exist at the plate cap or the plate terminal on the socket of the horizontal sweep amplifier. This is the reason for keeping away from this terminal, and especially for not attempting to make any plate voltage measurements with an ordinary voltmeter having a range only high enough for average plate voltage, not for the pulse peaks.

The voltage delivered from the power supply to the picture tube anode depends not only on the step-up ratio of the auto-transformer section but also on the amount by which the primary and secondary currents change in producing the voltage pulses. As mentioned before, the output voltage or anode voltage is varied by adjustment of the drive control and by adjustment of some types of negative peaking control. This voltage is varied also by voltage applied to the screen of the horizontal sweep amplifier, the greater the screen voltage the higher the picture tube anode voltage within limits. Still another factor is the d-c voltage applied from the low-voltage power supply to the deflecting coil circuit or to the damper tube. Higher d-c voltage will raise the picture tube anode voltage.

Some high-voltage output transformers have several taps on the secondary winding that connects to the deflecting coils. Moving the connection to a tap which includes more winding turns will raise the output voltage. This, ordinarily, is not a good way to change the picture tube anode voltage, because the taps are provided to allow matching the impedance of the transformer winding to the impedance of the deflecting coils. Matching these impedances allows maximum power transfer to the deflecting coils, and chang-

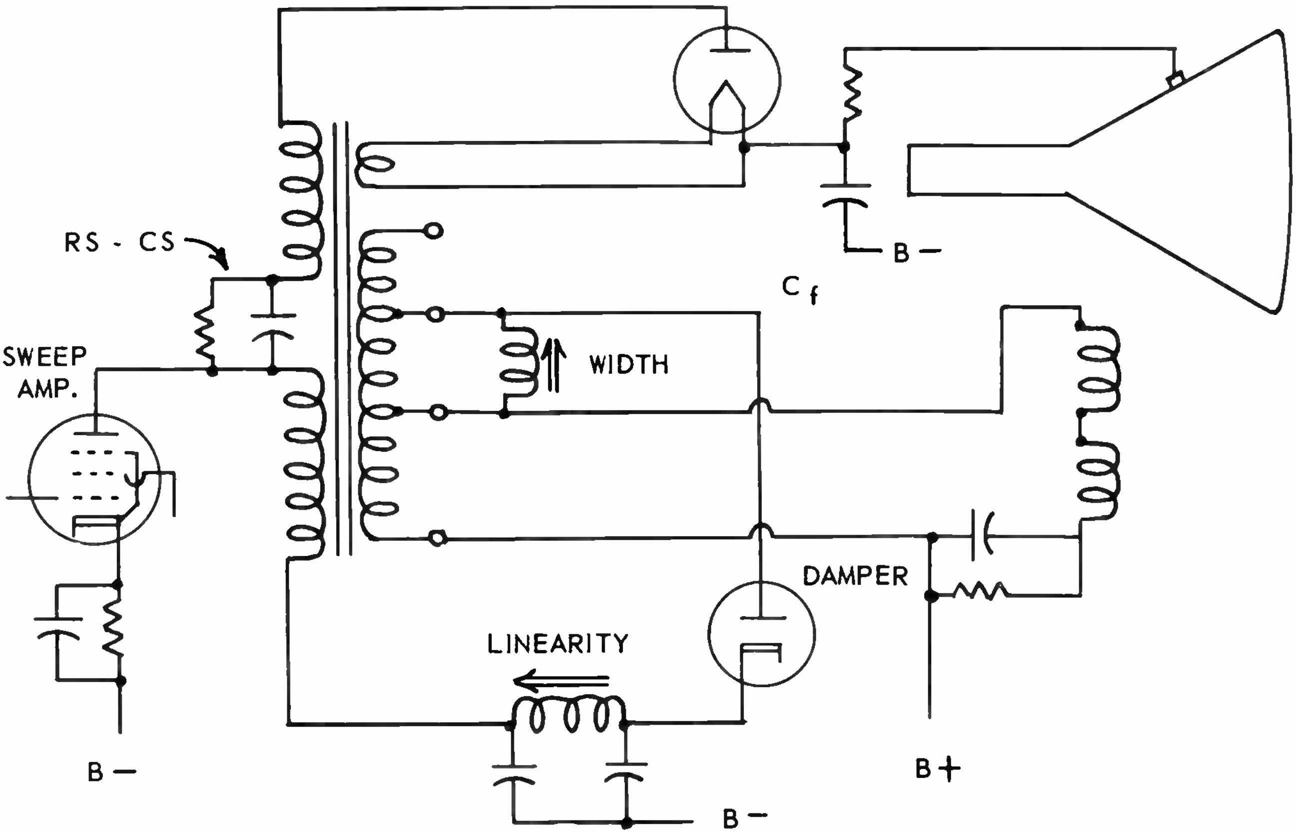


Fig. 62-4. Tapped secondary on a flyback output transformer.

ing the tap connection may cause deflection trouble. Fig. 62-4 shows one type of deflection circuit in which the transformer winding is provided with taps. This diagram shows also some of the variations to be expected in horizontal sweep circuits.

In transformerless or ac-dc television receivers the circuit points which otherwise might connect to chassis ground would connect to B-minus. This is illustrated in Fig. 62-4. In some such receivers there may be a resistor and capacitor, Rs and Cs, between the two primary sections of the output transformer. The capacitance may be about 2,000 mmf, providing a reactance of only 5,000 ohms around the half-megohm resistor, and thus completing the primary circuit for the horizontal line frequency. These units prevent a low-resistance short circuit from power line to chassis should the anode cable connector accidentally touch the chassis. When these safety units are not used the anode connector should be supported rigidly or in some other manner which prevents it from touching the chassis when disconnected from the picture tube.

VOLTAGE BOOSTING. With flyback circuits such as shown by Figs. 62-3 and 62-4 there is available a "boosted" positive voltage which is greater than the B plus voltage applied from the low-voltage power supply. The voltage boost results from the following action.

The damper tube conducts during a considerable portion of each sawtooth cycle in the deflecting coil circuit. There are peaks of conduction current when the first oscillation cycle is damped during each retrace period. We learned about this conduction when studying the subject of damping. The damper conduction current, charges the capacitor which is between the damper cathode and ground or B-minus, the capacitor that is on the damper side of the linearity control choke coil. Conduction current passes through the choke to also charge the capacitor at the other end.

Because of conduction current peaks during retrace periods these capacitors are charged to voltages higher than the B plus voltage applied to the circuit. The inductance of the linearity choke combined

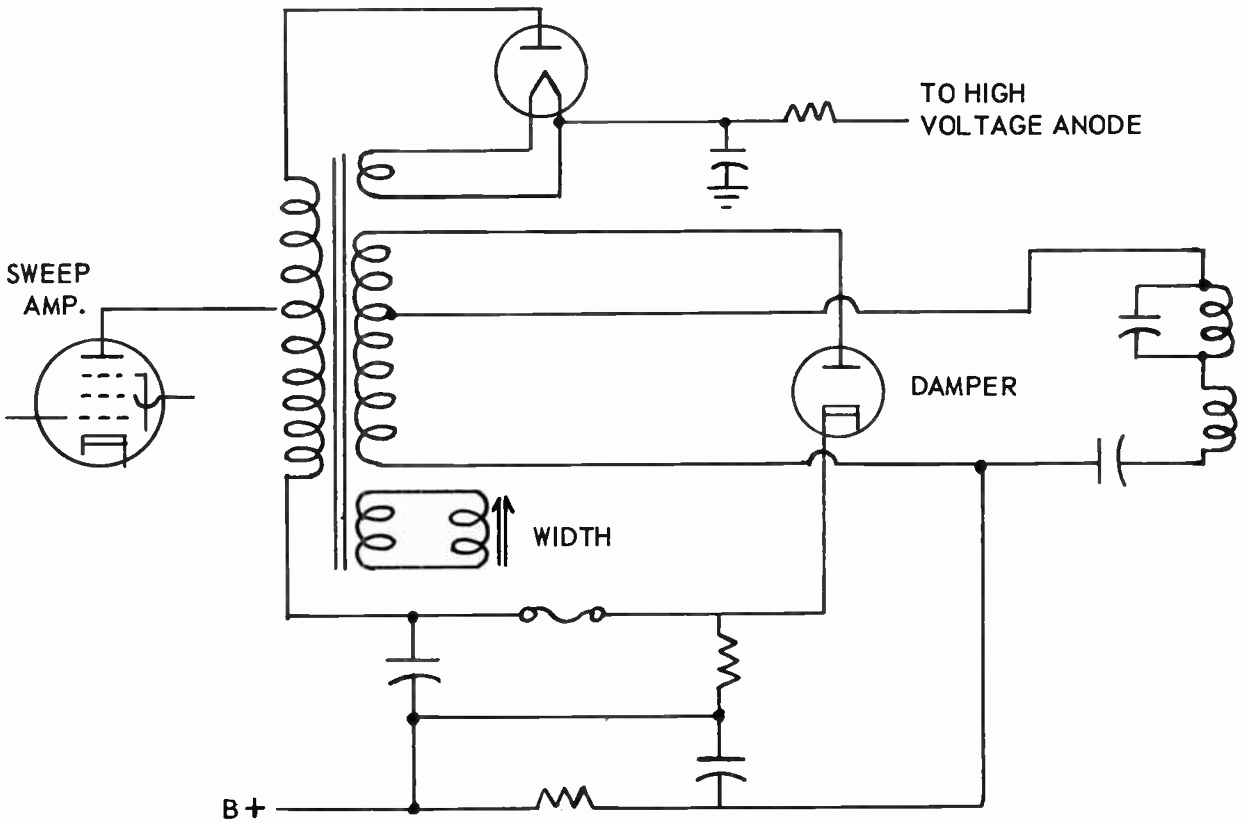


Fig. 62-5. Voltage boosting circuit without linearity adjustment choke.

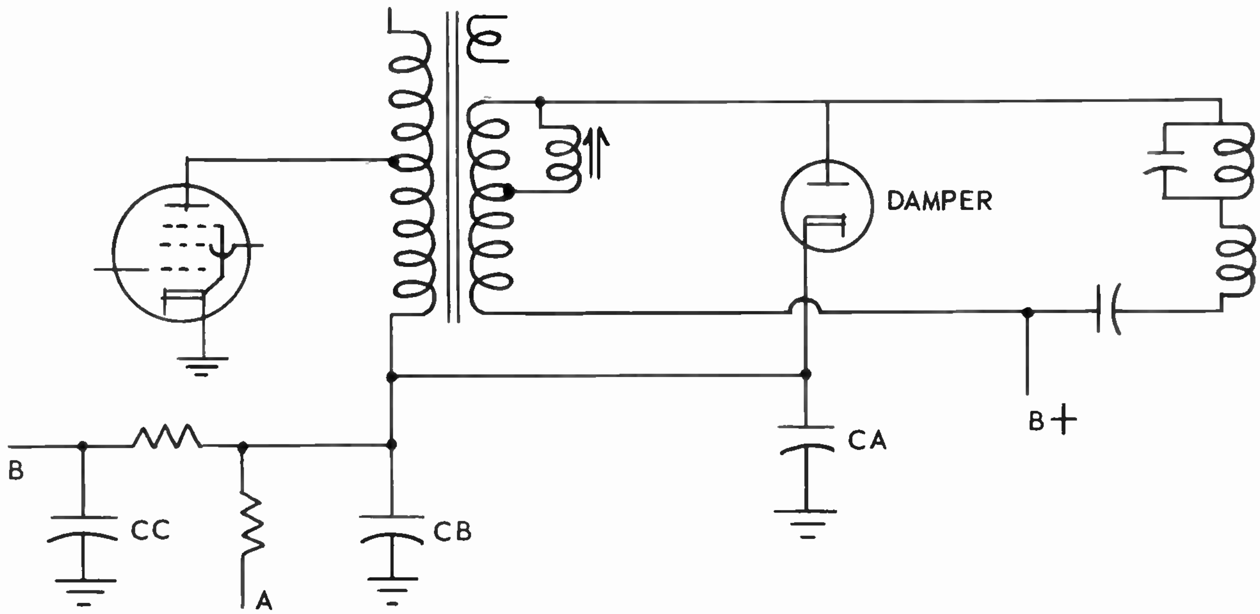


Fig. 62-6. Voltage boosting with only capacitors on the damper cathode line.

with the two capacitances form what amounts to a "low-pass" filter similar in action to a choke-capacitor filter used in low-voltage power supplies. The filter partially smooths the boosted d-c voltage.

The amount of boost may be almost anything from 40 to 250 volts, depending on the circuit design. This relatively high voltage is applied to the plate of the horizontal sweep amplifier through the output transformer primary. The boosted voltage often is used also for the sweep amplifier screen, for horizontal and vertical sweep oscillators, and sometimes for the vertical sweep amplifier plate.

The choke of the linearity control is not essential for voltage boosting. It is the capacitors which store the energy from current pulses, the choke merely helps to smooth or to vary the waveform of boosted voltage. One style of booster circuit using only capacitors and resistors is shown by Fig. 62-5. The capacitors here are connected between the damper cathode and the B plus line, just as are the capacitors at the ends of some of the linearity control chokes which we have seen in other circuits. This effectively returns one side of the capacitors to ground or B-minus because the positive side of the B plus lines always is bypassed through large capacitances to ground or to B-minus at one or more points.

Fig. 62-6 shows another voltage boosting circuit. Here there is energy storage in capacitors Ca and Cb, but there is neither a choke nor a resistor between damper cathode and the transformer primary. The boosted voltage is applied to the sweep amplifier plate and is taken also through dropping resistors to other tubes at connections such as A and B. Additional filtering or smoothing for the other circuits is provided by capacitor Cc and by other capacitors between the various leads and ground or B-minus.

VOLTAGE DOUBLING. In some receivers which have 16-inch and larger picture tubes, and in a few which have smaller tubes, the voltage from the high side of the horizontal output transformer is approximately doubled for the picture tube anode. This is accomplished by using two high-voltage rectifiers in a voltage doubler circuit. Such a power supply system is shown by Fig. 62-7. The two rectifiers are alike. Each has its own separately insulated filament winding on the transformer. Capacitors Ca, Cb, and Cc usually are of about 500 mmf capacitance, but may be as large as 1,000 mmf. All are high-voltage types, rated at 10,000 to 15,000 volts or more, according to the voltages at which the system works. Resistor Rd may be of some value between one and four megohms, often consisting of several resistors in series to lessen the chances of breakdown.

Because this doubler system is supplied with pulsating voltage rather than with alternating voltage of approximately sine wave form the circuit and its performance are not quite the same as with the doublers studied in connection with low-voltage power supplies.

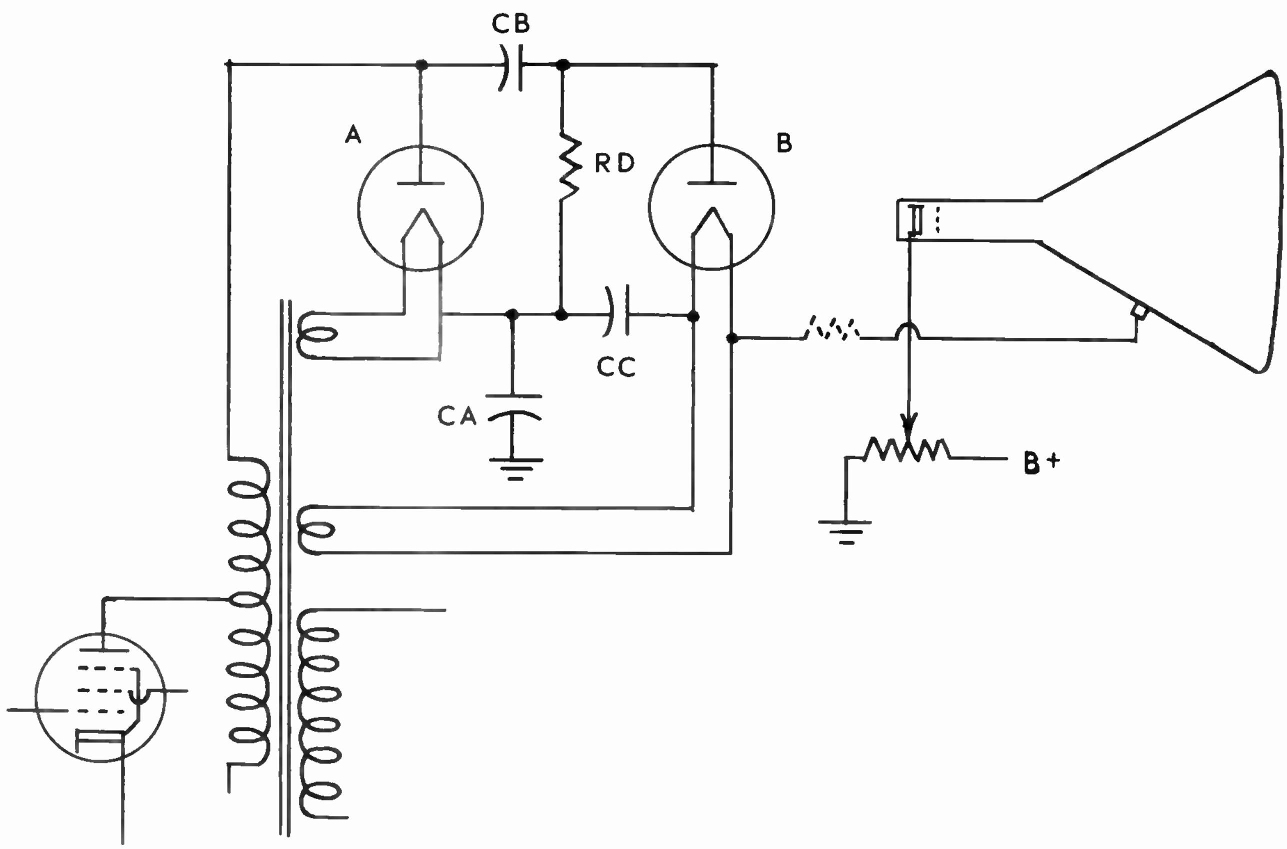


Fig. 62-7. Connections for voltage doubling with a flyback power supply.

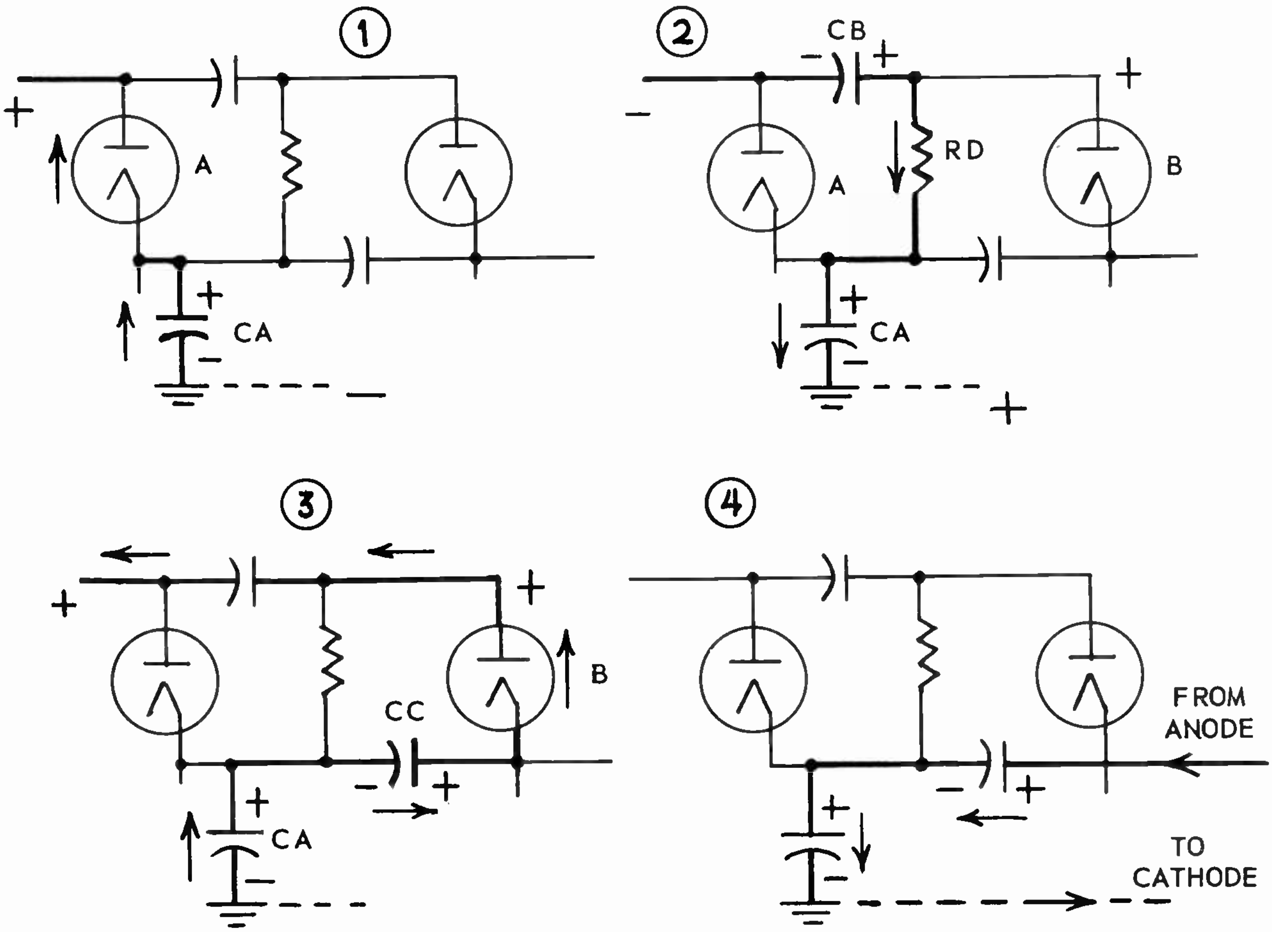


Fig. 62-8. Electron flows during voltage doubling with positive pulse voltage as the source.

The doubling action may be explained with the help of Fig. 62-8. Diagram 1 represents a period during which a positive voltage pulse is being applied from the transformer to the plate of rectifier A. This rectifier becomes conductive, and there is electron flow as shown by arrows, from the relatively negative ground connection through capacitor Ca and rectifier A. Capacitor Ca thus is charged in the marked polarity.

After the positive pulse has ceased, the plate of rectifier A becomes negative with reference to ground, as in diagram 2. The charge voltage which is on capacitor Ca now causes electron flow through capacitor Cb, resistor Rd, and capacitor Ca as shown by arrows. The polarity of the charge put into capacitor Cb is such as to make its right-hand terminal positive, and since this terminal is connected to the plate of rectifier B the plate is made positive and B may conduct.

At the next positive pulse we have the conditions shown by diagram 3. Although the charge voltage on capacitor Ca is acting in such polarity on the cathode of rectifier B as to hold this rectifier non-conductive, the combined strength of the new positive pulse and of the voltage which has been put on capacitor Cb overcomes the opposing voltage of Ca, and rectifier B conducts. The resulting electron flow in the direction of the arrows charges capacitor Cc in the marked polarity.

The result of the actions which have been described is to charge both Ca and Cc to voltages very nearly equal to the peak pulse voltage from the transformer. The two charged capacitors are in series with each other and the picture tube anode circuit, as shown by diagram 4. The voltages of the series capacitors add together, and apply between the picture tube anode and cathode a total potential difference almost double the pulse voltage. Electron flow in the anode-cathode circuit then is as shown by arrows in diagram 4.

You will realize, of course, that the actions illustrated by diagrams 1 and 3 of Fig. 62-8 do not occur during alternate pulse periods, both of them occur every time there is a positive pulse. Also, the anode-cathode electron flow of diagram 4 is continual because the two capacitors are being continually recharged by the voltage pulses and are continually losing some of their charge through the picture tube.

FILTERING FOR FLYBACK POWER SUPPLIES. As you will recall from our earlier studies of filtering for d-c power supply systems, the problem is made easier when only small currents are required. In the case of the anode supply for picture tubes the current varies with intensity of the electron beam or with variations of control grid voltage which control the beam. For even the brightest parts of any picture, when operating voltages are those ordinarily used, the anode current is unlikely to reach 150 microamperes, and the average current will be much less. Accordingly, the filtering for the anode supply may be quite simple.

We learned also when first studying the matter of filters that the output becomes more nearly a smooth direct current as the supply frequency is increased. With flyback power supplies the frequency is 15,750 cycles per second, which further adds to the ease of filtering.

With the commonly employed circuits of Figs. 62-3, 62-4, and 62-9 the filter capacitor Cf usually is of 500 mmf capacitance, and is rated at 10,000 volts or more according to the actual working voltage of the system. The filter resistance at Rf most often is between a half-megohm and one megohm, although it may be less than a half-megohm and in some cases there is no filter resistor at all.

When an all-glass picture tube has both external and internal conductive coatings, as in Fig. 62-9, the capacitance of these coatings, with the glass for dielectric, forms a second filter capacitor because the external coating is connected to ground. This capacitance of the conductive coatings increases with picture tube diameter. It never is less than 500 mmf, and in some of the largest all-glass tubes may go as high as 4,000 mmf.

In Fig. 62-9 the low end of filter capacitor Cf is shown connected to the line on which is the damper plate instead of to ground as in many of our other circuit diagrams. Connection to the damper or deflecting coil circuit raises the voltage applied to the picture tube anode because the potential of the entire high-voltage filter system thus is raised above ground potential by the amount of voltage in the damper or coil circuit. Voltage peaks in this circuit may be far higher than the average value of B-plus voltage coming to it from the low-voltage supply, and picture tube anode voltage may thus be raised by as much as 1,000 volts above the value secured with the filter capacitor to ground or B-minus.

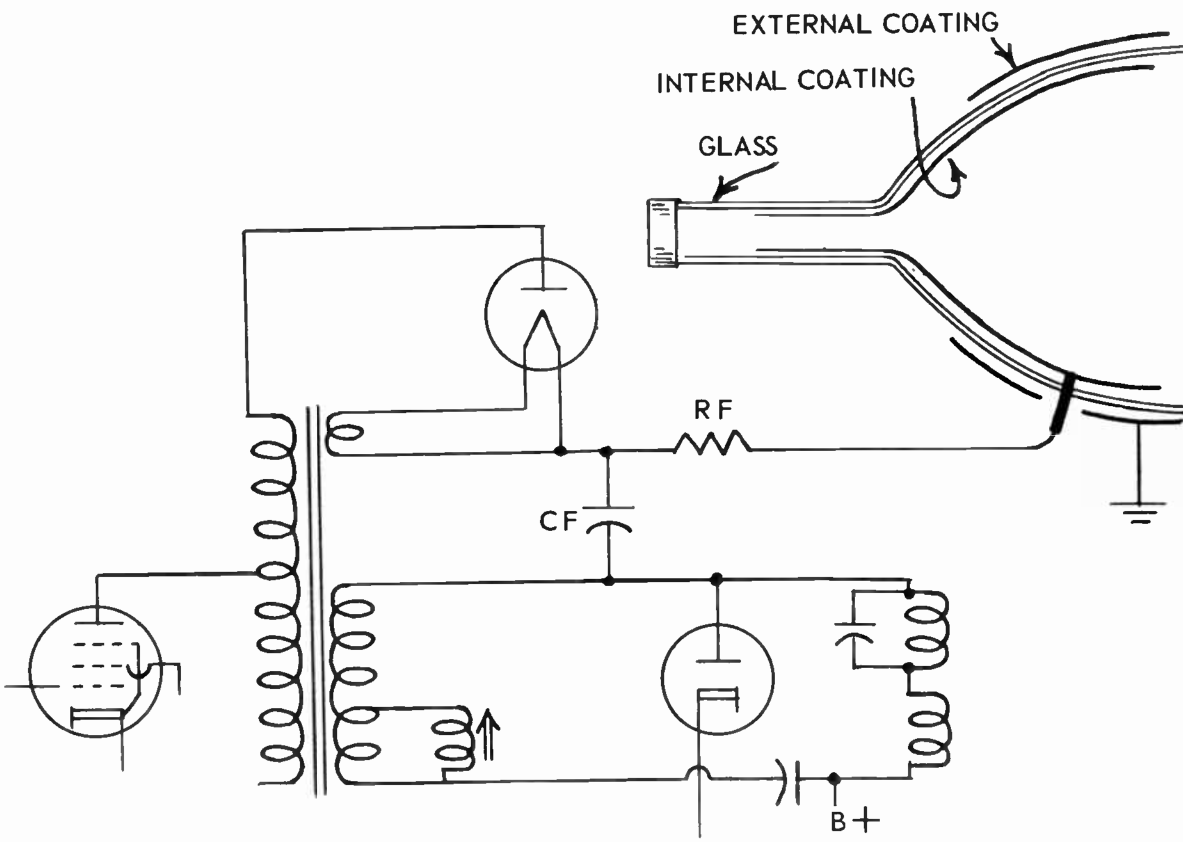


Fig. 62-9. How the coatings and glass envelope of a picture tube act as a power supply filter capacitance.

It is interesting to note the results of certain tube failures in a receiver which has a flyback type high-voltage power supply. When plate voltage for the horizontal sweep amplifier is taken from the boosted supply at the damper cathode, failure of the amplifier will stop the voltage pulses in the output transformer. Then there will be no high accelerating voltage applied to the picture tube anode to draw the electron beam, and the screen will remain dark. Naturally, the same thing happens with failure of the damper tube, for then there is no boosted voltage.

If the horizontal sweep oscillator or the discharge tube fails to deliver sawtooth voltage to the sweep amplifier there can be no voltage pulses to operate the flyback system, and the picture tube screen will remain dark because there will be no accelerating voltage for the electron beam. Should the vertical sweep circuit develop such faults as to prevent vertical deflection, while the horizontal sweep circuit continues to operate, there will be only a horizontal line of light across the picture tube screen.

It is fortunate that faults which would stop deflection of the electron beam will prevent the beam from reaching the picture tube screen. An undeflected beam of any intensity would maintain a bright stationary spot of light on the screen, and would leave a permanent blemish after only a short period of such operation.

Interesting also is the fact that sudden changes of currents and voltages in the flyback power system occur during horizontal retrace periods, while the electron beam is blanked. Consequently, these pulses do not interfere with picture reproduction. These systems can, however, radiate interference waves which may be troublesome for nearby radio receivers. This is one reason for enclosing the power supply elements within a metal cover grounded to the chassis.

HIGH-VOLTAGE RECTIFIERS. Doubtless you have noticed that in all the circuit diagrams of high-voltage power supplies the rectifier tubes are of the half-wave type with a filament-cathode and a plate as the only elements. A half-wave rectifier is entirely satisfactory in this service, because of the easy filtering requirements. A much more intricate pulse-voltage supply circuit would be needed to operate a full-wave rectifier.

For the three types of high-voltage rectifier tubes most commonly used Fig. 62-10 shows relative sizes and proportions of the bulbs and also the positions and connections of the base pins. The 1B3-GT is largest of the three types illustrated and is rated for highest maximum voltage peaks, highest maximum current peaks, and highest average current. This tube has an octal base with six pins, of which numbers 2 and 7 connect to the filament. The plate is connected to a top cap. The remaining pins may be connected to pin 7, but otherwise must not be used.

Fig. 62-11 is a picture of the 1B3-GT tube. The plate is a hollow metal cylinder attached to the top cap. The filament is well up inside the plate. Below the plate and above the base is a cylindrical metal shield which is internally connected to pin 7.

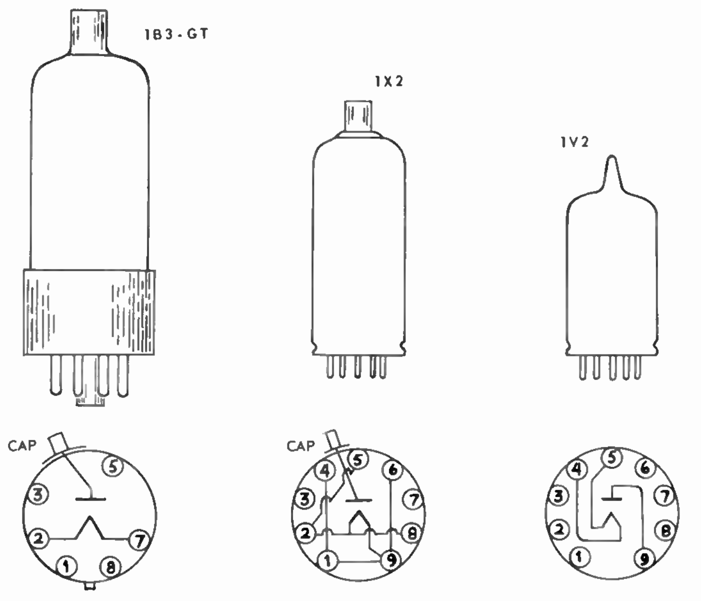


Fig. 62-10. Relative sizes and base pin connections of three high-voltage rectifier tubes.

Fig. 62-11. A 1B3-GT high-voltage half-wave rectifier tube.

The 1X2 and 1V2 rectifiers are miniature types, both with 9-pin bases. The plate of the 1X2 is connected to a top cap which is of smaller size than the cap on the 1B3-GT. One side of the filament is internally connected to pins 2, 5, and 8. The other side of the filament is connected to pins 9. 1, 4, and 6. The internal shield is connected to pins 1 and 9. Only pins 3 and 7 have no internal connection.

The 1V2 rectifier is the smallest of the types shown by Fig. 62-10. It has no top cap, rather the plate is connected to pin 9. The filament is between pins 4 and 5. This tube is rated for 7,500 volts maximum peak inverse plate voltage. The similar rating for the 1X2 is 18,000 volts, and for the 1B3-GT is 30,000 volts. Any of these tubes may be mounted in any position.

Filament power for all these high-voltage rectifiers is secured from one or two turns of heavily insulated wire on the horizontal output transformer. The 1B3-GT and 1X2 filaments operate at 1.25 volts and 0.2 ampere. The 1V2 filament operates with 0.625 volt or 5/8 volt and 0.3 ampere. The small number of turns on the output transformer is required because of the very great step-down ratio needed for the low filament voltages.

The filaments are heated by voltage and current at the horizontal line frequency of 15,750 cycles per second, but they are so large and heavy that they lose almost no heat and undergo negligible change of temperature between voltage and current pulses. Such filaments are said to have high "thermal inertia".

It is not practicable nor is it safe to make measurements of rectifier filament voltage with any ordinary equipment, because of the very high voltages at the filaments while the receiver is in operation. Should it ever be necessary to check the filament operation of either the 1B3-GT or the 1X2 it is done by comparison of the brightness with that of a filament known to be good. The recommended method is to apply to the filament of a good tube a direct voltage or a low-frequency alternating voltage of 1.25 volts and to look at the illumination reflected from the surface of the internal shield. This must be done in a darkened room. When the reflection from the good filament is apparently equal to that from the filament being checked their operating voltages and currents are nearly enough equal for satisfactory and safe operation.

The filament of the 1V2 rectifier cannot be checked with the method just described. With any of these tubes which you suspect of faulty operation the normal service procedure is to try a new tube known to be in good condition. If this corrects the trouble there is no need for making intricate checks, and if the trouble remains it is not the fault of the rectifier tube.

MEASURING HIGH VOLTAGES. You must not attempt to measure voltages at the anodes or in the anode circuits of picture tubes without equipment designed for, or especially arranged for, such measurements. You also must observe many precautions which are not necessary when measuring lower voltages. The purpose of these precautions is to avoid electric shock.

You can get a severe shock in spite of receiver design features which have been adopted to lessen the danger. All types of high-voltage power supplies now used in television receivers store very little energy in their capacitors. Also, the filter and voltage-divider resistances are high enough to cause, intentionally, very poor voltage regulation. This means that additional current drain, such as your body would draw should you get a shock, almost instantly drops the voltage to a relatively low value.

For anyone in normally good physical condition a shock from a high-voltage power supply circuit will not be fatal nor even dangerous in itself, but it can be most uncomfortable. You are almost certain to jerk your hands and arms violently, and quite probably this will wreck some of the equipment with which you are working.

The first precautions are these. Your hands, your shoes, the work bench, and the floor on which you stand must be dry. Before the plug on the receiver power cord is inserted in a line receptacle you should wipe away any films of moisture or dust from around the points at which voltage is to be measured. Make a preliminary examination to locate all exposed points at which there may be high voltage, preferably with the help of a circuit diagram. Then you can keep clear of all such points except the one to be checked.

Make no test connections with wire having ordinary insulation. Use only flexible test lead which is insulated for 15,000 volts or more. Such lead wire is available from all radio supply houses. When actually making a voltage measurement, with power applied to the receiver, use only one hand. It is safest to keep the other hand in your pocket. Then no shock current can go through the vital organs in your body, and your working hand cannot be hurt to any great extent.

Following is the general procedure when making voltage tests without using a high-voltage probe designed for such work.

1. Pull the power cord plug out of the line receptacle. Do not merely turn off the switch on the receiver.

2. Place the test connections so that nothing need be held by hand. Use insulated spring clips, or wedge the connections in place, or solder them temporarily.

3. Insert the plug in the line receptacle, turn on the receiver switch, and let the receiver warm up for a few minutes. Then make any desired observations or meter readings.

4. Do not remove any test connections until you pull the line plug.

If other tests are to be made at different points, follow the same procedure all over again. It is most important never to touch any test connections while the power plug is inserted in a line receptacle.

If test connections are to be made or changed immediately after a receiver has been in operation it is rather common practice to make sure that all power supply filter capacitances are discharged. This can be done as follows, in a set using a magnetic-deflection picture tube. Remove the high-voltage anode lead by grasping the rubber cover for the terminal. Touch the cable terminal to chassis metal. With a piece of insulated wire bared at both ends hold one end on chassis metal and then touch the other end to the anode connector on the picture tube. In sets having electrostatic-deflection picture tubes the filter capacitors discharge through the voltage divider.

Make it a rule never to handle the lead to a picture tube anode unless the power cord plug is withdrawn from the line receptacle. Remember that the first or focusing anode of a picture tube operating with electrostatic deflection is at a potential around 1,000 volts, or maybe much more. Treat this anode and its connections just like the high-voltage anode or second anode. Do not attempt voltage measurements at the caps of a high-voltage rectifier in any power system, nor at the cap of a sweep amplifier where there is a flyback power supply. There are high alternating or pulsating frequencies at these caps.

One difficulty in making high-voltage measurements is that you won't know just what voltage should be present unless you have access to service data for the particular receiver. In a very general way the typical or average anode voltages for magnetic-deflection picture tubes are as follows.

10-inch tubes	7,500 to 9,000 volts
12-inch tubes	8,000 to 12,000 volts
14-inch tubes	9,000 to 12,000 volts
16-inch tubes	11,000 to 14,000 volts

The anode voltages seldom will be higher than those listed, but still may be satisfactory when as much as 20 per cent lower. Tubes larger than the 16-inch size ordinarily take about the same anode voltage as 16-inch tubes, or not much more. Second-anode voltage for electrostatic tubes usually is between 3,500 and 4,500 volts. Voltage at the first or focusing anode will be around 20 to 25 per cent of second anode voltage.

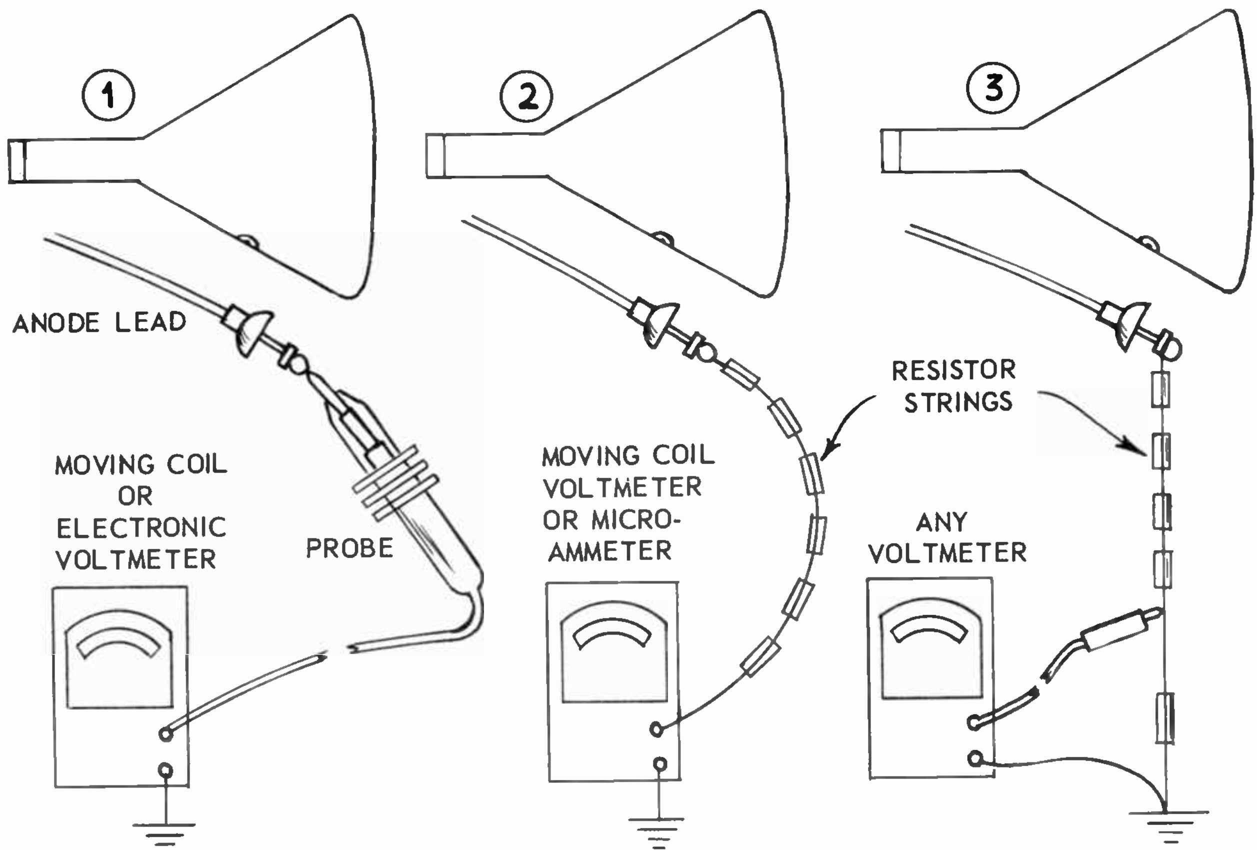


Fig. 62-12. Connections of instruments for measuring voltage at anode cable terminals.

Actual anode voltage varies with anode current and electron beam current. It is maximum with the beam cut off by highly negative grid bias, and is minimum with maximum beam current and maximum screen brightness. As a consequence, the anode voltage varies with setting of the brightness control. You will find maximum voltage with the brightness control turned all the way down, and voltage will be lowest with the brightness control turned all the way up. The difference between minimum and maximum may be from 1,000 to 5,000 volts, with the greater variations found with the larger picture tubes. It is common practice to measure anode voltages with the brightness control turned all the way down.

Fig. 62-12 illustrates methods commonly used for making high-voltage measurements. At 1 is represented either a moving coil voltmeter or an electronic voltmeter equipped with a special high-voltage probe in which are several hundred megohms of resistance to reduce the amount of current forced through the meter by the high voltage from the anode lead. At 2 is represented either a moving coil voltmeter or a microammeter in series with a string of resistors totaling something like 50 megohms to more than 100 megohms. At 3 is either an electronic or moving coil voltmeter connected across one resistor of a string between the anode lead and ground.

With any method of voltage measurement some power and some current are required for operation of the meter. Because of poor regulation (high resistance) in the power supply this current will cause the supply voltage to be somewhat less than without the measuring apparatus connected. The meter may take more or less current than normally taken by the picture tube. Less meter current would allow a higher voltage than normally would exist with the lead connected to the picture tube and the tube operating. Almost always the meter will take more current than the picture tube, and the measured voltage will be less than actual operating voltage.

If you have a high-voltage probe for your voltmeter, the probe lead is connected to a meter terminal, the tip is touched to the anode cable terminal, and the ground terminal of the meter is connected to chassis ground or to B-minus. The tip of the probe extends out from an insulating enclosure back of which are several insulating barrier discs to increase the length of current leakage path. Back of the discs is the insulating handle. All parts of the probe must be free from dust or dirt, and especially free of moisture film. Hold the probe only by its handle, with all fingers back of the barrier discs.

Where you expect a potential of not more than 10,000 volts it is all right to hold the probe in your hand while touching the tip to points at which voltage is to be measured. For higher voltages proceed according to the earlier instructions; pull the line plug, make connections, connect the power line, and then take a reading. Connect the tip of the probe to the cable terminal with an insulated spring clip or fasten them together in any other convenient manner. Support the cable end and the probe on an inch or more of clean, dry wood, or on plastic insulation with the live metal parts at least two inches from any other metal.

Resistor strings as used in diagrams 2 and 3 of Fig. 62-12 always should contain at least five or six separate resistors with their pigtail leads soldered together for the series connection. The resistors should be of the size rated at 2 watts preferably, although the 1-watt size often is used. Always connect the meter at the grounded end of the resistor string, never at the high-voltage end. The cable terminal, the resistors, and all their oonnections must be supported so that they will remain at least two inches from any other metal. Use clean, dry wood or solid plastic insulation for supports.

Not more than 50 microamperes should be drawn through resistors and meter if you expect to make measurements which have any real meaning. This will require 20 megohms resistance for every 1,000 volts to be measured. For example, for measurement of 10,000 volts the total resistance should be 10 times 20 megohms, or 200 megohms. Part of the total resistance will be the internal resistance of the meter. The remainder must be in the resistor string.

A moving coil meter is the type used in service volt-ohm-milliammeters and other types which require no connection to a power line for their operation. The movement of such a meter is pictured at the left in Fig. 62-13. There is a large and strong permanent magnet between the poles of which is supported a small coil of wire. Current from the measured circuit flows through the coil. Reaction between fields of the permanent magnet and the coil causes the coil and its support to rotate. The pointer is attached to the coil support, which is called the armature.

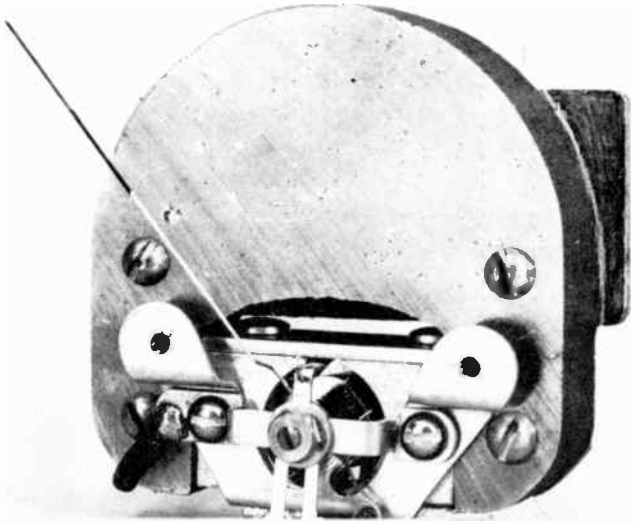


Fig. 62-13. The "movement" of a moving coil meter, which might be used in either a voltmeter or a current meter.

The moving coil of the meter has negligible resistance. In series with the coil is connected enough resistance to allow the measured voltage to force enough current through the coil to deflect the pointer across the meter scale. Sensitivity of the voltmeter is specified as the number of ohms resistance for each volt of reading or of pointer deflection. For high-voltage measurements you should use a meter of no less senstivity than 20,000 ohms per volt. This sensitivity means a meter current of 50 microamperes when the pointer is moved all the way across its scale.

If the full-scale reading of a 20,000 ohms per volt meter is 1,000 volts the total resistance will be 1,000 times 20,000 ohms, which comes to 20,000,000 ohms or 20 megohms. If the meter has a range or a scale going to 5,000 volts the total resistance will be 100 megohms.

Always use the highest range of a moving coil voltmeter. Determine the meter resistance by multiplying together the full-scale volts of this range and 20,000 ohms, assuming a sensitivity of 20,000 ohms per volt. Then determine the total required resistance by multiplying together 20 megohms and the number of thousands of volts you wish to measure with a full-scale reading. If this total is greater than the meter resistance, subtract the meter resistance and use the remainder in the external resistor string.

As an example, assume that you will use a 1,000-volt scale of a 20,000 ohms per volt meter for measuring up to 15,000 volts. Total required resistance will be 15 (thousands) times 20 megohms, or 300 megohms. The meter resistance will be 1,000 (full scale volts) times 20,000 ohms, or 20 megohms. Subtracting this 20 megohms from the required total of 300 megohms leaves 280 megohms for the resistor string.

An electronic voltmeter or vacuum tube voltmeter is a rather intricate piece of apparatus. The internal construction of one such instrument is shown by Fig. 62-14. The internal resistance of electronic voltmeters is not easily computed nor measured, and they are not used with a series resistor string as shown at 2 in Fig. 62-12. Resistance in the high-voltage probe of diagram 1 in that figure is equivalent to the external resistor string, but the probe to be used with any particular electronic voltmeter must be the type recommended by the manufacturer of the meter or the probe.

A microammeter may be used in series with an external resistor string as shown by diagram 2 of Fig. 62-12. The microammeter is a moving coil meter without the added internal resistance that would make it a voltmeter. The internal resistance of the microammeter is entirely negligible, and does not enter into determinations of total required resistance.

The microammeter should have a full-scale range of 100 to 500 microamperes. To determine the total

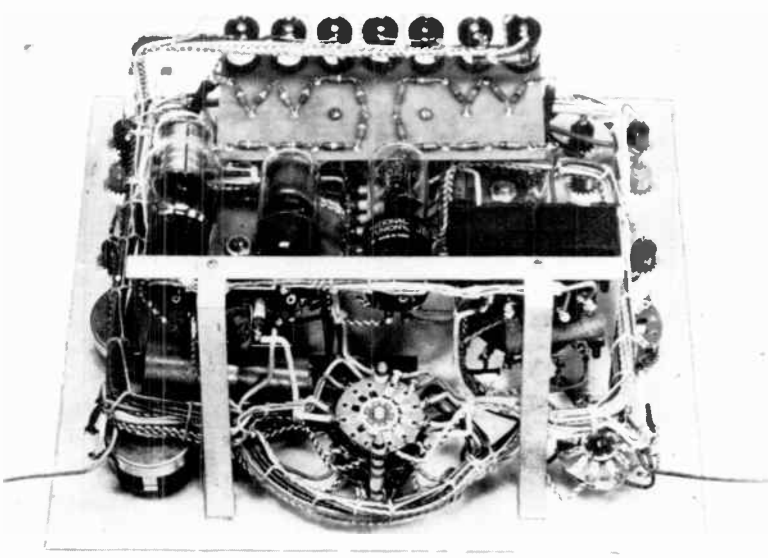


Fig. 62-14. The internal construction of an electronic voltmeter or vacuum tube voltmeter.

required series resistance in megohms, divide the highest voltage that you intend to read by some number less than the full-scale microamperes. For example, to read up to 15,000 volts on a 150 microampere meter you might divide 15,000 by 100, the latter number being less than the full-scale microamperes. The quotient of the division would be 150, which is the number of megohms to be used in the resistor string.

When using the microammeter you translate microamperes of scale reading into volts by multiplying together the indicated number of microamperes and the number of megohms in the resistor string. Were you using 150 megohms resistance, and were the meter to read 60 microamperes, you would multiply 60 by 150 to learn that the measured voltage is 9,000.

When using the measurement method illustrated at 3 in Fig. 62-12 you may use in the resistor string any total resistance which does not allow excessive current to be drawn from the high-voltage power supply or the anode cable. The resistor at the grounded end of the string, the one across which the meter is connected, should be of only one megohm. The meter resistance must be much greater than the current-carrying resistor across which connected.

Total resistance from the anode cable to ground, and also the resistance across which the meter is connected, must be measured accurately. Then voltage at the anode cable terminal is determined with the following formula.

$$\text{Anode volts} = \text{meter reading, volts} \times \frac{\text{total resistance, cable to ground}}{\text{resistance across meter}}$$

Resistance must be in the same unit, ohms or megohms in both places.

Here is an example. Total resistance, 26.2 megohms. Resistance across meter, 1.2 megohms. Meter reading, 355 volts.

$$\text{Anode volts} = 355 \times \frac{26.2}{1.2} = 355 \times 21.8 \text{ (approx.)} = 7739 \text{ or } 7740$$

TELEVISION

LESSON NO. 63

POWER SUPPLIES - II

Training Your Mind...
C.T.I.
Perfecting Your Skill

DO NOT TEAR – CUT ALONG THIS LINE AND SEND IN FOR GRADING.

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind
C.T.I.
Perfecting Your Skill

COMMERCIAL
TRADES
INSTITUTE

Chicago, Ill.

LESSON NO. 63

POWER SUPPLIES - II

R-F HIGH-VOLTAGE POWER SUPPLIES. An r-f type of high-voltage supply first changes d-c energy to oscillating voltage and current at some rather low radio frequency, then steps up the oscillating voltage, rectifies this voltage, and delivers the resulting direct current at high voltage to the picture tube anode or anodes. The essential tubes are an oscillator and a rectifier.

R-f high-voltage power supplies can be used with either magnetic-deflection or electrostatic-deflection picture tubes, for these power systems do not depend for their operation on voltage or current pulses from any other circuits. The flyback power supply can be used only with picture tubes employing magnetic deflection, for this type of supply depends primarily on sudden changes of current in the deflecting coil circuit.

R-f power supplies have been used with most of the electrostatic picture tubes during all the years in which television has been popular, and now are found in all the new sets which contain 7-inch electrostatic picture tubes. R-f power supplies are sometimes used also in receivers having magnetic-deflection picture tubes, and can be used in all sizes from 10-inch to 16-inch, operating at 8,000 to 14,000 anode volts.

Fig. 63-1 shows fairly typical circuit connections for an r-f power supply. Direct voltage and current for the oscillator plate and screen come through the input filter from the low-voltage B supply of the receiver. The rectifier is a halfwave high-voltage type such as used also in flyback power supplies. Between the oscillator and rectifier is connected the step-up transformer, with its primary winding in the oscillator plate circuit and its secondary in the rectifier plate-to-ground circuit. On the transformer are also one or two turns of insulated wire for furnishing voltage and current which heat the rectifier filament.

Frequency of oscillation is determined by the self-resonant frequency of the transformer secondary circuit, or by the combination of inductance, distributed capacitance, tube capacitance, and stray capacitances in this circuit. Actual oscillation frequencies range all the way from 90 to 285 kilocycles in various receivers. Frequencies are no higher than 300 kilocycles because the rectifier tube nearly always used, a 1B3-GT, is not designed for operation at any higher frequencies.

As mentioned, the oscillator tube most often is a beam power type. A few power pentodes are used, as are also a few power triodes. Quite a few r-f power supplies use twin triodes with all their elements in parallel to act like a single triode of greater power handling ability. Beam power tubes and pentodes oc-

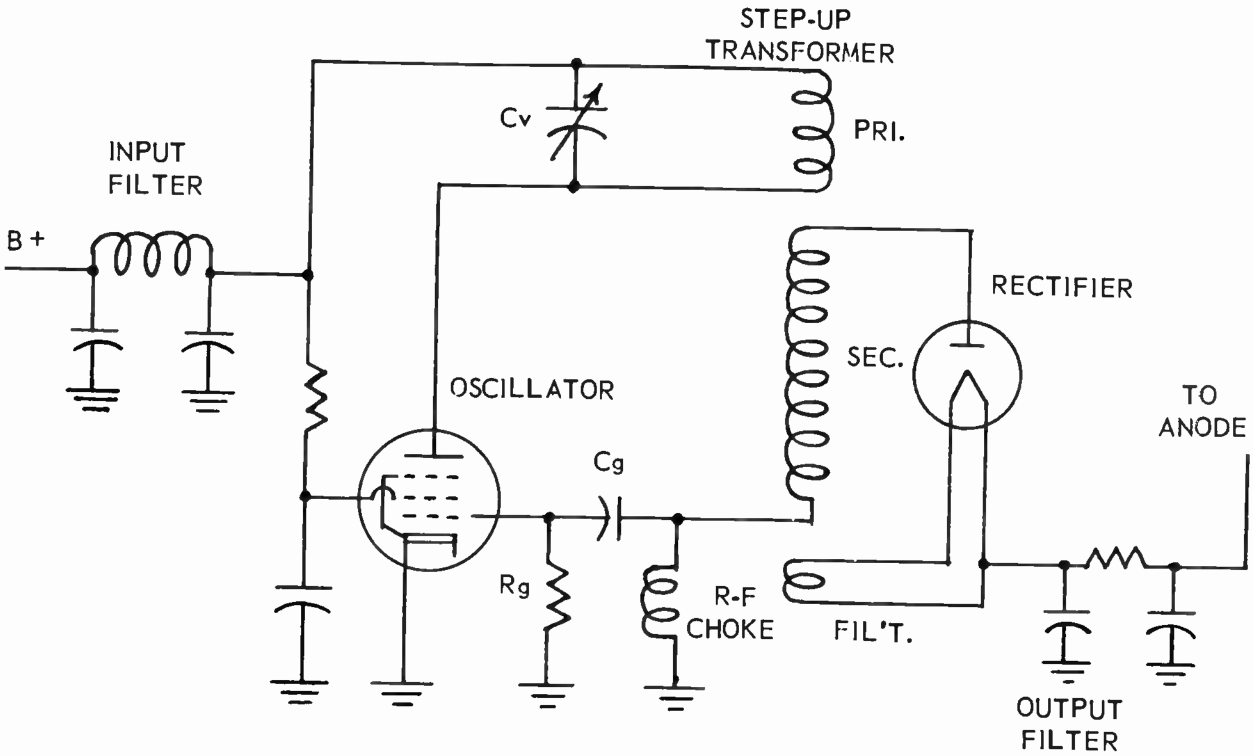


Fig. 63-1. Circuits of an r-f power supply with tuned plate winding.

casionally are connected as triodes, with their plates and screens directly connected together at the socket. Most oscillators of the various types mentioned are of GT or G sizes with octal bases, but again they may be miniatures. To increase the power handling capacity it is sometimes the practice to use two beam power tubes with their elements connected in parallel.

The high-voltage transformer is of the air-core type, with windings arranged as shown by Fig. 63-2. Energy losses at the oscillation frequency are less than they would be with iron-core construction. Nearest the bottom of the tubular form is the one- or two-turn winding that furnishes heating power for the rectifier filament. Then comes the primary winding, which is connected into the oscillator plate circuit. Above the primary is the high-voltage secondary winding, shown here as constructed of six series connected pies.

The primary winding usually is a duolateral type. Each of the secondary pies is a duolateral or honeycomb coil. This method of forming the secondary reduces its distributed capacitance and allows the self-resonant operating frequency to be up in the radio-frequency range even with the large inductance which results from the great number of secondary turns needed for the great step-up ratio.

In a typical power supply for a transformerless receiver the d-c B voltage to the rectifier plate may be on the order of 125 to 135 volts. Peak-to-peak voltage at the oscillator plate may be 300 to 400 volts, and rectified d-c output to the picture tube anode circuit may be about 9,000 volts. Such operation calls for transformer turns ratios of between 30 and 50 to 1.

6 Above the entire pie winding which is the secondary of the transformer in Fig. 63-2 is still another relatively small winding. This is the tickler coil furnishing feedback energy to the oscillator grid. The additional tickler coil would be used in a power supply circuit such as shown by Fig. 63-3. This diagram illustrates also how the two sections of a twin triode tube may be connected in parallel to act as a heavy-duty oscillator.

One of the most noticeable differences between r-f power supply circuits is the method of feeding energy from the output back to the input of the oscillator. With the tickler feedback of Fig. 63-3 the tickler coil is in the magnetic field of the secondary winding or plate winding and these two windings are inductively coupled. One end of the tickler coil is connected to the oscillator grid or grids. The other end is connected to the oscillator cathode. The oscillator is negatively biased by the grid-leak method. The grid capacitor Cg and grid resistor Rg are connected in parallel with each other and in series with the grid circuit.

Looking back at Fig. 63-1 you will see that the oscillator again is grid-leak biased by means of grid capacitor Cg and grid resistor Rg. Energy feedback from plate to grid of the oscillator is through a type of coupling commonly used for amplifier circuits. The oscillator plate winding or transformer primary induces high-frequency emf and current in the secondary. The high-frequency secondary current cannot complete its circuit to ground through the r-f choke, which has very great impedance at radio frequencies, but this current must go through capacitor Cg and resistor Rg. The accompanying high-frequency voltage drop across resistor Rg is applied between grid and cathode of the oscillator. Thus we have a transfer of high-frequency energy from the plate circuit back to the grid circuit of the oscillator.

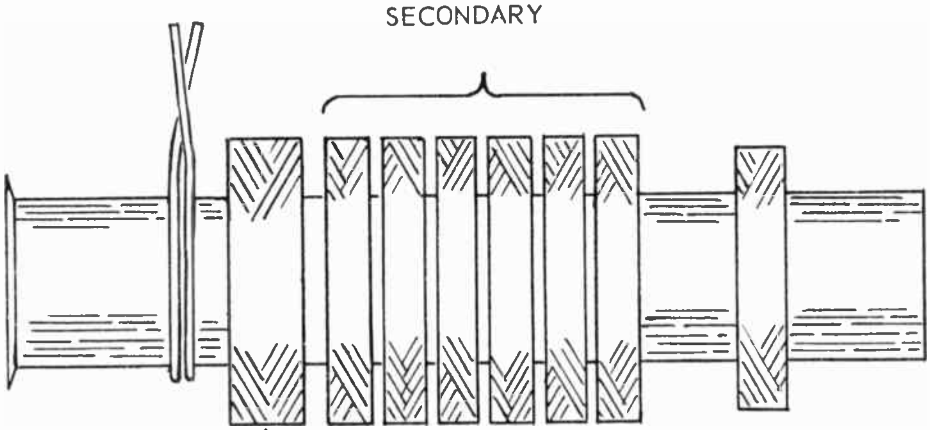


Fig. 63-2. Arrangement of windings on air-core transformer for r-f power supply.

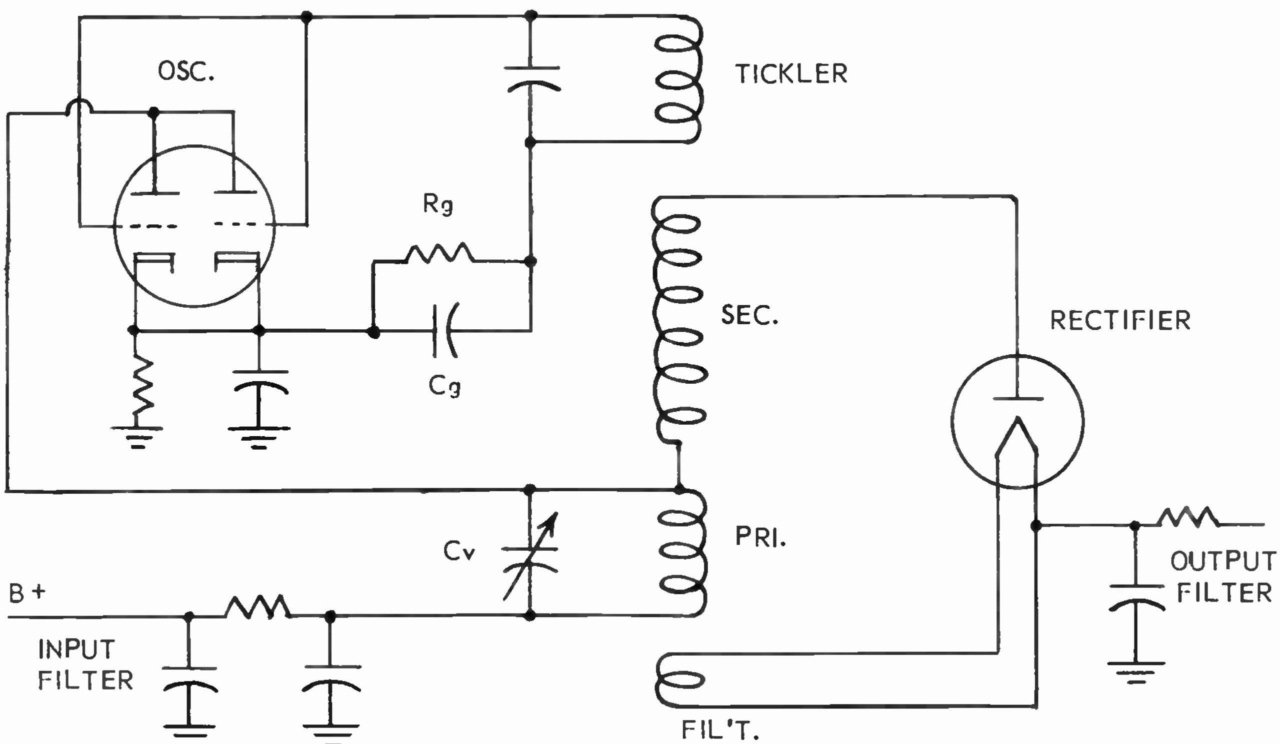


Fig. 63-3. An r-f high-voltage power supply with tickler feedback for oscillator grid.

Another method of energy feedback for the oscillator is illustrated by Fig. 63-4. The oscillator grid is connected to a feedback ring placed around the outside of the glass envelope of the high-voltage rectifier. There is capacitive coupling from the high-frequency voltage on the rectifier plate to the ring and back to the oscillator grid. What amounts to a feedback capacitor is composed of the metal of the oscillator plate and the metal of the external ring, with the glass envelope and vacuum space acting as dielectric. The oscillator is grid-leak biased, with the feedback capacitance at the rectifier taking the place of the usual grid capacitor and with resistor Rg acting as the grid leak resistance.

The feedback ring usually is a coiled spring of small diameter formed into a ring that slides down over the outside of the rectifier tube. Fig. 63-5 shows such a ring on a 1B3-GT rectifier. The wire that attaches to this ring runs in behind the transformer and down through an insulating bushing to the grid lug on the oscillator socket.

In the r-f power supply circuits of Figs. 63-1 and 63-3 there are adjustable capacitors Cv across the oscillator plate winding or the primary winding of the transformer. These capacitors are adjusted to make the resonant frequency of the plate winding the same or very nearly the same as the self-resonant frequency of the secondary. When the two resonant frequencies are alike there is greatest possible transfer of power from the oscillator into the rectifier circuit. This increases the amplitude of high-frequency secondary voltage being rectified, and increases the direct voltage delivered to picture tube anode circuits.

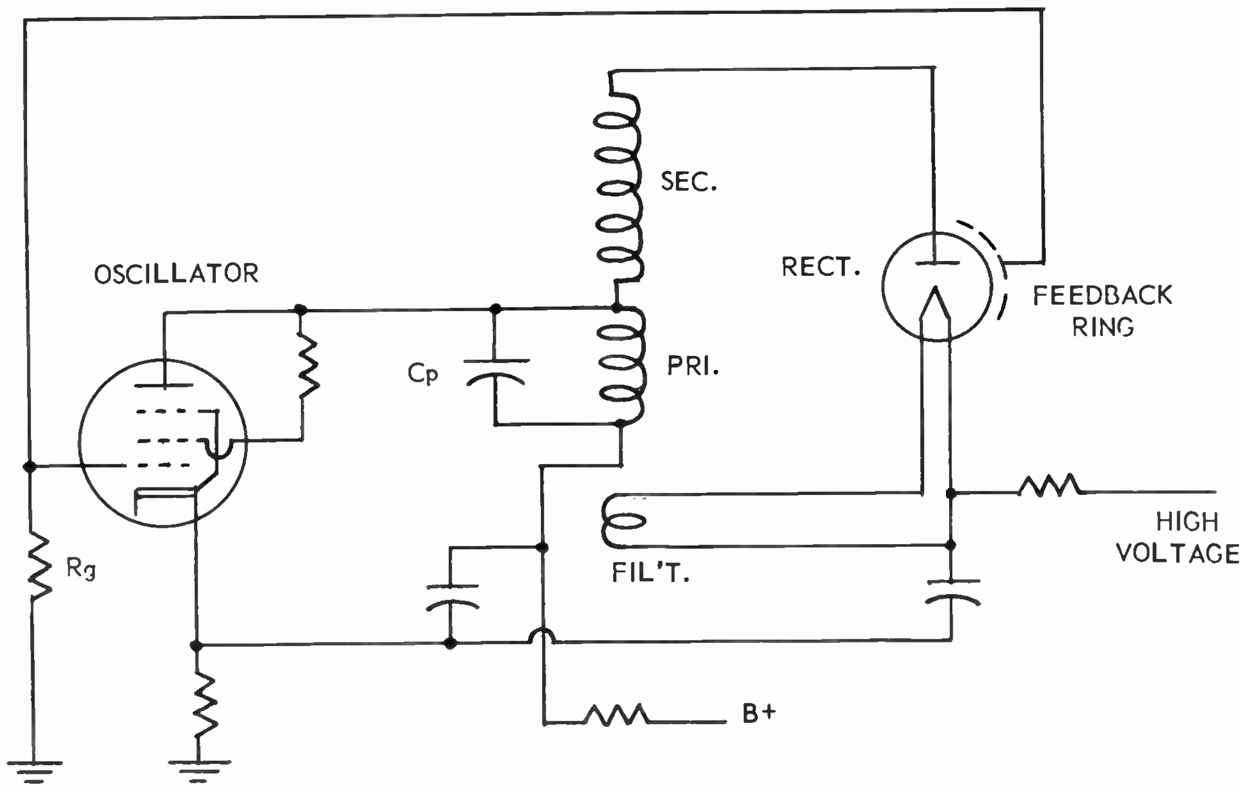


Fig. 63-4. Circuits of an r-f power supply with feedback from ring on rectifier tube to oscillator grid.

Oscillator plate tuning usually is by means of a trimmer type capacitor with mica dielectric and a screw adjustment. The required capacitance depends on inductance of the winding and the frequency to which it is to be tuned, but often is variable between about 400 and 1,500 mmf. In a few cases the plate winding is tuned by means of a movable molded iron core on a threaded stud having a screw driver slot.

In Fig. 63-4 the plate or primary winding on the transformer is tuned to approximate resonance with the secondary frequency by means of a fixed capacitor Cp. With such a design there is no special adjustment for output voltage, although this voltage is affected to some extent by the position of the feedback ring on the rectifier tube envelope.

As a general rule all of the r-f power supply parts shown in our circuit diagrams are enclosed within a metal shield. There are only four external leads, one for the B plus voltage and current from the low-voltage supply and another for the high voltage going to picture tube anodes or other circuits requiring high voltage. The remaining two are for oscillator filament voltage. In some receivers the r-f oscillator is outside the shield that encloses the transformer and rectifier tube.

1 The purpose of the input filter is to prevent high-frequency voltages and currents from getting back into the low-voltage B supply wiring from the power supply. This filter ordinarily consists of two capacitors of about 500 mmf each and of either a choke coil or a resistor in series with the line between the capacitors.

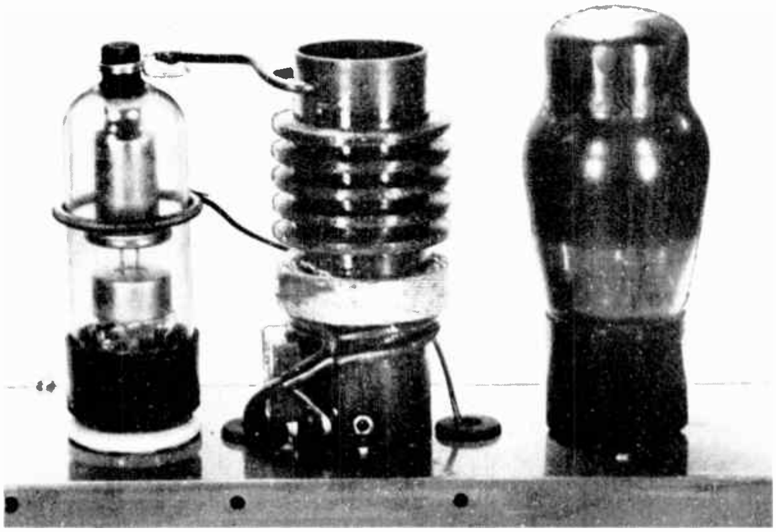


Fig. 63-5. R-f high-voltage power supply with feedback ring on rectifier.

2

The purpose of the output filter is to smooth the rectified direct current and voltage going to the anode circuits or other high-voltage circuits. This filter consists of one or more capacitors to ground and usually of a single resistor. Additional filter capacitance may be furnished by the external coating on the picture

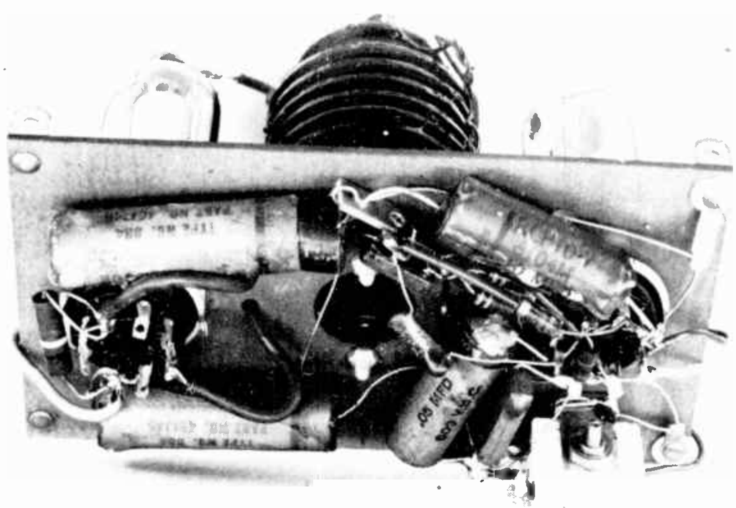


Fig. 63-6. Capacitors and resistors under the tube shelf of an r-f power supply unit.

tube. Fig. 63-6 is a picture of filtering and voltage dropping resistors and capacitors mounted underneath the tube and transformer shelf in a typical r-f power supply. At the lower right you can see the voltage adjusting trimmer capacitor. Everything shown in this picture is enclosed within the metal shielding cover.

The power supply shield may or may not be directly grounded to the chassis. When the shield is not directly grounded it usually has what is called an r-f ground through a fixed capacitor between shield and chassis. There will also be a resistor of 100,000 ohms or more between shield and chassis, this for the purpose of grounding the electric charges which tend to form on the shield because of radiation from the enclosed parts.

When an r-f power supply is used in a receiver having an electrostatic picture tube the load on the power supply will be of the general type shown by Fig. 63-7. At the outer end of the filter system, connected to the filament of the high-voltage rectifier, we may have about 4,500 d-c volts. This voltage goes to the second anode or high-voltage anode of the picture tube.

Between the high-voltage output of the power supply filter and ground there always are resistances totaling many megohms. These resistances are in the centering control system, the focusing control potentiometer, and the voltage dropping resistors which maintain correct potentials on the picture tube elements. In the load illustrated there is a total resistance of 30.8 megohms, of which 1.8 megohms is the combined resistance of the paralleled resistors in the centering controls. Current through the voltage divider portion of the power supply load usually is on the order of 150 to 175 microamperes.

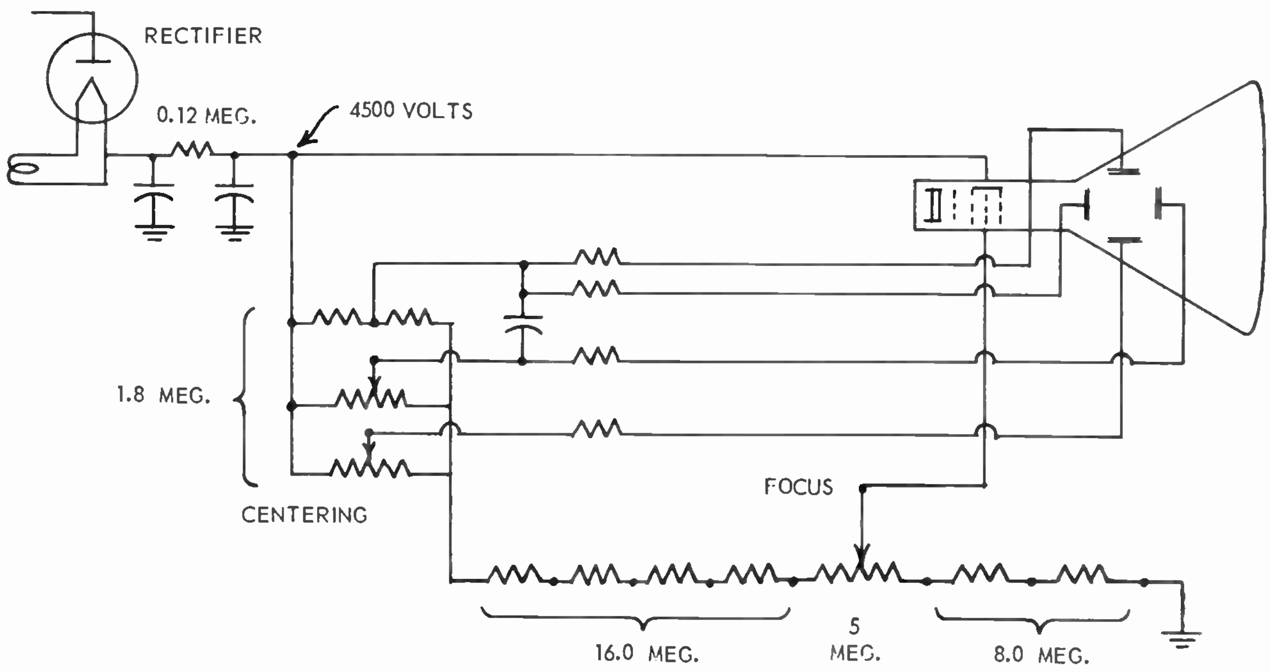


Fig. 63-7. Typical load circuits on an r-f high-voltage supply for an electrostatic picture tube.

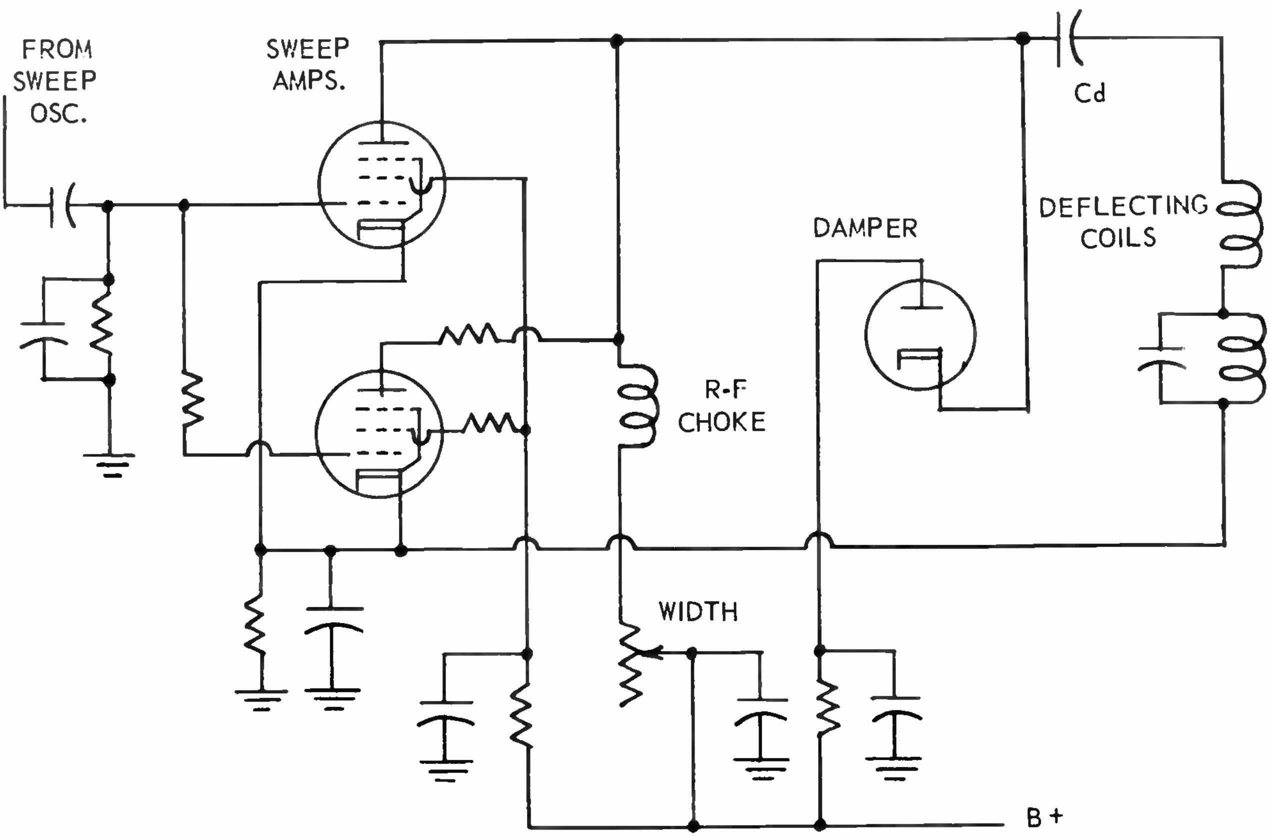


Fig. 63-8. One type of horizontal sweep circuit used in a receiver having an r-f high voltage power supply.

You will note that the larger resistances, such as those of 16 and 8 megohms, are each made up of several smaller resistance units connected in series. This divides the voltage drop among all the resistors and lessens the danger of voltage breakdown on any one. Because of the very small current the total power dissipation in all the resistors together is less than one watt, this in spite of the high total resistance. To further lessen the chances of breakdown, and to insure that the resistors remain cool during operation, each unit ordinarily is of 1-watt rating.

When an r-f high-voltage power supply is used in a receiver having magnetic-deflection picture tube there is no need for having a "flyback" type output transformer between the horizontal sweep amplifier and the deflecting coils. In such receivers you will find horizontal sweep circuits generally similar to the one shown by Fig. 63-8. The two sweep amplifiers have all their elements connected in parallel to act like a single tube of greater power-handling ability. The small resistors in series with the plate, screen, and grid of the lower amplifier are of about 50 to 100 ohms each. They help to suppress tendency toward parasitic oscillation in the tube circuits and also help equalize the load currents in the two plate circuits.

The plates of the two sweep amplifiers are connected through capacitor Cd to one end of the deflecting coils. The other end of the deflecting coils connects back to the amplifier cathodes. In this circuit there is no sweep output transformer. Note that the deflection frequency or horizontal line frequency is kept within the deflecting coil circuit and kept out of the B-plus connections by means of an r-f choke. The width control or horizontal size control varies the plate voltage applied to the amplifiers. The r-f choke offers no reactance to direct voltage and current for the plates. In other receivers the width control is in series with the amplifier screens, and varies the amplification or the transconductance by varying the screens, and varies the amplification or the transconductance by varying the screen voltage.

The damper tube in the circuit of Fig. 63-8 performs the same function as in any other magnetic deflection system. The damper plate is connected to B plus, tending to make the tube conductive, but the cathode also is connected to B plus through the r-f choke and the width control. The damper actually does become conductive when there is a sudden pulse of negative voltage on its cathode as plate current is cut off in the amplifiers. Then the damper places a heavy load on the deflecting coil circuit and causes the first alternation of self-oscillation to die out during the first part of the sawtooth wave, just as in other damper circuits which have been explained in detail.

Another horizontal magnetic deflection circuit used with an r-f high-voltage supply is shown by Fig. 63-9. Here there is a single sweep amplifier tube. Width is controlled by simultaneous variation of both plate and screen voltages on the amplifier. Choke coils L-L-L isolate the horizontal line frequency of the deflection circuits from the B plus low-voltage power supply lines. Centering is by means of direct current taken in either polarity from a center-tapped potentiometer inserted in the B plus line.

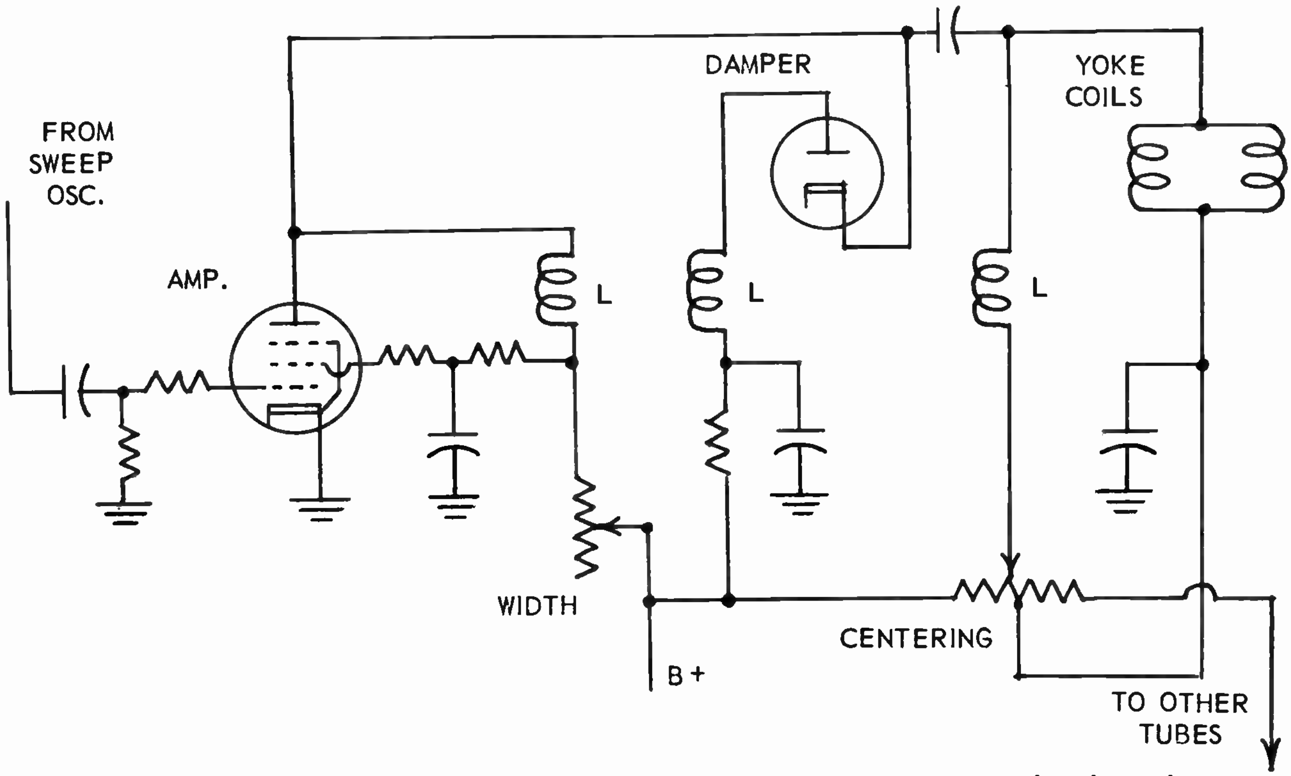


Fig. 63-9. Horizontal sweep circuit using a single amplifier tube and choke isolation from low-voltage B-power lines.

The power supply and deflection circuits which have been shown in this lesson illustrate various features which often are combined as shown but which also may be used in different combinations. We have shown a pentode oscillator with capacitor and grid resistor feedback, a twin triode with tickler feedback, and a triode-connected pentode with a feedback ring on the rectifier. These particular types of tubes are not always used with the feedbacks of the diagrams, the combinations may be quite different. This is true also of the filters and other circuit elements.

HIGH-VOLTAGE ADJUSTMENT. Our only power supply diagram which shows no tuning adjustment for the oscillator plate circuit is the one with the feedback ring on the rectifier. There will be the highest oscillation voltage and highest d-c output voltage when this ring is somewhere near the center of the long cup-shaped plate or anode of the rectifier. Output voltage usually will drop to a slight extent when the ring is moved to a point level with the top of the plate, and will do likewise when the ring is level with the rim around the bottom of the plate. Moving the ring still farther down toward the internal shield will cause a great decrease of output voltage, and the output may cease altogether as oscillation stops due to insufficient feedback to the oscillator grid.

4 The following instructions apply to power supplies in which there is an adjustable capacitor or a movable coil core for tuning the oscillator plate circuit. As stated before, the closer the plate circuit frequency is brought to the natural frequency of the secondary winding the higher will be the d-c output voltage. However, when tuning is to exact equality of frequencies or is adjusted for highest possible output the oscillator circuit is likely to be unstable and to allow voltage fluctuation. For this reason the plate circuit usually is left tuned slightly off resonance, which allows some reduction of voltage but insures a more constant voltage.

You should keep in mind that any increase of current through the output filter resistance and any following voltage-divider resistances will increase the voltage drop across these resistances and will lower the voltage measured at such points as the picture tube anodes. Beam current and anode current are increased by increasing the brightness on the screen, and are reduced by less brightness. Consequently, you will find the highest anode voltages with the brightness control turned all the way down, and minimum voltages with this control turned to maximum. The contrast control also varies the current and anode voltage to some extent because this control can vary the average brightness. Between minimum and maximum brightness the increase of anode voltage on a magnetic-deflection picture tube may be 1,000 volts or more.

Trimmer or core adjustment screws always are accessible from the outside of the power supply shield, so that voltage may be varied without removing the shield. To avoid incorrect tuning, and more especially as a safety precaution, you should invariably use a non-metallic screw driver for this adjustment. A metallic tool might touch parts of the high-voltage circuit, with unpleasant results.

If you have an electronic voltmeter equipped with a high-voltage prod, one insulated for at least 15,000 volts, it is possible to make direct measurements between the picture tube anode connector or terminal and either chassis ground or B-minus. Observe all the usual precautions for high-voltage measurements. It is possible also to employ any of the methods for high-voltage measurement which were described in a preceding lesson. As mentioned there, do not attempt measurements at the cap of the high-voltage rectifier, where there are strong potentials at radio frequency.

The high voltage may be adjusted as follows.

1. Turn both the brightness and the contrast controls to their lowest settings.

2. Turn the trimmer capacitor screw all the way in, clockwise, for maximum capacitance and lowest tuned frequency. If the position of a movable core can be determined, turn it all the way into the coil winding.

3. While watching whatever type of voltage indicator is being used, turn the trimmer capacitor or coil core screw very slowly in the direction that raises the frequency. Decrease the capacitance or the inductance. Leave the adjustment at the point giving the desired voltage.

4. When working with a trimmer capacitor it is possibly better practice to turn the adjustment a little farther out than the point of desired voltage, or turn it for a slightly higher voltage, then turn back to the desired voltage. This helps insure that the trimmer plates will remain as adjusted, without tending to spring farther apart.

Voltages at the socket lugs of the power supply oscillator may be made without using high-voltage test methods. With transformerless receivers the B plus voltage furnished to the power unit usually is about 125 volts. With other types of receivers this voltage usually is between 200 and 350. If the oscillator is operating correctly its grid voltage will be highly negative with reference to ground or to B-minus. About 125 volts negative is fairly typical for the oscillator grid. If this voltage is measured with other than an electronic voltmeter the reading will be much lower. The meter loading may stop oscillation, whereupon the reading will be only a few volts negative.

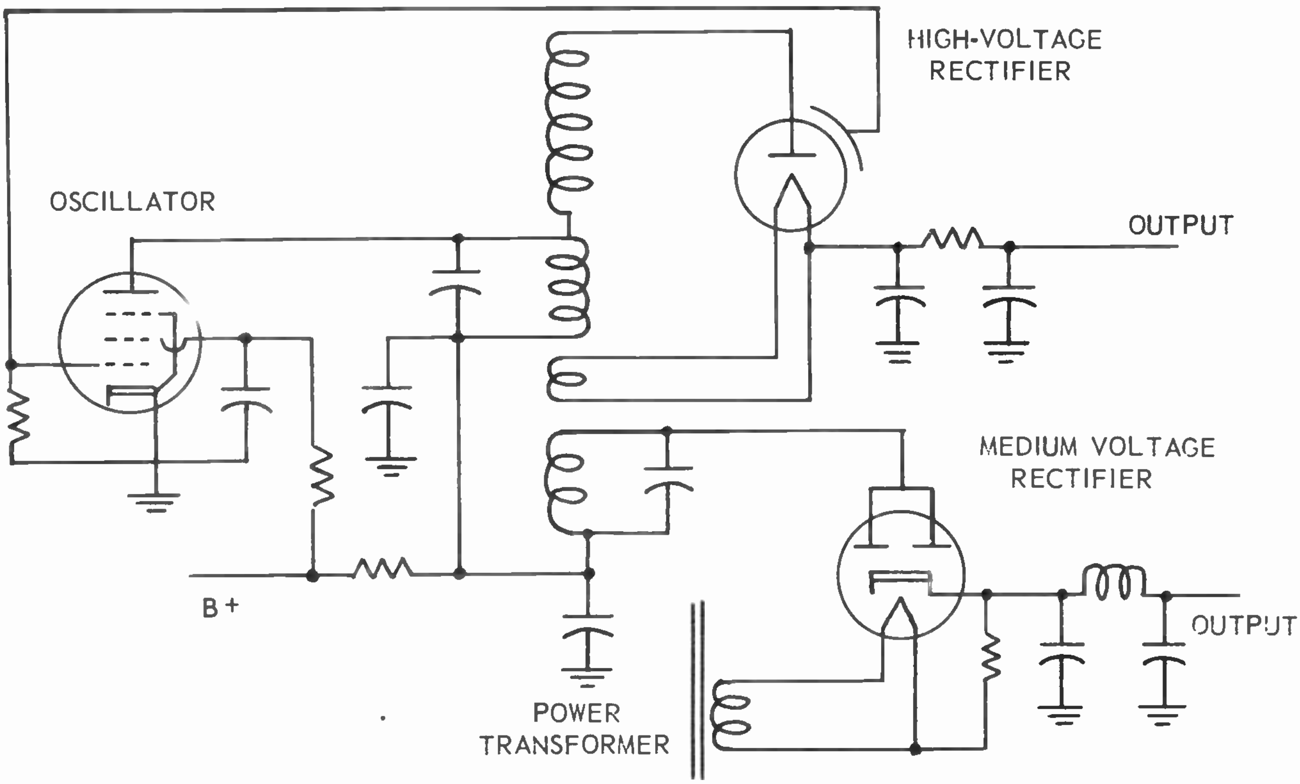


Fig. 63-10. A medium-voltage transformer winding and rectifier added to an r-f high-voltage power supply.

1 *DUAL VOLTAGE R-F POWER SUPPLY.* Fig. 63-10 shows circuits for an r-f power supply system which furnishes the usual high voltage for picture tube anodes and also a lower voltagefor plates and screens of other tubes. The circuits for the r-f oscillator and the high-voltage rectifier are like some of those shown earlier.

On the transformer is a fourth winding, shown below the other three in the diagram. The r-f field induces an emf in this winding just as in the othersecondaries. The radio-frequency voltage is rectified by a separate medium-voltage rectifier whose d-c output goes through a choke-capacitor filter to plate and screen circuits. The medium voltage rectifier is a heater-cathode type, with its heater supplied with power from a separate insulated winding on the main power transformer.

This dual voltage system is used in transformerless receivers which employ an r-f high-voltage power supply. When there is no voltage doubling in the low-voltage B supply of such receivers the maximum B voltage is about 125. From the additional rectifier of Fig. 63-10 we may obtain 500 or more d-c volts which, after dropping in resistors leading to plates and screens of various tubes, still will furnish 300 or more volts at these plates and screens. Thus we obtain, by adding one rectifier tube, a voltage step-up which would require a voltage tripling system in the B supply that is fed from the a-c power line.

PULSE RECTIFIERS. Fig. 63-11 shows how an additional B voltage may be obtained from a magnetic deflection system in a receiver that uses a sawtooth or pulse type high-voltage supply, instead of a flyback type. The deflection circuits are generally similar to those of Figs. 63-8 and 63-9, except that here are shown three paralleled sweep amplifier tubes instead of the one or two sweep amplifiers of those other circuits. The damper tube is connected similarly to dampers in all circuits of this general type and carries out the functions of damping in the usual way.

A tapped inductor L is connected across the deflecting coils. During each retrace period there appears across this inductor and the deflecting coils a pulse of about 1,000 volts. Pulses of this kind were explained in connection with operation of flyback high-voltage supplies. From the tap on inductor L is taken a small fraction of the pulse voltage, usually something between 110 and 130 volts. This voltage is applied to the plate of the pulse rectifier tube. This tube rectifies the applied voltage, in which is an alternating component as well as the pulses. The rectified voltage from the cathode of the pulse amplifier is added to the regular B plus voltage at point a on the diagram. The sum of the two voltages forms the boosted output voltage.

In transformerless receivers the low-voltage B supply furnishes about 125 d-c volts. Voltage added by the pulse rectifier brings the boosted value to between 225 and 250 volts. The value of the added voltage depends, of course, on the position of the tap on inductor L.

SIXTY-CYCLE HIGH-VOLTAGE SUPPLIES. In very early television receivers, those built around 1940 to 1943, the high-voltage power supplies for picture tube anodes utilized the same principles asthecommon type of low-voltage B supply, and operated at power line frequency, usually 60 cycles per second. This practice was continued in a few models until as late as 1947.

The 60-cycle high-voltage power supply most often is fed from a separate power transformer and consists of parts shown at 1 in Fig. 63-12. The rectifier tube is a half-wave filament-cathode type designed for high-voltage operation. The filter between the rectifier cathode and the lead to the picture tube anode usually contains one series resistor of about a half-megohm and either one or two capacitors to ground.

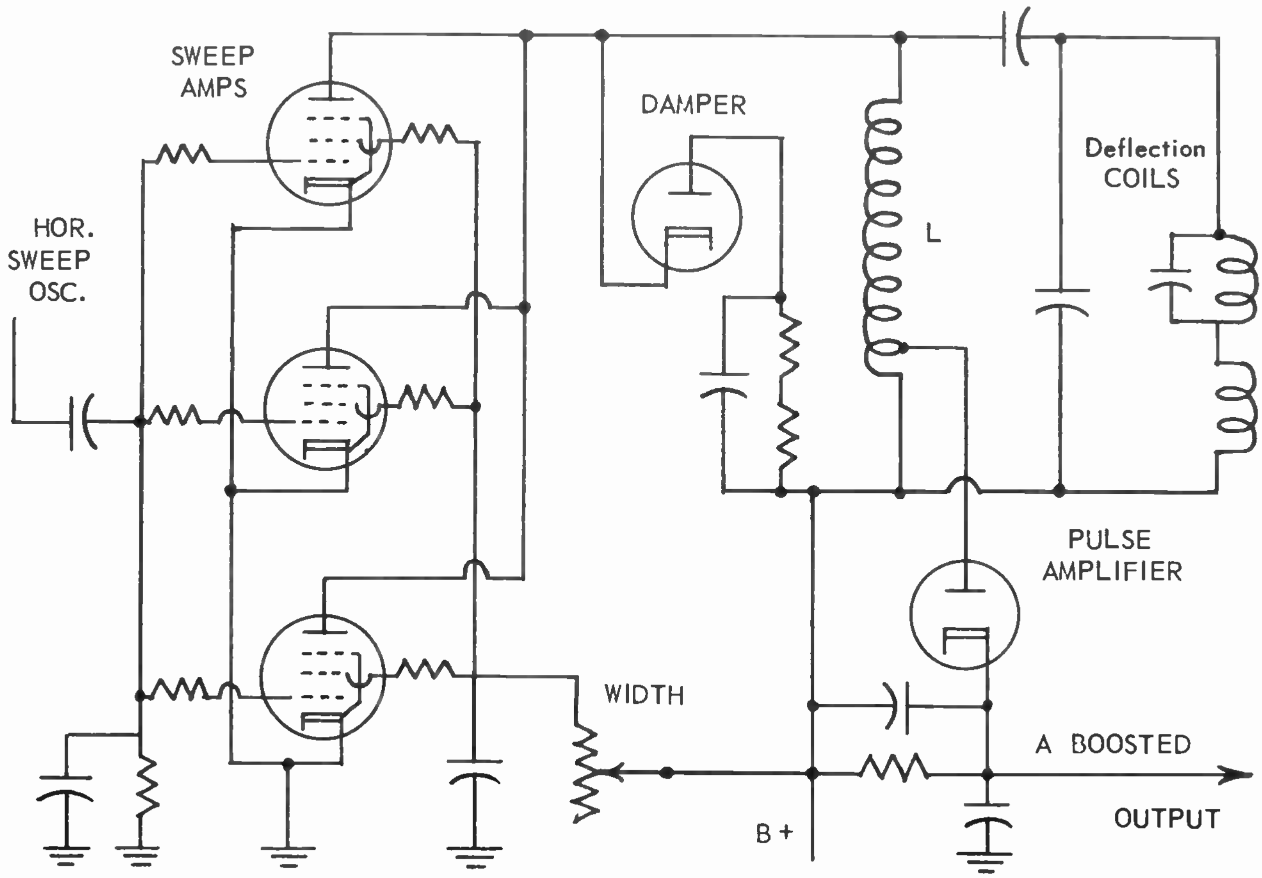


Fig. 63-11. A pulse amplifier for obtaining a boosted B-voltage from a horizontal sweep circuit without a flyback transformer.

For picture tubes having electrostatic focus the filter output feeds also a string of voltage-divider resistors ending at ground. Near the ground end of this divider system is the focus control potentiometer.

Because of the low (60 cycle) frequency at which these power supplies operate, the filter capacitors have values most often between 0.02 and 0.05 mfd, but in a few cases as great as 0.20 mfd. These capacitances are 40 to 400 times as great as used with the relatively high frequencies of present day flyback and r-f high-voltage supplies. The large capacitances, charged to high voltages, store enough energy to give a possibly fatal shock should a person come in bodily contact with their terminal connections or the anode lead connector. This very real danger was the chief reason for discontinuing the use of low-frequency high-voltage power supplies in all modern television receivers.

Some 60-cycle power supplies are connected as shown at 2 in Fig. 63-12. The power transformer has windings for both high- and low-voltage B supplies. Instead of grounding the negative end of the high-voltage winding it is connected to the positive output of the low-voltage B supply. Thus the lower B-voltage is added to the high voltage. Resistor R may be used to limit the current and prevent burnout of the high-voltage winding in case of overload.

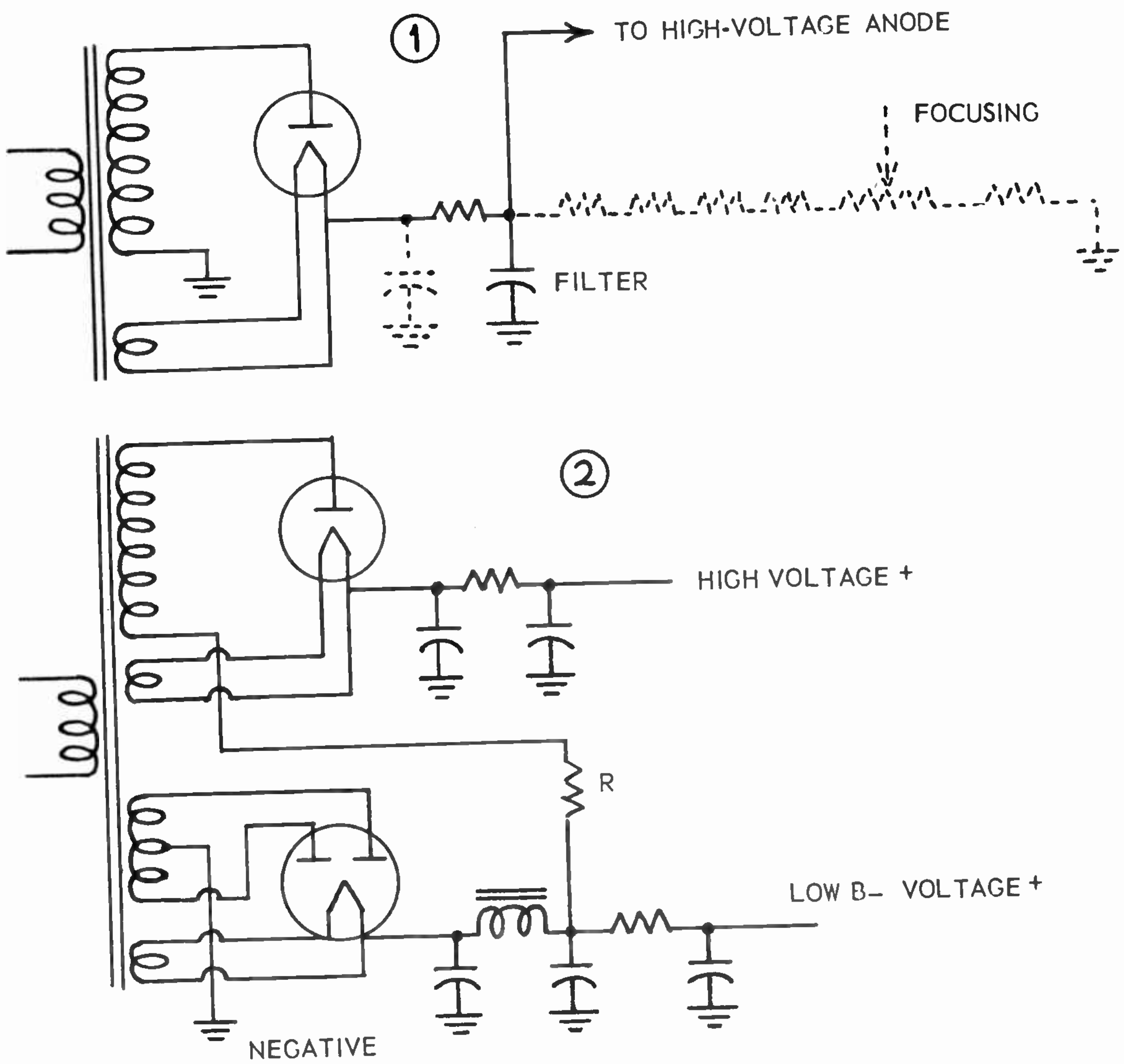


Fig. 63-12. High-voltage power supplies operating at power line frequency.

HANDLING THE PICTURE TUBE. Having learned a good deal about picture tubes and the voltages applied to them we may proceed now to the problem of removing and replacing these tubes during service operations. In the majority of receivers the picture tube is supported wholly on the chassis, and when the chassis is taken out of the cabinet the picture tube comes along. There are, however, many receivers in which the picture tube is supported on the inside of the cabinet. In a few cases the tube is supported on the chassis and also is clamped to the inside of the cabinet.

When the picture tube is supported entirely from the chassis, the arrangement of parts is shown in a general way by Fig. 63-13. The face end of the tube rests on a padded cradle at the front end of the chassis, and is held down on the cradle by a strap. This mounting strap may be of metal lined with felt, or may be of woven fabric, or of some semi-elastic material. If the picture tube is a metal-cone type the outer edge of the cone is enclosed by a plastic insulating ring held in place by the mounting strap and possibly by an additional rubber band.

The rear end of the picture tube flare is supported by a cushion, usually rubber, which is carried by the bracket that mounts the deflecting yoke and the focus coil or magnet. The inner openings through the yoke and coil or magnet do not touch the tube neck. The rear part of the neck carries the ion trap magnet, if used. On the tube base is the socket.

The mounting strap is held in some manner which allows it to be loosened and then raised or moved forward or back when the picture tube is to be dismounted from the chassis. Usually there is some means for adjustment of strap tension. On either side of a metal strap may be a clamping screw with a thumb nut, or both ends of the strap may have threaded members which pass through holes in the chassis or in brackets, where the ends are held and tightened by nuts. One end of a fabric strap usually is clamped or otherwise fastened securely to the chassis, while the other end is held to the chassis or a bracket by a clamp which may be loosened for tube removal. Otherwise there may be an adjustable screw clamp anywhere along the strap. Straps sometimes are held by hooks on one or both ends.

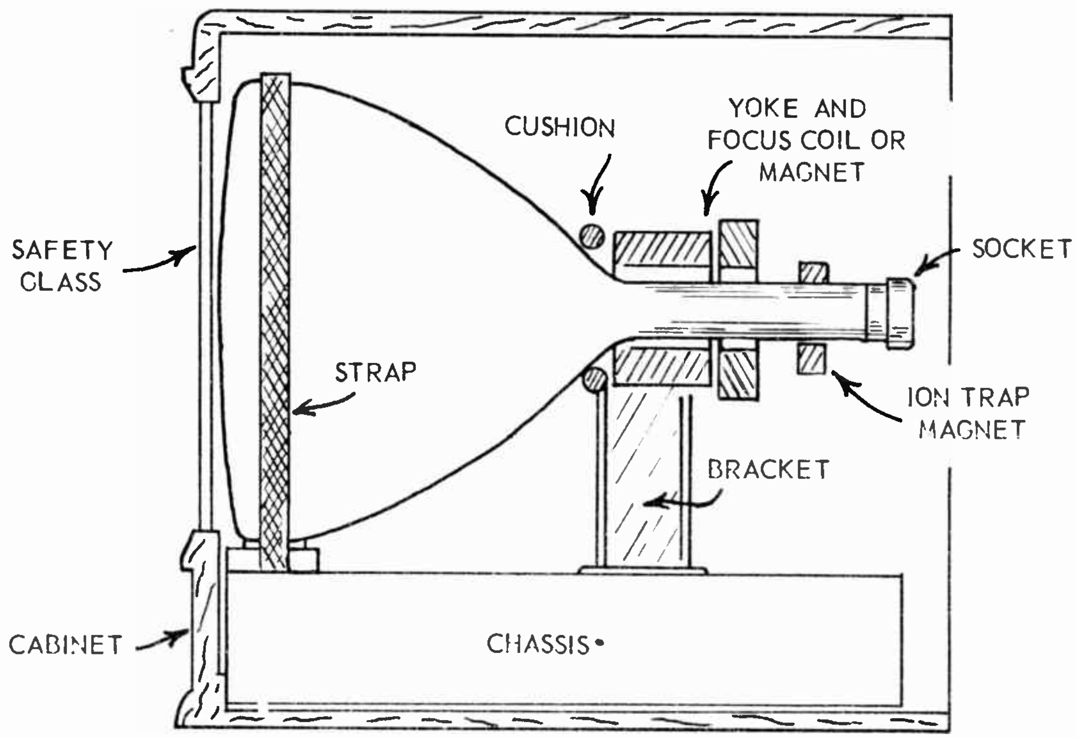


Fig. 63-13. A picture tube mounted on and wholly supported by the chassis.

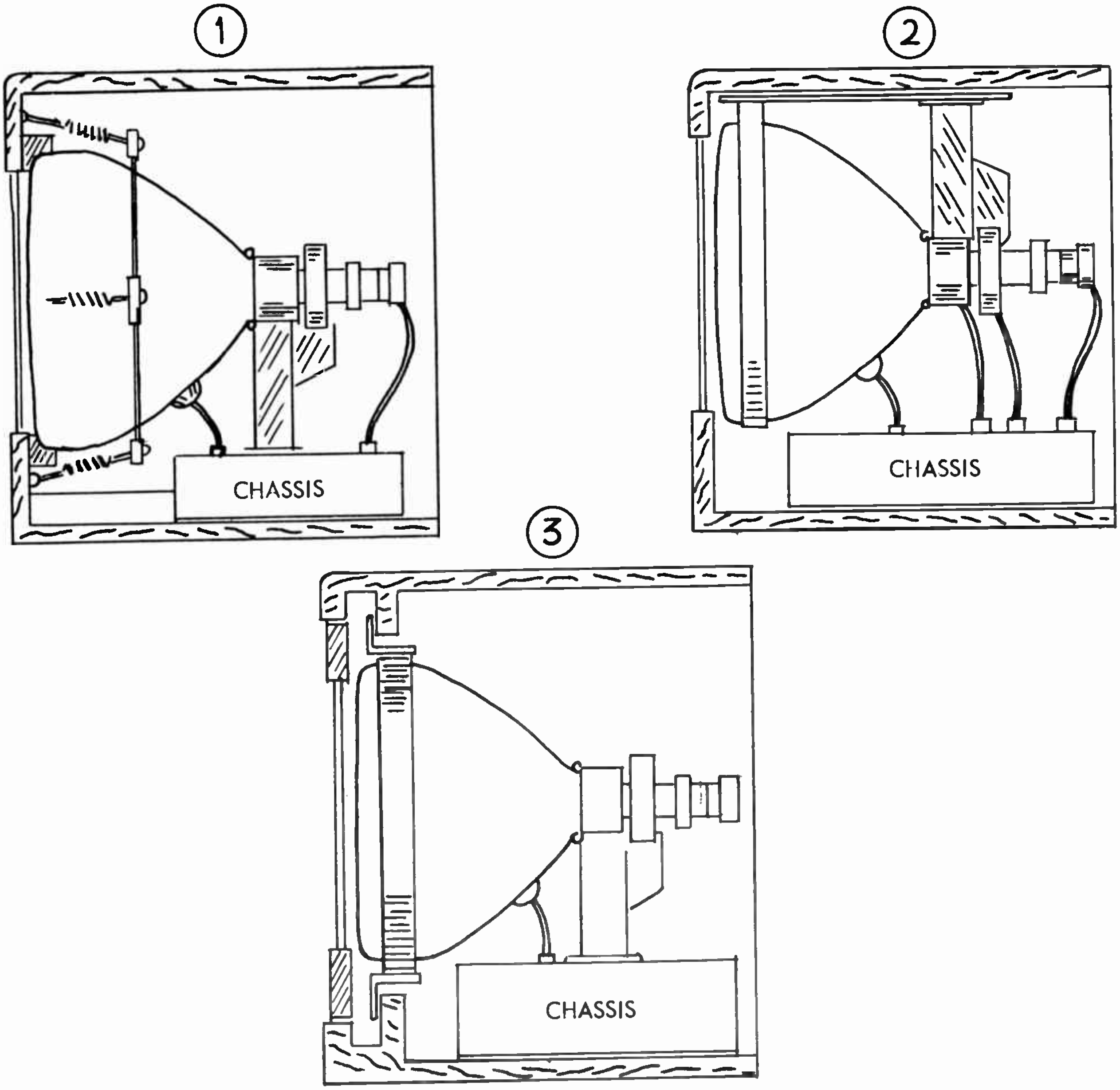


Fig. 63-14. Various methods of mounting picture tubes in cabinets instead of on the chassis.

When the picture tube is supported from the cabinet rather than on the chassis there is great variety of design and of mechanical fastenings. Unless you are familiar with the construction of a particular receiver you must make careful examination of all the details if the tube and chassis are to be removed without serious damage to any of the members, and to avoid personal injury should the picture tube implode.

Some of the more common principles used for separate tube mountings are illustrated by Fig. 63-14. Diagram 1 shows the picture tube held against cushions on the inside front of the cabinet by some form of harness which grips around the bell of the tube and holds it securely by tension of three or more coiled springs. The picture tube remains in the cabinet when the chassis is removed. Before the chassis can be taken out of the cabinet you must disconnect the anode lead, take the socket off the tube base, and remove the ion trap magnet if one is used. Then adjustments or fastenings which maintain the yoke and focus coil or magnet in correct operating positions are loosened so that, when taking the chassis out of the cabinet, these members will move off the base end of the picture tube without straining the neck. The picture tube is later removed through the rear of the cabinet after its harness is loosened or disconnected.

Diagram 2 shows the picture tube, the yoke, and the focus coil or magnet mounted on a separate plate or platform. This tube platform may be fastened to the inside top of the cabinet, as shown, or it may be fastened to either vertical side of the cabinet, or on the bottom of the cabinet alongside the chassis. It may or may not be possible to remove the picture tube platform and the chassis independently of each other. Sometimes the chassis has to come out first because the tube will not pass it. Before either the picture tube or the chassis may be taken out of the cabinet, it is necessary to disconnect the anode lead, also a cable containing deflecting yoke leads, the wires going to a focus coil, and to take the socket off the tube base or disconnect a socket cable at its chassis end.

There are other designs with which the frame carrying the picture tube strap carries also the bracket for yoke and focus coil, with this entire assembly held by screws or studs to the inside front of the cabinet. After disconnecting all leads between the tube assembly and the chassis, the chassis may be taken out from the back of the cabinet. Then the picture tube assembly may be loosened from the cabinet and removed through the rear of the cabinet.

Diagram 3 of Fig. 63-14 illustrates a picture tube held by brackets on its mounting strap to a frame just inside the front opening of the cabinet. After taking the protective glass and mask assembly off the front outside of the cabinet the tube mounting may be loosened and the tube taken out through the front opening of the cabinet. It is necessary first to disconnect the anode lead. Since the tube neck will come forward through the openings in the yoke and focus coil or magnet the positioning adjustments for the yoke assembly must be loosened to avoid binding on the neck. Also, it is necessary to take the socket off the tube base and to remove the ion trap magnet if one is used.

A 7-inch electrostatic picture tube usually is supported at its large end or face end by a recessed cushion around the mask in the cabinet front, with the rear end of the tube carried by a padded clamp around the neck. The neck clamp is permissible because of the small size and weight of this tube, and because the neck is much larger and stronger than on magnetic-deflection tubes. All or part of the length of an electrostatic tube may be enclosed by a metallic shield which may have a separate fastening.

To remove the chassis from the cabinet you first take off the cabinet back, which may take with it the connectors for the power cord and for the transmission line coming from an outdoor antenna. Then disconnect any section of the transmission line that is fastened to both the cabinet and the tuner on the chassis. On any set of fairly recent design there probably is a built-in antenna fastened to the inside of the cabinet. The leads must be disconnected from chassis or antenna, and any antenna tuning controls that interfere with chassis removal must be loosened or moved out of the way. The loud speaker nearly always is mounted in the cabinet. Disconnect the speaker from the chassis, usually by pulling pins or a plug out of clips or a jack that is on the speaker frame. If there is a pilot lamp, disconnect it from the chassis lead. A-m, f-m, or audio amplifier chasses which are separate from the main television chassis will have to be disconnected.

If cabinet holes for shafts of front control knobs will not allow the knobs to go back through them, as usually is the case, take off all the knobs. Usually the knobs slide straight off their shafts when pulled forward, but it is well to look carefully for set screws which may hold some of the knobs in place. Dual control knobs, one large and one small, often come off easiest if you grasp the large one and pull both off at once. Finally, take out the four or more screws which pass upward through the bottom of the cabinet into the metal of the chassis.

Before withdrawing the chassis, if it carries with it the picture tube, put on shatterproof goggles, heavy gloves, and roll down your sleeves. The picture tube might be broken during removal. Withdraw the chassis from the cabinet slowly, while watching for the possibility of chassis parts and cabinet parts striking one another and for the possibility that you have not disconnected all cables and wires that must come off.

To remove the picture tube from the chassis proceed as follows. Many of these instructions apply also to removal of the tube from any mounting which holds it in the cabinet. Always wear shatterproof goggles and heavy gloves. Allow no one not similarly protected to be in the room where you are working.

1. If the receiver has been operated during the preceding 10 minutes or so, discharge the filter and tube capacitances as explained in connection with high-voltage measurements. Discharge the glass faceplate of a metal-cone tube. Leave the anode cable terminal or clip disconnected from the anode connector or tube rim.

2. Carefully pull the socket off the tube base pins. Don't twist the socket, that might break the key on the tube base. Do not remove the socket cover that encloses the lugs and cable connections.

3. If the ion trap magnet is held in place by clamp nuts loosen these nuts. Slide the trap magnet back over the tube base. If the magnet is a coil type leave its connecting wires attached, if they are long enough to allow removing the magnet, and lay the magnet-coil assembly on the chassis.

4. The yoke and focus coil or magnet will remain on the chassis when later you slide the tube neck forward through them. It is advisable to loosen the mounting or adjusting screws of these units to avoid any possibility of the neck binding in their openings.

5. So that the present tube, or a new one, may be replaced in the original or correct position make a pencilled note of the position of the anode connector on a glass tube, or with any tube note the position of the key on the base extension and of the flags on the electron gun if such flags are present.

6. Loosen or remove the mounting strap and any other fastenings for the large or face end of the picture tube.

7. Slide the picture tube forward out of the yoke and focus coil or magnet. Support the weight of the tube only by holding its large end, never the neck. Support the neck or base only enough to guide the tube. Be especially careful not to strike or put pressure on the rather sharp bend at the front end of the flare or on the neck or base of the tube.

While picture tubes are removed from their chassis or other mountings observe the following precautions. Remember that external atmospheric pressure exceeds internal (vacuum) pressure by between 3½ and 4 tons on a 16-inch tube and by about 2½ tons on even a 10-inch tube.

1. While handling a picture tube wear shatterproof goggles and heavy gloves, and do not hold the tube very close to your body.

2. Don't let the tube and any hard objects bump together. Use no metal tools near the tube. Avoid scratching the glass, and do not use a tube that has become noticeably scratched. Be especially careful of joints between glass and metal on tubes of the metal-cone and metal shell types.

3. Rest the tube (other than small electrostatic types) only face down and neck up on cloth or other soft, firm surface. Don't leave the tube where it possibly may fall or be struck, and don't leave it near a radiator or where there may be great changes of temperature.

4. When removing a tube from a regular tube carton lift at the sides of the large end, with the tube face up. If at all possible, keep tubes in regular cartons while the tubes are removed from receivers. Retain all cartons for this purpose.

When you are ready to replace a tube or install a new one on the chassis proceed as follows. Also proceed similarly for separate tube mountings so far as these instructions apply.

1. If any fibre or cardboard cones or cylinders originally were on or around the picture tube make sure that they are correctly replaced. These parts are for insulation, personal protection, and protective spacing. Do not assume that they are not needed.

2. Slide the tube neck back through the yoke and focus coil or magnet until its front or face end will rest on whatever supports are provided. Remember to carry the weight of the tube by holding it near the face end, supporting the neck only enough to guide it into place. Don't let go of the tube until you are sure that it will remain in place. Never use any appreciable force in getting the tube into position; if it doesn't move freely, investigate and remove the obstruction.

3. Shift the tube into its correct "angular" position, so that the anode connector, the key on the base, or some other identifying mark is where it should be.

4. Make certain that a grounding spring or springs make good contact with an external coating on a glass tube. With metal-cone or metal shell tubes there may be a grounding spring or contact for the glass faceplate, watch for it. When a tube is supported separately from the main chassis there will be a grounding wire from the metal tube supports to the chassis—something else to replace.

5. The rear end of the flare or cone must fit closely against the cushion that is on the yoke support, and the yoke must be as far forward as it will go.

6. Check the centering of the tube neck in the openings through the yoke and the focus coil or magnet.

7. Tighten the mounting strap and any other supports for the front or face end of the picture tube.

8. Attach the terminal or clip of the high-voltage anode cable to the tube connector or rim.

9. Replace the ion trap magnet, if used, in approximately its correct position. This magnet will have to be adjusted later on.

10. Carefully push the socket onto the base pins of the tube.

11. When the tube or the chassis with tube mounted is to be replaced in the cabinet, first clean the safety glass or plastic plate in the cabinet and also the face of the picture tube. Use a clean, soft cloth and any of the liquids made especially for cleaning of windows.

12. Make sure that the face of the picture tube cannot be pushed against the protective glass or plastic of the cabinet. There should be stops to prevent this.

You must not dispose of defective picture tubes without first destroying the internal vacuum. Otherwise you may face large legal penalties if someone gets hurt. To destroy a tube, place it in a tube carton, close the carton, and drive any heavy metal rod through the carton and the side or face of the tube.

To destroy the vacuum while saving the tube, possibly for display purposes, proceed thus. Put the tube face down in a carton with enough soft padding under the face to bring the base to the top edge of the carton. Pack around the tube sides with wadded paper, cloth, or anything which holds the tube upright. Around the exposed end of the neck place heavy cloth to leave only the base in view. Using a twist drill no more than 1/8-inch diameter, drill slowly down through the center of the extension on the middle of the base. The tip of the drill will puncture the exhaust tip and allow air to enter the tube. The smaller the drill, without being so small as to break off, the more slowly air will enter the tube and the less is the likelihood of displacing the internal parts or part of the screen coating.

Do Not Tear – Cut Along This Line and Send In For Grading.

TELEVISION

LESSON NO. 64

CARRIER TRANSMISSION AND RECEPTION

Training Your Mind...
C. T. I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

COMMERCIAL TRADES INSTITUTE

Training Your Mind
C.T.I.
Perfecting Your Skill

Chicago, Ill.

LESSON NO. 64

CARRIER TRANSMISSION AND RECEPTION

Every television receiver which is more than a few miles from transmitters requires some kind of antenna for pickup of signal energy from carrier waves. The antenna may be an outdoor type, or it may be indoors, or built into the cabinet, but always the antenna is there. Every radio receiver used to require an elaborate antenna for anything more than local reception, but nowadays most of the standard broadcast sets pick up distant stations with a small self-contained loop or from the power line. Someday all television receivers may pick up all signals without elaborate antennas, but the difficulties are great and it unlikely to happen soon.

We encounter one difficulty in the fact that a television carrier must spread its power over a frequency range 600 times as great as that of a standard broadcast carrier. To provide comparable signal field strength at receivers the television carrier power should be hundreds of times greater than that of the standard broadcast carrier, but their powers actually are about the same.

Another difficulty is that effective use of power for carrier radiation goes down as frequency goes up, and the average television carrier frequency in channels 2 to 13 is nearly 150 times as high as the average standard broadcast carrier frequency. Ultra-high frequencies are far more difficult to handle with any reasonable efficiency.

Added amplification in the receiver is not the whole answer to these problems. This is because carrier waves from which reception is detected are accompanied in space by many kinds of unwanted electrical impulses which cause tiny flecks (snow) all over the picture. These impulses are called "noise". More amplification increases noise and signal together. Furthermore, added amplifying stages introduce noise of their own making, which becomes worse as we go from the antenna input toward the video detector. Noise cannot be tuned out because it extends throughout every channel.

A good antenna, more than anything else, will improve the ratio of signals to noise. A good antenna will bring up the signal strength on higher frequency channels, where greatest difficulty is encountered. It will cut off many kinds of electrical interference, also undesired signals from other types of transmission. A good antenna can clear the picture of "ghosts" due to reflected waves. All in all, it is easier to improve reception with a highly efficient antenna than in any other way.

Although we shall examine the generally used forms of television receiving antennas we cannot look at every style, for new designs are continually appearing. It is fortunate that all of them must conform to certain basic principles, and if we become acquainted with these principles there will be no mystery in their application to various designs and structures.

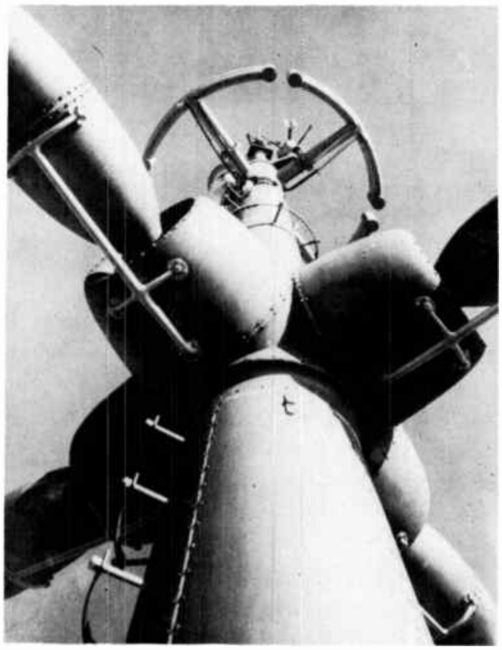


Fig. 64-1. One of the NBC television transmitting antennas.

To understand the behavior of receiving antennas, we first must know something about the carrier waves that bring signal energy from the transmitter. Then we shall look at practical antenna structures which extract signal energy from the waves. Finally, we shall learn about transmission lines through which signals pass from antenna to receiver.

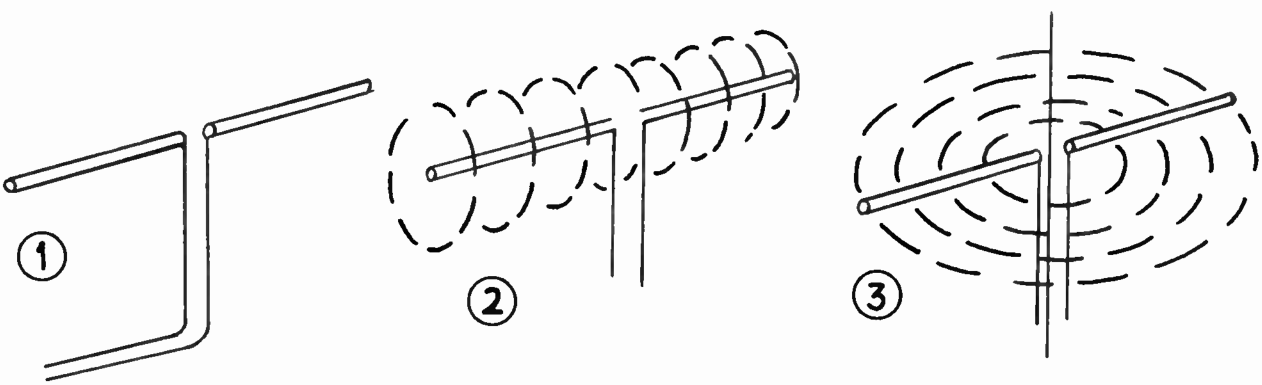


Fig. 64-2. A dipole antenna (1) radiates a magnetic field (2) and an electric field (3).

RADIATED WAVES. A transmitting antenna has, fundamentally, the form shown at 1 in Fig. 64-2. Two horizontal conductors are mounted end to end and connected to the transmitter circuits in which are generated currents and voltages at carrier frequencies. Electrons in the antenna conductors are caused to shift back and forth without leaving the conductors. These electron movements constitute a varying electric current in the conductors.

Around the outside of any conductor in which there is electric current is formed a magnetic field. The invisible and imaginary magnetic lines of force circle around the antenna conductors as in diagram 2, just as they would circle around any current-carrying wire. Because the antenna current is alternating, the magnetic field is continually reversing its polarity. Because antenna current is varying in strength as it alternates, the magnetic field strength varies likewise. The magnetic field varies in strength and reverses its polarity at the frequency of the antenna currents.

6 When a magnetic field is changing in strength or intensity, the change causes an electric field to appear in the same space. Why an electric field always accompanies a changing or moving magnetic field is a long story which goes back to the fact that the moving free electrons in the antenna conductors are electric charges, and all electric charges are surrounded by electric lines of force that form a field. The direction of the electric force or the electric field lines in the space around the antenna conductors is at right angles to the direction of the magnetic field lines, as shown at 3 in Fig. 64-2.

Energy which exists at any one instant in the form of a magnetic field during part of a current cycle will change its form and become an electric field during other parts of the cycle. Then the electric field will disappear and the energy will go back into the form of a magnetic field. This change of energy from magnetic to electric forms, and back again, will continue indefinitely. We saw one example of this action when studying resonance, where energy continually was exchanged between magnetic fields appearing and disappearing around the inductor, and the electric fields appearing and disappearing between the plates of a capacitor.

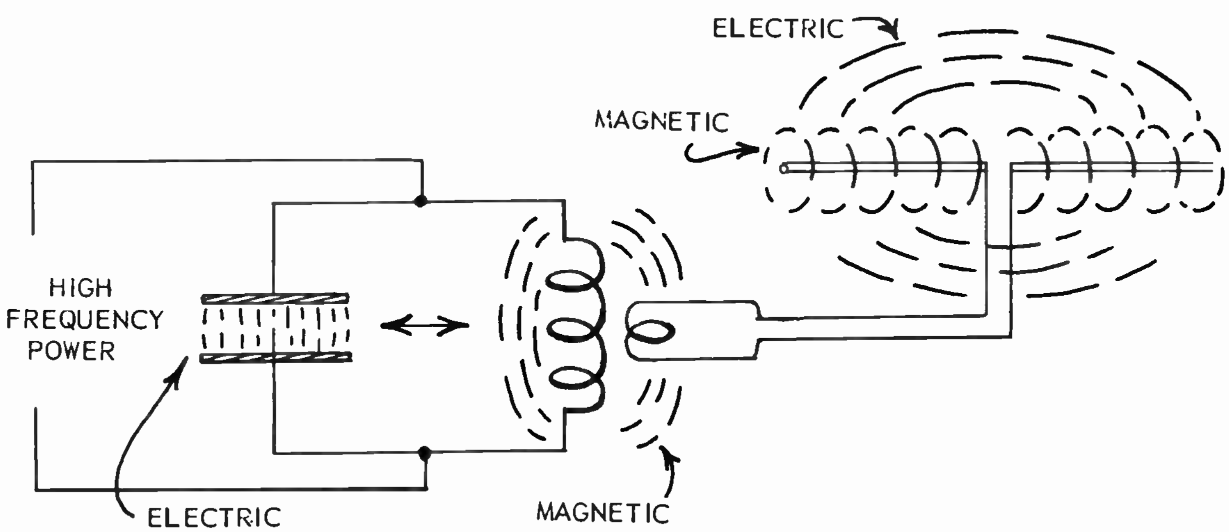


Fig. 64-3. Energy passes back and forth between magnetic and electric fields.

Some exchanges of energy are illustrated by Fig. 64-3. Energy in a resonant circuit exists at one moment in the space around the inductor, and at the next moment in the space between capacitor plates. The energy moves from one space to the other as electrons flow through the circuit conductors. But if we get a varying magnetic field and a varying electric field into the same space, as around the transmitting antenna, the interchange of energy from one form to the other can continue independently of any conductors at all. The two kinds of fields free themselves from the antenna conductors and move away through the surrounding space while, in effect, supporting each other.

As alternations of current in the antenna conductors produce successive magnetic and electric fields in the surrounding space all these fields which free themselves from the conductors fly away from the antenna at the speed of light, which is 186,000 miles per second. In every successive pair of magnetic and electric fields is a certain quantity of energy which has been taken out of the transmitting antenna. Although this energy spreads out thinner and thinner as it expands through space, some of it will reach the antenna at your receiver.

We may represent the varying strengths or intensities of the two kinds of fields as in Fig. 64-4. The alternating directions of the magnetic field are arbitrarily marked north and south, and of the electric field are marked positive and negative. At instant a all the energy is in the magnetic field, shown as of maximum strength in north polarity. At instant b the magnetic field has dropped to zero strength while all the energy has gone over into the electric field, which at b is shown as of maximum strength in positive polarity. At c the electric field has fallen to zero and all energy has gone back into the magnetic field, now of maximum strength in south polarity. At d all the energy has returned to the electric field form, where strength is maximum in negative polarity. At e the conditions are the same as at a. There has been one complete cycle of magnetic and electric field changes.

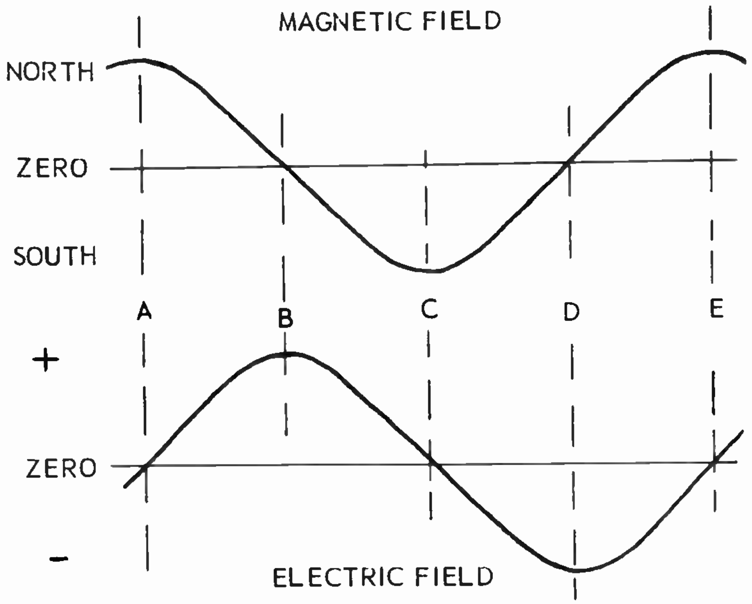


Fig. 64-4. Time or phase relations of the two fields in an electromagnetic wave.

COMMERCIAL TRADES INSTITUTE

TELEVISION

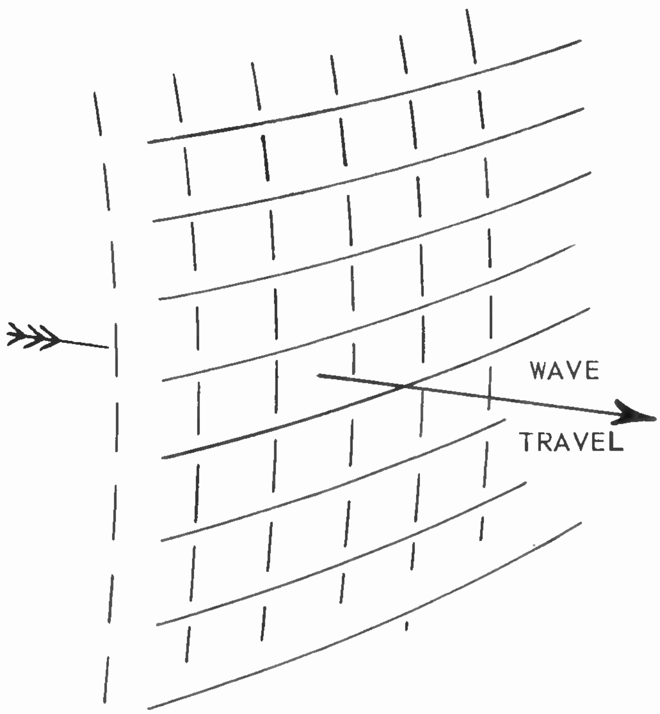


Fig. 64-5. Relative directions of the two fields of a radiated wave, and their direction of travel.

(8) We may think of the magnetic and electric fields as flying together through space somewhat as illustrated by Fig. 64-5. When the conductors of the transmitting antenna are on a horizontal line, parallel to the surface of the earth, the magnetic lines are vertical and the electric lines are horizontal. As the fields travel out through space their lines of force normally remain in these relative positions. The lines of the two fields are at right angles to each other. It is apparent also from the diagram that the lines of both fields are at right angles to the direction in which they travel through space.

The electric and magnetic fields between which there is exchange of energy form what is called an electromagnetic wave. This is the carrier wave with which we are so much concerned. The electromagnetic wave really is a surge of energy whose form continually is changing between electric and magnetic.

Were the conductors of the transmitting antenna vertical instead of horizontal we should have a radiated electromagnetic wave in which the electric lines are vertical and the magnetic lines horizontal. This is because the electric lines of the moving wave always are parallel to the conductors of the transmitting antenna, and the magnetic lines are at right angles.

The electromagnetic wave is said to be polarized in the direction of the radiated electric lines of force. With a horizontal antenna we have horizontally polarized waves, as at 1 in Fig. 64-6. With a vertical transmitting antenna we would have vertically polarized waves, as at 2.

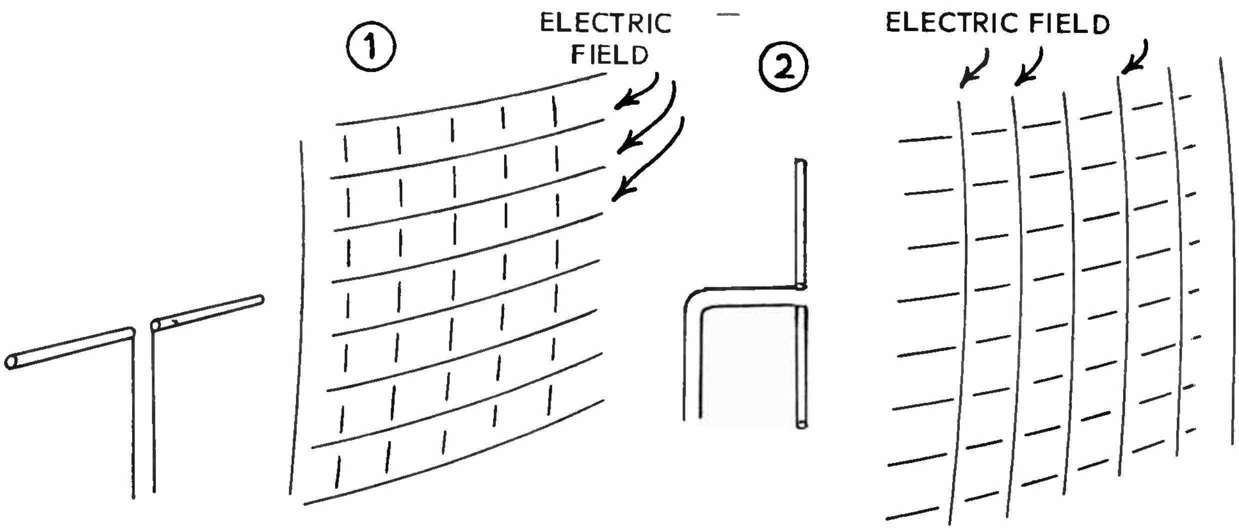


Fig. 64-6. Horizontal polarization (1) and vertical polarization (2).

It is standard to have horizontally polarized carrier waves for television and for f-m broadcast transmission. The direction or polarization of the receiving antennas must be the same as that of the transmitting antennas. Therefore, television and f-m broadcast receiving antennas are mounted so that their conductors extend horizontally. One reason for using horizontal polarization is that radiation from sparking electrical machines and from many other sources of interference is vertically polarized in most cases, and does not so strongly affect a horizontally polarized receiving antenna.

Vertical polarization is used for standard a-m sound broadcasting, also for the lower short-wave broadcasting and general communication. Vertical polarization may also be used for high-frequency transmission and reception when desired.

If we could look down on a horizontally polarized transmitting antenna, and were it possible to see the electric lines of force, they would appear as in Fig. 64-7. Where the lines are drawn closest together the field is of maximum strength, and midway between these points the field strength is zero. Then all the energy would be in the accompany magnetic field.

Since high-frequency currents in the transmitting antenna continually are reversing their direction, or are alternating, the radiated field lines must be doing likewise. Looking at the portion of the wave traveling toward the right, at distance a from the antenna, the field lines are in one direction or polarity. At b they are in the opposite direction or polarity. At c they are back in the first direction, and so they continue to reverse directions all the way to your receiving antenna.

From a to c or from b to d in Fig. 64-7 the electromagnetic wave goes through one complete cycle of changes, both electric and magnetic. The time for completing this cycle must be the same as the time period of one cycle of alternating high-frequency current in the transmitting antenna. If, for example, the carrier frequency were 50 megacycles or 50,000,000 cycles per second, the time from a to c or from b to d would be 1/50,000,000 second.

The wave, with its electric and magnetic fields, is traveling away from the antenna at the tremendous speed of light. This speed is 186,000 miles per second. As usually measured in the radio sciences, with the meter as the unit of distance, this is a speed of 300,000,000 meters per second. One meter is equal to 39.37 inches or to 3.281 feet.

To determine how far the wave has traveled during one cycle we divide 300,000,000 meters (distance per second) by 50,000,000 (cycles per second) and find the answer to be 6 meters. The 50-megacycle wave travels 6 meters through space during every complete cycle. Then the length of one 50-megacycle wave in space is 6 meters, and we may say that transmission is being carried out at a wavelength of 6 meters.

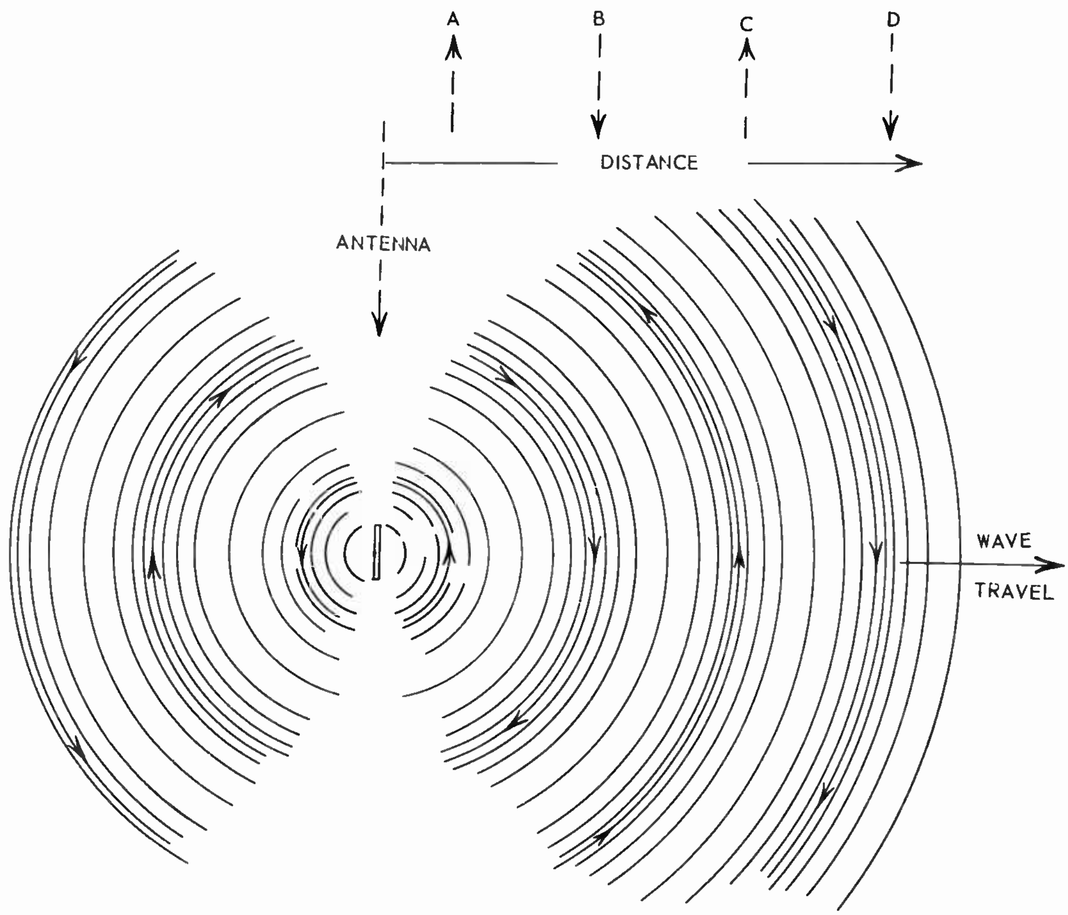


Fig. 64-7. How intensity varies in the electric field of a moving electromagnetic wave.

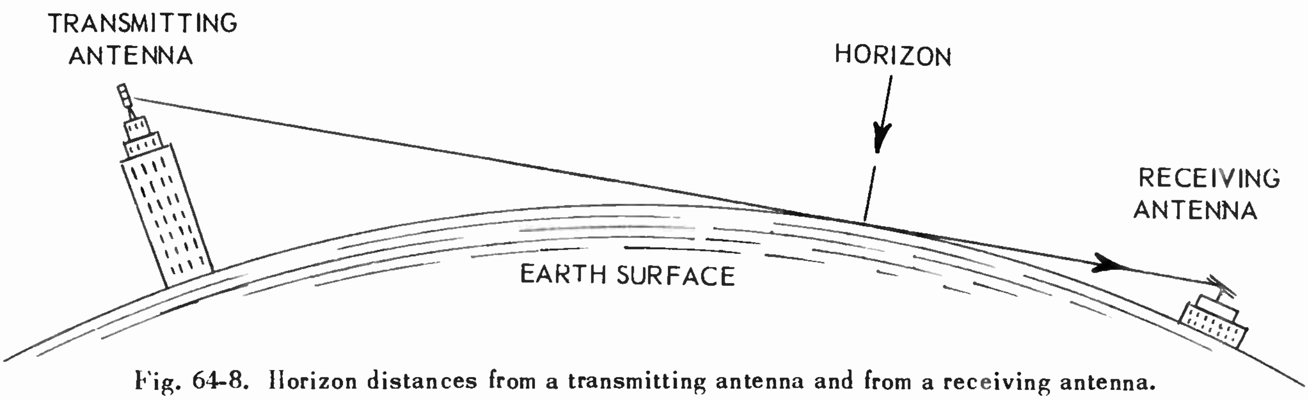


Fig. 64-8. Horizon distances from a transmitting antenna and from a receiving antenna.

With a frequency of 50 megacycles the wavelength must be 6 meters. You can translate any frequency in megacycles to the corresponding wavelength in meters by dividing 300 by the number of megacycles. As an example, dividing 300 by 50 (megacycles) gives 6, the wavelength in meters. Also, dividing 300 by any wavelength in meters gives the corresponding frequency in megacycles. If you divide 300 by 6 (meters wavelength) the answer is 50, the frequency in megacycles.

TRANSMISSION DISTANCES. Radiation waves at very-high and ultra-high frequencies not only travel at the speed of light but they behave like light in many other ways. For one thing, these carrier waves travel on a "line-of-sight" or on a straight line with only very slight bending under any conditions. Theoretically, a downwardly inclined wave from a transmitting antenna, as in Fig. 64-8, could not pass beyond the horizon, which is the farthest point on the earth's surface that could be seen from the position of the antenna.

If this wave passes just over the surface of the earth at the horizon, as seen from the transmitter, it could proceed and be picked up by a receiving antenna at some point beyond the horizon. In our illustration, were the receiving antenna any lower than shown, the wave would pass completely above it and there would be no reception. Also, a wave leaving the transmitting antenna at any higher angle or inclination would pass above this receiving antenna. These more elevated waves could be received only by raising the receiving antenna, which would have the effect of extending the horizon distance from this antenna.

It is apparent that the higher the transmitting antenna and also the higher the receiving antenna the greater may be the distance between the two at which reception is possible. The total distance is the sum of the distance from the transmitting antenna to its horizon plus the distance from the receiving antenna to its own horizon.

Actually the maximum reception distance is a little greater than the sum of the two horizon distances because of a phenomenon called diffraction. Diffraction is a slight bending of the carrier waves as they pass the boundary of any solid object along their path. Diffraction may occur when the wave comes to the earth at the horizon, and the wave will be bent a trifle downward as it goes over the horizon. This diffracted and downwardly bent wave could be picked up by a somewhat lower receiving antenna beyond the horizon.

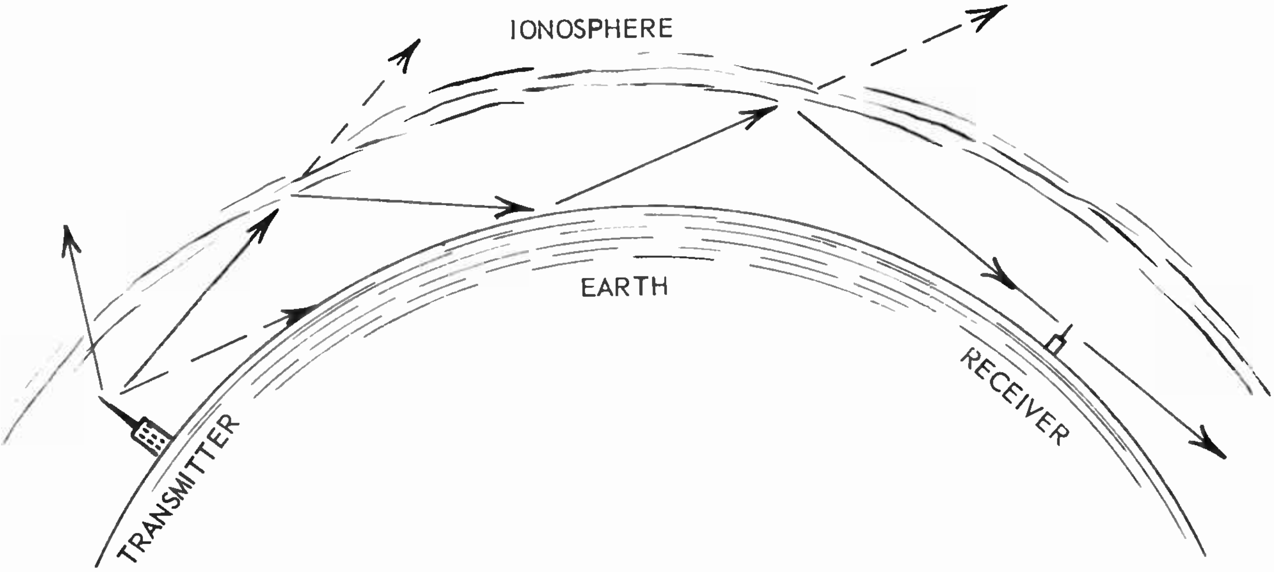


Fig. 64-9. Wave reflection between the ionosphere and the earth at the lower radio transmitting frequencies.

Radio waves at standard broadcast and the lower short-wave frequencies travel much farther than waves at very-high and ultra-high frequencies. This is because the lower-frequency waves can be reflected from a layer of ionized gases about 70 to 80 miles above the surface of the earth. This is illustrated by Fig. 64-9.

Part of the energy in an upwardly inclined radio wave striking the "ionosphere" is reflected back toward the earth and toward receivers possibly hundreds of miles from the transmitter. When a reflected wave comes back to earth it may again be reflected in part strength from the earth or water and go back up to the ionosphere for a second downward reflection. This can keep on until some part of the original wave energy reaches distances thousands of miles from the transmitter.

Waves at very-high and ultra-high frequencies undergo negligible reflection from the ionosphere. Practically all the energy radiated upward goes right through the ionized gases to interplanetary space. Therefore, satisfactory and dependable television reception is limited to but little more than line-of-sight distances.

The horizon distance, plus a liberal allowance for diffraction effect, is approximately equal in miles to the square root of twice the antenna height in feet. Such distances may be read without computation from the graph of Fig. 64-10. On the left-hand vertical scale find the antenna height in feet above average ground level. Trace across to the diagonal line, thence downward to the scale of transmission distances in miles.

This graph applies to both transmitting and receiving antennas. Total theoretical reception distances is the sum of the graph distances for the transmitting antenna and the receiving antenna. For example, assume a transmitting antenna 1,000 feet high and a receiving antenna 30 feet high. The distance for the

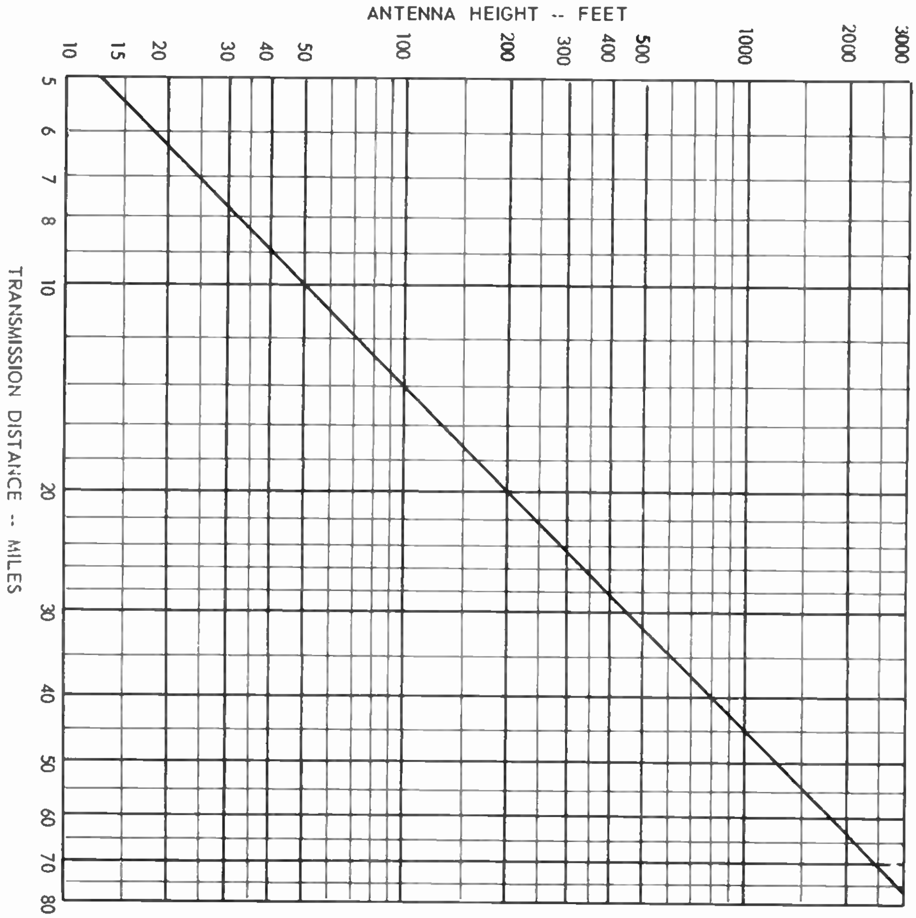


Fig. 64-10. Television reception or transmission distances in relation to the heights of transmitting and receiving antennas.

transmitting antenna is 45 miles and for the receiving antenna it is about 7.8 miles, so maximum possible reception distance between the two antennas is the sum, 52.8 miles.

3 Practicable increases of receiving antenna height add but little to the total reception distances based on horizons. For instance, increasing the height from 30 to 50 feet would add only about 2.2 miles, the gain being from 7.8 to 10 miles. Yet such an increase of receiving antenna height nearly always would allow an immense improvement in reception. This is because the higher antenna would be far above most objects which absorb and deflect the wave energy, and it would be much farther removed from most sources of electrical interference. Almost surely there would be great improvement in the ratio of signal to noise.

Radio waves travel through empty space almost without loss of energy, as witness the fact that radar waves have been sent to the moon and received on earth when reflected back. But the space between a television transmitting antenna and your receiving antenna is far from empty. In addition to gases which form air, and the suspended moisture, there are trees, wires, and all manner of things which absorb energy. As a result of energy losses it is found that maximum distance for satisfactory picture reception seldom is more than 60 miles with any ordinary heights of transmitting and receiving antennas.

Extreme height of the transmitting antenna can increase reception distance. From the many transmitting antennas on Mount Wilson, near Los Angeles, there is regular reception in cities such as Santa Barbara and Bakersfield, nearly 100 miles away, and even at San Diego which is about 115 miles distant. Reception distance may be temporarily increased by "freak" conditions. Such conditions include downward refraction where the waves pass through the boundary between air masses at different temperatures, densities, and humidities, just as when light passes between air and glass or between pieces of glass having different indexes of refraction.

REFLECTIONS AND SHADOWS. Another way in which very-high and ultra-high television waves behave like light waves is that both may be reflected from many objects on earth, although both go right through the ionosphere. When the television carrier waves strike the surface of some object whose dielectric constant is markedly different from that of air or a vacuum, there usually is reflection of a considerable portion of the wave energy. This happens at building surfaces of brick, cement, terra cotta, and stone. There are reflections also from the steep sides of hills and cliffs. Your own body may act as a reflector.

The strongest reflections are from metal surfaces, such as large or small tanks for water or gas, from structures such as bridges, and from metal sheathed buildings. Reflections are most numerous in the built-up sections of cities, especially in manufacturing districts where there are many metal structures and other large objects of metal. Reflection is much less common in the suburbs and in comparatively open country.

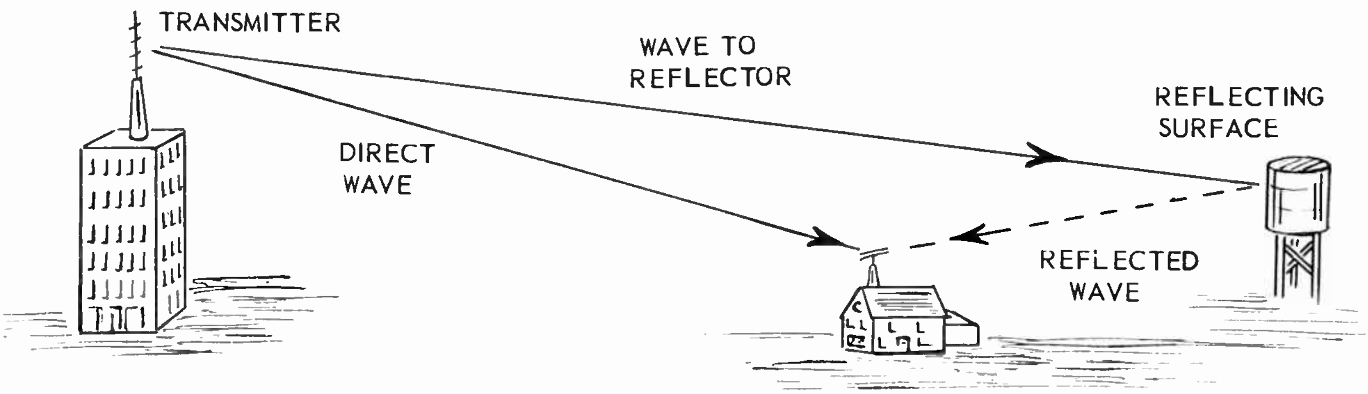


Fig. 64-11. Simultaneous pickup by a receiving antenna of a direct wave and a reflected wave.

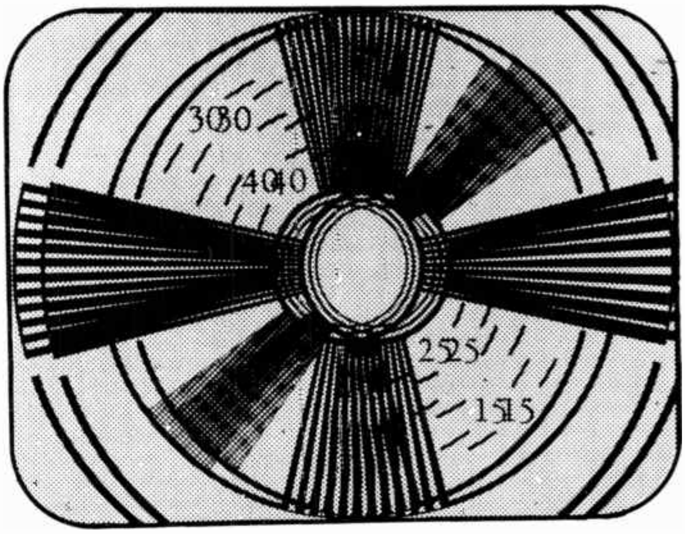


Fig. 64-12. Wave reflection can produce a second pattern or picture, called a ghost.

Your receiving antenna can pick up a reflected signal or several reflected signals at the same time it collects energy from a direct wave. This is illustrated by Fig. 64-11. Time for wave travel to the reflecting surface and from there to the receiving antenna is, of course, longer than the time taken by a direct wave coming to the receiving antenna. The reflected wave has traveled a greater total distance, consequently arrives at the receiving antenna a fraction of a microsecond or a few microseconds after the direct wave.

The two received signals, one direct and the other reflected, produce two pictures or patterns. Since the reflected wave arrives later than the direct wave, and the electron beam moves from left to right in the picture tube, the reflection picture will be displaced to the right of the direct picture on the screen.

5 Fig. 64-12 illustrates what could happen were the two signals of nearly equal strength, and were there a large difference between their arrival times. Ordinarily the reflection signal is weaker than the direct one, and the picture produced by reflection is not so far displaced from the direct one. The picture or pattern resulting from wave reflection is called a "ghost".

1 Fig. 64-13 illustrates some facts relating to wave reflection. While the electron beam in any picture tube completes one horizontal trace, in approximately 53 microseconds, a carrier wave in space travels about 9.8 miles. If a reflected wave has traveled a total of 1,000 feet farther than a direct wave, the ghost will be displaced to the right of the direct reproduction by about 1/50 of the screen width. The displacement measured in fractions of an inch becomes greater with increase of screen width or picture tube size. If the reflected wave has traveled only a small additional distance it will arrive so soon after the direct wave as to cause no distinctly separate image, but only a blurring and loss of detail. It is easy to suspect the focusing adjustments when reflections are the real trouble.

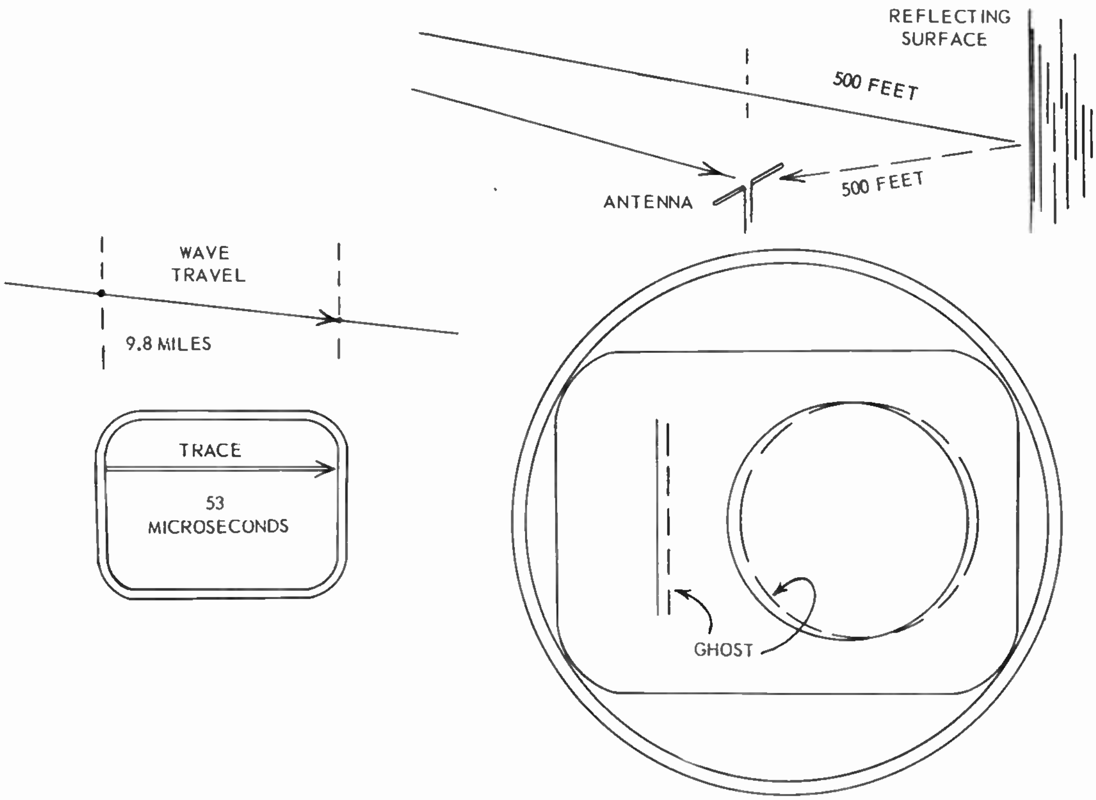


Fig. 64-13. Why a ghost image results from reception of direct and reflected waves.

Reflections are more troublesome from large surfaces than small ones, because a greater area of the "wave front" is reflected. The higher the carrier frequency the smaller may be a surface that causes troublesome reflection. Reflected signals have no noticeable effect on reproduction or quality of sound. Our ears do not detect the slight difference in timing or in phase of the direct and reflected sound signals.

Because very-high and ultra-high frequency waves travel practically along a line-of-sight they are completely cut off on the far side of any large solid masses, such as hills, large buildings, and large structures in general. Such objects cast a signal shadow, and when they are between the transmitting and receiving antennas no direct signal can be received.

When a receiving antenna is in a wave shadow it may be possible to pick up a satisfactory signal from a strong reflected wave. A difficulty which may be encountered under such conditions is pickup of several reflections of nearly equal strengths, with resulting blurring of pictures. Later we shall learn how reflection effects may be eliminated or greatly reduced by suitable antennas and antenna accessories.

THE RECEIVING ANTENNA. The fundamental form of receiving antenna is the same as that of the transmitting antenna, two conductors of similar length and diameter placed end to end, supported horizontally, and connected at their inner ends to the receiver. Such an antenna is called a dipole. Even a structure such as shown by Fig. 64-14 can be traced back to the elementary form, although it has undergone numerous modifications which improve the performance.

The two conductors of a simple dipole are separated at the center only so that we conveniently may take current and power from them to the receiver. So far as the actions of fields and electric charges are concerned we could use a single continuous conductor, as will be done in much of the following explanation.

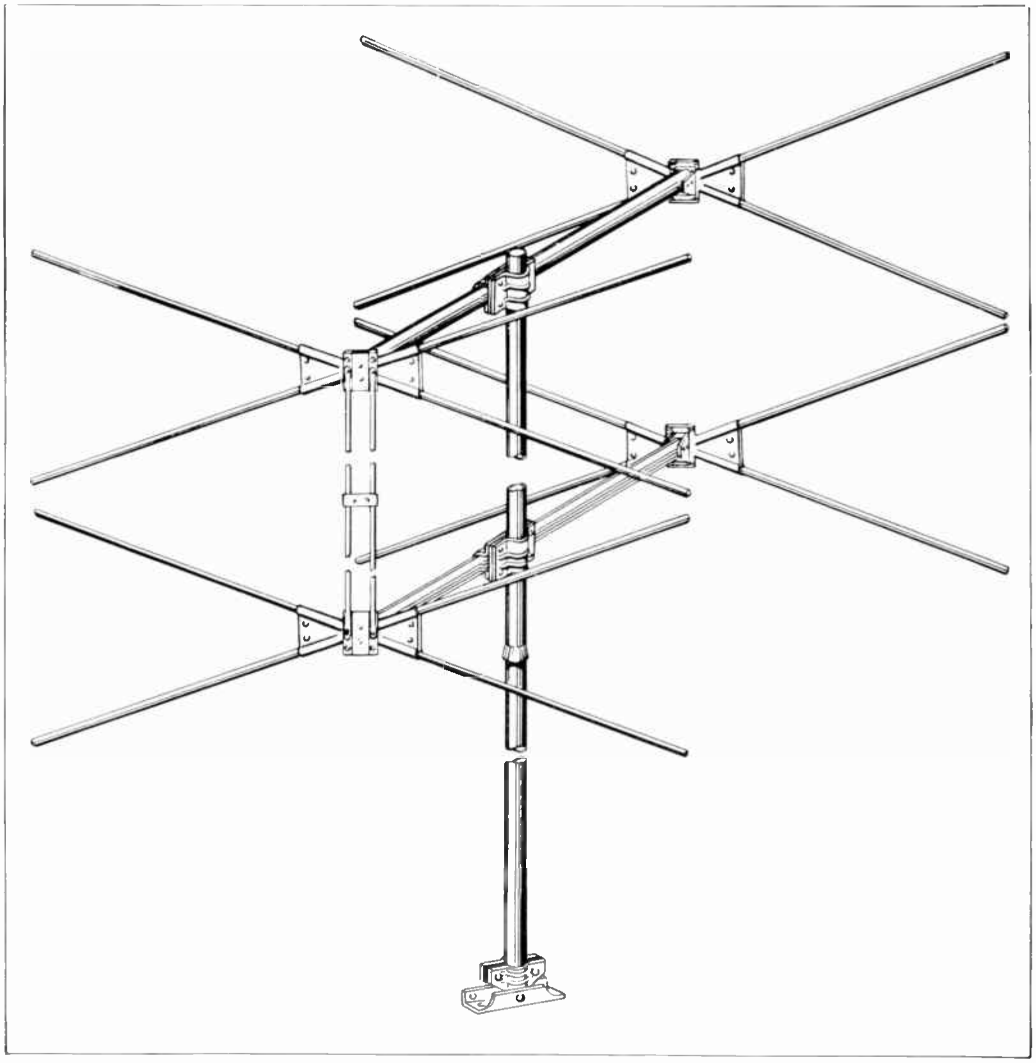


Fig. 64-14. A Taco stacked semi-conical antenna with reflectors.

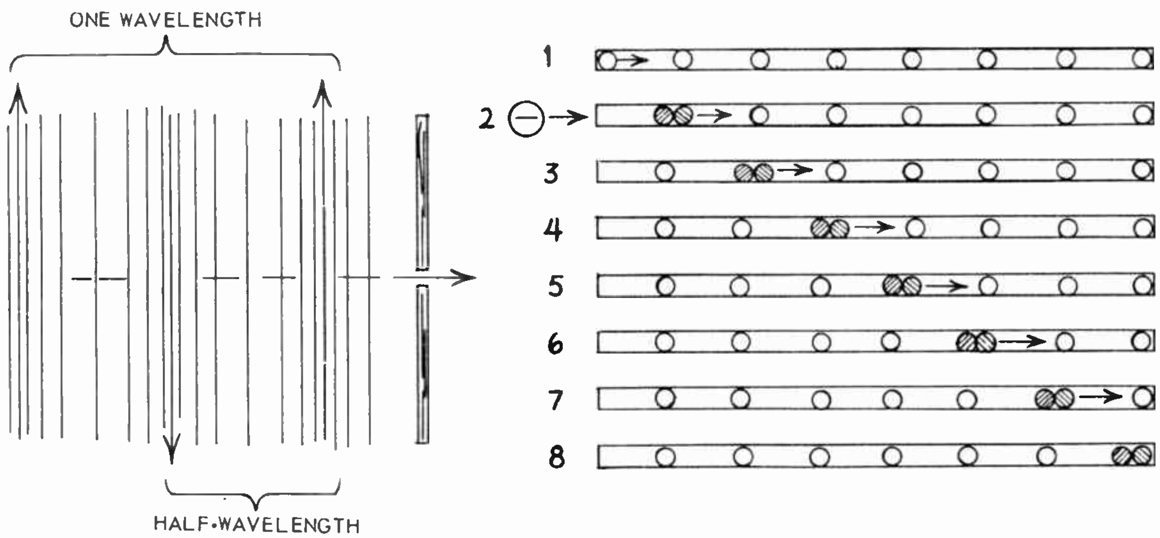


Fig. 64-15. A moving electric charge in a receiving antenna conductor.

Now let's see what happens in a receiving antenna when it is exposed to the successive moving electric fields of a carrier wave, as at the left inFig. 64-15. To begin with, consider the conductor represented at 1, on the right. This conductor is not being affected by any external forces and its free electrons are uniformly distributed. The small circles stand for some of the billions of free electrons in the conductor.

At 2 one end of the conductor has been approached by an external negative electric charge, or by the electric field lines which we imagine as associated with such a charge. The negative electron at that end of the conductor is repelled by the external charge and moves away. This brings the repelled electron close to the one next farther along in the conductor. The two negative electrons which now are closer together repel each other. The first one cannot go back, because of the external negative charge or field, so the second electron is pushed away as at 3. The same action takes place all the way along the conductor, as shown by successively numbered diagrams, until the farthest electrons are pushed together at 8.

Wherever two circles are shown pushed close together there is a concentration of free negative electrons in excess of the normal uniform distribution. A concentration of free negative electrons forms a negative electric charge, as you well know. Therefore, we actually have had a negative charge moving from one end to the other of the conductor. No electrons have moved from one end to the other, it is only a charge that has moved.

Here is a simple comparison which illustrates a moving charge. Assume that you have a string of billiard balls or bowling balls against one another in a trough, and strike the first ball a sharp blow. The last ball of the string instantly moves away, but the intervening balls shift hardly at all. A force, equivalent to our electric charge, has traveled almost instantly from the first ball to the last one, although each individual ball, except the last one, remains almost unmoved.

The last free electron at the far end of the conductor in diagram 8 cannot escape to relieve the concentration or charge at that point. Then this negative charge repels the negative free electrons just back of it. The whole performance now repeats in reverse, with the electron concentration or charge moving back toward its starting point. If the external negative change originally applied has been removed, we end up with an electron concentration or a negative charge as the left-hand end of the conductor, as shown by Fig. 64-16.

As you know, no free electron can progress very fast nor very far through the molecules or atoms of a conductor, and it doesn't have to. But the charge can move exceedingly fast, for the free electrons are so close together that the repulsion force of one acts almost instantly on the next one. The charge moves from end to end of the conductor very nearly as fast as an electromagnetic wave travels through space.

In Fig. 64-16 the backward moving charge reaches the left-hand end of the conductor. This charge then acts in exactly the same way as the original external charge at that end, and in the same way as the charge earlier developed at the right-hand end. The charge is forced back to the right. The charge is being reflected back and forth between the ends of the conductor. This will continue until all the extra energy put into the conductor from the original external charge or field is dissipated in overcoming resistance and other losses.

Instead of speaking of the moving concentration of electrons as a charge we often call it a wave. Then we say that there is wave reflection at the ends of the conductor.

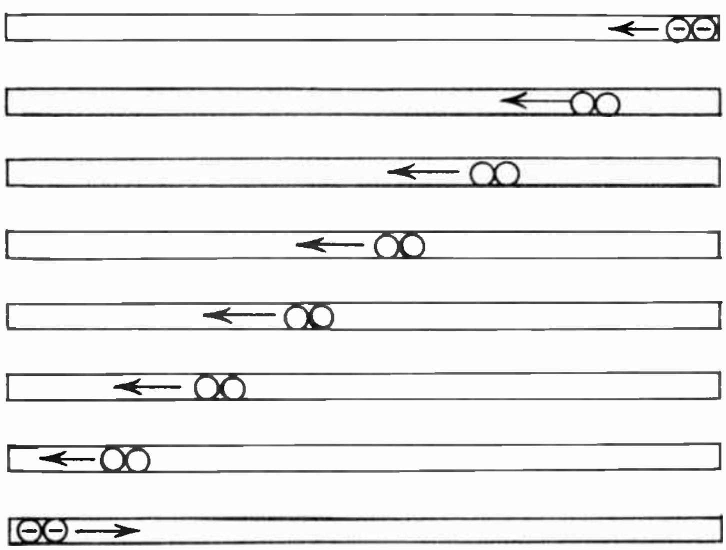


Fig. 64-16. The charge is reflected at the two ends of the conductor.

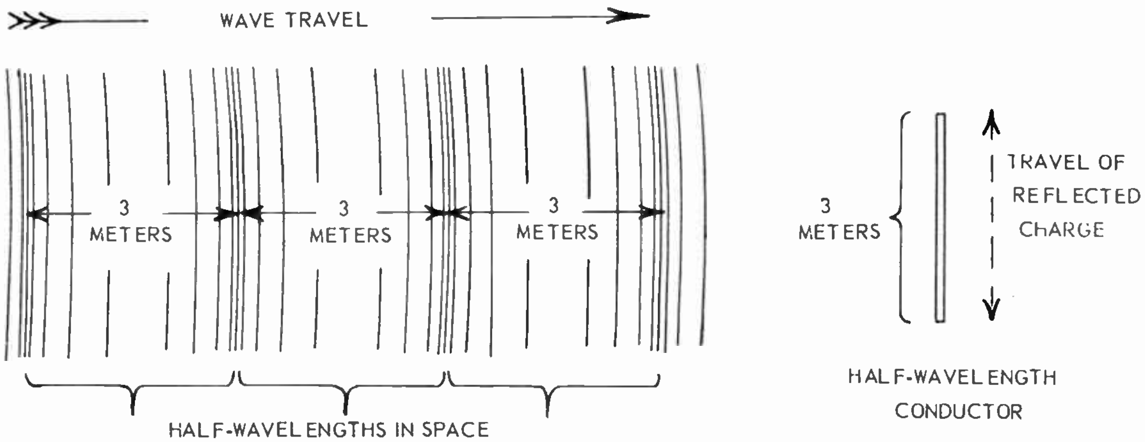


Fig. 64-17. The length of the antenna conductor is equal to a half-wavelength of a space wave.

Although the electric charge moves along the conductor at slightly less speed than a radiated wave moves through space, we shall assume for the time being that the charge can move and does move at the same speed as an electromagnetic wave in space, and that the reflected charge swings back and forth between conductor ends at this speed.

Supposing that the conductor is 6 meters in length, or as long as a 50-megacycle wave in space. Then, assuming that a charge in the conductor can travel at the same speed as a wave in space, the charge will go from one end to the other of the conductor in exactly the same time required for a 50-megacycle or 6-meter wave to travel one wavelength in space. This is in accordance with the relations between frequency, wavelength, and distance previously explained.

Next, supposing that we cut the conductor in half, making it 3 meters long. Now a charge can travel from one end to the other during the time in which a radiation field travels a half-wavelength in space. Fig. 64-17 shows the relations between wave travel, movement of a reflected charge, and conductor length.

Now we shall see what happens in an antenna conductor which is in the path of moving electromagnetic waves. At the top of Fig. 64-18 are represented the changes of electric field intensity in a radiation wave at successive instants of time as the wave moves across the antenna conductor. Relative field intensities are shown by variations of spacing between the electric lines of force. The antenna conductor, a half-wavelength long, is shown as it is reached alternately by maximum and minimum field intensities at equally spaced instants of time numbered from 1 to 5.

Down below are two curves. One represents the varying intensity and reversing polarity of the electric field, which furnishes the potential or voltage being applied to the antenna conductor. The other curve represents current in the antenna conductor. This current really consists of a moving charge which, since it is a concentration of electrons, may be considered as electron flow at a rate of so many coulombs per second past any given point. Such a flow is, of course, an electric current.

When examining curves representing voltage and current, as at the bottom of Fig. 64-18, we must keep in mind that time is assumed to elapse from left to right. Anything shown toward the left occurs before anything toward the right, and anything toward the right occurs later than anything on the left. Instant 1 is at an earlier time than instant 2. Instant 2 occurs before instant 3, and so on. This is the conventional and accepted way of indicating time relations or phase relations between voltages and currents.

We commence, at instant 1, with the charge or electron concentration momentarily at rest as it reaches one end of the conductor and is about to be reflected. Since there is now no electron movement, antenna current must be shown as of zero value. Electric field intensity, which is the potential or voltage applied to the conductor, here is of maximum negative amplitude. The charge now starts along the conductor, reaching a maximum rate of flow at instant 2, where antenna current is shown as of maximum amplitude and negative polarity. Field intensity or voltage now is zero.

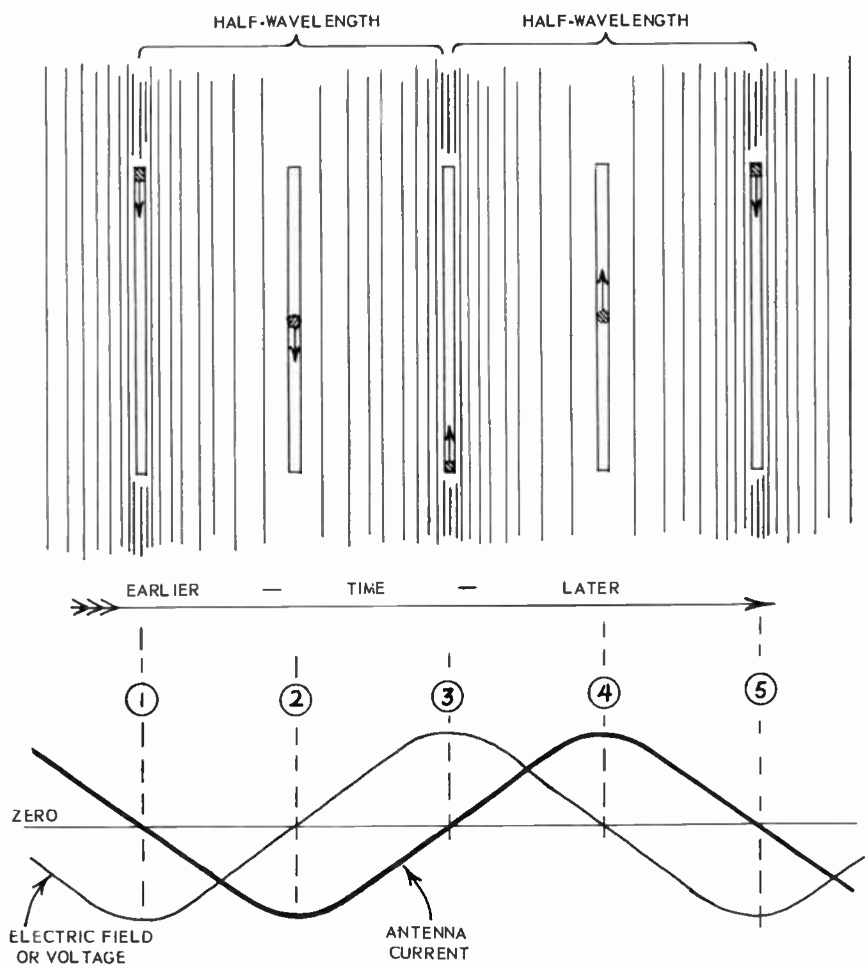


Fig. 64-18. A charge naturally reflected back and forth is aided by energy taken from the carrier wave.

At instant 3 the charge has arrived at the end of the antenna conductor, is momentarily at rest, and is about to be reflected. This means zero electron movement or zero current. Field intensity or voltage is maximum in a polarity opposite to that of instant 1. Now the charge or electron concentration begins its return trip, and at instant 4 we have maximum rate of flow or maximum current in positive polarity, accompanied by zero field intensity or voltage.

At instant 5 one full cycle of voltage and current has been completed, and we are again at the conditions of instant 1. The electromagnetic wave has traveled two half-wavelengths or one full wavelength, which is one full cycle of wave motion.

We have been assuming that a charge moves from end to end of the antenna conductor in the same time that a radiation field travels a half-wavelength in space, although actually the speed of the charge in the conductor must be somewhat less than that of the field in space. The explanation is that we do not cut the conductor to the exact length of a half-wave in space, but make it a little shorter. Then the charge, in spite of its slower speed in the conductor, can travel the shorter distance in the allotted time.

Changes of antenna voltage and current as shown by curves at the bottom of Fig. 64-18 keep time with changes of intensity in the electric and magnetic fields of the radiated wave as the wave moves past a fixed point in space, at which is the antenna conductor. This is illustrated by Fig. 64-19, at the top of which are represented the electromagnetic fields during one complete cycle of a radiated wave traveling from right to left. At point 1 there is maximum intensity of the electric field, whose lines of force are moving toward us. At this point all the energy of the wave is in the electric field, so we show the voltage curve, down below, as of maximum amplitude and negative polarity. This choice of polarity is merely one of convenience, so that we may be able to identify one direction of field lines as opposite to the other direction. Note that electric fields always are associated with potentials or voltages.

At point 2 all the wave energy has gone over into the magnetic field, whose lines of force are shown in a downward direction. Here the antenna current is shown as of maximum amplitude in negative polarity. Again the choice of polarity is only a matter of convenience. Note that magnetic fields always are associated with currents.

At point 3 the wave energy has returned to the electric form, but the lines of force have reversed their direction. Accordingly the voltage curve is shown as of maximum positive amplitude. At 4 the wave energy once more is in the form of a magnetic field, but with lines of force in an upward direction. To conform with this magnetic field the curve of antenna current is shown as of maximum amplitude but of positive polarity.

At point 5 the electric and magnetic fields of the radiation wave have completed one cycle, and are back where we started at point 1. Also, the antenna voltage and current have completed one cycle, and have the same relations as at point 1.

Even though no radiated waves were acting on the antenna conductor, the charge, once in motion, would be reflected back and forth at the frequency corresponding to conductor length and velocity with which the charge travels. However, the charge soon would cease to move because all its energy would be dissipated. Energy thus lost is replaced by energy taken from the radiated wave, and the charge is enabled to maintain its movement. Wave energy may also add to the charge or increase the concentration

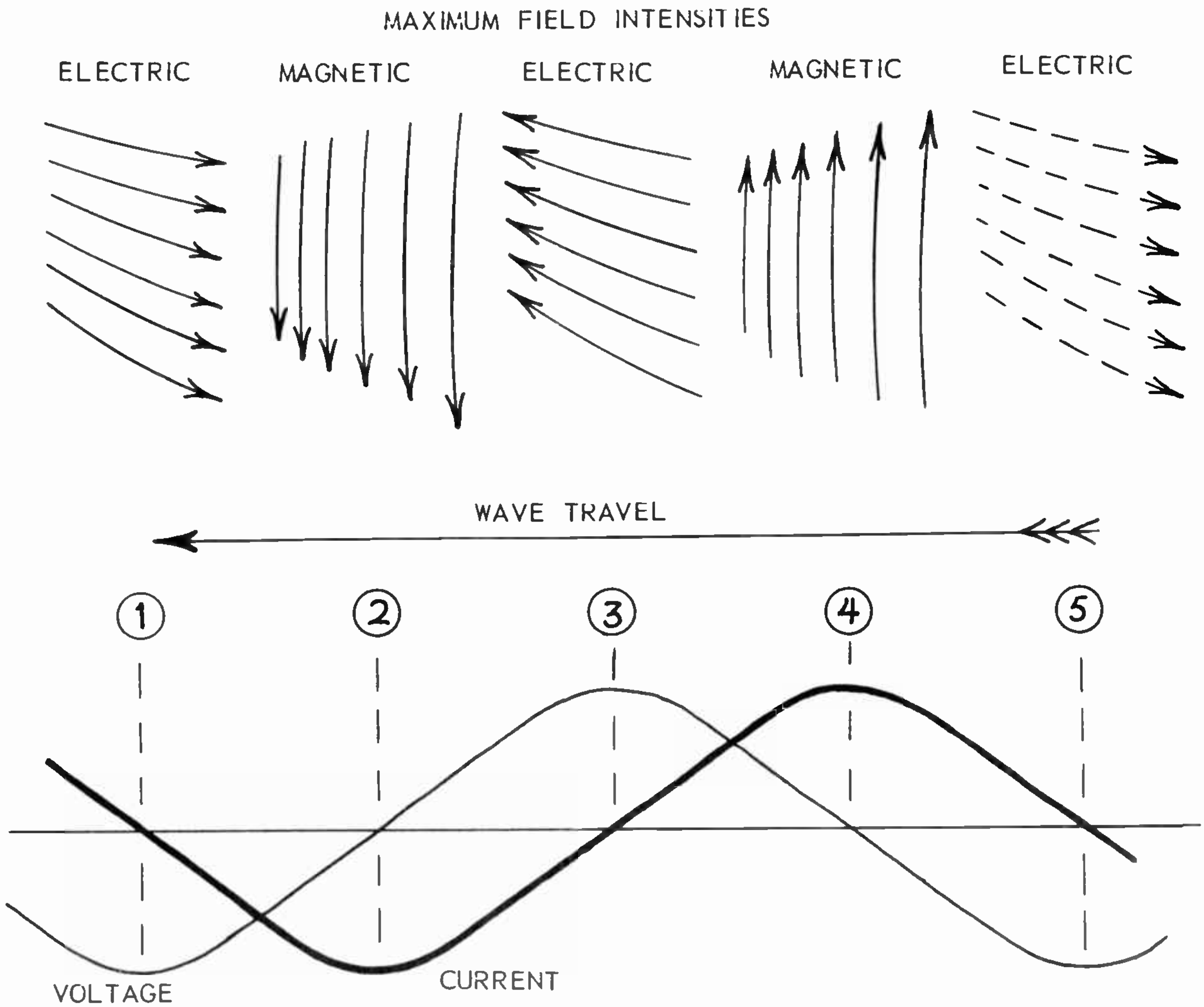


Fig. 64-19. Antenna voltage and current are in phase with changes of intensity in the electric and megnetic fields of the electromagnetic wave.

of electrons. Energy in the antenna conductor thus may be built up until there is a balance between that continually taken from carrier waves and the amount being dissipated and transferred to the receiver.

Antenna current which is maintained by wave energy will surge back and forth in the same manner whether the conductor is a single piece or is cnt at the center. Carrier waves act along both conductor sections when there is a center gap. If we connect two wires from the inner ends of the separated antenna conductors to the receiver input circuit, free electrons in these wires and in the receiver circuit will be forced to surge back and forth in time with movements of electrons in the antenna. Then, in the receiver input circuit, there will be current at the same frequency and same relative amplitudes as in the antenna of the transmitter far away.

DIPOLE ANTENNAS. An antenna of the type described is a dipole no matter what its length. When overall length of the antenna conductors is such that a charge or charges can travel from one outer end to the other in the same time that an electromagnetic wave travels a half-wavelength in space we have a half-wave dipole. The half-wave dipole is the fundamental form from which all television receiving antennas have been evolved.

A half-wave dipole antenna has many of the characteristics of tuned resonant circuits with which we are so familiar, because this antenna actually is resonant at the frequency or wavelength corresponding to conductor length. The half-wave dipole is tuned to a certain frequency or wavelength by making the conductors approximately as long as a half-wave at that frequency. An important characteristic of the half-wave dipole is its impedance. The impedance of any given antenna will vary with frequency of received carrier waves, just as the impedance of any other tuned circuit varies with applied frequency.

A half-wave dipole antenna possesses some certain Q-factor, as does any other resonant circuit. If we make the antenna of high-Q design and construction it will have a rather sharp peak of resonance, and will pick up sufficient signal energy in only a narrow range of frequencies. By lowering the Q-factor we may increase the bandwidth. Then, as with every resonant circuit, there will be some loss of gain.

The dipole antenna may be turned or "oriented" to receive signals in maximum strength from only one direction. The angle within which there is strong reception may be made very sharp, or, if we so desire, it may be widened to receive signals from any stations within a rather large directional spread. The dipole also may be rotated so that it refuses to accept signals from certain directions or, at least, picks up very little energy from these directions.

2 There are antenna attachments, called parasitic elements, which will practically cut off reception from one direction while greatly strengthening signal pickup from the opposite direction. We may erect and connect together two or more dipoles with or without parasitic elements to greatly increase the gain, somewhat as we might add stages of amplification in a receiver.

These are a few of the reasons why it is possible to have such improvement in picture reproduction by working with the antenna, and why reception is likely to be poor at any great distance from transmitters unless you know what can be done with antennas and how to do it. How to do it will be the subject for following lessons.

DO NOT TEAR — CUT ALONG THIS LINE AND SEND IN FOR GRADING.

TELEVISION

LESSON No. 65

ANTENNA CHARACTERISTICS

"Training Your Mind"

C.T.I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind
C.T.I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON No. 65

ANTENNA CHARACTERISTICS

Every television receiving antenna, no matter what its type, possesses five important characteristics. They are impedance, gain, bandwidth, directional properties, and front-to-back ratio. Impedance determines the degree to which signal currents and power are weakened at various received frequencies. An impedance curve for an antenna might appear as at 1 in Fig. 65-1. Gain is a measure of signal power from the antenna in question as compared to the power from the same carrier wave in some other type taken as a standard. Signal strength from the antenna at 2 might be several times greater than from a simple half-wave dipole.

9 Bandwidth refers to how great is the range of carrier frequencies satisfactorily received by the antenna. Bandwidth could be called frequency response, and shown as at 3 in Fig. 65-1. The directional property determines how wide or how narrow will be the angle within which carrier waves may come to the antenna for reception at given strengths, and from what direction the signals must come. Later we shall examine directional patterns such as the one shown at 4. Front-to-back ratio tells the relative signal powers obtained from carrier waves coming toward the front and toward the back of the antenna.

ANTENNA IMPEDANCE. Our attention will be given first to the matter of impedance, for all the other characteristics are affected by or related to impedance. According to our original definition of impedance it is opposition to flow of alternating current in a circuit containing resistance and either inductive reactance, capacitive reactance, or both kinds of reactance. Impedance, in ohms, is equal to the number of alternating volts applied to a circuit divided by the number of alternating amperes of resulting current flow.

You may think of the alternating voltage applied to the antenna as being the force in the electric fields of the carrier waves. The alternating current consists of the charges that surge back and forth within the antenna conductors.

If the antenna is of a total length that matches a half-wavelength of the carrier that is to be received, the charges within the antenna conductors surge back and forth within time periods that naturally match the times between successive electric fields. Then the charges and signal power are built up to the greatest possible strength.

In an ordinary electric circuit the greater the current produced by any value of applied voltage the less must be the impedance of the circuit. For example, were 10 volts to cause current of 2 amperes, the circuit impedance in ohms would be 10/2 or 5 ohms. Were the same 10 volts to cause 5 amperes of current the circuit impedance would be 10/5, or only 2 ohms. The same general rule holds true for voltage, current, and impedance of an antenna.

When there is maximum possible antenna current for any given carrier or field potential it means that the

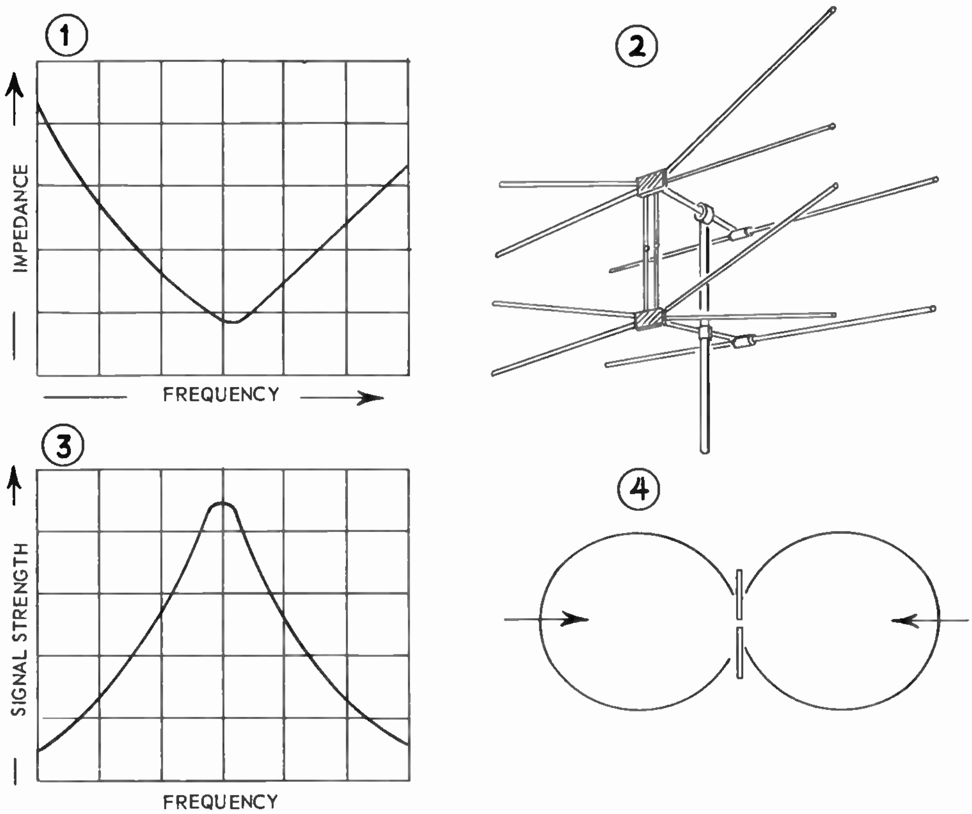


Fig. 65-1. Antenna characteristics may be shown with graphs.

antenna impedance must be of minimum possible value. Since there is maximum antenna current when the antenna is cut to match a half-wave-length of the carrier, this must be the length which allows minimum impedance.

The relations between antenna length and carrier wavelength for minimum antenna impedance are illustrated at 1 in Fig. 65-2. At the open ends of the antenna there will be zero movement of electrons, which means zero current, as the moving charge comes to rest and is reflected. When there is zero current at the end of the antenna there must be maximum voltage, because at any open circuit we have this condition. Then the reflected charge or maximum current will have come back to the center of the antenna as the voltage goes through its zero value.

If electric field strength of the carrier and current in the antenna can remain in such time relation or phase relation to each other as to satisfy the conditions just outlined the result will be maximum current and maximum signal power. This will be possible only when antenna length equals a half-wavelength of the carrier to be received, for only then can the current surge back and forth between ends of the antenna in exactly the time that elapses between arrival of successive carrier waves.

Supposing now that the antenna is made longer than a half-wavelength of the carrier, as at 2. The charges or current surges now have farther to travel between reflections, and they tend to fall behind the changes of carrier field strength. During the ends of each cycle the movement of charges in the antenna will be hindered rather than assisted by the carrier field. Although carrier field strength remains unchanged there will be less total antenna current. This means that antenna impedance has been increased.

Look next at diagram 3 of Fig. 65-2. Here the antenna has been made shorter than a half-wavelength of the carrier. Charges or current surges now have less distance to travel between reflections than in diagram 1, and they tend to get ahead of changes in carrier field strength. Antenna current will try to complete its cycle before the completion of a cycle of carrier field strength. The result of this out of phase condition is reduction of antenna current, which means that impedance has been increased. Impedance is increased and antenna current is lessened when the antenna length is either more or less than a half-wavelength of the carrier.

Now let's look at this matter of antenna impedance from another standpoint. At 1 in Fig. 65-3 a half-wavelength of the received carrier matches the length of the antenna. This is true also at 1 in the preceding figure. At 2 in Fig. 65-3 the antenna length is unchanged, but carrier frequency is higher and its wavelength shorter. This, in effect, makes the antenna longer than a half-wavelength of this higher-frequency carrier, just as at 2 in the preceding figure. Again we have the carrier field opposing movement of antenna charges during part of the cycle. There is reduced current and greater impedance. At 3 the antenna length still is unchanged, but now the carrier is of lower frequency and longer wavelength. This, in effect, makes the antenna shorter than a half-wavelength of this lower-frequency carrier. Conditions are the same as at 3 in Fig. 65-2, and antenna impedance is increased.

Antenna impedance is increased in Fig. 65-2 by making the antenna either longer or shorter than a half-wavelength of the carrier to be received. In Fig. 65-3 the antenna impedance is increased when carrier frequency is either higher or lower than that for which the antenna is cut. Note this: More antenna length has

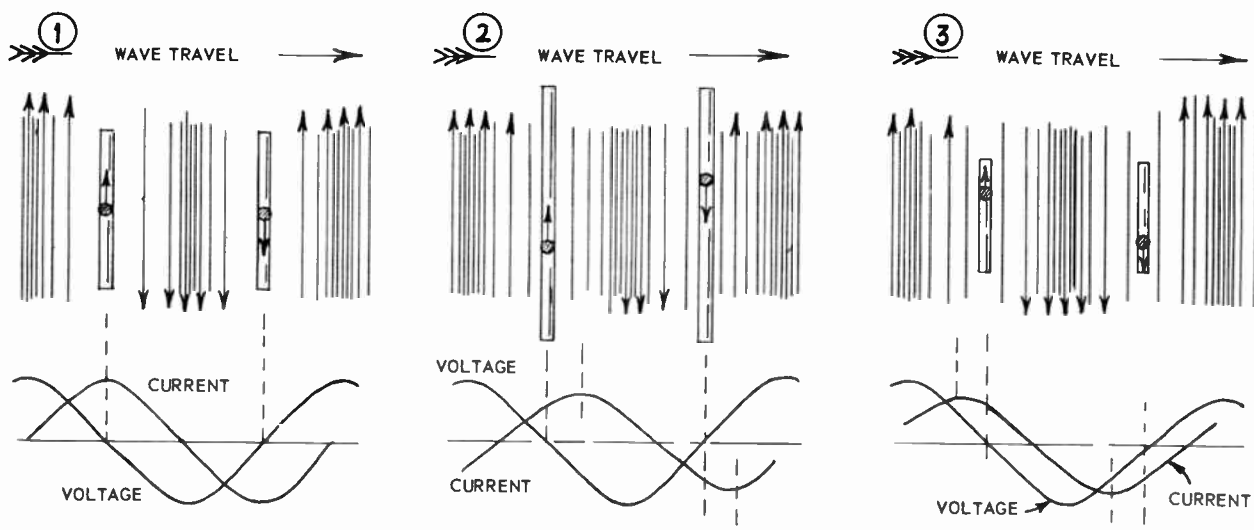


Fig. 65-2. Why changes of antenna length alter signal current and power.

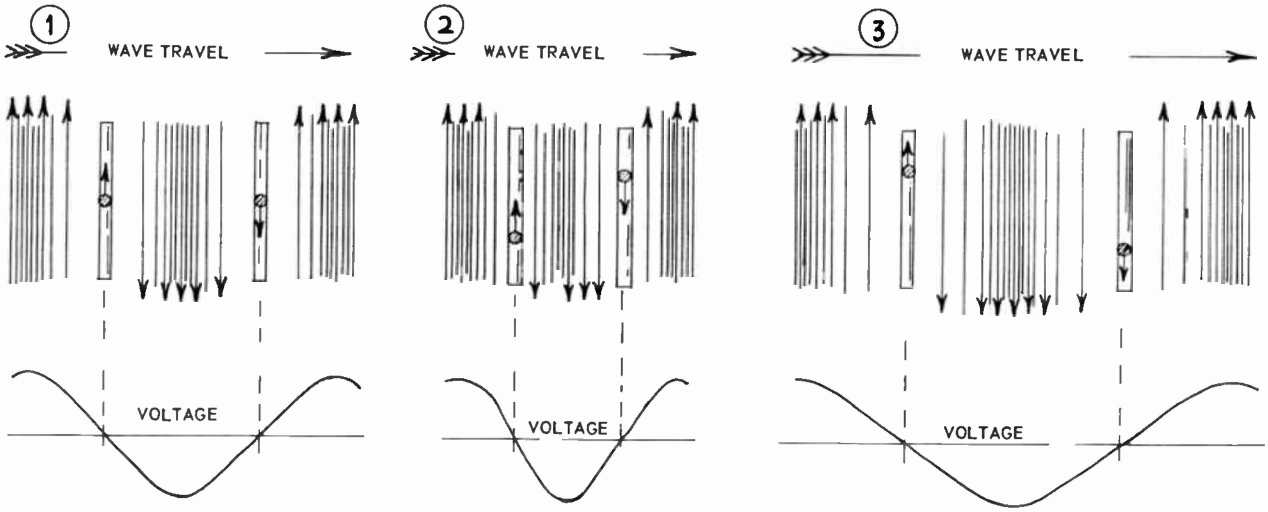


Fig. 65-3. **Why changes of carrier frequency alter signal current and power**

the same effect as higher carrier frequency, while less antenna length has the same effect as lower carrier frequency.

There is only one carrier frequency for which a half-wave dipole of any given length can have minimum impedance, and there is only one antenna length at which there can be minimum impedance for any given carrier frequency. When a simple half-wave dipole is well separated from all surrounding objects this type of antenna has minimum impedance of about 72 or 73 ohms. This impedance consists entirely of resistance, there are no reactances mixed in with it. You will say, correctly, that the antenna conductors are of such large diameter and of such little length in feet or inches that they could not possibly have anywhere near this much resistance to current flow.

The explanation for the seemingly high antenna resistance is that we are not talking about ordinary ohmic resistance, but about high-frequency resistance. As you will recall, high-frequency resistance is a measure of all kinds of energy losses. Here we are considering all the energy that is used for converting carrier field strength into signal currents. This lost energy is equal to that which would be lost, or changed to heat, in forcing the antenna current through 72 or 73 ohms of ordinary ohmic resistance. When we say that minimum antenna impedance is so many ohms, we are referring to the number of ohms that would use up the same energy that actually is being used to convert carrier waves into signal currents.

Many explanations of antenna performance are easier to understand if we think of the antenna as a resonant circuit which may be tuned to a given frequency by cutting the conductors to suitable length. This is the correct way to think of a television antenna, for it really is a resonant circuit. The transmission line which runs from antenna to receiver "sees" the antenna as a series resonant circuit.

A resonant circuit has inductance and capacitance, and has inductive and capacitive reactances. The television antenna possesses inductance because even a straight conductor of any kind has inductance directly proportional to its length. There is capacitance between different points along the antenna conductors. There must be capacitance because there are differences of potential or charge along the antenna. Between different potentials are electric lines of force acting through surrounding air as a dielectric. Where there are electric lines of force there is capacitance.

The antenna is resonant at a frequency whose half-wavelength is equal to the effective length of the antenna. At this resonant frequency the inductive and capacitive reactances balance out and leave only that minimum resistance which represents the losses in conversion of wave energy to signal currents. This condition is represented at 1 in Fig. 65-4.

If the received carrier frequency is higher than that for which the antenna is cut, or if the antenna is too long for the received frequency, as at 2, we have in the antenna an excess of inductive reactance. On the other hand, we have reduction of inductive reactance and a remaining excess of capacitive reactance when, as at 3, the carrier frequency is lower than that for which the antenna is cut, or when the antenna is too short for the received frequency.

The further we depart from the condition of resonance, either in received frequency or in length of antenna, the greater becomes the excess reactance of one kind or the other. Then there is increase of impe-

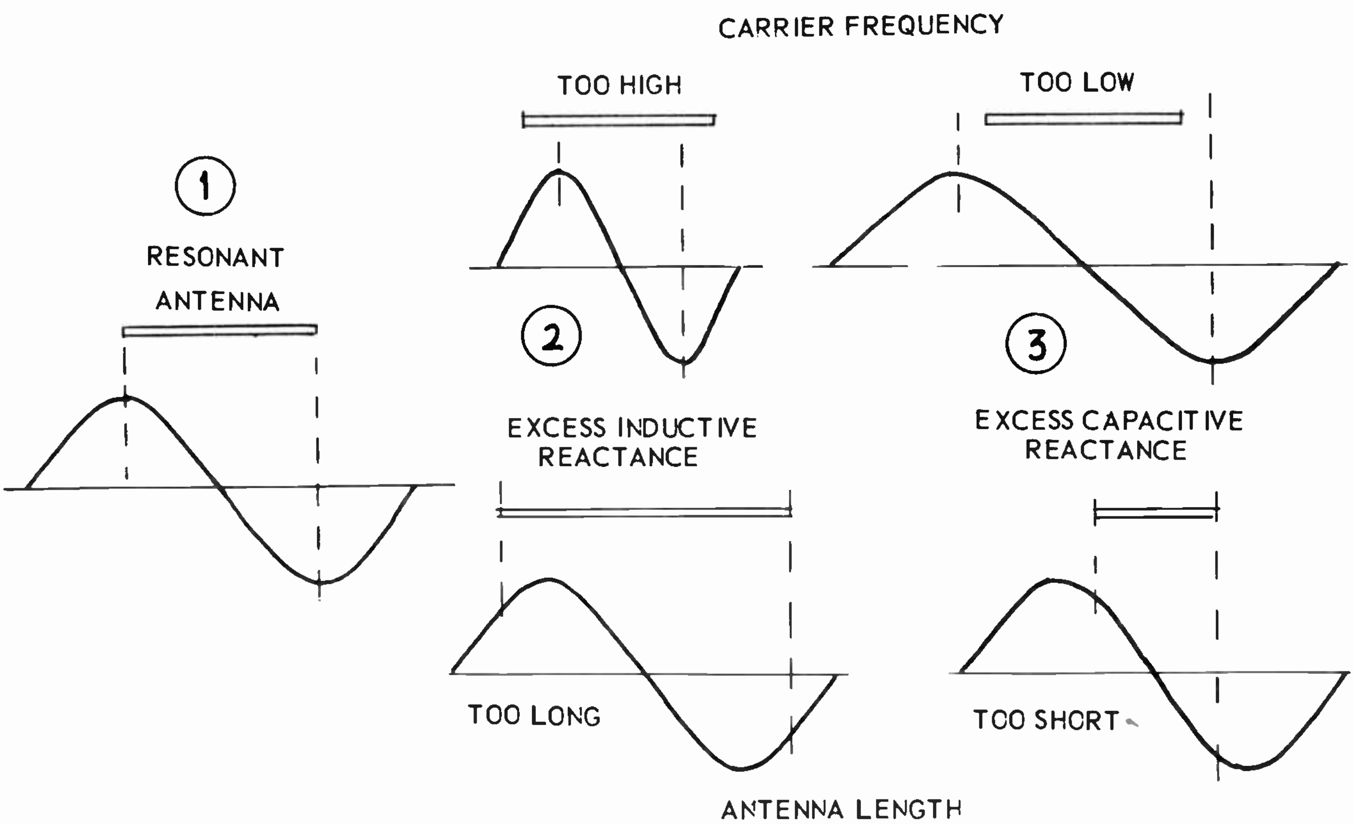


Fig. 65-4. Relations between carrier frequency, antenna length, and antenna reactances.

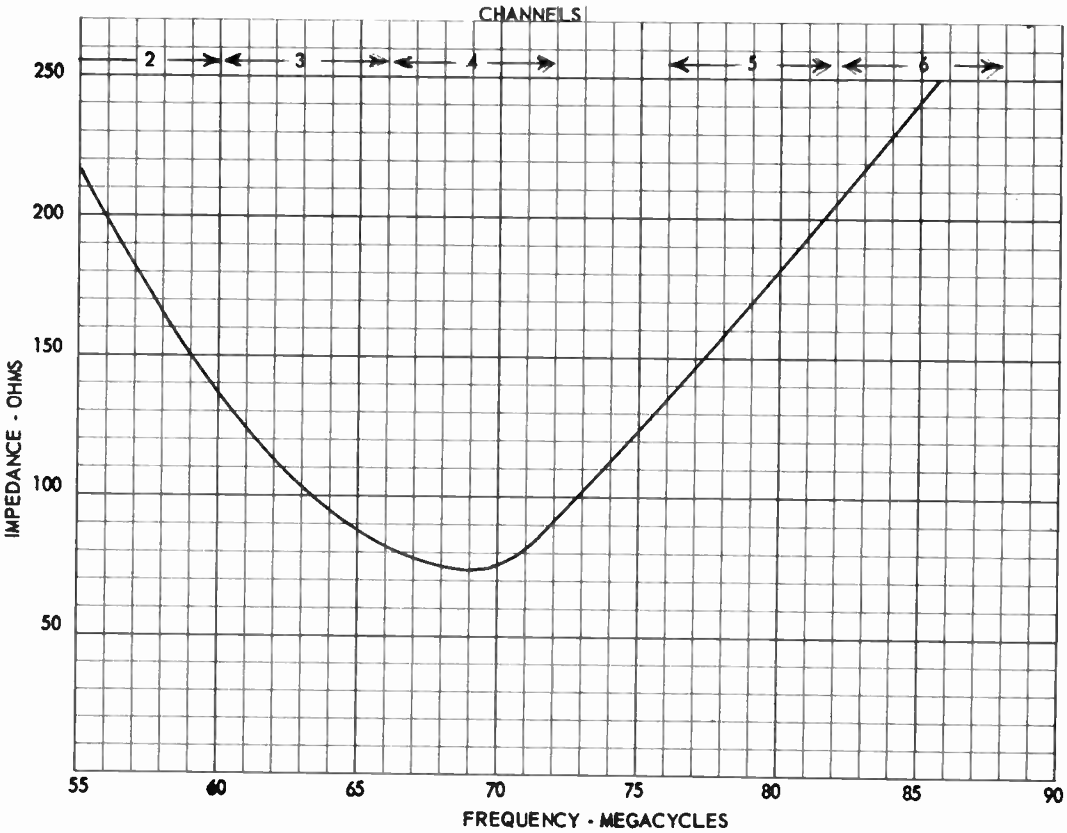


Fig. 65-5. An impedance curve for an antenna.

dance, and a reduction of signal current available for the receiver. Fig. 65-5 shows variation of impedance at various low-band channel frequencies in a particular antenna cut for resonance at the center frequency of channel 4.

The theoretical value of minimum impedance, 72 or 73 ohms for the simple half-wave dipole, is based on the assumption that the antenna conductors are well elevated above the ground, are far removed from all other conductors and also from dielectrics or non-conductors, and are used without any such parasitic elements as directors or reflectors, which will be examined later. Unless all these conditions are satisfied, the actual minimum impedance at the resonant frequency may be greater or less than the theoretical value.

ANTENNA LENGTH. Fig. 65-6 shows how or where we measure the physical length, in feet or inches, of a half-wave dipole in matching this length with the half-wavelength corresponding to some certain carrier frequency. The length is measured between the extreme outer ends of the two conductors. It is not the sum of the separate lengths of the conductors, because between them is a gap for connection of the transmission line that goes from antenna to receiver.

The width of the center gap has no bearing whatever on the effective length of the antenna as considered from the electrical standpoint. When the antenna conductors are attached to an insulation block carried by

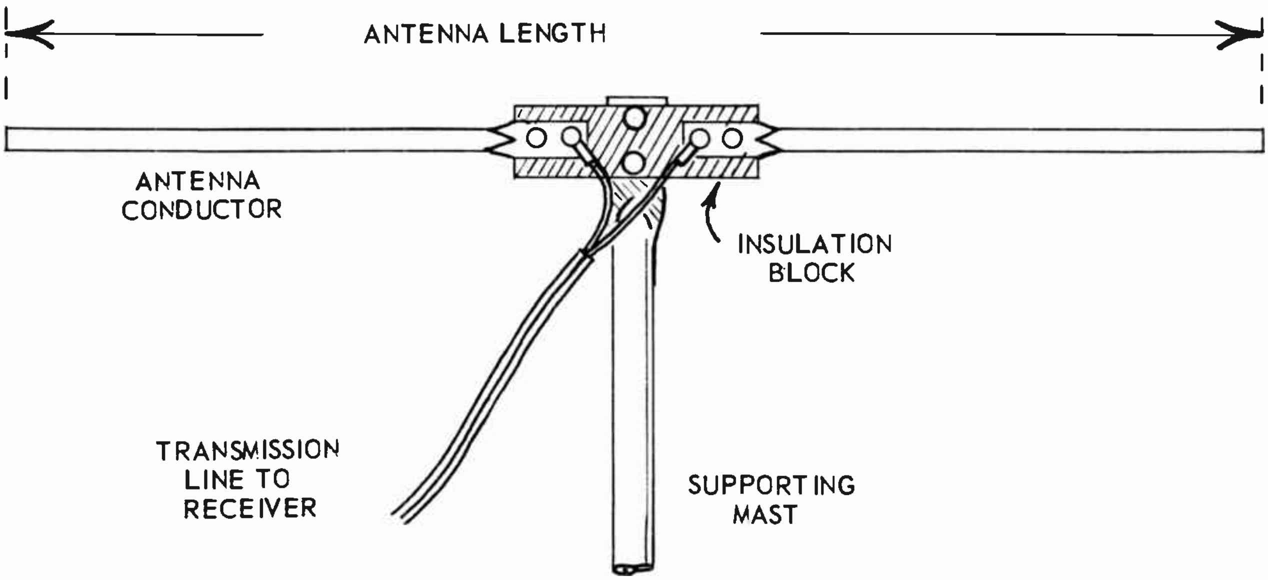


Fig. 65-6. Antenna length is the overall length.

the supporting mast, the gap must be wide enough to allow clearance for the metal parts with no possibility of short circuiting the conductor ends and transmission line. When there is no danger of a metallic short circuit the gap must be wide enough that accumulation of dust and wind-blown dirt will not form a path short enough for signal current leakage.

The frequency response or gain curve of a simple half-wave dipole is too sharply peaked for satisfactory reception of more than one channel, or two at the most. When you are more than six or eight miles from the transmitters, the antenna is cut for the one or two channels to be handled. When closer to the transmitters a simple dipole may give satisfactory reception throughout the entire low band or the entire high band of very-high frequencies. Then the antenna is cut for the mid-band frequency of whichever band is to be handled.

The accompanying table lists overall physical lengths for simple half-wave dipole antennas cut for center frequencies of the very-high frequency channels, also for mid-band frequencies in the low and high bands.

LENGTHS OF SIMPLE HALF-WAVE DIPOLE ANTENNAS

		Antenna Length - Inches	
Channel Numbers	Center Frequency	Decimal Measurement	Common Fractions
2	57 mc	99.90	99 29/32
3	63 mc	90.30	90 19/64
4	69 mc	82.30	82 19/64
5	79 mc	72.10	72 3/32
6	85 mc	67.00	67

7	177 mc	32.15	32 5/32
8	183 mc	31.10	31 3/32
9	189 mc	30.10	30 3/32
10	195 mc	29.20	29 13/64
11	201 mc	28.30	28 19/64
12	207 mc	27.50	27 1/2
13	213 mc	26.75	26 3/4
Mid-band			
Low	69½ mc	81.90	81 29/32
High	194 mc	29.33	29 21/64

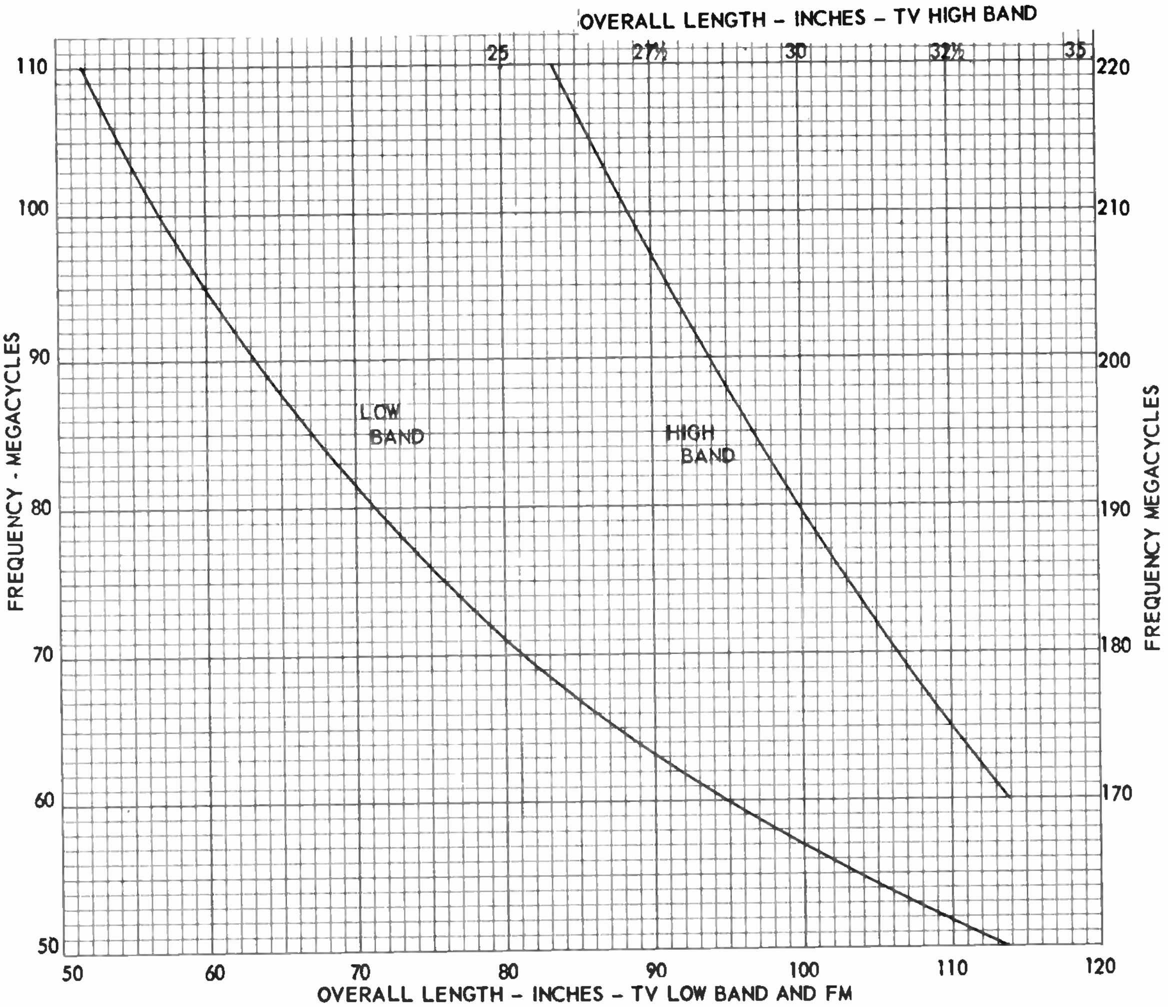


Fig. 65-7. Lengths of straight half-wave dipole antennas for resonance at various carrier frequencies.

Fig. 65-7 gives overall lengths for simple half-wave dipoles in the form of a graph. To obtain lengths in the low band of television frequencies use the left-hand curve with the left-hand vertical scale of frequency, and the bottom scale for length. To obtain lengths in the high band, use the right-hand curve with the right-hand vertical scale for frequency and the top scale for length.

There are a number of simple formulas which convert frequencies into corresponding wavelengths in inches or feet. Here are some of them.

One wavelength in space	inches $= \frac{11811}{mc}$	feet $= \frac{984}{mc}$
Half-wavelength in space	inches $= \frac{5905}{mc}$	feet $= \frac{492}{mc}$
Length of simple half-wave dipole	inches $= \frac{5690}{mc}$	feet $= \frac{468}{mc}$

BANDWIDTH AND GAIN. The frequency response or bandwidth of an antenna is affected by the same factors that determine frequency response of other kinds of resonant circuits. Especially important are changes of antenna impedance which occur with variations of carrier frequency. The frequency response of the antenna, or signal current at various frequencies, will be the reciprocal of antenna impedance.

As an illustration, assume that impedance of a certain antenna changes as shown at the left in Fig. 65-8. Then the relative response at the same frequencies will be as shown at the right. The response is shown in relative values or percentages of maximum, rather than in values of current, because actual current in amperes or microamperes will depend on signal strength and other things.

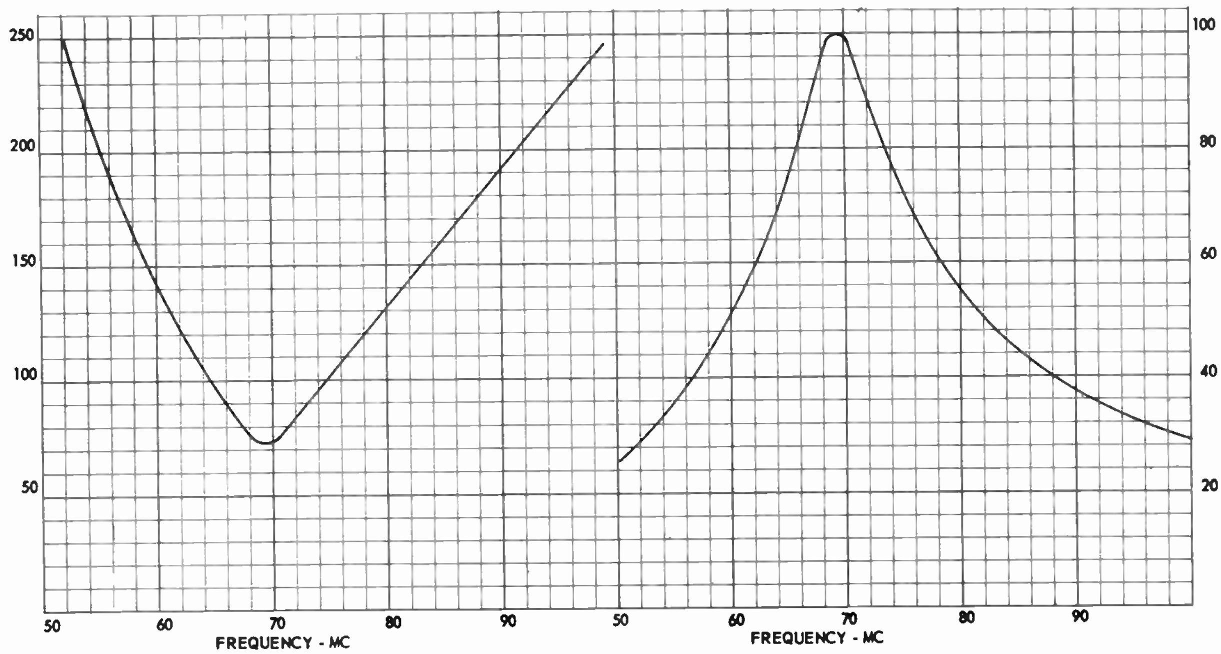


Fig. 65-8. Relations between antenna impedance and frequency response.

The response illustrated shows a sharp, narrow peak. This is because there is great change of impedance with small variations of carrier frequency. Such a response indicates an antenna designed, constructed, and used to have a high Q-factor – just as a sharply peaked resonance curve indicates a high Q-factor for any other kind of resonant circuit.

Supposing that, with this particular antenna, you wish to receive programs in channel 4, whose frequency range is from 66 to 72 mc. The response curve shows that at 66 mc we have 80 per cent of maximum response and at 72 mc have 88 per cent. Supposing further that receiver sensitivity allows for satisfactory reception only when the signal input is at least the amount realized from an 80 per cent response. Then you could have reception only from channel 4, because for all other channels the response would be less than 80 per cent of maximum. Remember, we are working with an antenna cut for one particular frequency or narrow range of frequencies.

If you move in closer to the transmitters, where field strength is greater, you might receive programs in channel 3 (60 to 66 mc) and possibly from channel 5 (76 to 82 mc). Also, you could have reception at the lower responses for these other channels by increasing receiver sensitivity, although interference and noise voltages then would be stronger in proportion to received signal strength.

Performance illustrated by Fig. 65-8 is that of a narrow band or single-channel antenna. To have satisfactory reception throughout the entire low band or entire high band at any great distance from transmitters, the antenna impedance would have to remain fairly constant over a much wider range of frequencies than it does with the antenna we are talking about. In order to receive programs in both the low and the high bands the antenna impedance would have to be even more constant with variations of frequency. The resulting broad band antenna would have to be of low-Q rather than high-Q design and construction. This would allow broader frequency response, but the gain or the signal currents would drop relatively low.

If we are to have satisfactory reception throughout a wider range of frequencies, and at greater distances from transmitters, the antenna must be modified or added to in some manner that allows a given field strength

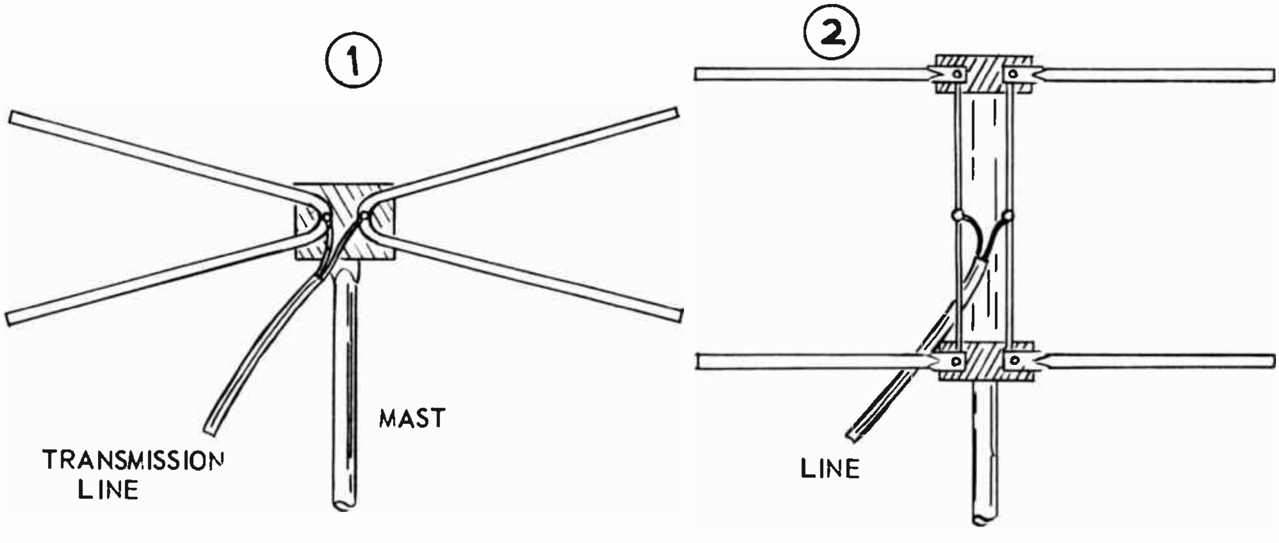


Fig. 65-9. Antennas which have greater pickup ability or greater gain than simple straight dipoles.

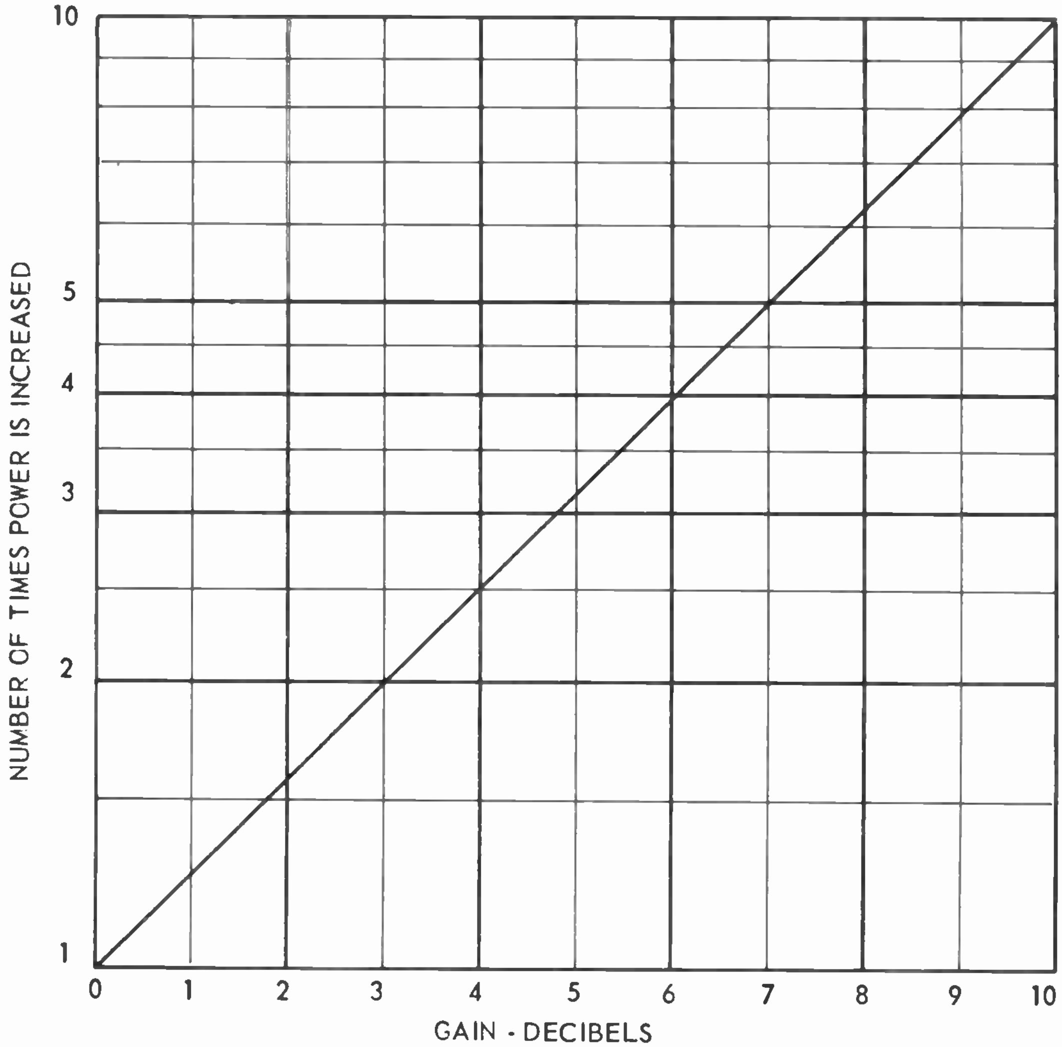


Fig. 65-10. A chart for making conversions between power gains in decibels and ratios of powers.

of carriers to produce more antenna current. That is, we must increase the gain of the antenna while retaining the broader frequency response. This is the principal object of most of the practical antenna designs which will be examined later.

Pickup ability may be increased in various ways. One way is to provide more conductors facing the oncoming carrier waves, possibly as at 1 in Fig. 65-9. This increases what we call the "frontal area" of the antenna. Another way, shown in one application at 2, is to use more than one set of antenna conductors in a "stacked" arrangement, with the sets of conductors or bays connected together and to the transmission, line so that their signal currents add together.

Large frontal areas lower the Q-factor and thus widen the frequency response while at the same time picking up more signal strength. The Q-factor may be lowered also by using conductors of larger diameter, as by using 1-inch tubing instead of a 1/2-inch or 3/8-inch diameter. Larger conductors reduce the Q-factor because they retain approximately the original inductance while increasing the capacitance because of greater surface area.

GAIN MEASUREMENTS. The gain of a particular television antenna is expressed as a number which shows how much stronger will be the signals obtained from this antenna than from some other type taken as a reference standard. As a general rule the reference standard for antenna comparisons is a simple half-wave dipole cut to suit the frequency at which the comparison is made. Both the measured and the reference antennas are presumed to be exposed to equal field strengths at the same frequency and to be so turned in relation to direction of wave travel that each has its maximum pickup. Unless the same reference standard is used for all comparisons the stated gains have no definite meaning, and they are not comparable.

Antenna gains are not stated as some number of times by which the signal power from a certain type exceeds the signal power from the reference antenna, but are given as so many decibels. A decibel is not a unit of power, like a watt, it is a unit that allows comparison of two powers as to their effectiveness in producing some desired result. Decibel measurements may be used also for comparing voltages or currents, but when dealing with antennas we use decibels for comparing their signal powers.

If antenna signal powers could be made to produce exactly proportional sound powers, the sound with a 2-decibel gain would seem twice as loud as with a 1-decibel gain, and it would take a 4-decibel gain to make the sound seem twice as loud as with a 2-decibel gain.

Fig. 65-10 allows conversion of gains expressed in decibels into equivalent ratios of two signal powers, power from the antenna in question and power from the reference standard. As an example, with an antenna whose gain is 4 decibels you can follow upward from 4 on the bottom scale to the diagonal line, then to the power ratio scale where it is shown that this antenna delivers 2½ times as much signal power as the reference standard. Reading the other way around, 5 times the power (on the left-hand scale) means a power gain of 7 decibels, on the bottom scale. The abbreviation for decibels is *db*, written with small letters, not capitals.

DIRECTIONAL PROPERTIES. At 1 in 65-11 a half-wave dipole antenna is in the path of carrier waves traveling at right angles to the length of the antenna. As we have learned, the wave energy is at all times assisting the antenna charges or current. At 2 the antenna has been turned parallel with the direction of wave travel. The carrier fields now can neither aid nor oppose the antenna current, and current quickly would die away even though some force were to start the charges moving in the first place. At 3 the antenna is at an angle of 45 degrees to the direction of wave travel. Since the waves across the antenna conductor diagonally, intervals between successive fields will not coincide with periods between natural reflections of charges. The charges or the antenna current will be assisted by wave energy only part of the time.

Antenna current will be maximum when the antenna is at right angles to wave travel, will be zero when antenna and wave travel are parallel, and will be something in between when the antenna is neither at right angles nor parallel to wave travel.

At 4 in Fig. 65-11 the antenna is "geographically" in the same position as at 1, but wave travel now is at an angle of 45 degrees to the antenna axis. Obviously, the signal strength from the antenna will be the same as at 3. At 5 the antenna is in the same position as at 1, but the wave travel is parallel to the antenna axis. This is exactly the same condition as at 2, and the antenna will pick up zero signal.

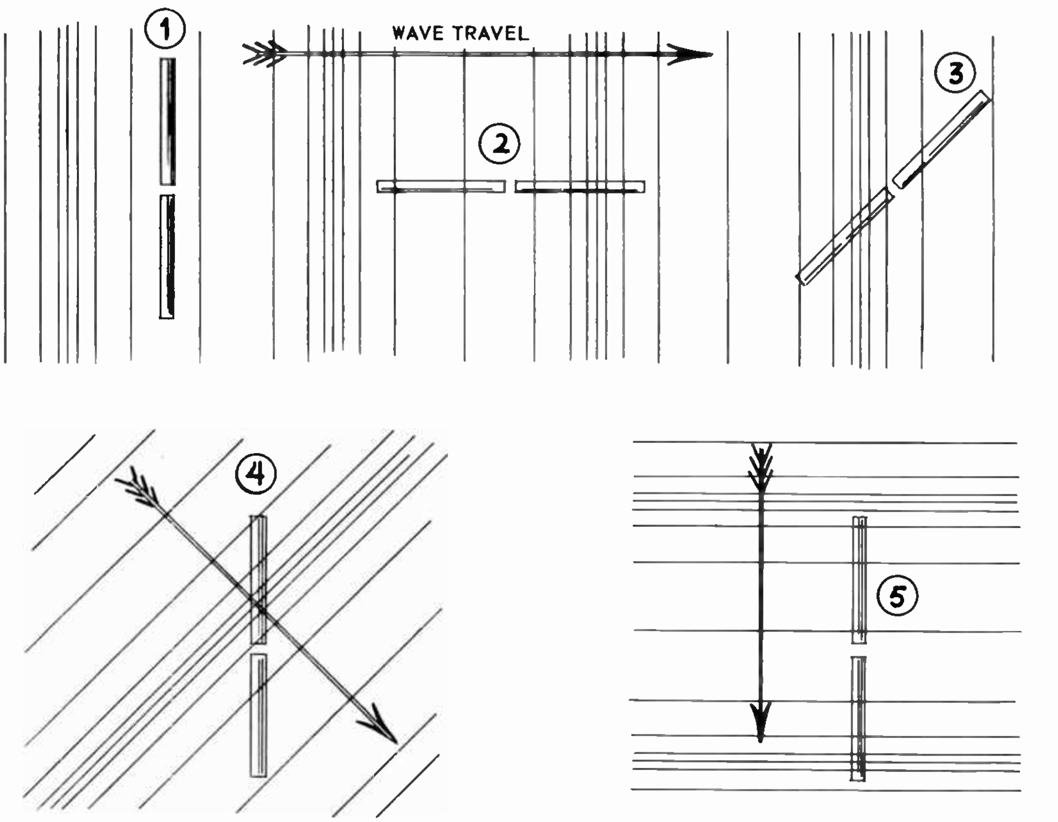


Fig. 65-11. How an antenna may be turned or oriented in relation to direction of carrier wave travel.

In the left-hand diagram of Fig. 65-12 the relative strengths of antenna signals with waves of equal strength approaching from the directions of the various arrows are proportional to the lengths of the lines on which are the arrowheads. Signal strengths from waves approaching in any directions are proportional to distances from the antenna center out to the broken-line curve.

If you look back at Fig. 65-11 it becomes apparent that relative signal pickups will not be altered if the direction of wave travel is reversed. This will be apparent upon turning that figure upside down. Then we may conclude that this antenna will pick up maximum energy from waves traveling in either of the two opposite directions with which wave direction is at right angles to the antenna axis. We must conclude also that pickup will theoretically be zero from waves traveling in either of the two opposite directions which are parallel to the antenna axis.

All this is shown by the "directional pattern" at the right in Fig. 65-12. The relative strengths of signals picked up from waves in all directions is indicated by the curved outlines, which are the same as the single broken-line curve around the arrows in the left-hand diagram. This is the pattern for a straight half-wave dipole used without any extra "parasitic" elements, and also for some of the other simpler kinds of antennas. It is the basic pattern for all television antennas, although great modifications may be made by altering the design and construction of antenna elements.

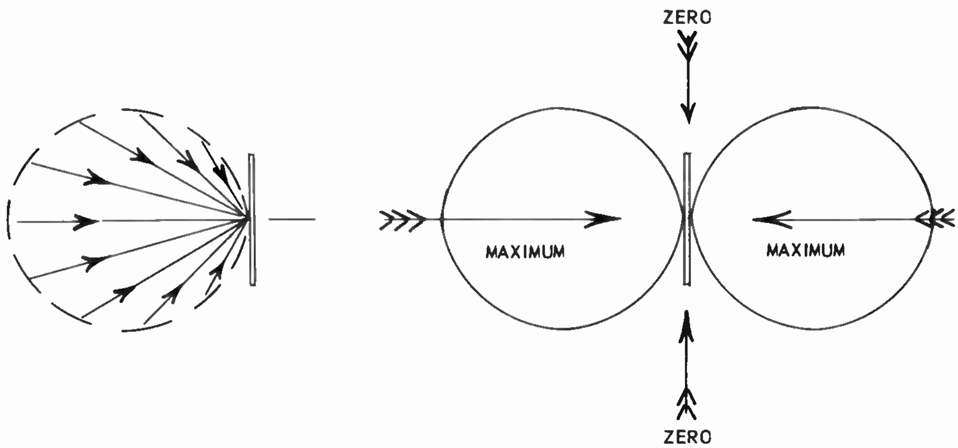


Fig. 65-12. Directional patterns for a straight half-wave dipole antenna.

If the conductors of a half-wave dipole were to extend north and south, as in Fig. 65-13, there would be the strongest possible signals from transmitters at A and at B. Were both transmitters of equal power there probably would be stronger signals from A than from B, because A is closer to the receiver. But in both cases the signals would be the strongest obtainable from these transmitters, because they are in directions at right angles to the antenna conductors.

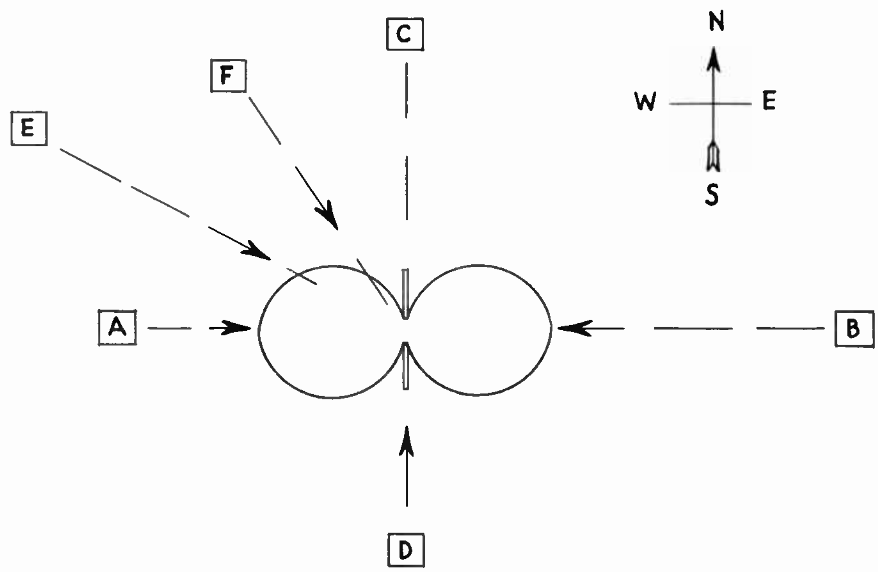


Fig. 65-13. How a directional pattern relates to signals from transmitters at various points of the compass.

There would be minimum possible signal pickup from transmitters at C and at D, because their directions from the receiver are on the axis of the antenna or are in line with the antenna conductors. Pickup from a transmitter at F would be weak, as you can tell by observing the point at which this signal line crosses the directional pattern. If we wish to have strong pickup from transmitter F the antenna may be rotated or oriented to bring its conductors at right angles to wave travel from this transmitter. Then there would be poor reception from transmitters A and B.

The antenna can be oriented for good pickup over quite wide angles from two opposite directions, but then the pickup will be weak for all transmitters not lying within these angles or arcs of circles. If sources of undesired radio signals are not in the same general direction from the receiver as are the television transmitters, the axis of the antenna conductors may be pointed toward the interfering signals. This will reduce or eliminate the interference while allowing reception of the television signals. As indicated by the directional pattern, the angle of zero and weak reception is much narrower than the angle of satisfactory reception. This often allows cutting out the interference while there is continued good reception from television transmitters lying in most directions from the receiver.

Directional patterns such as those of Figs. 65-11 and 65-12, and nearly all others applying to television antennas, are for carrier waves traveling on a line which is horizontal with respect to the position of the antenna in space. If the waves approach the antenna from an angle either above or below the horizontal, the pickup from directions in line with the conductors no longer is even approximately zero.

When a direct or reflected transmission wave comes to the antenna from a direction that is 5 degrees from a true horizontal line or plane through the antenna, the directional pattern will become about as shown at the left in Fig. 65-14. Still there is minimum pickup from directions in line with the antenna conductors, but this minimum is not very small. If a direct or deflected wave comes from 10 degrees away from a horizontal line the minimum pickup from directions in line with the antenna conductors will be still greater, as shown at the right.

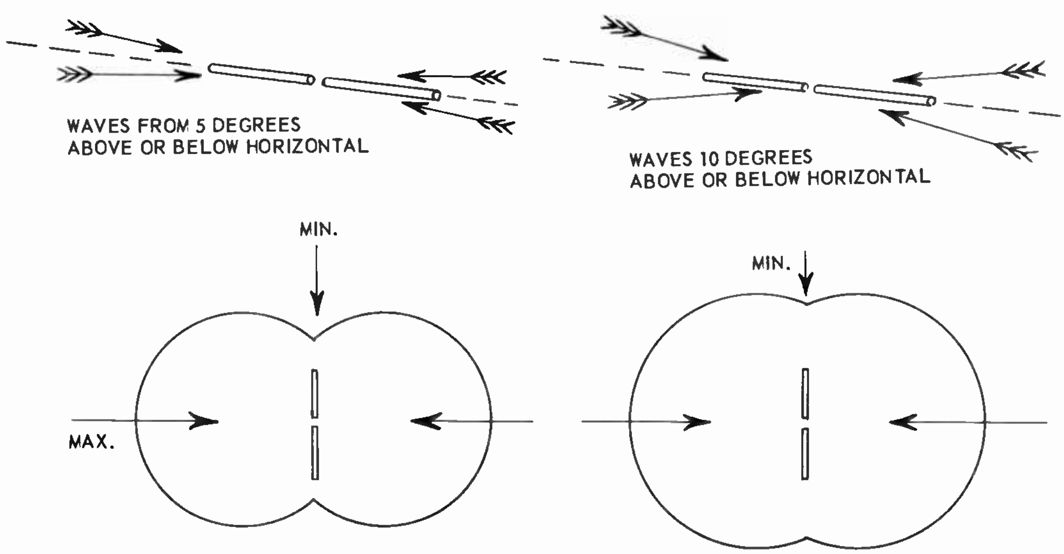


Fig. 65-14. The directional pattern has no sharp minimums for waves arriving from above or below the horizontal.

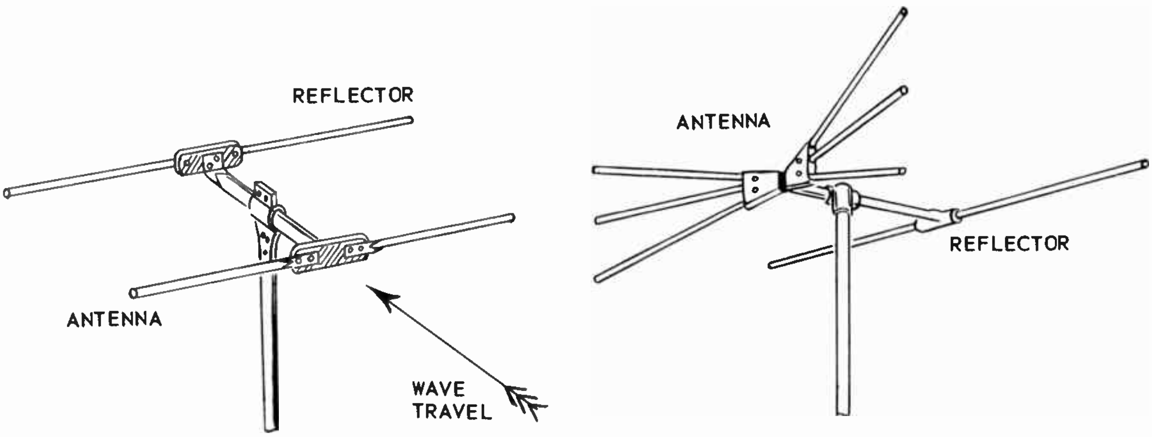


Fig. 65-15. How antennas and reflectors are mounted.

This explains why the effects of many kinds of electrical interference are reduced very little or not at all by attempts to orient the antenna so that its conductors are on a horizontal line through the source of interference. This type of interference seldom reaches the antenna along a horizontal line or plane, most often it comes from below the level of the antenna, and sometimes comes down from above.

REFLECTORS. An antenna which receives signals equally well from two opposite directions is said to be bi-directional. This characteristic is an advantage when stations whose signals are desired lie in approximately opposite directions from the receiver. But far more often most of the desired stations are in the same direction from the receiver or, at least, in the same general direction. Then we should like to increase the response to signals from these stations while reducing response in other directions to lessen the effects of any possible interference. Both these objects may be attained at the same time by using a reflector with the antenna.

5 A reflector, as shown by Fig. 65-15, is a conductor supported back of the antenna and parallel to the antenna conductors, or parallel to the mean direction of antenna conductors which are at symmetrical angles. The reflector is back of the antenna in relation to the direction of wave travel from transmitters which are to be received. That is, the reflector is placed so that desired carrier waves strike the antenna before they arrive at the reflector.

The reflector usually is made of the same kind and size of tubing as the antenna conductors. It may be in one continuous piece from end to end, or divided at the center if this is convenient for design and construction. A single straight reflector may be used with any form of antenna, or it may be of the same form as the antenna. The reflector may or may not be insulated from the mast and cross arm. If it grounds to the mast the electrical operation is not altered. The reflector is not conductively connected by any wire or cable to the antenna conductors, which are insulated from their supports.

The action of a reflector may be explained as follows. The antenna itself does not deliver to the receiver all the energy taken from passing carrier waves. Some of this energy, which has changed into antenna current, causes radiation fields around the receiving antenna just as such fields are caused around a transmitting antenna. The magnetic fields that move out and back around the antenna cut through the reflector and induce emf's and currents in the reflector. Then the reflector becomes surrounded with its own fields, which reradiate much of its received energy. A considerable portion of this reradiated energy goes back into the antenna.

If the reflector is suitably spaced from the antenna, and if the overall length of the reflector is suited to the spacing, the energy sent back to the antenna will arrive there in such phase or time relation to carrier wave energy that the two forces add in the antenna. The result is an increase of about three or possibly more decibels in antenna gain, as compared with the same form of antenna used without a reflector.

While the energy reradiated from reflector to antenna is reinforcing carrier wave energy coming from on front of the antenna, the reradiation arrives in such phase or time relation to wave energy from behind the antenna as to cause partial cancellation. For carrier waves which arrive first at the antenna, the direct and reradiated energies add in the antenna. For waves coming first to the reflector, their energy when reaching the antenna is largely cancelled by the out-of-phase reradiation from reflector to antenna. Were we to go into a detailed technical explanation of all these actions you would find that they depend on the fact that spacing between reflector and antenna allows energy from one wave or field to reinforce or cancel the energy from a following wave or field.

Addition of a reflector changes the directional pattern from the form at the left in Fig. 65-16 to the more nearly undirectional form at the right. The approximately circular outlines of these and all other directional patterns are called lobes. The reflector acts to enlarge one lobe and reduce the other. The enlarged lobe would be called the front lobe. It extends in the direction from which maximum signal strength is desired. There is a relatively small back lobe. A larger lobe may be called also a major lobe, and a smaller one a minor lobe.

The ratio between maximum pickup ability from the front and maximum from the back of the antenna is called the front-to-back ratio. The pattern at the left in Fig. 65-16 shows a front-to-back ratio of 1 to 1, because both lobes extend out to equal distances. The pattern at the right shows a front-to-back ratio of about $3\frac{1}{2}$ to 1. Carrier waves coming from in front would cause signal strength about $3\frac{1}{2}$ times as great as waves of equal field strength coming from the back.

Since a reflector increases signal pickup from the front and reduces it from the back, the antenna with reflector may be turned or oriented to lessen or eliminate ghost images caused by reflected waves from a direction approximately opposite to that of desired direct waves.

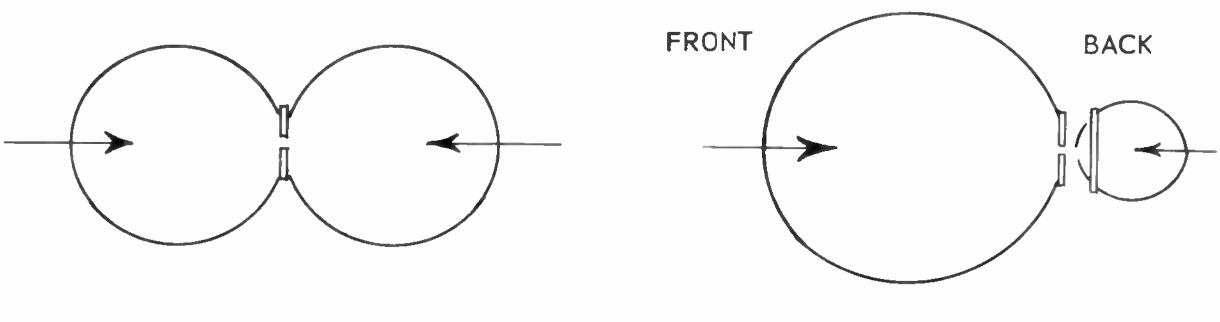
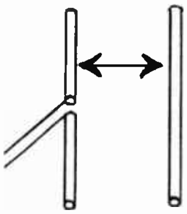
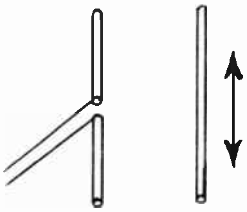


Fig. 65-16. A reflector improves the front-to-back ratio.





	REFLECTOR SPACING		REFLECTOR LENGTH	
	CLOSER	FARTHER	SHORTER	LONGER
GAIN	RAPID INCREASE	SLOW DECREASE	ACCORDING TO SPACING	
BANDWIDTH	NARROWER	WIDER	NARROWER	WIDER
FRONT-TO-BACK RATIO			POORER	BETTER
IMPEDANCE AT CENTER	DROPS RAPIDLY	DROPS SLOWLY		

Fig. 65-17. Effects on antenna characteristics of varying the reflector spacing and length.

Spacing between reflector and antenna elements is measured in fractions of a wavelength corresponding to that for which the antenna is cut, and at which it is resonant. The length of the antenna itself will be somewhat less than a half-wavelength at this frequency. The reflector may be placed anywhere from about 0.10 wavelength to more than 0.25 wavelength back of the antenna.

Spacing between reflector and antenna affects the gain, the bandwidth, the front-to-back ratio, and the minimum impedance of the antenna or the impedance at the gap where the transmission line is connected.

Fig. 65-17 shows in a very general way what happens to these characteristics when the reflector is moved closer to or farther from the antenna, and when the length of the reflector is altered. The chart relates to operation at or near the frequency for which the antenna is cut. Although a simple dipole and straight reflector are shown, the results of altering the spacing and reflector length would be much the same for other forms. You must keep in mind that effects listed on the chart are far from being independent of one another. When you change reflector spacing or length or both in an attempt to change one of the characteristics, all the others will be affected in greater or less degree at the same time.

Consider first the matter of gain. If the reflector is the same length as the antenna there will be maximum possible gain when the spacing is about 0.20 wavelength or a trifle more. Gain will drop quite rapidly as the reflector is moved closer to the antenna, and will drop less rapidly with greater spacings. But should you move the reflector closer to the antenna, and at the same time make the reflector longer than the antenna, there will be some certain length at which the gain becomes even greater than with the original 0.20 wavelength spacing.

If you move the reflector closer to the antenna in small steps, and at every change of spacing alter the reflector length until you obtain maximum gain at that spacing, the highest gain of all will be realized with a spacing of about 0.15 wavelength, and with the reflector maybe five to ten per cent longer than the antenna. Should you move the reflector farther than about 0.20 wavelength from the antenna, and at the same time make the reflector shorter than the antenna, the gain can be maintained at a value considerably higher than as though the reflector were not shortened.

Bandwidth or frequency response of the antenna becomes narrower as the reflector is moved toward the antenna, and also, to a limited extent, as the reflector is made shorter. Bandwidth tends to become wider as the reflector is moved away from the antenna, or is made longer than for maximum gain. As always happens, gain will drop when bandwidth increases, and will rise as bandwidth is narrowed.

Front-to-back ratio is affected chiefly by altering the reflector length. Length increase improves this ratio, but lessens the gain. A given alteration of length changes the front-to-back ratio much more than the gain, which allows obtaining a low back response without great loss of forward gain.

Impedance at the center of the antenna, for the frequency of resonance, is a highly important characteristic. When this impedance is the same as that of the transmission line and of the receiver input circuits there is maximum energy transfer from antenna to receiver. When all three impedances are not alike, and are not compensated by suitable matching devices, there is bound to be poor performance except where field strength is very high – and even then it is possible to have plenty of trouble from mismatched impedances between the line and receiver. We shall look into this matter of impedance matching a little later.

A reflector element always drops the antenna center impedance below the value without a reflector. A reflector may drop the impedance of a simple half-wave dipole antenna from 72 to about 60 ohms. Moving the reflector closer to the antenna drops the center impedance rapidly. Moving the reflector farther away drops the impedance more slowly.

Reflector spacing and length in ready-made antennas are such as will give most satisfactory performance with conditions which are assumed to be average. Unusual conditions cause less than optimum performance unless adjustments are made. The prime requirement might be reduction of pickup from behind the antenna, which would make front-to-back ratio all-important. At points far from transmitters the chief desire would be for high gain, at the sacrifice of bandwidth and front-to-back ratio. Closer to the transmitters, where gain is not needed, you might make adjustments for maximum bandwidth. Only expert technicians are capable of altering antenna adjustments to provide the most desired characteristics for customers living where conditions are not "average".

STRAIGHT DIPOLE ANTENNAS. Although the characteristics of the simple half-wave dipole have been explained as some length, a summary may be in order. When used without a reflector the rather limited pickup ability restricts this antenna to distances of something like five or six miles from transmitters. A reflector will extend the range to possibly eight or ten miles. These distance ranges assume that reception will be throughout either the low band or the high band, not both. At any greater distances the bandwidth narrows rapidly and, even with a reflector, the antenna becomes a single channel or two-channel type.

The simple straight dipole without a reflector has theoretical center impedance of between 72 and 73 ohms, and with a reflector about 60 ohms. Nearby objects of any material, and the details of antenna construction, may make the actual impedance almost anything between 50 and 100 ohms at the resonant frequency.

66 → 70

COMMERCIAL TRADES INSTITUTE

KNOWLEDGE + AMBITION = SUCCESS

LEARN . . .

Training Your Mind

C. T. I.

Perfecting Your Skill

ACT . . .

"The heights by great men reached and kept
Were not attained in sudden flight,
But they, while their companions slept,
Were climbing upward in the night."

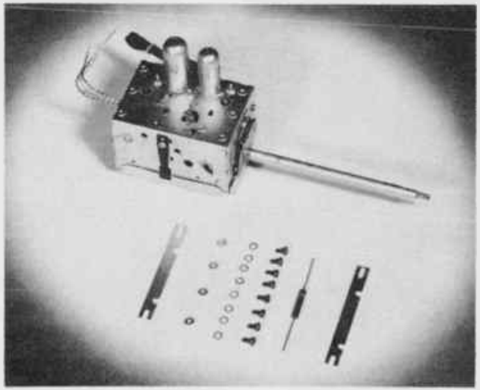
ENJOY . . .

COMMERCIAL TRADES INSTITUTE

CHICAGO, ILLINOIS

THE ORDER BLANK BELOW IS NOT TO BE USED UNTIL AFTER COMPLETION OF LESSON NO. 50

TELEVISION TUNER — KIT NO. 9



Television broadcasts throughout the country are being transmitted through the medium of a large number of channels allocated by the Federal Communications Commission. The tuner in a television receiver must have the ability to select the desired transmission and reject all others; this is a tremendous task when you consider the range of frequencies covered in these channels.

The Cascode Type Turret Tuner that you will receive with this kit provides more signal gain and less noise than many other types of tuners being presently used in television receivers. Its use in fringe areas is therefore highly recommendable. Easy adaptability to UHF reception is another of the many outstanding features of this tuner.

Kit No. 9 will be shipped upon receipt of your remittance of $3.00, which will be credited to your account. Please understand that while this payment reduces your balance to that extent, it is not considered as a monthly payment or part of a monthly installment.

Print your NAME, STUDENT NUMBER, and ADDRESS in INK. This is the Order Blank and Shipping Label for Kit 9.

TEAR-OFF TEAR-OFF

TELEVISION

LESSON NO. 66

TRANSMISSION LINES

COMMERCIAL TRADES INSTITUTE

Chicago, Illinois

Training Your Mind

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON NO. 66

TRANSMISSION LINES

No matter how effective an antenna may be in extracting signal power from passing carrier waves, this power is going to be of little use unless a good share of it is delivered to the tuner of the television receiver. As you may see from Fig. 66-1, signal power goes from the antenna to the tuner through a transmission line. The same signal current that is induced in the antenna moves in the transmission line and moves between the input connections of the tuner.

The antenna and transmission line are two parts of the same electrical circuit, and cannot well be studied as separate elements. For this reason we shall leave our examination of antennas for a short time while learning something about transmission lines. Only then will we be fully prepared to discuss the many varieties of antennas and the transmission line connections which must be used with them to have maximum transfer of signal power from carrier waves to tuner circuits.

Signal power is taken from a television antenna at the gap between opposite halves of the antenna. A conductor runs from each half to the tuner input circuit of the receiver. The two conductors, with necessary insulation and supports, form the transmission line. Because signal power taken from the carrier waves is so small, and because operating frequencies are so high, the design, construction, and installation of good transmission lines involves problems that don't exist in ordinary electric power lines, and that cause no serious difficulties even at standard broadcast frequencies.

4 There are three principal requirements for a good transmission line. First, it must allow signal power to pass in greatest possible amount from the antenna into the line and then from the line into the receiver. This requires matching the impedances of antenna, line, and receiver input. Second, there must be the least possible waste of power in getting signal currents through the line itself. This is a problem of reducing line attenuation. Third, the line must pick up in itself the least possible amount of interference and noise. This requirement is met by using a balanced, or else a shielded line.

The principle of a balanced transmission line is illustrated at 1 in Fig. 66-2. Note that the receiver input circuit is center-tapped, with the tap connected to ground. While current during any one portion of a signal cycle is moving one direction in one line conductor it is moving the opposite direction in the other line conductor, but in both halves of the tuner input this current moves in only one direction.

What happens with interference pickup is shown by diagram 2. Since the two conductors of the transmission line are close together, the interference fields induce equal emf's in both. Interference emf's

COMMERCIAL TRADES INSTITUTE

TELEVISION

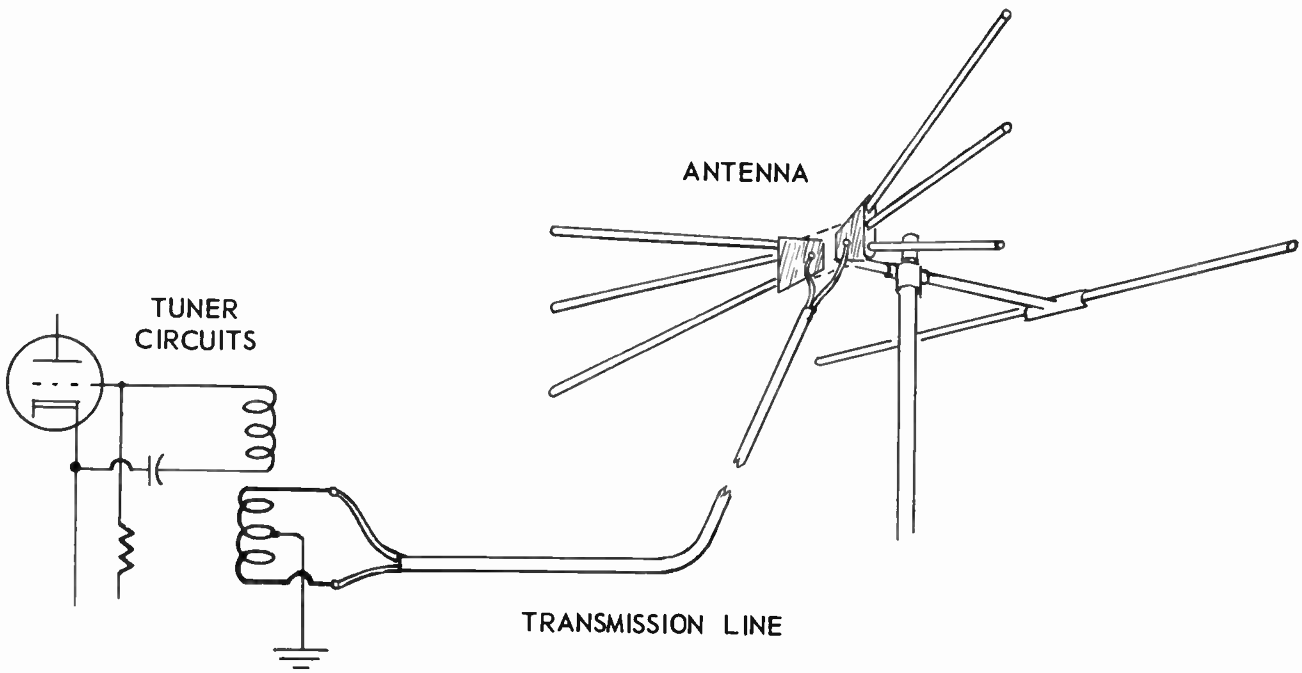


Fig. 66-1. The parts of an antenna circuit.

then act downward, or upward, at the same time in both conductors. These emf's oppose and balance in the two halves of the receiver input, or are dissipated to ground from the center tap.

The principle of an unbalanced shielded transmission line is illustrated by diagram 3 of Fig. 66-2. The transmission line here consists of a single central conductor surrounded by insulation. Around the outside of the insulation is braided copper which acts as the second conductor. At any one instant signal current from the antenna moves opposite directions in the central conductor and the braided conductor, but flows the same direction in the tuner input. The tuner input circuit is completed through ground to the braided conductor of the line.

When interference fields are present, as in diagram 4, they cannot reach the inner conductor because the outer braided conductor is acting as a grounded metallic shield. Interference currents induced to the braided shield pass to ground and do not affect reception.

The type of transmission line illustrated by diagrams 1 and 2 of Fig. 66-2 is called unshielded twin-conductor. The type shown by diagrams 3 and 4 is called coaxial line or coaxial cable, because the axes of both conductors lie together. Diagram 5 illustrates the principle of a shielded twin-conductor transmission line or a balanced and shielded line. The antenna is connected to the tuner input through two conductors, both insulated from, but wholly enclosed within a braided copper shield that is grounded. Signal current from the antenna flows in the same manner as at 1. Interference fields cannot pass through the shield, and their energy is carried to ground.

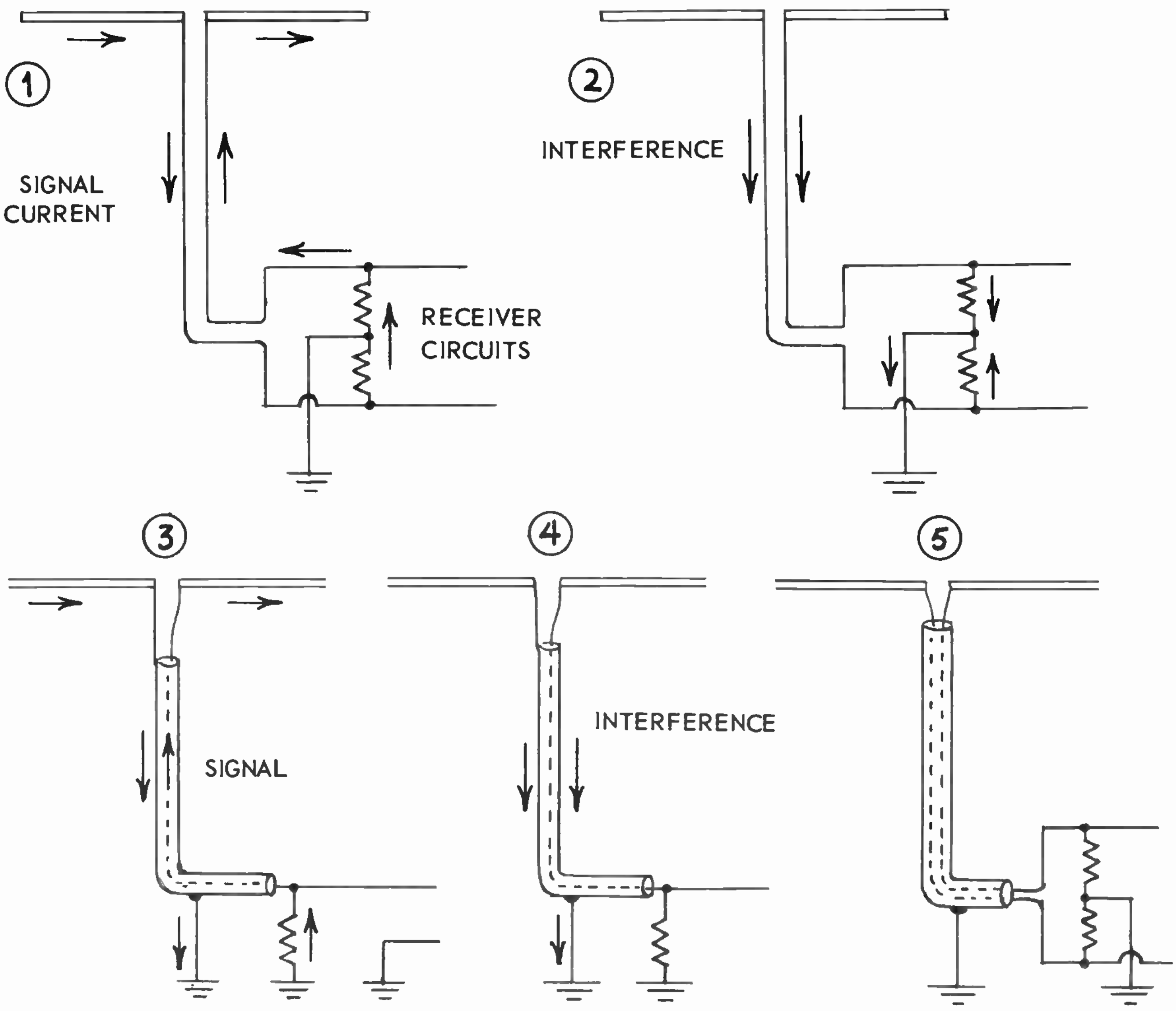


Fig. 66-2. Connections of balanced and unbalanced transmission lines.

Two kinds of unshielded twin-conductor transmission lines are illustrated at the top in Fig. 66-3, and down below are two kinds of coaxial cables. The ends of the conductors have been bared to show the construction. The twin conductors are stranded copper wires, each surrounded by a thin covering of low-loss polyethylene insulation, and with a thin web of this insulation between the conductors to maintain spacing and provide support. The coaxial line consists of a center conductor of either solid or

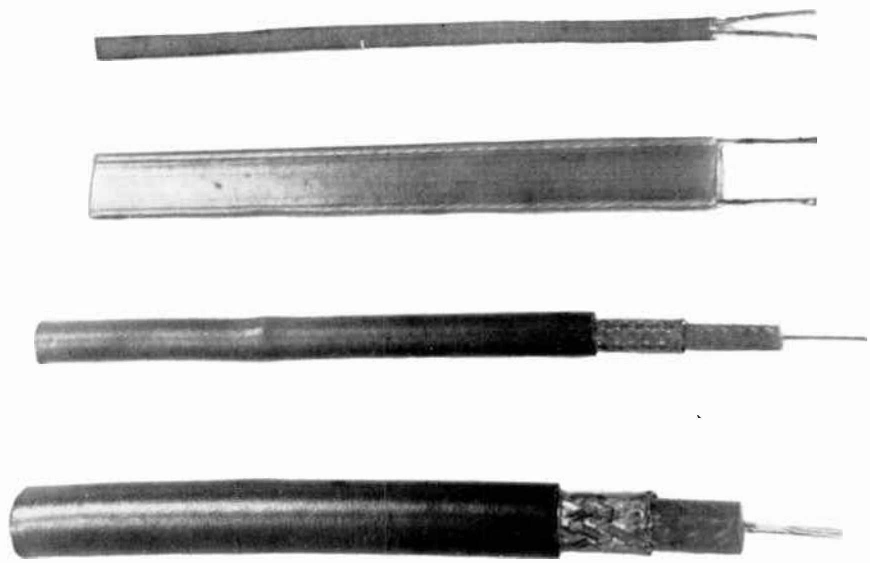


Fig. 66-3. Unshielded twin-conductor transmission lines (above) and coaxial cables (below).

stranded copper wire which is surrounded by polyethylene insulation. Over this insulation is the cylindrical copper braid, and on the outside of the braid is a covering of tough, waterproof vinyl plastic.

IMPEDANCE OF TRANSMISSION LINES. The first step toward understanding the behavior of transmission lines is a study of their highly important property called characteristic impedance. This impedance, like all others, is opposition to flow of alternating current, which in the present case is signal current. Impedance differs from ohmic resistance in that the opposition called impedance is due largely to inductive and capacitive reactances, as well as to ohmic resistance. Reactances vary with frequency. Therefore, ordinary impedances vary with frequency. Ohmic resistance, which accounts for part of ordinary impedances, varies with length of conductors. Therefore, ordinary impedances vary with conductor length.

In transmission lines there are inductance and capacitance, and we may use longer or shorter lengths. Yet the characteristic impedance of a transmission line does not vary with change of signal frequency, nor does it vary with length of line.

The paralleled conductors of a transmission line possess inductance, which may be measured in microhenrys or any similar unit. In a line of uniform structure along its whole length the inductance increases and decreases directly with increase and decrease of conductor length. We may say that there is some certain amount of inductance per foot of line, or per inch or any other unit of length.

The two conductors of the transmission line act like the two plates of a capacitor, with air or other insulation between them for the dielectric. Then the line must possess some certain amount of capacitance. If we double the length of line we have twice the capacitor plate area with the same kind and thickness of dielectric, so must have doubled the capacitance. Halving the length of line must halve the capacitance. Then it must be true that the line has some definite amount of capacitance per foot or other unit of length.

Let's consider one foot of transmission line. The inductance and capacitance in this unit length of line are represented by symbols for coils and a capacitor between points a and b of Fig. 66-4. In a transmission line made with any given size of conductors spaced a certain distance apart and separated by some particular kind of insulation or dielectric material, the characteristic impedance is proportional to the square root of the quotient of dividing henrys of inductance by farads of capacitance. We may substitute microhenrys of inductance, if we substitute microfarads of capacitance. Then we have,

$$\text{Approximate impedance in ohms} = \sqrt{\frac{\text{inductance}}{\text{capacitance}}}$$

Supposing that the inductance per foot of a certain transmission line is 0.54 microhenry, and that the capacitance per foot is 0.000006 microfarad. Dividing 0.5 (or 0.540000) by 0.000006 gives 90,000. The square root of 90,000 is 300, which is the approximate impedance of the foot of line, in ohms.

Now we shall add another foot of exactly the same kind of line, as from b to c in Fig. 66-4. This doubles the inductance, which becomes 1.08 microhenrys. The capacitance also is doubled, and becomes 0.000012 microfarad. Dividing this increased inductance by the increased capacitance gives 90,000. The square root of 90,000 is 300, just as before. So we haven't altered the characteristic impedance by adding more line.

No matter how much the line may be lengthened, and no matter how much it may be shortened, the characteristic impedance will remain unchanged so long as the line remains of the same dimensions and

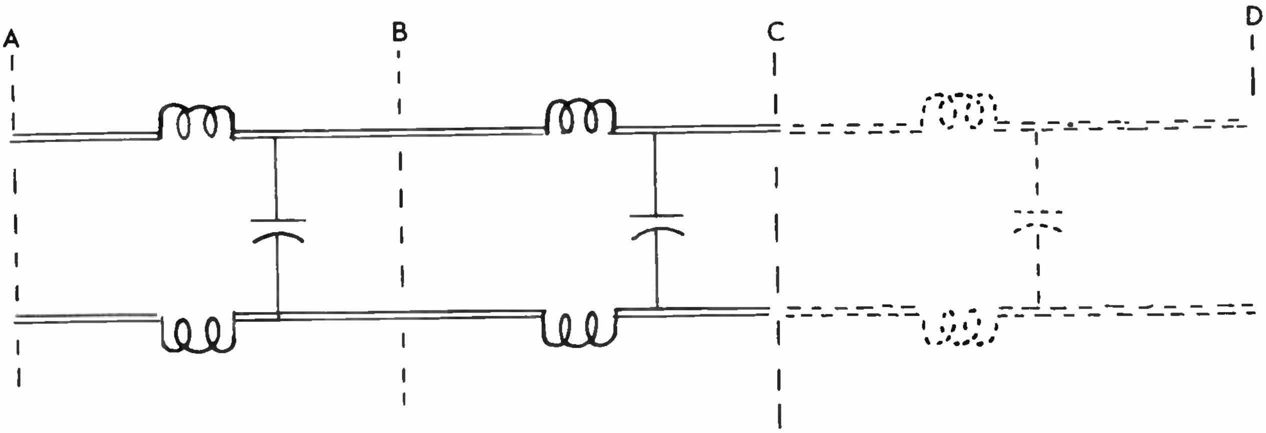


Fig. 66-4. Inductances and capacitances per unit lengths of transmission line.

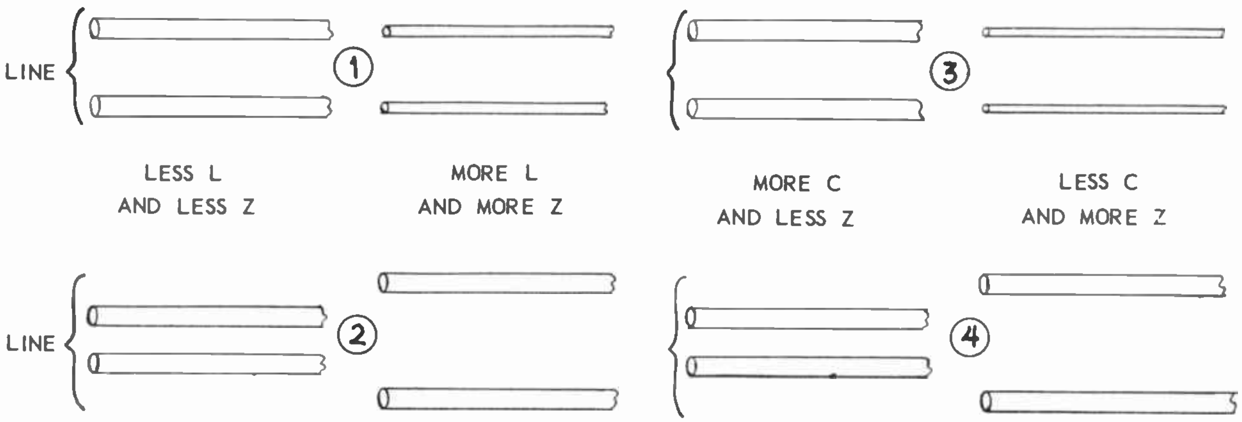


Fig. 66-5. How characteristic impedance is altered by varying the diameter and spacing of transmission line conductors.

materials. We might continue adding length, as from c to d, or might remove any portion of any total length, and always would have 300 ohms of characteristic impedance so long as there is no change of conductor diameter or spacing and no change in the kind of dielectric.

When computing the characteristic impedance of the transmission line we gave no consideration to freqnency. No term for frequency appears in the formula for characteristic impedance, and values derived from the formula could not be altered by differences of frequency. Therefore, we must conclude that characteristic impedance is not altered by any variations of frequency, it remains the same at all frequencies.

The characteristic impedance of a transmission line depends on the relations between inductance and capacitance in any unit length of line. The unit length may be any number of feet, inches, or fractions. As shown in Fig. 66-5, line impedance (Z) is reduced by less inductance and is increased by more inductance. As at 1 the inductance (L) is reduced by using larger conductors and is increased by smaller conductors when center-to-center spacing remains unchanged. As at 2, the inductance is reduced by bringing the two conductors closer together and is increased by moving them farther apart when conductor diameter is unchanged.

Characteristic impedance is reduced by more capacitance and is increased by less capacitance. At 3 in Fig. 66-5, capacitance is increased by nsing larger conductors and is decreased by using smaller conductors when center to center spacing between conductors remains unchanged. At 4 the capacitance is increased by moving the conductors closer together and is decreased by moving them farther apart when conductor diameter is unchanged.

Diagrams 1 and 3 are the same, as also are 2 and 4. Each is repeated to more clearly illustrate the the effects of conductor size and spacing on inductance, capacitance, and characteristic impedance.

Fig. 66-6 shows some relations between characteristic impedance, conductor diameter, and conductor spacing when the dielectric is air, or when the two conductors are supported with the least possible solid dielectric within their fields. Note that increasing the spacing between two number 8 wires (of about 1/8 inch diameter) from one-half inch to ten inches increases the characteristic impedance from about 250 ohms to more than 600 ohms. This graph shows that we may keep a constant center to center separation and increase the characteristic impedance by using smaller and smaller conductors. It is

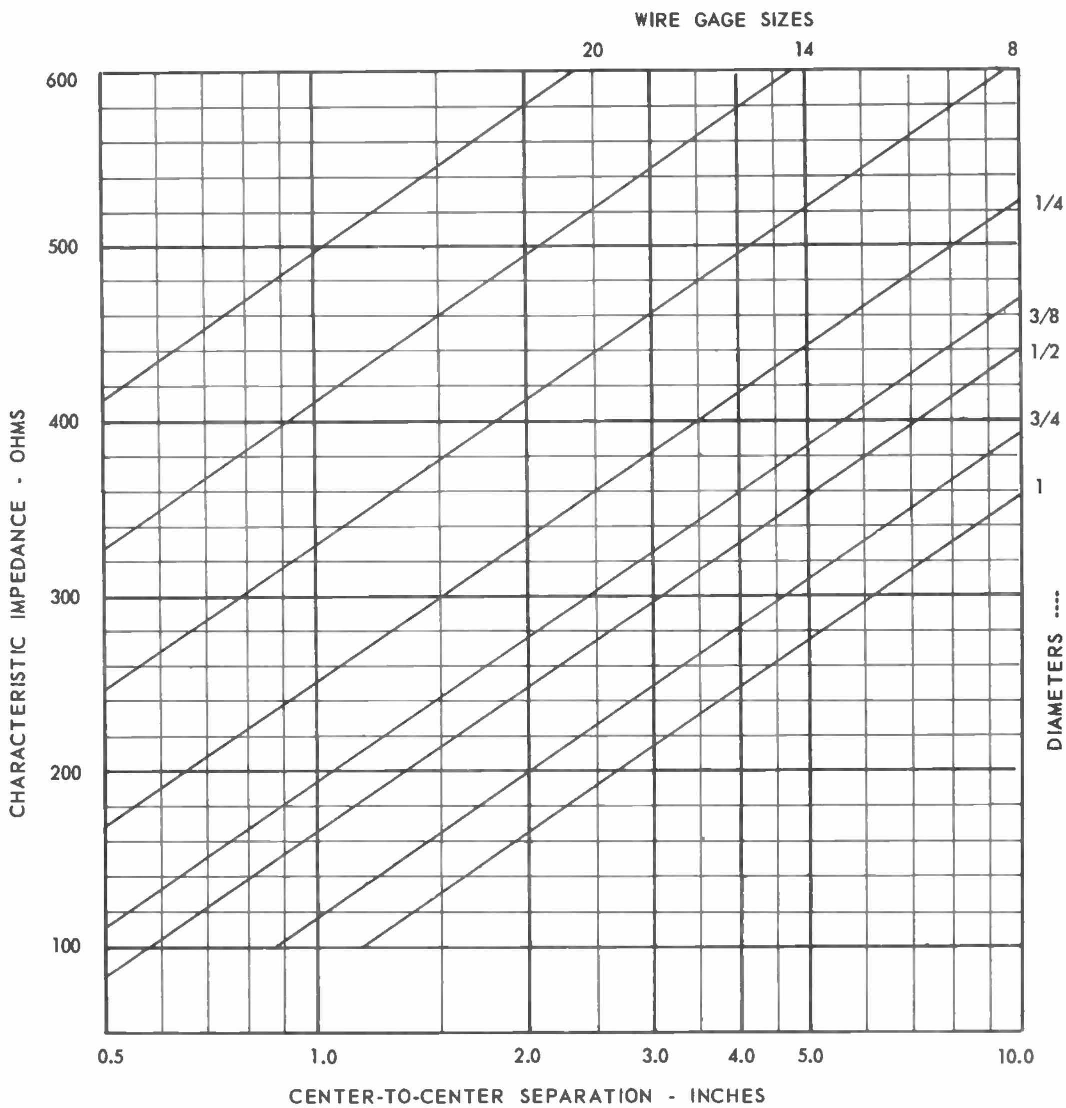


Fig. 66-6. Relations between characteristic impedance, spacing, and diameters of conductors in air-insulated transmission lines.

possible also to have the same impedance with larger and larger conductors if we keep increasing the spacing between them. Conductors of any diameter may be solid or may be of hollow tubing without affecting the values of characteristic impedance.*

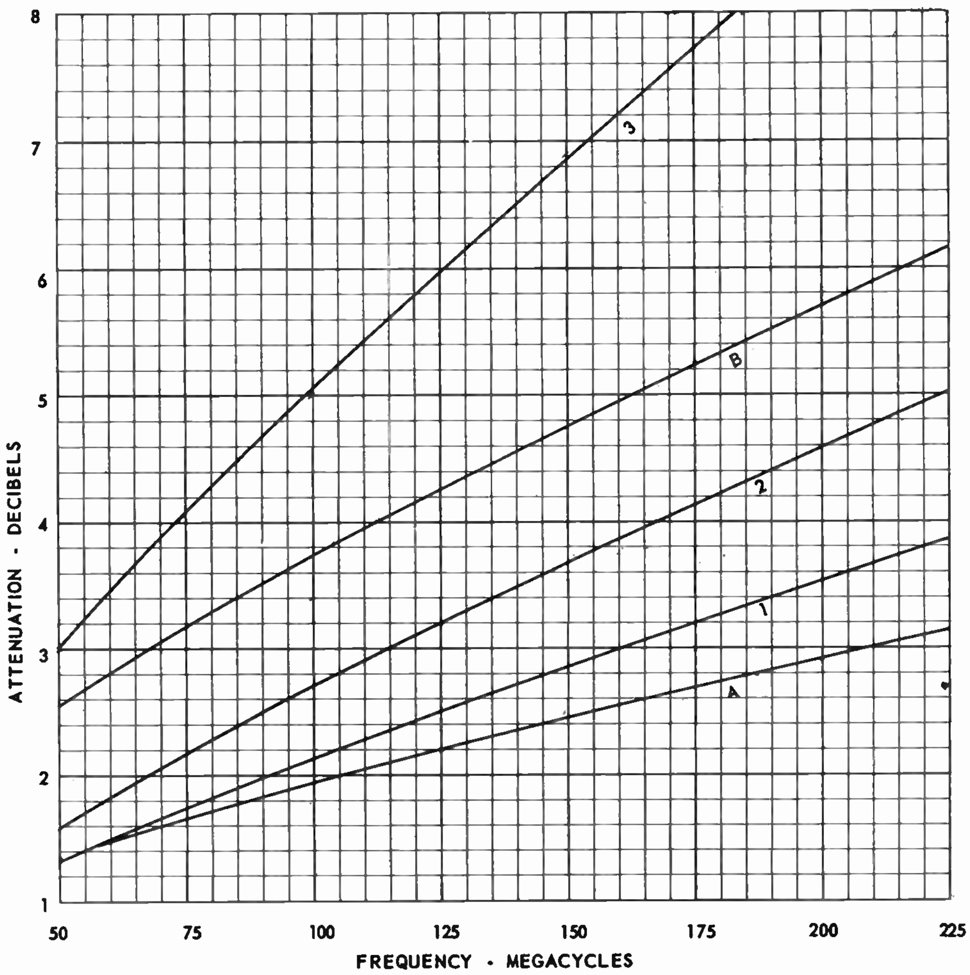


Fig. 66-7. Attenuations in decibels per 100 feet of several kinds of transmission lines.

* The graph of Fig. 66-6 is computed from this formula.

$$\text{Characteristic Z, ohms} = 276 \log \frac{2 \text{ x center to center spacing}}{\text{outside diameter}}$$

This means 276 times the common logarithm of the ratio of twice the spacing to the diameter, with all dimensions in the same unit, usually inches. With a table of logarithms you can compute values for any dimensions.

5 You will find that transmission lines most often are of types having characteristic impedance of 300 ohms. This is because the input impedance of television receivers is fairly well standardized at 300 ohms, and by making the impedances of line and receiver equal, all the signal power brought to the receiver input by the line goes into the tuner to produce pictures and sound.

6 It is true also that most receiving antennas are designed to have center impedances of 300 ohms or thereabouts, or are capable of working into a 300-ohm line. When impedances of antenna and line are equal there is maximum possible transfer of signal power from antenna to line. With maximum transfer, half the signal power picked up by the antenna from carrier waves goes into the transmission line. The other half is reradiated from the antenna.

ATTENUATION OF TRANSMISSION LINES. When signal currents flow in the conductors of a transmission line a very small part of the signal power is used to overcome conductor resistance. Much more power disappears in the various high-frequency losses which are discussed in other lessons. If insulation between the conductors is other than air, as almost always is the case, the greater part of the line loss occurs in the insulation, considered as a dielectric. Regardless of how great or how small may be the line loss, it is strictly proportional to length of line. Doubling the length of line doubles this loss, while halving the length halves the loss.

Signal power loss in the transmission line is called attenuation of the line. Attenuation usually is measured and specified as so many decibels of power per 100 feet of line. If the attenuation is 2 decibels in 100 feet of line it is 1 decibel in 50 feet and is 4 decibels in 200 feet. Fig. 66-7 shows approximate attenuations in decibels per 100 feet for several kinds of transmission line.

Curve A applies to a coaxial line having an outside diameter of about 0.4 inch, as illustrated at the bottom of Fig. 66-7. Curve B applies to the smaller coaxial line having an outside diameter of about 0.25 inch. Both these lines have characteristic impedance of 73 to 75 ohms. Curves 1, 2, and 3 apply respectively to unshielded twin-conductor lines of 300-ohm, 150-ohm, and 75-ohm impedance. Whatever the kind of line and whatever its impedance, the attenuation always increases as operating frequency becomes higher.

Fig. 66-8 allows conversion of attenuations in decibels into equivalent percentages of original power which has been lost or of percentages of the original power which remain. You may determine the percentage of loss by using the left-hand vertical scale with the bottom scale of decibels. As an example, with attenuation of 4 decibels the loss is 60 per cent of the original power. For power percentages remaining you should read the right-hand vertical scale. With attenuation of 4 decibels the remaining power is 40 per cent of the original power.

It is much easier to determine the total effect of several losses by using decibels than percentages. Several separate and successive attenuations in decibels are simply added together to find the total loss in decibels. Percentages must be multiplied together to find the total. This is illustrated by the follow-example.

Assume that there is a loss of 37 per cent or 0.37 of the original power. This leaves 63 per cent or 0.63 of the original power. Now supposing there is a second loss of 37 per cent in the same circuit. This second loss will not be of so many watts as the first one, for in the second case we take off only

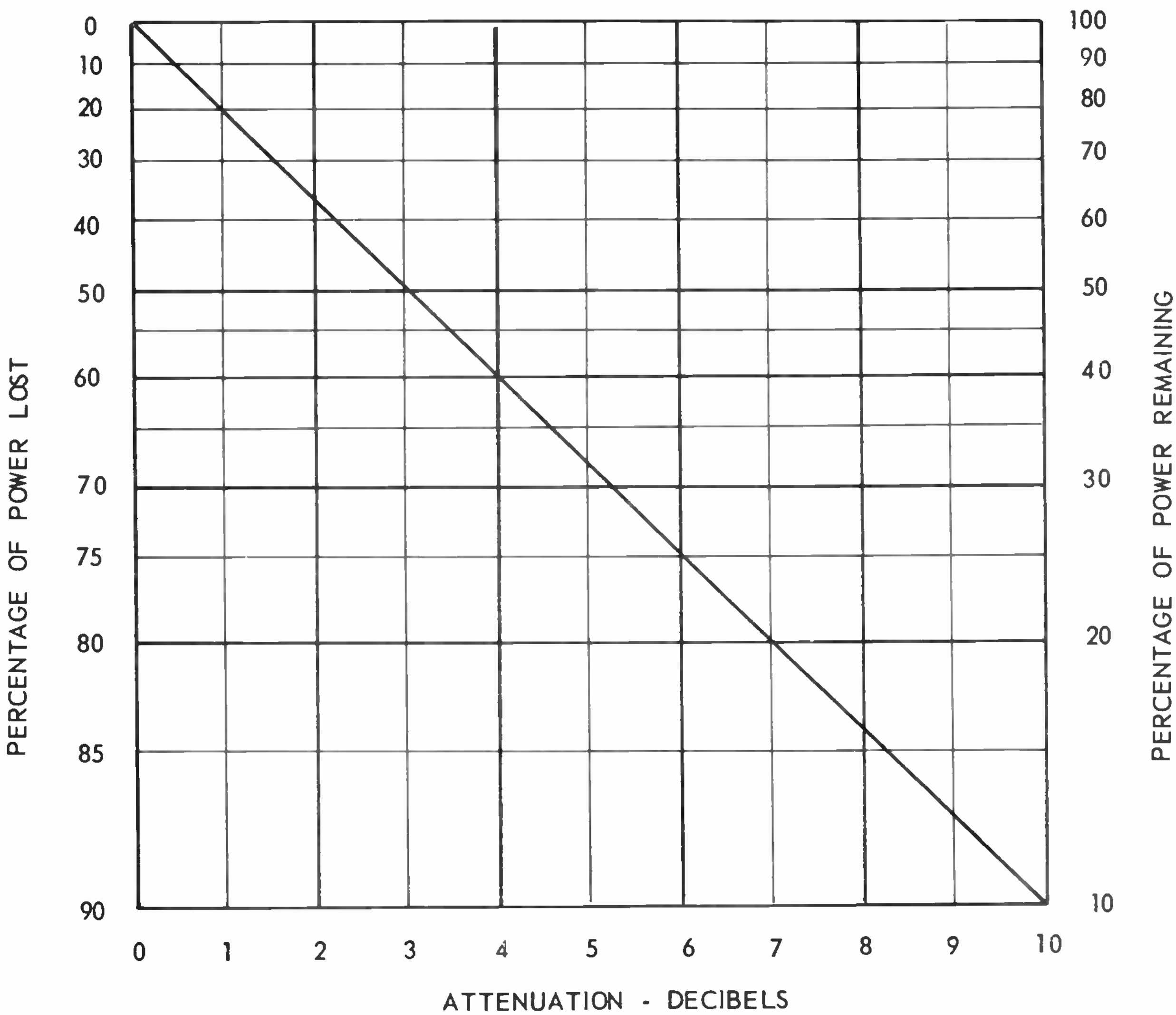


Fig. 66-8. Conversion chart for attenuation in decibels and percentage of power lost or remaining.

37 per cent of the 63 per cent that remained after the first loss. Now 37 per cent of 63 per cent, or 0.37 times 0.63, is 23 per cent or is 0.23. Subtracting 0.23 from 0.63 leaves 0.40 or 40 per cent of the original power after the two losses or attenuations.

An attenuation of 37 per cent is a loss of 2 decibels, from Fig. 66-8. If there is a following loss of another 2 decibels we may find the total loss in decibels by adding 2 to 2 for a total of 4 decibels. The graph shows that a loss of 4 decibels leaves 40 per cent of the original power, as was determined by the longer process of multiplying and subtracting percentages or decimal fractions.

NEED FOR IMPEDANCE MATCHING. There will be less than maximum possible transfer of signal power from antenna to transmission line when the impedances of these two parts are not alike. Also, there will be less than maximum possible power transfer from line to receiver when impedances of line and receiver are unequal. The reason is illustrated in a general way by Fig. 66-9 and explained as follows.

Consider the antenna to be our source of power, like a battery, a generator, or any other source. Every power source produces within itself an emf. Emf in the antenna results from energy taken from passing carrier waves. Every power source has more or less internal resistance or impedance, which here is the impedance of the antenna. Our power source is connected to a load, which is the transmission line. All loads have some certain resistance or impedance, which here is the characteristic impedance of the transmission line.

In diagram 1 we are assuming that carrier waves cause an emf of 18 volts in the antenna or source. This is an impossibly high emf, but it will avoid unwieldly decimal fractions and won't affect the principles involved. Impedances of antenna and line, or source and load, are matched. Both are 300 ohms. Total circuit impedance in source and load is 600 ohms. This is the impedance that opposes signal current in the source and the load. An emf of 18 volts acting in 600 ohms total impedance causes current of 0.03 ampere in the source and in the load.

Power in watts is equal to current squared and multiplied by impedance. Working this out for the source or antenna shows that this part of our circuit is consuming 0.27 watt of power. Since current and impedance in the load or the line are the same as in the source, the line must be taking 0.27 watt of power. Note, incidentally, that the line is taking exactly half the total power. The other half is used up in the antenna itself, in the impedance of the antenna or in causing reradiation of signal energy into surrounding space.

In diagram 2 we have the same emf as before, since carrier field strength is supposed to remain constant, but now we have a 600-ohm line or load connected to the 300-ohm source. Total circuit impedance is increased to 900 ohms. Emf of 18 volts in 900 ohms causes current of 0.02 ampere in both source and load. Power used in the source now is 0.12 watt, and in the load is 0.24 watt. This is less total power than in diagram 1, and although the load now takes more power than the source, the absolute value of power into the load or line is less than before.

Now look at diagram 3. Impedance of the source or antenna has been increased to 400 ohms. Load or line impedance has been dropped to 200 ohms. Total impedance is 600 ohms, just as in diagram 1, and we have the original current of 0.03 ampere in source and load. Total power is the same as in diagram 1, it is 0.54 watt. Power consumed in the source has gone up to 0.36 watt, while power put into the load or line has dropped to 0.18 watt - much less than in diagram 1.

If we were to keep the impedance of the antenna or source at 400 ohms, and increase the impedance of the line or load to 400 ohms, the total power would be about 0.40 watt. Half would be used in the a. tenna and half would go into the line. You never can get more than half the total power into a load, and you can get half into the load only when impedances of source and load are equal.

The percentage of total available power put into a load from a source will decrease whenever the two impedances are unequal. This is true whether load impedance is greater or less than source impedance.

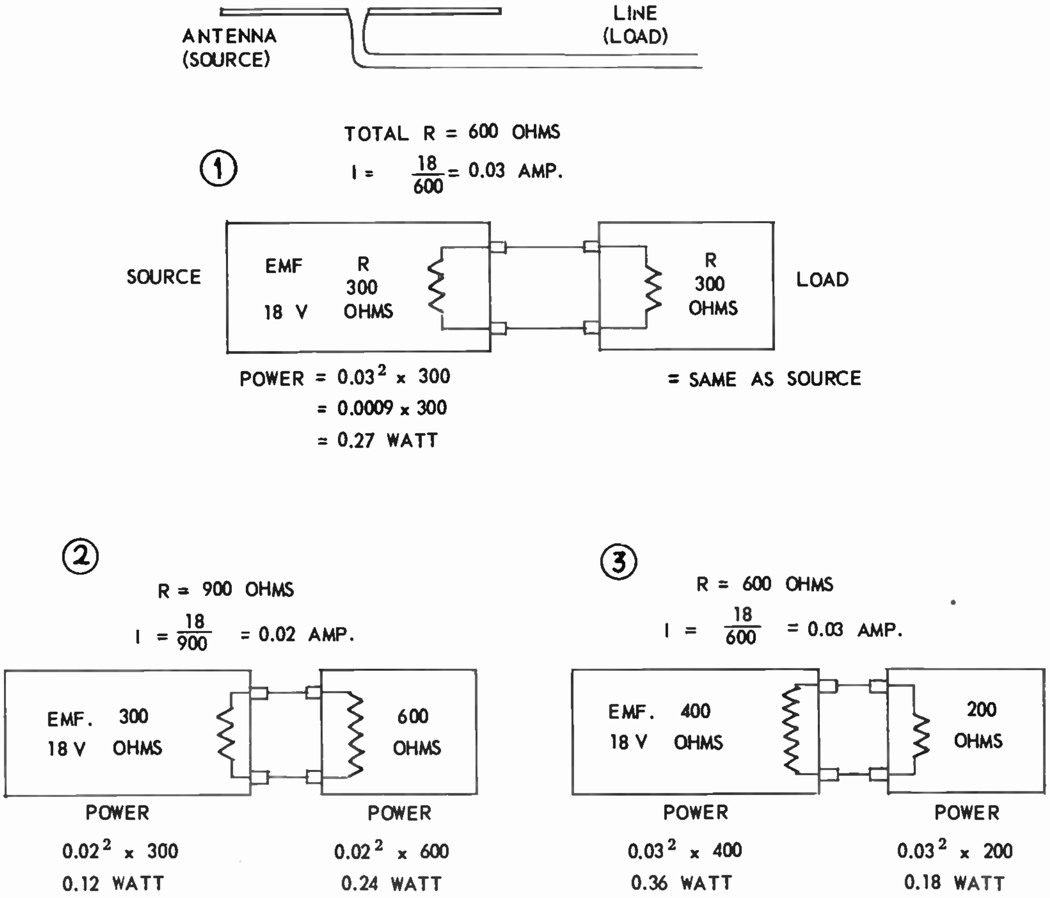


Fig. 66-9. Why the impedances of antenna and line, and of line and receiver, must be equal to have maximum possible transfer of signal power.

Because half of the total power taken from passing carrier waves is the maximum power that can be taken from the antenna into the transmission line we may as well consider this to be 100 per cent of the available signal power. Fig. 66-10 shows how power into the line is reduced by mismatch between impedances of antenna and line.

The bottom scale shows the ratio of the two impedances, or the ratio of mismatch, which is the number of times that either impedance exceeds the other impedance. It makes no difference whether the line impedance is greater or less than the antenna impedance, only their ratio counts in loss of available power. To read the percentage of available power transferred into the line we use the bottom scale and the left-hand vertical scale. To read the percentage of available power lost due to mismatched impedances we use the bottom scale and the right-hand vertical scale.

As an example, assume that you connect a 300-ohm transmission line to an antenna whose center impedance is 75 ohms. This is a mismatch of 4 to 1, because 300 is 4 times 75. The graph shows that this amount of mismatch drops the power transfer to 36 per cent. Were you to use a 150-ohm line with the 75-ohm antenna the mismatch would be 2 to 1. Then about 88.5 per cent of available signal power would go into the line, and only 11.5 per cent would be lost. This is not a serious loss, but the loss with the 4 to 1 mismatch can be serious.

How much mismatch may exist while still having satisfactory reception depends largely on average signal field strength or carrier strength at the antenna location. Ordinarily, if you are within five to eight miles of transmitters the mismatch might be as great as 4 to 1 and still leave plenty of signal power for the line and receiver. At greater distances or wherever the field strength is low the mismatch should be little if any greater than 2 to 1.

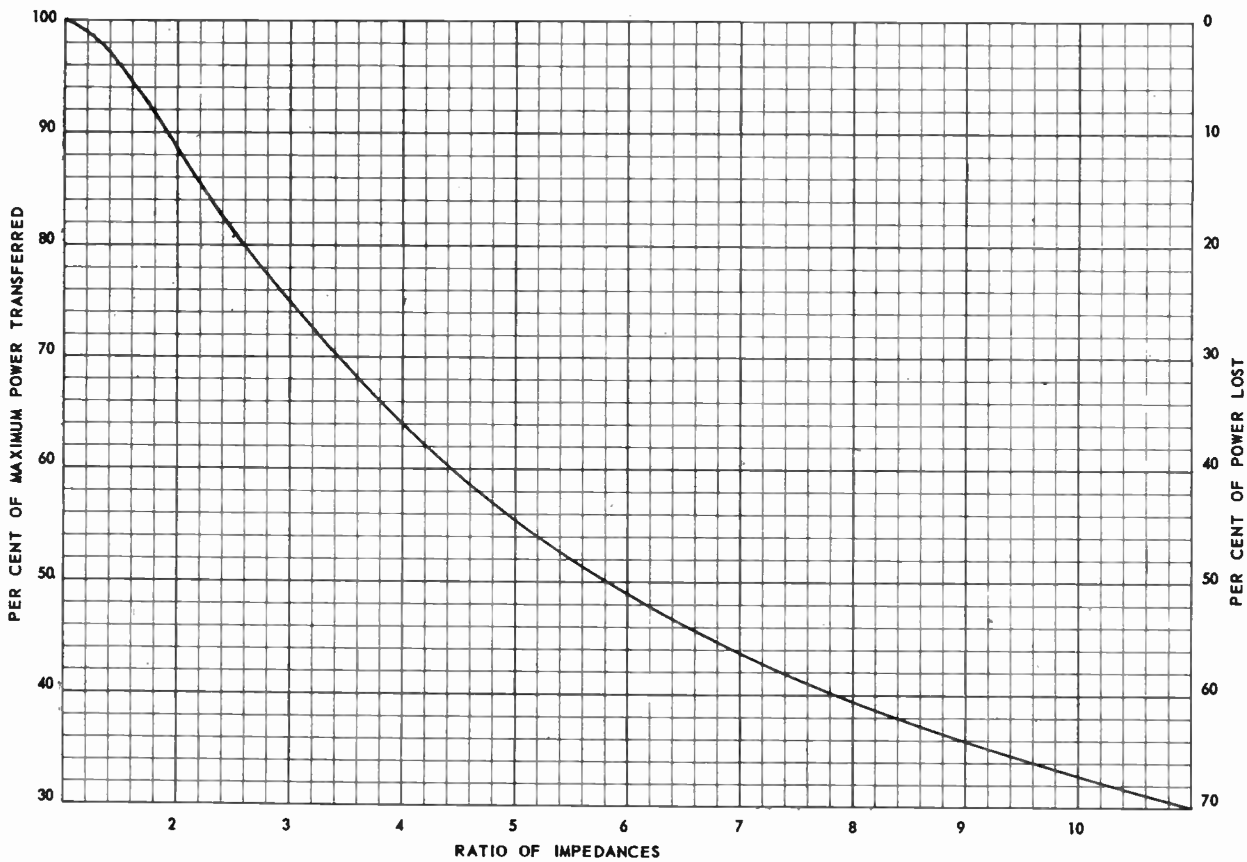


Fig. 66-10. Loss of signal power due to mismatching of impedances.

EFFECTS OF ANTENNA FREQUENCY RESPONSE. In talking about signal power loss that results from mismatch of antenna and line impedances we have assumed that there is no loss at any frequency so long as the two impedances are equal. It is perfectly true that there would be no loss due to mismatching if the impedances could remain equal. But they will not remain equal because antenna impedance varies with received frequency. Although we may make a perfect impedance match at the frequency for which the antenna is cut or is resonant, the match will not remain exact at other received frequencies.

To illustrate what happens, assume that we are working with a simple half-wave dipole which is cut for a frequency of 69 mc, and whose impedance varies as shown by curve 1 of Fig. 66-11. At the center frequency of 69 mc the antenna impedance is approximately 72 ohms. At 50 mc and again at 90 mc the impedance becomes more than 300 ohms. We shall connect to this antenna a transmission line whose characteristic impedance is 72 ohms. The match is perfect at 69 mc, but becomes progressively worse as frequency goes higher or lower.

You can figure out the mismatch ratio by dividing the constant line impedance of 72 ohms into the antenna impedance at any frequency. Then, from Fig. 66-10, you may determine the loss due to mismatch at each frequency. When the losses are plotted against frequency on a graph the curve will appear as at 2 in Fig. 66-11. This looks very good, for apparently the signal into the line will be more than 72 per cent of maximum all the way through channels 2, 3, 4, and 5, and will be 90 per cent or better for channels 3 and 4.

In arriving at these small losses we are forgetting something of great importance; as the antenna impedance goes up its signal pickup ability goes down. This drop in antenna response occurs regardless of impedance matching, and would occur even were no transmission line connected to the antenna.

The reduction of signal pickup by the antenna causes a big reduction of signal power put into the transmission line. Curve 3 of Fig. 66-11 shows power put into a transmission line which is perfectly matched for all frequencies, with line impedance changed to match the varying antenna impedance at every frequency - something which would be exceedingly difficult to do in practice. Power is shown as percentages of that going into the line at 69 mc, the center frequency of the antenna. Now, even within the 6-mc frequency range for a single channel, power to the line is down to about 87 per cent of minimum.

Curve 4 shows percentages of maximum signal power going into the transmission line at various frequencies when the line impedance is not varied, when it matches the antenna impedance at the center frequency only. There is not much difference between curves 3 and 4 until we get so far from the resonant frequency of the antenna that power into the line is too small to be of much use in any case.

If you make a fairly close impedance match between line and antenna at the center frequency of a narrow-band antenna, like the one being discussed, the decrease of signal power into the transmission line and receiver will be almost entirely the fault of the antenna pickup. If you start out with a poor impedance match, the peaks of all our power curves will drop proportionately to the mismatch loss, and reception will be poor in all channels.

SIGNAL REFLECTIONS. If the receiver end of the transmission line were to be connected across a pure resistance of the same value in ohms as the line impedance, this resistance would act just like

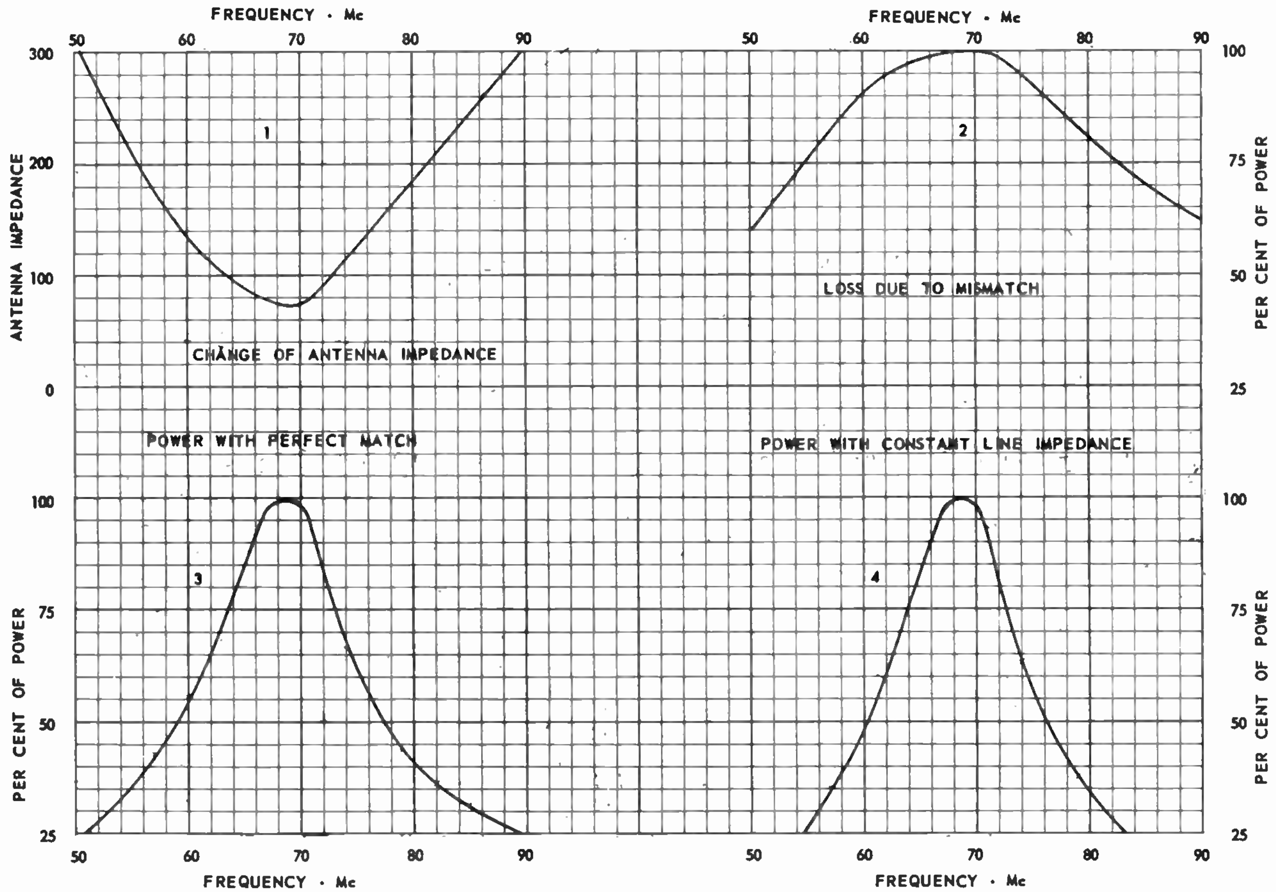


Fig. 66-11. How variations of antenna impedance which occur with change of frequency affect signal power going into the transmission line.

additional length of the transmission line. Signal power put into the line from the antenna would pass into the pure resistance just as it would pass from one part of the transmission line to a following part of the line. All the power delivered by the line would be dissipated in the resistance.

The transmission line does end in what amounts to a pure resistance if the input circuit of the tuner is made resonant at or near the middle frequency of each received channel, and if the frequency response of the tuner is practically flat throughout the range of frequencies for the one channel. Inductive and capacitive reactances of the input circuit then cancel out to leave only resistance. There is a good approximation of pure resistance when the input circuit consists chiefly of non-inductive resistors having suitable values and when there is not too much stray capacitance in wiring connections.

Unfortunately, a receiver input circuit may contain excess inductance instead of being resonant or purely resistive at received frequencies. Yon know that inductance itself does not absorb power. The power goes to build up a magnetic field, then the field collapses andreturns its power to the line. On the other hand, the input circuit may contain excess capacitance. Part of the signal power from the line then builds up electrostatic fields in the dielectric as the capacitance is charged. Then the capacitance discharges and this part of the power is returned to the transmission line.

Under such conditions there is what we call signal reflection from the receiver input back into the line. Signal reflection might be represented as in Fig. 66-12. Excess inductance or capacitance which causes signal reflection will possess inductive or capacitive reactance which varies with frequency. Changes of reactance will cause corresponding variations of receiver input impedance with change of frequency. This makes it impossible to maintain an impedauce match between transmission line and receiver, and there is excessive reduction of signal power retained in the receiver. In effect, the receiver is refusing to accept signal power brought to it by the transmission line, and is throwing part of this power right back into the line.

Reflections due to excess inductance or capacitance occur at the receiver end of the transmission line, not at the antenna end. The greater is the unbalanced inductive or capacitive reactance of the input circuit the less becomes the resistive portion of the impedance in comparison. This increases the proportion of power reflected, and decreases the proportion useful in the receiver.

There will be reflection and reflection loss also at any intermediate point on the transmission line where there is a change of line inductance, capacitance, or resistance, and consequently a change of line impedance. This will happen at a poorly made joint, or where capacitance is altered by spreading the conductors farther apart or by squeezing them together, or where there is a change of insulating material and dielectric constant.

If impedances are well matched between transmission line and receiver, and if the receiver input has no excess of inductive or capacitive reactance, the only effect of changing the length of transmission line is to change the amount of attenuation. Under such conditions a change of line length will not alter impedance matching nor will it cause reflections. We say that the line is flat, or is non-resonant, or is untuned.

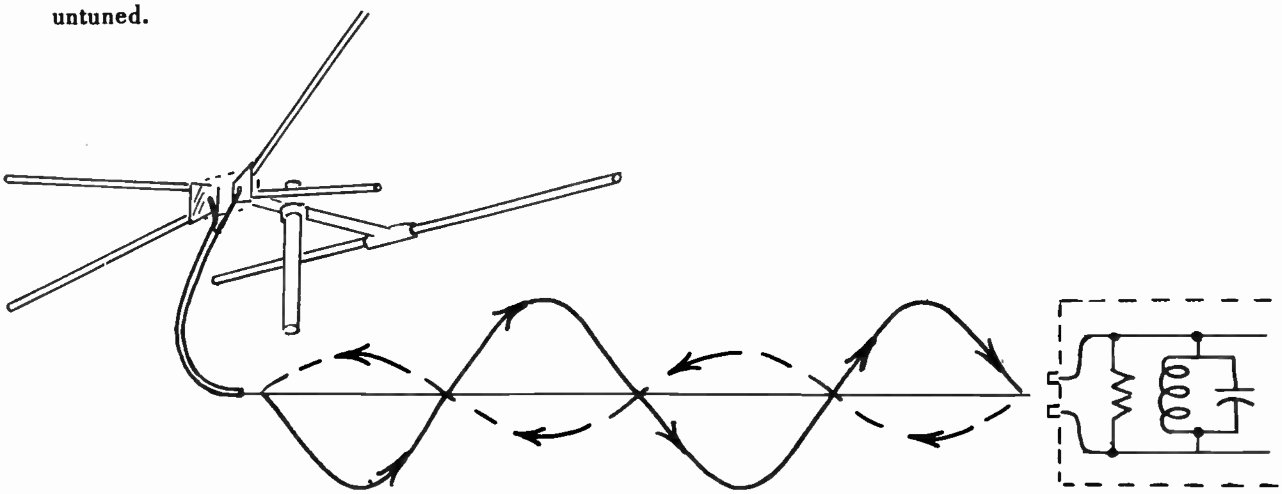


Fig. 66-12. Signal current and power may be reflected back into the transmission line by the receiver input circuit.

If the receiver input circuit is such as to cause reflection, the line length would have to be some exact multiple of a half-wavelength at the received frequency to prevent reflection losses at the receiver input. Since wavelength varies with every change of received frequency it would be nearly impossible to satisfy this requirement when more than one frequency is to be received.

SELECTING A TRANSMISSION LINE. The type of transmission line to be used for a given installation depends on the factors represented by Fig. 66-13 and others. If possible, the line should have characteristic impedance closely matching the impedances of both antenna and receiver, although, as we shall see later, unequal impedances may be joined in such a way as to cause very little loss, by using simple matching transformers.

In selecting a type of line it is necessary to consider also how much signal power may be wasted in attentuation and mismatch while still having acceptable reception. This depends on signal field strength at the point of reception, or on the distance between receiver and transmitters. If there is high field strength, much may be wasted and still leave enough for reception.

If the transmission line must pass through an area where there are strong interference fields or likelihood of noise pickup, we must use a type of line that does not readily pick up such effects. Finally, there is the matter of cost. Some types of transmission line cost up to ten times as much as others.

Unshielded twin conductor line is regularly available with characteristic impedance of 75, 150, or 300 ohms. Shielded twin-conductor line is most commonly used in the 300-ohm type. Coaxial line is regularly available with characteristic impedances of 52, 72, or 95 ohms.

Twin-conductor line, shielded or unshielded, of 300-ohm impedance is used for receivers having 300-ohm balanced input circuits, and where antenna center impedance is 300 ohms or of approximately this value. Twin-conductor or coaxial line of 75 or 72 ohms impedance is used with straight half-wave dipole antennas whose center impedance is 72 or 73 ohms. Most receivers are designed with 300-ohm balanced input. Some are designed with a 75-ohm unbalanced input circuit. Still others have separate input terminals leading into either 75-ohm unbalanced or 300-ohm balanced circuits. Lines with impedances such as 52, 95, 100, 150, and 200 ohms ordinarily are used only for impedance matching transformers or for some experimental purpose.

Possible noise pickup presents a serious problem in many places. This possibility exists where there is electrical machinery of any kind having sparking contacts, as found in d-c motors, sign flashers, and many other devices. Noise interference is prevalent in all factory and commercial buildings, in most apartment houses, and in tall buildings generally. The longer is the transmission line the greater is the chance that noise will be picked up.

In theory, the balanced unshielded type of transmission line connected to a balanced receiver input allows cancellation of interference voltages, and this works out when interference fields are horizontally polarized. But most radio signals and also the fields of spark-type interference, are vertically polarized, which allows vertical sections of transmission line to act as fair antennas for the interference.

A long unshielded transmission line is likely to pick up radio interference at all frequencies where signal field strength is high, and it is likely to pick up wave reflections that cause ghost images where there are strong reflections from nearby objects.

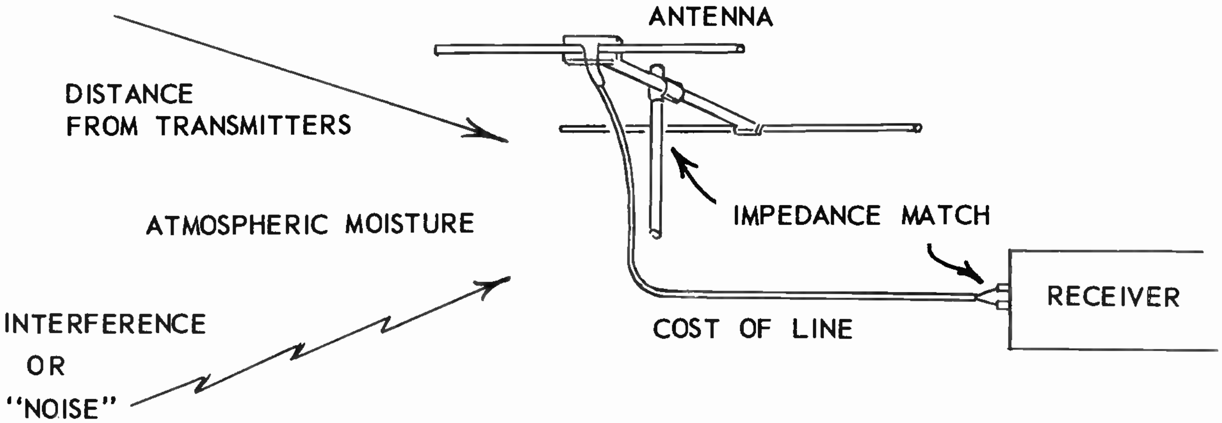


Fig. 66-13. Some of the things which influence the choice of kind of transmission line to be used.

In any of the cases just mentioned you should provide for adequate protection against unwanted pickup by the transmission line. Most often the protection is provided by using a coaxial line or else a shielded twin-conductor line. Reduction of interference may be had also with a type of unshielded transmission line having three side by side conductors. The two outside conductors are connected to carry the signals as with a twin-conductor line. The middle conductor is securely grounded to the receiver chassis, and tends to drain off interference voltages.

Of course, you must remember that interference or noise picked up by the antenna cannot be removed by any kind of transmission line. Then the extra cost for a shielded line would be wasted.

Shielded transmission line, with its practically weatherproof outer covering, may be required where there is excessive moisture in the atmosphere. This is especially true when an installation is near the ocean where salt deposits could damage the insulation of an unshielded line.

With unshielded twin-conductor line having a thin web of plastic insulation between the conductors, part of the dielectric is the insulation and the remainder is air. Moisture which collects and remains on such a line will alter the capacitance and impedance, and may cause temporary mismatch. If there is a good match originally, moisture will cause relatively little trouble, but it makes a poor match worse. There is less trouble from moisture on the narrow 75-ohm and 150-ohm ribbon lines than on the wider 300-ohm type, with which more of the dielectric is air.

There are two possible objections to the use of shielded transmission line. In the coaxial lines of smaller diameter the cost per foot is two to three times that of unshielded flat ribbon line, and the attenuation is greater than with unshielded twin-conductor of the same impedance. The larger diameter coaxial line has little if any greater attenuation than unshielded ribbon line, but it costs six to eight times as much per foot. However, if signals are weak and interference is strong, the high-cost shielded line may be the only way of obtaining good reception.

Transmission line of the twisted or flat pair type with high-quality or "low-loss" rubber insulation sometimes is used for very short runs where low cost is important and signal field strength is high. The attenuation of such lines is about 7 decibels per 100 feet at 70 megacycles and about 20 decibels at 195 megacycles.

TELEVISION

LESSON NO. 67

TYPES OF TELEVISION ANTENNAS

"Training Your Mind"

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind...

C.T.I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON NO. 67

TYPES OF TELEVISION ANTENNAS

Fig. 67-1 illustrates a type of antenna called a folded half-wave dipole or simply a folded dipole. It is shown at the left without a reflector and at the right with one. A reflector for a folded dipole most often is a single straight conductor, but sometimes is of the same form as the signal pickup element, another folded dipole.

A folded dipole most often consists of a single continuous piece of tubing bent as shown, with the ends brought back toward the center where they are left separated and insulated from each other to take the transmission line connections. This type of antenna is mounted with its two sides in the same vertical plane, one above the other. Whether the gap is above or below the continuous side makes no difference electrically, it is a matter of convenience in mounting and in making transmission line connections. At the center of the continuous side of this antenna, or at a point opposite the gap, the signal voltage is zero, This point then may be grounded without affecting reception, and as a consequence the antenna may be supported by attaching this center point directly to the metal mast or a metal cross arm.

5 Center to center separation between the top and bottom of the folded dipole usually is little if any greater than 2 inches in an element cut for high-band reception, and is between 2½ and 3½ inches in one cut for the low band. The overall length of a folded dipole antenna is made somewhat less than the overall length of a straight dipole which is resonant at the same frequency. This is because the end bends or connections between top and bottom add to the electrical length of the two sides. Suitable overall lengths for folded dipoles are shown by Fig. 67-2. This graph is used in the same manner as the earlier one applying to straight half-wave dipoles.

The directional pattern of the folded half-wave dipole is like that of the straight half-wave dipole, and addition of a reflector alters the pattern in a similar way. Signal pickup ability is somewhat greater than that of a straight dipole, because the folded type intercepts a little more of the wave front.

a An advantage of the folded dipole over the straight type is broader frequency response or greater bandwidth. Under conditions of signal strength with which a simple straight dipole would provide satisfactory reception in two adjacent channels, but not in three, a folded dipole ordinarily would be satisfactory throughout the entire low band or the entire high band. The broader frequency response is due to fairly constant impedance over a considerable range of frequencies.

Another advantage of the folded dipole is that its center impedance or minimum impedance at resonance

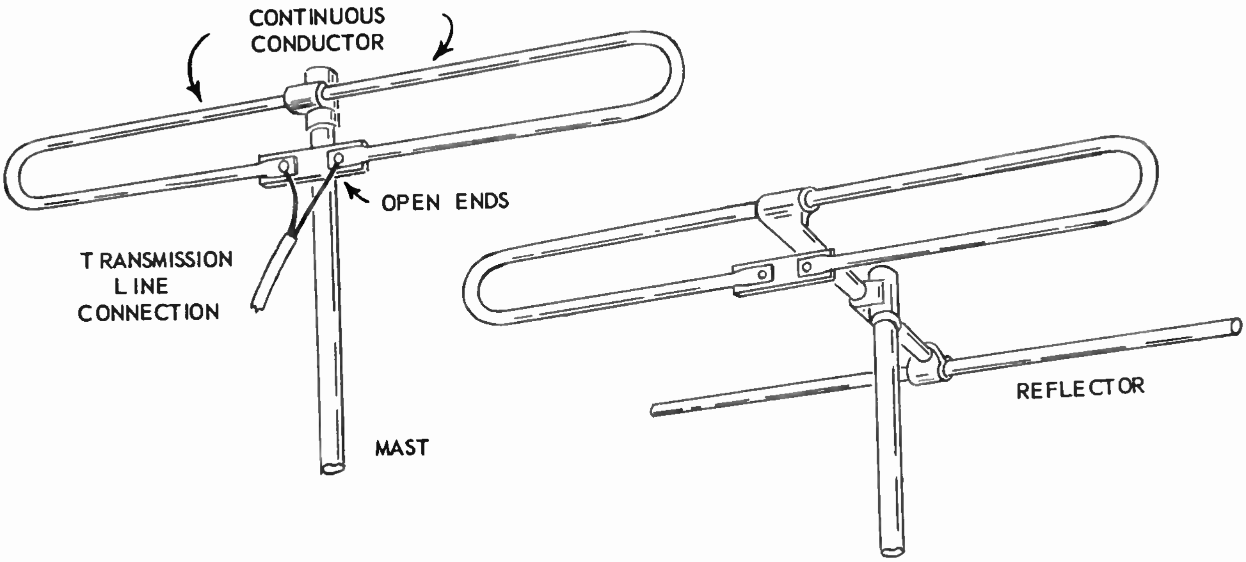


Fig. 67-1. Folded dipole antennas.

is approximately 300 ohms rather than the 72 to 73 ohms of the straight dipole. This is an advantage because the input impedance of he majority of receivers is 300 ohms, and there is good match by using a 300-ohm transmission line between the folded dipole and receiver.

A third advantage of the folded dipole is the possibility of increasing its center impedance to practically any value in excess of 300 ohms by changing the relative diameters of the two sides or by adding one or more additional conduct rs between the outer ends. This is an advantage because the addition of a reflector and of other "parasitic' elements to be described later causes a very considerable drop of impedance. Then, by alterin the design of the folded element, its own impedance may be increased so that in combination with one or more parasitic elements the combined impedance may be made to match that of standard transmissio lines and receivers.

Although the folded dipole antenna usually consists of only two parallel conductors of equal diameter, as in Fig 67-1, it may be mad with three or more conductors. A folded dipole element consisting of three conductors held together at their outer ends by metal clamps is shown at the left in Fig. 67-3. Instead of clamps, the end sup orts might be long screws with spacer sleeves between adjacent conductors, or of any other form which provides electrically conductive end connections. The divided conductor may be in the center.

The center impedance of a folded dipole is equal to 72 ohms multiplied by the square of the number of conductors. With the usual two conductors we multiply 72 by 4 (the square of 2) to find that the impedance is 288 ohms, or approximately 300 ohms. If there are three conductors we multiply 72 by the square

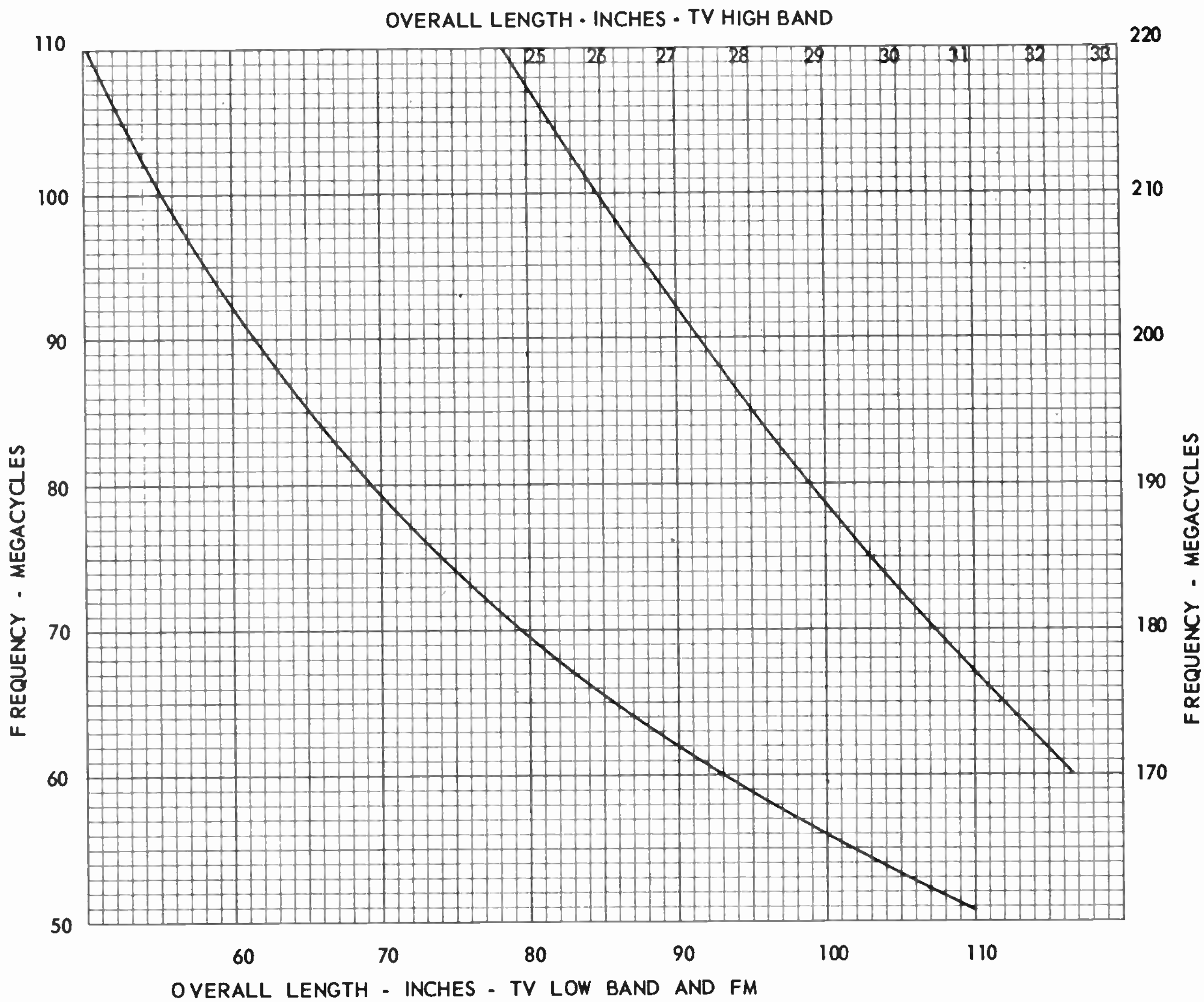


Fig. 67-2. Lengths of folded dipole antennas for resonance at various frequencies.

of 3, or 9, to find that the center impedance is 648 ohms, or approximately 600 ohms. This formula for impedance holds true when all the conductors are of the same diameter.

At the right in Fig 67-3 is illustrated one method of constructing a folded dipole with two conductors of different diameters. The outer ends are electrically connected together by screws passing through spacer sleeves. Connections for the transmission line are provided at the gap between inner ends of the smaller conductor. These inner ends are shown supported from the larger conductor by pieces of insulation.

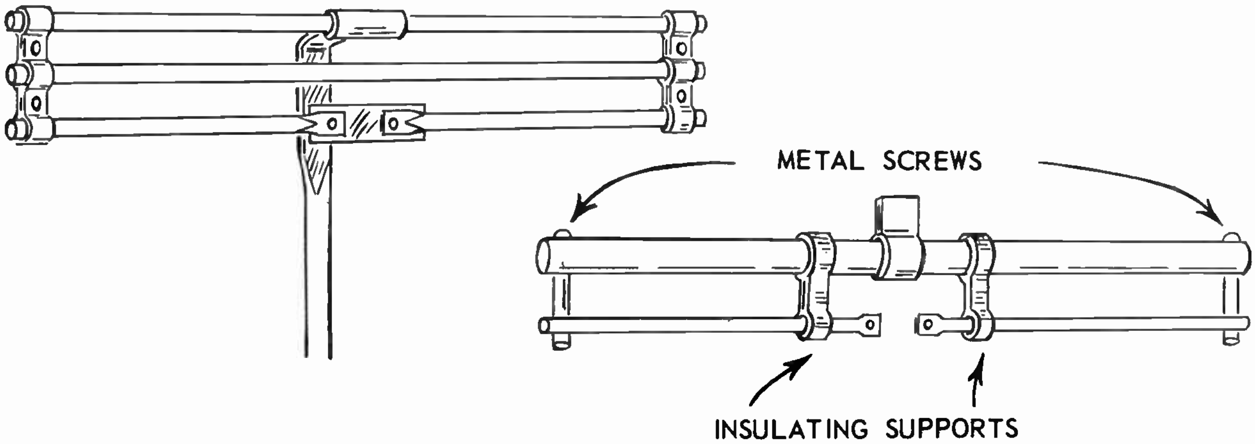


Fig. 67-3. A three-conductor folded dipole (left) and one with conductors of different diameters (right).

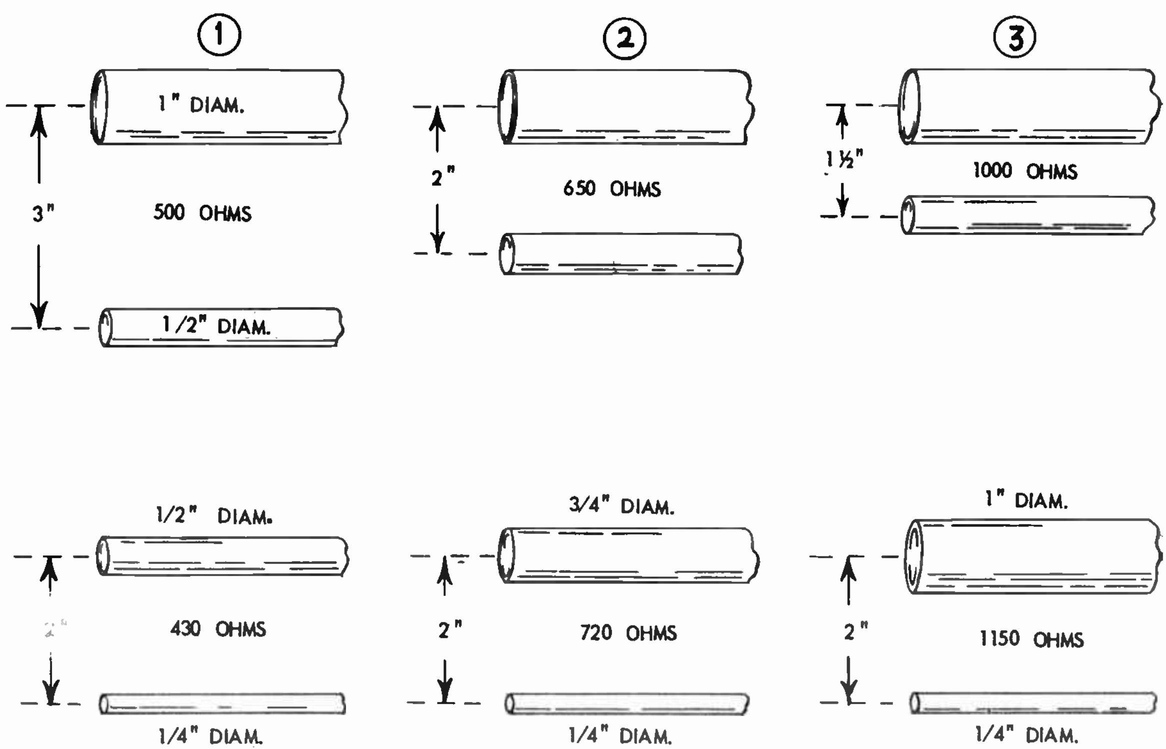


Fig. 67-4. How center impedance is varied by using different spacings and diameters for the sides of folded dipoles.

Fig. 67-4 shows a few examples of what may be accomplished by changing the relative diameters and the center-to-center spacing of two conductors in a folded dipole. At 1 the continuous conductor is 1 inch in outside diameter, the conductor containing the gap is 1/2 inch in outside diameter, and center-to-center separation is 3 inches. The center impedance becomes approximately 500 ohms. At 2 the separation between the same two conductors is reduced to 2 inches, and impedance goes up to about 650 ohms. At 3 the separation is still further reduced to 1½ inches, and the center impedance becomes 1,000 ohms.

The diagrams in the lower part of the figure show results of maintaining a constant center-to-center separation, here 2 inches, and of varying the outside diameter of the continuous conductor while the conductor containing the gap remains of 1/4 inch diameter. The greater the difference between diameters the higher becomes the center impedance.

Still another way of increasing the center impedance of a folded dipole is shown by Fig. 67-5. To the continuous conductor of each folded element is clamped or bolted a straight conductor of the same diameter, and of a length that would be suitable for a straight dipole cut for the same resonant frequency as the folded element.

This increases the center impedance of each folded dipole from approximately 300 ohms to 600 ohms.

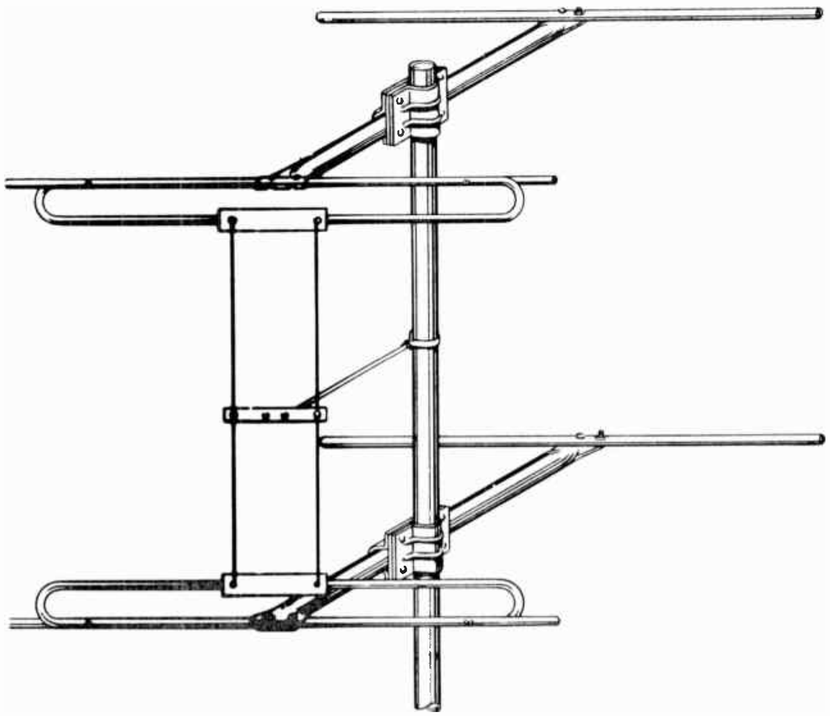


Fig. 67-5. Taco stacked folded dipole with 600-ohm elements.

In this particular design the two folded dipoles, each with 600-ohm center impedance, are connected together by vertical wires or rods which themselves are of such diameter and spacing as to have 600 ohms impedance for matching the dipoles. At the midpoint along the vertical connections we have two 600-ohm impedances in parallel for a parallel impedance of 300 ohms. This provides a match for a 300-ohm transmission line connected at this point.

HIGH-LOW ANTENNAS. A single folded or straight dipole antenna is suitable for reception from any number of transmitters in either the high band or low band provided all the transmitters are located in the same general direction from the receiver or are located within such an angle on the directional pattern that all carriers are picked up with necessary signal strength.

If low-band and high-band transmitters are in decidedly different directions we may use a so-called high-low combination of two antennas such as illustrated by Fig. 67-6. A folded dipole and reflector cut for the high band are mounted on the same mast with another pair cut for the low band. The fittings allow each antenna to be turned or oriented independently of the other, so that each may best receive from transmitters in its own band.

The two antennas may be of any types. Both may be folded dipoles, or both may be straight dipoles, or both may be of any other one type. The high-band antenna may be of one style and the low-band unit

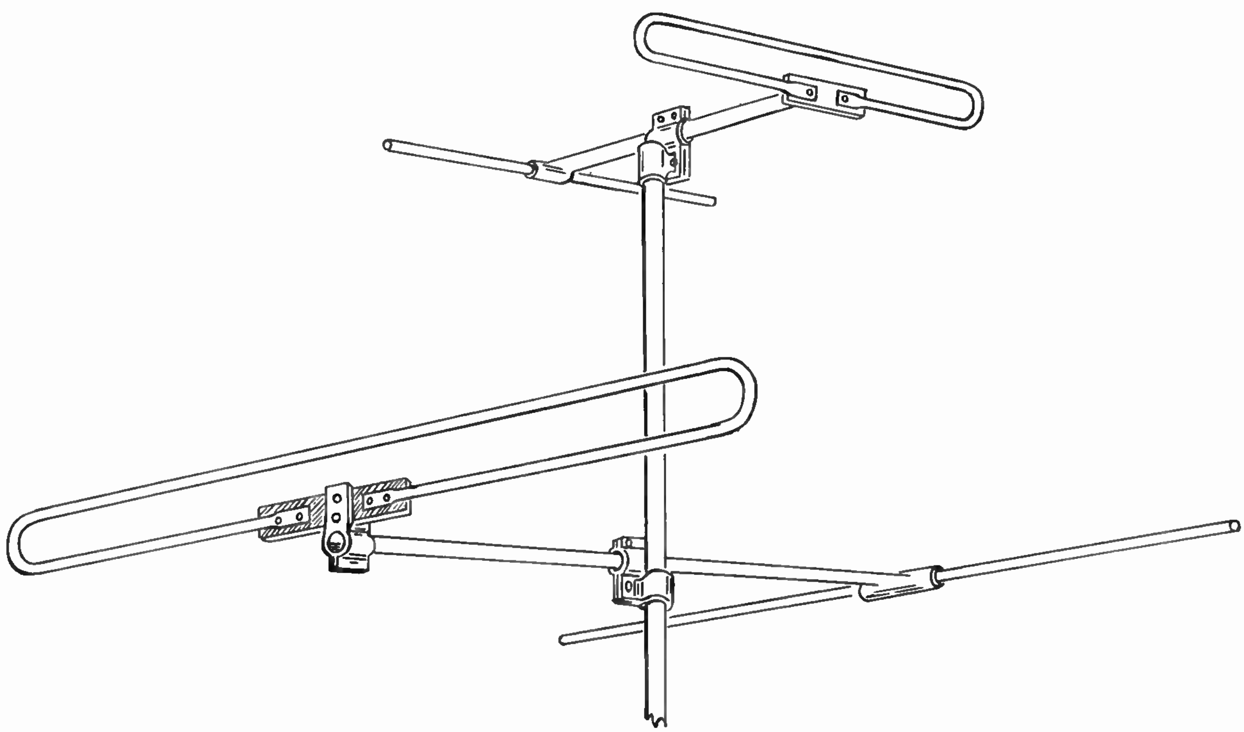


Fig. 67-6. A high-low antenna with folded dipoles.

of a different style. The high-band antenna may be either above or below the one for the low band. Either or both of the antennas may have reflectors or other parasitic elements.

The ideal way of using two antennas, one cut for each band, is to run a separate transmission line from each antenna to a changeover switch located at the receiver. Such a switch is built into the tuners of some receivers, so that there is automat c change of antennas in going from channel 6 to channel 7 or vice versa. For connecting the two antennas to a changeover switch there is a special type of four-conductor flat ribbon transmission line. Two conductors are connected to one antenna and the remaining two to the other antenna.

More often the high-band and low-band antennas are connected through a single two-conductor transmission line to the tuner input terminals of the receiver. In this case there arises a special problem, because the low-band antenna has some pickup for signals in the high band, and its connections can take away signal power picked up by the high-band antenna. This may be prevented by using suitable "phasing" connections at the antennas, between the antennas and the transmission line, and by using suitable vertical spacing between the two antennas. Instructions for correct installation are furnished with manufactured high-low antennas.

Fig. 67-7 shows some phasing connections wh'ch may be made with pieces of 300-ohm unsh elded transmission line when both antennas are folded dipoles. At 1 the transmission line going to the receiver is connected to the high-band antenna. From the high-band to the low-band antenna is connected an additional piece of transmission line approximately six feet long, the correct length being that which allows best reception as determined by trial.

Because the two antennas usually are mounted with vertical center-to-center separation of about 30 inches there will be some slack in the line between them. The slack is taken up by supporting this line from the mast or a cross arm by me ns of one or more standoff insulators.

To the center terminals of the low-band antenna are connected also the two conductors at one end of still another piece of transmission line about 12½ inches long. At the free end of this short "stub' its two conductors are left open, not connected together or to anything else. Inductance and capacitance of the stub act like a se ies resonant circuit tuned to the center of the high-band frequency range. This practically short circuits whatever power at these frequencies is picked up by the low-band antenna. Otherwise th's signal power could go to the transmission line in an out-of-phase relationship with power picked up by the high-band antenna.

In diagram 2 we have vertical center-to-center separation of 30 inches. A 12½-inch piece of transmission line is connected to the terminals of the high-band antenna. The regular transmission line which runs to the receiver is connected to the low-band antenna, and to this antenna is connected also an open stub as previously described. Now the free end of the short piece of line connected to the high-band antenna is temporarily connected at various places along the regular transmission line for the receiver, starting about 45 inches from the connection to the low-band antenna and working toward this antenna until best reception is obtained in high-band channels. The temporary connection may be made with needle points attached to the conductors of the short piece of line and pierced through the insulation on the regular transmission line. A permanent soldered connection is made when the best position is located.

COMMERCIAL TRADES INSTITUTE

TELEVISION

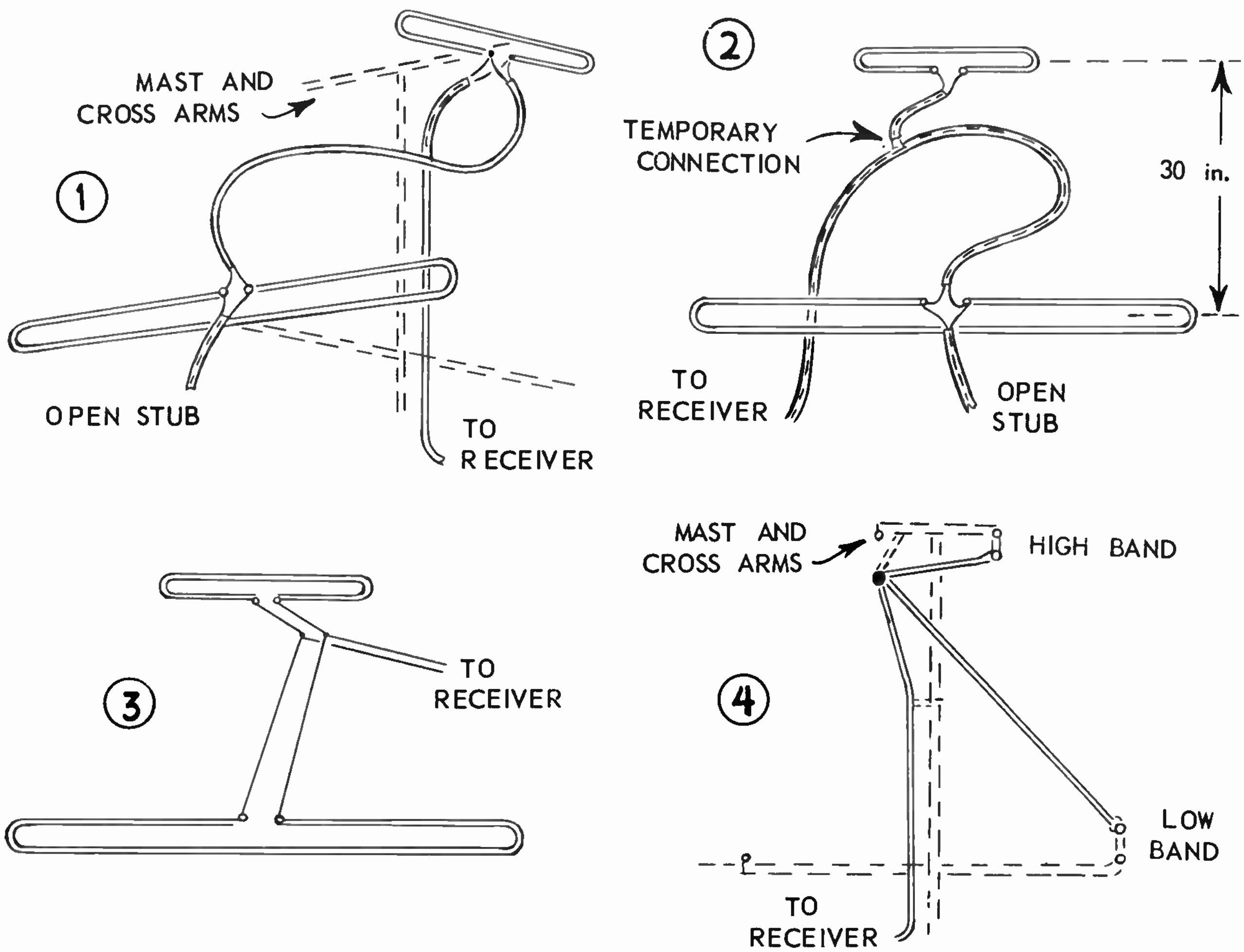


Fig. 67-7. Phasing connections for high-low antennas.

Sometimes it is easier to get the connections correctly phased by connecting the receiver transmission line to the terminals of the high-band antenna to begin with. Then a piece of line about 38 or 39 inches long is connected to the low-band antenna. The free end of this piece is temporarily connected at places somewhat more and less than 12 inches from the high-band antenna, along the transmission line connected to this antenna.

When the total length of phasing connection between center terminals of the two antennas is much greater than the vertical separation between antennas, as in diagrams 3 and 4 of Fig. 67-7, you can take up the slack by bringing the connecting line to the upper cross arm and holding it there with a standoff insulator. This also provides support for the end of the transmission line that goes to the receiver.

If both antennas are straight dipoles, with 72 to 73 ohms center impedance, use 75-ohm unshielded transmission line for all connections and runs. Then make the open stub on the low band antenna about 10½ inches long, and use this length also for the short connection on the high-band antenna in diagram 2. Separation between the two antennas remains 30 inches.

Should you have trouble in making correct phasing connections, leave these connections off both antennas while running the receiver transmission line first to the high-band unit and then to the low-band unit. Then check reception from stations in both bands. If reception is good in both cases, the original fault is in your phasing connections. If reception still is poor, the trouble is not in the phasing connections, but still could be in the antenna connections or in the orientation of the antennas.

DIRECTOR ELEMENTS. The term "parasitic element" has been used quite often while talking about antennas. Earlier we learned about the most commonly used parasitic element, a reflector, and now we shall learn about the other kind, which is called a director. As shown by Fig. 67-8, a director is mounted in front of the antenna, between the antenna and transmitters from which reception is desired, while a reflector is mounted behind the antenna. A director most often is a straight piece of tubing, although it may have the form of a folded dipole when the antenna itself is of this type.

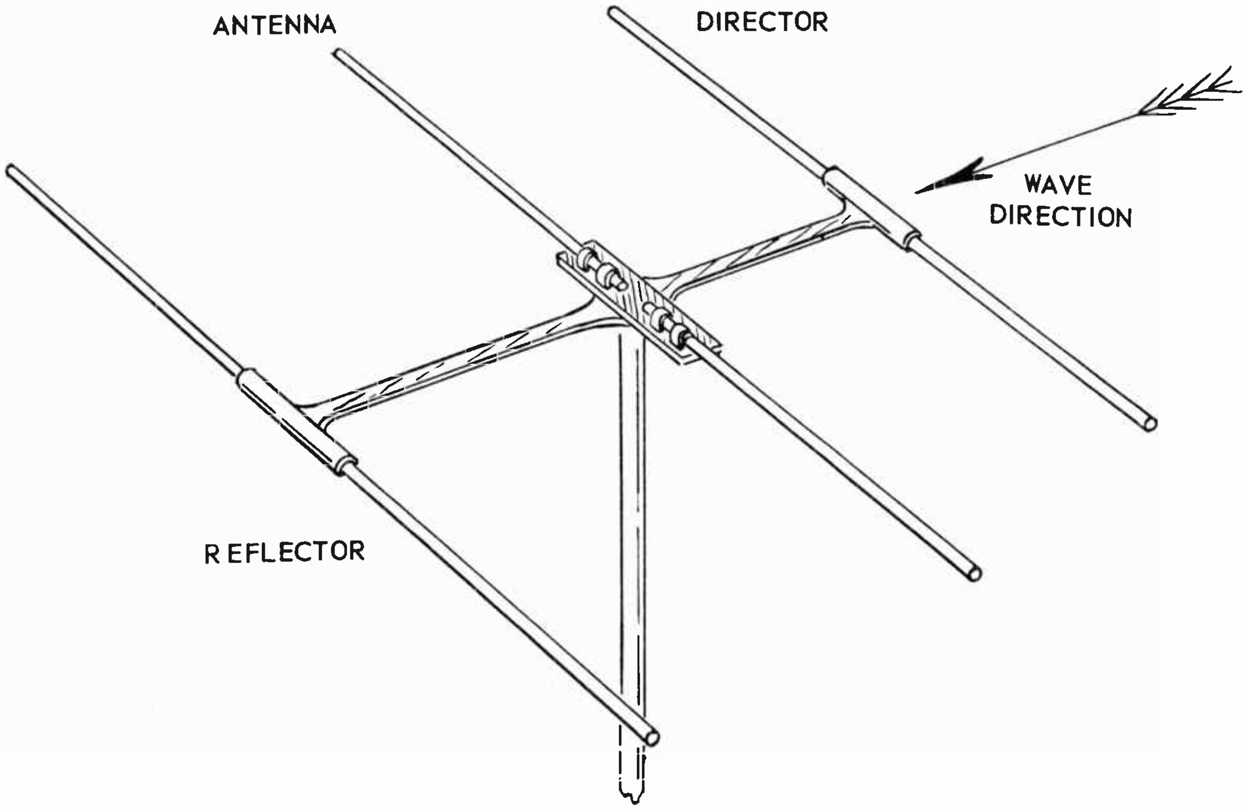


Fig. 67-8. Positions of a director and a reflector in relation to wave travel.

The director depends for its action on reradiation, as does also the reflector. The director is spaced at such a distance in front of the antenna that fields reradiated from the director to the antenna induce antenna emf's which are in phase with those produced by carrier waves arriving from the front. When this spacing is used, carrier waves coming from the rear of the antenna cause director reradiation in such phase as opposes antenna emf's which are due to these waves arriving from the rear. Spacing between director and antenna is measured in fractions of a wavelength corresponding to the frequency for which the antenna is cut.

A director ordinarily is closer to the antenna than is a reflector. It is the difference between spacings of director and reflector, and their lengths in relation to that of the antenna, which cause each of these parasitic elements to perform its desired functions. That is, whether a parasitic element acts as a director or a reflector depends on its spacing from the antenna and on its length. Directors and reflectors both act to increase signal power taken by the antenna from carrier waves coming from one direction, and both these elements decrease pickup from the opposite direction.

It is entirely possible to use a director without a reflector on the same antenna. But were only one parasitic element to be employed, ordinarily there would be greater advantage in using it as a reflector rather than a director. Consequently we find directors commonly used only where there is also a reflector on the same antenna.

Varying the spacing between a director and its antenna affects all the characteristics of the antenna; gain, bandwidth, directional pattern, front-to-back ratio, and characteristic impedance. Every change of spacing alters all these things to a greater or less extent.

A director causes maximum possible gain when this element is the same length as the antenna element and is spaced approximately 0.10 wavelength in front of the antenna. Gain decreases with either more or less spacing. The amount by which gain is decreased may be lessened if the length of the director is altered to suit each change of spacing. Fig. 67-9 illustrates some variations of element spacing and the general manner in which lengths of a director and a reflector are altered in order to have the best gain with each spacing.

When a director is more than 0.10 wavelength in front of the antenna the director must be made somewhat shorter than the antenna to have the best gain attainable with such spacing, but this gain will be less than with 0.10 wavelength spacing and equal element lengths. If the director is less than 0.10 wavelength in front of the antenna the director must be somewhat longer than the antenna to attain the best gain for this spacing, but again the gain will be less than with 0.10 wavelength spacing and equal lengths.

What we really are doing by changing the length of a director, or of a reflector, is to alter the resonant frequency of the parasitic element. With equal lengths the antenna and the parasitic element are resonant at the same frequency. Shortening the parasitic element reduces its inductance and raises its resonant frequency. Less inductance leaves relatively more capacitance. Then induced current and the resulting reradiation field lead the induced emf in the parasitic element, and the effect reaches the antenna in less time.

4 Lengthening the parasitic element increases its inductance and lowers its resonant frequency. The

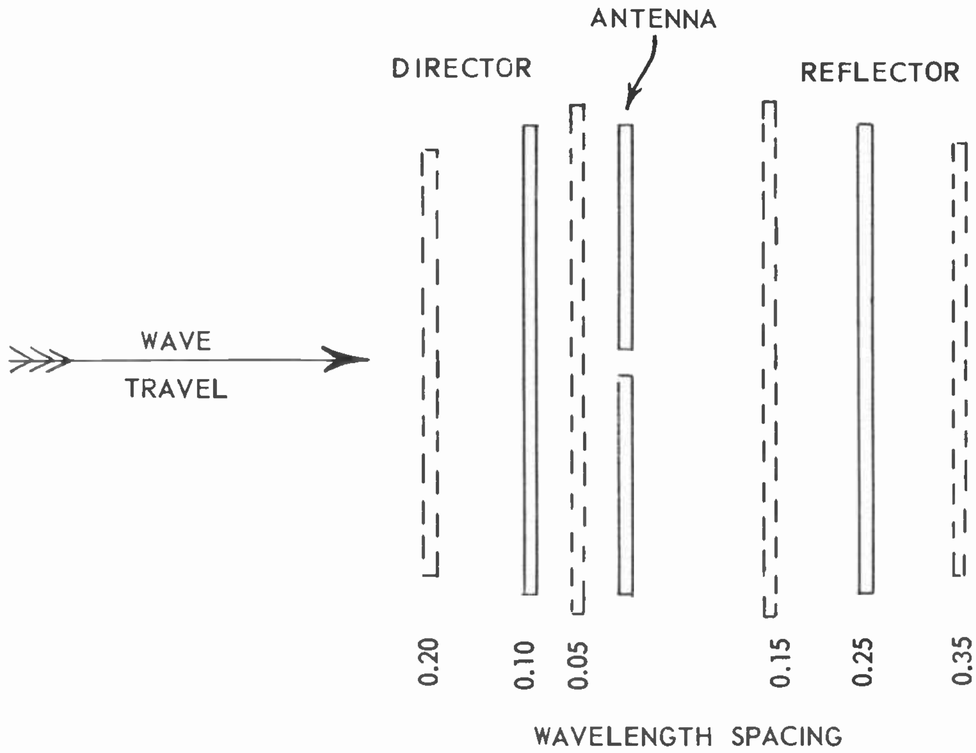


Fig. 67-9. How lengths of director and reflector are altered to suit their spacings when maximum gain is desired.

induced current and the reradiation field then lag the signal emf induced in the parasitic element, and the effect takes longer to reach the antenna. From Fig. 67-9 you may see that reradiation effects are speeded up by shortening the parasitic element when it is spaced farther from the antenna, and the effects are slowed down by lengthening this element when it is moved closer to the antenna. Obviously, the reradiation fields must be advanced or retarded with greater or less spacing if their phase is to remain unchanged in relation to that of current in the antenna.

The combined effect of a director and a reflector is to raise the antenna gain by four or five decibels or even more when spacings and element lengths are suitably adjusted. As you should expect, the increase of gain due to the director is accompanied by a narrowing of the frequency response or by less bandwidth. When a director and reflector are adjusted for high gain the bandwidth will effectively cover only one channel. Even were the response to remain fairly high throughout the range for two adjacent channels this would do us no good, because adjacent channels never are used by transmitters serving the same locality.

A director narrows the directional pattern of the antenna even more than does a reflector. Oftentimes this is the principal reason for using a director. As an example, at the left in Fig. 67-10 a wave reflec-

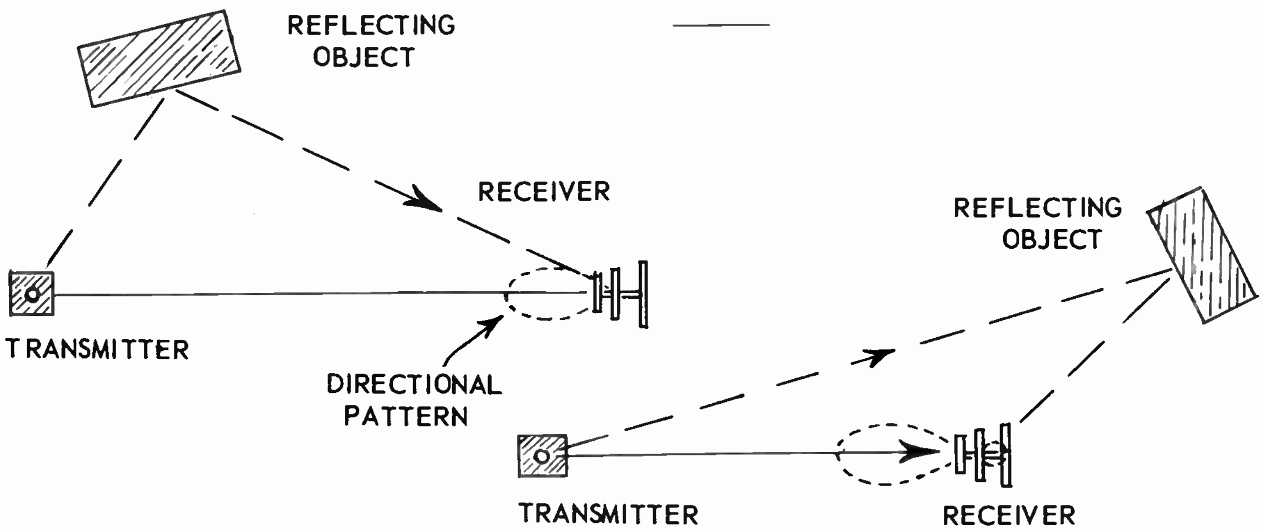


Fig. 67-10. How a director helps to exclude reflected waves and ghost signals.

tion comes to the receiving antenna from almost the same direction as the direct wave. A director can produce a directional pattern so narrow as to practically exclude the wave that would cause ghost images. Undesired radio waves of any kind may be similarly excluded.

A director is capable of causing a great increase of front-to-back ratio, and may be highly effective in reducing or eliminating pickup from carrier waves reflected toward the back of the antenna. This is illustrated at the right in Fig. 67-10.

Front to back ratio is increased by moving the director closer to the antenna. With spacing of about 0.10 wavelength, for maximum gain, the front-to-back ratio may be on the order of 10 to 1. With lesser spacing this ratio becomes rapidly greater, but gain goes down just about as fast. With a spacing of 0.05 wavelength the front-to-back ratio may double, but gain will be down to about two-thirds of maximum even with director length such as gives best gain for this spacing.

As spacing between director and antenna is made greater than 0.10 wavelength there is decrease of both gain and front-to-back ratio. With spacing as great as 0.20 to 0.25 wavelength the gain is down to about two-thirds of maximum and front-to-back ratio drops to about 5 to 1, so far as it is affected by the director. Whether a director is spaced for best gain or greater front-to-back ratio depends on which of these is more to be desired for a particular installation.

The addition of a director reduces the center impedance of the antenna below the impedance of the antenna alone or of the antenna with a reflector. The closer the director is placed to the antenna the greater is the reduction of antenna impedance. If the antenna is of the folded dipole type its center impedance may be raised to compensate for the drop due to a director, a reflector, or both, and there can be a good match for a 300-ohm transmission line. Otherwise it is possible to use transmission line of lower impedance or to use a matching transformer.

THREE-ELEMENT ANTENNAS. The combination of a straight or folded dipole antenna with a straight reflector and a straight director usually is called a three-element antenna or a three-element array. Such an array ordinarily is used for reception in only one channel in localities where signal field strength is low and high gain is essential.

Fig. 67-11 illustrates an array consisting of a folded dipole for the high-band antenna, another folded dipole for the low-band antenna, and a straight reflector for low-band frequencies. The three elements are so positioned in relation to one another that the low-band folded dipole acts also as a reflector for the small high-band antenna in front of it. This combination, often called an "in-line" antenna, is suitable where all transmitters from which reception is desired are in approximately the same direction from the receiver.

The relative spacings and lengths of elements for the in-line antenna are such as to maintain nearly constant impedance of 300 ohms throughout both the low band and high band frequencies, and to provide bandwidth great enough for reception in all the very-high frequency channels. This advantage in bandwidth over a three-element antenna is accompanied by a sacrifice in gain, which is two to three decibels or which averages about two decibels for all channels.

YAGI ANTENNAS. By using either two or three directors and one reflector with a straight or folded dipole antenna element we have what usually is called a yagi array. With two directors, one reflector, and an antenna element the combination is a four-element yagi. With one more director it is a five-element yagi array. A five-element array is illustrated by Fig. 67-12.

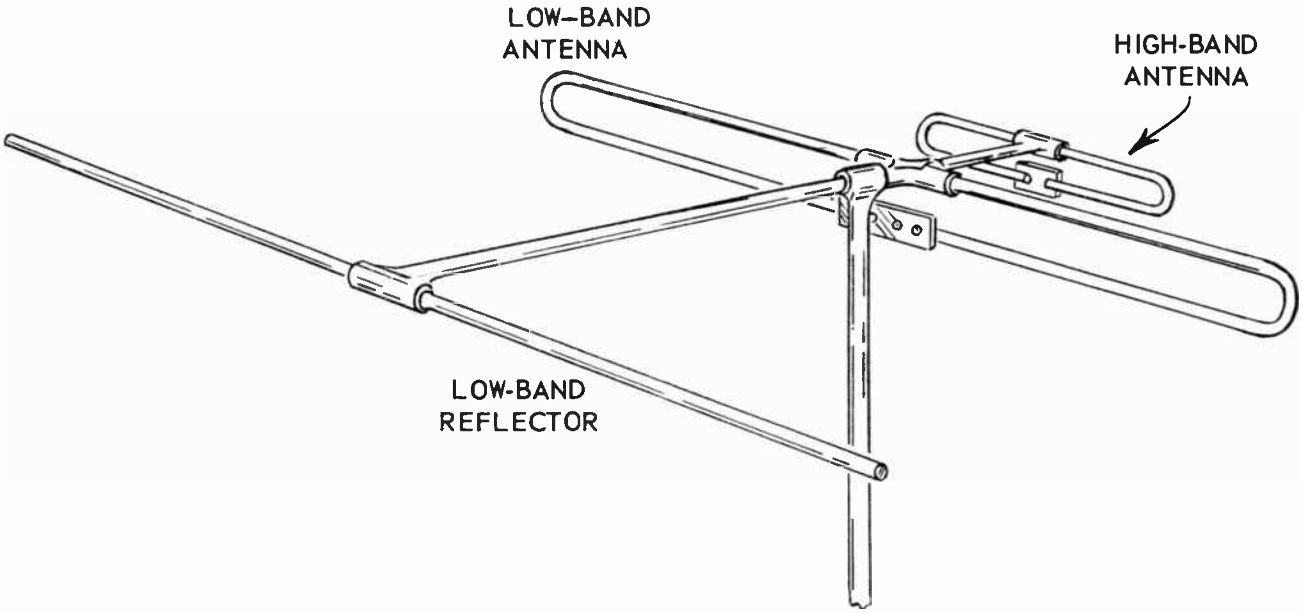


Fig. 67-11. A three-element in-line antenna.

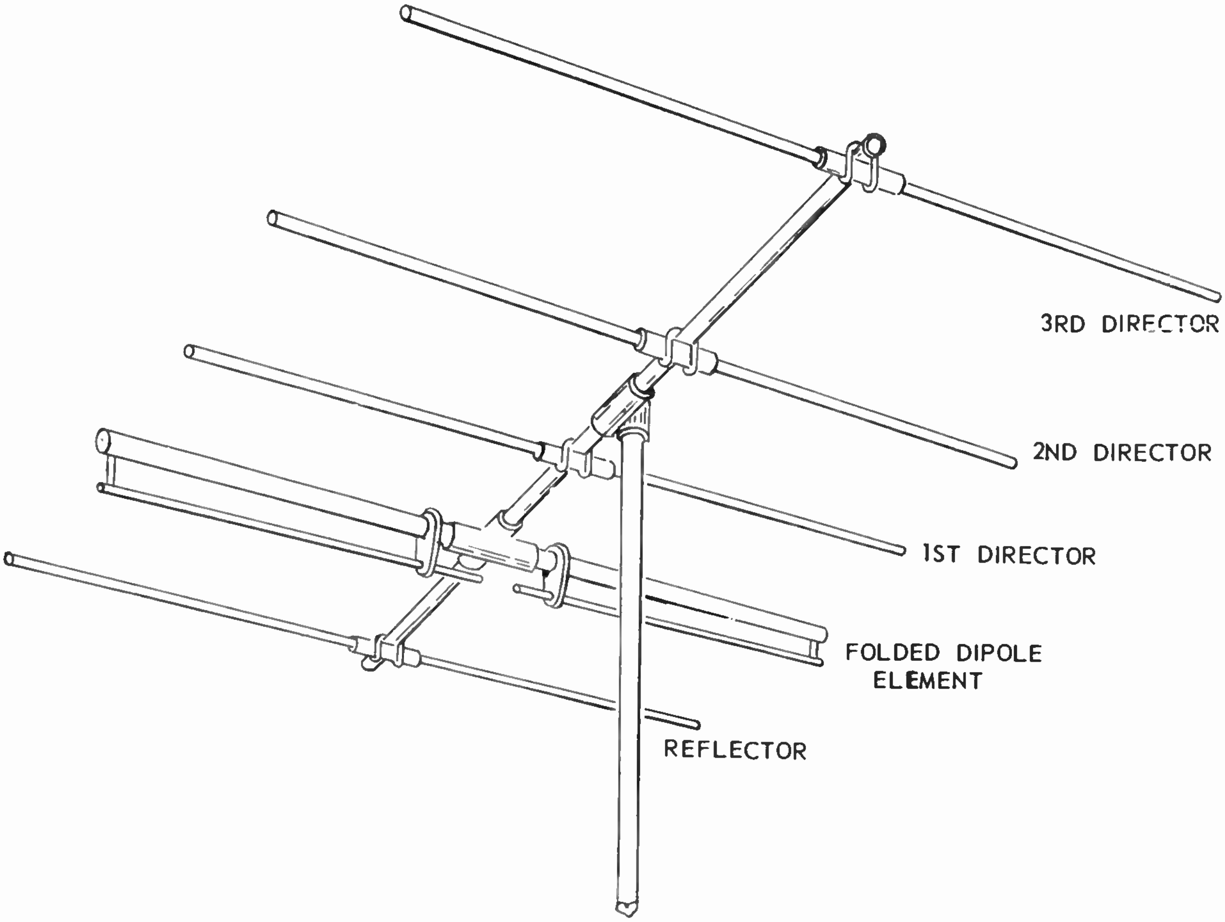


Fig. 67-12. Five-element yagi antenna.

The yagi is strictly a single-channel antenna. A separate array cut to correct length is required for each channel to be received. This type of antenna affords high gain, usually about 10 decibels more than a simple straight dipole, and at the same time provides a high front-to-back ratio. The directional pattern is sharp, with a reception angle of something like 20 degrees within which signal power remains at half or more of the maximum value secured in a straight forward direction.

Phasing of currents in the parasitic elements, and of fields reradiated from these elements to the antenna element, depend on relative element spacings and lengths. In addition to the inductive coupling between antenna and directors there is some coupling and reradiation between the directors themselves. There are so many variable factors that design of a yagi to best suit an individual application usually is arrived at by cut and try methods. Manufactured antennas of this type are carefully engineered for best performance under average conditions where signal field strength is low.

Spacings and relative lengths of elements which may be used for a four-element yagi array are shown at 1 in Fig. 67-13. Wavelength spacings are converted to inch measurements by dividing 11,810 by the frequency in megacycles for which the antenna element is cut, then multiplying by the fractional wavelengths shown. Spacings and relative element lengths for two different five-element yagi arrays are shown at 2 and 3.

When a yagi array is designed for maximum possible gain the frequency response tends to be very narrow. This response or bandwidth is broadened by using a folded dipole as the antenna element. Bandwidth may be increased somewhat more by using dimensions such as shown by diagram 3, where

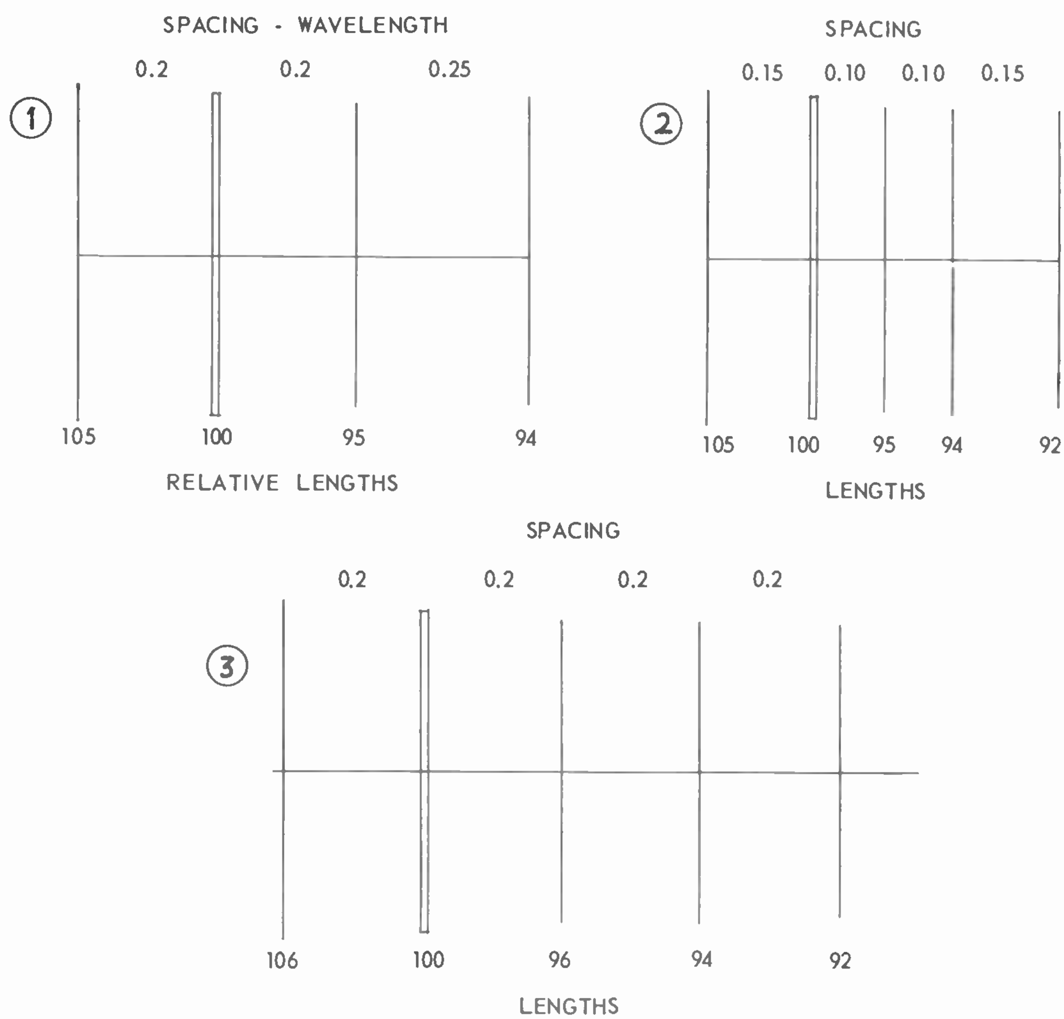


Fig. 67-13. Spacings and lengths of parasitic elements for typical yagi antennas.

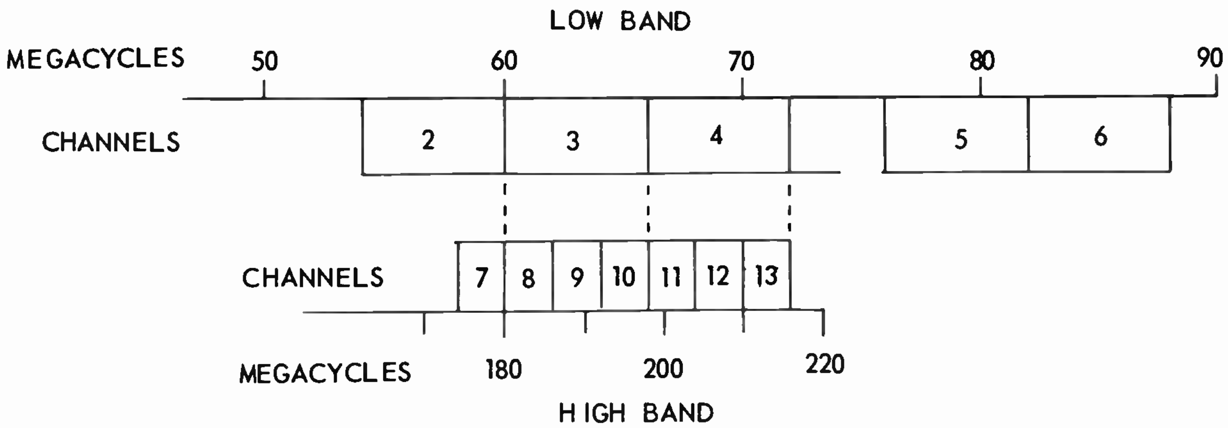


Fig. 67-14. All high-band frequencies are third harmonics of low-band frequencies.

there is increased spacing between elements while at the same time the reflector is made slightly longer and the directors slightly longer than lengths required for maximum gain.

The added parasitic elements tend to make the center impedance of a yagi array too low for matching commonly available transmission lines. This effect is compensated for by making the folded dipole element of large and small tubing, as shown in Fig. 67-12, and by suitably adjusting the spacing between these pieces of tubing. In this manner the center impedance may be brought up to a good match for 150-ohm or 300-ohm transmission lines.

HARMONIC OPERATION OF ANTENNAS. The shortest antenna which is resonant at any given frequency is one cut to a half-wavelength corresponding to that frequency. If you make an antenna any shorter it still may pick up signals at the frequency considered, but signal power will be relatively weak because the antenna is out of tune, it is no longer resonant.

When you cut an antenna to a half-wavelength at a certain frequency the antenna behaves as a series resonant circuit with low impedance as seen by the transmission line. If you increase the frequency of carrier waves reaching this antenna its impedance goes up. At double the original frequency, or at the second harmonic frequency, there is maximum impedance of several thousand ohms. The antenna then is acting as a parallel resonant circuit. It becomes practically useless for signal pickup.

7
Making the applied frequency still higher causes the impedance to drop from its maximum value, and at three times the original frequency the impedance again becomes minimum and there is good signal pickup ability. This "three times" frequency is the third harmonic of the first one. It is an odd harmonic. By experimenting with carrier frequencies you would discover that the antenna will have useful pickup at still higher odd harmonics of the original or fundamental frequency for which the antenna is cut.

If you multiply by three the frequencies used for the low-band television channels, to determine their third harmonics, and then compare these harmonics with frequencies in the high band you will see that every high-band frequency is a third harmonic of some frequency in the low band. Such a comparison is shown by Fig. 67-14. According to this, an antenna whose bandwidth is sufficient for the low band should respond also to all the high-band frequencies. In a few exceptional cases this actually is possible. It is possible if the high-band transmitters happen to be in a direction just about 50 angular degrees removed from the direction of low-band transmitters.

That last statement should be a clue to the fact that the directional pattern of an antenna working at a third harmonic high-band frequency is different from the pattern at the low-band fundamental frequencies. If a directional pattern appears as at the left in Fig. 67-15 for reception at the fundamental frequency for which the antenna is cut, then the pattern for the third harmonic frequency will appear about as shown at the right.

Instead of two wide lobes extending out at right angles to the axis of the antenna we have for the third harmonic frequency four major lobes and two minor ones. The insignificant minor lobes are at right angles to the antenna axis, but the major lobes are at an angle of about 40 degrees to this axis. Directions from which carrier waves must come for best reception are shown by arrows on the patterns. If low-band stations are in directions at right angles to the antenna axis, and high-band stations are in directions about 50 degrees to either side, we may have reception in both bands with the one antenna.

There is some slight change of directional pattern between lowest and highest frequencies in each television band, but the change is too little to affect reception. By adding a reflector we may greatly reduce the lobe or lobes on one side of the antenna while strengthening those on the opposite side, but this will not materially alter the directions for best reception.

CONICAL ANTENNAS. To have the harmonic frequency antenna receive both low-band and high-band

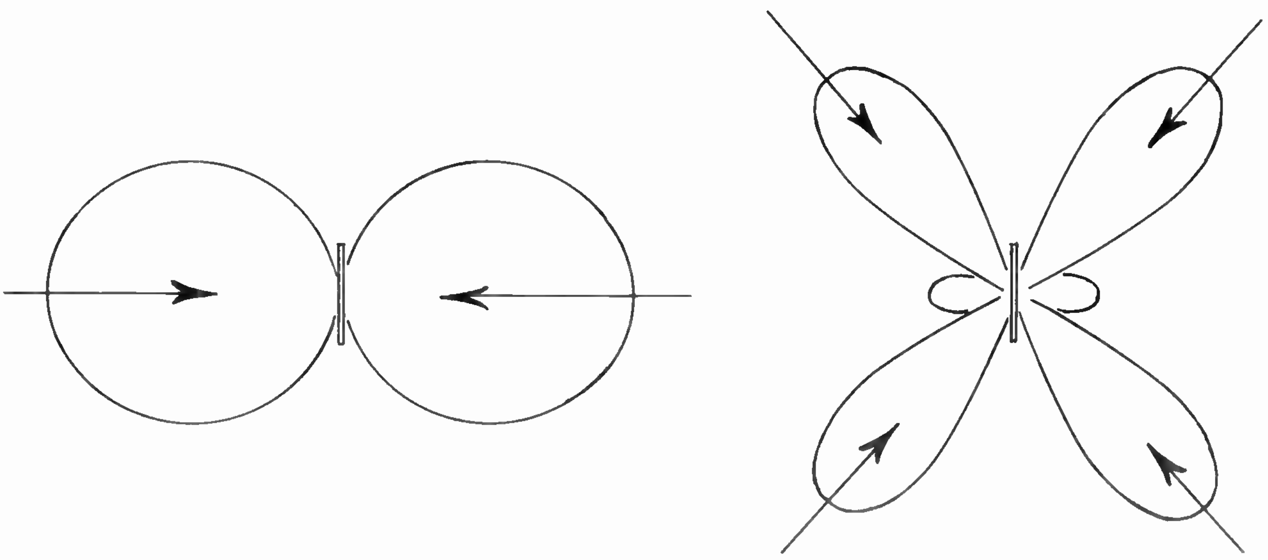


Fig. 67-15. Directional patterns of antenna operating at fundamental frequency (left) and at third harmonic frequency (right).

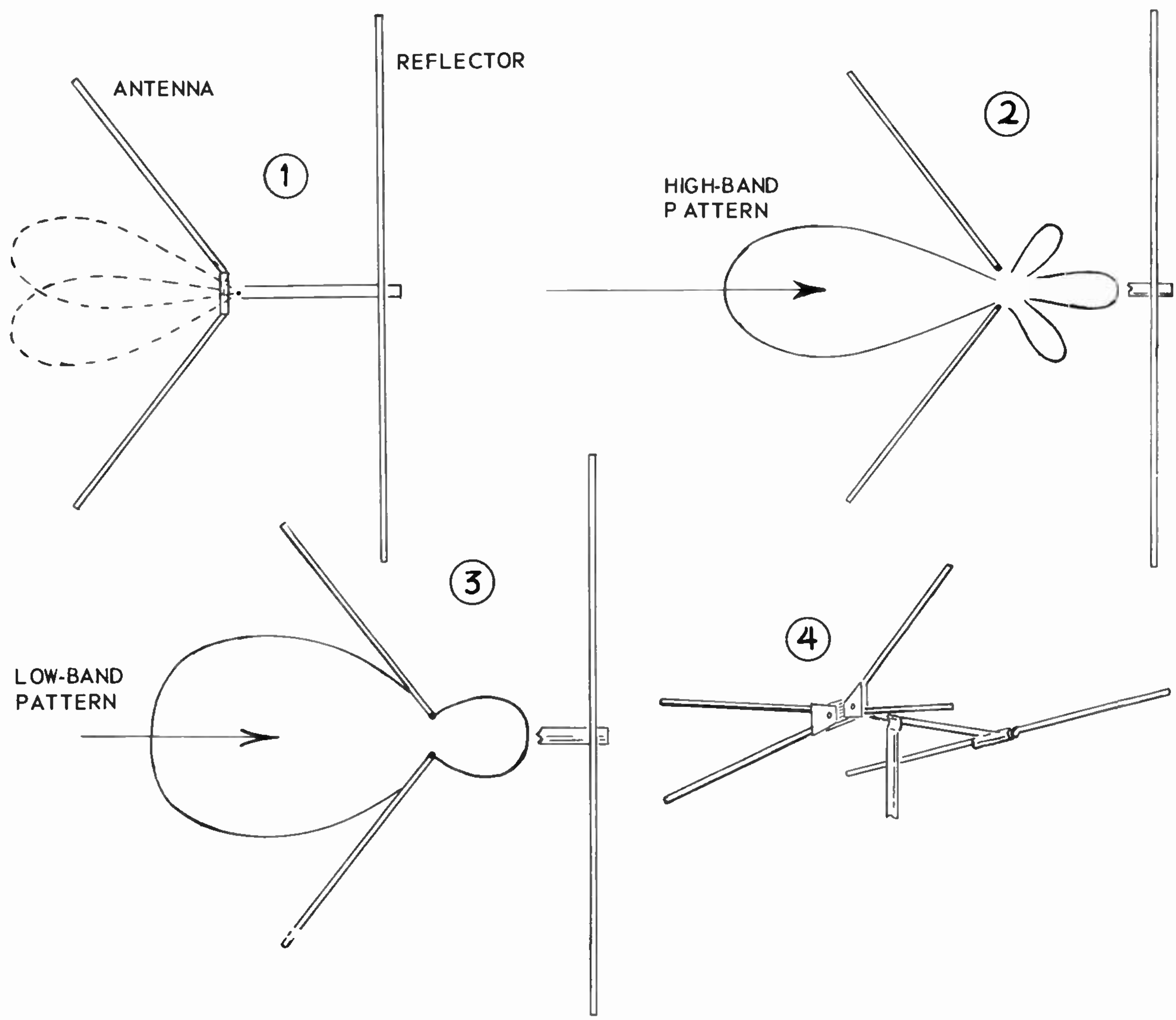


Fig. 67-16. How a conical antenna merges the high-band directional lobes.

signals from the same general direction the two opposite ends or halves of the antenna conductor may be brought forward into sort of a "V" formation as at 1 in Fig. 67-16. The two forward lobes will change positions along with the antenna conductors, and will be pushed together as shown by the broken-line patterns. The result is merging of the two lobes into a single forward pattern, as shown at 2. This becomes the directional pattern for high-band reception. The addition of a reflector reduces the back lobes about as indicated, although these minor lobes will change in number and size with variations of frequency throughout the band.

Bringing the antenna conductors forward into a V-formation has the effect of squeezing the low-band pattern into a somewhat narrower angle than with a straight antenna. The result is shown by diagram 3, which is the approximate directional pattern for low-band reception.

The complete array may appear about as illustrated at 4. Using the two pieces of tubing for each antenna conductor or in each side of the antenna helps to broaden the frequency response. Instead of two pieces of tubing there may be three on each side, as shown by some earlier illustrations. In this case the middle piece of tubing may be shorter than the other two. A director cut for high band operation sometimes is mounted in front of the antenna element. Arrays of this general style usually are called conical antennas, because the conductors which angle forward and outward from the center at the same time could be considered as sections taken from two cones having their apexes at the center of the structure.

Most conical antennas have a center impedance of approximately 150 ohms. There will be no serious mismatch when the transmission line has a characteristic impedance of 75, 150, or 300 ohms. The gain ordinarily is something between 2 and 5 decibels in the various channels.

DOUBLE-V ANTENNA. Another type of antenna making use of third harmonic response for reception

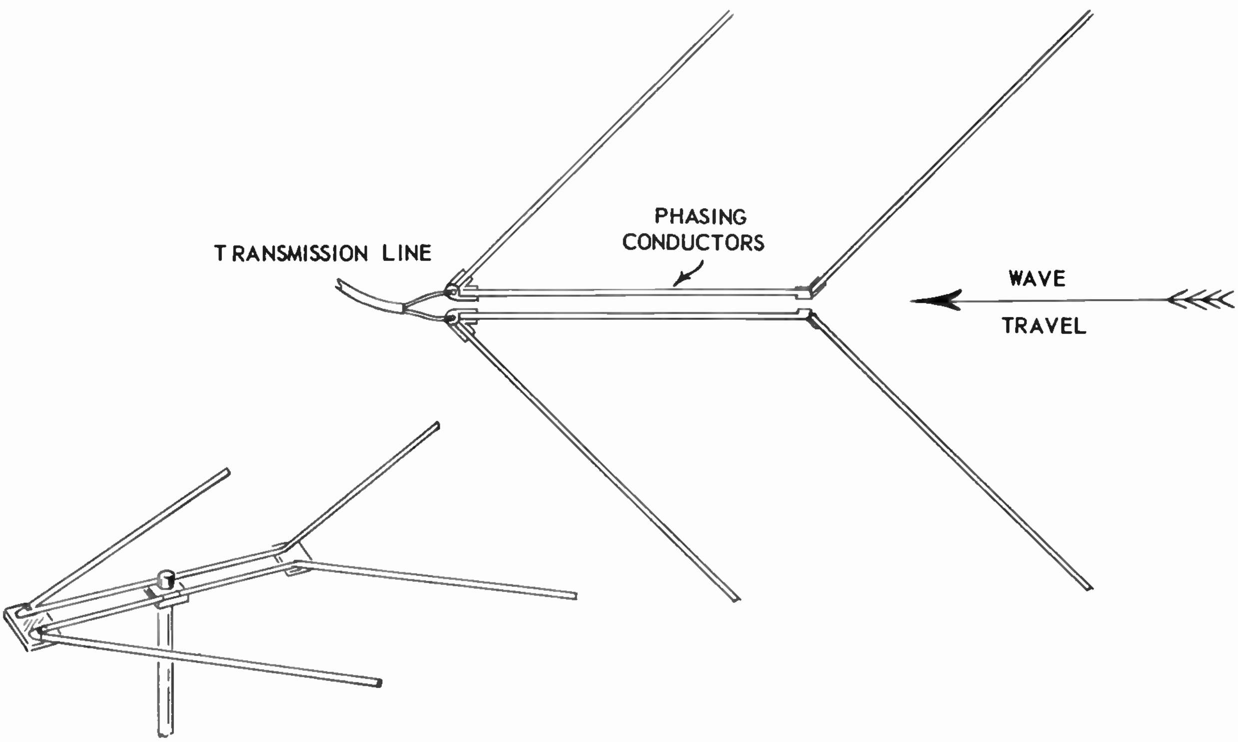


Fig. 67-17. A double-V antenna.

throughout both the high and the low television bands is illustrated by Fig. 67-17. This style may be called either a double-V or a V-beam array. The styles which we have referred to as conicals sometimes are called V-beam conicals.

The double-V antenna consists of a forward dipole and a rear dipole with all conductors in the same horizontal plane. Each of the V-dipoles is divided at its center for signal power takeoff to the transmission line, just as all dipoles are divided. The gap openings or center terminals of the two dipoles are connected together by two fore and aft phasing conductors. The right-hand side of each dipole is connected through one of these phasing conductors to the right-hand side of the other dipole, and the left-hand sides of the two dipoles are similarly connected through the other phasing conductor. Right-hand and left-hand here refer to the sides of the antenna as it is mounted in space, not as shown by the diagram. The two conductors of the transmission line are connected to the rear ends of the two phasing conductors, or to the gap terminals of the rear dipole.

Front to back separation between the two dipoles is such that their signal currents are in phase, or nearly so, and so that signal powers add at the transmission line takeoff. Directional patterns for high-band and low-band reception are quite similar to those shown in Fig. 67-16 for a conical antenna. The angle included between the forwardly inclined conical conductors in that figure is about 105 degrees, while the included angle of the double V-antenna illustrated is 90 degrees. As a consequence of the smaller angle the directional lobes of this particular double-V antenna would be somewhat narrower than for the conical type illustrated. Narrowing the included angle of either style of antenna will narrow the directional lobes while giving somewhat greater gain in a straight forward direction.

a *STACKED ANTENNAS.* A stacked array, as shown by Fig. 67-18, consists of either two or four similar antennas mounted one above another, connected together and to the transmission line in such manner as to combine their signal currents and powers. The prime object of stacking is to provide greater gain. Any style of antenna, with or without any number and kind of parasitic elements, may be stacked. Each antenna element or each combination of elements in a stacked array usually is called a bay.

All the buys in a stacked array are of similar design, construction, and frequency response. For this reason we do not call a high-low antenna a stacked array, the two sections have different frequency responses. Neither would any other two dissimilar elements on the same mast form a stacked array. You could, however, use two similar elements for one two-bay stack cut for the low band, and on the same mast mount another two-bay stack of the same or a different style cut for the high band. Then you would have two separate stacked arrays on the one mast. Any stacked array will have essentially the same frequency response or bandwidth as one of its bays used alone.

The horizontal directional pattern is not materially altered by stacking. The entire stacked array will have practically the same horizontal pattern as one of its bays. But the vertical directional pattern is greatly improved by stacking. That is to say, a stacked array has much less pickup than one of its bays used alone for waves arriving from either below or above the horizontal plane in which the antenna lies. This may be a decided advantage where there is electrical and radio interference in the area where the antenna must be located. Vertical pickup was discussed in another lesson, where directional patterns were shown for waves arriving at angles below or above the horizontal. Each bay of a stacked array seems to act for the other bays somewhat like a shield against waves not traveling in a nearly horizontal direction.

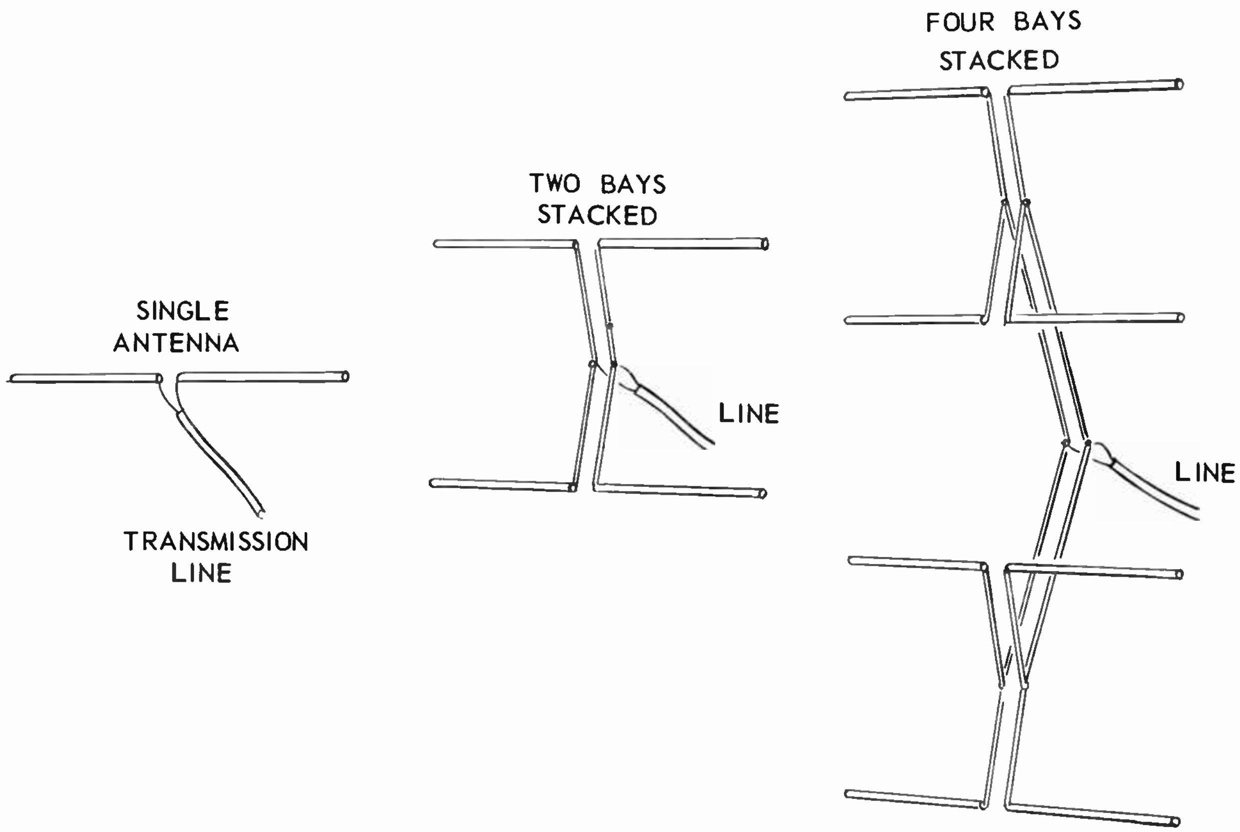


Fig. 67-18. Single and stacked antennas with their phasing connections.

Increases of gain due to stacking are illustrated by Fig. 67-19, which shows fairly representative gains in decibels when using a single conical antenna with reflector, when using the same kind of antenna and reflector for each of two bays in a stacked array, and when using four similar bays. The zero line at the bottom of this graph represents signal power from a simple half-wave dipole without reflector, which is the reference standard.

Make sure that you correctly interpret the additional gains due to stacking. As an example, the graph shows a two-bay stacked array as providing gain of about 6.7 db in comparison with a simple dipole on channel 5. This does not mean that you will have 6.7 db more signal power than from a single conical antenna and reflector, because the gain with a single such unit would be about 4.0 db as compared with a simple dipole. Extra gain due to the two-bay stacking is the difference between 6.7 db and 4.0 db, or is 2.7 db. This is a signal power increase of about 87 per cent of the power from one bay. Power is nearly but not quite doubled. Using four bays would give a signal power increase of about two-thirds of that from two bays, and about three times the power from a single bay.

To realize the extra gain which is possible with stacking there must be as much vertical separation as is practicable between the bays. Separation usually is limited by overall size and mounting difficulties. Also, the bays must be connected together and to the transmission line for correct phasing of signal currents and satisfactory matching of impedances.

If a stacked array were designed for reception in only one channel, as would be the case with two or four yagis or with three-element antennas for the bays, a vertical separation of one wavelength would allow maximum feasible gain, and with separation of a half-wavelength there still would be high gain. But were a stacked array designed for broad band reception the one-wavelength separation would throw the high-band signals away out of phase.

To have reception in both television bands there must be a compromise in vertical separation between bays. One method employs separation of about a quarter-wavelength measured at the lower frequencies of the low band, which is a separation of somewhat less than a full wavelength at the higher frequencies of the high band. This separation favors high-band reception. Greater separation would favor the low band.

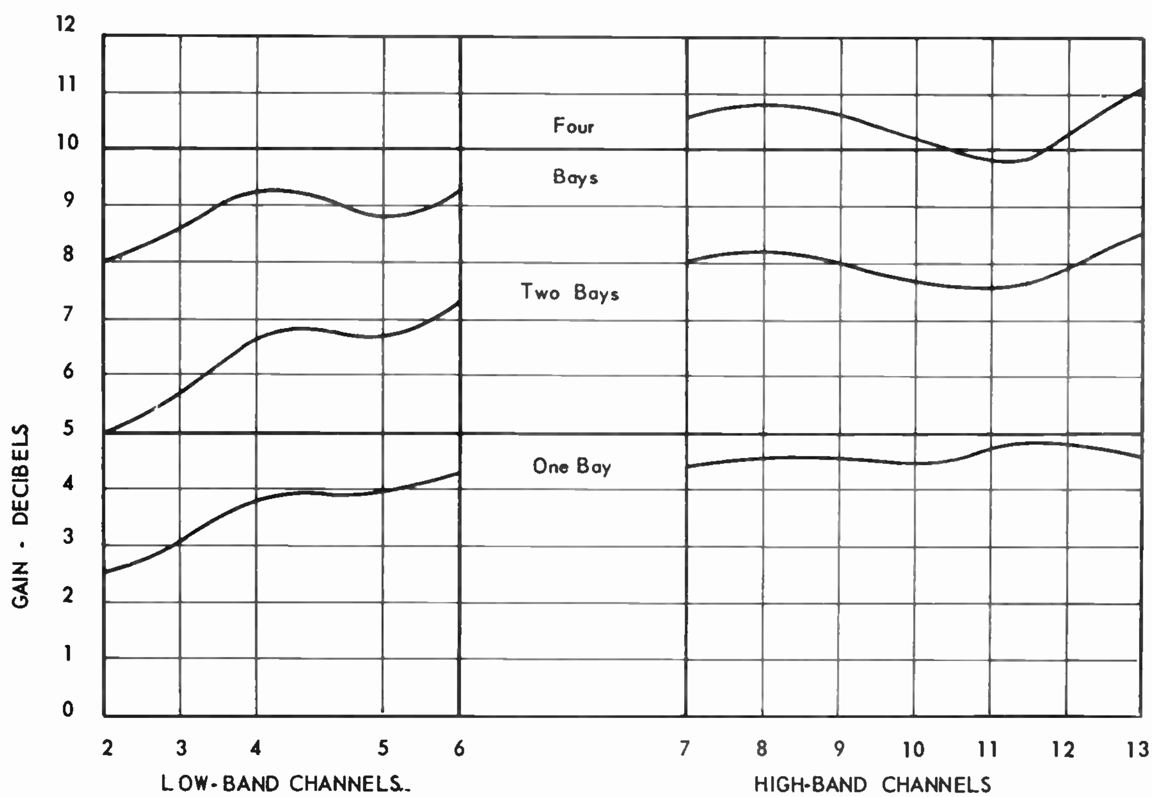


Fig. 67-19. Typical gains for single and stacked conical antennas.

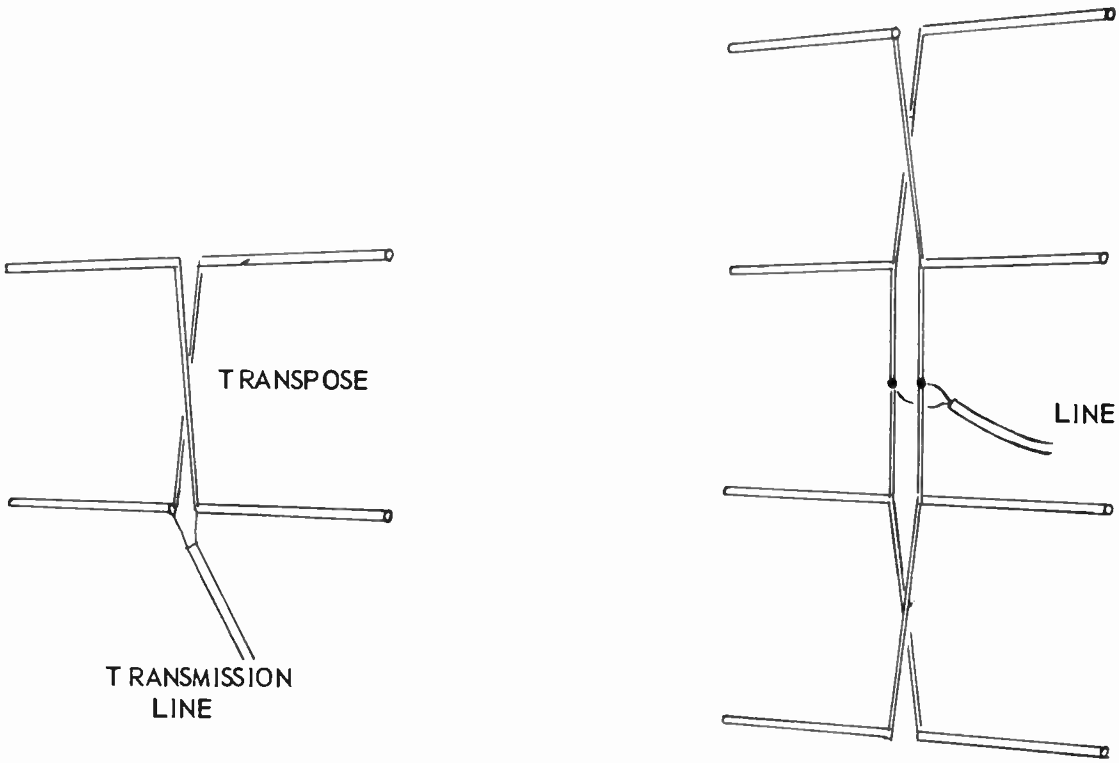


Fig. 67-20. End takeoffs for phasing of stacked antennas.

Center takeoff for transmission lines on stacked antennas is illustrated by Fig. 67-18. With only two bays the vertical phasing conductors are parallel, with the transmission line connected at the centers of the two conductors. These conductors may be either metal tubing or pieces of transmission line. With four bays the upper two are connected to each other with center-tapped phasing conductors, and the lower two are similarly connected to each other. These center-tap points then are connected together by two more phasing conductors, and from midpoints on these latter conductors the transmission line goes to the receiver.

End takeoff for stacked arrays is illustrated by Fig. 67-20. With only two bays the transmission line may be connected to the gap terminals of either bay. Then the phasing connections between bays must be transposed as shown. With four bays the top two are connected together with transposed phasing conductors and the bottom two are similarly connected together. The top and bottom pairs then are connected together with parallel phasing conductors, and from the centers of these latter conductors the transmission line is run to the receiver.

The stacked arrays are shown as simple straight dipoles to clearly illustrate the methods of interconnection. In practice there nearly always would be reflectors, possibly directors as well, and the bays could be made up with any other types of antenna elements.

H-ANTENNAS. Fig. 67-21 illustrates some styles of stacked antennas which have long been in common use. All are called H-antennas or lazy-H antennas, because they appear like a capital letter H lying on its side.

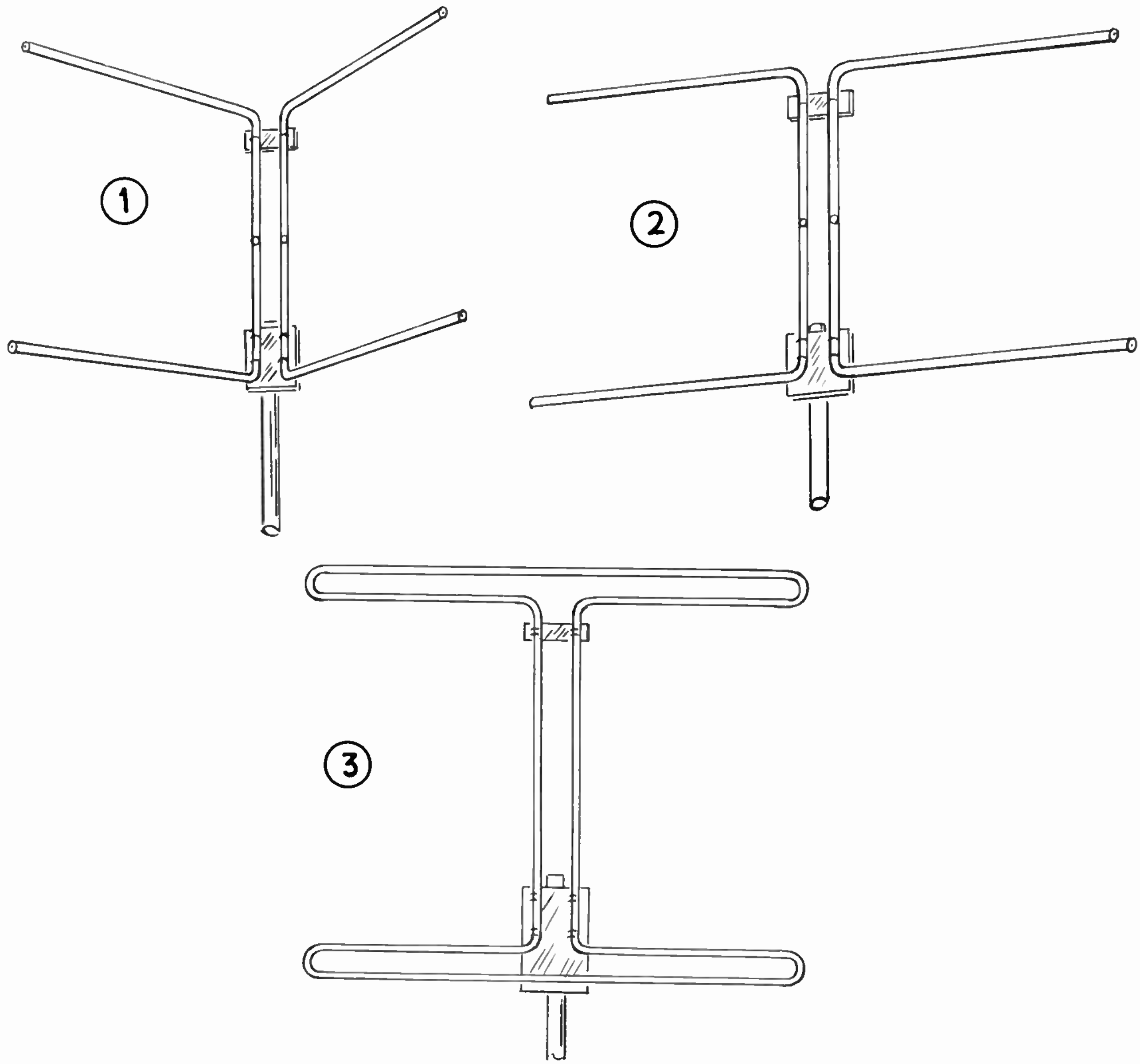


Fig. 67-21. Various H-type or lazy-H stacked antennas.

At 1 the top and bottom elements are half-wave dipoles with the outer ends brought forward into a V-formation. When this design is used with straight reflectors back of each dipole the bandwidth is great enough for either the low band or the high band where signal field strength is fairly high, but it is not enough for both bands. Gain is around 3 to 5 decibels. Since this is a single band antenna, cut for reception in either one band or the other, the principal effect of the V-formation is a narrowing of the directional pattern rather than merging of separate lobes. This narrowing may be an advantage where front reception is wanted only within a rather narrow angle.

At 2 the top and bottom elements are straight half-wave dipoles. The directional pattern is like that of a single straight dipole, wider than when the V-formation is used. This straight dipole array ordinarily would include a director behind each element. Two or more directors sometimes are added to increase the gain, with proportional reduction of bandwidth and narrowing of the directional pattern.

Diagram 3 of Fig. 67-21 shows two folded dipoles in a stacked array. Where field strength is fairly good the bandwidth is great enough for reception throughout either band for which the antennas are cut, and in areas of high signal strength this antenna will cover both bands when cut for a frequency of about 150 megacycles.

When any of the antennas illustrated by Fig. 67-21 are cut for high-band reception the vertical separation between bays usually is about a half-wavelength at the middle of that band. When cut for the low band the vertical separation usually is about a quarter wavelength at the middle of the low band. This lesser separation for the low band is only to reduce overall dimensions and make for easier mounting. As with all stacked antennas, greater vertical separation will increase the gain quite materially.

Although there are many forms of antennas in addition to those which we have discussed, enough types have been illustrated and described to bring out the general principles employed for all antennas. Other varieties should hold no mysteries.

TELEVISION

LESSON NO. 68

RESONANT LINES AND MATCHING STUBS

"Training Your Mind"

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

-A

Chicago, Illinois

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON NO. 68

RESONANT LINES AND MATCHING SETS

Imagine someone coming to you and claiming to have in their pocket a wonderful invention that will:

1. Compensate for mismatch of impedances between antennas, transmission lines, and receivers with negligible energy loss.

2. Bring up the response on a weak channel when connected to the receiver input.

3. Provide either inductive or capacitive reactance, and acts like either a coil or a capacitor for constructing such things as frequency filters.

4. Act as a precisely tuned series resonant circuit or a parallel resonant circuit, whichever you want. This makes it useful as an interference trap for undesired television, f-m, and high-frequency radio signals.

5. Either in its present form or with less insulation be used as a television tuner, and also in making built-in antennas for television receivers.

On top of all this, the invention has far better efficiency or higher Q-factor at very-high and ultra-high frequencies than the coils and capacitors you have been using. When you get a look at this remarkable gadget it turns out to be nothing but a piece of transmission line something less than four feet long - yet every claim is well justified.

When a piece of transmission line is used for purposes just listed we call it a resonant line, a tuned section, a linear circuit, or a stub. Because resonant lines or stubs are so commonly used at present, and because at ultra-high frequencies they are the most practicable means for avoiding excessive energy loss, it is none too soon to get acquainted with some of their elementary principles.

To begin with, we must realize that currents and voltages at very-high and ultra-high frequencies are not of constant values at all points along a line. We may have several cycles on a line or a conductor at the same time. Supposing, for example, that you have a source of voltage at 197 megacycles connected to one end of a line whose other end is open, as at 1 in Fig. 68-1. An alternation or pulse of voltage from the source will start a charge moving along the line. For the present we shall assume that the charge travels at the same velocity as radiant waves in space. Because the wavelength corresponding to 197

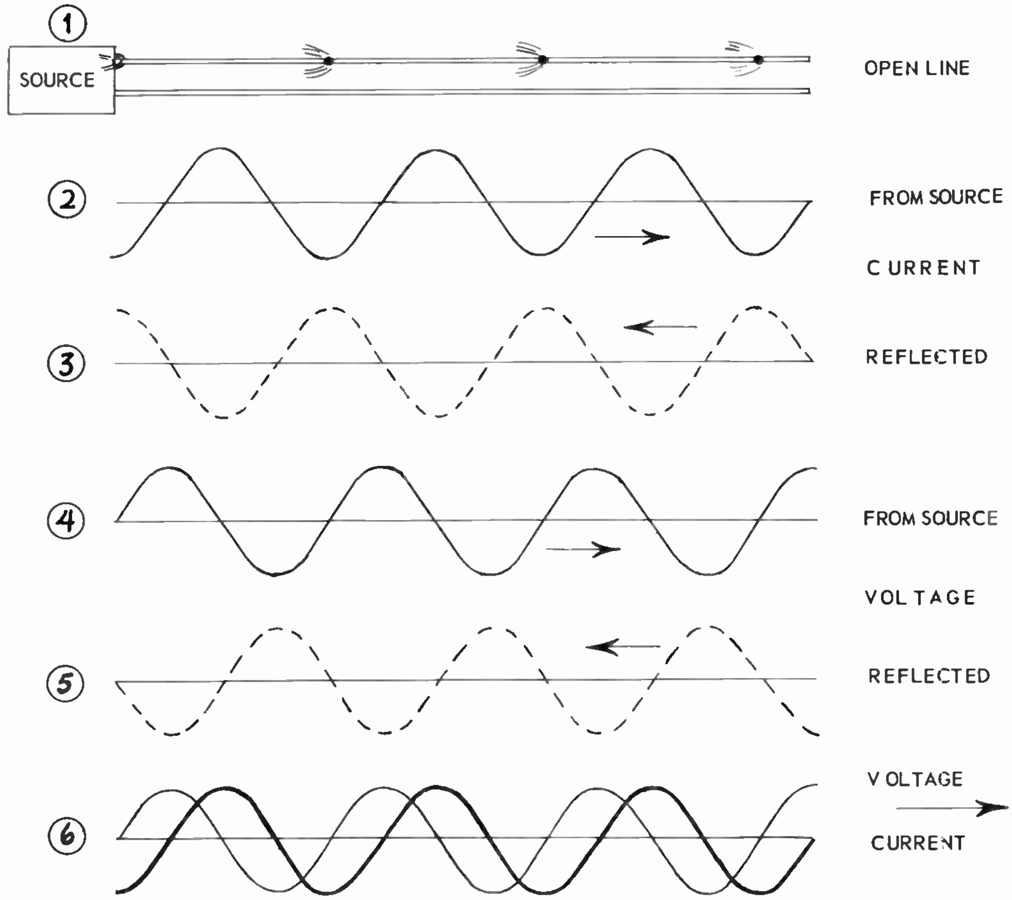


Fig. 68-1. Advancing and reflected current and voltage on an open transmission line.

megacycles is 5 feet, this charge will have progressed 5 feet before the next charge is started along the line. The next charge will be followed by a third and by a continuing succession of charges at intervals of 5 feet. Of course, the same thing happens on the other side of the line as alternating voltage from the source reverses during each cycle.

Since the moving charges are concentrations of electrons or peaks of current we may represent current along the line as at 2. When each charge reaches the open ends it will be reflected back toward the source. The reflected current is represented at 3. It is shown in opposite polarity to the source current because the reflected current is moving in the opposite direction, and because positive and negative alternating polarity are merely convenient terms which indicate that electron movement is in one direction or the opposite direction. Because there can be no current across an open circuit we must show the current as zero at the open end of the line or at the point of reflection.

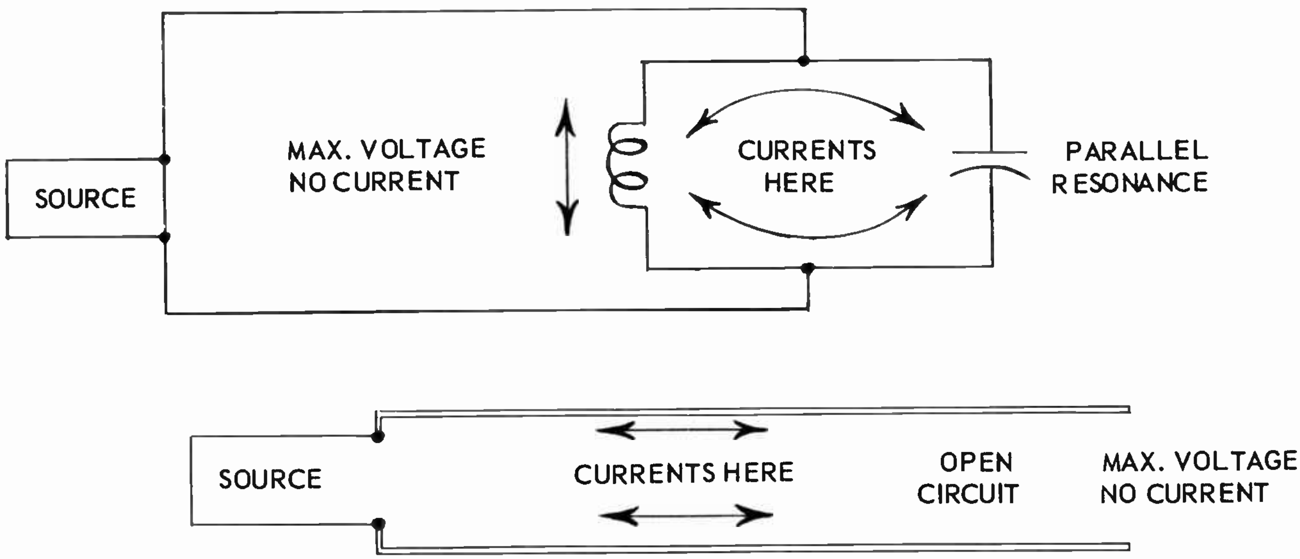


Fig. 68-2. Why an open-ended line is electrically equivalent to a parallel resonant circuit.

Changes of current must be accompanied by changes of potential, or by a voltage. We show this voltage at 4 in Fig. 68-1. It is of maximum value at the open end of the line because at any open circuit where no current can move there is nothing to cause loss of potential energy or drop of voltage. Reflected voltage, as represented at 5, is in opposite polarity to voltage from the source, because the reflected voltage is acting in the opposite direction.

Diagram 6 shows current from the source by a heavy-line curve and shows voltage from the source by a light-line curve. These are the same curves as at 2 and 4, here shown together to illustrate phase relations between current and voltage. The reflected current and voltage would have similar phase relation, but would be of opposite polarities.

It is shown by Fig. 68-2 that the open end of our line is electrically equivalent to a parallel resonant circuit containing only inductance and capacitance. When voltage at the resonant frequency is applied across a parallel resonant circuit there will be circulating currents in the inductance and capacitance, but no current from the source will pass through the resonant circuit. We have the same conditions with the open line, because currents flow in the inductance and capacitance of the line but not across the open end.

Now look at Fig. 68-3. In order that our open-ended line may act toward the source as would a parallel resonant circuit we must have zero current and maximum voltage at the source as well as at the far end of the line. This will happen only when the length of line between source and open end is equal to some number of half-wavelengths for the frequency being used. It is a half-wavelength from the open end at a back to b, another half-wavelength from b to c, and another half-wavelength from c to the source

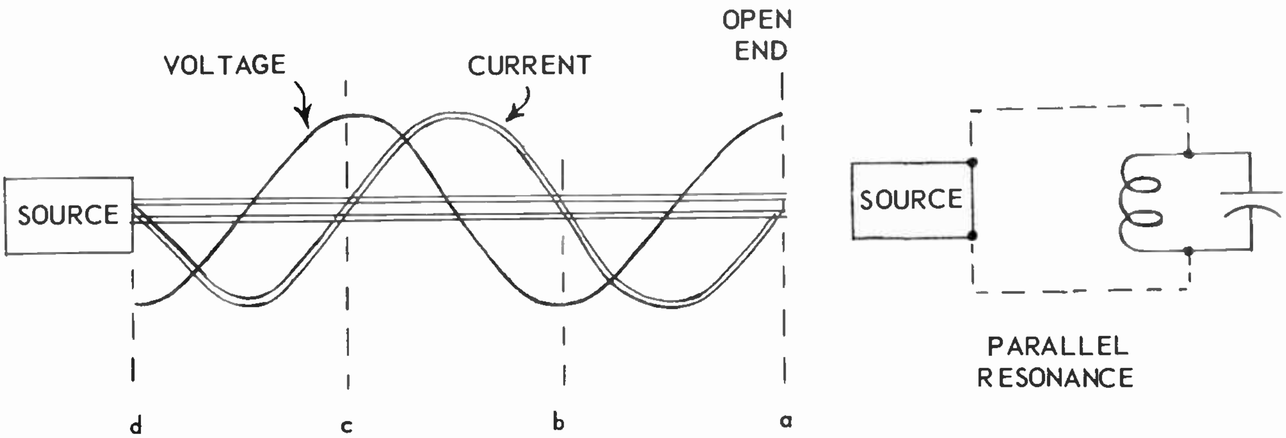


Fig. 68-3. Relations of current and voltage at half-wave intervals along an open line.

at d. You could place either the open end of the line or else the source at any of these half-wave intervals and always have maximum voltage and zero current at both ends of the line.

This allows writing the first rule for resonant lines. When length of an open-ended line is equal to any number of half-wavelengths, the source will see the line as an open circuit and will see the equivalent of a parallel resonant circuit.

Now let's see what will happen when the far end of the line is short circuited rather than being left open. The conditions are shown by Fig. 68-4. The short circuit is electrically equivalent to a series resonant circuit, for the following reasons. Flowing across a short circuit there is maximum possible current. This is true also of a series resonant circuit. There is reflection from a series resonant circuit

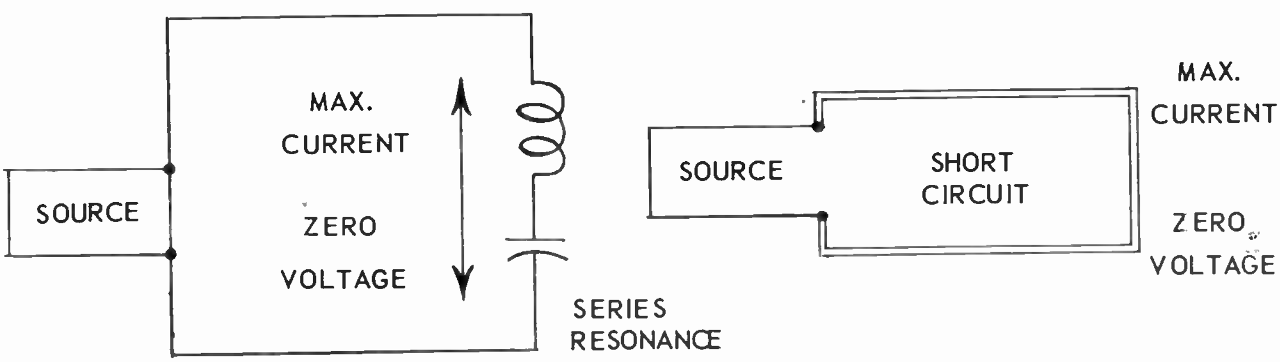


Fig. 68-4. Why a short circuited line is electrically equivalent to a series resonant circuit.

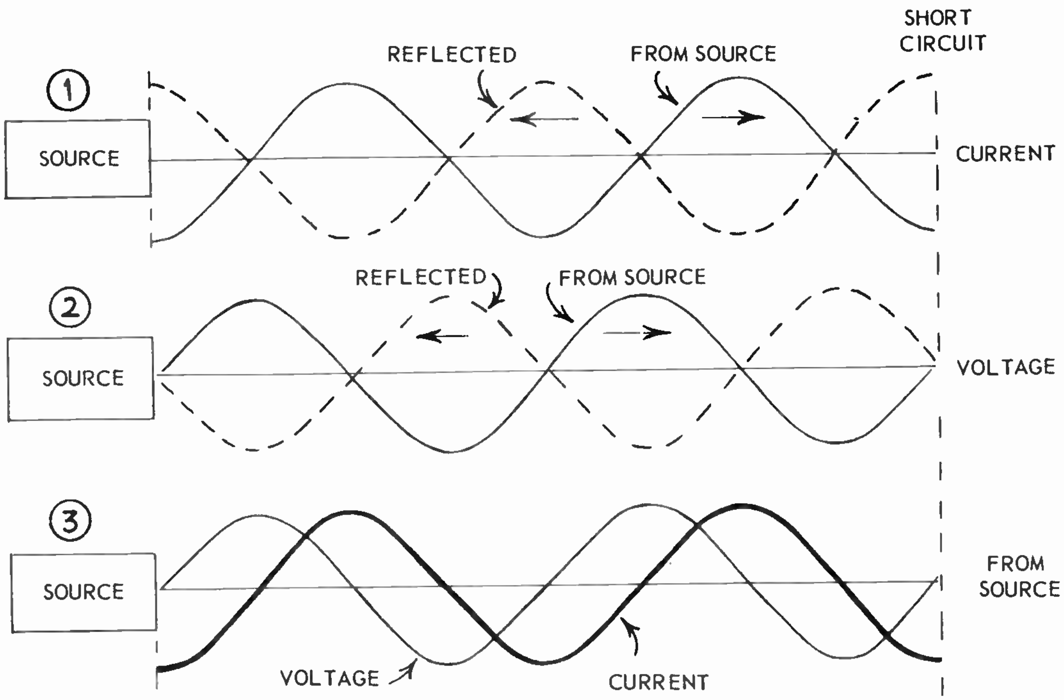


Fig. 68-5. Advancing and reflected current and voltage on a short circuited transmission line.

because neither its inductance nor its capacitance absorb any energy. They temporarily change received energy into magnetic and electric fields, which disappear as their total energy returns to the line. There is total reflection also at a short circuit if its resistance is zero.

Current from the source in a shorted line is represented by the full-line curve at 1 in Fig. 68-5. Reflected current is shown by the broken-line curve. Reflected current travels oppositely to current from the source, so is shown in opposite polarity. In order to indicate zero voltage at the short circuit the voltage wave must be shown as at 2. Reflected voltage acts oppositely, and is shown as of opposite polarity by the broken-line curve. Current and voltage from the source are shown together at 3. Reflected current and voltage could be shown by inverting the polarities.

6 In order that the source may see the shorted line as a series resonant circuit, the current at the source must be maximum when voltage is zero, just as at the shorted end of the line. This will happen only when the line between source and shorted end is some number of half-wavelengths long, as you may see from Fig. 68-6. Points, a, b, c, and d are separated by half-wavelengths. At each of these points there is maximum current and zero voltage. Either the source or the shorted end of the line could be at any one of these points and the source still would see the shorted line as a series resonant circuit or as a short circuit.

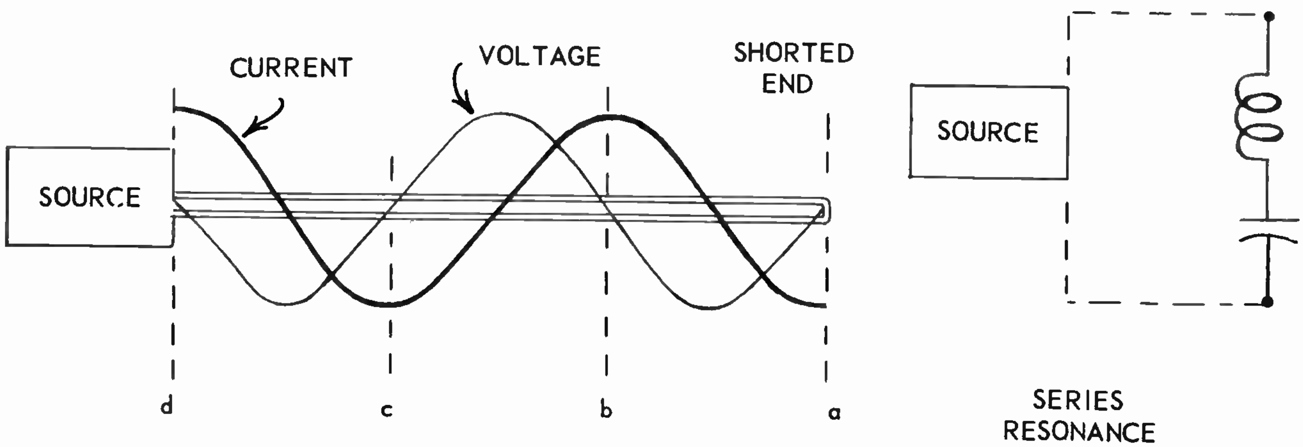


Fig. 68-6. Relations of current and voltage at half-wave intervals along a shorted line.

Now we have a second rule for resonant lines. With length of a shorted line equal to any number of half-wavelengths the source will see the equivalent of a series resonant circuit or will see a short circuit.

Next we shall see what happens when open and shorted lines are made only a quarter-wavelength long. As at 1 in Fig. 68-7, we have zero current and maximum voltage at the end of an open line, just as in Fig. 68-1. At point a, a quarter-wavelength back from the open end, we have zero voltage and maximum current. But these are the relations of voltage and current existing at a short circuit or at a series resonant circuit. When the open line is cut down to a single quarter-wavelength, as at 2, and the source is connected as shown, there is zero voltage and maximum current at the source. Then the source sees this quarter-wave open line as a series resonant circuit.

At 3 in Fig. 68-7 are shown zero voltage and maximum current at the end of a shorted line, just as in Fig. 68-5. At a distance of a quarter-wavelength back from the shorted end of the line there is maximum voltage and zero current. This is the voltage-current relation that exists at a parallel resonant circuit. Therefore, if the shorted line is made just a quarter-wavelength long, as at 4, the source will see this quarter-wave shorted line as a parallel resonant circuit.

Fig. 68-8 summarizes what we have learned about half-wave and quarter-wave open and shorted lines, and adds other important information. Voltage sources are indicated by the usual symbol containing a wavy line representing one cycle.

1. Half-wave open lines appear to the source as parallel resonant circuits or open circuits.

2. Quarter-wave open lines appear to the source as series resonant circuits or short circuits.

3. Open lines less than a quarter-wavelength long appear to the source as capacitors or capacitive

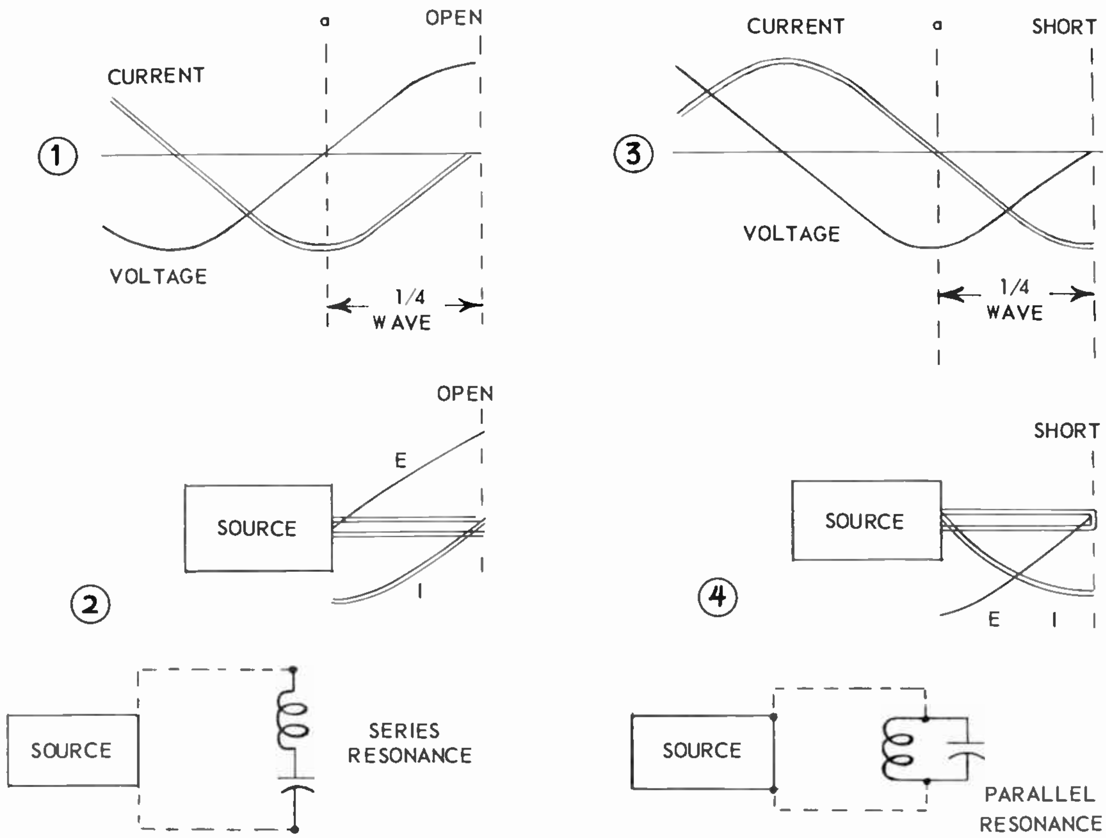


Fig. 68-7. Relations between values of current and voltage are inverted by quarter-wave lines.

reactances. Reactance increases, or capacitance decreases as the line is made shorter and shorter below a quarter-wavelength.

4. Open lines between a quarter-wavelength and a half-wavelength long appear to the source as inductors or inductive reactances. Reactance and inductance increase as the line is made longer than a quarter-wavelength, and become maximum at a half-wavelength where there is the equivalent of an open circuit.

5. Half-wave shorted lines appear to the source as series resonant circuits or short circuits.

6. Quarter-wave shorted lines appear to the source as parallel resonant circuits or open circuits.

a 7. Shorted lines of less than a quarter-wavelength appear to the source as inductors or inductive reactances. The shorter the line the less becomes the inductance or inductive reactance.

8. Shorted lines between a quarter-wavelength and a half-wavelength long appear to the source as capacitors or capacitive reactances. Reactance decreases or capacitance increases as length is increased from a quarter-wavelength toward a half-wavelength.

ADDING HALF-WAVE SECTIONS. In Figs. 68-3 and 68-6 we observed that increasing or decreasing the length of either an open or shorted line by any number of half-wavelengths has no effect on the way in which the source sees the line. It is just as true that adding any number of half-wavelengths to any of the line sections of Fig. 68-8 will not alter the way in which the source sees the line.

For example, at 2 and at 6 in that figure there are quarter-wave lines. The total lengths could be made 3/4 wavelength or 1¼ wavelengths, or of any other length consisting of the original quarter-wavelength plus any number of half-wavelengths without altering the way in which the source sees the line. The same general principle applies to lengths shown in Fig. 68-8 as less than a quarter-wavelength and to those between a quarter- and a half-wavelength, you may add any number of half-wavelengths without altering the way in which the source sees the line.

Should you wish to know how a line of any random length will behave, you could measure off as many half-wavelengths as would leave a final half-wavelength or something less. This remaining part will tell the story, as shown by Fig. 68-8.

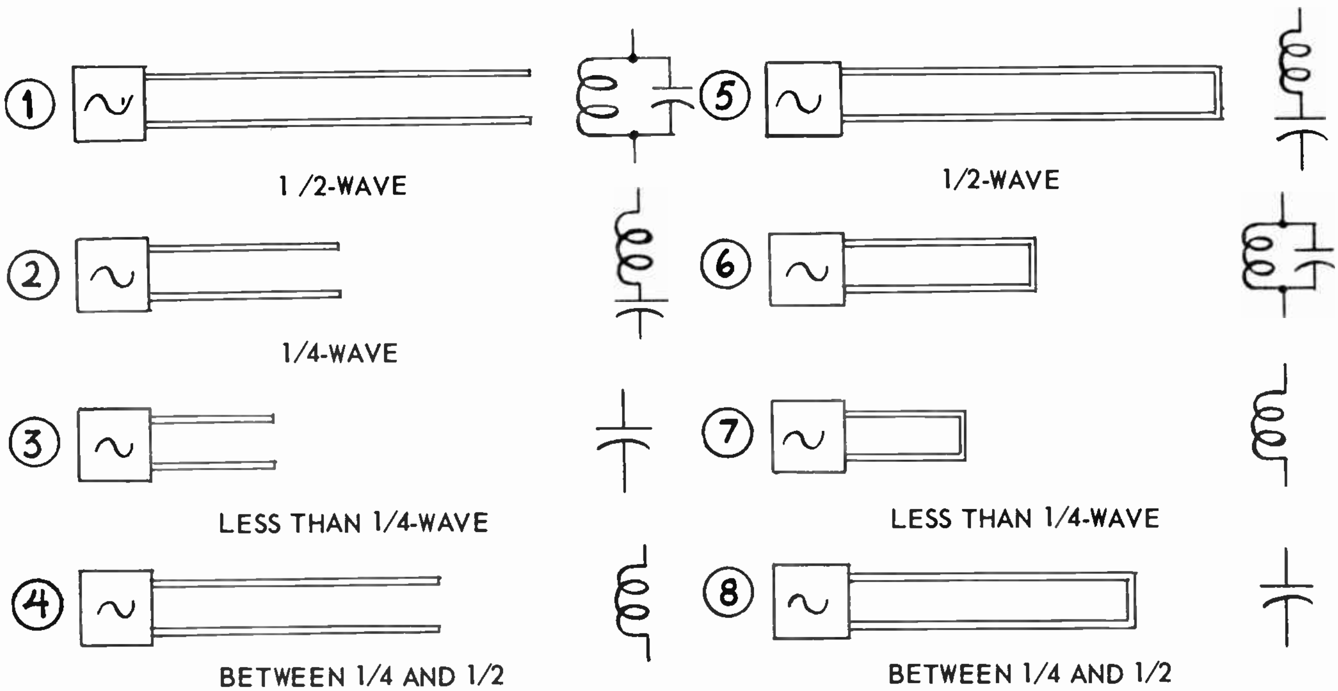


Fig. 68-8. How the source sees open and shorted lines of various lengths between a half-wavelength and zero.

4 We may draw an important conclusion from the repeating effect of half-wavelengths. If the total length of a resonant line is any exact multiple of a half-wavelength at the operating frequency, then the impedance or the reactance at the load end will appear at the RF source end without any change. We say that a half-wavelength resonant line repeats the load.

TUNING OF RESONANT LINES. When tuning a circuit which consists of an inductor and capacitor you adjust either one or both to vary the frequency of resonance. Reducing the inductance, the capacitance, or both, will raise the frequency of resonance, while increasing inductance, capacitance, or both, will lower the resonant frequency.

3 When you use a greater length of transmission line as a tuned section you add to both its inductance and capacitance, and thus lower the frequency of resonance. Using a shorter length reduces both inductance and capacitance, and raises the resonant frequency.

The actual length of a resonant line, in inches, corresponding to any fraction of a wavelength depends entirely on the operating frequency. The line must be relatively short for high frequencies and longer for lower frequencies.

As a matter of convenience in supporting or mounting it is usually desirable to use short length of line. With this in mind look back at Fig. 68-8 and decide whether you would use an open line or a shorted line to obtain the effect of parallel resonance. The shorted line would have to be only a quarter-wavelength long, while the open line would have to be a half-wavelength long. But should you wish to have the effect of series resonance the length of the open line would be only half that of the required shorted line.

VELOCITY FACTORS. When a charge or a wave of current moves along a transmission line or any short section of line the charge carries along with it an electric field. And because the charge is a peak of current it carries also a magnetic field. The charge or current, and its accompanying fields, must stay together and travel at the same speed. Electrons composing the charge or current move in the copper conductors of the line, but the electric and magnetic fields which are around the conductor have to move along through whatever kind of dielectric material is around and between the conductors.

The electric and magnetic fields can travel through space at 300,000,000 meters per second, and through air at practically the same speed. But these fields cannot travel at this speed through solid or plastic dielectric materials. They are slowed down to a greater or less extent even by the least amount of insulation that can provide the supports for "air-insulated" lines made of wire or of metal tubing.

At any given frequency there must be a corresponding number of waves per second. When the waves slow down a greater number crowd into every mile or other unit of travel distance. Then each wavelength, or any fraction of a wavelength, is shorter in any kind of transmission line than in air. Some comparisons are shown by Fig. 68-9.

A half-wave or quarter-wave line must be shorter in inches than a half-wavelength or quarter-wavelength in air. The fraction of the air or space wavelength to which transmission line must be cut is called the velocity factor or velocity constant of the line. Really it is a fraction of the wave velocity in air, but this amounts to the same thing.

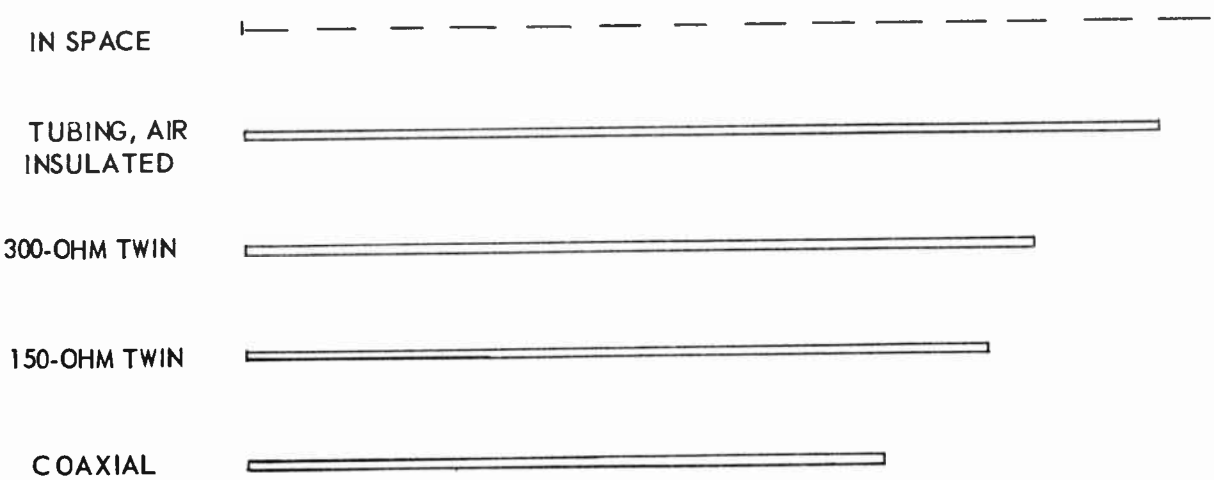


Fig. 68-9. Relative lengths in inches of several kinds of transmission line having equal electrical lengths or equal fractional wavelengths.

The accompanying table lists velocity factors for several kinds of transmission line commonly used for resonant sections or stubs. This table gives also the numbers which may be divided by megacycles of frequency to determine how many inches of actual length correspond to wavelength dimensions usually employed. Line insulation is assumed to be polyethylene in all cases. Factors for coaxial line refer to any such line as ordinarily used for television service work.

As an example in using the table, supposing you wish to cut a half-wave stub from 150-ohm twin-conductor to use at a frequency of 195 megacycles. In the half-wave column, opposite twin-conductor of 150-ohms, is the number 4,547. Dividing this number by 195 (megacycles) gives 23.3 inches as the required length.

VELOCITY FACTORS AND LINE LENGTH FACTORS

Kind of Transmission	Velocity Factor	Line Length Factors		
		1-wave	½-wave	¼-wave
Waves in space, no line used.	1.00	11810	5905	2953
Wire line, air insulated	.97	11457	5728	2864
Tubing line, air insulated	.95	11220	5610	2805
Twin-conductor, unshielded, 300-ohm	.82	9684	4842	2421
” ” 150-ohm	.77	9094	4547	2274
” ” 75-ohm	.69	8150	4075	2037
Coaxial cable, 50- to 75-ohm	.66	7795	3898	1949

MATCHING WITH INSERTED SECTIONS. One of the most useful service applications of line sections and stubs is in the form called impedance matching transformers, which connect unequal impedances of antennas, transmission lines, and receivers for maximum attainable power transfer. Such transformers often are necessities when the antenna has two or more parasitic elements which drop its impedance below that of any available transmission line. They are needed also for matching at connections between bays and for transmission line connections of stacked antennas. There is a special need for matching transformers when any of these antennas are used in areas of low carrier strength, where every bit of signal power must be conserved.

One of the simplest matching transformers consists of a quarter-wave section of line inserted between the mismatched elements, as shown by Fig. 68-10. If the antenna impedance does not match transmission line impedance, but the line does match the receiver, the quarter-wave section is connected between antenna and transmission line. If there is satisfactory matching between antenna and transmission line, but not between line and receiver, the matching section is connected between line and receiver. If there is mismatch at both ends of the transmission line, quarter-wave matching sections are connected between antenna and line and between line and receiver.

Both the antenna and the line, or the line and receiver, must see their own impedances at their connection to the inserted section in order to have correct matching. This will occur when characteristic impedance of the matching section is correctly chosen. A detailed explanation of why it happens would not be of any great help, provided we know how to get the final answer. We could begin an explanation with these familiar formulas involving a-c voltage, impedance, and power.

$$\text{Impedance} = \frac{(\text{volts})^2}{\text{Power, watts}} \qquad \text{Volts} = \sqrt{\text{Power} \times \text{impedance}}$$

And finish with these presently useful formulas.

$$(1)\quad \text{Matching impedance} = \sqrt{\text{antenna impedance} \times \text{line impedance}}$$

$$(2)\quad \text{Impedance seen by antenna} = \frac{(\text{matching impedance})^2}{\text{line impedance}} \qquad (3)\quad \text{Impedance seen by line} = \frac{(\text{matching impedance})^2}{\text{antenna impedance}}$$

Formula 1 allows determining the characteristic impedance in ohms of the quarter-wave section to be inserted between elements whose impedances are unequal. For a transformer connected at the receiver end of the transmission line we would change the term "antenna impedance" to "receiver impedance" in all the formulas.

For an example, assume that you have an antenna whose center impedance is 72 ohms, and wish to use a 300-ohm transmission line. What should be the characteristic impedance of the quarter-wave inserted section? Here is the solution, using formula 1.

$$\text{Matching impedance} = \sqrt{72 \times 300} = \sqrt{21600} = 145.5 \text{ ohms, approximately.}$$

This required impedance is so close to 150 ohms that the quarter-wave section could be made of 150-ohm transmission line and be entirely satisfactory. Supposing that we do use a 150-ohm matching section, what impedance will be seen by the antenna? To find out we use formula 2, thus,

$$\text{Impedance seen by antenna} = \frac{150^2}{300} = \frac{22500}{300} = 75 \text{ ohms.}$$

This is so close to the antenna impedance of 72 ohms as to make a practically perfect match. Next, what impedance will be seen by the line? For this we use formula 3.

$$\text{Impedance seen by line} = \frac{150^2}{72} = \frac{22500}{72} = 313 \text{ ohms, approximately.}$$

This is very close to the actual impedance of the line, 300 ohms. The line sees itself matched to the antenna, and the antenna sees itself matched to the line.

Briefly, the explanation is as follows. By looking back at Fig. 68-8 you will see that a quarter-wave line makes an actual open circuit look like a series resonant circuit or a short circuit (at 2) and makes an

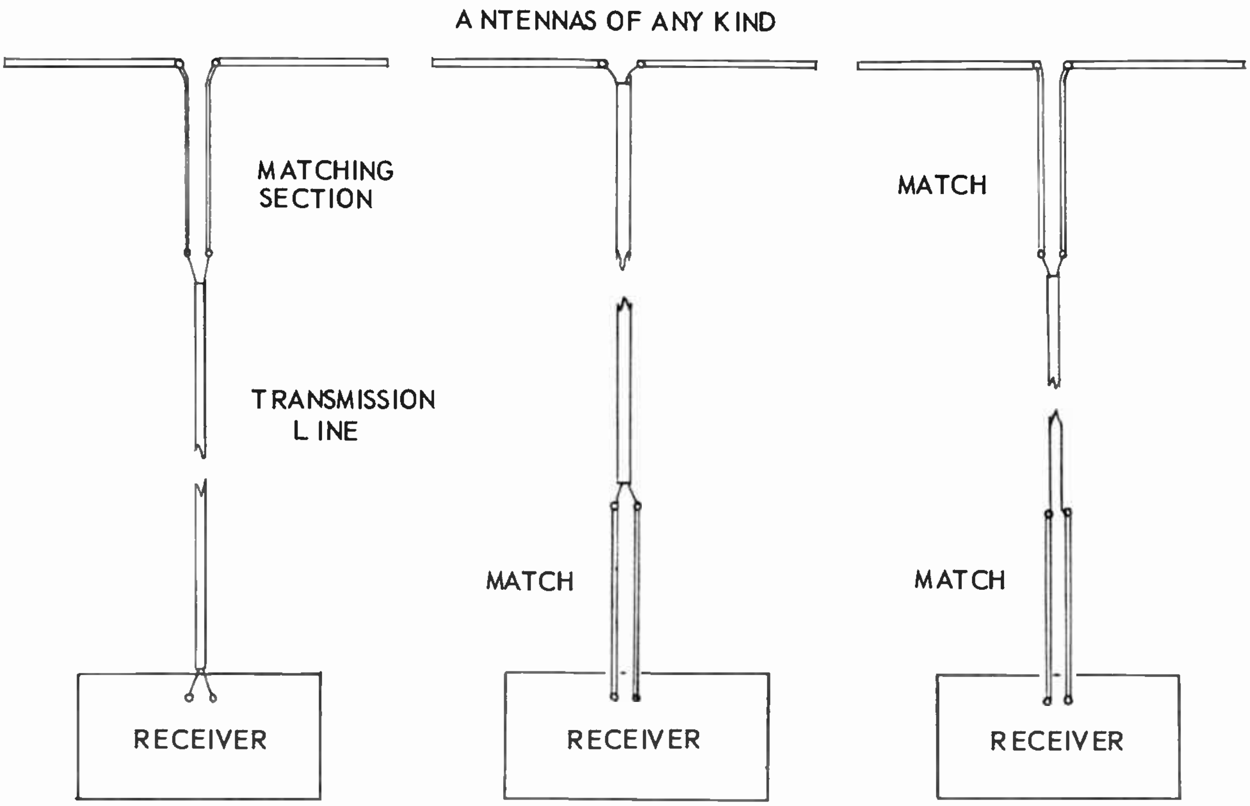


Fig. 68-10. Where inserted sections may be placed when used as matching transformers.

actual short circuit look like a parallel resonant circuit or an open circuit (at 6). The quarter-wave line is an inverter. Anything at one of its ends appears as the opposite effect when seen from the other end. This ability to invert allows the low impedance of an antenna at one end of the quarter-wave section to look into a high impedance at the line end. It allows the high impedance of the transmission line at one end to look like low impedance at the antenna end.

The inserted section may be made from any kind of transmission line, shielded or unshielded. Usually it is of the same general type as the regular transmission line except for characteristic impedance. The two conductors on one end of the inserted section are attached to the center terminals of the antenna, or to the receiver terminals, and at the other end are connected to the transmission line.

The length in inches of the inserted section must correspond to a quarter-wavelength at some certain frequency. If you are matching a single-channel antenna you will use the center frequency of that channel. For matching either the low band or high band you will use the frequency at the center of that band. For both bands or all channels you will use a frequency of 130 to 150 mc, depending on whether you wish to favor the low band or the high band. When cutting the inserted section do not forget to allow for the velocity factor of the kind of line being used. The correct length can be figured with the help of the preceding table of velocity factors in this lesson.

If no available kind of transmission line comes within about 20 per cent of the required characteristic impedance you can make the inserted section of air-insulated wires or metallic tubing. Tubing attached to the antenna terminals makes a good looking installation. Conductor diameters and spacings for the needed impedance may be determined from a chart in another lesson. Maintain the spacing by using additional supports of Bakelite, polystyrene, or other solid insulation.

MATCHING WITH STUBS. A different type of matching transformer makes use of the fact that effective impedance of a quarter-wave section varies along its length. How this property is employed with a quarter-wave open-ended stub is illustrated by fig. 68-11. In the upper diagram we have the same relations between voltage and current that are shown at 2 in Fig. 68-7.

Impedance always is equal to voltage divided by current. At the open end of the stub there is maximum voltage and minimum current. Therefore, at this open end we must have greatest possible impedance. It may be several thousand ohms. At the source end of the open stub there is minimum voltage and maximum current, so here the impedance must be of least possible value. At intermediate points along the quarter-wave open stub there are varying impedances, from greatest to least.

In the lower diagram an antenna of low center impedance has been substituted for the power source. The stub must be made from a kind of line whose characteristic impedance closely matches the antenna impedance. A transmission line of relatively high impedance is connected to the stub between antenna and open end, at a point where impedance is greater than at the antenna but less than at the open end.

By moving the transmission line takeoff along the stub it will be possible to find a position where characteristic impedance of the transmission line is matched by impedance of the stub at that particular point. Then we have an impedance match at the line takeoff and also at the antenna. This is a means for matching a transmission line to any antenna whose impedance is less than that of the line. It is a

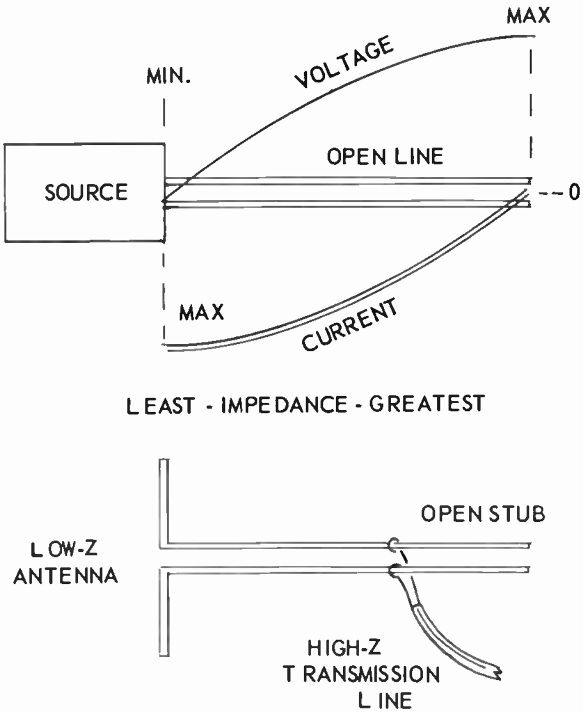


Fig. 68-11. An open stub used as a matching transformer.

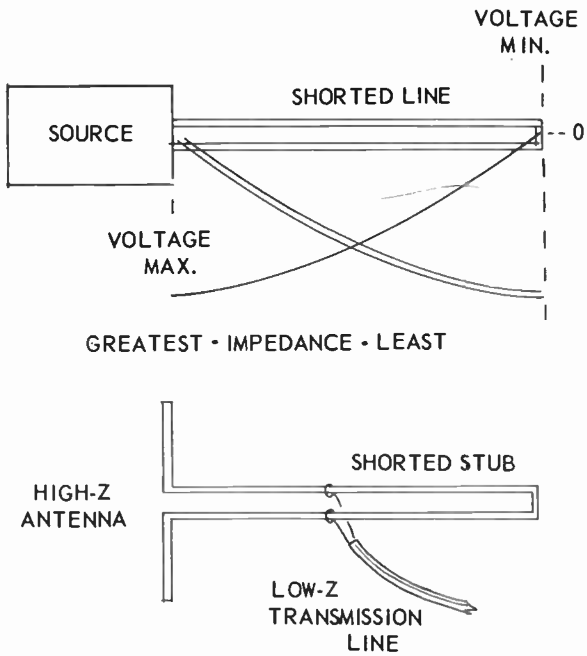


Fig. 68-12. A shorted stub used as a matching transformer.

common method of connection for antennas having several parasitic elements which drop the antenna impedance.

Supposing now that we wish to match a transmission line to an antenna whose center impedance is higher than the characteristic impedance of the line. We may use a quarter-wave shorted stub as in Fig. 68-12. The upper diagram is like the one at 4 in Fig. 68-7. There is minimum voltage at the shorted end of the line and maximum voltage at the source. Therefore, we have minimum impedance at the shorted end, maximum impedance at the source, and intermediate values of impedance along the stub.

In the lower diagram a high impedance antenna becomes the power source. The quarter-wave shorted stub must have characteristic impedance that closely matches the antenna center impedance. The low impedance transmission line is connected to a point on the stub where impedance is lower than at the antenna, to the point where transmission line impedance is matched by stub impedance. This kind of matching transformer could be used with a 300-ohm antenna and a 52-ohm or 72-ohm coaxial transmission line, or in any similar situation.

Quarter-wave stubs such as shown by Figs. 68-11 and 68-12 commonly are made of metallic tubing of the same kind and diameter as used for antenna conductors. The frequency for which the stub is cut may be arrived at as previously explained. In cutting the quarter-wave stubs their velocity factor should be considered, although with tubing the length will be only 5 per cent less than a quarter-wave in space.

If it is more convenient from the construction standpoint, as for supporting the antenna, the matching sections may be 3/4 wave long, or might be a quarter-wavelength plus any number of half-wavelengths. For making final adjustments the transmission line is moved along the stub until reception is best, and is clamped or soldered in this position.

9 *BALANCING OF REACTANCES.* If the antenna input circuit of a receiver is resonant at each received frequency, or at the middle of each channel, the receiver input will appear as a pure resistance to the transmission line, since reactances are balanced out. Then there will be nothing to cause reflections back into the line. If also the impedance at resonance is equal to the characteristic impedance of the line there will be maximum possible signal power transfer. Results will be the same if any other means are taken to make the receiver input impedance a pure resistance equal to line impedance.

If the receiver input should contain unbalanced inductive reactance it could be balanced out at one channel frequency by adding a suitable amount of opposing capacitive reactance. Should there be unbalanced capacitive reactance at the receiver input a suitable amount of inductive reactance could be added to obtain a balance. The necessary capacitive or inductive reactance may be added at the receiver input by connecting to the antenna or transmission line terminals of the receiver either an open or shorted stub of correct length, leaving the transmission line from the antenna connected as usual.

At the left in Fig. 68-13 is shown an open stub on the receiver input terminals. When this open stub is made a half-wavelength long it acts like a parallel resonant circuit or open circuit, and has no effect on reproduction. As its length is gradually lessened the stub becomes an inductive reactance of continually decreasing value. By the time the stub is reduced to a quarter-wavelength it becomes the equivalent of a series resonant circuit or a short circuit, and pictures disappear. Still more shortening makes the stub act like a capacitive reactance which first is of small value and then becomes greater as stub length is reduced toward zero. All this we learned in connection with Fig. 68-8.

At the right are shown the effects of a shorted stub. At a half-wavelength there is the effect of series resonance or a short circuit, with no pictures. As the length of the stub is reduced there is first the effect of a small capacitive reactance, which then increases in value until at a quarter-wavelength there is an equivalent parallel resonant circuit or an open circuit, with no effect on receiver performance. Still more shortening in length makes this stub act like inductive reactance, which first is of large value and then becomes less and less as the stub is cut down to zero length.

Although any degree of inductive or capacitive unbalance could be compensated for with a stub, it would take extensive measurements to determine just what fault is present. The practical thing to do is attach to the antenna terminals of the receiver one end of a piece of transmission line an inch or two longer than a quarter-wavelength at the mid-frequency of the channel giving most trouble. The length in inches may be determined from the table which lists velocity factors and line lengths. Leaving the free ends of this piece of line unconnected makes it an open stub.

Turn the receiver brightness control as low as leaves a visible picture, so that improvement will be readily noticeable. Then take your wire cutters and clip off not much more than a quarter inch of the

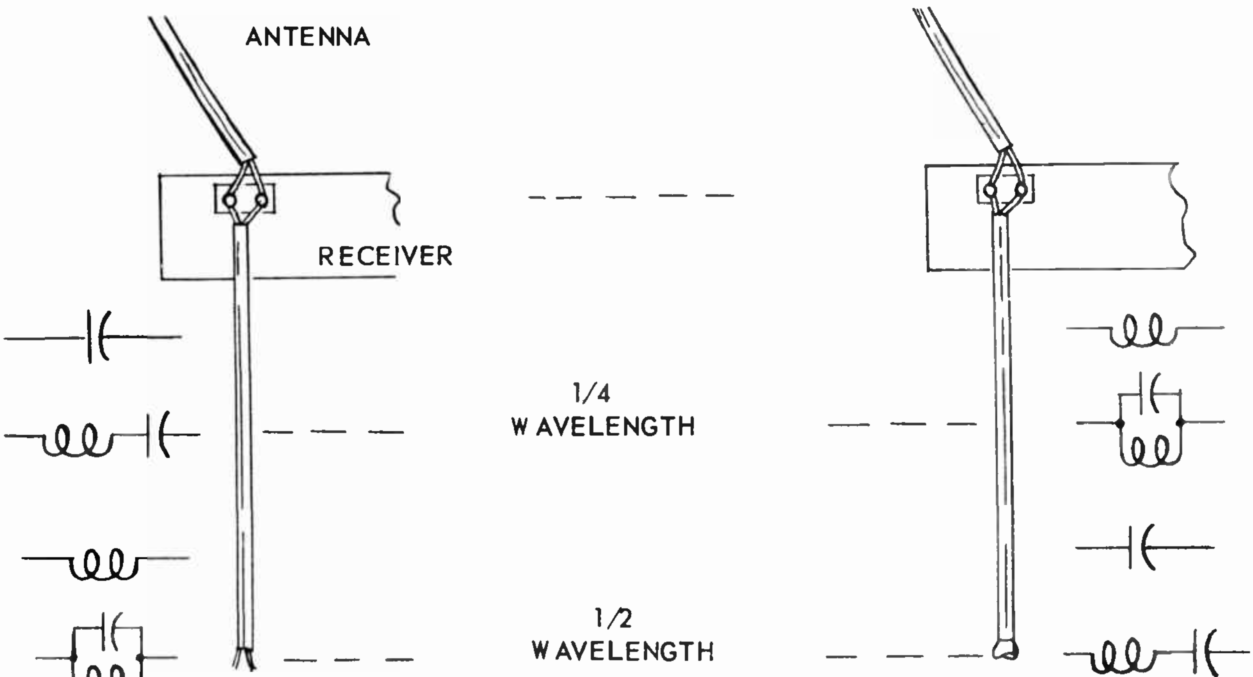


Fig. 68-13. How the length of a stub on the receiver affects the stub characteristics for preventing reflections.

stub at a time, observing picture reproduction after each clip. As you may see at the left in Fig. 68-13, this adds capacitive reactance in increasing amounts. If reproduction becomes better with any length of stub, then deteriorates with less length, you have corrected the trouble. Cut another open stub of the length which gave best results and connect it to the receiver terminals.

Should there be no improvement, start over again with the same original length of line and try a shorted stub. This adds inductive reactance to the receiver input. The inductive reactance is lessened as the shorted end of the stub comes closer to the receiver terminals.

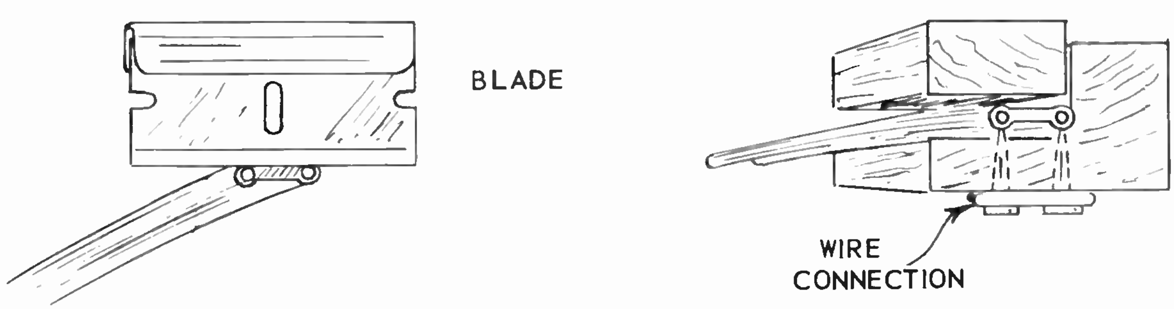


Fig. 68-14. Methods of temporarily short circuiting a stub at various points along its length.

The effective length of a shorted stub made of unshielded twin conductor may be temporarily reduced for testing as illustrated in Fig. 68-14. You may use a razor blade to cut through the plastic covering of the line to make electrical connection between the two conductors, but without cutting into either of them. Otherwise you may put the sharp points of two bright tacks or brads through a small block of wood or other insulation so that insulation on the stub will be pierced and electrical connections made when the piece of line is pressed down on the points. The tacks or brads must be electrically connected together.

Start short circuiting the stub close to the receiver, then farther and farther away at intervals not much more than a quarter inch. The portion of stub line hanging beyond the shorted place has no material effect on your tests. When you find a position of the short circuit that gives improved pictures, cut the stub about 3/8 inch longer than at this position. Bare this much of the two conductors, twist them tightly together, and solder the joint.

An open or shorted stub which allows improved pictures on the weakest channel ordinarily won't affect reproduction to any extent on other channels. Should it weaken the other channels too much the stub will have to be connected and disconnected or switched on and off as the channel selector is changed.

STANDING WAVES. When addition of an open or shorted stub improves reception, the improvement is because the stub balances excess inductive or capacitive reactance of the receiver input at some one

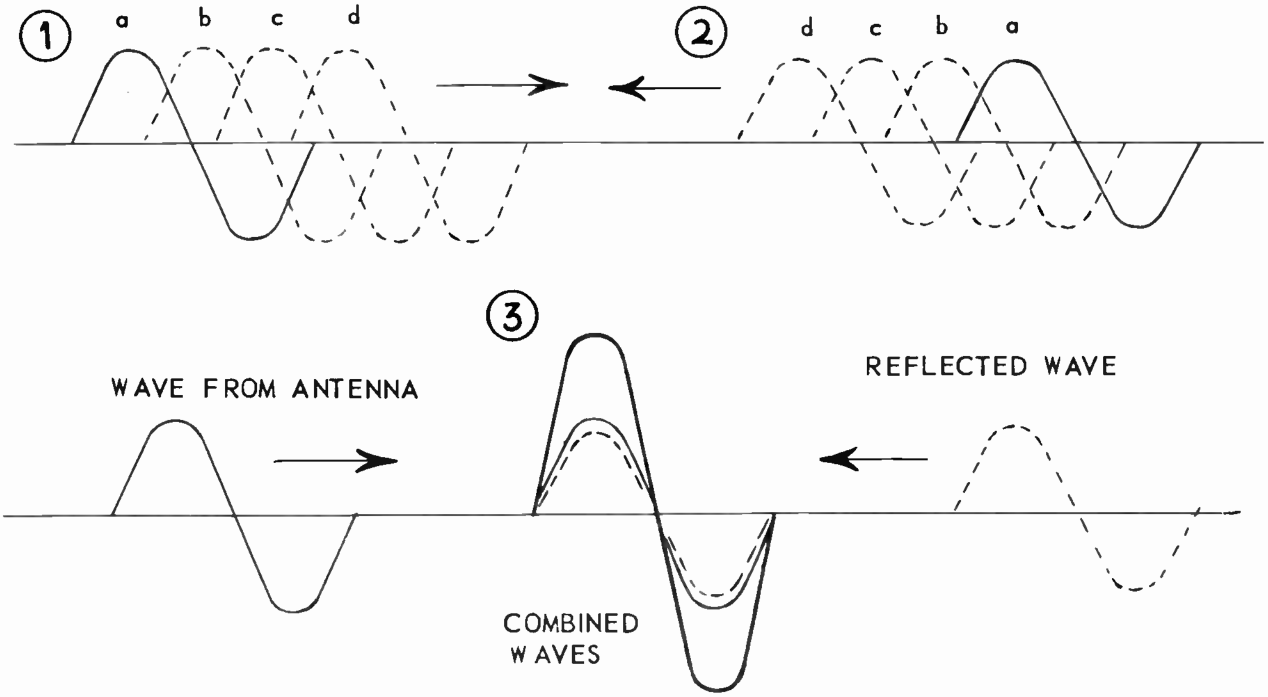


Fig. 68-15. Advancing and reflected waves combine their amplitudes where they come into phase.

frequency or band of frequencies. Such reactances cause reflections back into the line. The reflected waves combine with advancing waves to form what are called standing waves on the transmission line. These are waves that remain stationary at certain points on the line. How standing waves come into existence may be explained as follows.

When a wave leaves the antenna the wave travels progressively along the transmission line as at 1 in Fig. 68-15. At some one instant of time the wave has reached position a. A tiny fraction of a second later the wave will have gone on to b, then to c, and d, and so on until it reaches the receiver. A reflected wave, at 2, moves in the same progressive manner but in the opposite direction.

In diagram 3 are shown one wave advancing from the antenna toward the receiver and another wave which has been reflected. These two waves must meet at an intermediate position along the line. When the two waves get to this position, and are in phase with each other, their forces combine to produce a wave stronger than either the advancing or the reflected wave alone.

Of course, the oppositely traveling waves do not suddenly change from separate waves to the in-phase condition, they merge gradually. As they do so, the resulting peaks becomes higher and higher, about as illustrated at 1 in Fig. 68-16.

Waves start from the antenna at equal intervals, because each cycle is as long as every other cycle. All the waves, both advancing and reflected, travel at the same speed along the transmission line. All have to travel equal distances to reach any given point on the line, either before or after reflection. Consequently, advancing and reflected waves always will come into phase with each other at the same points along the transmission line. Just where these points will be located with reference to receiver and antenna terminals depends on frequency and on total length of line as measured in wavelength fractions.

Instead of showing the advancing and reflected waves as at 3 in Fig. 68-15 they might be shown in opposite polarity, as at 2 in Fig. 68-16, and results would be the same. Then, as at 3 in this latter figure, there will be a whole series of continually rising and falling waves at fixed points along the transmission line. These are standing waves. Although outgoing and reflected waves move continually along the transmission line, maximum amplitudes combine at only certain fixed positions.

The greater the excess of inductance or capacitance at the receiver input circuit the greater will be the reflection effect, the stronger will be the reflected waves, and the stronger will be the standing waves which are the sum of advancing and reflected amplitudes. Should the receiver input present a pure resistance, with no excess of either kind of reactance, there would be no reflections and no standing waves. The strength of the standing waves as compared with minimum amplitudes between these waves is a measure of the signal energy being wasted due to reflection.

The fact that standing waves actually exist under certain conditions, and that they do remain stationary on the transmission line is easily proven if the transmission line is an unshielded type. The proof happens to lie in the service technician's usual test for standing waves and reflection losses. The test is made as follows.

1. Turn down the contrast control, and the brightness control if necessary, to make pictures appear quite weak or dim.

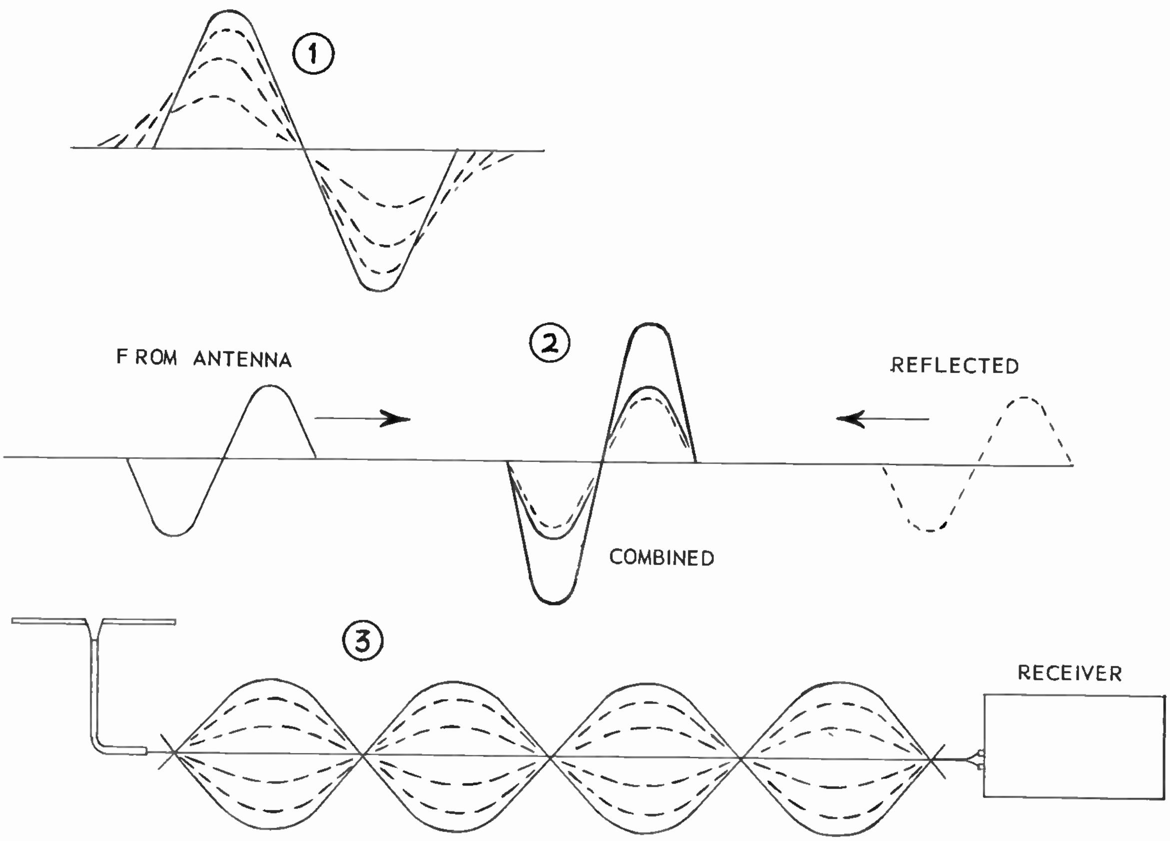


Fig. 68-16. The advancing and reflected waves form standing waves at fixed positions along the transmission line.

2. Starting close to the transmission line connection on the receiver, grasp the line firmly with one hand at successive points no more than four or five inches apart until you get about seven feet from the receiver. Keep your other hand off the line.

3. If the picture becomes very dim, or blacks out, or goes out of synchronization, there are reflections and standing waves. There is maximum amplitude of a standing wave at the point being grasped when the picture goes out as described.

MATCHING WITH RESISTORS. It is possible to match unequal impedances of antennas, transmission lines, and receivers by means of "matching pads" made up with resistors of suitable values. The connections of the type of pad most often employed for this purpose are shown by Fig. 68-17. In series with each side of the circuit is one of two series resistors, which are of equal resistance values. Across the circuit is a single shunt resistor, of different value in most cases. The end of the pad at which is the shunt resistor always is placed toward the lesser of the two impedances to be matched, and the end with the series resistors always is toward the greater impedance.

Suitable values for shunt and series resistors are listed in the accompanying table. Pads made with these resistance values allow both the source and the load to see resistances equal approximately to their own impedances. The listed resistance values are those conforming to preferred values, which are readily available from supply houses. Somewhat more exact matches might be made with different resistances, but the advantage would not be worth the trouble of obtaining the odd values.

PADS FOR IMPEDANCE MATCHING

Impedances To Be Matched, Ohms		One Shunt Resistance, Ohms	Each of Two Series Resistances, Ohms
150	300	220	110
95	300	120	120
72	300	82	130
52	300	56	130
95	150	160	43
72	150	100	56
52	150	68	62
72	95	150	22
52	72	91	20

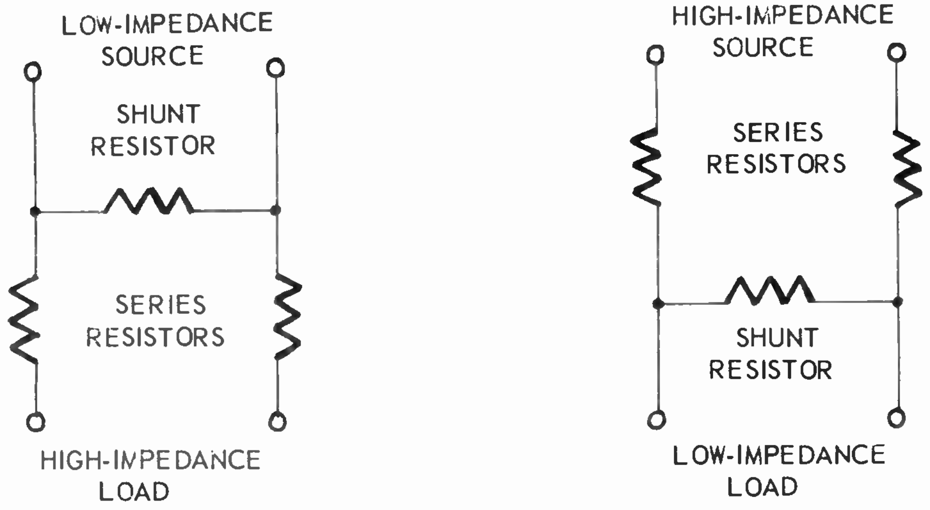


Fig. 68-17. Connections for resistors used as matching pads.

To learn how both the unequal impedances are matched by means of a pad you might write on either of the diagrams in Fig. 68-17 any of the combinations of unequal impedances and matching resistances given in the table. At the high impedance end of the pad a source or load sees resistance equal to the sum of the two series resistances plus the parallel resistance of the shunt resistor and the low impedance at the other end. At the low impedance end of the pad there is resistance equal to the shunt resistance paralleled with the sum of the two series resistances plus the high impedance at the other end. You will find that both the load and the source see resistances very nearly equal in ohms to their own impedances.

Fig. 68-18 shows some of the ways in which matching pads are used. Transmission lines may be of any style. At 1 a low impedance line is matched to a receiver of higher input impedance, and at 2 a higher impedance line is matched to a low impedance receiver. Note that the pads are reversed, end for end. At 3 a low impedance antenna is matched to a higher impedance transmission line, and at 4 a high impedance antenna is matched to a lower impedance line. Pads may be used at both ends of the line, as, for example, when a 300-ohm antenna is to be connected through a 72-ohm coaxial line to a 300-ohm receiver.

In diagram 5 a pad is used to match an unbalanced coaxial transmission line to the balanced input of a receiver. At 6 a pad is used between a shielded twin-conductor transmission line and a receiver having balanced input.

When making up your own matching pads use 1/4-watt non-inductive carbon resistors of 5 per cent tolerance. Cut the pigtails as short as allows soldering them to each other, to the line, and to lugs which will fit the terminals of the antenna or receiver. Pads used out of doors should be well coated with low-loss cement, such as that made from Polystyrene, or preferably wrapped with cellophane tape and then coated with cement.

Although resistance pads allow satisfactory matching of unequal impedances, the pad resistances introduce considerable loss of signal energy. The table lists pad resistances for matching 72 and 95 ohms and for matching 52 and 72 ohms. These mismatches are so slight that the loss due to resistance would be greater than the gain due to improved matching. Unless the degree of mismatch is great, or unless the carrier field strength is high, matching with resistance pads seldom is advisable.

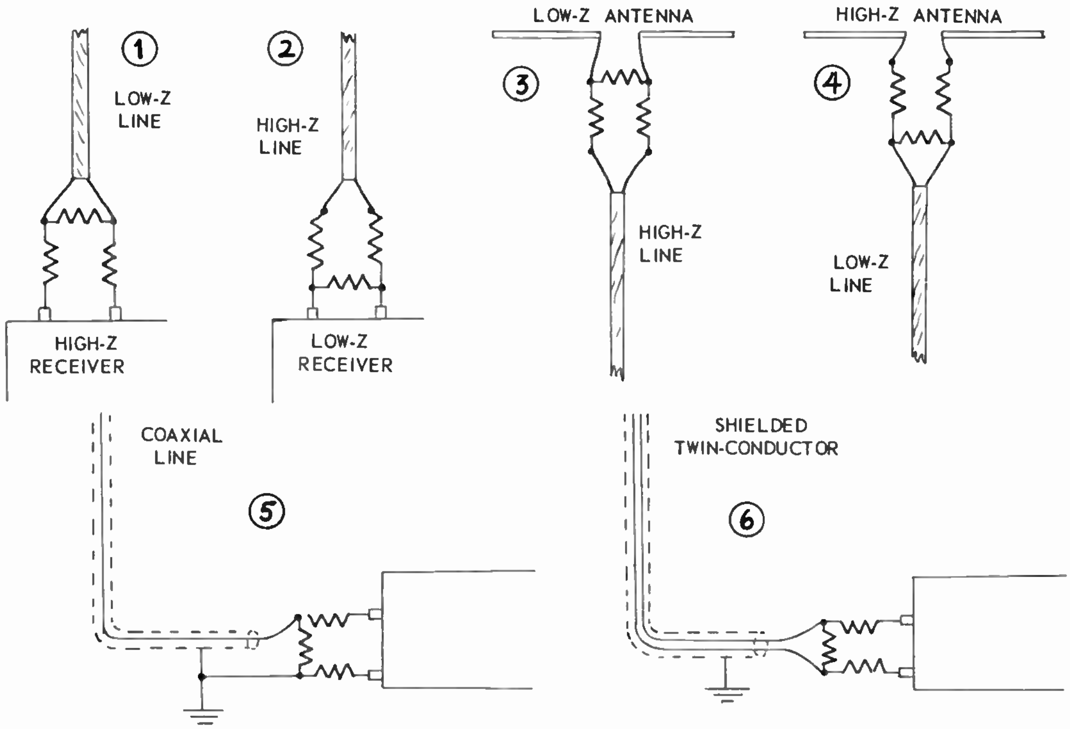


Fig. 68-18. How matching pads may be connected between transmission lines and receivers or antennas.

TELEVISION

LESSON NO. 69

ANTENNA INSTALLATION

Training Your Mind...

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind
C.T.I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON NO. 69

ANTENNA INSTALLATION

The signal gathering ability of the finest antenna may be largely wasted by putting it in a poor location, while a relatively poor antenna may work very well when erected in a favorable location. Whether you have picked a good location or a poor one can be finally decided only by tests of signal strength and picture quality, but you may be fairly sure of success if the following requirements are satisfied.

1. As far as possible from walls, poles, trees, and buildings, and especially from large metal objects whether or not any of these things will come directly between your antenna and the transmitters.

2. Away from electric signs and displays, and from electrical machinery and devices in general.

3. No closer than necessary to streets or roads carrying heavy automobile, bus, or motor truck traffic.

4. At least five feet above roof surfaces, rain gutters, parapet walls, and other structural parts of buildings which will be underneath the antenna.

5. At least 30 feet above the surrounding average ground level. The ratio of signal to noise is determined largely by antenna height.

6. If the mast below the antenna elements will be more than five to eight feet high it will require the added support of guy wires. Then the location must permit fastening the lower ends of these wires at least five feet out from the base of the mast in three directions at approximately equal spacings.

7. Having selected one or more locations which meet the foregoing requirements you should obtain permission from the building owner, preferably in writing, for installation of the antenna at one or any of these places. Do this before testing for signal strength and picture quality.

TESTING THE SIGNALS. Signal strength or picture quality may be checked with any of the following as an indicator.

The receiver which is to be used with the antenna.

A d-c microammeter and high-frequency rectifier connected to the video amplifier system of the receiver to be used with the antenna.

A commercial type of field strength meter or a portable receiver designed especially for this work.

A small television receiver made portable and otherwise adapted for signal tests.

Unless you expect to make antenna installation a part of your business the simplest and least costly method is to use the regular receiver for testing. The job then requires two persons, one to move and rotate the antenna while the other watches results at the receiver. The procedure is as follows.

1. Attach the antenna to a section of mast at least five or six feet long so that one person may move the antenna about and rotate it for orientation.

9 2. Connect the antenna terminals to the receiver input terminals with the same kind and with approximately the same length of transmission line which finally will be installed.

3. It will be necessary to communicate between receiver and antenna locations. If the distance is too great for talking or shouting you may enlist the help of a third person to relay the information, or you may arrange some system of signalling, or you may employ a pair of sound-powered phones connected through a long length of wire. Do not connect phones through the transmission line.

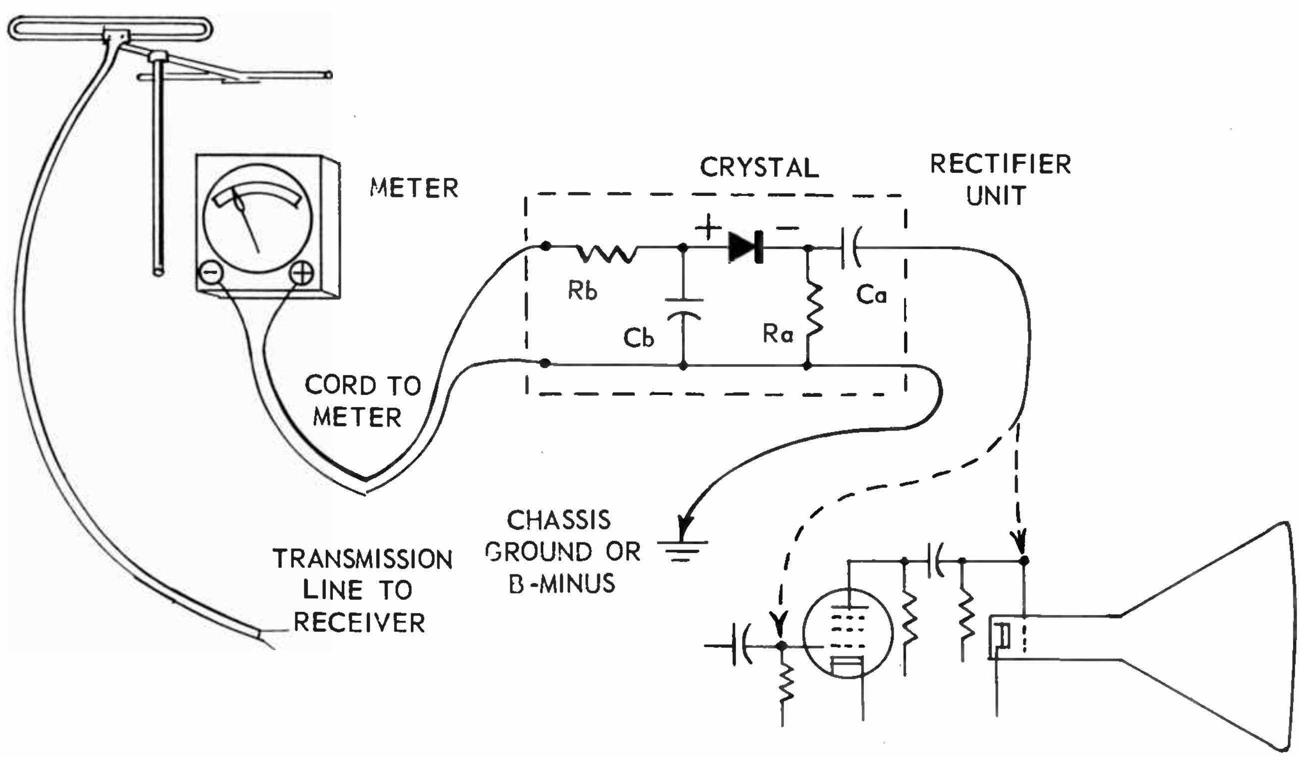


Fig. 69-1. Rectifier unit and d-c meter used for locating and orienting an antenna.

4. Tune the receiver to the weakest channel, if it is known. Set contrast and brightness controls as low as will allow a recognizable picture, so that any improvement may be apparent.

5. Hold the section of antenna mast in a vertical position with the antenna elements as high as possible while rotating the mast and antenna very, very slowly to allow noting changes of picture strength and quality at the receiver. Try all possible locations. Try different heights. If the antenna is of a style which can be tilted from the horizontal when permanently mounted, and if you are in a closely built up district, try tilting the antenna elements at various angles.

6. When you find a location and orientation giving best results on the weakest channel, keep the antenna in this position while the receiver is tuned and adjusted for each of the other active channels. A compromise orientation may be required to obtain satisfactory reception on all channels.

Now for the second kind of signal indicator. You can do a one-man job with the help of a d-c microammeter having a full-scale range no greater than 500 microamperes or a half milliampere and a rectifier unit including a crystal diode. The equipment and connections are shown by Fig. 69-1.

The rectifier unit, which remains at the receiver, may be built into any small non-metallic housing. The crystal may be a type 1N34, 1N48, or any generally similar germanium diode. Capacitors Ca and Cb are each 0.01 mf 400-volt paper tubular types. Resistor Ra is 100,000 ohms 1/4-watt. Resistor Rb, also a 1/4-watt size, protects the meter against overload. Start with about 50,000 ohms at Rb. If meter readings never go high enough try smaller values, if they tend to go off scale try more resistance.

The automatic gain control of the receiver should be overridden or made inoperative. The high-side clip of the rectifier unit is attached to the grid lead of any video amplifier or to the signal input element of the picture tube, which may be either the control grid or the cathode. The ground clip is attached to chassis ground or B-minus. The meter connection cord, long enough to reach from receiver to antenna location, may be any two-conductor type, such as lamp cord or transmission line. The microammeter is connected to the far end of this cord and taken to the antenna. To use this indicator proceed as follows.

1. Tune the receiver to the channel which is to be checked first. Set the contrast control about midway of its range and turn the brightness all the way down.

2. Take the meter to the antenna location and set or hold it where readings may be easily watched.

3. Move the antenna from place to place and orient it to obtain maximum meter reading, following the same general procedure as when one person watches at the receiver. If meter readings persist in going too high, reduce the contrast control at the receiver, and if they always are too low try turning the contrast control higher.

4. Temporarily support the antenna in the position giving maximum meter reading. Go to the receiver and disconnect the high side of the rectifier unit. Try all channels for picture strength and quality. If results are satisfactory the antenna should be erected where it is now placed. Otherwise tune the receiver to the poorest channel, reconnect the rectifier unit, and repeat the operations at the antenna.

Commercial types of field strength meters usually include a regular television tuner, one or two stages of video amplification, a video detector, and possibly a video amplifier stage. All these parts are fully shielded. The output of the detector or amplifier goes to a microammeter or to an electronic voltmeter circuit for indication of signal field strengths. There may or may not be a picture tube in the meter unit. There will be a complete power supply which is connected through a long power cord to any convenient outlet. An a-c voltmeter indicates line voltage.

The field strength meter is connected to the antenna terminals through a transmission line of the same kind and of about the same length as will be finally used for permanent installation. Meter readings or picture reproduction are observed on each active channel while the antenna is moved about and oriented in the usual way. The field strength meter may be calibrated to read microvolts of field strength. Most receivers are designed for normal inputs of 150 to 250 microvolts and for minimum useful signal of something like 50 to 100 microvolts.

Instead of purchasing a field strength meter some service specialists construct their own from a low-cost television receiver, usually a transformerless type to save weight. The picture tube may be either an electrostatic or magnetic deflection type. The loud speaker and all or part of the sound system are removed, again to save weight. The agc system should be disconnected and all grids returned to suitable fixed or manually adjustable biases. A d-c microammeter of 200 to 500 microampere full-scale range may be connected directly across the video detector load resistor. If the meter is connected to a video amplifier or the picture tube a rectifier is used in the same way as in Fig. 69-1.

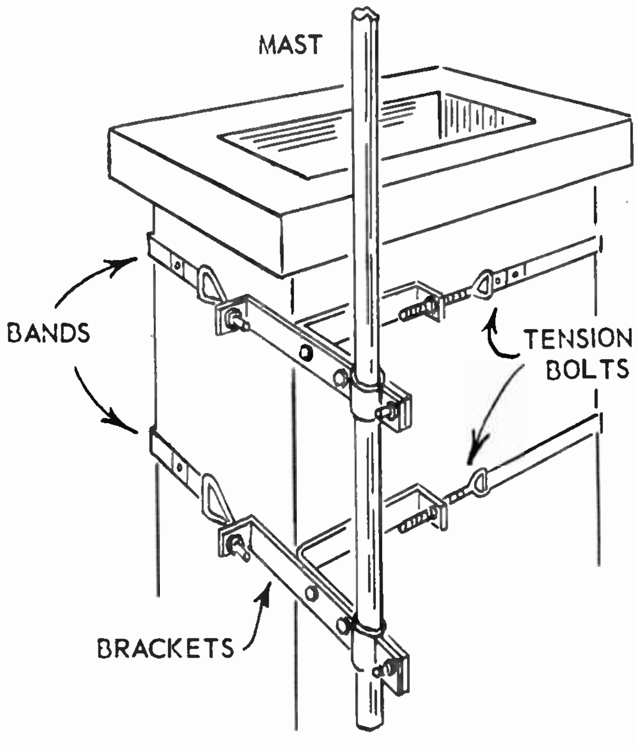


Fig. 69-2. Chimney mount for an antenna mast.

An a-c meter of 150-volt range may be added for observing line voltage. A deep hood around the picture tube screen will make outdoor observations much easier. The entire cabinet must be covered or lined with sheet aluminum connected to the receiver chassis or B-minus for shielding. This outfit is connected and used in the same way as a commercial field strength meter.

Unless a signal strength indicator is correctly calibrated to read field strength in microvolts at all carrier frequencies you cannot make dependable comparisons between one channel and others. You can check only for best location and orientation of the antenna for each individual channel. When using the regular receiver or one remodeled as a signal indicator you have no certainty that response is the same for equal field strengths at the different channel frequencies. You cannot positively determine that signals from one transmitter come in stronger or weaker than from another at a given location, for apparent differences usually are due to shortcomings in your measuring equipment.

ERECTING THE ANTENNA. All common types of antennas are available in kit form, complete with instructions necessary for assembly and erection. Masts and mountings for masts may or may not be included. Mast mounts purchased separately come with instructions for installation. All antenna supports and accessory hardware used out of doors should be rust proofed with heavy plating of cadmium, zinc, or some equivalent coating. Otherwise these things should be carefully painted after erection.

Masts usually are of either seamless or seamed steel tubing, but sometimes of lighter but more costly aluminum alloy. Standard lengths commonly are 5 feet or 10 feet, but may be anything between 4 and 12 feet. Thin wall electrical conduit often is used for antenna masts, or even standard weight galvanized steel pipe of the kind found in plumbing.

Mast diameters range from 3/4 inch to 2 inches. Light weight tubing in diameters up to 1 inch is used for antennas weighing no more than 10 to 12 pounds and not over 6 to 8 feet high. Greater weights and heights call for proportionately larger diameter tubing.

A mast support that is both sturdy and easily installed consists of a pair of chimney mounts of the general style shown in Fig. 69-2. There are many minor variations in design details, but all these brackets are held against a corner of the chimney by steel bands or straps drawn tight with tension bolts or turnbuckles. The two brackets should have greatest practicable vertical separation to make the mast rigid. Be sure to select mounts which extend the mast out far enough to clear the chimney cap or cornice. The antenna should be elevated several feet above the top of a chimney which is in use for smoke or fumes. A somewhat similar support for the mast is on a vent stack which extends above the roof of the building. One example is illustrated by Fig. 69-3.

Wall mounts of the general style illustrated by Fig. 69-4 are commonly used on the sides of buildings having extended roofs, as in a gable or any peaked construction. Brackets of standard size hold the mast at 8 to 18 inches from the wall. Types having extension arms allow clearances as great as 30 inches or more from the wall surface. One or both brackets should have a third leg that slopes from the mast end down and inward to the wall, in order to provide vertical support for the weight of mast and antenna. Also available are brackets which hold the mast only 2 or 3 inches from either a flat surface or a corner. These are used for either wall or chimney mounts.

If a wall is of all wooden construction any brackets may be fastened to it with lag screws provided you

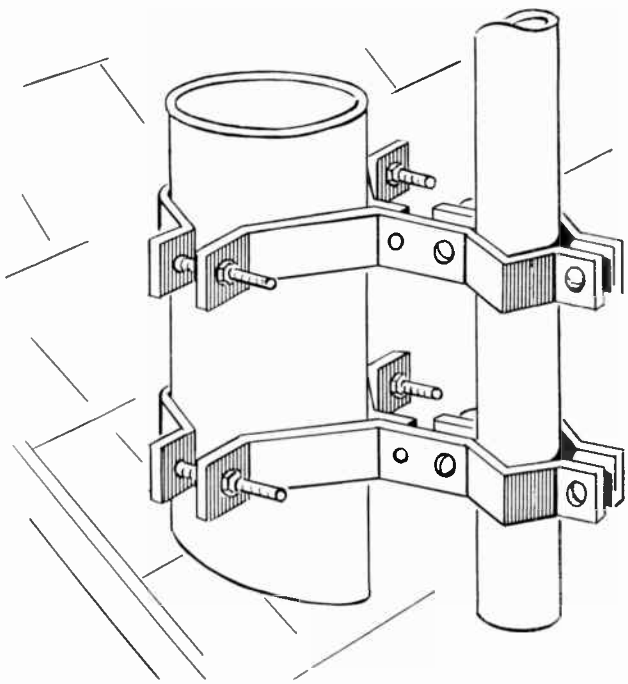


Fig. 69-3. Vent pipe mounting for a mast.

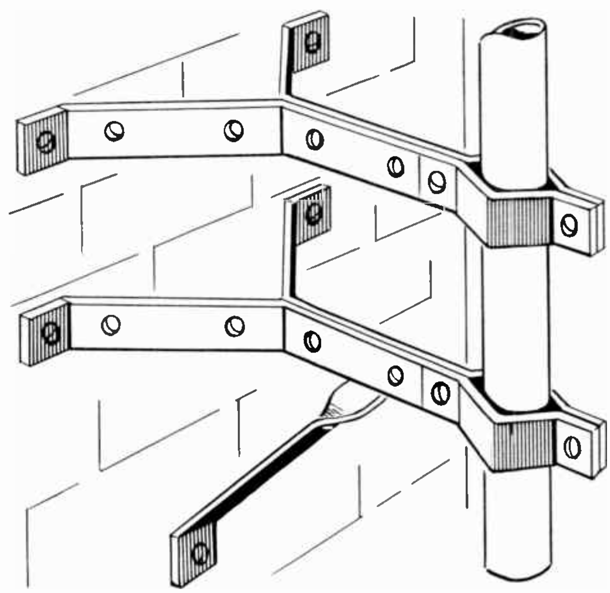


Fig. 69-4. Wall brackets supporting an antenna mast.

can be sure that the inner ends of the screws pass into a stud or a header, not merely through the sheathing. Carriage bolts with large washers may be used when you can reach the inside of the wall to put on the nuts. Various items useful in making wall mountings are pictured by Fig. 69-5.

Any kind of masonry wall construction should be pierced part way or all the way through by using a star drill, which is somewhat like a long cold chisel but with two cutting edges at right angles. A hole is made by hammering on the drill while continually rotating it. Expansion sleeves which take a lag screw may be inserted in a hole drilled only 2 to 3 inches deep in masonry. Turning the screw into place expands the split sleeve so that it grips securely in the hole. A less satisfactory way is to drive wooden plugs into the drilled holes, then use lag screws in the wood.

If holes are drilled all the way through masonry the brackets may be held with long carriage bolts or machine bolts. The outdoor ends of these holes should be made weather tight with caulking compound or mastic. Such compounds, also all kinds of screws, bolts, other hardware, and necessary tools are obtainable from radio supply houses or from hardware stores.

The bottom of the mast often may be supported on a flat or pitched roof surface or on the ridge of a peaked roof. There are many styles of base mounts which allow the mast to be vertical when the supporting surface has any degree of pitch. Two kinds are illustrated by Fig. 69-6.

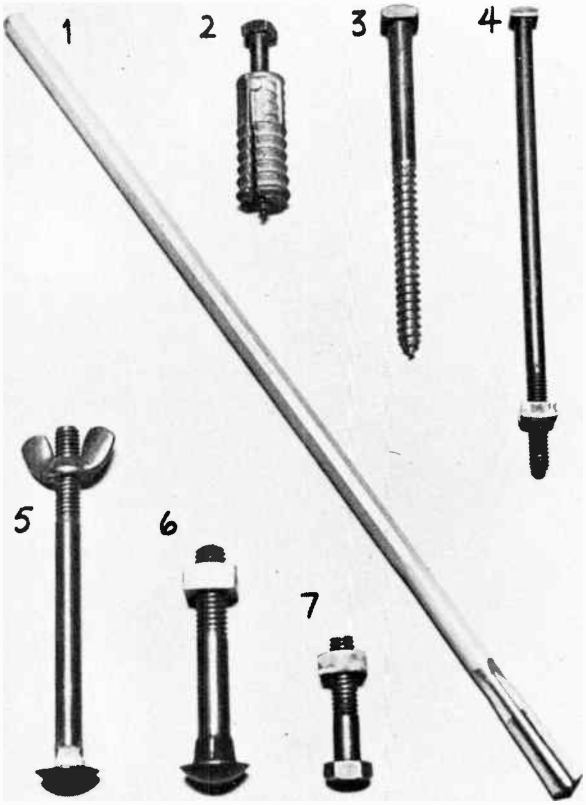


Fig. 69-5. Items used when mounting antennas.
1, star drill. 2, expansion sleeve with lag screw. 3, lag screw. 4, carriage bolt with wing nut. 5, carriage bolt with square nut. 6, cap screw with hexagon nut.

Base mounts may be fastened with large wood screws or with lag screws to surfaces covered with roll roofing, composition shingles, wood shingles, or metal. Do not attempt putting such mounts on tile or slate roofs. When the base mount is held on the supporting surface with any type of screws or bolts be sure to apply a liberal coating of roofing cement over the screw heads and around the edges of the brackets.

If a roof is flat it is advisable to attach the mounting bracket to a piece of board a foot or more square, then simply lay the board on the roof where it will be held by friction. If the roof has only a slight pitch you can drive four or more nails through the board so that the points protrude underneath no more than 1/8 to 3/16 inch, to catch on the roof surface. Any board must be well painted all over to prevent warping and rotting with moisture.

Puncturing any kind of roofing is more than likely to void the guarantee on that roof, even though you make no holes all the way through the covering. You may be called upon to pay damages unless permission has been obtained from the building owner.

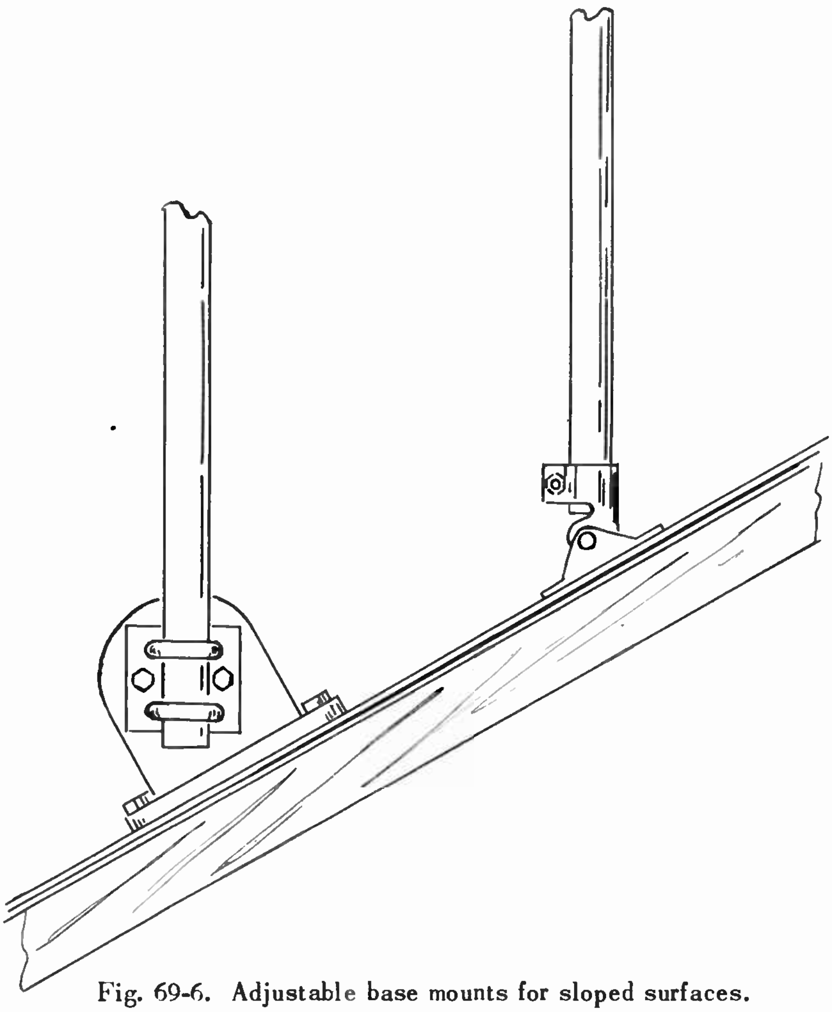


Fig. 69-6. Adjustable base mounts for sloped surfaces.

When a mast must be higher than allowed by a single length of tubing any number of pieces may be fastened together by methods such as illustrated in Fig. 69-7. At 1 a few inches on the end of one length of tubing are contracted to smaller diameter so that this end may be forced into the end of another length to make a telescoped joint. When telescoped tubing is of the seamed variety the seams in the two lengths should be in line. One telescoped joint in a mast works well, but two or more in the one mast tend to allow tilting or bending, since the fits cannot be made very tight and true.

The ends of upper and lower sections may be butted together as at 2 and held in place with a split tubular clamp which is tightened by means of screws and nuts. Clamps of this kind may have openings at one end for attachment of guy wires. Two lengths of mast tubing may be held side by side with two clamps separated vertically by 8 inches or more. One such clamp and the manner of using it is illustrated at 3. Lengths of tubing may be held side by side also with two or more U-bolts long enough to pass over both diameters with the threaded ends extending through a cross bar that forms a bearing for the clamping nuts.

When a mast is carried at its base by mounts of the general type shown by Fig. 69-6, guy wires must be used. If the bracket supports clamp the mast at two points separated vertically by 8 inches or more, guying may not be necessary when the topmost part of the antenna is no more than 6 to 8 feet above the highest clamp.

As a general rule, any mast whose length is 10 to 20 feet below the antenna elements must be supported with one set of guy wires, and for every additional 10 feet of length there should be one additional set of guys. This is necessary because of stresses due to wind, snow, and ice which might collapse an unguyed structure, also to prevent slight swaying which may cause fluttering and loss of synchronization in pictures.

As shown by Fig. 69-8 there should be three guy wires in each set or group, spaced as equally as possible around the bottom of the mast, and coming far enough from the bottom that the angle between wires and mast is not less than 30 degrees, as shown by full lines. It is preferable to make the angle about 45 degrees, as shown by the single broken line.

If the upper ends of the guy wires must come within 8 feet or less from the lowermost antenna element

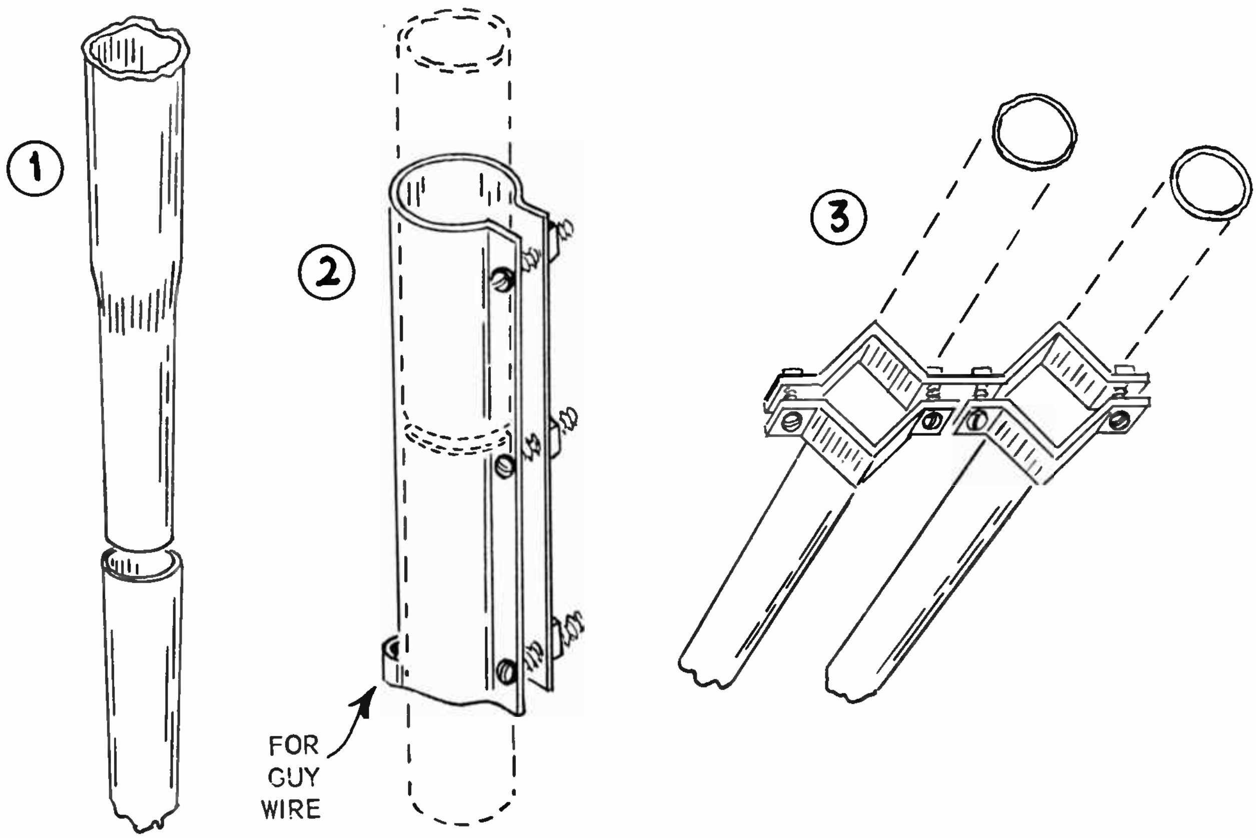


Fig. 69-7. Methods for supporting mast extensions.

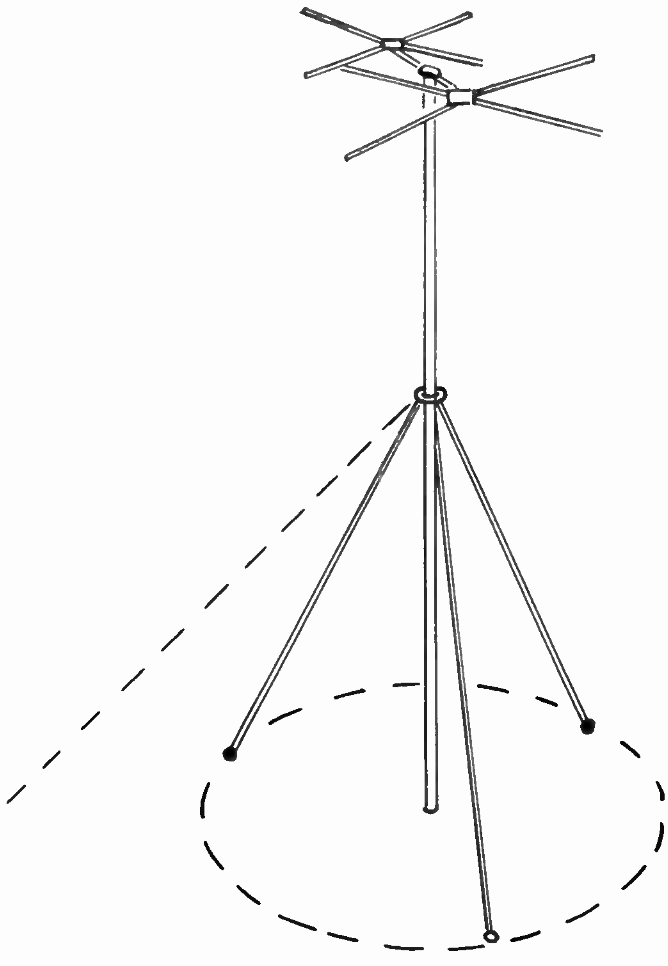


Fig. 69-8. Positions for guy wires on a mast.

the ends of the guys toward the antenna should be made up with two or three 2-foot lengths joined with strain insulators such as used for radio broadcast antennas. This is to prevent wave reflections from the guy wires, and possibility of smears in the pictures.

Wire for guying usually is stranded galvanized steel with six or seven strands, each of number 18 or 20 gage size. Another choice is stranded bronze radio antenna wire, which does not rust. Steel baling wire may be used for guying small antennas. Some guy wire fittings are pictured by Fig. 69-9. At the bottom is a heavy-duty ring which clamps onto the mast and provides attachment points for three guy wires.

The lower ends of guy wires are held by large screw eyes, screw hooks, ring bolts, or equivalent devices. The upper ends may be fastened to extension clamps such as shown at 2 and 3 of Fig. 69-7 or to any of the many available kinds of guy wire rings and clamps. Wires less than 10 feet long are unlikely to stretch in any great degree and may be fastened by twisting the free end back around the span a number of

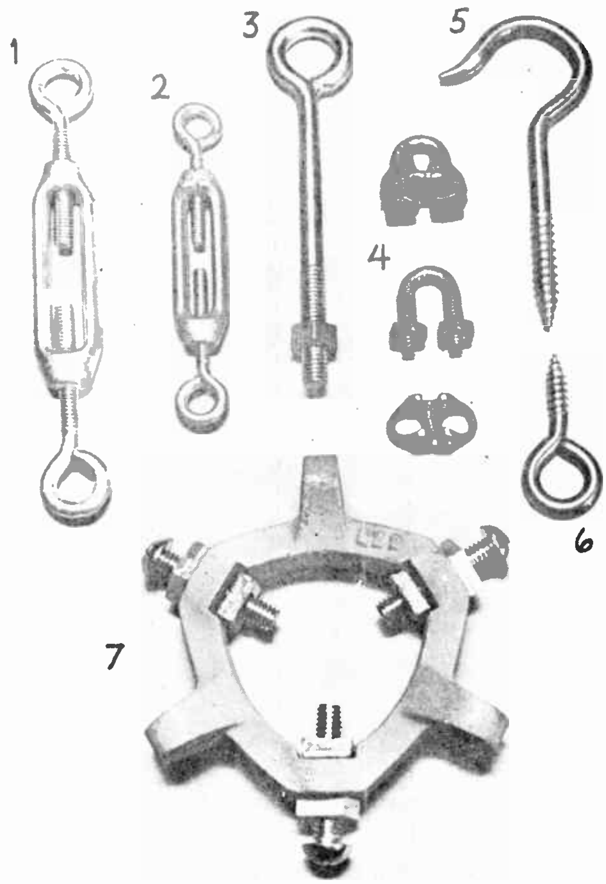


Fig. 69-9. Guy wire fittings.
1 and 2, turnbuckles. 3, ring bolt. 4, cable clamp assembled and separated. 5, screw hook. 6, screw eye. 7, guy wire clamp.

times. It is, however, much better practice to use regular U-bolt clamps or small cable clamps made for this purpose. Longer guy wires should each be fitted with a turnbuckle to take up any slack. These come in lengths of 3 to 8 inches as measured when fully closed. The longer the wire the longer should be the turnbuckle used.

TRANSMISSION LINE INSTALLATION. Before commencing permanent installation of the transmission line pick out the shortest path, to minimize attenuation and interference pickup. Some of the principles to be observed are illustrated by Fig. 69-10. The line should be held on the mast with a standoff insulator within a foot or two of the antenna terminals. Additional standoffs should be placed every 6 to 8 feet all along the line, or wherever needed to prevent swaying or to carry the line around corners. Avoid sharp bends, make the radius an inch or more.

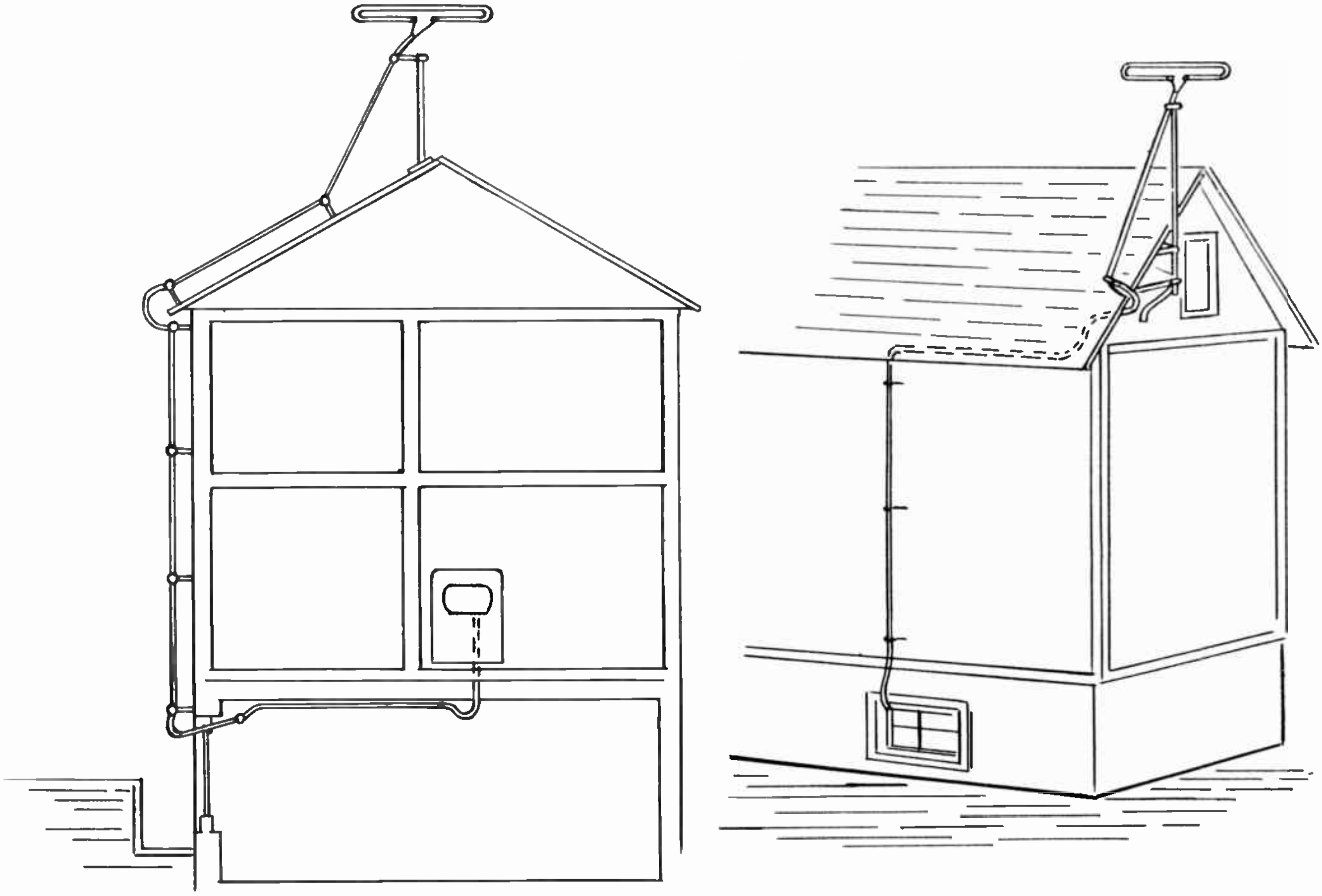


Fig. 69-10. How transmission lines may be run from the antenna.

Where the line must cross the edge of a roof or any other protrusion, use standoffs which will hold the line away from all these things. Fig. 69-11 shows a few of the many types of standoff insulators which may be used. Some have clamps for fastening to the mast. Others have wood screw tips. Still others are designed for driving, like nails, into either wood or the mortar joints of masonry.

In most standoffs the opening through the insulation is designed to take either flat unshielded line or the smaller sizes of shielded line. Otherwise the insulators may have only a plain round opening for large diameter coaxial cable. The metal rings around the plastic insulators have open slots which admit the line without having to thread it through for the entire length. After the runs are completed these metal rings may be squeezed tight with pliers to securely clamp the line against endwise slipping. Be sure to do this on all vertical runs.

The antenna ends of both conductors in the transmission line should be fitted with soldered lugs which will attach to the terminals of the antenna elements. If you do not have lugs, loop the conductor ends to

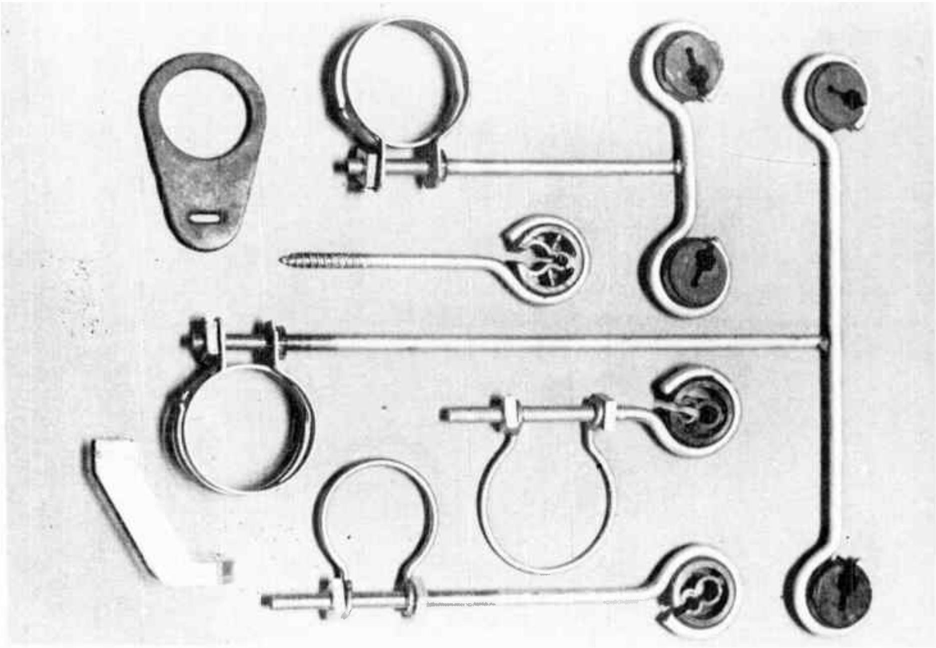


Fig. 69-11. Various styles of single and double standoffs with insulators.

fit the terminals and fill the strands with solder. Make clean, tight joints and coat them with low-loss cement or lacquer. Do not draw the line so taut at the antenna as to strain the terminal connections.

Where the line is brought indoors through a window casing or door casing, or through any kind of wall, make the outdoor end of the hole an inch or more lower than the indoor end, to keep water from entering. Unshielded line should be protected with a porcelain tube or circular loom where it goes through the hole. Do not squeeze this type of line to change the conductor spacing. Make the outdoor end of the hole weather tight with mastic or caulking compound.

The following instructions apply especially to unshielded transmission line.

Keep the line at least two feet from all large metal objects, including rain gutters, downspouts, vent pipes, conduits, water or gas piping, and metal roofs. Never run unshielded line through metal conduit or metal tubing of any kind. Do not run the line close to and parallel with the antenna mast any farther than necessary. From the first standoff, close to the antenna terminals, it is desirable to bring the line away from the mast, but not at an angle greater than 45 degrees. If the line must come down the mast, use long standoffs which keep the line at least four inches from the mast.

Make horizontal runs as few and as short as possible, chiefly because rain, snow, and sleet remain on such runs to temporarily alter the line impedance, and also because horizontal runs tend to pick up interference. When possible place any horizontal runs under the eaves or where they are otherwise protected. Any part of the line parallel to a roof surface should be supported high enough to remain above snow levels.

Twist unshielded line one turn about every foot of its entire length from antenna to building entrance. This helps prevent interference pickup. Thin ribbon line may be brought indoors over the sill or over the top of the sash of a closed window provided there is no metal weather stripping and no metal in the sash or frame. Never use window lead-in strips such as employed with radio antennas.

Unshielded line may be protected from dielectric effects of surface moisture with a thin coat of silicone compound, which does not wash off. A substitute, not so good, is a coating of automobile polishing wax. Either kind of coating causes the moisture to gather into small drops instead of remaining as a uniformly distributed film. Do not paint unshielded line, especially with any paint having lead or other metallic base.

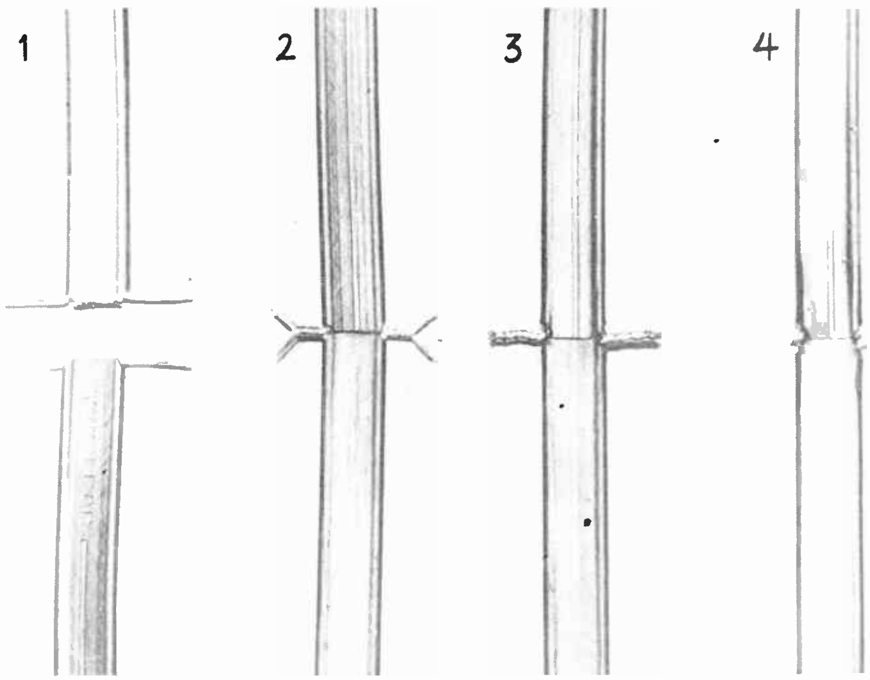


Fig. 69-12. Splicing an unshielded twin-conductor transmission line.

Splices in outdoor transmission line should be avoided, the line should be continuous from antenna to building entrance, and preferably all the way to the receiver. If an outdoor splice is necessary, make it as shown by Fig. 69-12.

1. Remove the plastic insulation to bare 3/8 to 1/2 inch of the conductors. Cut straight and true across the center web. Bend the bared wires outward.

2. Twist the opposite pairs of wire ends together until the center webs of insulation are drawn together.

3. Solder the two joints. Hold your iron near the free ends of the wires and let solder flow all the way through the joints. Try not to melt the insulation.

4. Clip off the protruding wire ends as close as will not unduly weaken the joints. Securely clamped, standoff insulators should be placed within a few inches of the splice on both sides, to prevent the soldered joints from being stressed.

Indoor runs of unshielded transmission line must be kept clear of radiators, pipes, and other large metal objects whether they are exposed or concealed in walls or floors. The line may be fastened along moldings, baseboards, or similar interior trim with small headed tacks, preferably with special insulated tacks made for this purpose. Small plastic standoffs instead of tacks are desirable if their appearance is not objected to. Unshielded line should not be tucked behind moldings, nor run for more than five or six feet underneath carpets or rugs, and in no case should it be painted.

At the receiver end of the line leave only enough slack to allow moving the receiver for cleaning. Splices in indoor sections of line may be made with small connectors designed for such use. These connectors have bodies of low-loss plastic with pairs of push pins correctly spaced. The bared ends of the line conductors are held in the connectors with small set screws.

5 Installation of shielded transmission line is relatively simple, for this kind of line may be run practically anywhere. It may be clamped tight to the antenna mast, run down inside the mast, or inside of any kind of metal conduit, tubing, or pipe. It is unnecessary to avoid metal objects of any kind except steam pipes or other surfaces which may be heated. Shielded line may be run through wall spaces and may be laid along interior masonry walls.

The central conductor of coaxial transmission line may be connected to either terminal of the antenna, with the braid conductor to the other terminal. Preparation of the connections is illustrated by Fig. 69-13.

1. Remove about 3 inches of the outer protective covering, or enough so that the conductor ends will reach the antenna terminals when spread apart.

2. With a blunt pointed tool loosen the braid around the free end and push the braid away from the polyethylene insulation and as far back as possible. With the blunt tool work an opening into one side of the braid close to the remaining outer covering. Make this opening somewhat larger than the polyethylene insulation, but without unnecessary breakage of braid strands.

3. Bend the insulated inner conductor at the opening through the braid. Work the insulated conductor

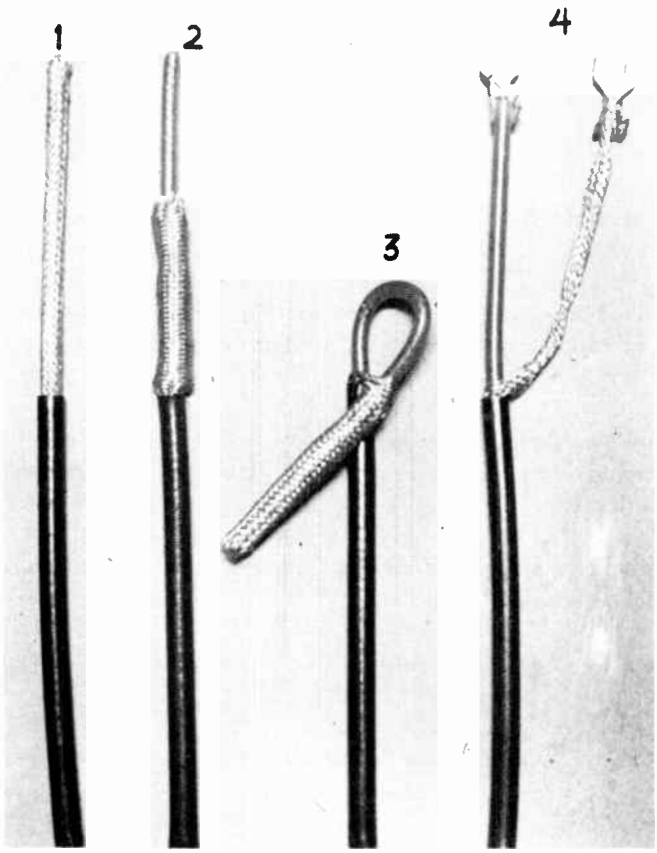


Fig. 69-13. Making terminal connections on coaxial cable.

out through the opening while working the braid over the polyethylene insulation until you get the insulated inner conductor out through the opening in the braid.

4. Pull the braid out to its full length and twist it tightly. To the free end of the braid solder a lug which will fit an antenna terminal, then fill the remainder of the braid with solder to stiffen it. Bare enough of the inner conductor to take a similar soldered lug for the other antenna terminal. Note that the picture shows suitable lugs in place, but not yet soldered. The lugs must clamp securely over the braid and over the polyethylene insulation to withstand pulling.

The other end of the coaxial line may be similarly prepared for connection to the input terminals of the receiver.

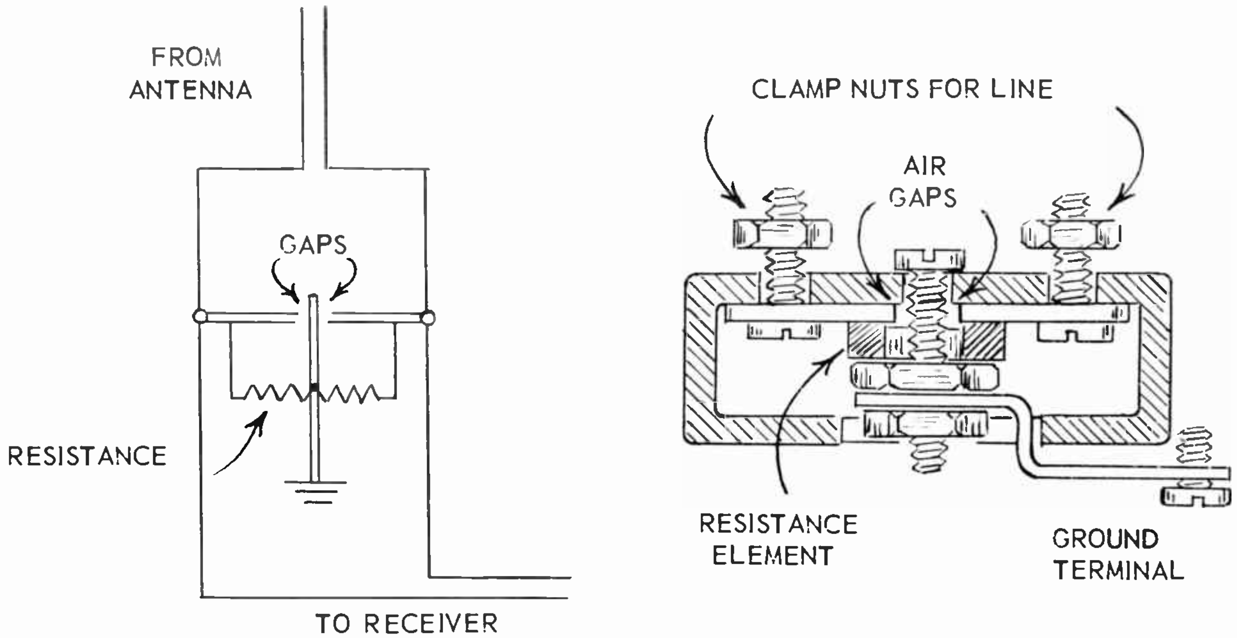


Fig. 69-14. The principle and some construction details of a lightning arrestor.

LIGHTNING PROTECTION. The combination of a metal mast and an elevated antenna connected through a transmission line to an indoor receiver forms a dangerous lightning and fire hazard unless suitable protective measures are taken. These measures include making a direct low resistance connection from the mast to ground, and using an Underwriters' approved lightning arrester on the transmission line. Then the mast becomes an effective lightning rod to protect the building.

The principle of a television lightning arrester is illustrated at the left in Fig. 69-14, and some typical construction details are shown at the right. Each of the transmission line conductors is connected to a piece of metal whose inner end is spaced from a grounded internal member by a narrow air gap. Should lightning strike the antenna elements the charge will jump these gaps and go to ground rather than on through the line, since lightning takes the shortest path to earth.

In the particular design illustrated, each of the line conductors connects also through a resistance element to the grounded conductor. Electrostatic charges which may build up to rather high voltage on a fully insulated antenna will pass off through the resistance element. The resistance between each line conductor and ground is one or two megohms, and between the two sides of the line it is two to four megohms, a value much too great to allow appreciable signal leakage.

In some lighnting arresters designed for attachment to the antenna mast the grounded member, which may be a clamp, makes firm electrical connection with the mast. Then a single grounding wire from the arrester grounds the mast while providing lightning protection for antenna and line. An arrangement of this kind is shown at the left in Fig. 69-15.

3 A lightning arrester to which only the transmission line is connected may be mounted anywhere along the line between antenna and building entrance, but usually is at the building entrance as at the right in Fig. 69-15. A grounding wire then is run from the arrester to a connection into the earth, and an extension or a separate grounding wire is run to a clamp on the antenna mast. No matter where the lightning arrester is mounted, the unbroken grounding wire should extend from the mast to the earth connection with no intervening gap.

The earth end of the grounding wire must be securely clamped or soldered to a low resistance conductor that extends unbroken into permanently moist earth, which means earth three feet or more below ground level. Cold water pipes on the street side of any water meter are preferred, although any cold water pipe or any fixture on such a pipe, as an outdoor hose bib for example, usually is satisfactory. Do not make ground connections to pipes carrying gas, oil, hot water, or steam—they do not extend unbroken into the earth. An effective ground connection may be made with a regular steel grounding rod four feet or more in length and preferably coated with copper or zinc. The rod is driven into the earth at any convenient point.

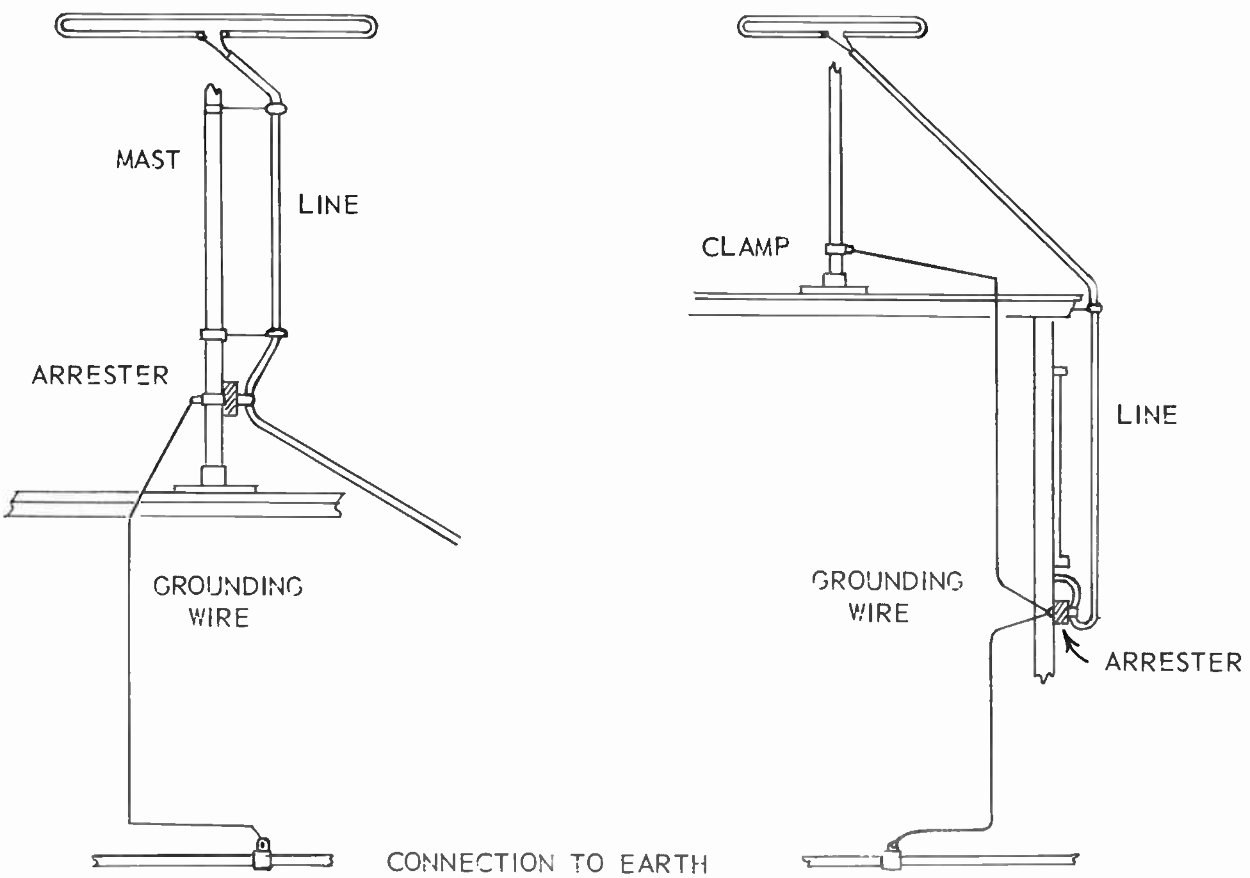


Fig. 69-15. Connections for lightning arresters and grounding wires.

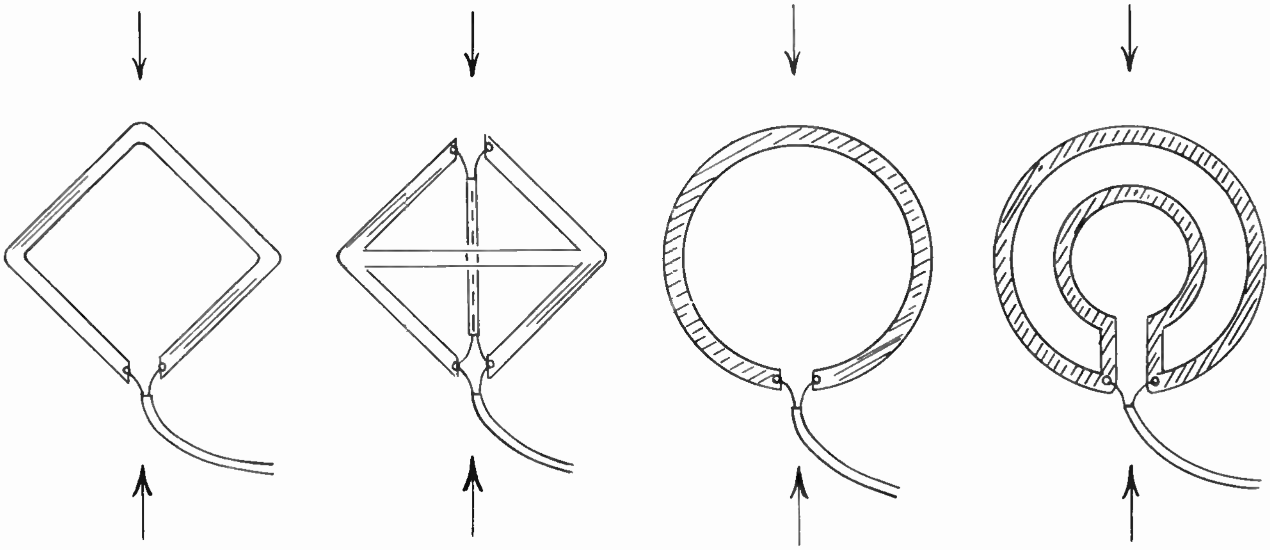


Fig. 69-16. Forms of built-in antenna elements.

BUILT-IN ANTENNAS. Current models of the majority of receivers are equipped with built-in antennas which, when conditions are favorable, allow reception at distances usually up to six or eight miles from transmitters. Provision always is made for disconnecting the built-in antenna and substituting an outdoor type when necessary. Since a built-in antenna is wholly within the receiver cabinet or console, commonly tacked or stapled in place, the antenna location depends on where the set is used. The higher the location above surrounding buildings the better are the chances for good reception. Also, the fewer the walls between the receiver and outdoors the more satisfactory will be the signals. Reception is almost impossible in basement apartments below ground level.

Signal attenuation and wave reflections may be bad in buildings having steel frames or beams. Results are better in structures with wood framing, with brick or stone veneer or stucco on wood framing, and with solid masonry construction. If, however, these latter buildings have metal lath, a metal roof, metal window or door frames, or have metal foil thermal insulation, there are likely to be difficulties. There may be trouble also if the receiver with its built-in antenna is placed close to exposed or concealed pipes for water, steam, or gas, or to long metal curtain rods, or metal venetian blinds, or if there is any kind of radiant heating system in walls, floors or ceilings.

Electrical interference often is troublesome. It arises from operation of home appliances, especially when they are not in good order, and from defective switches or connections in lighting and power circuits. Distances and the number of localities in which built-in antennas are satisfactory are being extended as television transmitters are allowed to operate with increased carrier powers.

In addition to the difficulties due to receiver location, the limited cabinet space imposes a number of serious design problems. First, of course, is the matter of obtaining sufficient signal pickup or gain. High gain requires sharply peaked resonant curves and limited frequency response. When frequency response is limited it becomes necessary to tune or resonant the antenna for each channel or, at least, for each band of frequencies. Another difficulty is that of orientation. Built-in antennas, like all other television antennas, are basically half-wave dipoles horizontally polarized. As such they have maximum signal pickup from two opposite directions and have minimum pickup from directions at right angles.

However the orientation may be handled, it is desirable to have the widest possible directional pattern. Such a pattern may be attained by making the antenna in the form of a single or double square, rectangle, circle, or oval, and mounting it so that all the elements lie in a horizontal plane, as when attached to the inside top or bottom of the cabinet. Such forms are shown by Fig. 69-16. All have maximum response in the directions of the arrows, there may be reasonably strong response through wide angles on both side, but always there is a minimum at exact right angles.

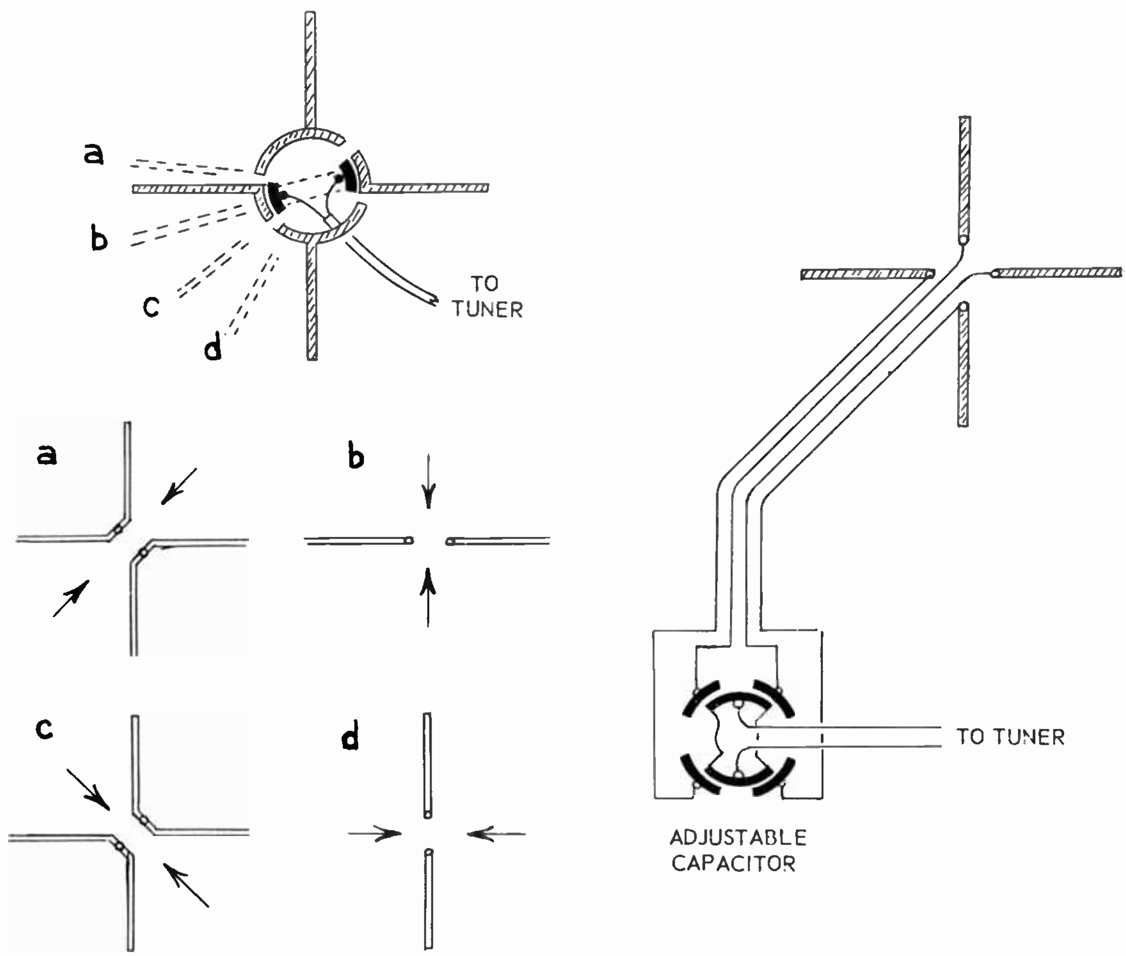


Fig. 69-17. Electrical orientation of built-in antennas.

The simplest solution of the orientation problem is to fasten the antenna in a fixed position and leave it to the operator to turn the entire receiver with its cabinet into a position that allows reception. Probably the least simple solution, mechanically, is actual rotation of the entire antenna structure inside the cabinet by means of an external control knob. It is possible to secure very nearly the same results electrically.

One method of electrical orientation is shown in principle at the left in Fig. 69-17. Two dipoles are at right angles to each other, with the inner ends of their conductors attached to stationary segments of a rotary switch. Two contacts on the switch rotor are connected to the line that runs to the tuner. Turning the rotor by means of a lever serves to connect the four antenna conductors in any of the combinations shown below the switch diagram. Arrows indicate directions of maximum response for each combination.

The principle of a different kind of electrical orientation is illustrated at the right. Again there are two

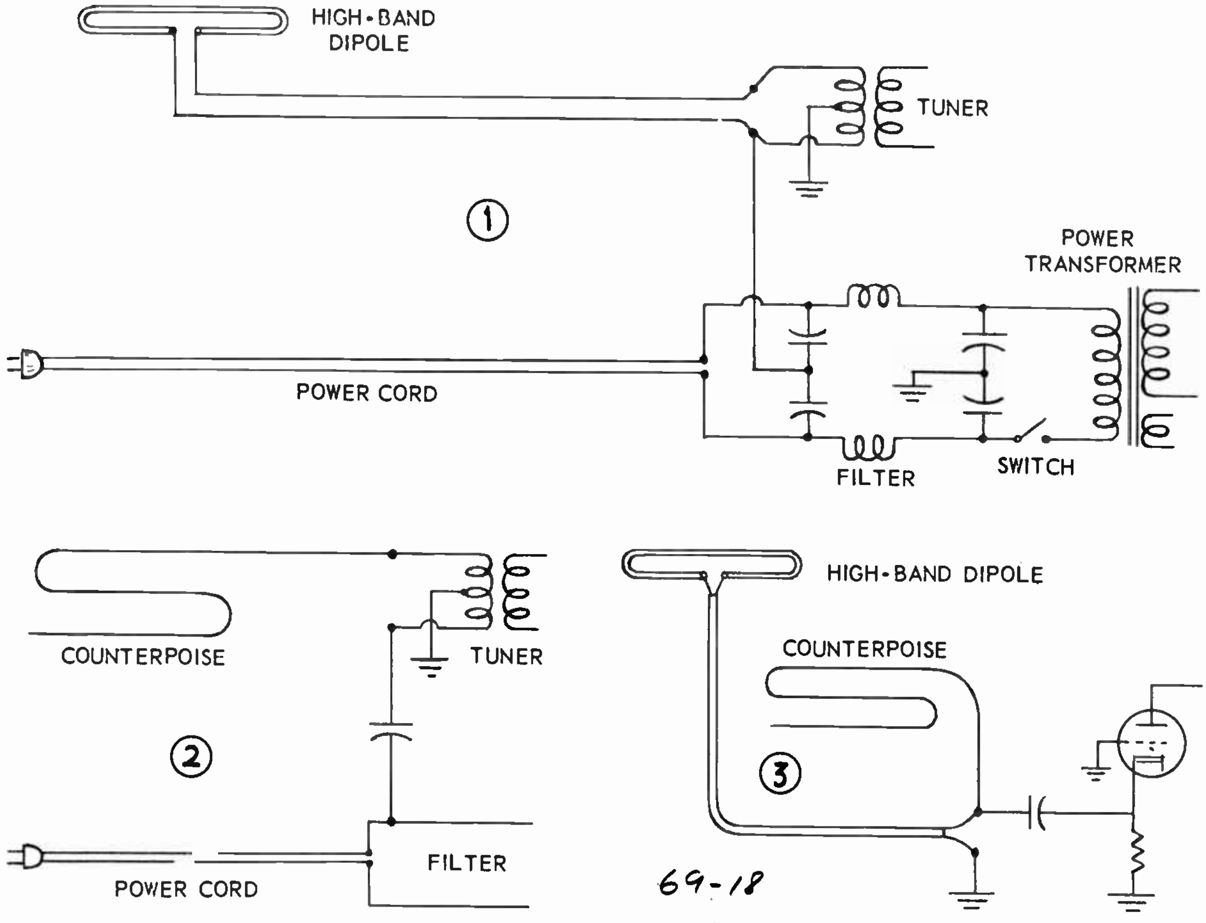


Fig. 69-18. Power cords and counterpoises used for, or in connection with, built-in antennas.

dipoles at right angles, but connection to the tuner line here is through the capacitance existing between four stationary plates connected to the antenna elements and two rotor plates carrying the twin conductors for the tuner. Which pairs of antenna elements are capacitively coupled to the tuner depends on the position to which the rotor plates are turned by means of a control knob. The several possible combinations of elements and the variable capacitive couplings allow maximum response from various compass directions.

The problem of tuning or resonating a built-in antenna is simplified by two facts. First, low-band signals of given field strength are easier to pick up than high-band signals of equal strength. Second, a high-band antenna is small enough to fit into small cabinet space. As a consequence, most built-in antennas are primarily high-band types, usually some modification of a straight or folded dipole.

^ Many receivers employ all or part of the built-in antenna connections shown at 1 of Fig. 69-18. Here there is a small dipole for high-band reception. The principal pickup for low-band signals is through the power cord, from which connection to the tuner is made through the two capacitors shown or, in many cases, through a single capacitor from only one side of the power circuit. These coupling capacitors may be as small as 30 mmf or larger than 1,000 mmf.

Between the signal takeoff point and the primary of the power transformer is a filter consisting of r-f chokes and bypass capacitors to ground. The capacitors usually are of 0.01 mf size. Sometimes there is only one bypass capacitor, and in a few cases there will be only a single r-f choke and one bypass in the filter. When a power cord is used as an antenna, the cord should be kept away from large metallic objects, should not be kinked, and may be spread out or moved about to obtain best reception.

As at 2 in Fig. 69-18, a counterpoise wire may be used instead of the high-band dipole in conjunction with a power cord antenna. The counterpoise is a long wire tacked or stapled inside the cabinet. In other cases, illustrated by diagram 3, a similar counterpoise may be used to provide improved low-band pickup when a small dipole is used for the high band.

Many built-in antennas utilize the principles of tunable stubs and impedance matching stubs. Some examples are shown by Fig. 69-19. In examining these devices it should be remembered that a resonant half-wave dipole acts like a series resonant circuit, having excess capacitance at frequencies below resonance, and excess inductive reactance at frequencies above resonance. To tune such an antenna to a lower frequency it is possible to add inductive reactance, or for higher frequency to add capacitive reactance. Variable capacitance for channel tuning usually is provided by an adjustable capacitor, while inductance may be added by small coils or suitable stubs.

These built-in antennas consist of a short dipole whose naturally high resonant frequency is reduced by end loading, which refers to almost any means for increasing the capacitance by enlarging the outer ends of the dipole conductors. The conductors may be round, but more often are sheet metal or metal foil. Tuning capacitors have minimum capacitances of 2 to 4 mmf and maximums of 30 to 70 mmf or even more.

Control knobs or handles for tuning or orientation may be above or below the picture tube on the front of the cabinet, or on the back, one side, or top of the cabinet. Tune the antenna after adjusting the contrast and fine tuning controls. Do this without getting your arm or body closer than necessary to any part of the cabinet inside of which the antenna is mounted.

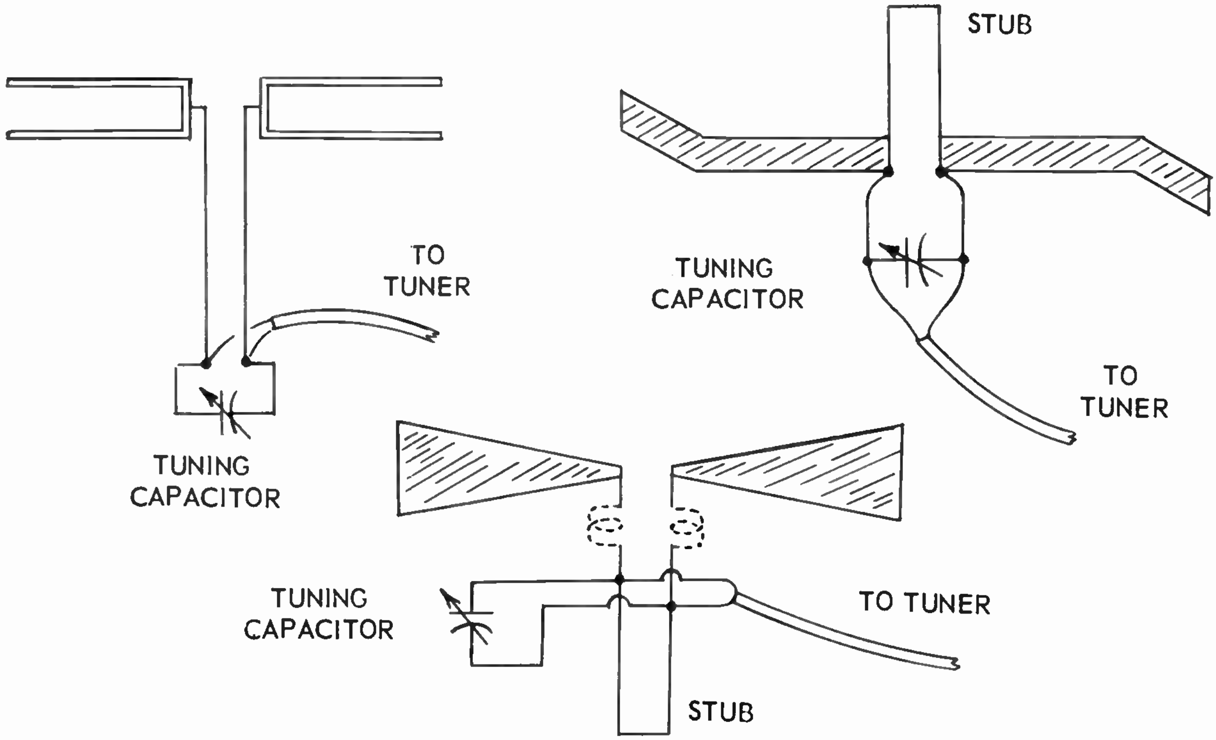


Fig. 69-19. Built-in antennas fitted with tuning capacitors.

For best reception try turning or orienting the cabinet even when there are adjustable orientation controls. Try different locations in the room. Close to a window usually is good. Ghosts or smears often may be reduced by moving the receiver to another location.

Troubles most often are due to antenna conductors touching the chassis or other metal, to loose or dirty connections or switch contacts, and to the antenna becoming loose in its supports. No lamps or other large metallic objects should be placed on top of the cabinet.

INDOOR ANTENNAS. Indoor antennas, as distinguished from the built-in variety, may be moved about independently of the receiver while remaining connected through a short transmission line. All of the troubles resulting from location of receivers having built-in antennas apply to the location of any indoor antenna. A popular type of indoor antenna is illustrated by Fig. 69-20. Telescoping dipole rods allow varying the length for tuning or resonating to various carrier frequencies. The rods may be raised or lowered to broaden the frequency response. The entire antenna may be rotated for best orientation on each channel. An indoor antenna may be placed on top of or alongside the receiver, but usually works best when near a window. Some types can be hung on a wall, the higher the better. Never coil the transmission line between an indoor antenna and the receiver.

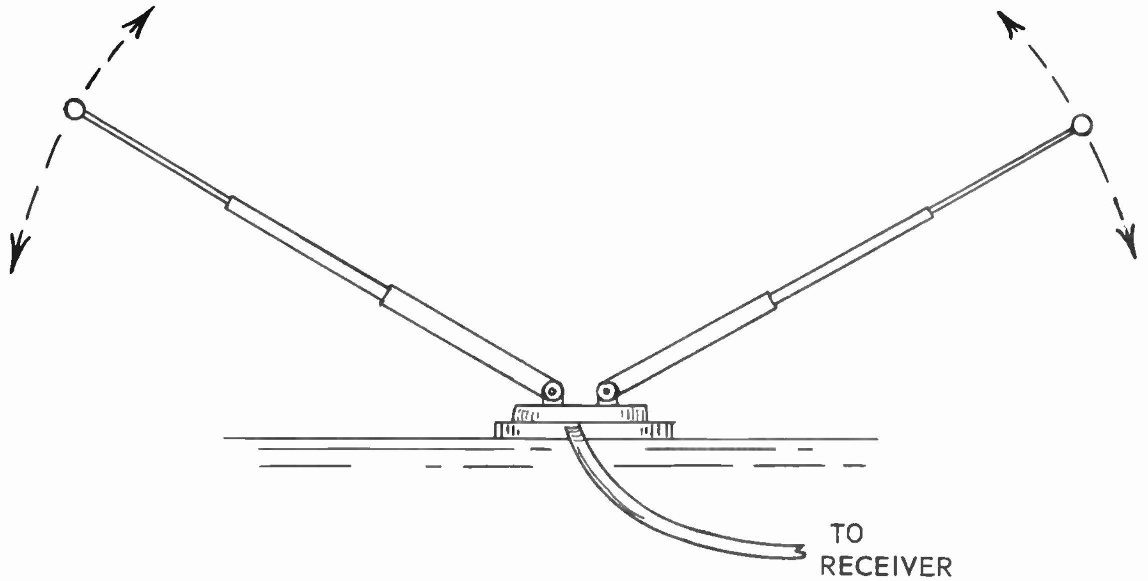


Fig. 69-20. One style of indoor antenna which is portable.

2 To tune and orient an indoor antenna commence with the arms nearly horizontal and at right angles to the direction of wave travel. For the lowest channels pull the rods all the way out and for the higher channels shorten them proportionately. Keep both rods of the same length, approximately. Slowly rotate the antenna for best reception. Try raising or lowering one or both rods into various V-formations. If signal field strength is high a single adjustment usually is satisfactory for an entire band, but rod lengths usually must be changed when going from one band to the other.

Folded dipole indoor antennas may be made with a length of twin-conductor transmission line as shown by Fig. 69-21. Bare the outer ends, twist them together, and solder. At the exact center of the overall length bare about 1½ inches of one conductor, and cut this conductor at the center of the bared part. Bare the two conductors for about 3/4 inch at the antenna end of the transmission line which is to go to the receiver. Twist the antenna and line conductors together as shown by the illustration and solder the joints. The overall length of the antenna section is determined in the same manner as for any other folded dipole, with due allowance for velocity factor of the kind of line being used.

The antenna of Fig. 69-21 may be concealed behind drapes or in any other inconspicuous place where it may be suitably oriented. This type or any other kind of antenna may be mounted in an attic. If the location is not too far from transmitters, and if the building does not have a metal roof, an attic is an excellent location. It allows greater height above ground level than any other location, and installation is simple because there is no need for substantial brackets or guying.

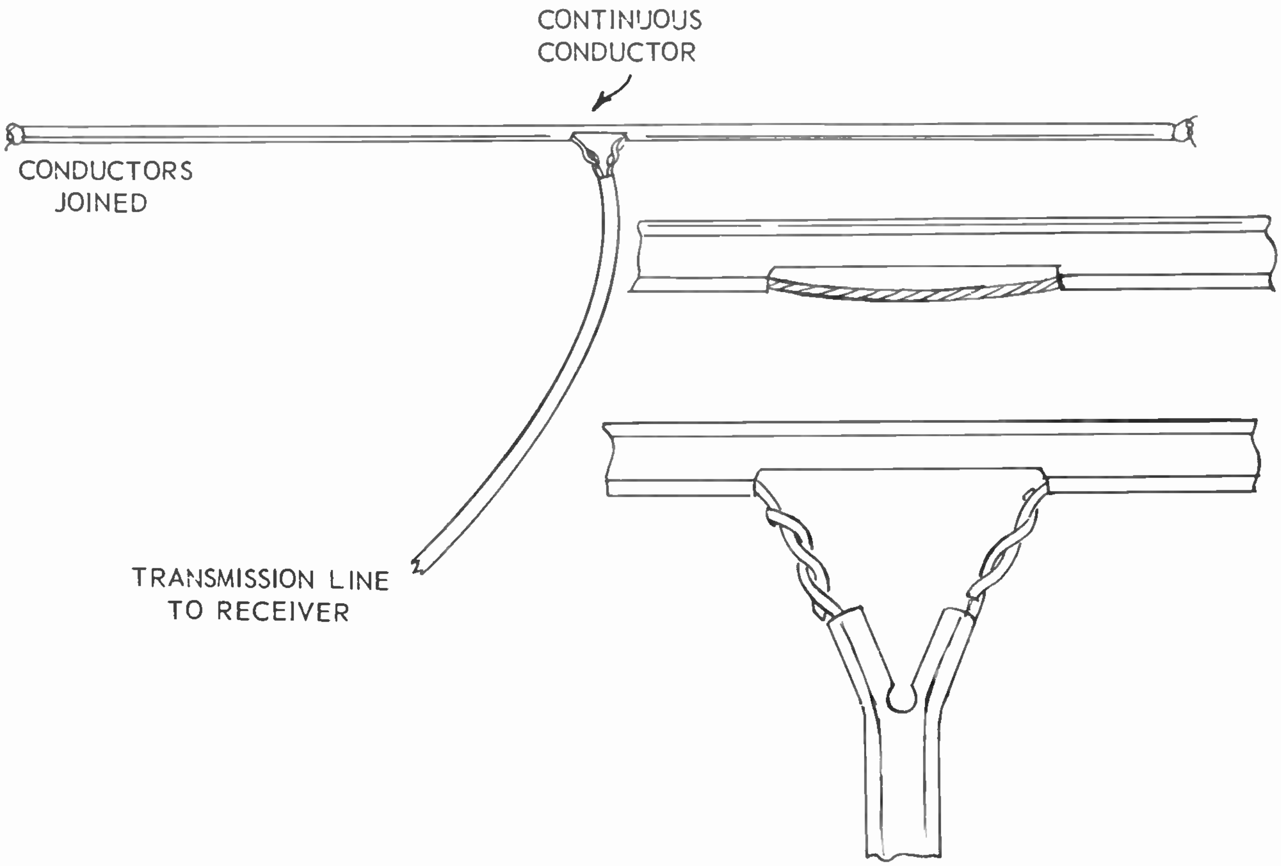


Fig. 69-21. Constructing an indoor folded dipole from transmission line.

TELEVISION

LESSON NO. 70

MEASURING VOLTAGE AND CURRENT

"Training Your Mind"
C. T. I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

DO NOT TEAR – CUT ALONG THIS LINE AND SEND IN FOR GRADING.

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON NO. 70

MEASURING VOLTAGE AND CURRENT

When trouble occurs in a television or radio receiver you try first to identify the probable cause by noting the kind of defects which are apparent in picture or sound reproduction. Failing in this, you may use any of a rather wide variety of specialized testing instruments to trace the difficulty to some certain section or circuit. But to pin-point the fault as existing in a particular part, connection, or tube you nearly always must finally resort to measurements of voltage, current, resistance, or capacitance.

As you might expect, there are quite a few different instruments and methods which may be used for measurements of these properties. We could use a bridge, or possible a dynamometer instrument, or maybe a vane type, or ever an electrostatic instrument in some cases. But it is safe to say that more than ninety-five per cent of all the television and radio service instruments that measure voltage, current, resistance, or capacitance contain a permanent magnet moving coil meter as the indicator. Since this type of meter is so universally employed we should become acquainted with its characteristics.

The face of a typical permanent magnet moving coil instrument which has been adapted for use as a voltmeter is pictured by Fig. 70-1. Concealed by the dial is a large permanent magnet with a cylindrical gap in which is mounted the "movement", as pictured at the left in Fig. 70-2. At the right is shown one of these magnets built up from a number of thin laminations with all the north poles on one side of the gap and with all the south poles on the opposite side.

A meter movement removed from the magnet is pictured at the left in Fig. 70-3. An important part of the movement is the armature, shown at the right. The armature consists of a number of turns of small wire wound on a bobbin of non-magnetic metal. The armature bobbin is carried by pointed pivots at the front and back. These pivots fit into adjustable jewel bearings, usually of natural or artificial sapphires or of special kinds of glass, which are supported from the frame of the movement. The indicating pointer is fastened to the armature bobbin. The armature and pointer may turn in the jewels, but turning is opposed by small coiled hair springs whose outer ends are attached to the stationary part of the movement.

One end of the armature winding is connected to the front hair spring and the other end to the rear hair spring. Direct current from the circuit being measured is brought to the stationary end of one hair spring, through the armature winding and the other hair spring back to the circuit. Fig. 70-4 shows how the front and rear hair springs, in series with the armature winding, are connected to the two large terminal screws that protrude through the back of the meter case.

When no current is flowing in the armature winding and the pointer stands at zero on the dial scale the relative positions of the armature and the pole pieces of the permanent magnet are as shown at 1 in Fig. 70-5. Within the air gap of the permanent magnet, where the armature is located, is a strong magnetic field as shown

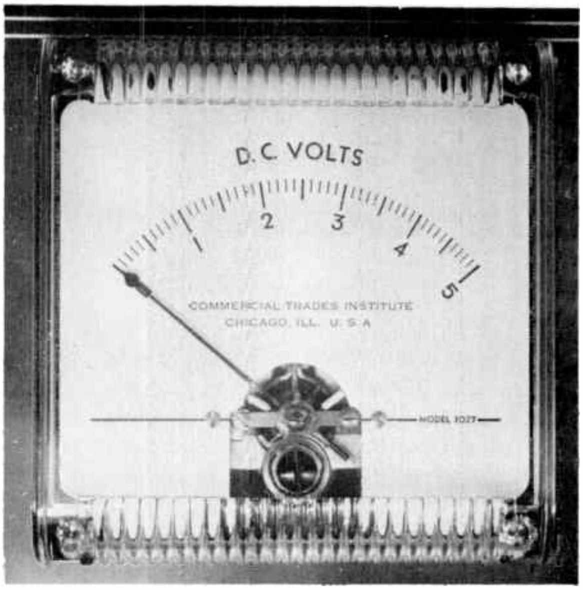


Fig. 70-1. A permanent magnet moving coil meter for measuring voltages.

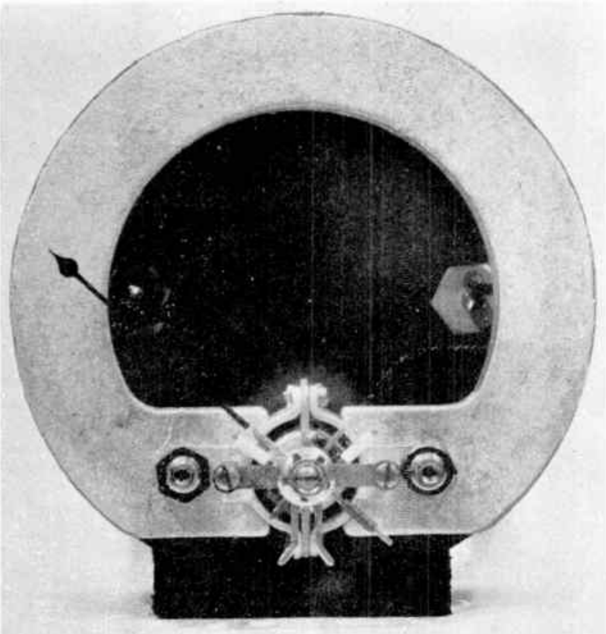
at 2. Not shown in the diagram is an additional cylindrical piece of iron which nearly fills the gap and allows on either side a space only wide enough for free movement of the armature. When current flows in the armature winding this winding becomes an electromagnet with poles and magnetic lines of its own, as shown at 3. Direction of armature current must be such that its south pole is toward the south pole of the permanent magnet, and the two north poles are toward each other.

(5) Like magnetic poles repel each other, while unlike poles attract. Consequently, the permanent and electromagnetic fields react on each other to turn the armature and attached pointer in the direction of the arrow on diagram 3. This turning is opposed by the hair springs, but in spite of this opposition the armature does turn and the pointer moves away from zero and "up scale". The greater the armature current and the stronger its resulting electromagnetic field the farther the pointer is moved.

A permanent magnet moving coil meter gives indications only when there is direct current in its armature, and only when this current flows in the correct direction. Alternating current would produce an electromagnetic field of rapidly alternating polarity. The armature would try to turn first one way and then the opposite, and the result would be merely a vibration of the pointer at the frequency of the applied current. When alternating currents are to be measured they first must be rectified, and the resulting direct current put into the armature.

This type of meter responds only to current in its armature. When the instrument is used as a voltmeter we actually are measuring the current which is caused by the applied voltage, but the dial scale is marked to read the number of volts causing the current to flow.

To move the armature and pointer, and hold them in position against tension of the hair springs, requires producing and maintaining some certain strength of electromagnetic field by continual flow of armature cur-



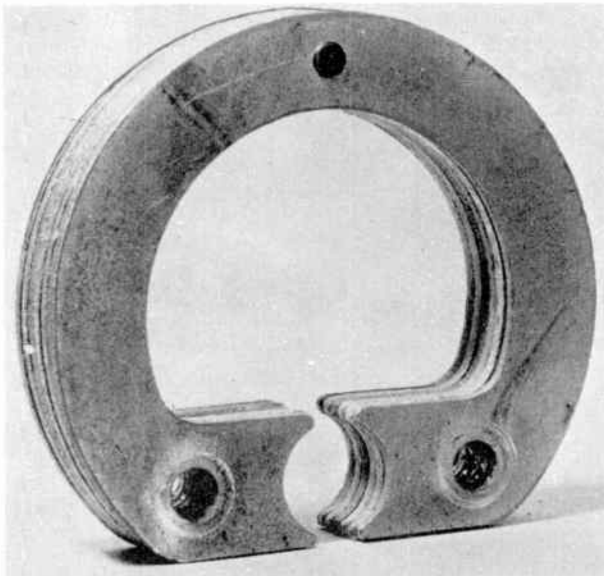


Fig. 70-2. The movement supported in the permanent magnet (left) and the laminated permanent magnet (right).

rent. There are meters in which 10 microamperes or less of armature current will deflect the pointer all the way across the dial scale, but the least armature current for such full-scale deflection in meters for service instruments usually is 50 microamperes. Other low-current ranges in general use include 100, 200, 250, 400, and 500 microamperes, also 1, 5, and 10 milliamperes.

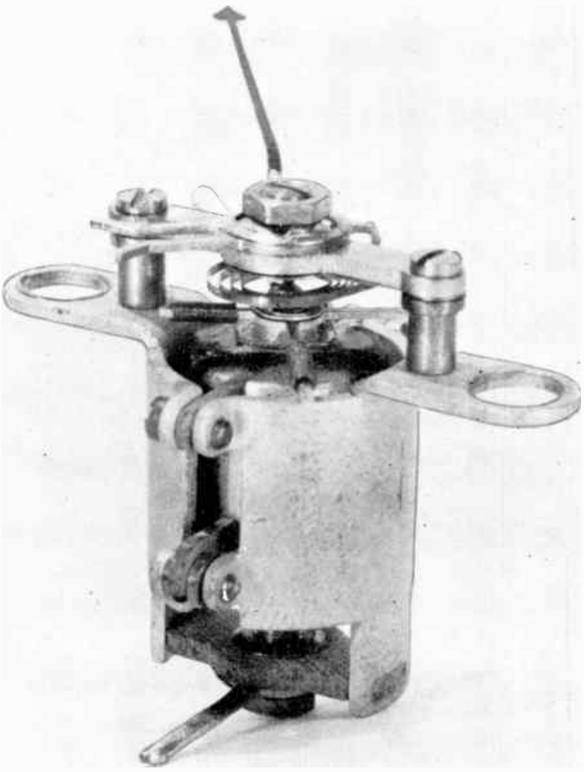
The greater the armature current for full-scale deflection the stronger is the electromagnetic field for any given number of wire turns on the armature, the greater is the power available for moving and holding the pointer, and the more rugged may be the construction. A 1-milliampere meter easily will withstand the treatment usually given to portable testing instruments with little chance for damage, but a 50-microampere meter is a delicate instrument and must be handled with great care.

VOLTMETERS AND MULTIPLIERS. A permanent magnet moving coil instrument may be used for indicating very small voltages provided we know the full-scale current and the armature resistance or internal resistance of the meter. As an example, assume that we have a meter giving full-scale deflection with current of 1 milliampere, and whose internal resistance is 50 ohms. Our regular formula for voltage drop is,

$$\text{Volts} = \frac{\text{milliamperes x ohms}}{1000}$$

Substituting the values for the meter of our example gives,

$$\text{Volts} = \frac{1 \times 50}{1000} = \frac{50}{1000} = \frac{1}{20} = 0.050 \text{ volt, or 50 millivolts.}$$



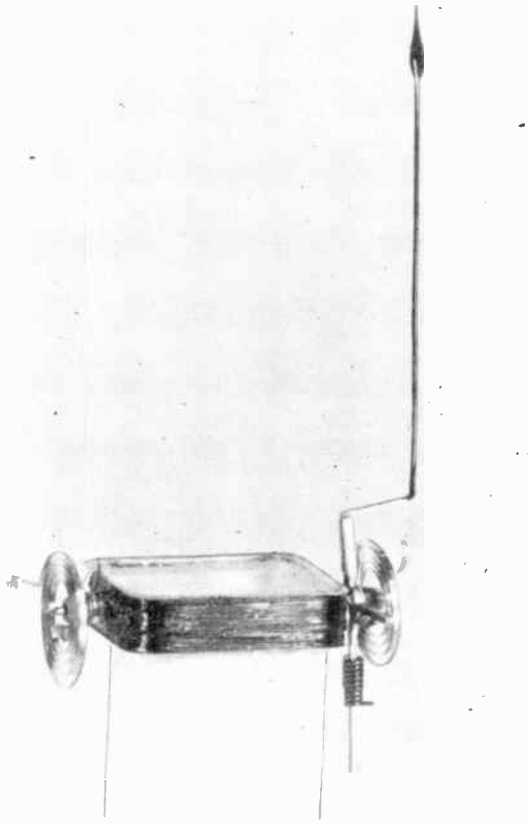


Fig. 70-3. A movement for a permanent magnet moving coil meter (left) and the armature, hair springs, and pointer (right).

Thus we learn that the meter will give a full-scale reading when a potential difference of 50 millivolts is applied to its terminals, because this will cause 1 milliampere of current to flow in 50 ohms of resistance. The scale could be graduated to read uniformly from zero to a maximum of 50 millivolts. Any other current meter could similarly be fitted with a scale reading in fractions of a volt, with the full-scale reading determined from the formula used in our example.

If the meter is to be adapted for reading higher voltages it is necessary to connect in series with the armature a multiplier resistor which will limit the current to that causing full-scale pointer deflection when applying the maximum voltage to be measured. Connections are shown at the left in Fig. 70-6.

The required total resistance of multiplier and armature together may be determined from our regular formula for resistance, which is,

$$\text{Ohms} = \frac{1000 \times \text{volts}}{\text{milliamperes}}$$

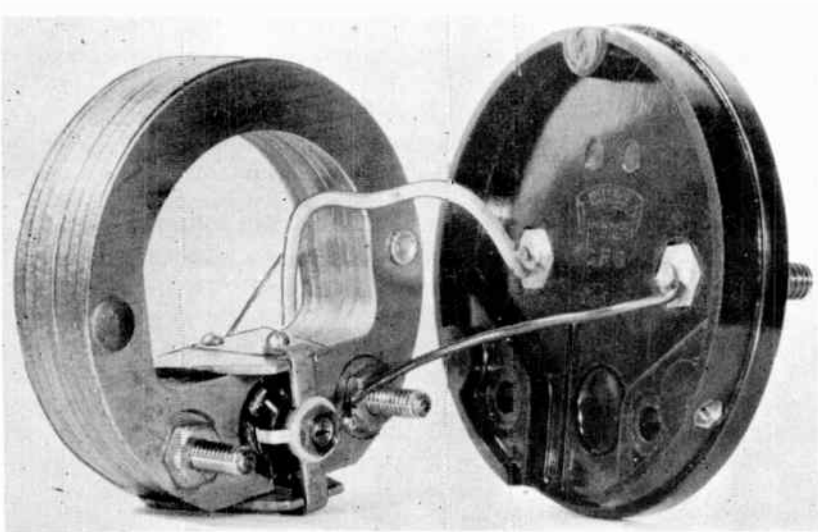


Fig. 70-4. Terminals on the meter case are connected through internal leads to the ends of the hair springs.

Supposing that we wish to have our 1-milliampere 50-ohm current meter indicate d-c voltages from zero to 500, what should be the total resistance of multiplier and meter? We use the formula thus.

$$\text{Ohms} = \frac{1000 \times 500 \text{ (max volts)}}{1 \text{ (full-scale milliamps)}} = \frac{500,000}{1} = 500,000 \text{ ohms}$$

Since the internal resistance of the meter is 50 ohms we could subtract this much from the total required resistance and use 499,950 ohms in the multiplier. But this meter or movement resistance is only one-hun-

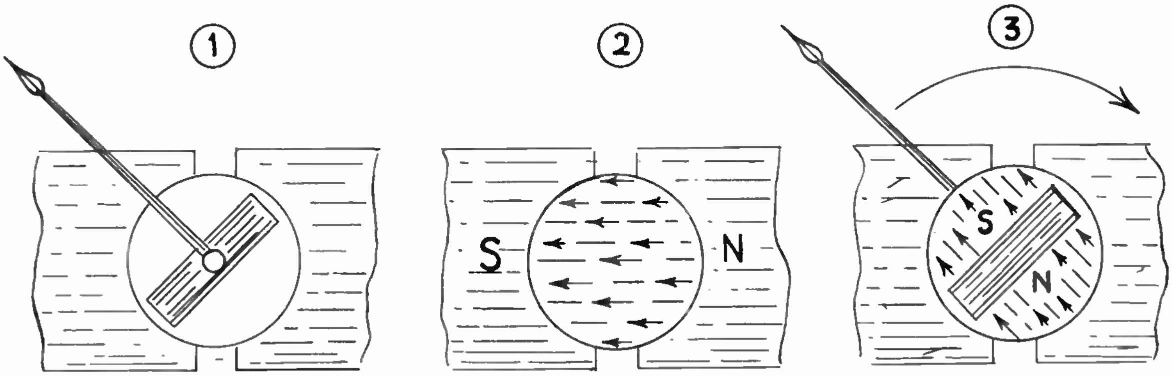


Fig. 70-5. Permanent and electromagnetic fields in a permanent magnet moving coil meter.

dredth of one per cent of the total, and inasmuch as the meter itself hardly ever will be accurate to better than one per cent it would be a useless refinement to subtract the internal resistance in this case. A 500, 000-ohm multiplier would be O.K.

Supposing now that we wish the full-scale reading to be only 1 volt. By using the same formula we find that the total resistance should be 1,000 ohms. The internal resistance of the meter now is 1/20 or 5 per cent of the total, and to preserve reasonable accuracy we should subtract the internal resistance from the required total, and use a multiplier of 950 ohms. In general, if the internal resistance of the meter is more than 1/200 of the total required resistance, the internal resistance should be subtracted from the total in determining required multiplier resistance.

If the full-scale readingof the original currentmeter is measured in microamperes the procedure is similar except that we use this formula.

$$\text{Total ohms} = \frac{1\ 000\ 000 \times \text{full-scale volts}}{\text{full-scale microamperes}}$$

When a meter is originally graduated to read volts, or is a voltmeter, you may extend its range by adding an external multiplier resistor as at the right in Fig. 70-6. A voltmeter nearly always has one self-contained multiplier resistor inside the case. To determine the value of the added external multiplier it is necessary to know the full internal resistance of the voltmeter, not only its armature resistance but the resistance of

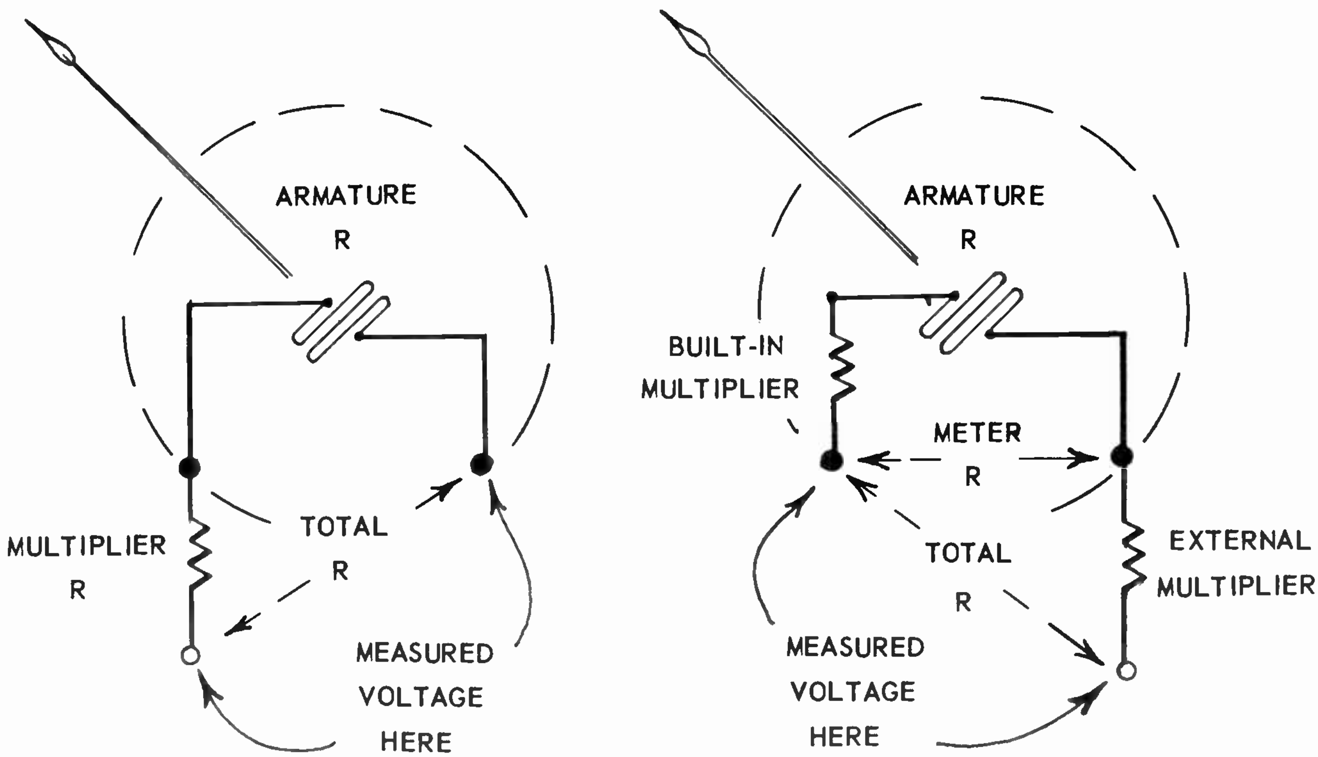


Fig. 70-6. How external multipliers are connected to a current meter and to a voltmeter.

the armature plus the built-in multiplier. Then we proceed to compute the resistance required in the multiplier plus the meter, and from this subtract the meter resistance to find the remaining resistance required in the multiplier. The formula for total resistance of multiplier plus meter is,

$$\text{Total ohms} = \frac{\text{required full-scale reading, volts}}{\text{original full-scale reading, volts}} \times \frac{\text{meter resistance,}}{\text{ohms}}$$

For an example, say that we have a voltmeter whose full-scale reading originally is 5 volts, whose resistance is 5,000 ohms, and that we wish to make the full-scale reading 100 volts. With these values in the formula we have,

$$\text{Total ohms} = \frac{100}{5} \times 5000 = 20 \times 5000 = 100{,}000 \text{ ohms}$$

From this total we must subtract the amount of resistance already in the meter, which is 5,000 ohms in our example. Taking 5,000 from the total of 100,000 ohms leaves 95,000 ohms as the required resistance for the external multiplier.

To make a multi-range voltmeter it would be necessary only to arrange for switching multipliers of various values in series with the meter. To the 5-volt meter we might add ranges of 50 volts and 500 volts. A meter might be provided with ranges of 2½, 10, 50, 250, and 1,000 volts, or with any other desired combination.

2 *CURRENT METERS AND SHUNTS.* A meter will make direct measurements of currents no greater than those which cause full-scale deflection of the pointer. Larger currents may be measured by sending most of the current through a shunt resistor and letting only part of the total flow in the meter armature. Connections are shown by Fig. 70-7.

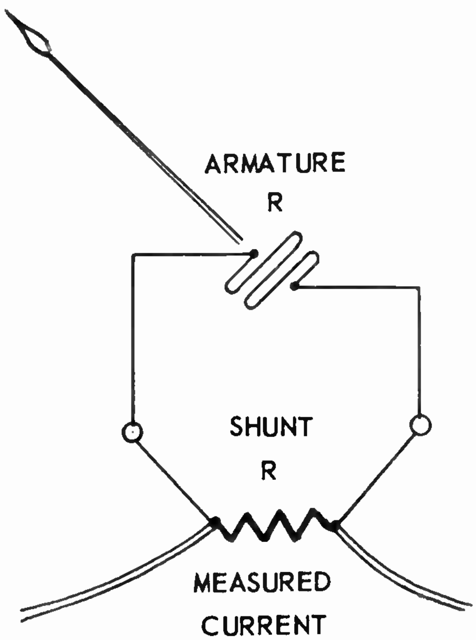


Fig. 70-7. Connection of a shunt to a current meter.

The value of shunt resistance is most easily figured out in steps, as follows.

a. Divide the desired larger full-scale current by the smaller original full-scale current. Use the same unit for both currents; microamperes, milliamperes, or ampers.
b. From the quotient obtained in step *a* subtract 1.
c. Divide the number of ohms of meter or armature resistance by the number obtained in step *b*. The answer is the required shunt resistance, in ohms.

For an example assume that we wish to measure from zero to 10 milliamperes with a meter originally having full-scale deflection on 2 milliamperes. Assume further that the resistance of this meter is 25 ohms. Here are the steps of computation.

a. Divide. $\dfrac{10 \text{ (milliamperes)}}{2 \text{ (milliamperes)}} = 5$

b. Subtract. $5 - 1 = 4$

c. Divide. $\dfrac{25 \text{ (ohms in meter)}}{4} = 6\frac{1}{4}$ ohms or 6.25 ohms for the shunt.

Note that this shunt resistance is 1/4 of the meter resistance. Therefore, the shunt will carry 4 times as much current as the meter. When the meter itself is carrying 2 milliamperes, for full-scale deflection, the shunt will be carrying 4 times 2, or 8 milliamperes. Then the total of meter and shunt currents will be 10 milliamperes, which is the current desired for full-scale deflection.

RESISTORS FOR SHUNTS AND MULTIPLIERS. In addition to determining the value of resistance required for shunts or multipliers we must consider the watts of power which they will have to dissipate as heat. The safe and easy way to figure the power dissipation for multipliers is to square the number of volts at full-scale reading with the multiplier in use, then divide this result by the number of ohms in the multiplier.

For example, assume a full-scale reading of 100 volts with multiplier resistance of 200,000 ohms. We proceed thus.

a. Squaring the full-scale volts, 100, gives 10,000.
b. Dividing 10,000 by 200,000 (multiplier ohms) gives 1/20 watt as the actual power dissipation in the form of heat.

Always use a multiplier resistor rated for at least twice, and preferably for three to four times the number of watts thus computed, in order that the resistor may remain cool. In our present example the multiplier resistor might be selected as of 1/4 or 1/2 watt rating. Unless you are measuring more than 100 volts with a meter taking 1 milliampere or more of current for full-scale deflection, a 1/2-watt multiplier resistor is amply large.

For a shunt resistor, when current is measured in milliamperes, the watts of power dissipation may be found as follows.

a. Square the number of full-scale milliamperes, with the shunt in use.
b. Multiply the result of step *a* by the number of ohms in the shunt resistor.
c. Divide the result of step *b* by 1,000,000.

Here is an example. We have a 6¼-ohm shunt and full-scale current of 10 milliamperes when the shunt is in use. The steps are as follows.

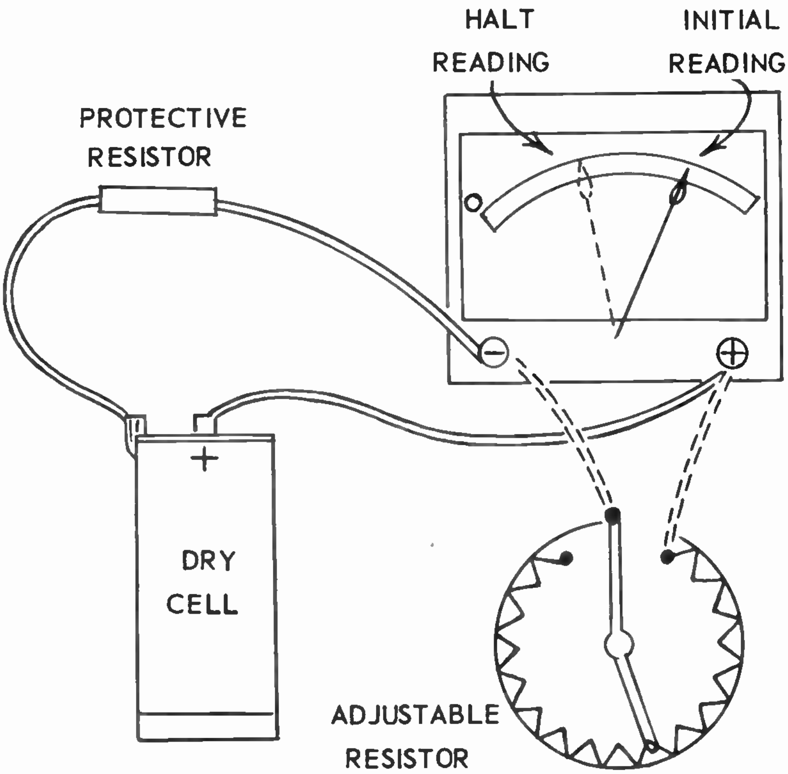


Fig. 70-8. Measuring approximate internal resistance of a meter.

a. Square 10 (full-scale milliamperes), which gives 100.
b. Multiply 100 by 6¼ (ohms in shunt), which gives 625.
c. Divide 625 by 1,000,000, which gives 0.000625 watt as the dissipation.

If current is measured in microamperes we should use the number of full-scale microamperes in step *a*, and in step *c* should divide by 1,000,000,000,000. It becomes fairly obvious that a 1/2-watt resistor will be large enough for practically any shunt used in service instruments.

The tolerance of multiplier and shunt resistors should be no more than half of the rated accuracy of the meter itself, with both expressed as percentages. That is, with a meter of 2 per cent accuracy the multiplier or shunt resistor should have tolerance of no more than 1 per cent. Such tolerances are found only in precision resistors or else in ordinary resistors which have been carefully measured and selected for suitable accuracy.

The resistors may be wire wound, carbon, or composition types when measurements are to be of direct currents or voltages. When permanent magnet moving coil meters are used with rectifiers for measuring alternating currents or voltages, multiplier and shunt resistors should not be of wire wound types, unless of special non-inductive construction, for frequencies higher than about 1,000 cycles. The inductance of ordinary wire wound resistors would seriously affect readings at higher frequencies.

METER RESISTANCE. When it is necessary to measure the internal resistance of a current meter it may be done as shown by Fig. 70-8. First, connect in series with the meter a single dry cell and a protective resistor which will limit the pointer deflection to something less than full scale. For a reading of 3/4 scale you can compute the value of protective resistance by dividing 2,000 by the number of milliamperes full-scale reading. Then carefully note the initial reading of the meter.

With the dry cell and protective resistor still connected to the meter, connect across the meter terminals an additional adjustable resistor of 500 ohms or less. Set this adjustable resistor to bring the meter reading down to exactly half of the initial reading. Resistance of the adjustable unit now is approximately equal to the meter resistance. The adjusted resistor may be measured with a resistance bridge or an ohmmeter. This is not a precision method for measuring meter resistance, because actual resistance of the meter will be somewhat greater than that of the adjusted resistor. The results will be more nearly correct when using a relatively large number of dry cells in series and inserting enough additional protective resistance to limit the meter current to less than the full-scale value.

Never attempt measuring the internal resistance of a current meter by connecting a resistance bridge or an ohmmeter directly to the meter terminals. Current from either of these measuring instruments might easily exceed the maximum safe current for the meter being checked.

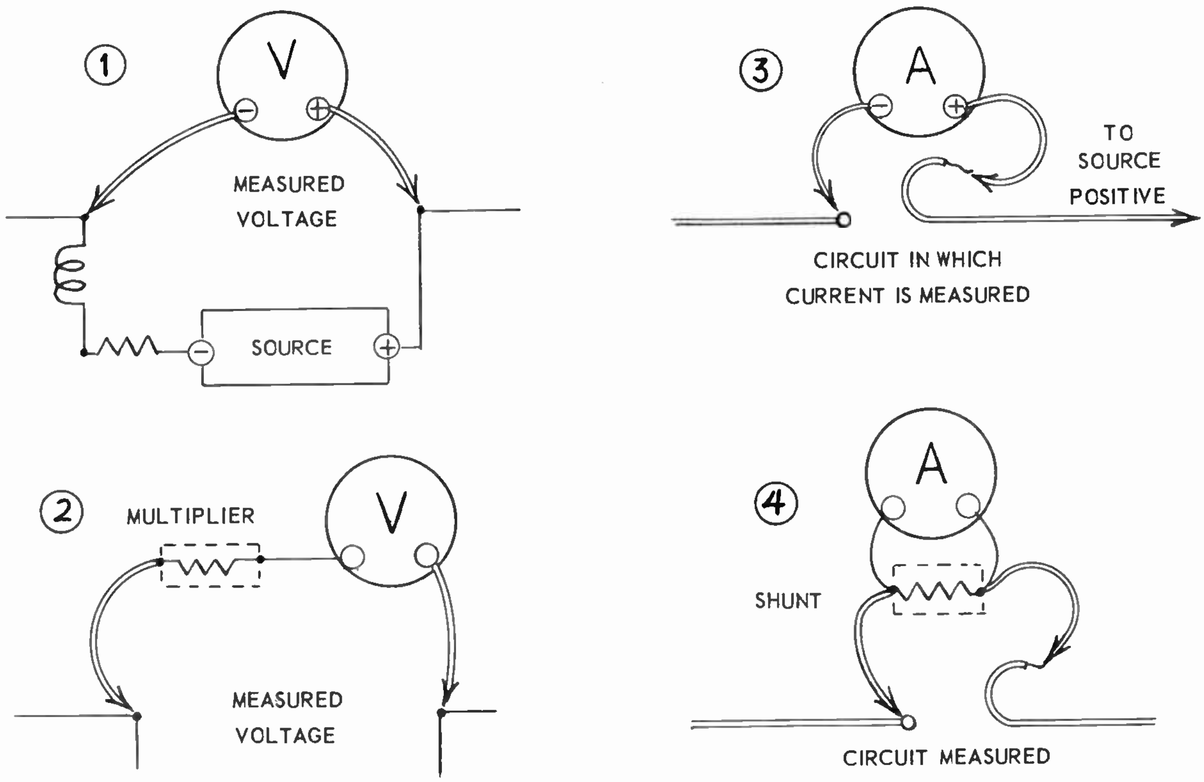


Fig. 70-9. Connections of voltmeters (V) and of current meters (A) to measured circuits.

If a voltmeter has a full-scale range of 30 or more volts, its internal resistance usually may be measured by direct connection of an ohmmeter to the voltmeter terminals. The 30-volt limit is specified because few ohmmeters contain batteries furnishing more than this voltage. If the ohmmeter is known to contain a battery of less voltage it may be used for checking voltmeters having full-scale range equal to or greater than voltage of the ohmmeter battery.

METER CONNECTIONS. As shown at *1* in Fig. 70-9, a voltmeter always is connected to the points between which voltage drop is to be measured. This places the meter in parallel with the circuit or portion of the circuit wherein voltage drop is to be measured. The measured circuit is not opened, and the meter never is connected in series with the circuit unless the purpose is only to check the voltage or emf of the source which is acting in the circuit. Placing the high resistance of a voltmeter in series with a circuit prevents flow of normal current and allows only as much current as may pass through the voltmeter resistance plus the circuit resistance. Then the meter will indicate the source voltage under no-load conditions, or with practically no current being taken from the source.

The positive terminal of the voltmeter should go to the point which connects more or less directly to the positive of the source, if this is known. Service types of voltmeters will withstand application of reversed voltage as great as the full-scale range in the normal polarity. If the meter pointer moves off scale below zero it is simply necessary to reverse the connections to the meter or to the measured circuit. The meter will not have been harmed.

When an external multiplier is used on a voltmeter, as at *2* in Fig. 70-9, one end of the multiplier is connected to either one of the meter terminals and the other end of the multiplier is connected to the measured circuit. It makes no difference to which terminal of the meter you connect any multiplier, it still will be in series with the meter.

1 To use any current measuring meter the measured circuit must be opened, as in diagram *3*, and the meter connected between the opened ends of the circuit. That is, a current meter must be connected in series with the circuit whose current is to be measured. The positive terminal of the meter should go toward the positive side of any source acting in the measured circuit, if this is known. A reversed connection will do no harm if current does not exceed the full-scale meter current in normal polarity.

When an external shunt is used with a current meter, as at *4* in Fig. 70-9, the shunt is connected in series with the circuit whose current is to be measured. Then the meter is connected across the shunt or in parallel with the shunt. Connections between the external shunt and the current meter must be clean and tight. If a shunt has been furnished as part of the meter assembly, use no connections between meter and shunt other than those originally supplied, since the resistance of these connections or leads has been considered when calibrating the meter.

8 If a d-c meter of any kind is connected to or into an a-c circuit the meter will give no reading, but may be burned out by excessive voltage or current. A d-c meter ordinarily will suffer no damage when connected to or into a circuit wherein the direct current or voltage is accompanied by an alternating component, as occurs when measuring in circuits connected to B-power supplies. The peaks of alternating component will not exceed the normal overload capacity of the meter.

READING THE METERS. Unless permanent magnet moving coil meters are fully enclosed within iron or steel cases or housings, you must keep them separated by six inches or more. Otherwise the permanent magnet fields of each instrument will affect the readings of others. Keep meters away from power transformers, chokes, and other parts producing strong magnetic fields, whether or not the meters are in use. A-c fields tend to weaken the permanent magnets which are in the meters.

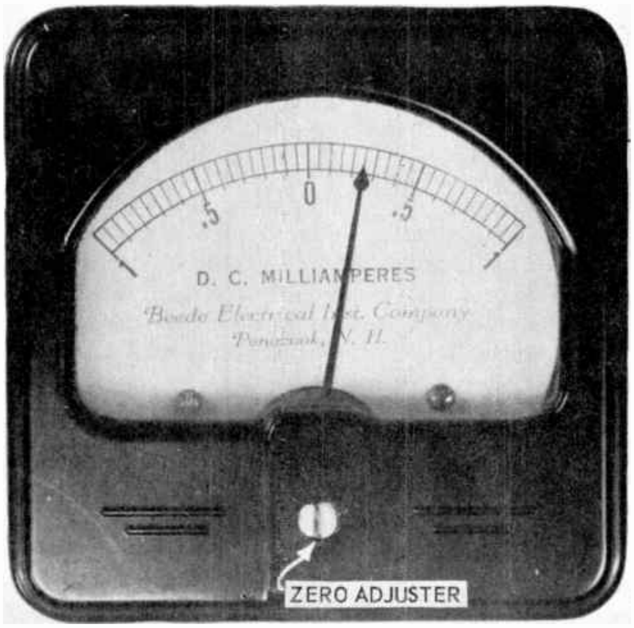


Fig. 70-10. A zero adjuster and the manner in which it moves the pointer.

Take readings only while the meter is in the position for which it is designed, or only while any instrument containing a meter is in this position. If a meter or an instrument is designed to stand upright, with the meter face vertical, take readings only in this position. If the meter is intended to be horizontal or at a slant, take readings only in the appropriate position.

If a meter is originally mounted in a steel panel it has been calibrated to read correctly under this condition, and will give incorrect readings when not so mounted. A meter which has not been calibrated for mounting in a steel panel must be mounted only in some kind of non-magnetic panel, aluminum for example.

Practically all meters are provided with zero adjusters on the front of their housings or cases. The adjuster usually is in the form of a small screw head, such as the one you can see in Fig. 70-10. The screw head has been painted white to make it show up more clearly. Turning the adjuster will bring the pointer to the exact zero position or will move it away from zero in either direction. The adjuster pictured has been turned intentionally to the wrong position, and the pointer has been moved to the right of its correct zero position.

Before voltage or current is applied to any meter you should observe whether or not the pointer stands at zero. It is a good idea to jar the meter very slightly, in case the pointer should be sticking or dragging. If the pointer then is not at zero, carefully turn the adjuster to bring it there.

The meter of Fig. 70-10 is a zero-center type. In all such meters the pointer normally stands at the center of the dial scale, where the reading is zero. There are equal graduations and equal ranges of voltage or current measurements on both sides of zero.

When using any meter you must take every reasonable precaution to protect it from damaging overloads. Most meters in service instruments will withstand current or voltage up to at least one and one-half times their full-scale range, and many will withstand twice the full-scale value without damage. Any greater current or voltage probably will burn out the armature winding, or an internal shunt or multiplier or, at least, will cause bending of the pointer. Any of these accidents means a costly repair job by a meter specialist.

a If you are working with any multi-range meter, of which one style is pictured by Fig. 70-11, make it an invariable rule to set the range selector for the highest range of voltage or current when commencing a test. If the reading then is within the next lower range you may change the selector to that lower range, or to a still lower one if it covers the indicated voltage or current. One other rule, which cannot be overstressed, is to leave the range selector in the position for measuring highest d-c voltage – then you or someone else making some following test won't inadvertently connect a 10-volt range into a 500-volt circuit.

With a single range meter it ordinarily is a case of your knowing the circuits well enough to be sure that the meter won't be overloaded when the voltage or current is applied. Technicians who work with costly instruments often fit them with a shunt which, when connected, will drop the reading to something like a tenth to a fifth of the normal indication. They have this shunt connected when first applying current to the meter. If the reading does not go above a tenth to a fifth of full scale, the shunt is disconnected and readings are taken in the usual way. A voltmeter may be similarly protected with a high-resistance multiplier.

SENSITIVITY OF METERS. The sensitivity of a current measuring meter is specified as the number of amperes, milliamperes, or microamperes which will cause full-scale deflection of the pointer. When, for example, we speak of a one-milliampere meter we are referring to the sensitivity of an instrument which will give full-scale deflection when carrying current of one milliampere.

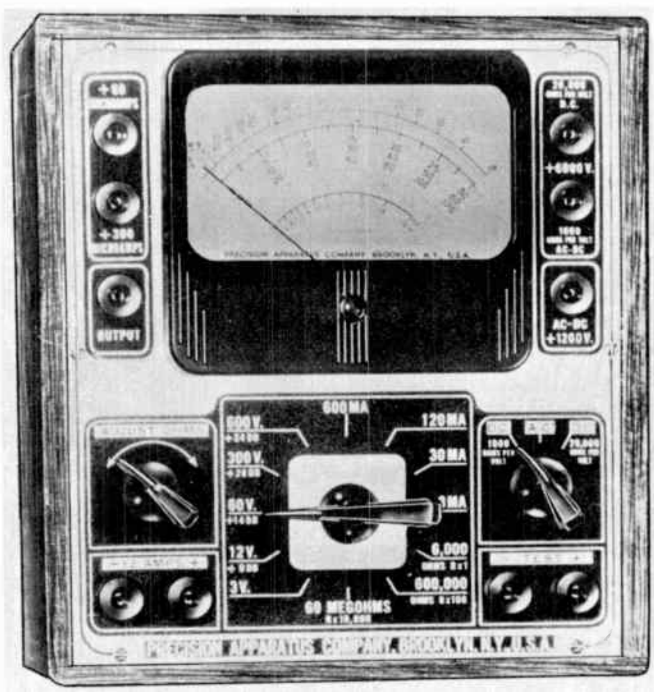


Fig. 70-11. The range selector switch of this multi-range instrument is directly below the meter.

TELEVISION

Sensitivity of d-c voltmeters usually is expressed as the number of ohms resistance per volt of full-scale deflection. When we say that a voltmeter has sensitivity of 1,000 ohms per volt we mean that the total internal resistance, including any built-in multiplier, is 1,000 ohms multiplied by the number of volts at full scale. If this meter has full-scale deflection of 10 volts the total internal resistance is equal to 1,000 (ohms per volt) times 10 (full-scale volts), or is 10,000 ohms. Were the sensitivity 20,000 ohms per volt, and the full-scale reading 50 volts, the internal resistance would be 20,000 times 50, or would be 1,000,000 ohms. All this applies to each range of a multi-range voltmeter, the resistance for each range is found from the sensitivity and the full-scale voltage of that particular range.

When knowing the sensitivity of a voltmeter in ohms per volt the full-scale current in milliamperes is determined upon dividing 1,000 by the number of ohms per volt. For instance, with any voltmeter rated at 1,000 ohms per volt we divide 1,000 by 1,000 to find that the full-scale current must be 1 milliampere. If the sensitivity is 20,000 ohms per volt we divide 1,000 by 20,000 to find that full-scale current is 1/20 milliampere or 50 microamperes. When using this method the voltage range of the meter has no bearing on full-scale current. That is, a meter whose sensitivity is 1,000 ohms per volt always takes 1 milliampere at full-scale no matter what the full-scale voltage may be. A meter rated at 20,000 ohms per volt always takes 50 microamperes at full scale, for any voltage range, and so on.

We have used sensitivities of 1,000 ohms per volt and of 20,000 ohms per volt in many examples because most service voltmeters have either one or the other of these two ratings. There are some meters with sensitivities of 5,000 ohms, 10,000 ohms, and 25,000 ohms per volt, but they are not very common.

The greater the sensitivity of a voltmeter the more nearly its readings will approach the circuit and element voltages which exist with no meter connected, while a low-sensitivity meter may introduce a considerable error. The reason for this is illustrated by Fig. 70-12. At *1* is shown a series circuit containing a source of 250 volts and two resistors of 75,000 ohms and 50,000 ohms, across which the voltage drops are respectively 150 volts and 100 volts in normal operation. Current is 2.00 milliamperes everywhere in this series circuit.

We will now work out the actual reading which would be obtained when using a voltmeter having a sensitivity of 1,000 ohms per volt, and the voltmeter is set on the 250 volt range.

The meter multiplier will now be 250 x 1,000 = 250,000 ohms, and this will be placed in parallel across the 75,000 ohms, and the total resistance will now equal:

$$\frac{250{,}000 \times 75{,}000}{250{,}000 + 75{,}000} = \frac{18{,}750{,}000{,}000}{325{,}000} = 57{,}692 \text{ ohms}$$

The total resistance in the circuit will now be:

$$57{,}692 + 50{,}000 = 107{,}692$$

The current flowing in the circuit will equal:

$$\frac{1{,}000 \times E}{R} = \frac{1{,}000 \times 250}{107{,}692} = \frac{250{,}000}{107{,}692} = 2.3 \text{ MA}$$

The voltage drop across the 75,000 ohm resistor and the meter will be:

$$\frac{IR}{1{,}000} = \frac{57{,}692 \times 2.3}{1{,}000} = 132 \text{ volts approximate}$$

The percentage error will equal the difference between the actual 150 volts present with no voltmeter in the circuit, and the 132 volts which is indicated by the voltmeter, which will equal a difference of:

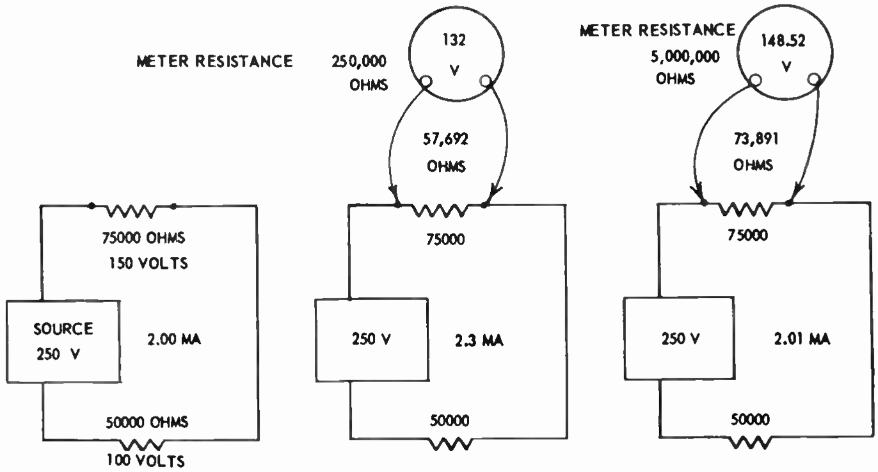


Fig. 70-12. There is considerable error when measuring high-resistance circuits with a voltmeter of low sensitivity.

$$150 - 132 = 18$$

$$\frac{18}{150} \times 100 = 12\%$$

The voltmeter which will now be used will be 20,000 ohms per volt, and the meter multiplier will now equal 250 x 20,000 = 5,000,000. The multiplier of the voltmeter will be placed in parallel with the 75,000 ohm resistor, which will now equal:

$$\frac{5{,}000{,}000 \times 75{,}000}{5{,}000{,}000 + 75{,}000} = \frac{375{,}000{,}000{,}000}{5{,}075{,}000} = 73{,}891 \text{ ohms approximate.}$$

The total resistance in the circuit equals 73,891 + 50,000 = 123,891

Current in the circuit equals:

$$\frac{1{,}000 \times E}{R} = \frac{1{,}000 \times 250}{123{,}891} = \frac{250{,}000}{123{,}891} = 2.01 \text{ MA}$$

The voltage drop across the 75,000 ohm resistor and the meter will be:

$$\frac{IR}{1{,}000} = \frac{73{,}891 \times 2.01}{1{,}000} = 148.52 \text{ volts}$$

The percentage error of the actual voltage measured equals:

$$\frac{1.48 \times 100}{150} = .98\%$$

3 The high-sensitivity meter causes relatively little error because its own resistance is far greater than the resistance across which the measurement is taken, while the low-sensitivity instrument causes large error because its own resistance is comparable with the resistance across which voltage is measured. A high-sensitivity voltmeter is needed when measuring high-resistance circuits, but a meter of lower sensitivity may be satisfactory for measurements across low resistances. Were there a resistor of 10,000 ohms in the circuit of Fig. 70-12, the 1,000 ohms per volt meter would measure the drop with an error of only about 3 per cent.

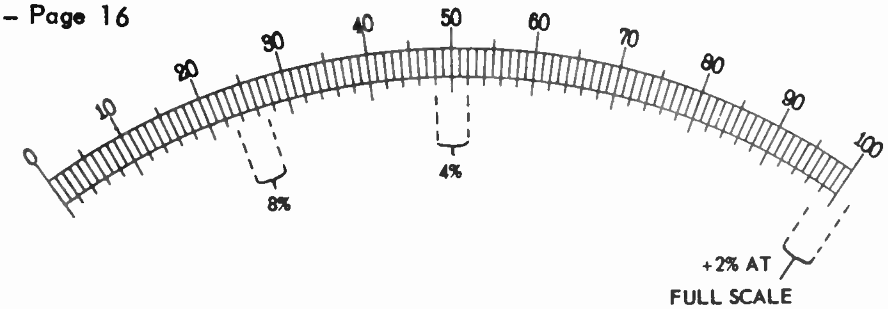


Fig. 70-13. Two per cent accuracy at full scale, and what it may mean at lower readings.

ACCURACY OF METERS. The accuracy of a meter usually is expressed as the percentage of the full-scale reading by which indications may be in error. The percentage of accuracy ordinarily is "plus or minus", meaning that readings may be either too low or too high by the stated percentage. Let's see what this may mean in practice.

Assume that we have a meter of 2 per cent accuracy. The full-scale reading is 100 of any unit; volts, milliamperes, microamperes, or anything else. Such a dial scale is shown by Fig. 70-13. The possible deviation of the pointer from a correct reading may be 2 units high or 2 units low, not only at full scale but also anywhere else. Should the meter be indicating 50 units the correct reading might be anything between 48 and 52. When indicating 25 units the correct reading might be between 23 and 27. The closer we come to zero the greater may be the error, as a percentage of the indicated value.

Accuracy at the low end of the scale seldom is so bad as you might expect from this interpretation of the possible error, but it is advisable to take readings well up on the scale when the meter ranges are such as to allow doing so. Accuracy is not uniform at all points nor does it vary uniformly from end to end of the scale. There could be less error low down on the scale than higher up; it all depends on how the meter has been calibrated.

Service types of d-c voltage and current meters most often are of 2 per cent accuracy, although meters with 1 per cent accuracy are fairly common. Accuracies of 1 per cent and of 1/2 of one per cent are commonly used for factory production testing, while 1/4 to 1/10 of one per cent accuracy will be found in laboratory instruments and in meters used for checking the accuracy of other meters.

ALTERNATING CURRENT METERS. There are many types of meters well suited for measuring voltage and current at power line frequencies in low resistance circuits. Most of these meters, in the lower price ranges, contain a coil of wire in which flows alternating current taken from the measured circuit. The alternating magnetic field magnetizes two iron vanes in like polarities at every instant. Since like polarities repel, the vanes push apart. One vane is stationary, while the other is attached to a part of the movement that carries the pointer. Deflection of the pointer due to the magnetic repulsion is opposed by hair springs as the pointer moves across the dial scale.

Meters of this general kind are called iron vane types or may be called moving iron types. Their principal use in television and radio servicing is for measurement of a-c power line voltages and the low a-c voltages in heater and filament circuits. The inductive reactance of the meter coil and iron vanes is too great to allow using these meters at audio frequencies, and they are entirely useless at radio frequencies. A rather large current is required for full-scale deflection, which makes this type of instrument unsuitable for measurements in high-resistance low-current circuits.

RECTIFIER METERS. Reasonably accurate measurements at audio frequencies as well as at power line frequencies may be made by rectifying the a-c voltage taken from the measured circuit and putting the rectified direct current through a permanent magnet moving coil meter. The combination is called a rectifier meter. Rectifier meters may have current sensitivities as low as 100 microamperes, but for use in service instruments they usually require approximately 1 milliampere for full-scale deflection.

Instrument rectifiers most often are contact types employing copper oxide or selenium discs. A unit containing four complete rectifier elements is shown at the left in Fig. 70-14. This unit actually is only 1/2 inch high, 1/4 inch wide, and 1/4 inch from front to back. At the right is a picture of a small rectifier unit mounted on the inside of the back of a meter case, and connected to the armature circuit of a permanent magnet moving coil instrument, which thus becomes a rectifier meter. The permanent magnet has been removed for making this picture.

Instrument rectifiers are rated as to a-c voltage and direct current. The voltage is the maximum effective or r-m-s voltage which may be applied to the rectifier, while current is the resulting d-c output for the meter movement.

Four rectifier elements may be used in a full-wave circuit as at *1* in Fig. 70-15. When alternating voltage at the input is momentarily of the polarity indicated by full-line arrows, electron flow is through rectifier *a*, the d-c meter movement, and rectifier *b*. When alternating polarity reverses at the input the electron flow is as shown by broken-line arrows, through rectifier *c*, the meter movement, and rectifier *d*. Electron flow through the d-c meter movement is in the same direction with both polarities of alternating input, consequently the meter always carries direct current in its movement.

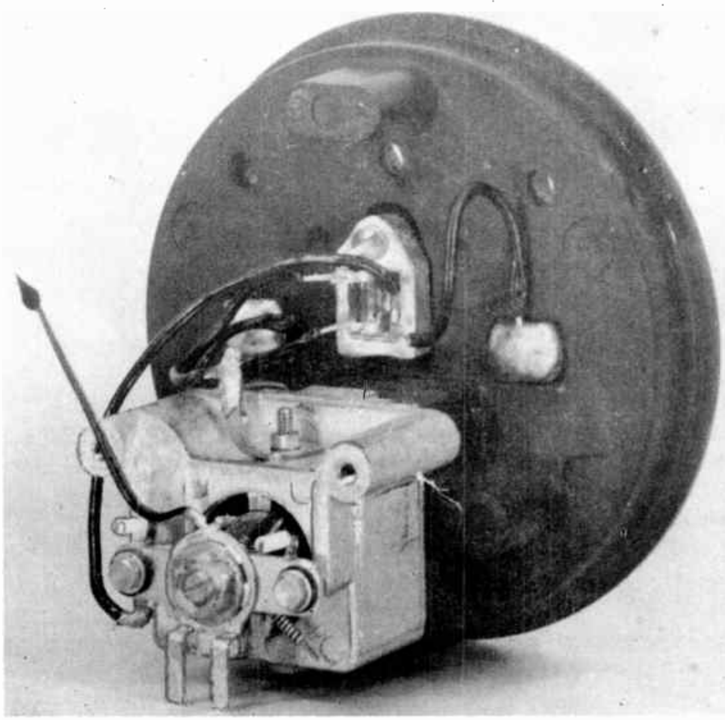


Fig. 70-14. A four-element full-wave instrument rectifier (left) and how such a rectifier is mounted inside a meter (right).

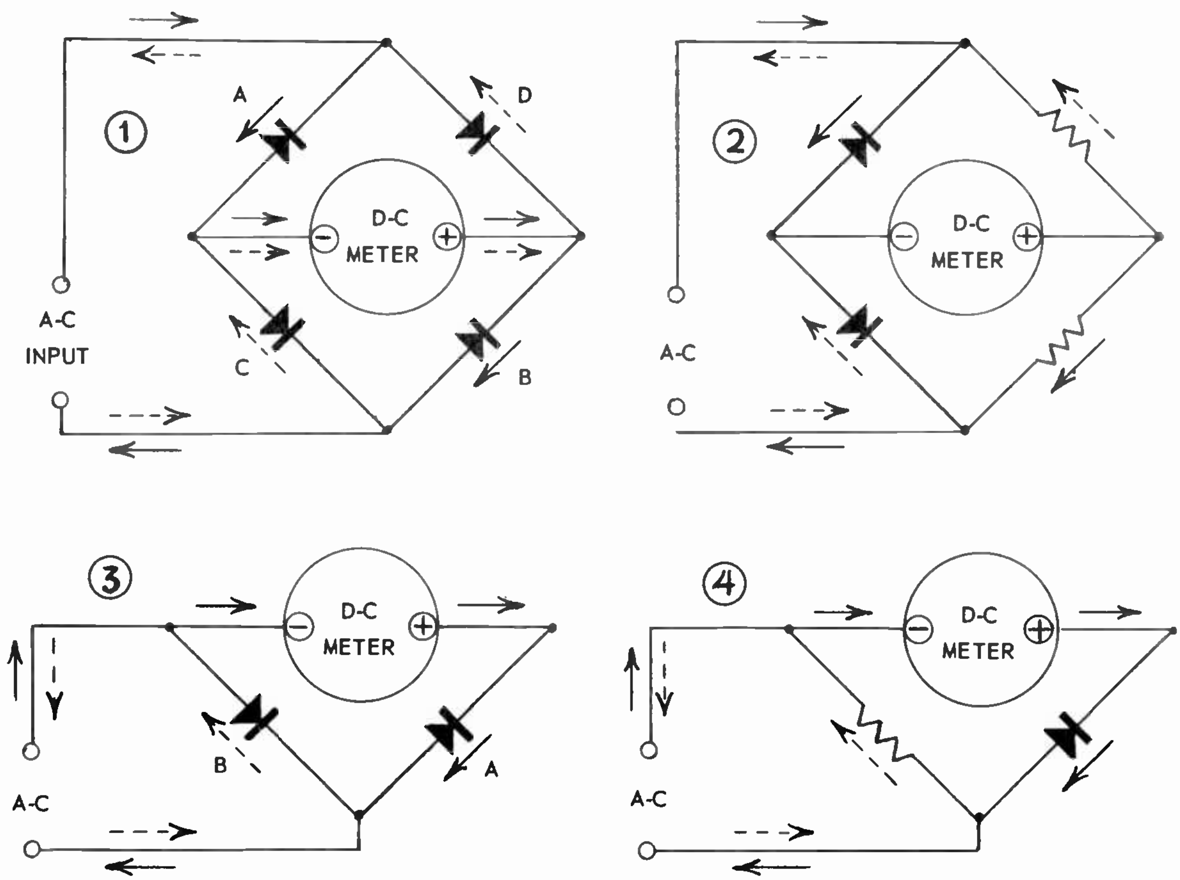


Fig. 70-15. Full-wave and half-wave connections for rectifier meters.

A full-wave circuit using only two rectifier elements is shown by diagram *2*. Two of the rectifier elements of the first diagram have been replaced with resistors to complete the circuits for both alternating polarities while preventing short circuiting of the a-c input.

A half-wave circuit using two rectifiers sometimes is arranged as in diagram *3*. Electrons flow during the alternating polarity indicated by full-line arrows goes through the d-c meter movement and rectifier *a*. During the opposite polarity the electron flow is through rectifier *b*, but not through the meter movement. Another half-wave circuit which employs only one rectifier element is shown at *4*. Only one of the alternating polarities is rectified to furnish direct current for the meter movement. Electron flow during the opposite polarity goes through the resistor.

RECTIFIER METER SCALES. Resistance of a contact type instrument rectifier is somewhat greater for low currents than for high ones which are within the rated limit for the unit. The result is crowding or cramping of the low end of the dial scale. The effects on dial markings for a particular rectifier meter are illustrated by Fig. 70-16. The upper scale, for d-c volts or current, is uniformly divided from zero to maximum.

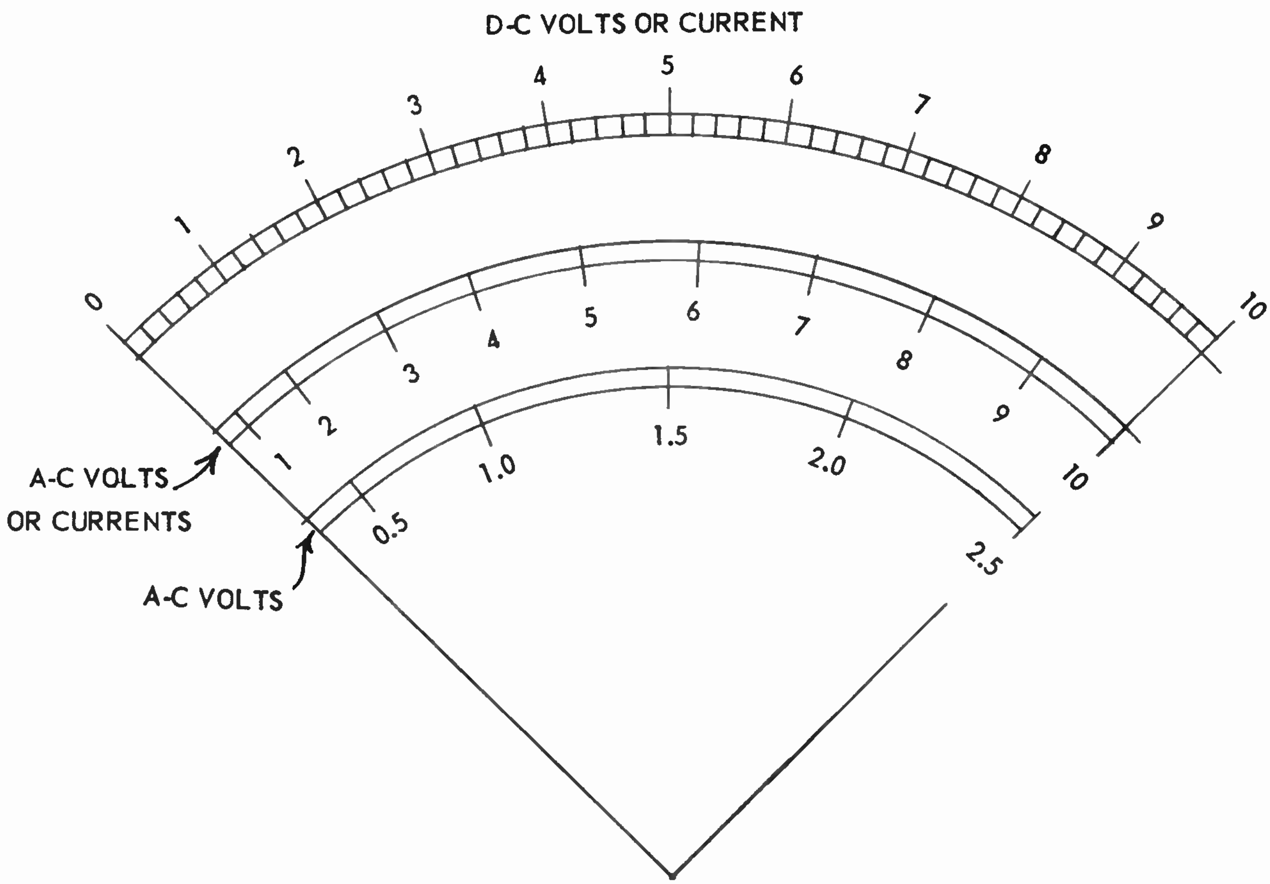


Fig. 70-16. Crowding at the lower ends of scales for rectifier meters.

The bottom scale, for 2½ a-c volts, shows very decided crowding at the low end and spreading at the high end. With this scale in use there will be only small multiplier resistance, or maybe only the rectifier resistance in series with the meter movement. Then the unequal graduations at opposite ends of the scale are due almost entirely to varying resistance in the rectifier unit.

The middle scale, for 10 a-c volts or 10 a-c milliamperes, still is not uniform but it is not so crowded at its low end as the 2½-volt scale. This is because we now will have in series or in parallel with the rectifier a considerable amount of multiplier or shunt resistance, whose value does not change with variations of current. The unvarying series or shunt resistance maintains the total effective resistance fairly constant in spite of changing resistance in the rectifier unit. Scales for still greater voltages or currents will have about the same distribution of graduations as on this 10-volt scale.

Multiplier and shunt resistors may be used with rectifier meters for increasing the range. The required resistances cannot be computed with the simple formulas used for d-c meters. This is partly because pointer travel actually is proportional to the average value of a-c voltage, although the dial is graduated in effective or r-m-s values, and also because energy loss in the rectifier unit will vary between low and high ranges.

Multipliers and shunts always must be connected to the a-c input terminals as marked in Fig. 70-15, never directly to the terminals of the d-c movement. This latter connection would allow the full applied voltage or current to act on the rectifiers, and they would be destroyed.

RECTIFIER METER SENSITIVITY AND ACCURACY. Rectifier voltmeters ordinarily have sensitivities no greater than 1,000 ohms per volt, although sensitivities of 2,000 and even as high as 5,000 ohms per volt sometimes are found. The low sensitivity allows using a meter movement and rectifier units which will carry approximately one milliampere at full scale. When current is much smaller than this, the relations between a-c voltage applied to the rectifier and its d-c output are likely to be rather erratic, and the indications less reliable.

The accuracy of a rectifier meter can, of course, be no better than that of the d-c movement which is part of the meter. Accuracy becomes poorer with increase of frequency, or when there is a considerable d-c component, or when applied voltage or current is not of sine wave form, and when the meter is used at very low or very high temperatures. Even under favorable conditions the accuracy seldom is better than plus or minus 5 per cent of the full-scale reading.

Frequency error results chiefly from the fact that the rectifier elements act like small capacitors to by-pass more and more current as frequency increases. Readings are lowered by about 1/2 per cent per 1,000 cycles. In spite of this frequency error, rectifier meters are commonly used for all audio-frequency measurements where only approximate values or comparative values are required.

If used on pure direct voltage or current a rectifier meter will read about 11 per cent too high. This is because these meters are actuated by average a-c values (0.636 of peak value), are calibrated to read effective sine wave values (0.707 of peak value), and the effective value is about 11 per cent greater than the average value. When there is a greater or less d-c component along with the a-c voltage or current the meter will tend to read somewhat too high.

Rectifier meters are calibrated for sine wave voltages or currents. On other wave-forms there may be serious error. If the peaks of actual voltage or current are flatter than those of a sine wave we are approaching a direct current, and the meter will tend to read too high. If peaks are sharper than those of a sine wave the meter will tend to read too low. This waveform error is increased by presence of harmonic frequencies along with the fundamental.

Rectifier meters are calibrated for use at some certain temperature, usually at 68° F or at 77° F. At higher temperatures the instrument rectifier delivers more direct current for a given applied voltage, but its resistance increases. There are opposite effects at lower temperatures. The changes combine to make the meter read too low at temperatures which are either above or below the calibration temperature. These meters usually are most reliable when used at temperatures between 55° F and 90° F.

71→75

COMMERCIAL TRADES INSTITUTE

KNOWLEDGE + AMBITION = SUCCESS

LEARN . . .

Training Your Mind

C.T.I.

Perfecting Your Skill

ACT . . .

"The heights by great men reached and kept
Were not attained in sudden flight,
But they, while their companions slept,
Were climbing upward in the night."

ENJOY . . .

COMMERCIAL TRADES INSTITUTE

CHICAGO, ILLINOIS

TELEVISION

LESSON NO. 71

USING THE VOLT-OHM-MILLIAMMETER

"Training Your Mind"

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON NO. 71

USING THE VOLT-OHM-MILLIAMMETER

We have learned how to measure voltages and currents, both direct and alternating. In what may be called the basic methods of trouble shooting in television and radio circuits you will measure voltages as often as all the other circuit properties combined. We discussed current measurements along with those of voltage largely because the same general type of meter is used in both cases, but current measurements are not very often required in service work. What we do require, next to voltage measurements, is the ability to make rapid and reasonably accurate measurements of resistance.

Although the requirements of rapidity and reasonable accuracy are best satisfied with the instrument called an ohmmeter, there are other ways of measuring resistance with only a voltmeter or with a voltmeter and a current meter. Before commencing work with ohmmeters we shall look very briefly at some of these other methods, for surely you will hear of them and will want to know how they work. Connections are shown by numbered diagrams in Fig. 71-1, and methods are described in following paragraphs of corresponding numbers.

1. Connect the unknown resistance in series with a known resistance and a source of voltage. Measure the voltage drop across the known resistance (full-line connections) then across the unknown (broken lines). Use this formula.

$$\text{Unknown ohms} = \frac{\text{volts across unknown resistance} \times \text{ohms of known resistance}}{\text{volts across known resistance}}$$

This method is quite satisfactory with a high-sensitivity voltmeter and when the two resistances are not very greatly different from each other.

2. Connect the unknown resistance in series with a voltage source and the voltmeter, and read the voltage. Short circuit the resistance and again read the voltage.

$$\text{Unknown ohms} = \frac{\text{meter resistance in ohms}}{\text{volts without short circuit}} \times \text{difference between the two reading volts}$$

Unless the unknown resistance is large in comparison with the meter resistance the difference between the two readings will be hard to determine with accuracy.

3. Connect one end of the unknown resistance to one terminal of a voltage source. Measure the source

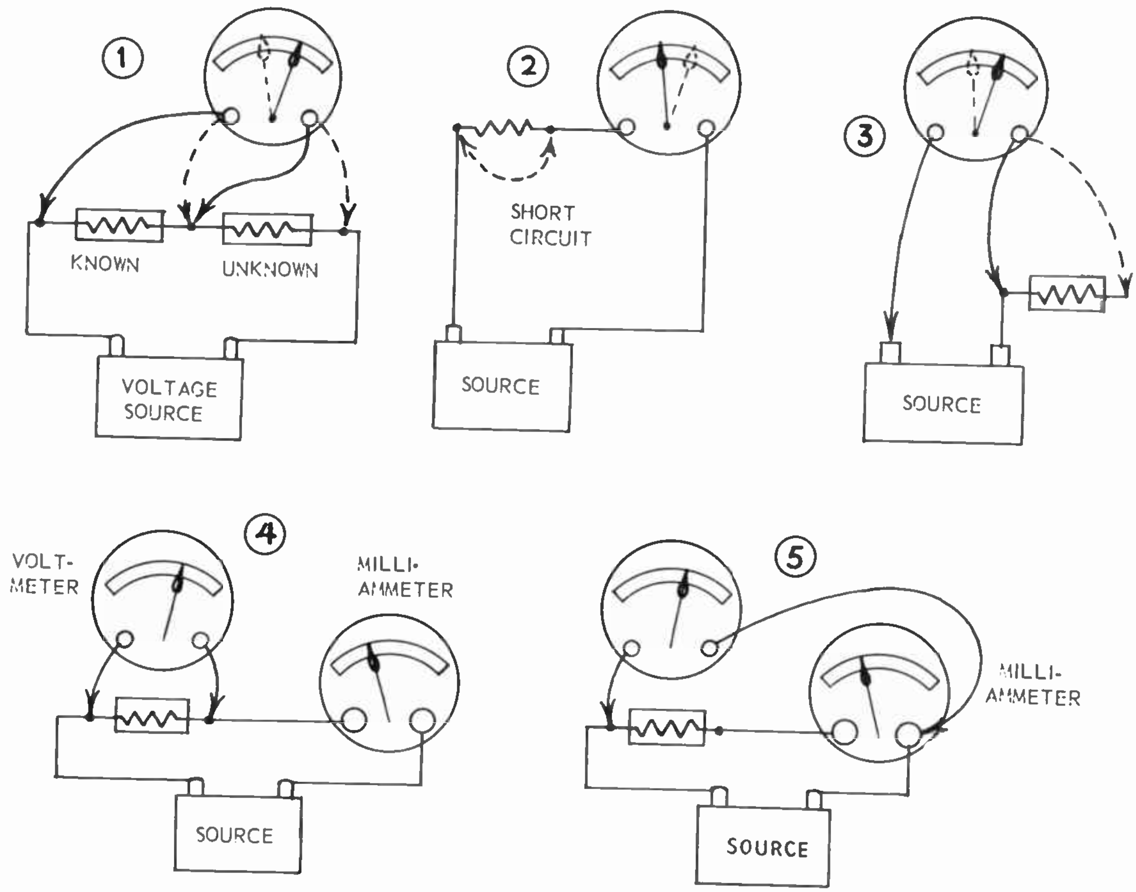


Fig. 71-1. How resistance may be measured with voltmeters and ammeters.

voltage, then measure the voltage across the source and unknown resistance in series. Make the computation as follows.

a. Divide the source voltage by the voltage measured across the source and unknown resistor.
b. Subtract 1 from the quotient obtained in step a.
c. Multiply the result of step b by the number of ohms of meter resistance. The answer is the unknown resistance.

4. Connect the unknown resistance in series with a milliammeter and a source of voltage. Connect a voltmeter across the resistance.

$$\text{Unknown ohms} = \frac{1000 \times \text{volts across resistor}}{\text{milliamperes}}$$

This connection is suitable for measuring small resistances. The milliammeter is carrying the voltmeter current in addition to current in the resistance.

5. The voltmeter here is connected across both the unknown resistance and the milliammeter. Computation of unknown resistance is the same as at 4. This connection is suitable for measuring large unknown resistance. The voltmeter measures the drop across both the resistance and the milliammeter.

All the foregoing methods are slow because you have to take two readings and use arithmetic to determine the unknown resistance. They are, however, useful when you have no ohmmeter at hand.

OHMMETERS. An ohmmeter is an instrument which gives direct readings of the number of ohms resistance in any part to which the ohmmeter terminals are connected. The principle employed with one type of ohmmeter is illustrated by Fig. 71-2. In diagram 1 we have a series circuit containing a 1-milliampere 50-ohm current meter, a fixed resistor of 1,450 ohms, and a dry cell whose potential difference is 1.5 volts. Neglecting the relatively small resistance of the dry cell, the total resistance of this series circuit is 1,500 ohms, as made up of the resistances of the meter and the fixed resistor. The circuit is open between terminals X and X. Therefore, there is no current flowing, and the meter pointer will not move. An open circuit is the equivalent of an infinitely great resistance, so we may say that a maximum reading of the meter indicates infinite resistance between terminals X and X.

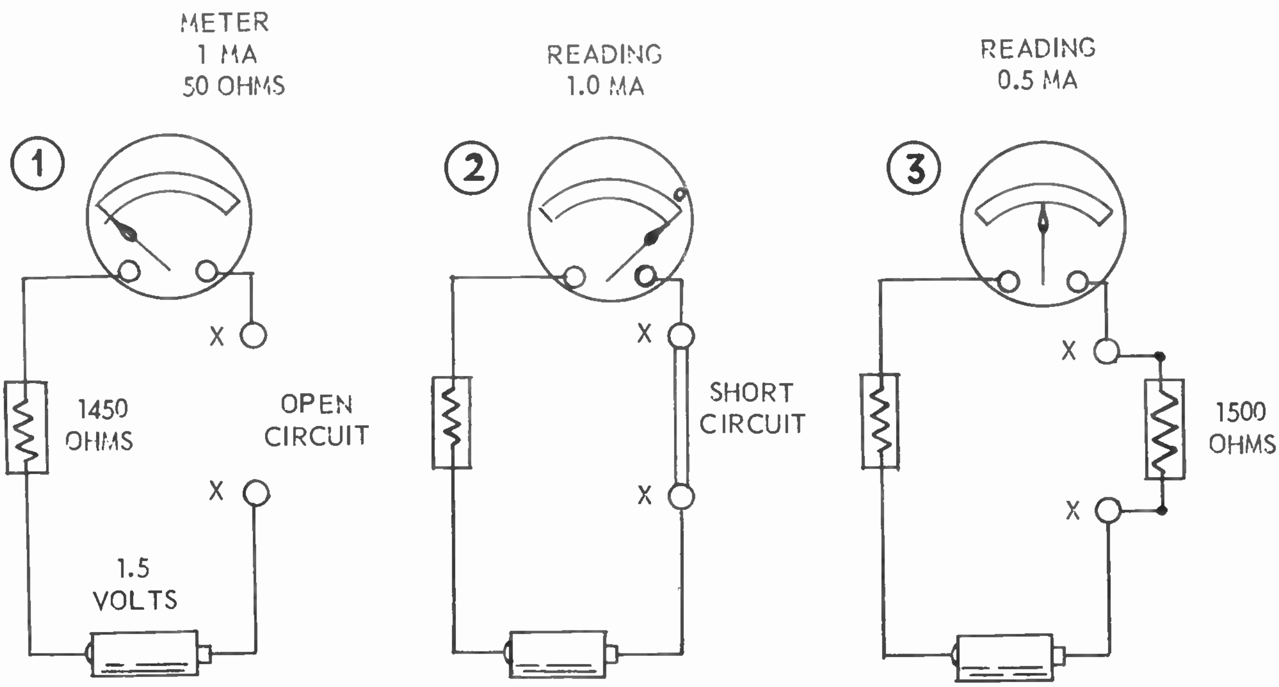


Fig. 71-2. The elementary principle of the "series" type ohmmeter.

In diagram 2 there is a short circuit or a connection of wholly negligible resistance between terminals X and X. Now we have 1.5 volts from the dry cell acting on the total circuit resistance of 1,500 ohms. Your regular formula for current will show that the meter, and the remainder of the circuit, will be carrying current of 1 milliampere. We may say that a 1-milliampere reading or a full-scale pointer swing of this meter indicates that there is zero resistance between the terminals.

In diagram 3 there is connected between terminals X and X a resistance of 1,500 ohms, which is exactly equal to the resistance of the remainder of the circuit. With 1.5 volts from the dry cell acting on this increased resistance of 3,000 ohms in the entire circuit the current will be 0.5 milliampere, as indicated by the meter. You may compute this value of current by using our current formula with 1.5 volts and 3,000 ohms. Then we may say that a meter reading of 0.5 milliampere indicates that resistance of 1,500 ohms is connected between the terminals.

We may compute the currents and meter readings for various other resistances connected between terminals X and X, always using 1.5 volts as the potential difference, and for total resistance adding together the initial circuit resistance of 1,500 ohms and whatever resistance is connected between the terminals. As examples, you will find that 500 connected ohms allows a reading of 0.75 milliampere, that 1,000 ohms allows 0.60 milliampere, that 3,000 ohms allows 0.33 milliampere, and so on.

It is apparent that the dial scale of the meter may be graduated to read in numbers of ohms connected between terminals X and X instead of being graduated in milliamperes. If we compute the values of current for a great many connected resistances, and mark these resistances on the meter scale the result will be as shown by Fig. 71-3. Thus we shall have constructed a direct reading ohmmeter.

The various resistance values are well spread out at the right-hand end of our ohmmeter scale, but become continually more crowded toward the left. When resistances are fairly small, say no greater than 2,000 ohms, it is quite easy to read or estimate changes as small or even smaller than 50 ohms. But up around 20,000 to 30,000 ohms we could estimate no closer than differences of several thousands of ohms, and above 30,000 ohms it is not worth while to insert any graduations. We would describe our ohmmeter as measuring from zero to 30,000 ohms, with a center-scale reading of 1,500 ohms.

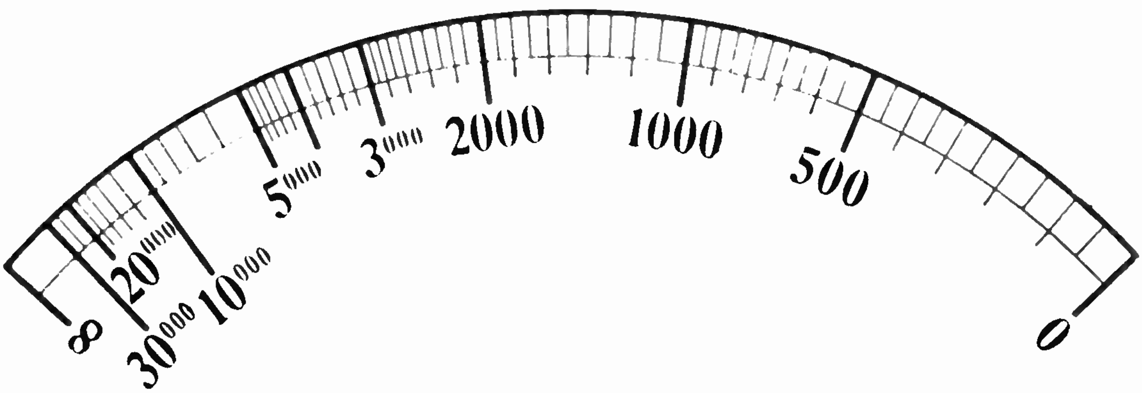


Fig. 71-3. The ohmmeter scale is open at low resistances, crowded at high resistances.

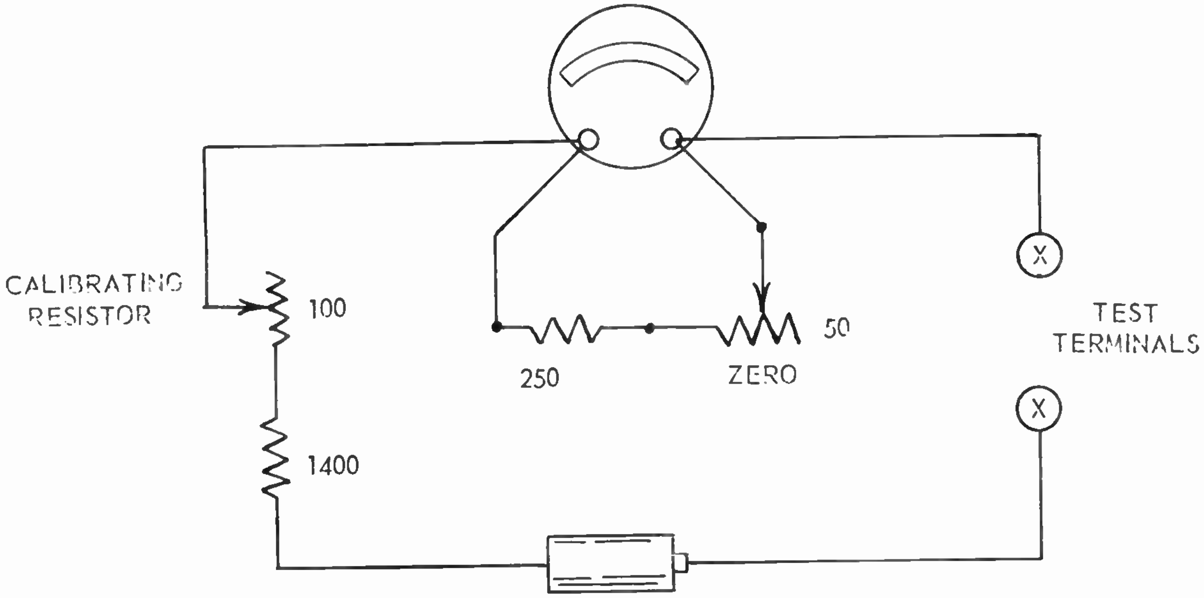


Fig. 71-4. The series ohmmeter circuit with added adjustments for zero ohms and for original calibration.

As the ohmmeter is continued in use the dry cell will become weaker and will force less and less current through the meter with any resistance tested. Readings then will indicate resistances greater than actual resistances of parts being tested. To compensate for a limited decrease of cell voltage we may alter the original circuit as shown by Fig. 71-4. Across the meter we have connected an adjustable shunt consisting of 250 ohms fixed resistance in series with a 50-ohm variable unit which will be called the "Zero Ohms" adjuster. Increasing the adjustable shunt resistance as the dry cell gets weaker will cause less current to flow in the shunt and more in the meter, thus preserving the original indications for measured resistances.

Parallel resistance of the meter and shunt is less than that of the meter alone. The decrease of circuit resistance is compensated for by providing an increased resistance in series with the dry cell and test terminals. This is done by replacing the original 1,450-ohm resistor with a 1,400-ohm fixed resistor in series with a 100-ohm adjustable unit, which is called the calibrating resistor. The calibrating resistor is adjusted permanently by inserting a fresh dry cell, shorting the test terminals, setting the zero ohms adjuster for minimum meter deflection, and adjusting the calibrating resistor to bring the meter to full scale or the zero ohms position on the dial.

1 Shunting resistance in Fig. 71-4 is suitable for a meter having 50 ohms internal resistance. When total series resistance is high in comparison with meter resistance, adjustment of the shunt as the dry cell weakens will not materially alter the accuracy of the instrument. Before proceeding to measure any resistance connected to the test terminals we short circuit these terminals while using the zero ohms adjuster to bring the meter pointer to zero on the ohms scale.

To alter the maximum and center-scale ohms ranges of the type of ohmmeter being examined it is necessary to use a meter of different sensitivity, or to use more or less battery voltage, or to do both at the same time. The maximum measurable ohms always will be 20 to 25 times the center-scale reading. Computations are most conveniently based on center-scale readings. When we select a certain number of ohms as the desired center-scale reading, this also will be the combined resistance of the meter and the resistance to be connected in series with the meter, battery, and test terminals. Here are formulas for required meter sensitivity in milliamperes (full-scale deflection) and for required battery voltage.

$$\text{Full-scale milliampers} = \frac{1000 \times \text{battery volts}}{\text{center-scale ohms}}$$

$$\text{Battery volts} = \frac{\text{meter full-scale milliamps} \times \text{center-scale ohms}}{1000}$$

As an example, supposing we wish a center-scale reading of 5,000 ohms, as shown on the dial of Fig. 71-5, when using a single 1.5-volt dry cell. We use the first formula thus.

$$\text{Full-scale milliamps} = \frac{1000 \times 1.5}{5000} = \frac{1500}{5000} = 0.3 \text{ milliamps, or } 300 \text{ microamperes}$$

Should we wish to use a meter of 1 milliampere (full-scale) sensitivity for 5,000 ohms center-scale reading the second formula is used.

$$\text{Battery volts} = \frac{1 \times 5000}{1000} = \frac{5000}{1000} = 5 \text{ volts from the battery}$$

Since we cannot make up a 5-volt battery from 1.5-volt dry cells it would be better to adopt a center-scale reading of 4,500 ohms and use a three-cell battery if the 1-milliampere meter must be employed.

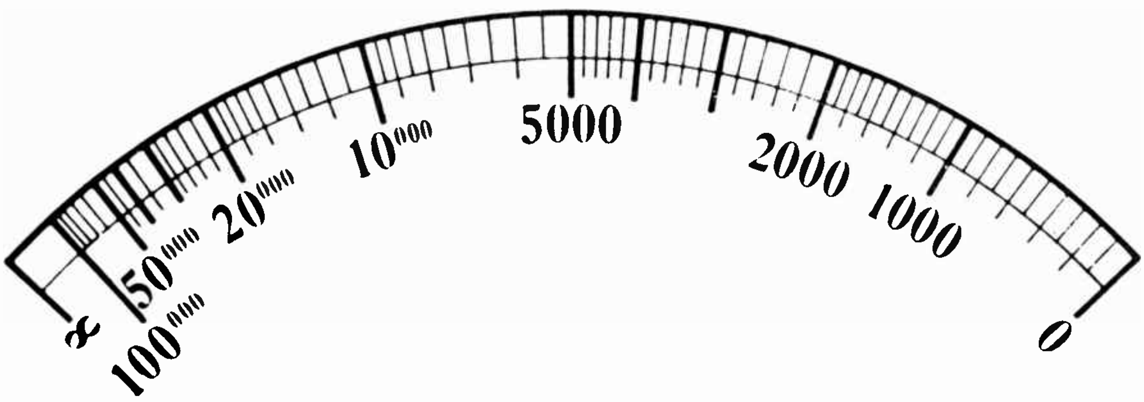


Fig. 71-5. An ohmmeter dial with 5000 ohms as the center-scale reading.

The principle of an entirely different ohmmeter circuit is illustrated by Fig. 71-6. Examination of diagram 1 shows that we are using a meter of 1 milliampere (full-scale) sensitivity and a voltage source consisting of a single 1.5-volt dry cell. Meter resistance is 50 ohms, and in series with the meter is a 1,450-ohm resistor which will limit the current to 1 milliampere with 1.5 volts applied. We might use a meter of any other sensitivity and internal resistance, and a source of any other voltage, provided the resistance in series with the meter will limit the current to the maximum or full-scale value for the meter selected. Measured resistors will be connected between terminals X and X. In this first diagram these test terminals are open, which opens the connection to one end of the dry cell. Consequently, no voltage can reach the meter and the pointer stands at the zero current position, which indicates infinite resistance between the test terminals.

a Diagram 2 shows the test terminals short circuited through practically zero resistance. One side of the dry cell is connected through the short to point a and the other side is connected directly to point b. The meter with its series protective resistor is connected to points a and b, therefore is connected across the dry cell. The 1.5 volts from the cell causes the meter current to become 1 milliampere. The pointer moves to the full-scale position for current, which always is the position for zero resistance between the test terminals. The extra resistor between a and b does not affect the zero ohms indication.

In diagram 3 we have removed the short from the test terminals and have put in its place a 600-ohm resistance. The resistance between points a and b is 1,000 ohms. Electron flow from the dry cell goes through the 600 ohms between the test terminals and to point a. Here the current divides, part going through the 1,000 ohms from a to b while the other part goes through the 1,500 ohms of meter and protective resistor from a to b. The parallel resistance of 1,000 ohms and 1,500 ohms is 600 ohms. Now we have across the dry cell the 600 ohms between the test terminals and the 600 ohms of the parallel paths. The 1.5-volt potential difference of the cell divides equally between these equal resistances, and we have 0.75 volt across the meter and its protective resistor. This half-voltage causes half the full-scale current in the meter, whose pointer goes to the position for 0.5 milliampere. Thus we have a center-scale indication when 600 ohms is between the test terminals, and the ohmmeter dial would be so graduated.

Diagram 4 shows 2,400 ohms of resistance between the test terminals. We still have the 600 ohms of parallel resistance between points a and b. Cell voltage divides inversely as these two resistances, with 0.30 volt between a and b, and 1.20 volts at the test terminals. Since the meter branch is connected to points a and b this branch is affected by 0.30 volt and the meter current becomes 0.2 milliampere. This establishes the position for a 2,400-volt reading on the ohmmeter scale. The completed scale will look about like others we have seen, except that the center-scale reading will be 600 ohms and the maximum 12,000 to 15,000 ohms.

In diagram 5 of Fig. 71-6 we have added a zero ohms adjuster and have changed the center-scale indication to 60 ohms. For the zero ohms adjuster the protective resistance for the meter has been divided. There is a 1,000-ohm fixed resistor and a 500-ohm adjustable resistor in series with each other and the meter. This adjuster will allow compensation for decrease of dry cell voltage with use or with aging.

To change the center-scale reading we have put a 62.5 ohm resistor between a and b instead of the original 1,000 ohms in this position. With 1,500 ohms in the meter branch between a and b, and with the new 62.5-ohm unit between these two points, the parallel resistance becomes 60 ohms. Cell voltage di-

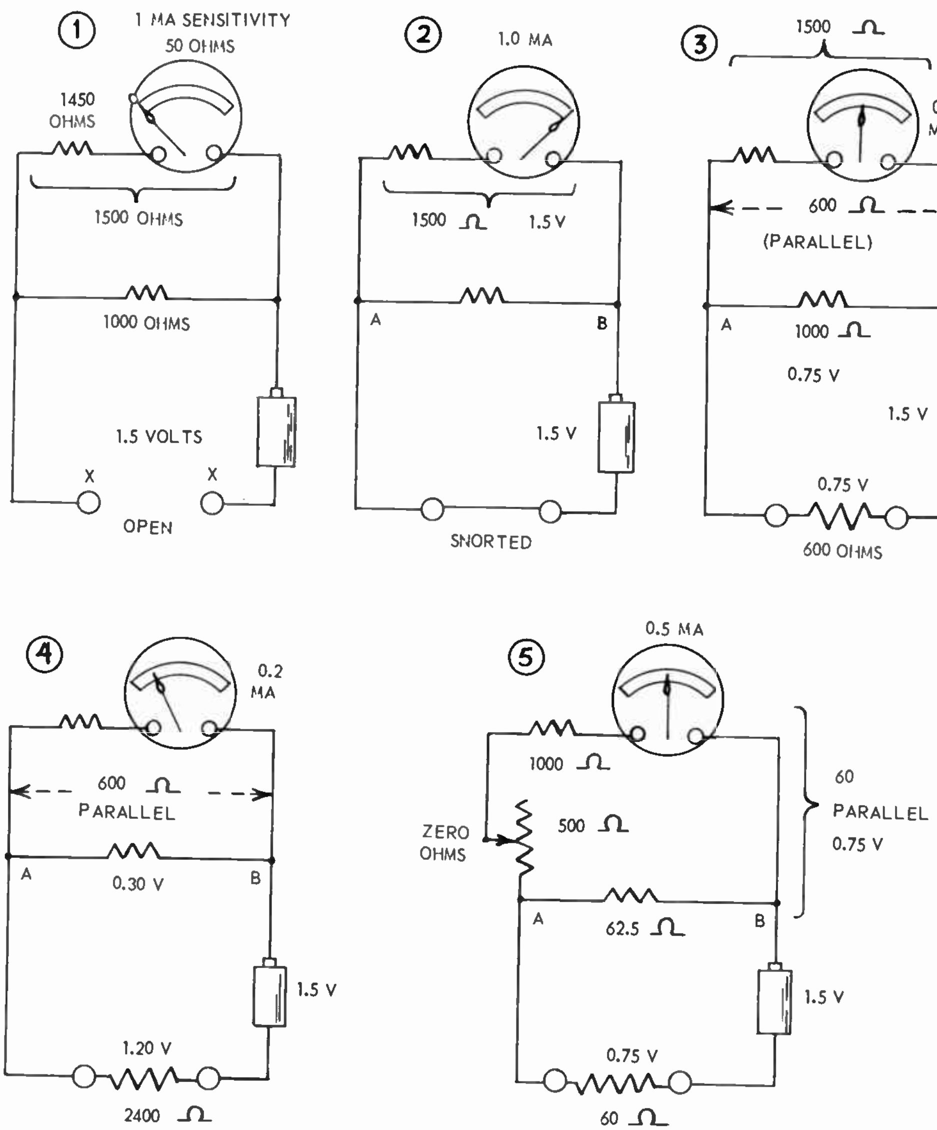


Fig. 71-6. The principle of an ohmmeter employing parallel circuits.

vides equally between this 60-ohm parallel resistance and the 60 ohms between the test terminals. The result is 0.75 volt across the meter branch and a center-scale reading just as in diagram 3.

Various other center-scale and maximum ohms readings might be provided with other resistances between a and b. To make a multi-range ohmmeter we would have to add only a range selector switch and several resistors between a and b. To provide for measuring resistances much greater than allowed with the 600-ohm center-scale arrangement it would be necessary to use either a more sensitive meter or else increased battery voltage, just as with other ohmmeter circuits.

An ohmmeter circuit suitable for measurement of very small resistances is shown by Fig. 71-7. When the test terminals are open there is a series circuit consisting of the battery, a current limiting resistor, an adjustable calibration resistor, a test switch that remains open except while taking readings, and the meter, which is shunted with an adjustable zero ohms arrangement that compensates for weakening of the battery. When the test switch is pressed to close this circuit the meter pointer will move to full scale or nearly there.

When a resistance to be measured is connected between the test terminals this resistance acts as an additional shunt on the meter, reduces the meter current, and brings the pointer farther to the left. The smaller the measured resistance the less becomes the meter current and the more nearly the pointer comes back toward the zero current position. The upper scale of Fig. 71-8 is designed for this type of ohmmeter. The lower scale, designed for one of the previously described circuits, is shown for comparison. Note that the low ohms scale increases from left to right, while the other, more common, scale increases from right to left.

USING THE OHMMETER. If a part whose resistance is to be measured is connected into a circuit, always disconnect at least one end before making a measurement. Otherwise, should the circuit be alive,

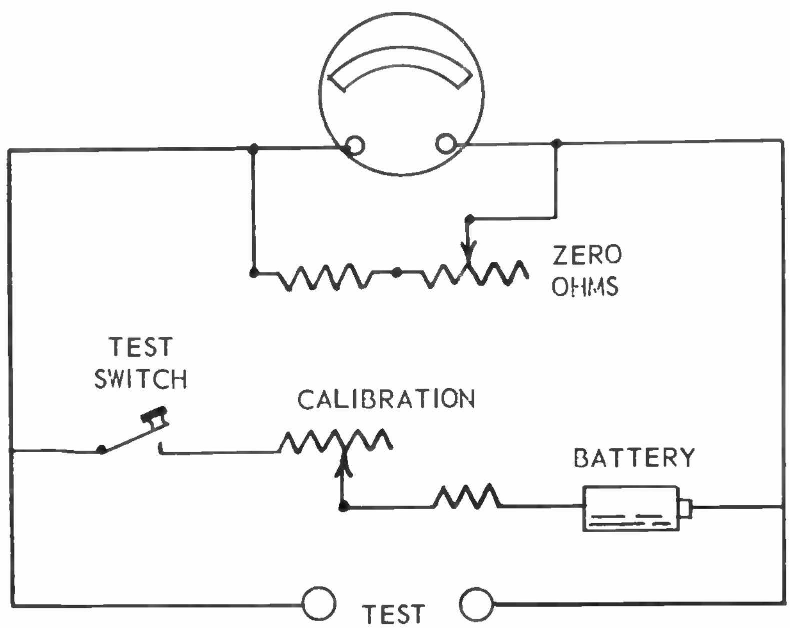


Fig. 71-7. Ohmmeter circuit for measuring small resistances.

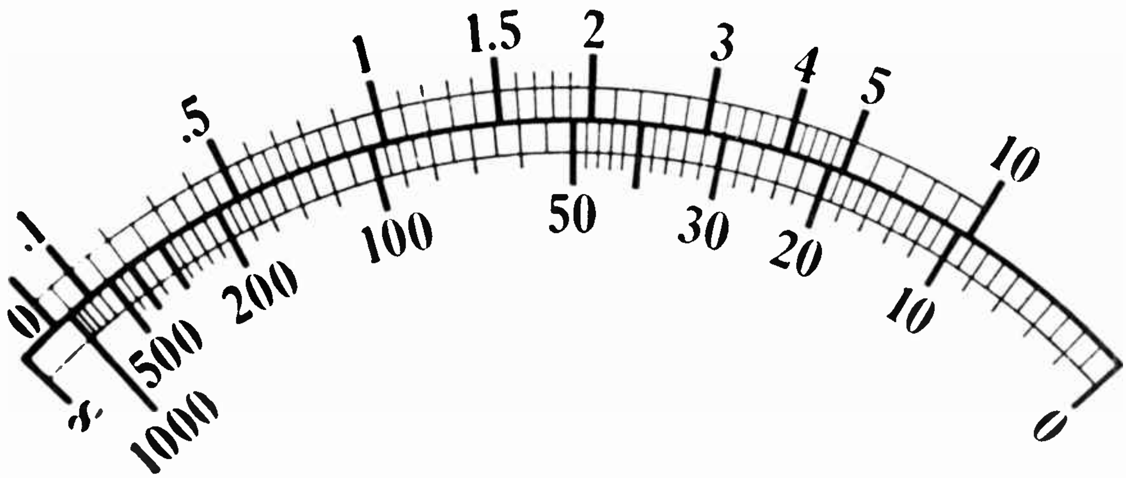


Fig. 71-8. A low-resistance ohmmeter scale (above) with a high-resistance scale (below).

6 the voltage is likely to wreck the ohmmeter. Another reason for disconnecting a measured part is that you never can be very sure of what other resistance elements may be parallel with the one you check, and when there are such parallel paths you would not be measuring resistance of the one unit.

When commencing to use an ohmmeter that has been idle for any length of time always short circuit the test terminals or test lead tips and use the zero ohms adjuster to bring the meter pointer to zero on the ohms scale. Make this zero setting also whenever you change from one range to another on a multi-range ohmmeter.

8 When resistance ranges allow it, take readings no higher on the ohms scale than about one-fifth of the maximum, which would be at about one-fourth of full-scale current. Then you will be using the ohms scale where it is fairly well spread out, and will be using the meter in the portion of its current range where there is likely to be greatest accuracy. Readings taken of resistances much greater than about ten times the center-scale value are likely to be misleading.

If the zero ohms adjuster will not bring the meter pointer to zero on the ohms scale while the test terminals are shorted, it is time to replace the dry cell or cells with new ones. They are too weak to be of further use.

As a final precaution, do not have your fingers or hands on both test terminals or both test lead clips or prods at the same time while making resistance measurements. Your body will act as a parallel resistance which, when measuring high resistances, would prevent accurate indications.

VOLT-OHM-MILLIAMMETERS. A volt-ohm-milliammeter is just what the name implies, it is a combination voltmeter, ohmmeter, and milliammeter. The rather long name usually is abbreviated to the letters "VOM". One such instrument is pictured by Fig. 71-9. There are many other styles which are equivalent in design and which serve the same purposes.

Fig. 71-9. A multi-range volt-ohm-milliammeter.

There is a permanent magnet moving coil meter of any sensitivity that suits the measurements to be made. By means of one or more rotary or press button switches, or with suitable pin jack terminals, the meter is connected into self-contained voltmeter circuits, current meter circuits, or ohmmeter circuits. Usually there are voltage and current measuring circuits for both direct and alternating quantities.

The several kinds of measurements may be called the functions of the instrument. The functions might include measurement of direct voltages, of direct currents, of alternating voltages, of alternating currents, and of resistances. For each of these functions there will be several ranges. For instance, we might have voltage ranges of zero to 2½, to 10, to 50, to 250, and to 1,000 volts, also similar current ranges in milliamperes. There may be one switch for selecting the desired function and another for selecting a suitable range, or both function and range may be selected by a single rotary switch, or possibly by a number of press button switches, or there may be separate pin jacks for functions and ranges. In addition there will be a zero ohms adjuster.

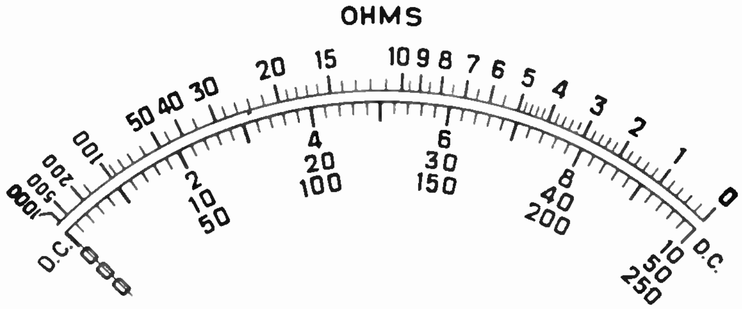


Fig. 71-10. A dial with single resistance scale and three scales for voltage and current.

The meter dial of the VOM may have a single scale for resistance and two or more scales for voltages and currents. Such a dial is shown by Fig. 71-10, where the single ohms scale is at the top and three direct voltage or direct current scales are down below. In practice there would be an additional set of voltage and current scales for measuring alternating quantities, which require different distribution of the markings. The pointer travels across all the scales at once. Which scale you read depends on the function and the range for which the instrument is set.

When there is only a single ohms scale the resistance range selector will be marked in some way to indicate the factor by which dial readings are to be multiplied. With the dial of Fig. 71-10 the several positions of the resistance selector or the several resistance pin jacks might have markings and interpretations as follows.

R X 1	Read the ohms scale directly as marked or graduated.
R X 10	Multiply every reading by 10.
R X 100	Multiply every reading by 100.
R X 10K	Multiply every reading by 10,000.
R X 1M	Scale markings are read as megohms of resistance.

There will be one selector position for each of the voltage and current scales appearing on the dial, and nearly always there will be additional ranges and selector positions with which readings on one or the other of the scales must be multiplied by a factor indicated by the selector. As an example, with the dial scales of Fig. 71-10 there would be selector positions for 10, 50, and 250 volts or milliamperes. In addition there might be a selector position for 500 volts or 500 milliamperes, with which you would read the 50 scale and multiply by 10. Another range might be for 1,000 volts or milliamperes, with which you would read the 10 scale and multiply by 100.

When controls of the VOM are set for voltage measurements the instrument is a voltmeter. Then you must use exactly the same methods and observe all the precautions which have been explained in connection with voltmeters. Similarly, when the VOM is set for current measurements you treat it just like a milliammeter, and when set for resistance measurements you work as with any ohmmeter. Remember to

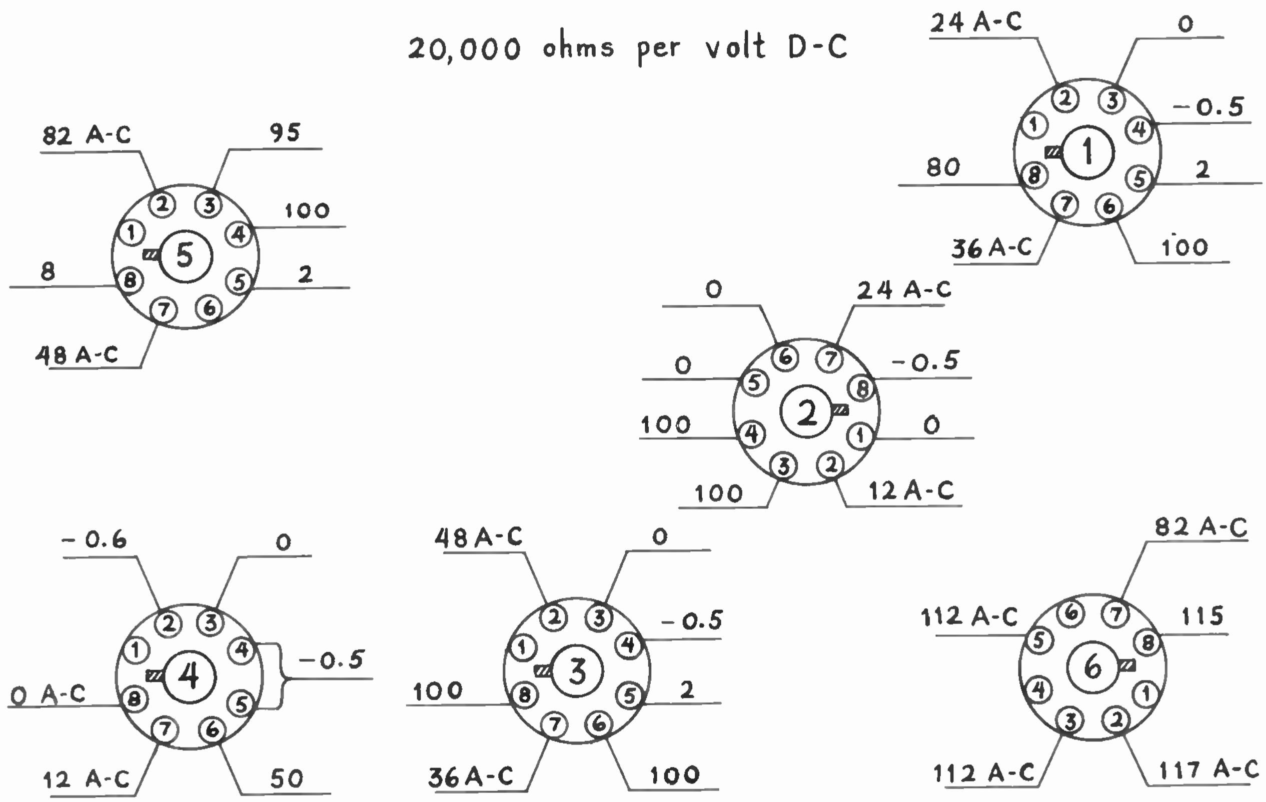


Fig. 71-11. Voltage diagram based on measurements with a meter of 20,000 ohms per volt sensitivity

commence a voltage or current measurement with the highest range, and drop to a lower range only when the indication is within that lower range. When measuring direct voltage or current it is advisable, when convenient, to cut off the power when changing from one selector switch position to another one. This avoids arcing at the switch contacts. One of the fairly common troubles with multi-range VOMs is contact roughness caused by arcing, or any kind of oxidation or dirt which increases the resistance of the switch contacts.

CHECKING VOLTAGES. The volt-ohm-milliammeter or a separate voltmeter and ohmmeter most often are used in connection with service diagrams or with tables of service information which show normal voltages and resistances as measured at tube base pins and at certain other easily identified points on a receiver. A voltage diagram for a six-tube radio receiver is illustrated by Fig. 71-11. A voltage diagram for a television receiver would be similar, except for having more tubes or sockets. On a voltage diagram are represented the socket lugs or tube base pins as they appear from underneath the chassis, in approximately the same relative positions as actually occupied.

Normal voltages are marked at each pin where a measurement may be made. Unless otherwise noted,

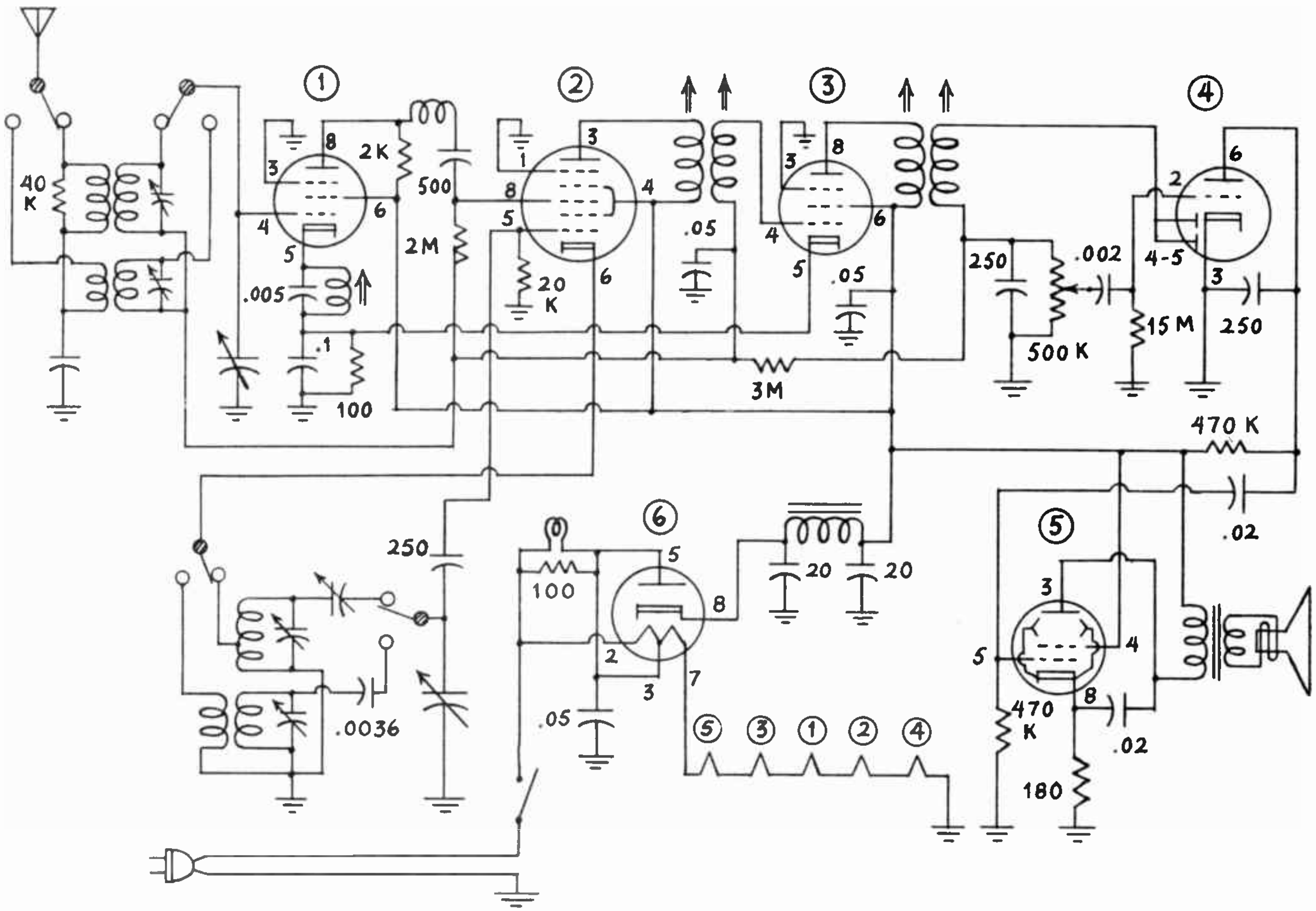


Fig. 71-12. Schematic circuit diagram for the receiver on which voltages and resistances are measured.

all voltages are measured from the pins to chassis ground, all are d-c voltages unless marked a-c, and all d-c voltages are positive at the pins unless marked negative.

Fig. 71-12 is a schematic service diagram for the same receiver to which the voltage diagram applies The tubes are similarly numbered from 1 to 6 on both diagrams, and base pin numbers are shown on both diagrams. The base pin numbers allow identifying the various tube element connections on the voltage diagram. Some voltage diagrams carry abbreviations for the tube elements to make this identification easier.

On the schematic service diagram are marked the values of all fixed resistors and capacitors, and of some adjustable units. Resistances to be read in megohms are followed by the letter "M", as 3M for 3 megohms. Resistances to be read in thousands of ohms are followed by the letter "K", as 20K for 20,000 ohms. Where no letter follows a resistance value it is to be read in ohms, as the number 180 followed by no letter meaning 180 ohms.

Capacitor values to be read in microfarads are shown by decimal fractions, as .002 for .002 mf. Capacitor values to be read in micro-microfarads are shown by whole numbers, not fractions. For instance, the number 250 near a capacitor symbol means 250 mmf. These are common practices for showing capacitance and resistance values on service diagrams, but they are not universally followed.

Voltages often are marked on the schematic service diagram for a receiver instead of being shown on a separate base pin diagram such as illustrated. Sometimes the same information is listed as in the accompanying table, which applies to the same receiver as the voltage diagram of Fig. 71-11.

Voltage measurements are made with all tubes in their sockets unless otherwise noted. Tubes sometimes are removed to prevent current flow in certain circuits while checking no-load voltages. Television service diagrams and tables usually specify the positions for all controls during tests. Similar information is given for some radio receivers. When no positions are mentioned, all the controls should be set as for normal reception.

SOCKET VOLTAGES

Tubes	Base Pins							
	1	2	3	4	5	6	7	8
1. R-F Amp		24 A-C	0	–0.5	2	100	36 A-C	80
2. Converter	0	12 A-C	100	100	0	0	24 A-C	–0.5
3. I-f Amp		48 A-C	0	–0.5	2	100	36 A-C	100
4. Det'r A-f Amp		–0.6	0	–0.5	–0.5	50	12 A-C	0 A-C
5. Output Amp		82 A-C	95	100	2		48 A-C	8
6. Rectifier		117 A-C	112 A-C		112 A-C		82 A-C	115

Notes: D-c volts measured with 20,000 ohms per volt meter.
A-c volts measured with 1,000 ohms per volt meter.
Power line 117 a-c volts, 60 cycles.

COMMERCIAL TRADES INSTITUTE

TELEVISION

Before commencing measurements you should check the power line voltage. Test voltages of service diagrams or tables are based on 117 line volts. If actual line voltage is lower or higher you will have to make some allowance for the difference. Many shops use voltage adjusting transformers which will furnish 117 volts to the receiver when actual line voltage is anything between about 95 and 130.

5 Service diagrams and tables always specify the sensitivity of the voltmeter used for obtaining the original average or normal values. Values shown on Fig. 71-11 are for use with a voltmeter sensitivity of 20,000 ohms per volt. The actual shunting error due to connection of the meter resistance across the measured points will vary with the range being used, since total meter resistance is the product of sensitivity and full-scale volts for each range. When measuring across very high resistances the indicated voltage is likely to be less on a low range than on a high range, because the meter resistance is less on the low range than on the high one.

Fig. 71-13 shows tube pin voltages as measured on our six-tube receiver with a meter having sensitivity of 1,000 ohms per volt. The a-c heater voltages are not shown here, since they would be the same as on the preceding diagram where a-c measurements are at a sensitivity of 1,000 ohms per volt.

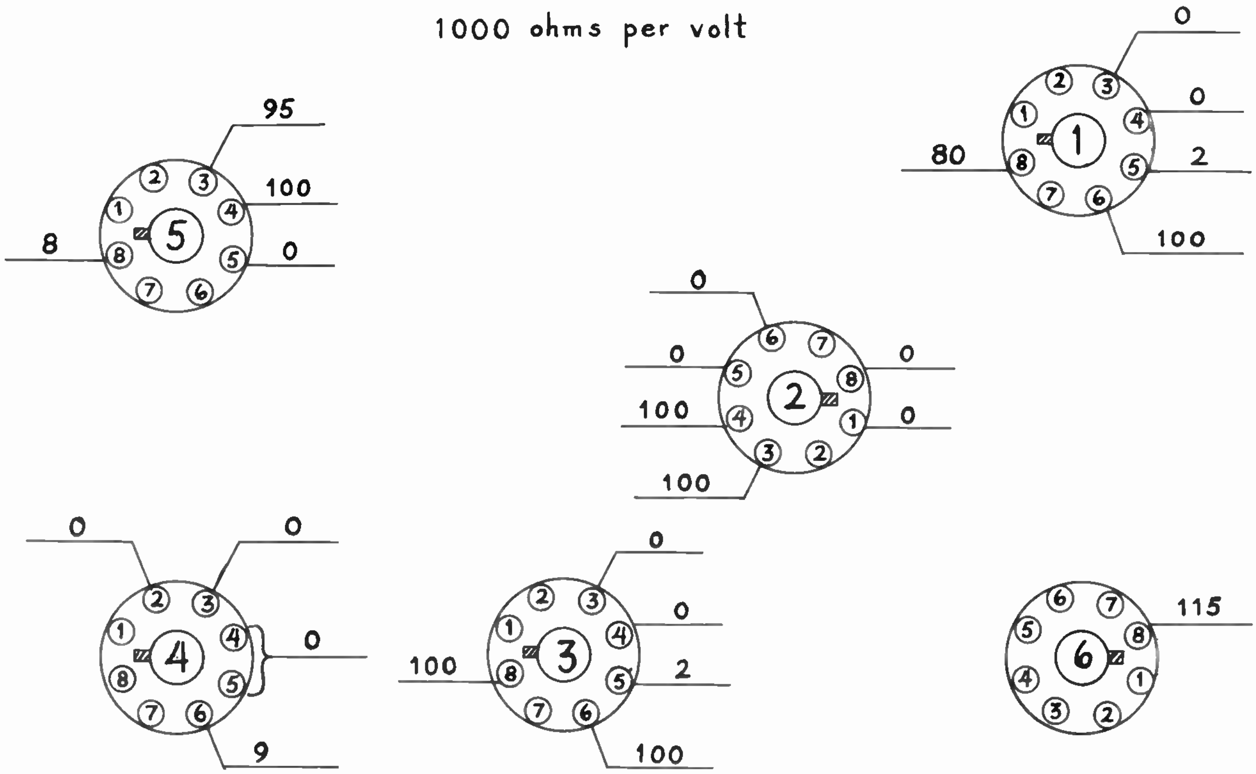


Fig. 71-13. Voltage diagram based on meter sensitivity of 1,000 ohms per volt.

With the meter of lower sensitivity all the control grid voltages are dropped to zero. These are the voltages at the following points. Pin 4 of tube 1. Pin 8 of tube 2. Pin 4 of tube 3. Pin 2 of tube 4. Pin 5 of tube 5. There is little if any use of attempting to measure control grid voltages or biases with a meter of low sensitivity. Even when using the higher sensitivity voltmeter the voltages at grids not connected to the avc line read lower than they actually are. For example, the grid of tube 4 actually is at about one-third greater negative voltage than shown by the 20,000 ohms per volt meter, the grid of tube 5 is at about 50 per cent greater positive voltage.

When measuring control grid voltages in any receiver having automatic volume or gain control there must be absolutely no signal coming into the receiver or else the automatic control must be disabled or overridden to provide constant grid voltage. Otherwise the grid voltage will vary with every change of received signal strength.

Note that the oscillator grid, pin 5 of tube 2, shows zero voltage with either voltmeter although this grid actually operates normally between 6 and 8 volts negative. This negative voltage is developed by grid-leak bias, which requires oscillating currents for its operation. There is so much bypassing capacitance in the leads and internal parts of any ordinary voltmeter that oscillation is stopped, and grid voltage actually does drop to very nearly zero.

Now compare the cathode voltages as indicated by the meters of high and low sensitivity at pin 5 of tube 1, at pin 5 of tube 3, and at pin 8 of tube 5. All these voltages measure alike with both meters. This is because all these cathodes connect to ground through resistances of less than 200 ohms, across which the shunting effect of either meter makes no measurable difference. The cathode of tube 2 grounds through the negligible resistance of the oscillator coils, while the cathode of tube 6 is directly grounded. In spite of the accuracy of cathode voltage measurements we cannot accurately determine the grid biases from the difference between cathode and grid voltages, because the measured grid voltages are incorrect.

Voltages do not have to be exactly as shown by diagrams and tables. Except in certain critical circuits, voltages which are as much as 20 per cent higher or lower than specified will not cause unsatisfactory operation of the receiver, and differences up to 10 per cent are to be expected in all cases.

When making any measurements, of either voltage or resistance, you must use care that the bare metal tip of a test prod or the metal of a spring clip does not touch any conductor other than the one at which the test is to be made. Spring clips used on the ends of test leads should be covered with rubber sleeves extending almost to the tips of the jaws.

CHECKING RESISTANCES. Another type of service diagram sometimes furnished for radio and television receivers is illustrated by Fig. 71-14. Again we have tube base pins or socket terminals in their approximate relative positions as seen from underneath the chassis, but now the marked values are those for resistance measurements with an ohmmeter or a volt-ohm-milliammeter.

Resistances are normal or average values as measured from the indicated points to chassis ground unless otherwise noted. Resistances in megohms are identified by the letter "M", those in thousands of ohms by the letter "K", and for resistances in ohms there is no letter after the value. The circuits in which the marked resistances exist may be traced on the schematic diagram of Fig. 71-12.

Resistance measurements may be used for an additional check when voltages appear to be incorrect, but they are even more useful when some circuit or section or the entire receiver goes completely dead, with no voltages to be measured. Resistance measurements are helpful also when application of line power causes overheating and smoking from parts which cannot be quickly identified and in which serious damage will occur unless power is immediately cut off.

1 Never make resistance tests while line power is applied to the receiver, always keep the power cord plug out of a receptacle. All tubes should be in their sockets unless otherwise noted. If all tubes are in good condition it should make no difference whether they are in or out of the sockets, but a defective tube might in some instances be located by a resistance check. All controls should be in the positions specified or, when no special instructions are given, the controls should be set as for normal reception.

When measuring very small resistances it is highly important that all the test connections are clean and tight. A thin film of any foreign matter may destroy all chance of accuracy. Blunt ended test prods often fail to make good contacts. It is better to use strong spring clips on both test leads or else to use a clip on one lead (to ground) and a needle pointed prod on the other lead. When measuring high resistances you must look out for leakage paths. Moisture may cause highly erroneous indications, although it takes a rather thick coating of dry dirty material to appreciably lower the resistance. Putting your fingers on both test lead connections is sure to cause difficulty.

A number of the resistances shown on Fig. 71-14 are marked with a star referring to the note that says, "Measure To Pin No. 8 of Tube 6". Should you measure these resistances to ground the ohmmeter pointer first will swing to a low reading and then will gradually move toward a reading of several hundreds of thousands of ohms. With the test connection to ground you would be reading the d-c resistance of the two 20-mf filter capacitors in the power supply, as will be evident from examination of Fig. 71-12.

All the test connections at which the ohmmeter might behave this way are to plates and screens, and all are on the B plus lines coming from the power supply. The indicated resistance increases as current from the ohmmeter charges the filter capacitors. Making the test to the cathode of the rectifier tube takes the filter capacitors out of the measured circuit. There may be similar behavior of the ohmmeter wherever large electrolytic capacitors are used, as in some cathode bypasses, in the loads of ratio detectors, and other places.

Incidentally, it is possible to make a fairly reliable test for leakage of electrolytic capacitors by connecting the ohmmeter from their positive side to ground or B-minus. The positive terminal of the ohmmeter must be connected to the positive side of the capacitor. If your ohmmeter is not marked for polarity, try both connections. The one with which the indication changes quite rapidly from lower to higher resistances is correct for testing. As a very general rule, if the resistance will not increase to more than a half megohm or to more than 500,000 ohms, there probably is serious leakage through the capacitor and it should be replaced.

When an electrolytic capacitor has been long idle it may take several minutes for its resistance to come

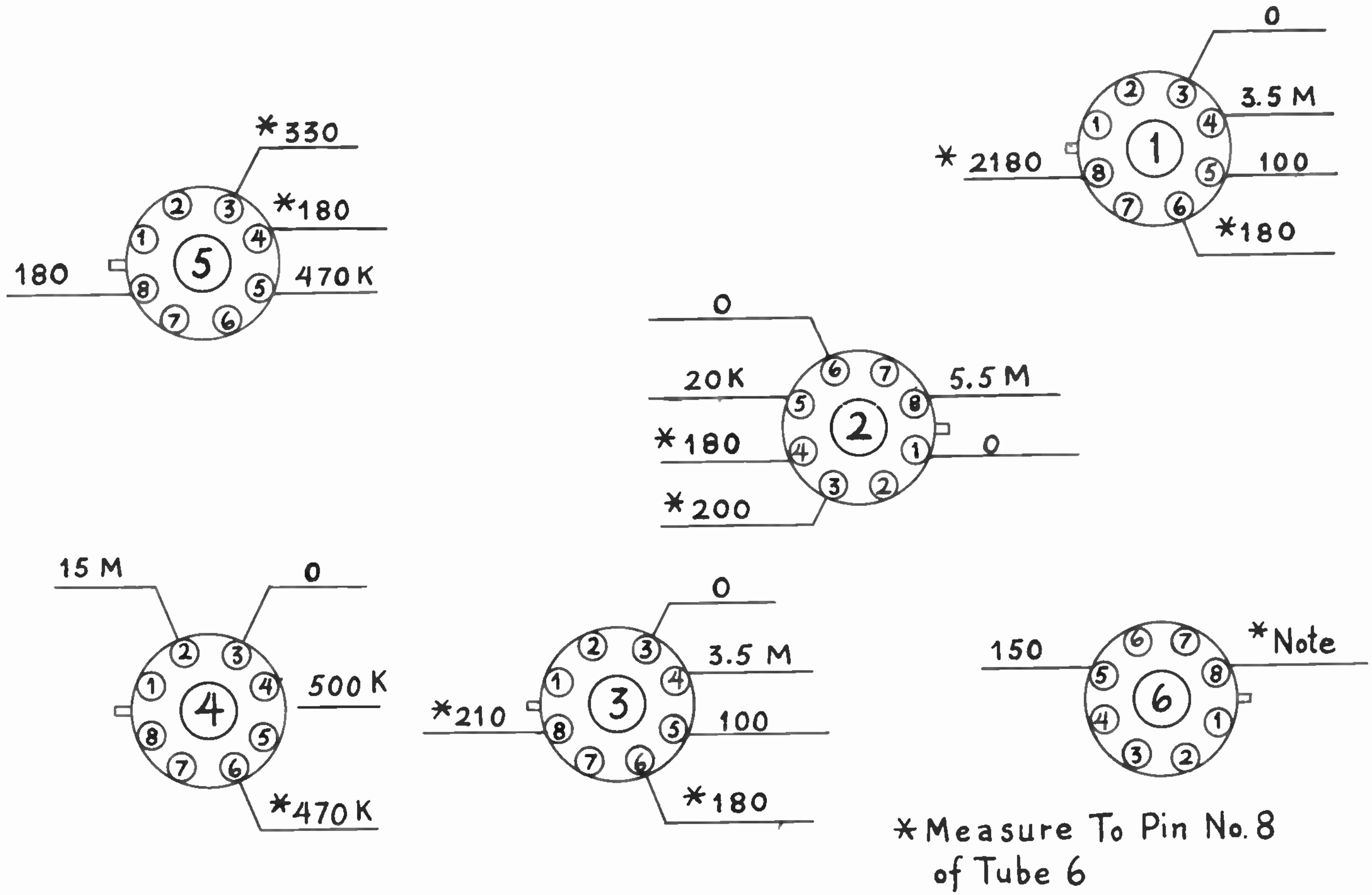


Fig. 71-14. A service diagram giving normal or average values of resistances.

up to a final high value. A leaky capacitor will show a slow rise of resistance. A capacitor in good condition, which has recently been used in a live receiver, usually will show a final resistance of several megohms.

Returning once more to our regular measurements of circuit resistances, it should be mentioned that measured values within 20 per cent plus or minus of marked or listed values usually are satisfactory. This is because most commercial resistors are of 20 per cent tolerance.

TELEVISION

LESSON NO. 72

CIRCUIT TESTING

"Training Your Mind"

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

COMMERCIAL TRADES INSTITUTE

Training Your Mind

C.T.I.

Perfecting Your Skill

Chicago, Ill.

LESSON NO. 72

CIRCUIT TESTING

Measurement of voltage and resistance is only one of the steps in trouble shooting, and it is never either the first step or the last one. When a receiver is in trouble no competent technician would first pull out the chassis, turn it upside down, and commence checking voltages and resistances. Unless he knows from experience with that particular kind of receiver what probably is wrong, he would begin by looking at the tubes.

When you can see the lighted heaters or filaments in glass tubes, and metal tubes are hot to the touch, the chances are fairly good that the tubes are working normally. If one tube is out, the obvious thing to do is replace it with one known to be good before doing anything else. If all the tubes are out you should make sure that the power cord is plugged into a live receptacle and that the receiver switch is turned on. If the set has series heaters or series filaments, and any one tube is open, it will put out all or a number of the others, and they will have to be checked one by one.

Before doing all this, or maybe right afterward, you would try moving all the controls that are accessible to the operator. Probably you would jar the receiver by striking the chassis with your knuckles. This might show that there is a loose or broken connection, by causing momentary sounds from the loud speaker or flashes on the picture tube. The point to all this is that you must use your head when commencing to shoot trouble. Make all the simple and easy tests first. If none of them give a clue to the fault, it is logical to proceed with checks of voltages and resistances.

When measured voltages or measured resistances show up as too great or too small you still have found only symptoms or results of some trouble. It remains to find the exact point or points at which the fault exists or to find the particular unit that has gone wrong.

4 If only one voltage or only one resistance is far out of line, your knowledge of the receiver circuits may allow immediate identification of the cause. Otherwise it always is a good idea to continue checking every voltage or every resistance. Then again you must apply common sense. For instance, if all d-c voltages are low, the trouble probably is in the B power supply. You should try replacing the power rectifier if it is a tube, or watch for excessive heating of a contact rectifier, and you should look for faulty resistors, poor connections, leaky filter capacitors, and such things.

5 When most of the voltages are within normal limits, with only one or two incorrect, it would be foolish to commence working on the power supply. You should confine your efforts to the sections or circuits in

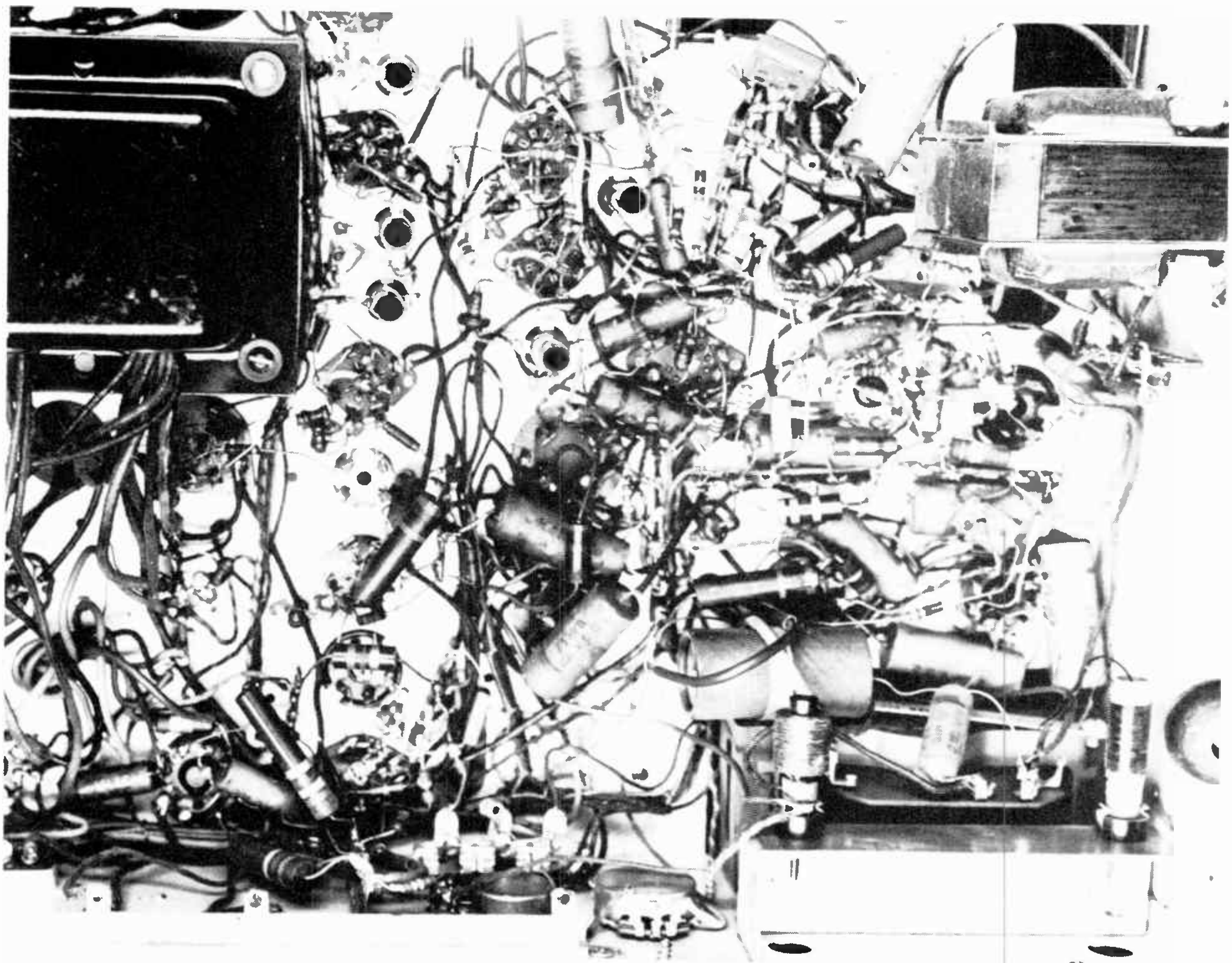


Fig. 72-1. Circuits like these are not too difficult to trace when you know how.

which incorrect voltages appear. If replacement of the tube or tubes in these sections or circuits fails to bring performance back to normal you probably have some real trouble on your hands.

Assuming that all the tubes as well as the power supply or voltage source are in good order, let's consider just a few of the things which may cause incorrect voltages or resistances. We shall commence with causes for low voltage.

Low voltage may result from a short circuit, often called simply a "short". This means any conductive path which allows current from the source to get back to the source without having gone through the normal load or through parts of circuits in which it should flow. We speak of a dead short circuit or a dead short when the path around the normal load is of negligible resistance, and of a partial short or a high-resistance short when there remains any considerable amount of resistance in the incorrect current path.

In Fig. 72-3 a short circuit is represented as existing in power supply filter capacitor Cf. Much of the current from the power supply will flow through this shorted capacitor and ground, completing a path from the power supply without going through the regular load circuits.

When a short circuit occurs between a live conductor and chassis metal it is called an accidental ground or simply a "ground". A ground is indicated on the cathode terminal of the amplifier tube in Fig. 72-3. Cathode current can flow through the ground connection instead of through bias resistor Rk.

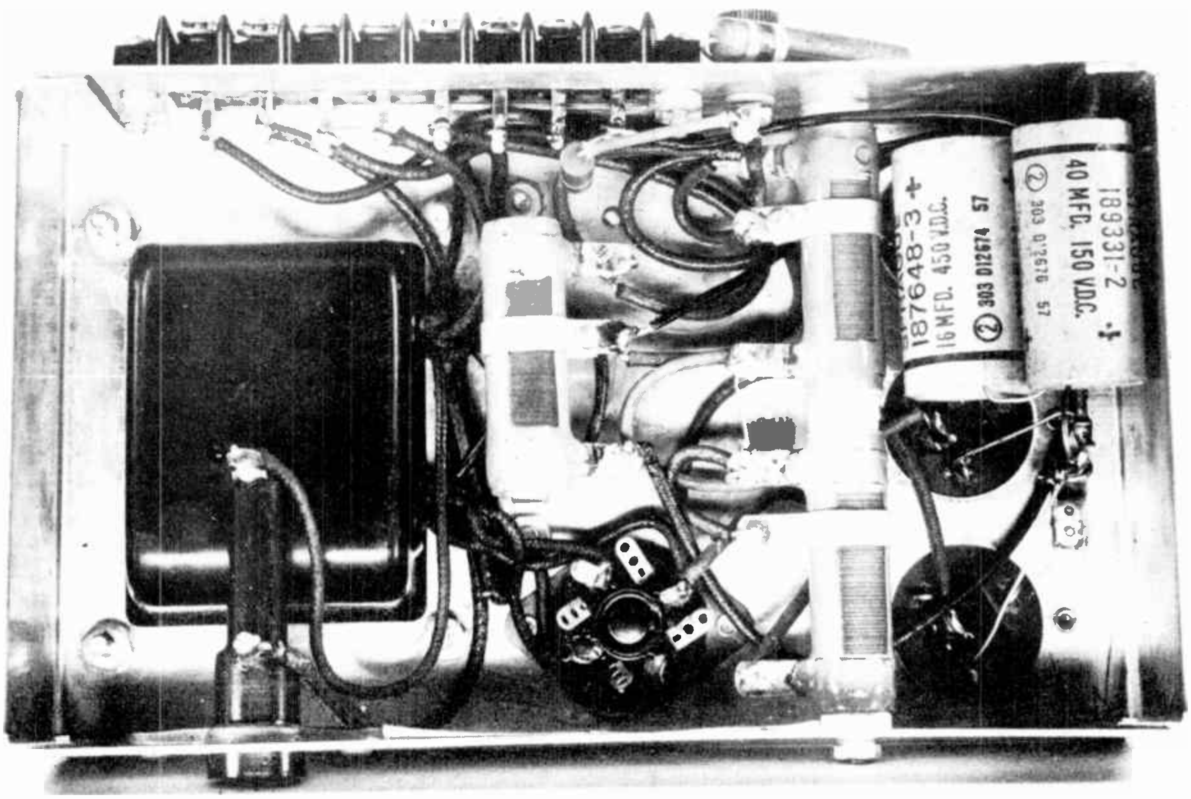


Fig. 72-2. A power supply may contain many resistors and capacitors in which troubles may develop.

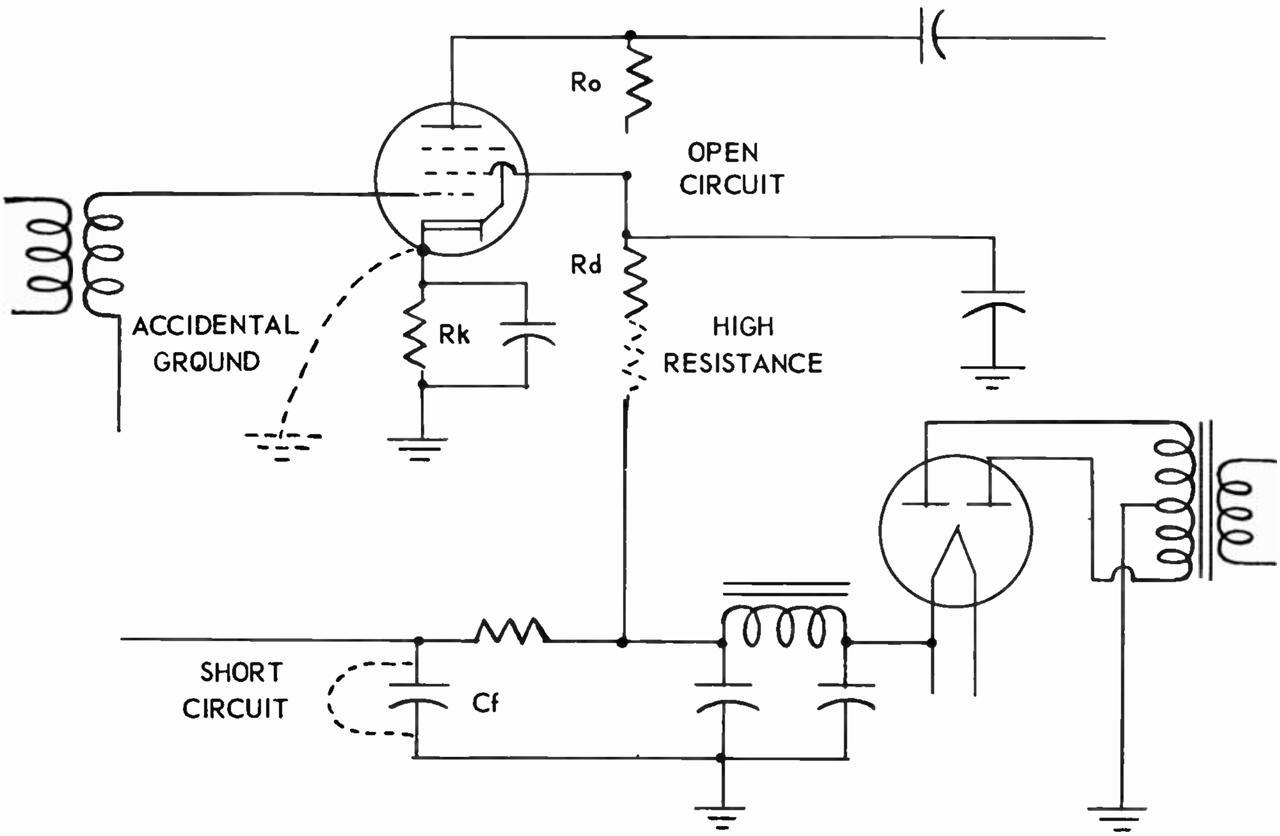


Fig. 72-3. Circuit troubles which might appear in B+ lines.

Shorts and grounds occur where insulation is broken, scraped away, or lacking for any other reason, and where bare conductors have been pushed out of place to make contact with other conductors. Shorts and grounds occur also through resistors which have been greatly overheated and whose resistance has decreased. These faults occur quite commonly through capacitors whose dielectric or internal insulation has broken down. As a general rule a capacitor which becomes slightly leaky quickly develops a nearly dead short circuit.

Most shorts and grounds, but not all of them, take excessive current from the source. Although the regular load or circuit may be getting little or no current, the source usually is heavily overloaded by current flowing through the short or ground. This is the reason why these troubles usually cause abnormally low voltages in all connected circuits, the excessive current is causing a large voltage drop in the internal resistance of the source.

A low voltage is often due to excessively great resistance in series between the high side of the source and the point at which low voltage appears. An open circuit is represented at the bottom of the plate load resistor Ro in Fig. 72-3. When a circuit is completely open there will be zero voltage on the side that is away from the source, as at the plate of the amplifier tube in our illustration.

2 Resistor Rd in the lead to the screen and plate of the amplifier tube is shown as having too great resistance. Someone may have misread the color coding and replaced a 10,000-ohm resistor with one of 100,000 ohms. In any excessively high resistance the voltage drop usually is so great as to leave abnormally low voltages at all points beyond the fault. In the present example, low voltage would be read at the amplifier screen. High resistance often is caused by such things as rosin joints or cold solder joints in wiring or by loose or dirty terminal connections.

Whereas an abnormally low voltage often indicates that there is an overload on the source, a voltage that is too high just as often indicates that the source is underloaded and is delivering less than normal current. This could result from either an open circuit or a very high resistance through which normal current cannot flow.

High voltage often is caused by a burned out or weak tube which no longer takes its normal plate and screen currents and acts like a partially open circuit to reduce the load. An open or high resistance in a resistor may cause high voltage at all points between the high side of the source and the point of trouble, and low voltage between this point and the low side of the source. An abnormally low voltage originally due to excessive current through a resistor may change to a high voltage when the resistor burns out to become an open circuit.

Since low voltages commonly result from the low resistances of shorts and grounds, it follows that resistance measurements in the same circuit are likely to indicate resistances lower than normal. Also, because high voltages so often result from an open point or an excessively high resistance in the measured circuit, resistance checks in this circuit frequently show abnormally high resistance.

CONTINUITY TESTING. The exact point at which there exists a short, an accidental ground, an open, or an excessively high resistance is not too hard to locate if we go about it in an orderly fashion rather than just stabbing while hoping for luck. The only thing we need in the way of a test instrument is something which will apply only very low voltage or will allow only very small current to flow in the

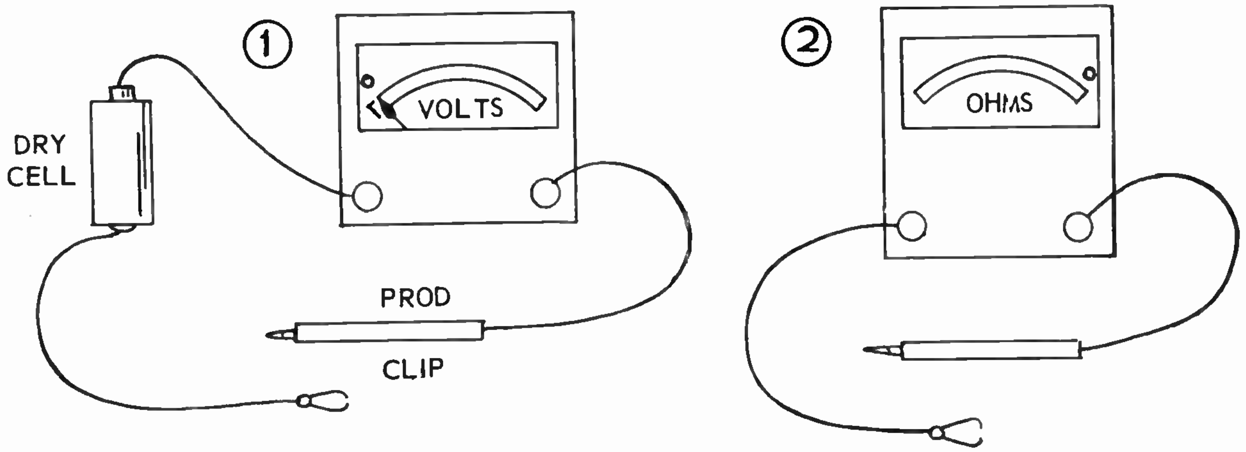


Fig. 72-4. Two forms of continuity testers.

tested circuit while indicating whether or not the circuit is continuous. Any device meeting these requirements may be called a continuity tester. The requirement of low voltage or small current is for protection of delicate coils or transformer windings which may be in the tested circuit.

A continuity tester may consist of a single dry cell in series with any low range voltmeter whose total resistance is 500 ohms or more, as shown at 1 in Fig. 72-4. Then the voltage applied to a circuit under test cannot be more than 1.5, and the current cannot exceed 3 milliamperes. Any ohmmeter with test prods, as at 2, makes an excellent continuity tester, since it will show whether the circuit is continuous or open and will indicate the amount of resistance between points tested. There are a number of specially designed continuity testers. However, lamps, buzzers, and other arrangements commonly used by electricians are not suitable for testing television and radio circuits. They operate with too great voltage, current, or both, and sooner or later will cause damage.

Now we shall examine some methods for locating faults with the help of a continuity tester, and other methods with which only a voltmeter is needed. Descriptions and diagrams of these methods serve to explain only the principles involved. You must understand television and radio circuits or be able to follow a circuit well enough to adapt the principles to individual cases.

LOCATING SHORTS OR GROUNDS. A short circuit or accidental ground usually places such a heavy load on the power supply and other parts that line power cannot be applied to the receiver without causing overheating and ruinous burning of circuit elements. To avoid such danger your continuity tester may be used as in Fig. 72-5. Here we have an accidental ground on the B+ line and also a shorted bypass capacitor from the amplifier screen to ground.

Either terminal of the continuity tester is connected to a cathode lug on the socket for the power

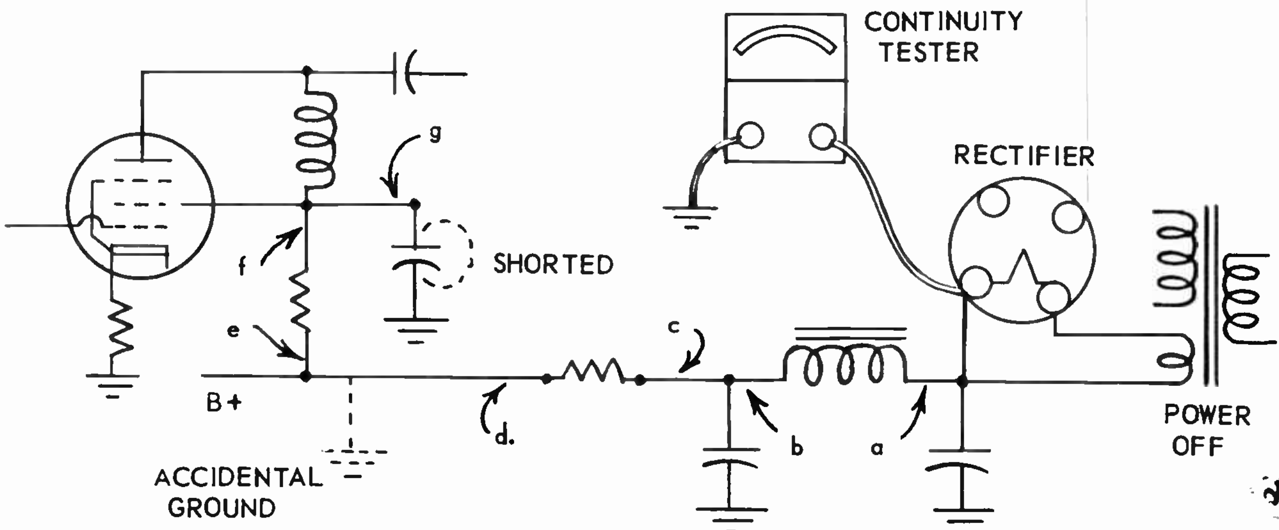


Fig. 72-5. Locating shorts and grounds with a continuity tester.

rectifier. The other tester terminal is connected to ground or to B-minus. The rectifier tube may remain in its socket, because line power is cut off. Current from the tester will flow through the accidental ground and the short circuit, and the tester will show this flow by indicating a complete circuit or a very low resistance.

The next step is to carefully lift or move all the wires and also all fixed resistors and capacitors supported by their pigtails in the suspected circuit. If you thus succeed in temporarily removing and thus locating a short or ground the tester will show that the trouble has been corrected, by indicating an open circuit or zero current. It is fairly easy to locate a single accidental ground or a single short on an exposed conductor in this way. But when there are two such faults at the same time one is likely to return while you remove the other. Also, this method will not locate shorts or grounds which are inside the parts in a circuit.

If you haven't located the trouble it becomes necessary to temporarily open the circuit at successive points, while working away from the tester, and watching for an indication of a closed circuit. The procedure is as follows.

First, temporarily disconnect conductor a. With faults as shown by Fig. 72-5 the tester now will indicate zero current or an open circuit. Reconnect the conductor a and temporarily disconnect conductor b. Still the tester will indicate zero current while the circuit is temporarily open. The same thing will happen with conductors c and d temporarily disconnected. But when you disconnect conductor e the tester will continue to indicate current or a complete circuit, because current actually continues to flow from the tester through the accidental ground on the B+ line.

When using this method a short or ground must exist somewhere between the last disconnection with which the tester indicates no current and the first disconnection with which the tester continues to indicate current.

After getting rid of the ground on the B+ line and reconnecting conductor e, the tester again would indicate the presence of a short or gronnd on the circuit being tested. Then you should proceed from e to f in the regular manner. At g you can go no farther without connecting both sides of the tester to ground, so the trouble must exist between g and ground, and must be in the shorted capacitor.

LOCATING OPENS AND HIGH RESISTANCES. An open circuit and a point of excessively high resistance may be considered as the same general kind of trouble, but in different degree, since an open circuit is merely an infinitely high resistance. Whether you can distinguish between the two varieties of trouble when making tests depends largely on the instrument being used as a tester and on how well you can interpret its indications. Often it makes no difference whether the fanlt is an open or a very high resistance so far as test methods are concerned, for when you locate the point at which the trouble exists it is relatively easy to determine just what has gone wrong. The chief object of your tests is to locate the exact point where something must be done.

6 Opens and high resistances may be located as illustrated by Fig. 72-6. If you use a self-powered tester, such as an ohmmeter or a self powered continuity tester, it is absolutely essential that line power be cut off the receiver. Otherwise you may use a high-range voltmeter with the receiver tnrned on as for regular operation. Either test instrument is connected across each part that is suspected of being in trouble.

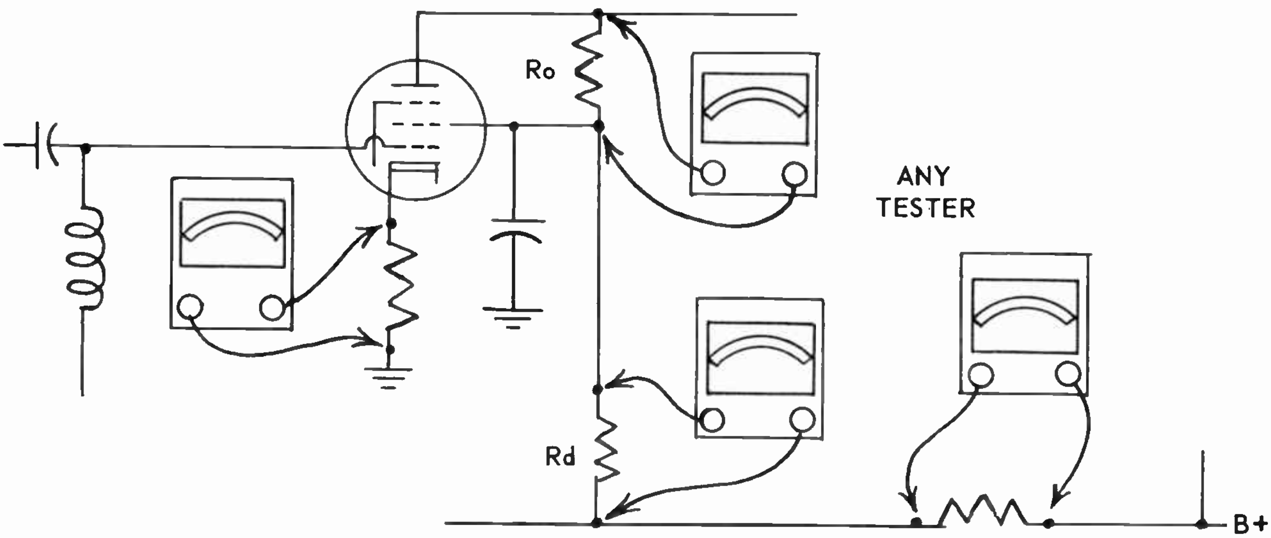


Fig. 72-6. Testing individual circuit elements for opens or high resistances.

A self-powered tester will indicate no current if the part is open, or high resistance if this is the kind of trouble that exists. A voltmeter will read higher when across an open circuit than when across a high resistance, and will read higher across an excessively high resistance than across a normal resistance. When the voltmeter bridges an open point the instrument indicates the no-load voltage of the source or of the circuit being tested, and when bridging any amount of resistance it indicates the voltage drop in that resistance while it is shunted by the meter.

To make these tests truly reliable, one end of the part or circuit being checked should be disconnected and left open when using any self-powered tester. When this is not done you are quite likely to be making tests across parallel circuits of which the suspected element is only one branch. Since the voltmeter depends on the power supply of the receiver for test readings it may fail or may indicate non-existant faults when there are two or more opens on the same line. For example, a voltmeter across resistor Rd of Fig. 72-6 will read high voltage to correctly indicate an open in this resistor, but across resistor Ro will read zero voltage because of the open circuit at Rd.

A self-powered continuity tester may be used for checking progressively along an entire circuit as shown by Fig. 72-7. Here we are working along a B+ line from the power supply to the plate of an amplifier. As in every case where a self-powered-tester is used, the a-c line power must be cut off the receiver to avoid possibility of ruining your test instrument. One end of the tested circuit must be opened, and one terminal of the tester connected to the opened end. This is conveniently done in the illustration by taking the rectifier tube out of its socket and connecting the tester to a cathode terminal of the rectifier socket.

With the other terminal of the tester connected to point a the indication will be of a closed circuit. This will be true also as we proceed to points b and c. But when the test connection is made to point

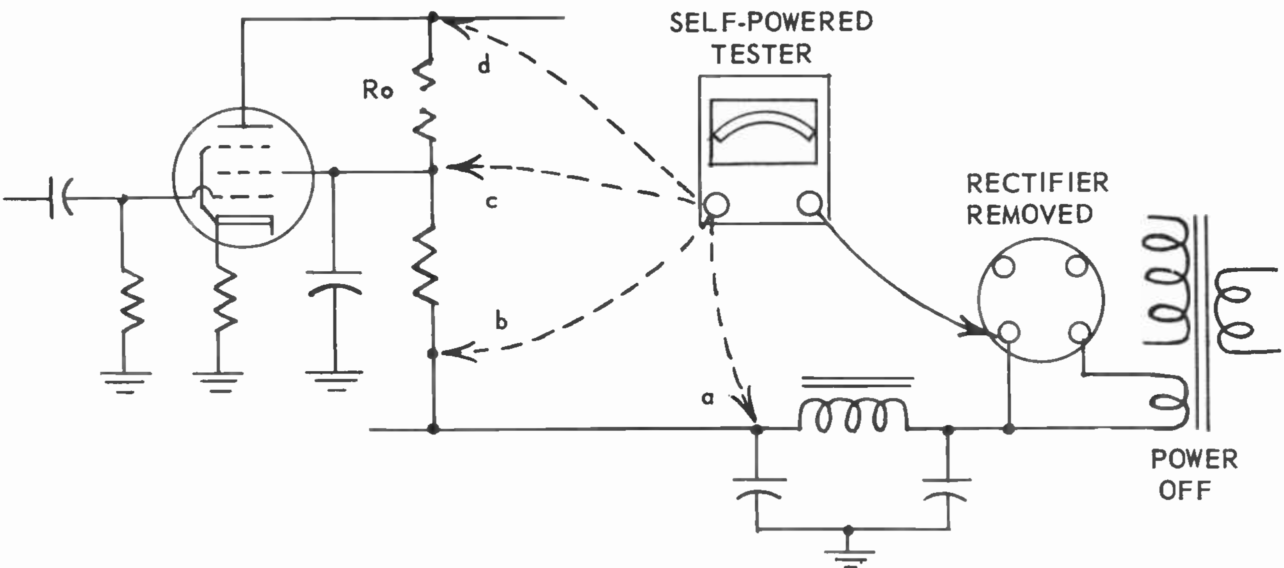


Fig. 72-7. Using a self-powered tester for locating opens and high resistances.

d the indication will be of an open circuit, which exists in resistor Ro. When using this method, the open circuit, or an excessively high resistance, will be found between the last test point allowing a closed circuit indication and the first point giving an open circuit or high resistance indication.

The particular test illustrated by Fig. 72-7 could not be extended beyond the B+ lines of the receiver. Any other section might be similarly checked by opening one end of the principal circuit and connecting one side of the tester to the opened conductor. Then successive tests would be made by working away from this connection toward the far end of the lines.

With a television receiver containing many sections whose B voltage is supplied through separate lines from the low-voltage power supply, the job of trouble location sometimes may be speeded up as illustrated by Fig. 72-8. The general method is the same as in Fig. 72-7, but instead of working progressively away from the power supply to the far end of each B+ line a preliminary check is made at the plate of only one tube in each major section of the receiver. For instance, you might check at the r-f amplifier in the tuner, at any one i-f amplifier, at any one tube in the sound section, and so on. If an open or very high resistance is indicated in only one section it is necessary to make progressive tests on only the B+lines in that section. But if trouble shows up in all sections the progressive tests should start at the power supply and go through the main B+ line that supplies all sections showing trouble.

Fig. 72-9 shows how a voltmeter may be used for progressively checking for open circuits along a B+ line, using the regular power supply as the source of testing voltage. The negative terminal of the meter is connected to chassis ground or to B-minus. The receiver power is turned on and the test prod, attached to the positive terminal of the meter, is touched to successive points as you work away from the power rectifier, as at a, b, c, and d in the diagram.

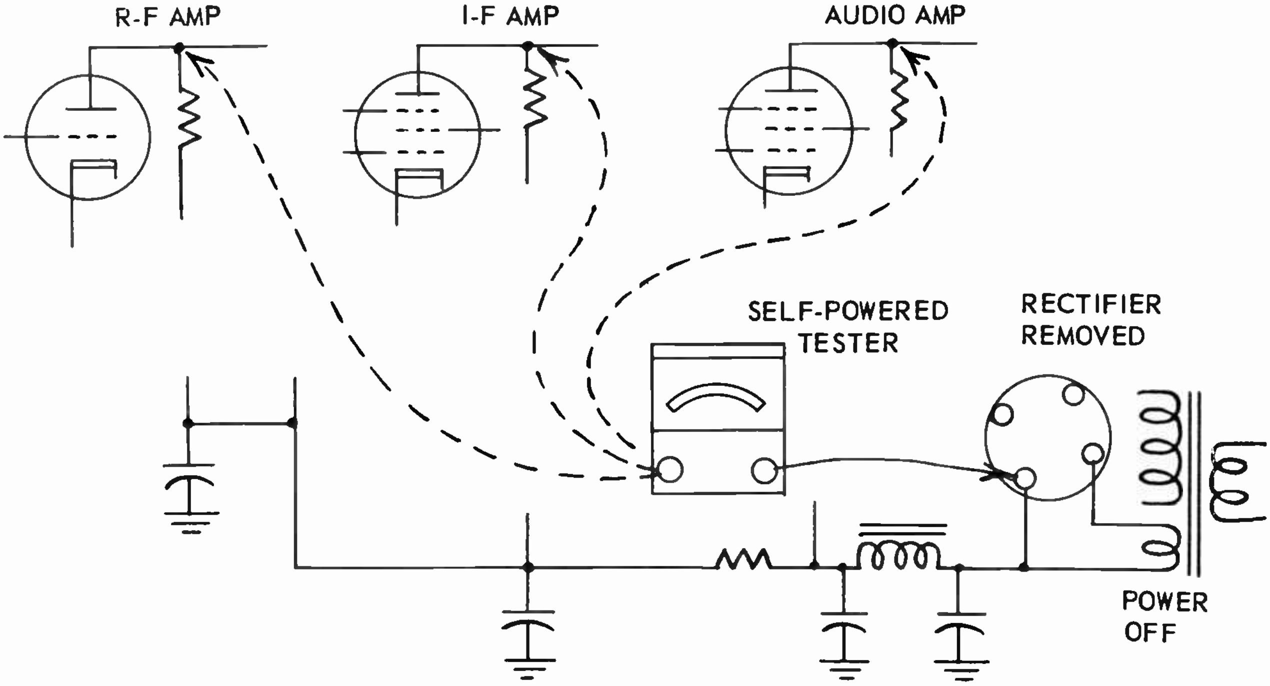


Fig. 72-8. One method of tracing trouble to some one section of a receiver.

The voltmeter will indicate high voltage at all test connections which are between the power supply and the first open circuit, as at a, b, and c. As soon as an open circuit has been passed the meter will indicate zero voltage. This would happen with a test connection at point d, because there is an open circuit in resistor Ro between test points c and d. This probably is the easiest of all tests for open circuits, but it will not indicate with certainty an excessively high resistance because enough current can flow through such a resistance to deflect the meter to about the same degree as with normal circuit resistance or no resistance at all.

TRACING CIRCUITS ON THE CHASSIS. It is easy to trace on a schematic service diagram, some circuit in which you wish to make tests for trouble, but to trace the same circuit through the actual parts and wiring on a receiver chassis is not quite so easy. To illustrate the difference we shall consider Fig. 72-10, which is a schematic diagram of the automatic frequency control circuit for a horizontal sweep oscillator. Here we have a small section of a complete receiver diagram.

Tube V1 is a sync amplifier which obtains its signal from a d-c restorer circuit and feeds sync pulses of opposite polarities to a phase detector tube, V2. The output of the phase detector goes to the input of tube V3, which is a multivibrator sweep oscillator. There are identification numbers for all the tube elements or pins, for all the resistors, all the capacitors, and for the winding of the frequency control. Values of resistances and capacitances which ordinarily would appear on a service diagram are omitted here for the sake of simplicity.

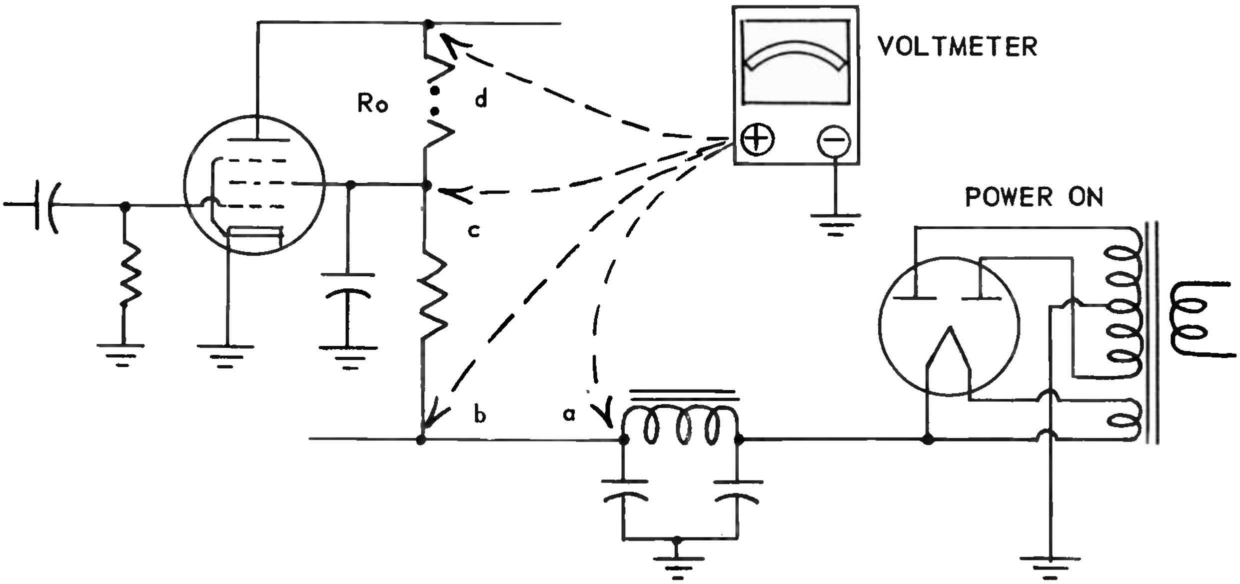


Fig. 72-9. Locating open circuits with a voltmeter and the B power supply.

The manner in which all these parts and connections are arranged on one small portion of the receiver chassis is illustrated by Fig. 72-11. Here everything is numbered to correspond with numbers on the schematic diagram. Of course, there would be no such numbers on the parts in a real receiver, instead you would have to identify each one by looking at its color coding and checking the indicated value with information on the schematic diagram or a parts list.

On the receiver there are two tie strips or terminal strips which provide the necessary junction points and supports for the various resistors and capacitors. These junction points are not the same as those shown between conductor lines on the schematic, rather they are chosen for convenience and for shortest connections in mounting and wiring the parts. The symbol X on both diagrams indicates that a wire goes from each of these points to one of the a-c heater voltage or current terminals of the power transformer.

If you trace any path on the schematic diagram it is possible to trace the same path on the chassis so far as current flow and voltage transfer are concerned. Also, you may examine any junction point on the chassis and find that on the schematic diagram the same parts are connected together by conductor lines.

During the innumerable times when you will trace through the wiring and parts on a chassis with the help of a schematic service diagram you will have to make use of easily identified starting points, from which it is possible to follow through the actual paths for current and voltage. Some of these identifications are listed by the following paragraphs.

1. Tube socket lug numbers or base pin numbers. These may be marked on the sockets, or you easily can count around in a clockwise direction from the locating key in octal and lock-in sockets, or from the

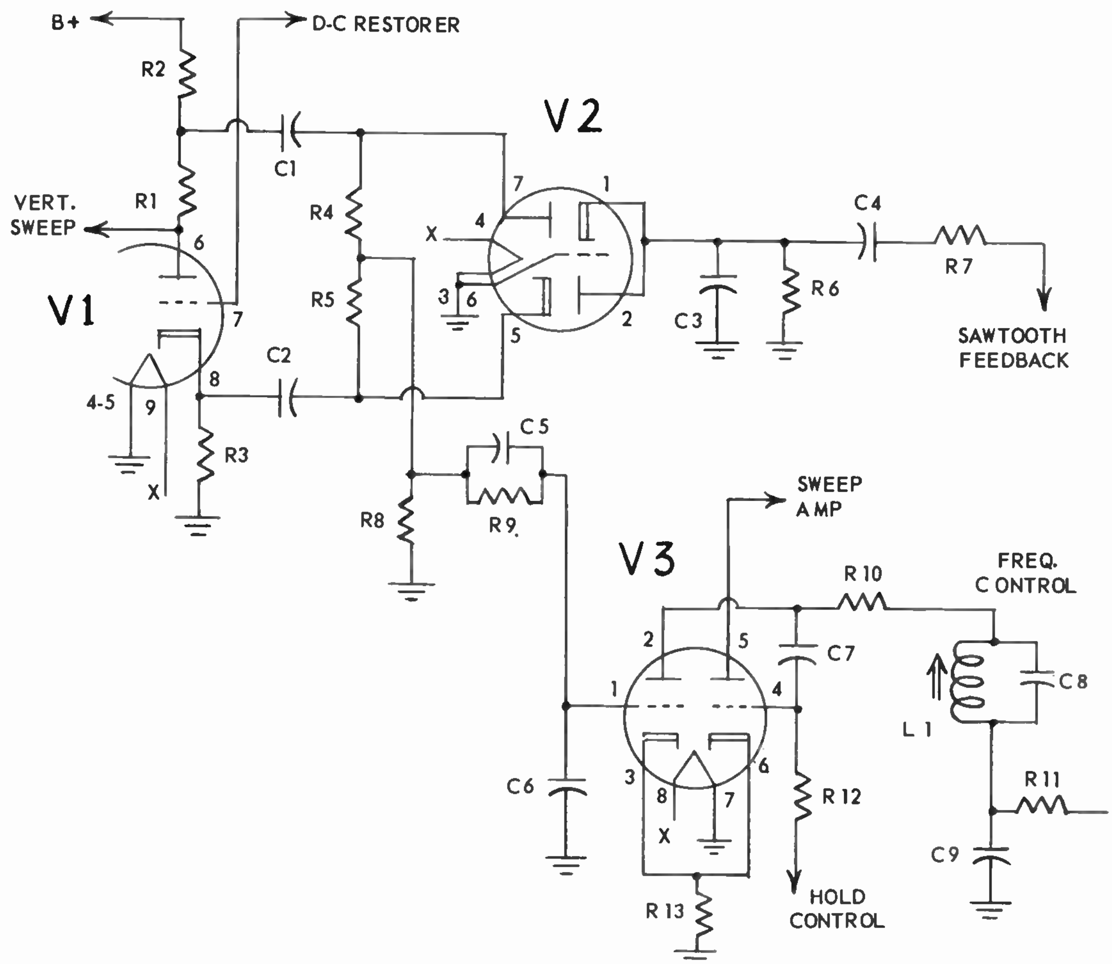


Fig. 72-10. A schematic service diagram for an automatic frequency control on a sweep oscillator.

space having no pin position on sockets for miniature tubes. The same numbers appear on the tube symbols of the schematic diagram, or you can find the correct numbers in any tube data book or list.

2. Terminals of control units, either the controls used by the set operator or those for service adjustments. These are especially helpful when working on television receivers, which have a great many such controls.

3. Terminals or openings for the leads of r-f, i-f, and a-f transformers and couplers. Sometimes these terminals are numbered or lettered on the unit itself and on the service schematic. In other cases the leads are of different colors which are identified on the service diagram.

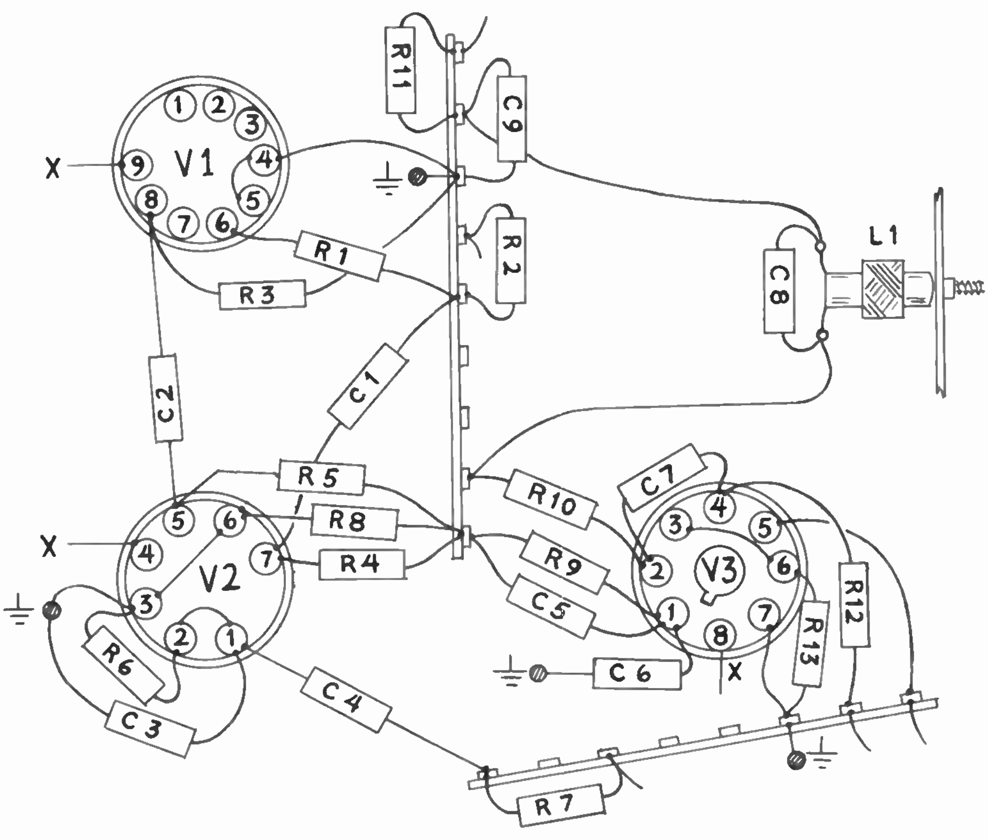


Fig. 72-11. Positions on the chassis of the parts and connections for the automatic frequency control.

4. The plug pins and the jack openings of connectors for multi-conductor cables used between various sections of some receivers, or for such parts as focus coils and deflecting yokes.

When you wish to locate some possibly defective unit, start from the nearest point that is easily identified. Then follow the wiring connections by eye, or by gently moving the connections with needle nosed pliers or with a small blunt ended prod of any kind. Most sets are wired with conductors having insulation of many different colors and color combinations. This makes it easy to identify the two ends of a wire connection. If a large number of leads are cabled together it may be necessary to loosen the cabling cord while tracing the circuits. Be sure to replace the cord when the job is finished.

CIRCUITS CONTAINING CAPACITORS. When checking voltages and resistances in circuits containing paper, mica, or ceramic capacitors you may run into indications which are confusing unless you keep in mind just how capacitors behave. If tests are being made with direct voltages from a power supply, or from an ohmmeter or other self-powered tester, every good paper, mica, or ceramic capacitor which is in series with the tested circuit will take a charge as the test connections are completed, then will act as an open circuit. The meter pointer may give a little jump, then go back to its open circuit position.

When using an ohmmeter in which the meter movement is of low sensitivity, and when the instrument is powered by only a single dry cell, this effect will be barely discernible with 0.25 mf of capacitance in the circuit, but will be decidedly noticeable with 0.5 mf or more. If the meter is of high sensitivity and is used on a range powered with 10 to 15 volts, the effect may be observed with capacitance as small as 0.005 mf. If the test connections are reversed, the capacitor may discharge through the testing meter and then take an opposite charge, making the pointer jump farther than before.

When you make tests with a sensitive voltmeter and a supply potential on the order of 250 volts d-c, the meter pointer will jump noticeably from zero with capacitance as small as 100 mmf in series with the circuit, but will immediately drop back to zero and remain there if the capacitor is not shorted.

Once you become familiar with the performance of some certain voltmeter, ohmmeter, or continuity tester it is possible to tell with fair certainty the condition of capacitors which are in series with the tested circuit. A momentary jump of the meter pointer and a return to the open circuit reading shows that the capacitor is neither shorted nor open. A continued reading indicates a shorted capacitor, while no jump at all may be an indication that the capacitor is internally open circuited or is disconnected. All this assumes that the capacitor is in series with the tested circuit and is of enough capacitance to make the meter pointer jump.

Earlier we noted that an electrolytic capacitor being checked with an ohmmeter will first show a rather low resistance, which quickly will increase to a much greater resistance if the capacitor is in good condition. If a high test voltage, as from a B power supply, is applied through a high sensitivity voltmeter to a circuit having an electrolytic capacitor in series, there will be a continued high voltage reading even though the capacitor is in perfectly good condition. With a low sensitivity voltmeter there will be a momentary high reading, but this will drop quickly to a relatively low reading if the electrolytic capacitor is in good condition. The longer you apply the test voltage the lower the reading will drop.

All this leads to the conclusion that tests with B power voltage are almost sure to be unreliable when there is an electrolytic capacitor in the tested circuit. Also, the high testing voltage may help to promote a leaky condition in electrolytics whose rated working voltage is much less than the B power voltage. Many power filter capacitors in transformerless sets are rated at only 150 d-c volts, and bypass capacitors or ratio detector load capacitors may have ratings as low as 10 d-c volts in many receivers.

To digress for a moment from the testing of circuits containing capacitors, there are likely to be confusing resistance indications when you check any circuit in which is a selenium rectifier such as used in the power supplies of transformerless receivers. Unless the positive voltage from the tester goes to the positive side of the rectifier the measured resistance will be low. When polarities are correct, the resistance normally will be several hundred thousand ohms in rectifiers of small current rating, but it will vary with the current rating and the condition of the rectifier. Rectifier resistance will vary also with the strength of voltage applied from the tester.

When it will not interfere with other tests, it is advisable to short out the selenium rectifier by making a direct connection from B plus to B-minus, or else to temporarily disconnect either side of the rectifier, keeping the power turned off.

Returning now to the subject of testing where capacitors are involved, it is well to remember some of the important facts relating to capacitances in series. Don't forget that any number of equal capacitances in series have combined capacitance equal to that of one unit divided by the number of units. Also, the combined capacitance of two unequal capacitances in series is equal to their product divided by their sum. That is,

$$\text{Series capacitance} = \frac{\text{capacitance A} \times \text{capacitance B}}{\text{capacitance A} + \text{capacitance B}}$$

Sometimes you will have on hand a capacitor of certain value and wish to determine what other capacitor can be connected in series to make a lower capacitance. To solve such a problem we need make only a slight change in the formula, like this.

$$\text{Required series capacitance} = \frac{\text{capacitance on hand} \times \text{desired capacitance}}{\text{capacitance on hand} - \text{desired capacitance}}$$

As an example, should you have on hand a 240-mmf capacitor and wish to obtain 80 mmf, the problem would be solved thus.

$$\text{Required series capacitance} = \frac{240 \times 80}{240 - 80} = \frac{19200}{160} = 120 \text{ mmf}$$

Always use the same unit for all capacitances in any of these formulas. Have everything in micro-microfarads or else everything in microfarads.

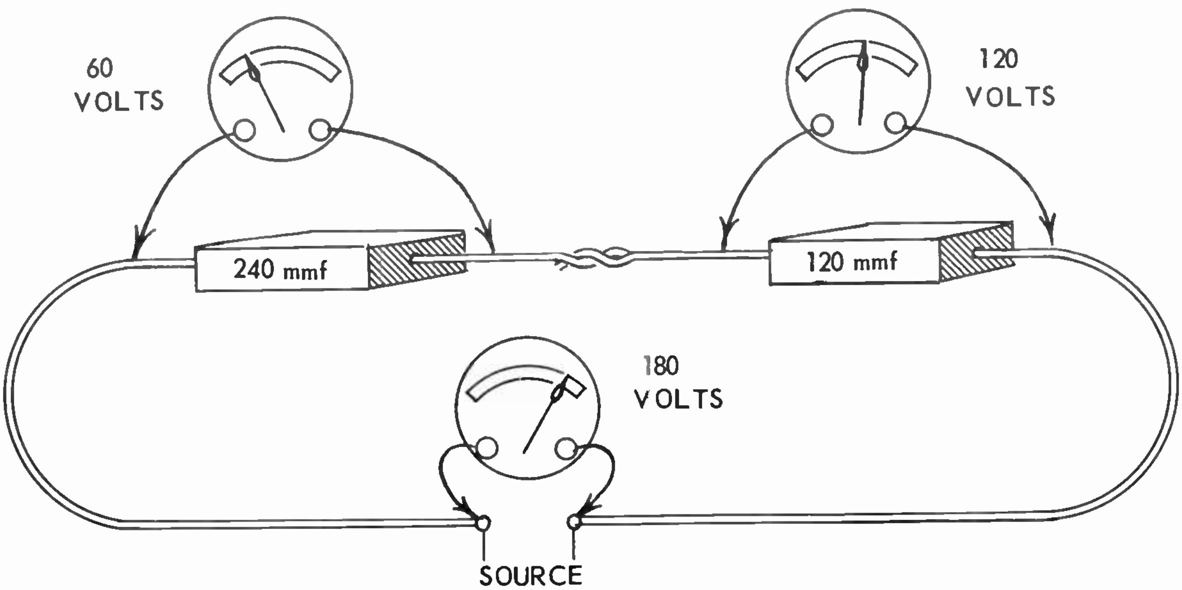


Fig. 72-12. How an applied voltage divides unequally between unequal series capacitances.

A Another thing to remember is that any voltage, either d-c or a-c, applied across unequal capacitances in series will divide inversely as the capacitances.

In Fig. 72-12 we have 180 volts across series connected capacitors of 240 mmf and 120 mmf. The unit having twice as much capacitance as the other one gets half as much voltage, while the unit having half as much capacitance gets twice as much voltage. With any number of series capacitors, the one with greatest capacitance always carries the least voltage drop, and the one with least capacitance carries the greatest voltage drop. By forgetting how voltage divides you quite easily may puncture a small capacitor which is in series with a larger one in a high-voltage circuit.

Division of capacitor voltages may help us out when we need a capacitor of higher voltage rating than anything on hand. Often it is possible to use two or even more than two lower voltage capacitors in series, Inasmuch as whatever voltage is applied to the series group will be divided between them, the effective working voltage becomes the sum of the voltage ratings of all the capacitors together - provided that all the capacitors are of the same capacitance and that all are rated for the same working voltage.

As an example, you might need capacitance of 0.1 mf to stand 600 volts, and have on hand only 400-volt units. By using two 400-volt 0.2 mf capacitors in series you will have double the working voltage, making it 800 volts, and one-half the capacitance, bringing it to the 0.1 mf that is needed.

The reason for using series capacitors of equal capacitances and voltage ratings is illustrated by Fig. 72-13. In diagram 1 we have 600 volts across series capacitors which each are of 0.1 mf capacitance and 400 working volts. The sum of the working voltages is 800, which is well in excess of the 600 volts applied. Each capacitor carries a drop of 300 volts, which is well within its rating.

In diagram 2 the working voltages still are alike for the two capacitors, and their sum exceeds the applied voltage. But the capacitances are not alike, one is three times the other, and the larger capacitor carries only one-third as much voltage drop as the smaller one. The smaller capacitor is being operated above its rated working voltage and is likely to puncture if this continues.

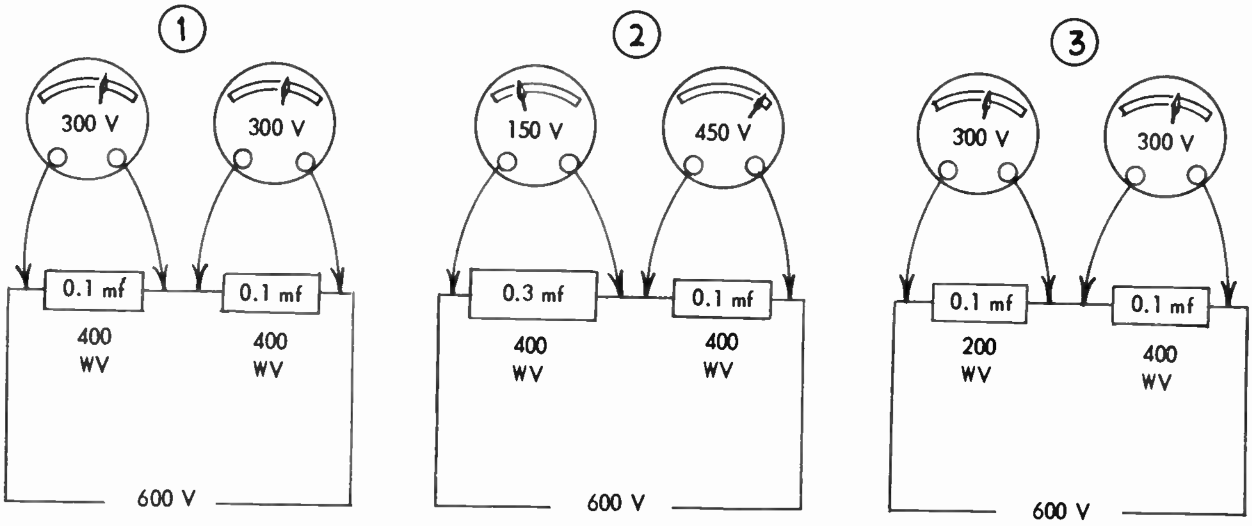


Fig. 72-13. How one capacitor may be overloaded even when the sum of rated voltages is more than the total applied voltage.

In diagram 3 we again have equal capacitances, but now the working voltages are not alike. Their sum is 600 volts, and only 600 volts is being applied to the two capacitors in series. But the capacitor rated at only 200 working volts actually carries a drop of 300 volts, which is half the total applied voltage, and almost surely will blow.

It would be possible to use small capacitors of high voltage rating in series with larger capacitors of relatively low voltage rating, and remain with in the rated limits of all the units. Then you would have to compute the inverse voltages, and a slip in arithmetic might mean the end of a good capacitor.

The actual division of a direct voltage between series capacitors can be measured only with an ultra-sensitive voltmeter, such as an electronic type, with a fairly high applied voltage, and with large capacitances. Even then only the initial momentary readings give fairly truthful indications. The measured voltage will drop almost immediately, because shunting of the meter resistance across the capacitor produces the effect of a very leaky capacitor. The voltage division across fairly large capacitances is easily measured with an a-c voltmeter of usual sensitivity when applying a-c voltage to the capacitors. Then the capacitors are being continually charged and recharged in opposite polarities.

Oftentimes you will need an a-c testing voltage lower than anything available from sources at hand, and when you have no suitable step-down transformer. Nearly always it is possible to obtain the lower

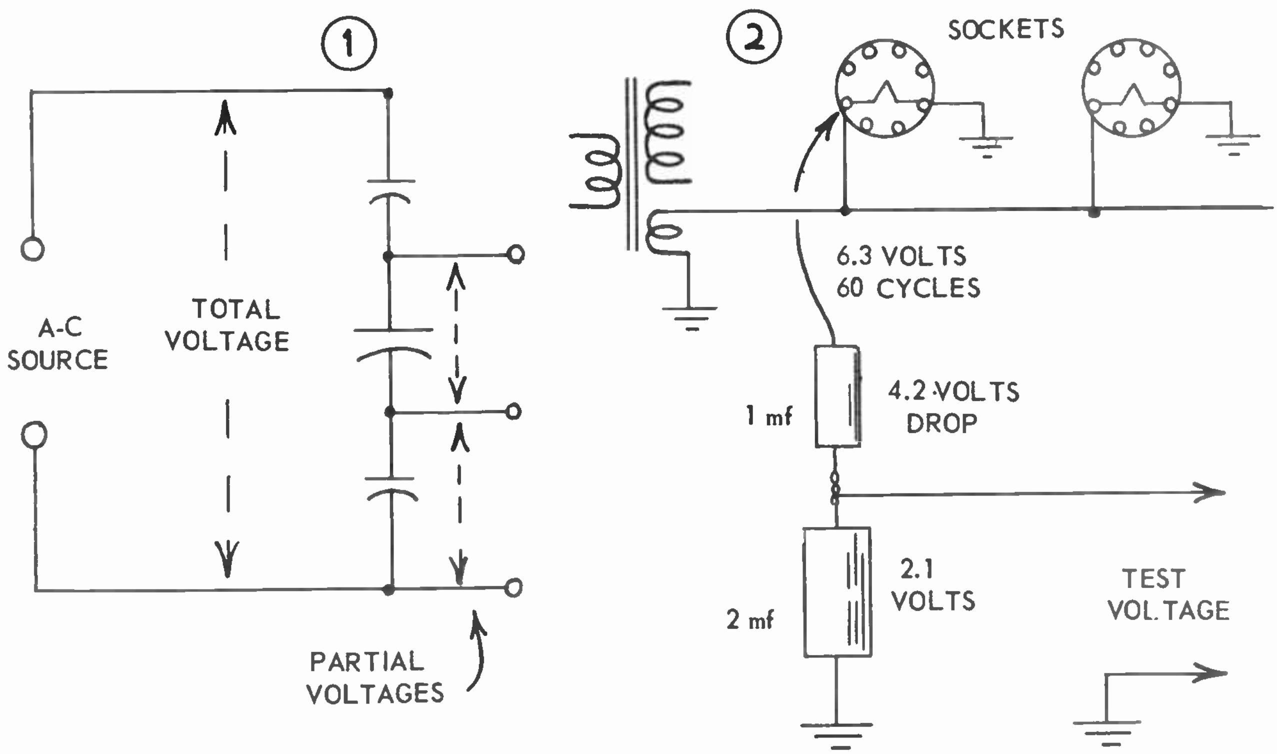


Fig. 72-14. A capacitance voltage divider may provide low a-c voltages for testing.

voltage by making up a capacitance voltage divider, as at 1 in Fig. 72-14, with two or more capacitors connected in series across the available source. Then any of the partial voltages existing across the capacitors may be employed.

Diagram 2 illustrates how 2.1 a-c volts may be taken from the 6.3-volt a-c heater circuit of any receiver. Two capacitors are connected in series from the ungrounded heater pin of any socket to ground. The heater voltage will divide between the two capacitors inversely as their capacitances. With the larger capacitor on the grounded end of the voltage divider the smaller testing voltage is taken between the top of this capacitor and ground. Any other test voltage less than 6.3 volts could be had by suitable choice of capacitances.

In applications like this it usually is necessary to give some consideration to the capacitive reactances of the capacitors. In the present case the supply frequency is 60 cycles. At this frequency the capacitive reactance of the 1-mf capacitor is about 2,650 ohms and that of the 2-mf capacitor about 1,325 ohms, for a total series reactance of around 3,975 ohms. The applied 6.3 volts will send about 1.5 milliamperes of alternating current through both capacitors at all times. Whatever test current is drawn will have to flow in the 1-mf capacitor in addition to the 1.5 ma, which will increase the voltage drop in that capacitor and will reduce the test voltage.

If no current were taken through the test connections we might have exactly the same division of voltage with two capacitors of 0.001 and 0.002 mf. Then the two reactances of 60 cycles would be respectively 2,650,000 ohms and 1,325,000 ohms. If the test leads then were connected to some low resistance circuit that circuit would attempt to take a large current. But the low resistance in parallel with the bottom capacitor would so decrease the impedance and resulting voltage drop that this voltage in the test leads would approach zero. Were the leads applied only across resistances of several megohms the voltage would remain around 2.0 even with the small capacitances in the divider.

TELEVISION

LESSON NO. 73

VACCUUM TUBE VOLTMETERS

Training Your Mind...

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind
C.T.I.
Perfecting Your Skill

COMMERCIAL
TRADES
INSTITUTE

Chicago, Ill.

LESSON NO. 73

VACCUUM TUBE VOLTMETERS

In a vacuum tube voltmeter the voltage to be measured does not act directly on the indicating meter, but is applied instead to the grid circuit of an amplifying tube. The resulting change of plate current in this tube then actuates a permanent magnet moving coil meter whose scale is graduated to read voltages applied to the grid circuit. The name "vacuum tube voltmeter" usually is abbreviated to "VTVM". This instrument is called also an electronic voltmeter, or it may be given various trade names by manufacturers.

10 The great advantage of the vacuum tube voltmeter over all other types employed for television and radio servicing is that the resistance of the instrument itself may be made exceedingly great, even on low ranges, while using a meter movement of low current sensitivity. For example, with a one-milliampere movement it is possible to have instrument resistance of 20 megohms on a 5-volt range. This means a voltage sensitivity of 4,000,000 ohms per volt. When measuring a 5-volt potential difference such an instrument would take only one-quarter of one microampere from the measured circuit, and as a consequence would give useful indications even where circuit resistances are very high or circuit currents very small.

The VTVM is the only service type voltage measuring instrument which may be used satisfactorily on almost any control grid circuit. It will measure d-c grid voltage on an r-f oscillator without stopping oscillation. Of course, the VTVM may be used with equal satisfaction on high-resistance plate and screen circuits, or on low resistance circuits, or anywhere else that any other type of voltmeter might be employed.

9 The usual type of vacuum tube voltmeter is designed primarily for measuring direct voltages. Most of these instruments have an auxiliary built-in rectifier which allows measurements of alternating voltages and relative a-c power levels at frequencies up through the audio range. For measurements at radio frequencies it is necessary to use an external detector probe, which we shall examine a little later. Many instruments have provision also for measuring resistances up to several hundreds of megohms, whereupon we have a vacuum tube volt-ohmmeter.

Fig. 73-1 is a picture of the under-chassis wiring of a vacuum tube voltmeter which will measure direct voltages in either polarity, will measure alternating voltages, and which provides for continuity testing. Fig. 73-2 shows the internal construction of an instrument having a self-contained power supply and provision for measuring resistances as well as either direct or alternating voltages in a number of ranges.

The basic principle of utilizing changes of plate current to indicate applied grid voltages may be em-

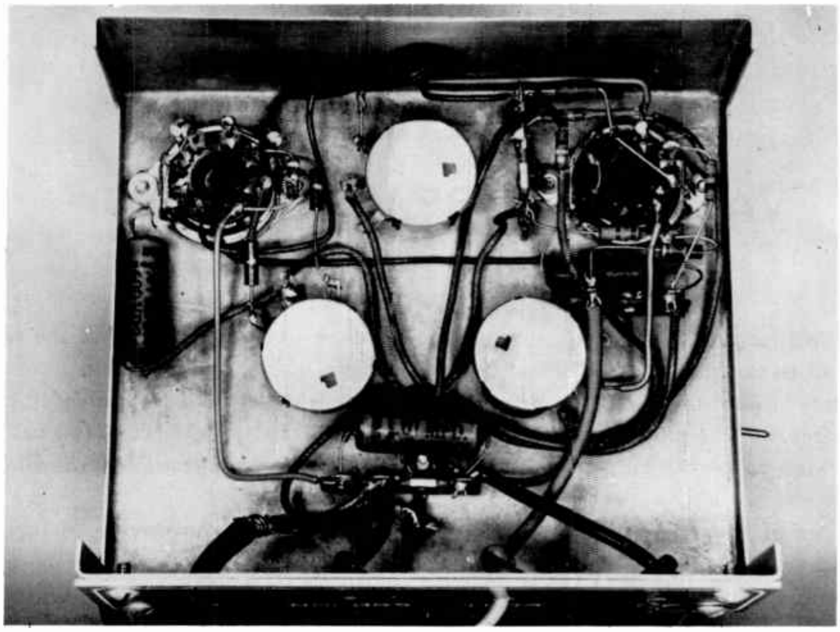


Fig. 73-1. Wiring for an ac-dc vacuum tube voltmeter, as seen from under the tube shelf.

ployed in practice by means of a wide variety of electrical and mechanical designs. Among formerly popular vacuum tube voltmeters were those of the grid rectification and plate rectification types, operating like detectors of these kinds. Another variety was the slide back VTVM, which compares the effect of an applied grid voltage with that of an adjustable measured battery voltage. These and other early designs require rather critical operating adjustments, such as for "bucking out" the normal steady plate current in order to read only the changes, and to compensate for variations of power supply voltage in their effect on measured plate current.

Most of the objections to the earlier types of vacuum tube voltmeters disappear when we use any of the many modifications of what is called a bridge circuit. The elementary principle of a bridge circuit for vacuum tube voltmeters is illustrated by Fig. 73-3. There is a twin triode tube, or there might be two separate triodes. The same B plus voltage is applied to both plates. Biasing resistors of equal values, Ka and Kb, are connected between the two cathodes and ground or B-minus.

If the two triodes have identical characteristics, and operate with equal grid biases, the equal plate voltages will cause both sections to carry equal plate or cathode currents. The two bias voltages are made alike in the following manner.

The grid of triode A is returned to ground through resistor Rg. Since there will be a negative bias from the voltage drop across resistor Ka, there will be no current and no potential difference in Rg, and the grid

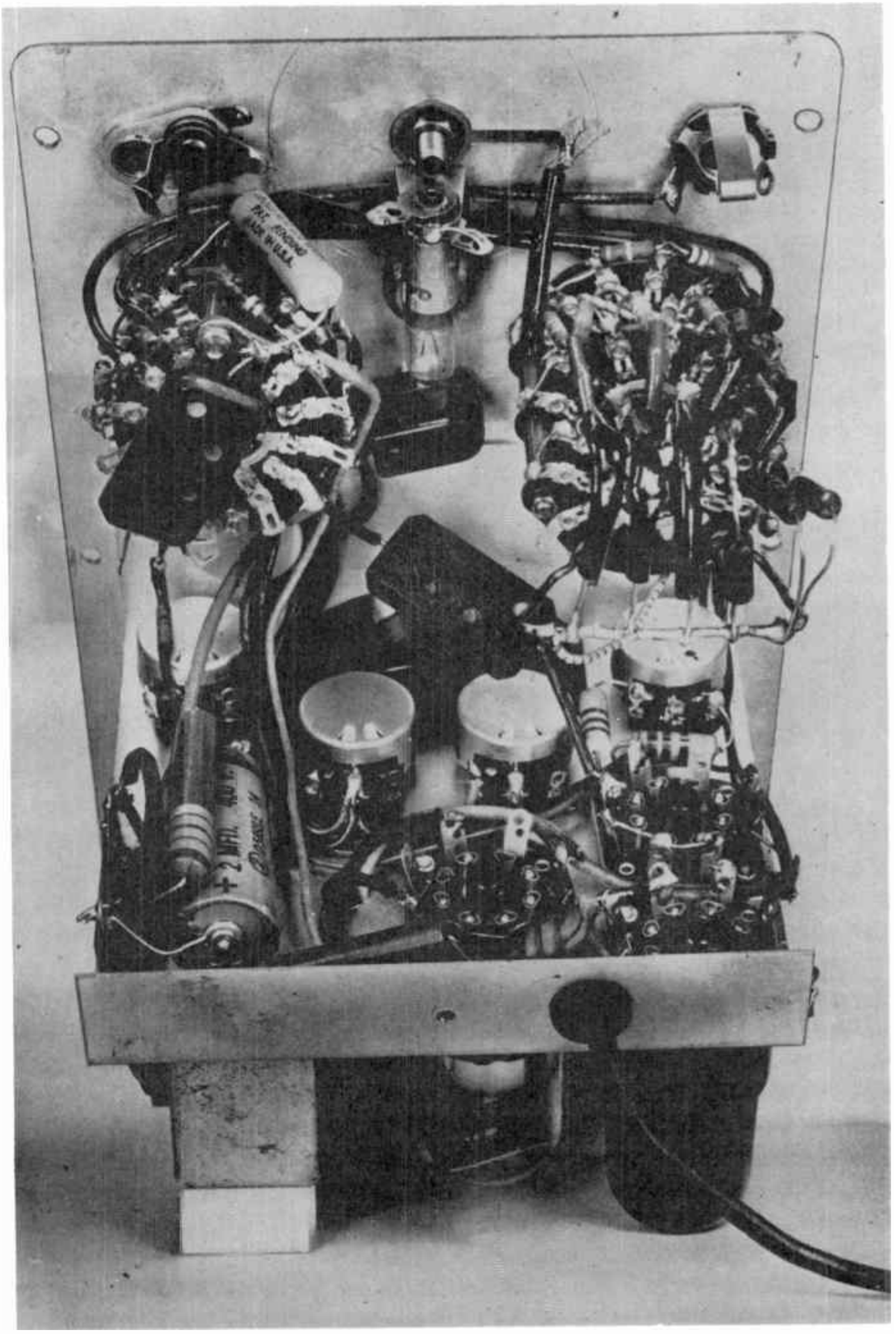


Fig. 73-2. Wiring and under-panel construction of a vacuum tube volt-ohmmeter with self-contained power supply.

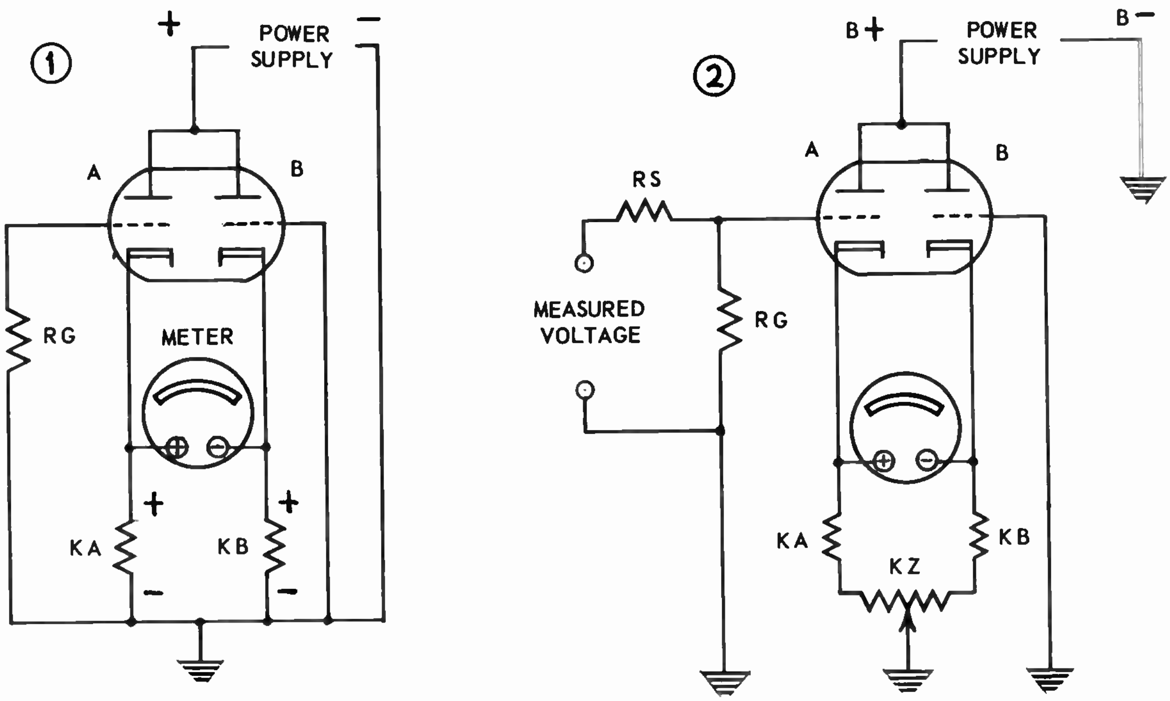


Fig. 73-3. The principle of a bridge circuit used for vacuum tube voltmeters.

of triode A will be at ground potential. The grid of triode B is returned directly to ground, so is also at ground potential. The cathodes of both triodes are at equal potentials above ground because equal plate currents in resistors Ka and Kb cause equal voltage drops between the cathodes and ground. Then, with both grids at ground potential, we must have equal grid biases.

We have established equal potentials at the two cathodes or at the cathode ends of the two bias resistors. The current meter is "bridged" between these points of equal potential. Then there is no difference of potential across the meter terminals and there is a reading of zero current.

The two triodes have been assumed as having identical characteristics, which seldom would be true of real tubes. To insure having equal potentials at both cathodes and a zero reading of the meter with ordinary commercial tubes we must provide, as in diagram 2, means for varying the cathode voltages and the drops across the biasing resistances.

Instead of connecting the lower ends of resistors Ka and Kb directly to ground or B-minus, this connection now is made through the outer ends of a potentiometer, Kz, whose slider connects to ground. Moving the slider one direction will increase the resistance in series with the cathode of one triode while reducing it on the other triode, and vice versa. The slider may be moved to a position where voltage drops in the two biasing resistances are such that both cathodes are at equal potentials, even though the two triodes have slightly different characteristics. This provides the desired zero difference of potential across the meter, and in the added potentiometer we have an adjustment for bringing the meter to an initial reading of zero.

In diagram 2 of Fig. 73-3 we have added also two terminals to which may be applied a voltage to be measured. This voltage will act through series resistor Rs and grid resistor Rg. The portion of the voltage which appears across Rg is in the grid circuit of triode A. If the measured voltage is positive at the upper terminal and negative at the bottom one it will make the grid of triode A more positive, or less negative, with reference to the cathode. Then triode A will carry an increased plate current or cathode current. There will be an increase of potential at the top of resistor Ka and at the positive terminal of the meter.

Application of the measured voltage has no effect on triode B, whose grid remains connected directly to ground. Plate or cathode current remains unchanged in triode B, and there is no change of potential at the top of resistor Kb or at the negative terminal of the meter. Now, entirely as a result of voltage applied at the measuring terminals, there is a potential difference across the meter and it will carry a corresponding current to make the pointer deflect. The meter scale is graduated to indicate the values of voltages applied to the measuring terminals.

It takes only a very small change of grid voltage on triode A to materially alter the plate current and the voltage drop across resistor Ka and part of Kz. The resulting difference of potential across a meter having only moderate current sensitivity results in a very considerable meter current, and there is a large deflection of the pointer. Since a small measured voltage causes a large deflection of the meter pointer we have here a highly sensitive voltmeter.

The only current taken from the measured circuit is that flowing through series resistors Rs and grid resistor Rg. Resistance at Rg may be as great as will not interfere with operation of the tube, and resistance at Rs may be much greater. The combined resistance of the two units across the measured voltage may be many megohms, so great that current through them is held to a fraction of a microampere. This is the only load placed on the measured circuit.

A-C MEASUREMENTS. The vacuum tube voltmeter which will measure direct voltages may have a rectifier connected ahead of its bridge circuit. Then the d-c output from the rectifier will cause deflection of the meter pointer proportionately to the value of alternating voltage applied to the input of the rectifier. The same bridge circuit is used for measuring either direct or rectified alternating voltages.

Fig. 73-4 shows the added parts which allow our d-c vacuum tube voltmeter to also measure alternating voltages when the function switch is moved from the d-c to the a-c position. To the grid circuit of bridge triode A we have added capacitor Ca, which bypasses to ground any slight alternations of voltage which may get through the preceding circuits. We also have rearranged resistors Rs and Rg for better filtering of such voltage alternations.

The added tube is a twin diode, which is our rectifier. Note that the cathode of diode 1 and the plate of diode 2 are connected to ground. While alternating voltage applied at the a-c input terminals is momentarily of such polarity as to make the upper terminal positive and the grounded terminal negative, the grounded cathode of diode 1 is made negative with reference to its plate, and this diode can conduct. But

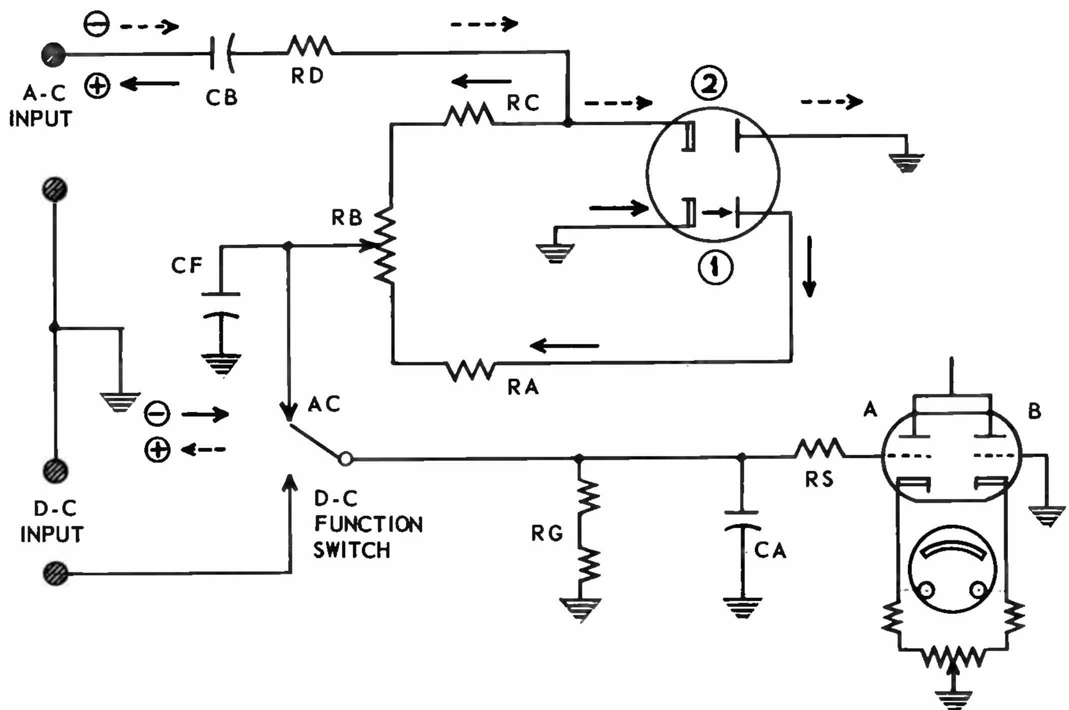


Fig. 73-4. Addition of a rectifier system for measurement of alternating voltages.

the grounded plate of diode 2 is made negative with reference to its cathode, and this diode cannot conduct.

While applied alternating voltage is of this momentary polarity the resulting electron flow is as shown by full-line arrows. The flow proceeds from the grounded input terminal to the cathode of diode 1 and through this diode to its plate. Thence the flow goes through resistor Ra, potentiometer Rb, resistors Rc and Rd in order, and to one side of capacitor Cb. From the other side of Cb the flow is to the upper input terminal. At the slider of potentiometer Rb the potential becomes positive with reference to the cathode of diode 1 and to ground. This positive potential is applied through the function switch to the grid circuit of triode A in the voltage measuring bridge.

When polarity of the applied alternating voltage reverses, there is reversal of polarities at the plates and cathodes of the diodes. The cathode of diode 1 and the plate of diode 2 becomes positive with reference to their opposite elements. This prevents conduction in diode 1, but allows conduction in diode 2. Electron flow now is as shown by broken-line arrows. This flow proceeds from the upper input terminal to one side of capacitor Cb, and from the other side of Cb through resistor Rd to the cathode of diode 2. Flow goes through this diode to its plate, thence through ground back to the grounded input terminal.

COMMERCIAL TRADES INSTITUTE

TELEVISION

Capacitor Cb is a blocking capacitor whose purpose is to prevent any d-c component of the measured a-c voltage from entering the rectifier circuit, so that readings of the bridge meter will be only those corresponding to measured alternating voltage. Were you to short circuit capacitor Cb, then apply positive direct voltage to the upper a-c input terminal, this voltage would cause electron flow through diode 1 and the bridge meter would deflect. The meter reading would not correctly indicate the applied voltage, because it is not calibrated for this manner of operation.

The purpose of capacitor Cf is to provide additional bypassing or filtering of voltage alternations, and keep them from going to the d-c bridge circuit. This capacitor, together with capacitor Ca and resistors Rg and Rs, provide quite complete filtering for voltage alternations and insure that practically smooth direct voltage reaches the grid of bridge triode A.

CONTACT POTENTIALS. Probably you have questioned why we need diode 2 and adjustable potentiometer Rb of Fig. 73-4, since neither of these seems necessary in rectifying the measured alternating voltage. Well, we need the second diode to furnish a contact potential which counteracts the contact potential developed in diode 1, and we need the a-c balancing potentiometer Rb to balance the two contact potentials against each other.

S The question now becomes, what and why is a contact potential? Here is the answer. Any element of a tube which is in the same evacuated space with a heated cathode collects some of the negative electrons which are emitted from the cathode into the space charge. If these electrons cannot readily escape from the element it acquires a negative charge. This negative charge gives rise to a so-called contact potential on the element. Although there is no physical contact between the cathode and the other element, we speak of a contact potential because this phenomenon first was observed between different metals in actual contact or conductively connected.

Depending on the type of tube, its age, the cathode temperature, and the resistance in series with an element, the contact potential may be almost anything from 0.02 volt to more than 1.5 volts. A contact potential can be measured only with a high-resistance vacuum tube voltmeter. It so happens that the d-c portion of the instrument that we are examining will measure the contact potential on the elements of the diodes in the a-c section. The measurement could be carried out as follows.

With the function switch in its d-c position, as in Fig. 73-5, and no connections to the a-c input terminal, the a-c section is electrically isolated. If the instrument is turned on and the ungrounded d-c input terminal connected to the plate of diode 1 while the diode cathode is hot, the contact potential at this plate will be found negative with reference to the cathode and ground. The contact potential will be fairly high because there is no conductive path from the diode plate to ground other than through the temporary testing connection, and most of the collected electrons remain on the plate.

If a similar measurement is to be made on diode 2 the d-c test connection must be made at the cathode of this diode, because its plate and also one side of the d-c measuring circuit are grounded. The cathode will be found positive with reference to ground, which shows that the plate is negative with reference to the cathode. Contact potential always makes the affected element negative with reference to the cathode.

If the twin diode tube is removed from its socket the contact potentials disappear. Also, were you to

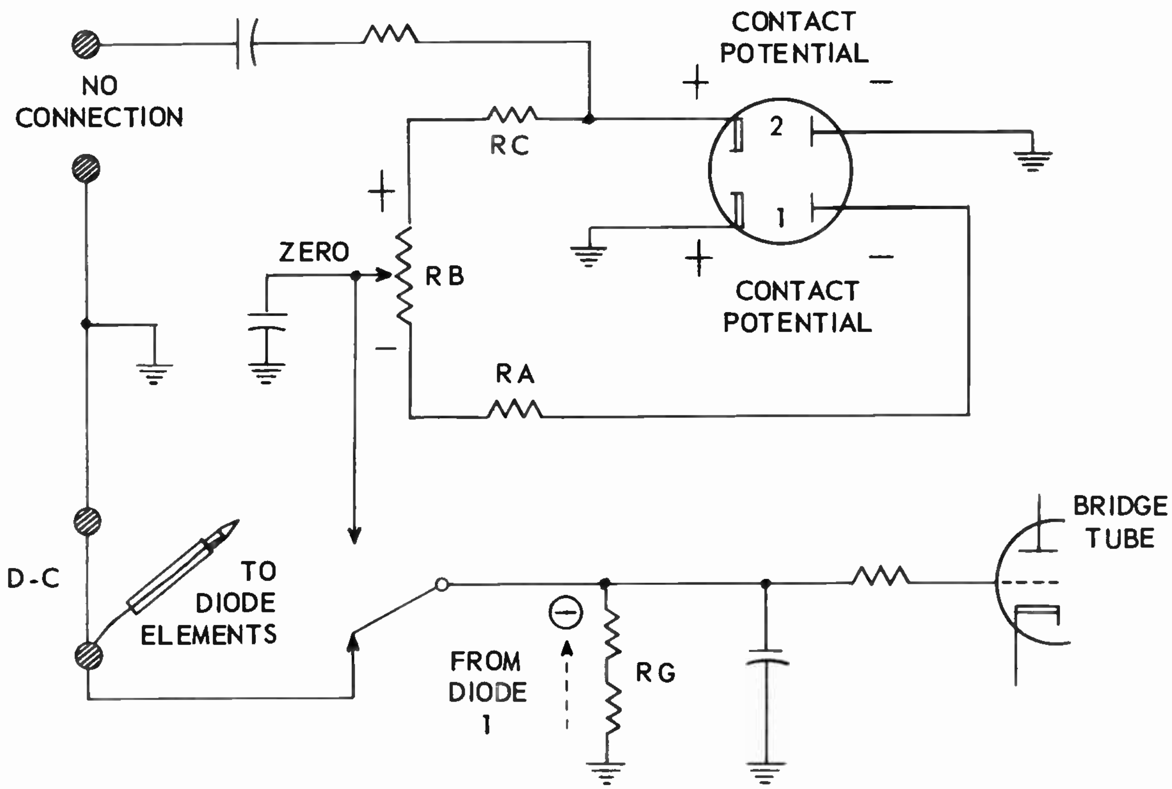


Fig. 73-5. Where the contact potentials appear in the rectifier circuits.

temporarily disconnect either side of the heater in the twin diode, to have a cold heater and no emission, there would be only zero voltages where before there were contact potentials.

(6) Unless you have a most unusual twin diode the two contact potentials will not be alike. This is where the a-c balancing potentiometer comes in. Note that the negative contact potential from the plate of diode 1 reaches the lower end of potentiometer Rb, also that the positive contact potential from the cathode of diode 2 is reaching the top of the potentiometer. Somewhere between the top and bottom the negative and positive contact potentials will balance out to zero. When you get the slider adjusted to this position there will be zero contact potential or no contact potential at all going to the bridge circuit when the function switch is in its a-c position.

Were there no balancing by means of diode 2 the contact potential from rectifying diode 1 would make the top of resistor Rg negative with reference to ground. This potential would go to the grid of bridge triode A, and all a-c readings would be inaccurate. Were the two contact potentials not exactly balanced at the potentiometer slider the zero settings for a-c and d-c measurements, and also for different ranges of a-c voltages could vary widely.

We are discussing contact potentials at some length because they have important effects not only in a-c vacuum tube voltmeters but also in nearly all measuring instruments which employ electron tubes in high resistance circuits, and in the grid circuits of all kinds of tubes for television and radio receivers.

In a triode or pentode there will be greatest contact potential at the control grid, because this is the element closest to the cathode where space charge electrons are thickest. There will be smaller contact potential at the screen, and one still smaller at the plate. Unless these contact potentials are cancelled by large voltages in external circuits they will affect operation of the tube. This explains why the relatively great contact potential at a control grid, in whose external circuit there may be a small biasing

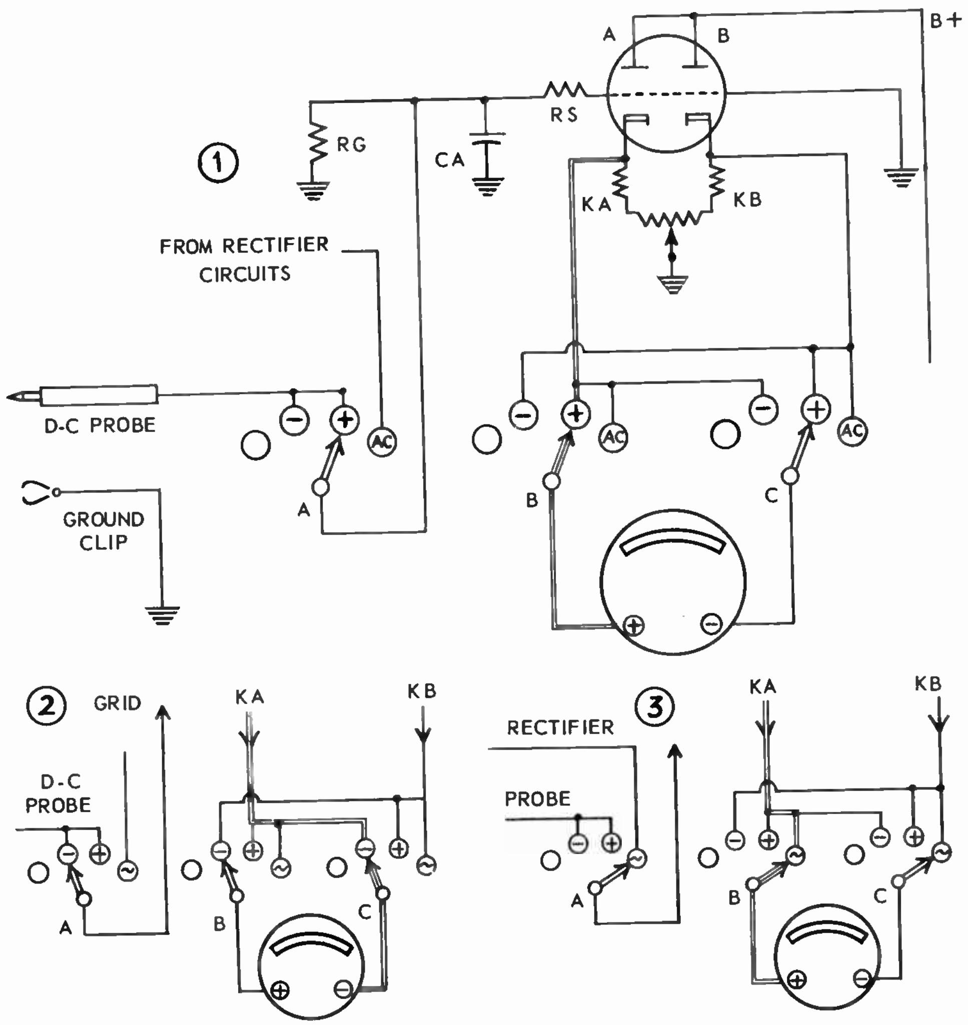


Fig. 73-6. Connections made by the function selector switch in various positions.

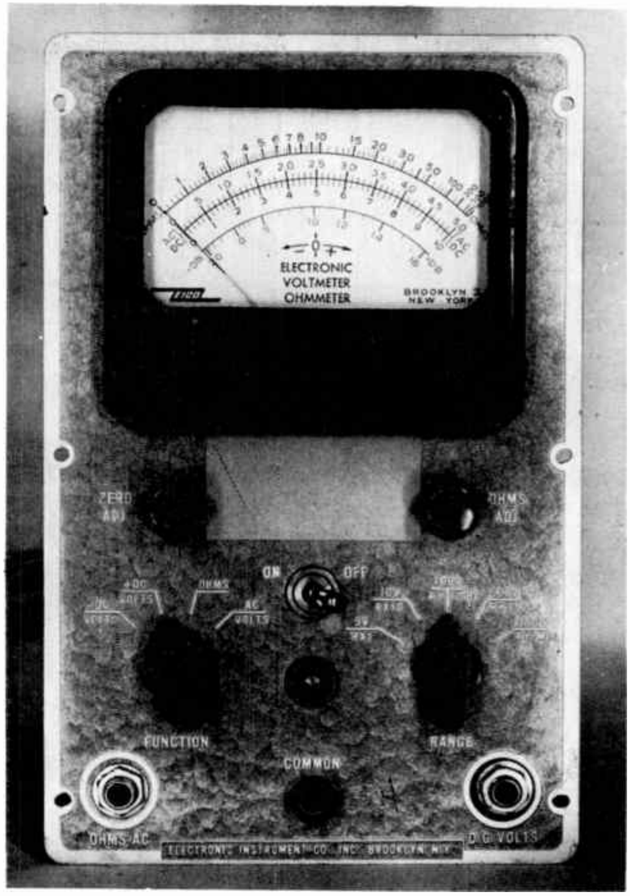


Fig. 73-7. A multi-range vacuum tube volt-ohmmeter.

voltage or none at all, can cause trouble. The actual biasing voltage on a grid may be due in greater measure to contact potential than to some small biasing voltage which is used intentionally.

INPUT CIRCUITS. Fig. 73-6 shows fairly typical connections for the function selector switch in a vacuum tube voltmeter. The switch has three sections, a, b, and c, with four positions on each section. One position is for measurement of a-c voltages. Another position is for measurement of d-c voltages which are positive with respect to ground. A third is for d-c voltages which are negative with respect to ground. A fourth position is shown vacant. It might be used for an ohmmeter function, for continuity testing connections, or any other desired purpose.

The upper diagram, 1, shows the switch sections in position for measurement of positive d-c voltages. The d-c probe, which is on the end of an external flexible cord, connects through switch section a to the grid circuit of amplifying triode A in the bridge system. The meter no longer is connected directly between the cathodes of the bridge tube, but is connected between the rotor contacts of switch sections b and c. The cathode of triode A is connected to the positive terminal of the meter through switch section b, while the cathode of triode B is connected to the negative side of the meter through section c.

Diagram 2 shows the function switch in position for measurement of negative d-c voltages. The d-c probe remains connected as before. The cathode of bridge triode A now has been switched over to the negative terminal of the meter, and the cathode of triode B is shifted to the positive side of the meter. When a negative d-c voltage is applied through the probe of the grid of triode A there is a decrease of plate or cathode current in this triode. Since the grid of triode B remains at its original voltage there is no change in its plate or cathode current, but we have the same effect as though this current were increased. The effective increase of current raises the relative potential at the cathode of triode B. Due to the reversed connections of the meter this rise of potential acts on the positive terminal of the meter and the meter reads up scale to indicate the value of applied negative d-c voltage.

The ability to read values of either positive or negative d-c voltages by merely shifting the function selector switch is a great convenience when checking plate, screen, and grid circuits, also automatic volume or gain control circuits and others where you sometimes encounter one d-c polarity and again the opposite polarity. Otherwise it would be necessary to reverse the connections of the probe and the ground clip for every change of polarity at points being checked.

Diagram 3 shows the function selector switch in position for measurement of a-c voltages. The d-c probe has been disconnected, and the grid circuit of bridge triode A is connected through switch section a to the output of the a-c rectifier. The rectifier circuit which we have examined delivers a positive rectified voltage from its output. Consequently, switch sections b and c now connect the cathodes of the bridge triodes to the meter terminals in a way suitable for measuring positive direct voltages. That

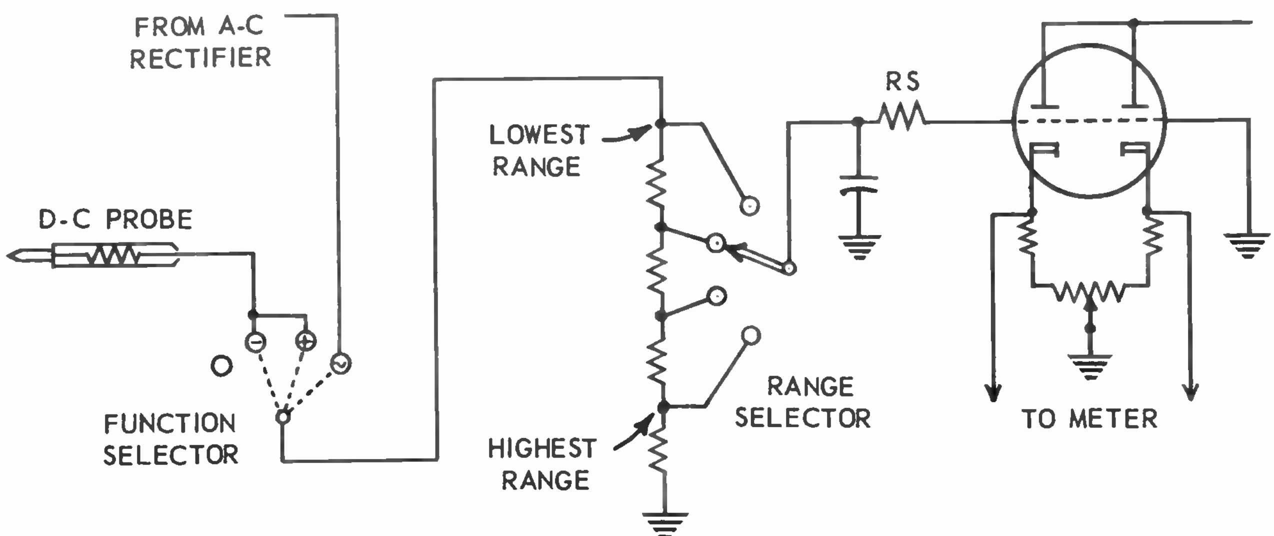


Fig. 73-8. Connections of a voltage range selector between the input and the bridge circuit.

REV – A

is, the connections between bridge cathodes and meter terminals are the same as for measurement of positive d-c voltages, although the switch rotors are in their a-c positions.

There are other ways of reversing the internal connections of the vacuum tube voltmeter for measurement of d-c potentials in either polarity. In some cases there is an exchange of connections to the grids of the bridge triodes. However the problem is handled the final result is the same - the meter reads up scale with a d-c voltage of either polarity being measured.

A vacuum tube voltmeter nearly always is a multi-range instrument. The unit of Fig. 73-7 has five voltage ranges with full-scale limits from 5 to 1,000 volts on both d-c and a-c measurements. There are meter scales for voltages and for resistances. A function selector may be set for d-c voltages of either polarity to ground, for a-c volts, or for resistance.

Fig. 73-8 shows connections for one kind of voltage range selector utilizing a four-position single-section switch. The range resistors take the place of grid resistor Rg, shown connected to the grid of a bridge triode in preceding diagrams. To the top of the string of resistors is applied the potential coming through a section of the function selector from either the d-c probe or the a-c rectifier. The bottom of the string is grounded. Taps at various points are connected to stationary contacts on the range selector switch, whose rotor is connected to the grid of the bridge triode.

The range selector resistors form a voltage divider. Total voltage across this divider is whatever comes from the function selector switch. Various fractions of the total voltage are taken off by operating the range selector, and these fractions of the total voltage are applied to the grid circuit of the bridge triode.

A voltage divider type of range selector may be designed as illustrated by Fig. 73-9. The same general procedure would be followed for any kind of voltmeter, electronic or otherwise, and also in laying out any kind of instrument, amplifier, or other device wherein certain fractions of the total applied voltage are required. The diagrams show an ordinary permanent magnet moving coil meter in series with a fixed resistor, Rs, which might represent the internal resistance of the moving coil instrument or the series resistor used in our vacuum tube voltmeter.

We shall assume that the problem is to design a range selector having an input resistance of 1,000,000 ohms or one megohm on all ranges. The ranges are to be 500 volts, 50 volts, 10 volts, and 5 volts. The meter, or the meter and series resistance, are to allow full-scale deflection on 5 volts.

Diagram 1 shows connections for the 500-volt range. The combined resistance of range resistor a and the meter circuit must be 1/100 of the total divider resistance, because 5 volts for the meter is 1/100 of the 500 volts to be measured at full-scale deflection on this range. Therefore, this first divider resistance must be 10,000 ohms.

Diagram 2 shows connections for the 50-volt range. The meter connection has been moved up to include divider resistors a and b. Since the 5 volts required at the meter is 1/10 of the 50 volts to be measured at full scale, the combined resistance at a and b must be 1/10 of the total 1,000,000 ohms resistance, or must be 100,000 ohms. Since we already have 10,000 ohms at a, an additional 90,000 ohms is required at b.

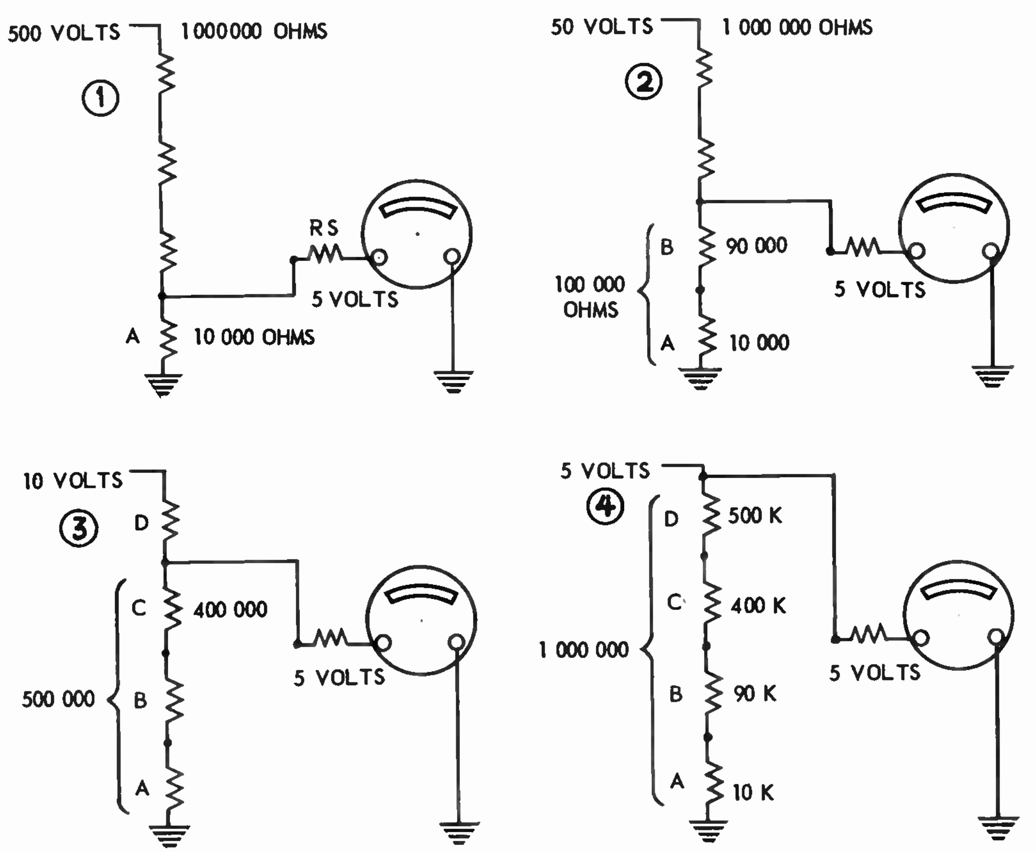


Fig. 73-9. How resistances for the range selector are determined.

Diagram 3 applies to the 10-volt range. The meter connection has been moved up to include another divider resistor, c. Because the 5 volts required at the meter is 1/2 of the 10 volts to be measured, the combined resistance of a, b, and c must be 1/2 of the total 1,000,000 ohms, or must be 500,000 ohms. In a and b we already have 100,000 ohms, so at c we need an additional 400,000 ohms.

Diagram 4 applies to the 5-volt range. Because the meter system requires 5 volts for full-scale deflection it is connected directly to the input of the divider. Note that on every range we have the full 1,000,000 ohms of all the divider resistances connected between the input and ground or connected across the applied voltage which is to be measured.

There are many modifications of this range selector. To compensate for the shunting effect of the meter or meter system at the higher ranges the series resistance, Rs, may be changed simultaneously with changes of divider resistance. This requires an added section on the range selector switch. In other cases the sensitivity of the meter or meter system is varied by a number of shunt resistances.

Total resistance in the range selector voltage divider for vacuum tube voltmeters usually is on the order of 5 to 10 megohms. Always there is some additional resistance in the d-c probe, as shown in Fig. 73-8. The probe resistance often is one megohm, but may be 10 to 15 megohms or even more. The d-c input resistance of the instrument is the sum of the resistances in the probe and in the range selector. This d-c input resistance remains the same for all voltage ranges, rather than becoming smaller on the lower ranges as in the case of a simple voltmeter of the permanent magnet moving coil type. On the basis of ohms per volt, the input resistance of the vacuum tube voltmeter is greatest on the lowest range and least on the highest range. The sensitivity on low ranges becomes so great that shielded cable with the shield grounded is used between the instrument and the d-c probe. Otherwise there is enough voltage pickup from any and all kinds of electric fields to drive the meter pointer all over the scale.

The input resistance when measuring a-c voltages is that of the range selector voltage divider plus all the resistances in series with the rectifier system. This a-c input resistance usually is somewhat less than the d-c input resistance because the a-c probe ordinarily contains no resistor.

With reference to measurement of a-c voltages we must consider the input impedance, which is the combined opposition due to resistance and capacitance to ground within the instrument and in the external cable and probe. With any ordinary construction there are many rather large internal capacitances. They exist between all the conductors in wires, resistors, capacitors, or other parts and the chassis metal. The diode rectifier elements possess capacitances, as do also the base pins and the socket lugs. All these capacitances to ground may add up to something like 100 to 200 mmf.

The internal capacitances are in parallel with the voltage applied to the measuring terminals, or rather with the measured circuit in which this voltage exists. The capacitive reactance decreases as the measured frequency increases, and the result is as though the instrument resistance were made lower and lower with increase of applied frequency.

The capacitive reactance of 150 mmf at a power frequency of 60 cycles is nearly 18 megohms, but at an audio frequency of 5,000 cycles it is down to about 210,000 ohms. At a broadcast radio frequency of 1,000 kilocycles the reactance would be only a little more than 1,000 ohms, and at a video intermediate frequency we would have a near short circuit, maybe 40 ohms or thereabouts. Regardless of how high the input resistance may be, the input impedance never can be higher than the capacitive reactance. This is why the a-c VTVM by itself is not good for measurements at frequencies much above the audio range. No other kind of a-c voltmeter, by itself, is any better, and most of them are not so good as the VTVM. There are voltage readings at the higher frequencies, but they are lower than the true value of measured voltages.

In addition to an error due to frequency the VTVM when used for alternating voltages is subject to waveform error, just as is any other instrument employing a rectifier. Some vacuum tube voltmeters are calibrated to indicate peak values of alternating voltages, but the majority of service type instruments are calibrated for r-m-s or effective values. The calibration is based on a sine wave voltage, and when the wave is of other than sine form the readings may be either higher or lower than the true r-m-s voltage.

The reading of an a-c vacuum tube voltmeter results from the sum of all the instantaneous voltages in the rectified alternations, as these instantaneous voltages are indicated for a sine wave at 1 in Fig. 73-10. The effective or r-m-s value is equal to a steady direct voltage at 0.707 times the peak a-c voltage, as at 2. If the rectified voltage is of some other form than a sine wave, as at 3, the meter reading still will result from the sum of all the instantaneous voltages, but this sum may not be the same as with a sine wave.

Alternating voltage waveforms represented at 1 and at 3 are symmetrical. That is, the forms are identical, although inverted, from beginning to end of each positive and negative cycle. The voltmeter would give the same reading with connections to the measured circuit reversed, because it would make no difference which alternation were rectified, the negative alternation would be just like the positive alternation.

The waveform shown at 4 is not symmetrical. The sum of the instantaneous voltages during the upper alternation is greater than during the lower alternation. With the meter connected to the measured circuit in a manner which would result in rectification of the upper alternation, the voltage reading would be higher than with connections reversed, whereupon the lower alternation would be rectified and measured. You may find quite different voltage readings when reversing the a-c meter connections during service operations.

Although waveform errors can prevent readings corresponding to r-m-s calibrations, this seldom is of any importance in service work. About the only time we wish to know true a-c voltages is when check-

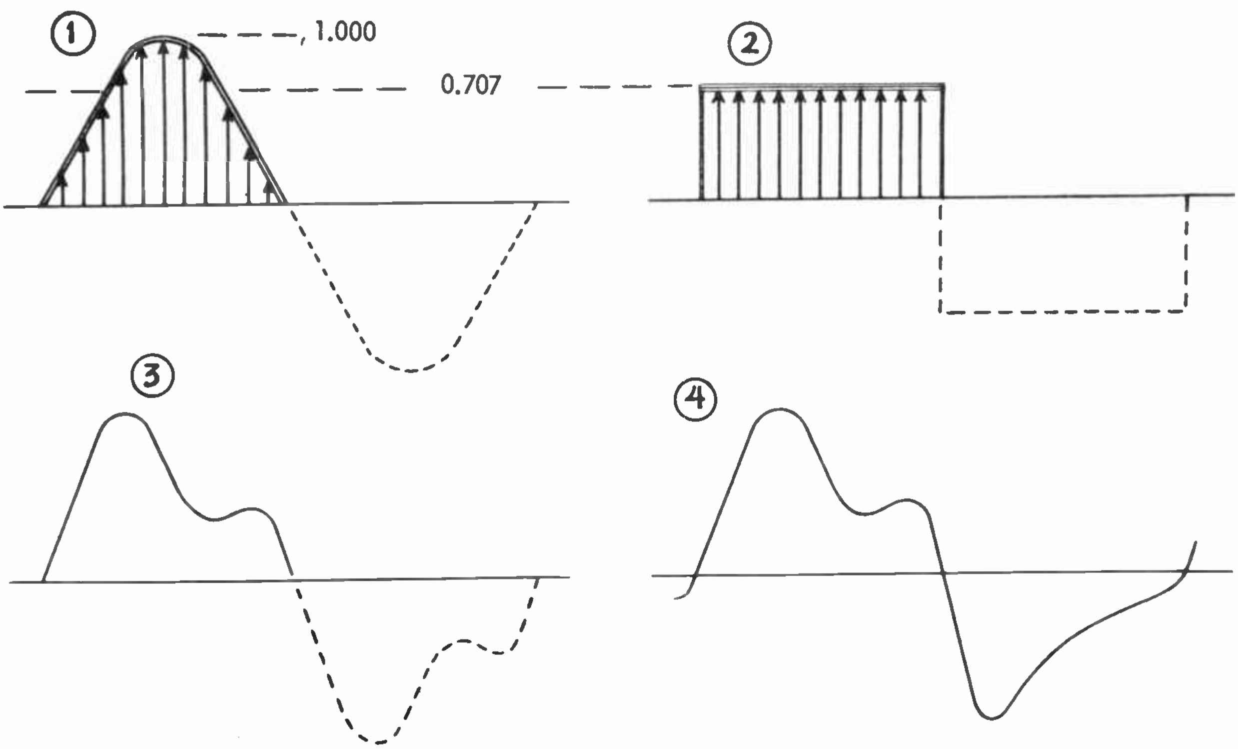


Fig. 73-10. The a-c vacuum tube voltmeter is subject to waveform errors.

ing power lines, and in these lines the voltage always is very close to sine wave form. When it comes to audio-frequency measurements there usually are irregular and non-symmetrical waveforms, but here we seldom are so concerned with absolute voltage levels as with whether the voltage in one part of an amplifier or other circuit is greater or less than in another part. The chances are that the waveforms will be nearly enough alike at the two points as to cause the same waveform error in both cases, and the comparative values will give us all needed information.

USING THE VTVM. All of the precautions which have been explained in connection with other types of meters must be observed when using the vacuum tube instrument for similar purposes. In addition, there are two other things which always must be done before proceeding to make measurements of any kind.

First you must allow an adequate warm-up period, which may be anything between two or three minutes and 15 minutes or more. This is necessary in order that the two triodes of the bridge circuit may settle down to their final rates of electron emission and of cathode currents. Remember, it takes only a very small difference between the two cathode currents to cause a large deflection of the meter pointer, and if there are slight changes of emission during a series of tests the measurements may mean little or nothing when compared one with another. The warm-up period is necessary also to allow all resistors and other circuit elements to reach normal operating temperatures and corresponding values.

The second operation which is peculiar to vacuum tube voltmeters is setting the zero adjuster, potentiometer Kz in some of our diagrams, while the tip of the test probe or prod is connected to the instrument ground or to the common lead. This is the only way to insure that no nearby electric or magnetic fields are producing a potential difference between the probe tip and the instrument ground. It is not enough to adjust the meter for a zero reading with the test probe open. After making a correct zero adjustment, pay no attention to any meter reading that occurs with the probe tip disconnected from the instrument ground. This indicates the pickup of potentials from surrounding fields, and it will not exist when you use the probe for regular checking while connected to measured circuits.

3 A fairly good way of determining when the warm-up period has been sufficient is to make a correct zero setting, then a few minutes later check this setting. If the meter pointer moves away from zero between two settings, there has not yet been enough warming up of the parts.

Always be sure to use the regular d-c probe, with its shielded cable and self-contained resistor, for making d-c measurements, but be equally sure not to use it for a-c measurements. Using the wrong probe will make meter readings meaningless.

To make the most reliable measurements of voltages, either direct or alternating, it is desirable that ground or the common lead of the VTVM always be connected to chassis ground of the receiver or other part being tested. This rule is easy to follow when using service types of voltage charts and tables, because they almost invariably show voltages to chassis ground.

In some cases you may wish to measure potential differences across certain parts of which neither end connects to ground. For example, with a tube circuit such as shown by Fig. 73-11, you might wish to know the voltage being applied across decoupling resistor Rd. The thing to do would be to measure from each end of this resistor to ground, then subtract the smaller from the larger voltage. Quite often

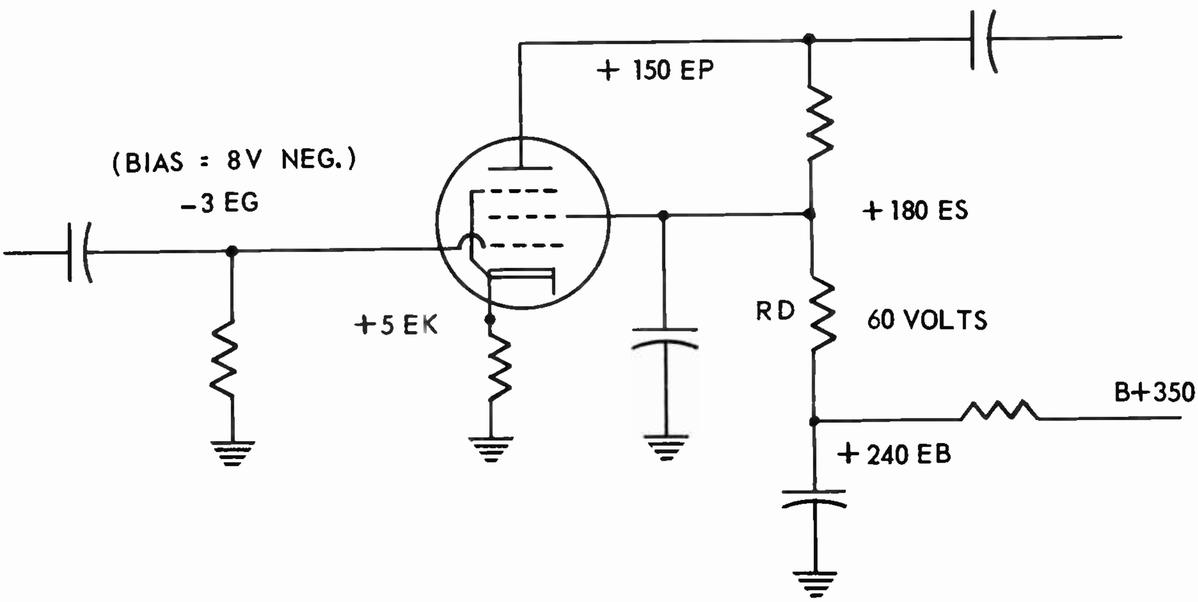


Fig. 73-11. Measured potential differences in tube circuits may be at points not connected to ground.

you will wish to know the true plate voltage, which is the potential difference between plate and cathode. Then you should measure from plate to ground and from cathode to ground, and subtract.

The real reason for this procedure is to prevent the entire VTVM from operating at some potential above ground, for this allows the large mass of the instrument to pick up strongpotentials from surrounding electric fields and the readings are likely to be highly inaccurate. It is especially important to keep one of the test leads connected to chassis ground when making measurements in high-resistance or low-current circuits, in grid circuits for example. To determine the grid bias in Fig. 73-11 the voltages of the grid and the cathode to ground would be measured separately. Since one element is negative to gronnd and the other positive you would add the two voltages to determine the actual potential difference, which is grid bias.

Some vacuum tube voltmeters have no internal blocking capacitor in series with the a-c input. This is capacitor Cb in Fig. 73-4. When you wish to measure only alternating voltage in a circuit where direct voltage also is present, the meter with no internal blocking capacitor requires connection of a fixed capacitor in series with the a-c test probe, somewhat as shown by Fig. 73-12. The series capacitance must be large enough, and its reactance low enough, not to seriously drop the a-c voltage going to the meter. This scheme would be used principally for measuring audio-frequency voltages in plate circuits where a direct B voltage is present.

An external capacitor in series with the a-c test lead sometimes is recommended for protecting the vacuum tube voltmeter against high d-c voltages which may be in the measured circuit along with the

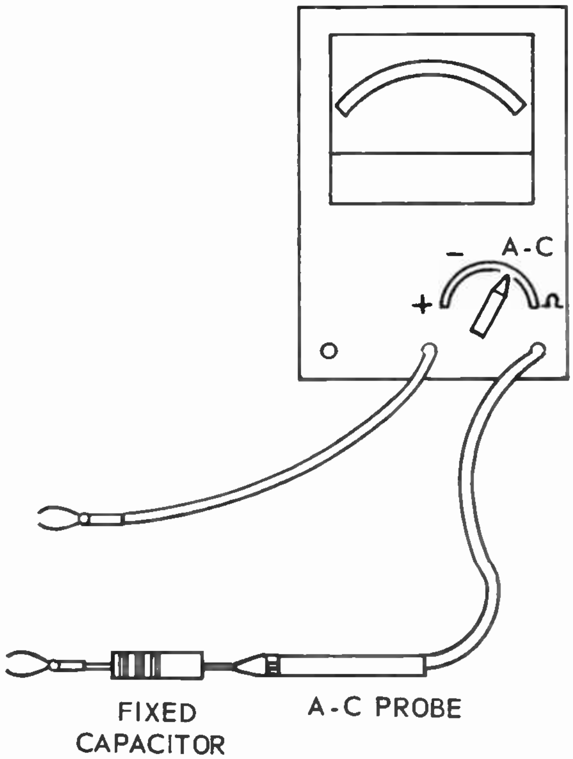


Fig. 73-12. Blocking of d-c components from the VTVM.

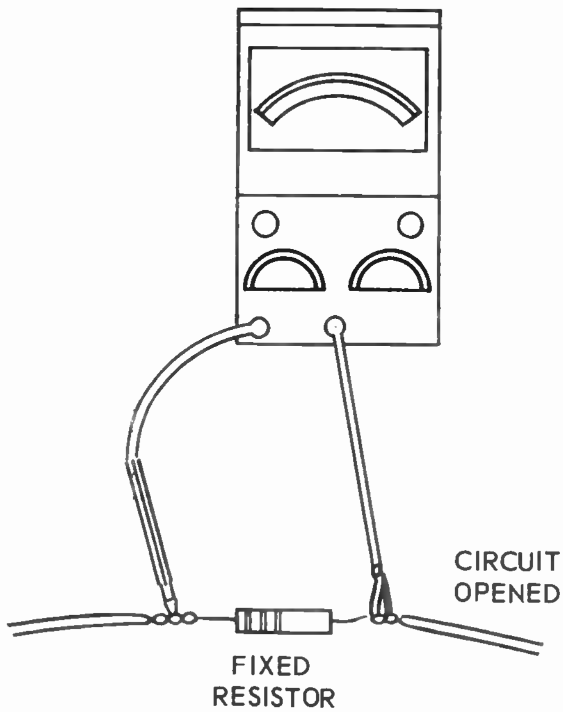


Fig. 73-13. How current may be measured with the VTVM.

a-c to be measured. The voltage rating of the series unit must be amply in excess of any d-c voltage likely to be encountered. Unless this series capacitance is large enough not to cause a great drop in a-c voltage the measurements will be incorrect.

A vacuum tube voltmeter may have built-in provisions for measuring direct currents, alternating currents, or both kinds. Direct currents may be measured without applying line power to the instrument, simply using the permanent magnet moving coil meter with suitable shunts for various current ranges.

Any vacuum tube voltmeter may be used for measuring either direct or alternating currents by means of connections shown in Fig. 73-13. It is necessary to open the circuit in which current is to be measured, then close it through a fixed resistor. The resistor should be carbon or composition (non-inductive) if alternating current is to be measured, and the resistance must be known with considerable accuracy. Then, with current flowing in the measured circuit, we use the VTVM to measure the voltage drop across the known resistance. Our regular rule for current gives the answer.

$$\text{Milliamperes} = \frac{1000 \times \text{volts measured across resistor}}{\text{ohms in resistor}}$$

The resistor should be of the lowest resistance in which existing current will give an easily and accurately read voltage drop. Values such as 1, 2, 10, and 100 ohms usually are satisfactory, and they have numbers with which division is easy to make when using the formula. It is no more trouble to open the measured circuit and insert the resistor than to open it and insert any kind of current measuring meter. The series resistance usually may be as small as the internal resistance of any meter which would measure the same current, so the circuit is little affected while testing.

CALIBRATING THE VTVM. To calibrate a vacuum tube voltmeter means to set the adjustable resistors at values which allow accurate indications of voltages and any other circuit properties or characteristics which may be measured. All instruments have several calibrating adjustments, usually on top of or underneath the tube shelf that is behind the panel. The adjustments may or may not be marked as to their purpose. If they are not marked, and you have no instructions applying to the particular instrument, there is nothing to do but experiment.

As a general rule there are two adjustments for d-c voltage. One may be for measurement of voltages which are positive to ground, and the other for voltages negative to ground. Sometimes one of these adjustments is for keeping the meter on zero when shifting from positive to negative voltages, with another for obtaining correct readings. There will be at least one adjustment for correct indication of a-c voltages, and when two rectifiers are used there will be another adjustment for balancing the contact potentials. See Figure 73-14

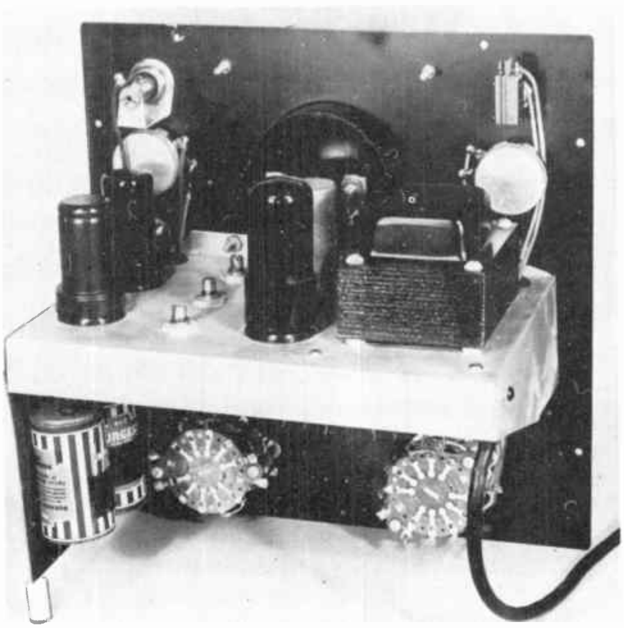


Fig. 73-14. Three calibration adjusters are between the tubes on top of the tube shelf.

The first step is to leave the line power turned off while, if necessary, setting the zero adjuster on the meter case to bring the pointer exactly to zero. The meter should be lightly tapped to make sure that the pointer swings freely during this adjustment. Then the line power should be turned on and the instrument allowed to warm up for at least 30 minutes.

The calibration adjustments are set while an accurately known voltage is applied through the test terminals or probes, bringing the meter pointer to the corresponding positions on the scale. It is best, when possible, to make the calibration with the VTVM connected in parallel with another meter of known accuracy on the source of calibrating voltage, as in Fig. 73-15. A d-c source might be a dry cell battery, or the B power supply in any other instrument or any receiver, at some point following the filter. An a-c source might be the heater supply in any instrument or receiver, making connections at the heater lugs of a socket. The a-c power line could be used if you take suitable precautions to avoid shocks and blowing of fuses because of accidental shorts in your test connections.

A d-c comparison meter may be of any sensitivity, since no matter how much or how little the source is loaded the comparison meter and the VTVM still are connected to the same voltage at the source. An a-c comparison meter is preferably of the moving vane type, since its inherent accuracy is better than that of rectifier meters.

If no comparison meter is available, d-c calibration may be made with excellent accuracy by using one or more fairly fresh dry cells, connected in series when more than one is employed. Since very little cell current will be taken by the VTVM there will be no appreciable internal drop in the battery, and the measured voltage may be figured as 1.55 volts per cell.

The a-c power line often is used for a-c voltage calibration, assuming it to furnish 117 a-c volts. It is best not to use the power line voltage when it may be somewhat low at times of peak loads. Peak loads occur at all meal times, while it still is dark during winter mornings, and as dusk comes on at any time of the year. A-c heater voltages from receivers or other testing instruments may be used for an approximate a-c calibration. Test instruments usually provide heater voltages fairly close to tube ratings, as 6.3 volts, although sometimes the heaters are intentionally run at a slightly lower voltage to have long life. Receivers often apply excessively high heater voltages.

Calibration adjustments usually are made on the lowest voltage range of the VTVM, since this keeps most or all of the range resistors in the circuit. If higher ranges then do not show correct calibration there is little or nothing to do about it, because these readings depend on accuracy of resistances in the range selector system.

Direct voltage always should be calibrated before alternating voltage. This is because the d-c bridge circuit or its equivalent nearly always is employed when measuring rectified alternating voltages, and incorrect d-c calibration would make it impossible to correctly calibrate for alternating voltages.

Re-calibration always will be needed after the tubes in the instrument age during normal use. If a VTVM is being regularly employed for service work the aging will take about two weeks, or it may be hastened by leaving the instrument turned on continually during three or four daytimes. Leaving the instrument on all night is not too safe, there could be overheating without prompt detection, and a possible fire.

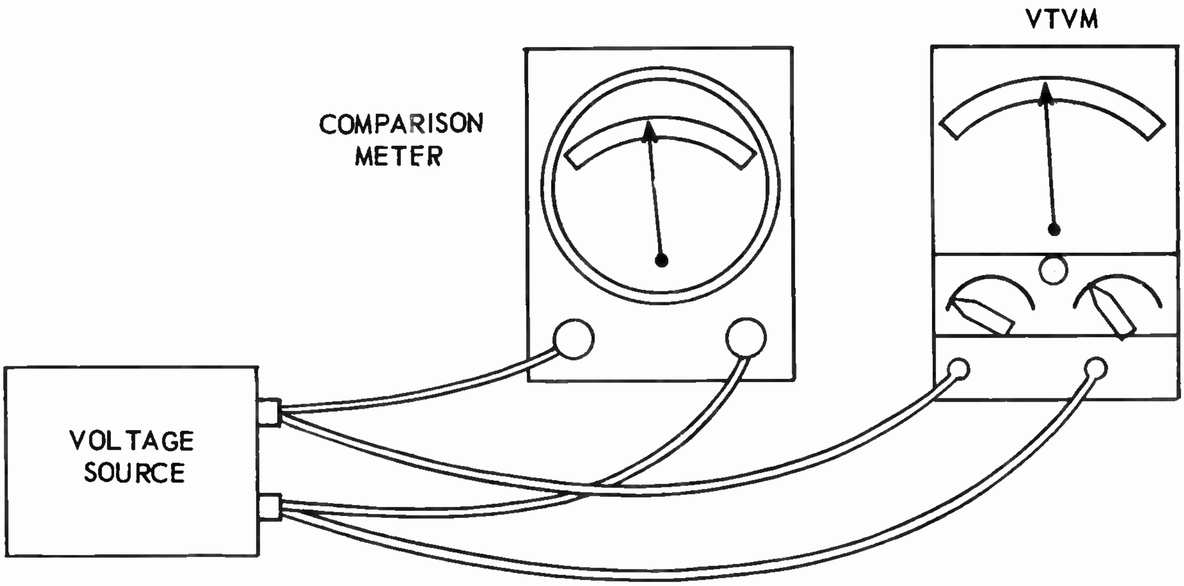


Fig. 73-15. Calibration of a vacuum tube voltmeter by comparison with a reference meter.

Replacing a rectifier always will call for re-callibration of a-c voltage measurements. Replacing a bridge tube will call for re-calibration of both d-c and a-c voltage readings.

TROUBLES IN THE VTVM. If it becomes impossible to bring the meter pointer to zero on the voltage functions by using the zero adjusting potentiometer, the bridge tube may need replacing. The two triodes or the two sections of a twin triode must be fairly well matched in emission, and usually the cathode of one will commence to give out while the other retains almost normal performance. Interchanging the grid connections will do no good, it will merely reverse the direction of error. This trouble sometimes results from too low B voltage, in which case the power supply rectifier probably should be replaced.

If the pointer goes all the way to one extreme or the other, and stays there, there is likely to be an open circuit, a defective joint in the wiring, a defective resistor in the bridge circuit, or the function selector switch contacts may be bent, corroded, or dirty.

No readings of voltage may result from open circuits and similar faults previously mentioned. There may be defective resistor or some form of open circuit in the range selector system. The testlead may not be making good contact if it is connected through a plug and jack, or there may be an open inside the prod.

If the instrument cannot be satisfactorily calibrated it is possible that the meter movement has been damaged through misuse, or the permanent magnet may have been weakened. More probably the trouble is in the range selector system, either shorts or opens, or poor connections causing excessively high resistances.

DO NOT TEAR — CUT ALONG THIS LINE AND SEND IN FOR GRADING.

TELEVISION

LESSON NO. 74

MEASURING CAPACITANCE AND INDUCTANCE

Training Your Mind
C.T.I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind
C.T.I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON NO. 74

MEASURING CAPACITANCE AND INDUCTANCE

When first we talked about trouble shooting, the statement was made that sooner or later it becomes necessary to measure voltages, currents, resistances, and capacitances on all the more difficult jobs. We have learned how to measure voltages, currents, and resistances, but capacitance measurements remain to be investigated.

Now it happens that most of the capacitor testers used in service work make use of some form of bridge circuit. Furthermore, it happens that a bridge circuit may be used not only for measuring capacitance, but also for resistance, inductance, reactance, impedance, and even for measuring frequencies. In addition to all these uses you will find that the fundamental principle of the bridge is employed for purposes other than measurement, in audio-frequency generators for instance.

Because a general purpose bridge, usually called an impedance bridge, may be employed for measuring so many different electrical quantities it is commonly used in development laboratories, but seldom for service work. This bridge is not quite fast enough for service operations, because a certain amount of computation may be needed in arriving at values which correspond to scale readings. But special purpose bridges, which measure only one kind of quantity and give direct readings, are common service instruments.

The simplest bridge circuit, used for measuring resistances, is shown by Fig. 74–1. There are four "arms", marked X, S, A, and B. Arms X and A are in series with each other across a voltage source, as are also arms S and B. Branch X-A is in parallel with branch S-B across the source. A voltage indicator, here shown as a meter, is connected from between X and A to between S and B.

In diagram 1 each of the four bridge arms has 20 ohms resistance, and the source is furnishing a potential difference of 10 volts. Half of this applied voltage is dropped in arm X and the other half in arm A of the upper parallel branch, and there is the same division of voltage in arms S and B of the lower branch. Consequently, there is a potential of 5 volts on both sides of the indicating meter, and the meter reads zero because there is no difference of potential across it.

When the indicator shows a zero difference of potential the bridge is said to be balanced. Whatever device is used to show zero potential difference may be called the null indicator, because the word null means nothing. The indicator still would read zero no matter what the voltage from the source, because voltage drops would remain equal in the bridge arms and the same potentials would be applied to both sides of the indicator.

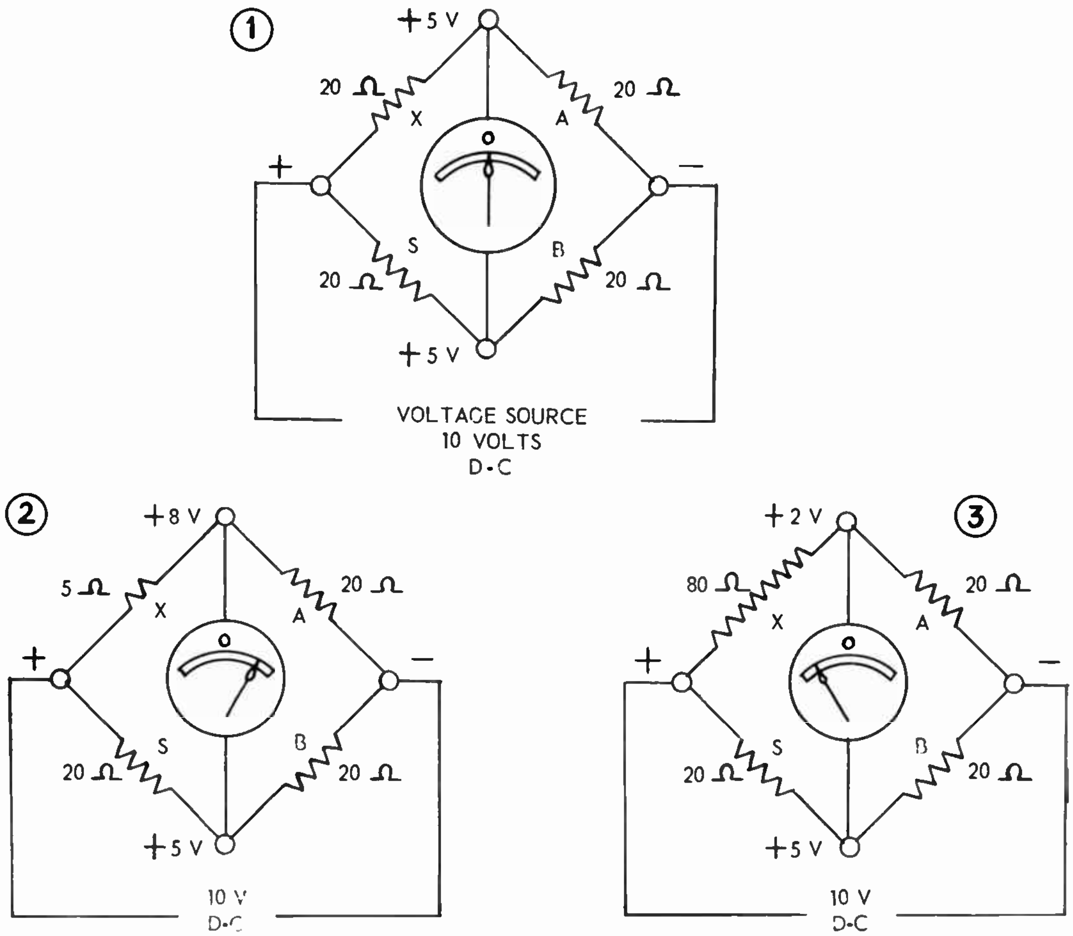


Fig. 74–1. The elementary circuit of a resistance bridge.

In diagram 2 the resistance in arm X has been changed to 5 ohms, with all other arms and also the applied voltage remaining as before. There is a drop of only 2 volts in the reduced resistance at X, leaving 8 volts across arm A and at the upper terminal of the indicator. We still have 5 volts at the lower terminal of the indicator. There is now a potential difference of 3 volts (8 minus 5) across the indicator, and the pointer deflects accordingly.

In diagram 3 the resistance at X has been changed to 80 ohms, with everything else as before. There is a drop of 8 volts in this 80 ohms at X, leaving 2 volts in the 20 ohms at A and at the top of the indicator. The original 5 volts remains at the bottom terminal of the indicator, and there is a potential difference of 3 volts (8 minus 5) across the indicator. The pointer deflection now is opposite to that of diagram 2, because here the upper terminal of the indicator is less positive or is effectively negative with reference to the lower terminal, while in diagram 2 the upper terminal is more positive than the lower terminal.

So long as we have 20 ohms at S, A, and B the bridge will balance only with 20 ohms at X. The indicator deflects one direction with smaller resistances at X, and deflects the opposite direction with greater resistances at X. We could use this bridge for separating resistors into three classes, those of exactly 20 ohms, of less than 20 ohms, and of more than 20 ohms resistance.

Fig. 74–2 shows one way in which a bridge may be used for measuring a wide range of resistances. Arm X has been provided with terminals between which may be connected any unknown resistance. The letter symbol "X" always stands for an unknown quantity, which here is a resistance to be measured. The resistance in arm S has been made adjustable. We shall refer to this arm as the standard resistance.

In diagram 1 the resistance at X is 5 ohms. The bridge will balance when the standard resistance is adjusted to 5 ohms, because when arms A and B are equal, and the standard is adjusted to equal the unknown resistance, there will be equal drops of voltage across X and S to make equal potentials on both sides of the indicator. In diagram 2 the resistance connected at X is 80 ohms. The bridge will balance when the standard is adjusted to 80 ohms. The range of unknown resistances which may be measured now is the same as the limits of minimum and maximum resistances to which the standard may be adjusted.

The accuracy of the bridge will be equal to the accuracy or the tolerance of the standard resistance, provided the resistances at A and B are precisely equal to each other. Measurements of ohmic resistance are most accurate when the source furnishes pure direct current, as from a battery. If the bridge arms and the unknown resistance are non-inductive the measurements may be made with alternating voltage of any frequency up to about 1,000 cycles, but at higher frequencies the distributed and stray capacitances may cause errors.

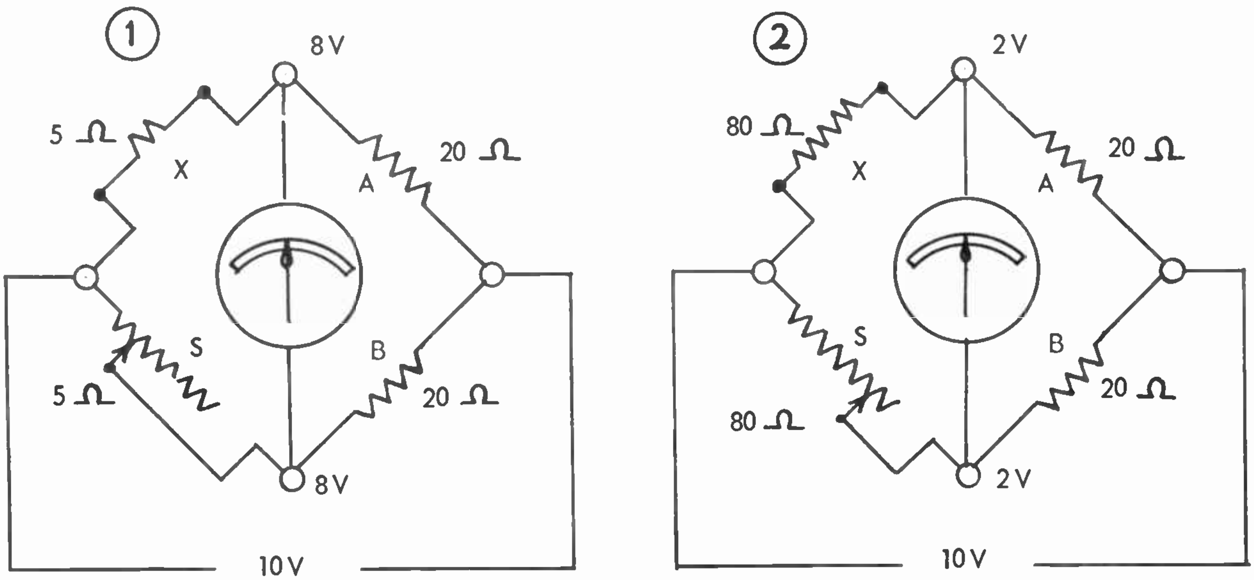


Fig. 74–2. An adjustable standard allows measuring a wide range of resistances.

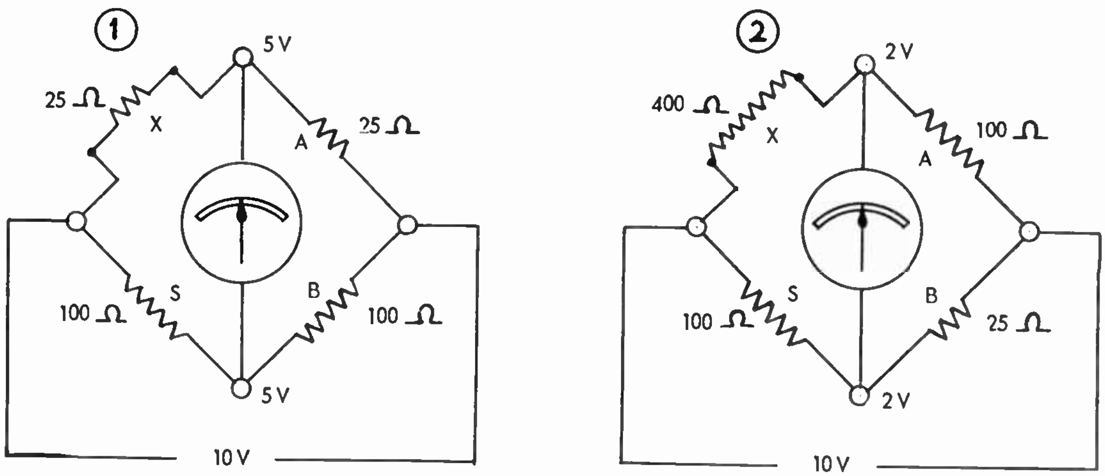


Fig. 74–3. The ratio of unknown to standard resistances is the same as that of the two ratio arms.

Adjustable non-inductive resistors of close tolerance and a wide range of values are quite costly. It is possible to use a single highly accurate standard resistance for a wide range of measurements by varying the relative resistances in arms A and B, which are called the ratio arms of the bridge.

What happens when we vary the ratio arm resistances is illustrated by Fig. 74–3, where the standard resistance remains of 100 ohms for all measurements. In diagram 1 we have 25 ohms in ratio arm A and have 100 ohms in ratio arm B. The bridge will balance when the unknown resistance is 25 ohms, for then there are equal voltage drops in arms X and A, and also in arms S and B. This causes equal potentials on both sides of the indicator, and it gives a null indication.

In diagram 2 the ratio arm resistances have been interchanged. The bridge will balance with 400 ohms connected at X, for then there will be 8 volts drop in X and also in S, and the remaining 2-volt drop will occur in A and also in B.

In diagram 1 the ratio of A to B is 25/100 or 1/4. Then the ratio of the unknown to the standard resistance must likewise be 1/4, and since the standard is 100 ohms the unknown must be 1/4 of 100, or 25 ohms. At 2 the ratio of A to B is 100/25 or 4, and the unknown resistance must be 4 times the standard, which is 400 ohms. Were the ratio arm resistances equal to each other the ratio would be 1/1 or 1, and the bridge would balance when the unknown resistance is equal to the standard.

Adjustable ratio arms often consist of a potentiometer connected as shown by Fig. 74–4. The unknown resistance, X, and the standard resistance, S, remain as before. The connection from X and the indicator which formerly went to one end of arm A now goes to one end of the potentiometer, while the connection from S and the indicator which formerly went to arm B now goes to the other end of the potentiometer. The slider of the potentiometer, which now is the junction between arms A and B, goes to the source.

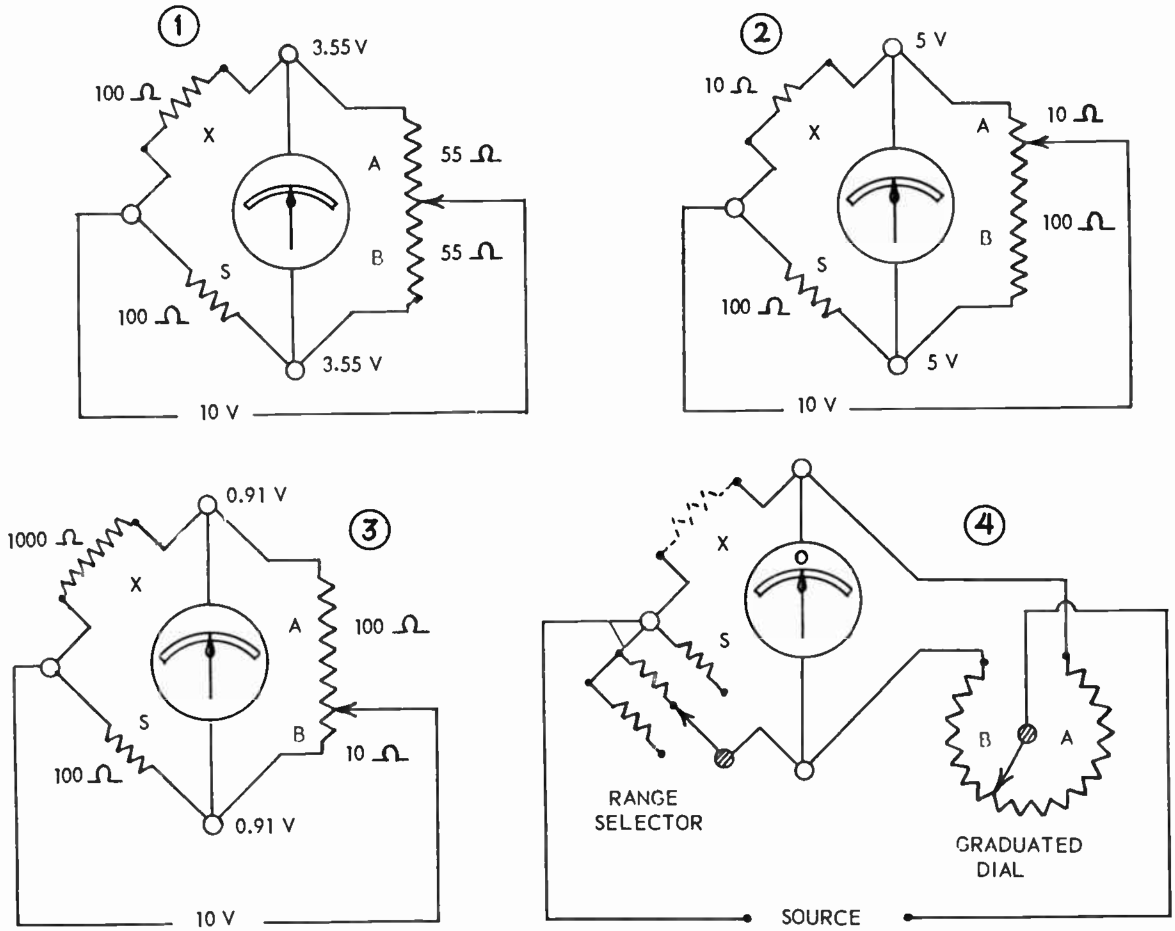


Fig. 74–4. An adjustable potentiometer used for the two ratio arms.

With the slider at the electrical center of the potentiometer resistance, as in diagram 1, arms A and B are equal and the bridge ratio is 1/1. If the slider is moved toward the connection of the unknown resistance, as at 2, the portion of the resistance which is arm A is decreased, while the portion which is arm B is increased. The diagram shows 10 ohms for A and 100 ohms for B. The ratio is 10/100 or 1/10. The unknown resistance at X must be 1/10 of the standard at S. Were the unknown resistance to be 1/5 of the standard, the bridge would balance with the potentiometer slider moved to a position giving a ratio of 1/5 between portions A and B.

In diagram 3 the unknown resistance is assumed to be 1,000 ohms, which is 10 times the standard resistance of 100 ohms. The ratio of unknown to standard is 10/1, and to balance the bridge the potentiometer slider must be moved to give a ratio of 10/1 between portions A and B of the potentiometer resistance.

With the slider of the potentiometer at various positions between those of diagrams 2 and 3 the bridge could be balanced for any unknown resistance between 10 and 1,000 ohms, employing only the single 100-ohm unit as a standard. Moving the slider still farther in either direction would allow balancing with unknown resistances of less than 10 ohms and of more than 1,000 ohms.

Without working close to the ends of the potentiometer we could measure unknown resistances between 10 and 1,000 ohms with a standard resistance of 100 ohms. Using a standard resistance of 1 ohm would allow measuring unknowns from 0.1 ohm to 10 ohms, and a standard of 10,000 ohms would handle unknowns of 1,000 to 100,000 ohms. To do all this our resistance bridge might be arranged as in diagram 4 of Fig. 74–4, where there is a range selector switch for cutting in various standard resistances, and a potentiometer dial graduated in ohms of unknown resistance. The graduation values would be multiplied or divided according to the range being used.

a The type of bridge which we have been examining, in which an unknown resistance is compared with a known or standard resistance is called a Wheatstone bridge, after its inventor, Sir Charles Wheatstone. There are many other types of bridges, such as Maxwell, Hay, Wein, Carey-Foster, and so on, all of which are named for their inventors. All these other types employ the basic Wheatstone circuit, but with modifications which suit the instrument for certain special purposes.

A-C BRIDGES. The Wheatstone bridge circuit may be used for comparing two capacitances, as at 1 in Fig. 74–5, or for comparing two inductances, as at 2. The unknown element is connected in arm X, the known or standard element is connected in arm S, and the ratio arms A and B are adjusted for a balance. Thereupon the ratio of A to B is the same as the ratio of X to S, just as in the resistance bridge.

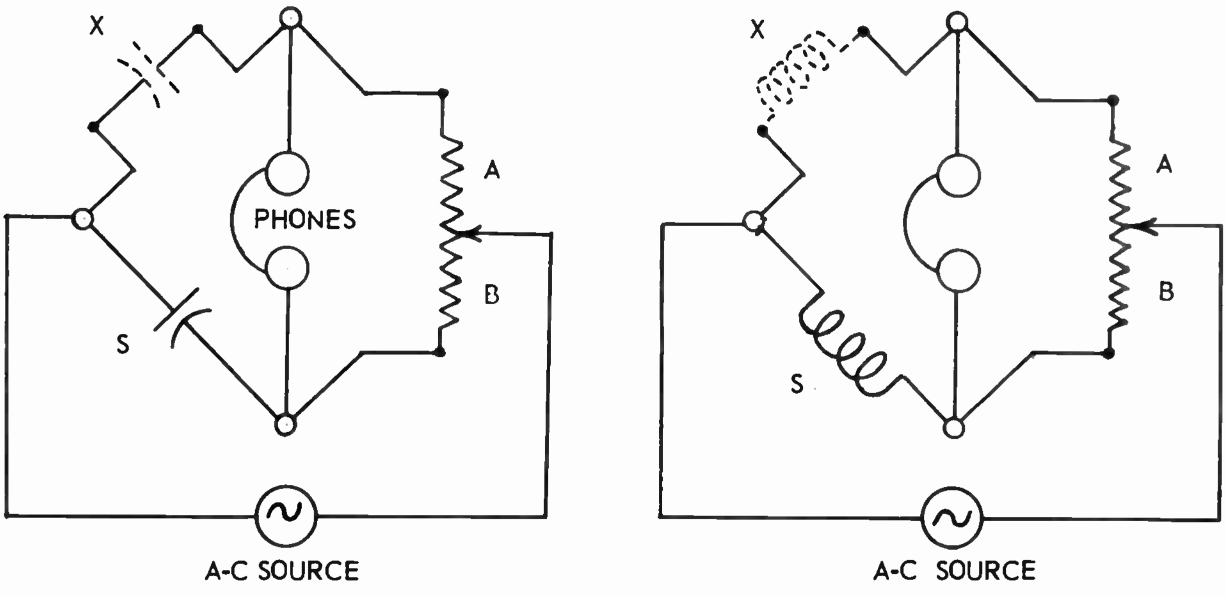


Fig. 74–5. A-c bridges for measuring capacitances and inductances.

In these and other kinds of a-c bridges we are not making direct measurements of capacitance and inductance, but rather of capacitive reactance and inductive reactance. Since reactances are measured in ohms they may be used in some of the arms of a bridge whose other arms are resistances, also measured in ohms. Reactance of any kind is a form of opposition to alternating current, and in order to have reactances appear in the unknown and standard elements the bridge must be powered from alternating voltage or current.

The voltage source for an a-c bridge may be obtained through a step-down transformer from the 60-cycle power line. It may be a type of vibrator, called a microphone hummer, which usually operates at 1,000 cycles. Any kind of oscillator may be used, operating at either audio frequencies or at the lower radio frequencies.

The kind of null indicator used with an a-c bridge depends to some extent on the frequency of the voltage source. With any audio frequency it is fairly common practice to use headphones, from which the audible tone falls to zero or to minimum loudness when the bridge is balanced. It is rather difficult to identify the point of minimum loudness with an exactness allowing measurement accuracy much better than 5 to 10 per cent.

Rectifier meters with full-scale deflection on no more than 500 microamperes often are used to provide a visible null indication. As the bridge is adjusted through the balance point the meter reading will drop to zero or to a minimum and then will rise again. On alternating current the meter pointer cannot swing from one side of zero to the other side, as with direct current, because electron flow through the rectifier can be in only one direction regardless of alternating polarity. Consequently, the minimum reading indicates a balance. A rectifier meter is satisfactory at power line frequencies, at all audio frequencies, and at low radio frequencies.

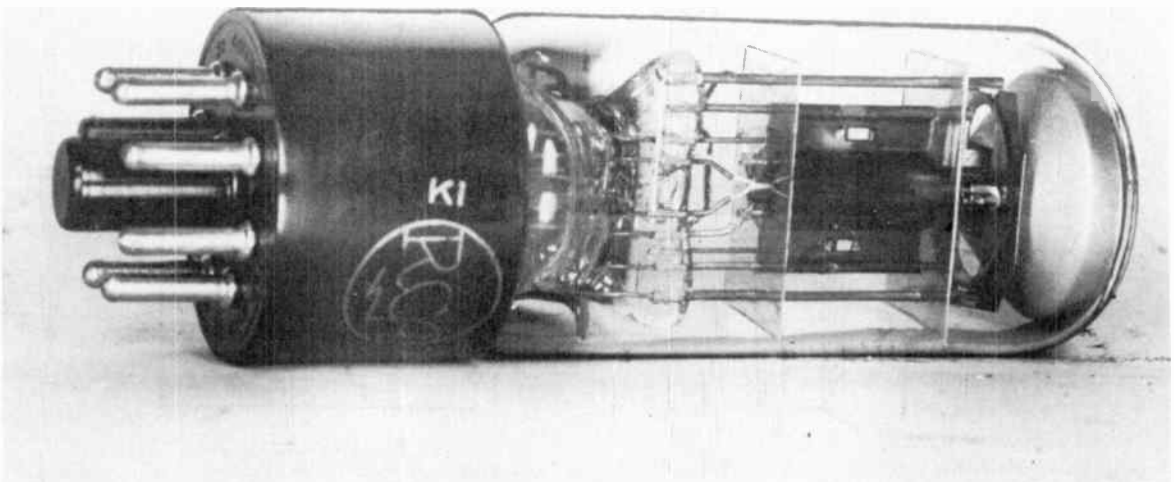


Fig. 74–6. An electron-ray tube with self-contained triode amplifier.

ELECTRON-RAY INDICATORS. A visual indicator widely used for all kinds of a-c bridges, as well as for other testing instruments of both service and laboratory types, is the electron-ray tube. This type of tube was developed originally for visually indicating whether or not a receiver is correctly tuned to resonance for a received signal, and still is commonly used for this purpose. A popular name for the electron-ray tube is "magic eye".

One type of electron-ray tube is pictured by Fig. 74–6. In the end of the glass bulb farthest from the base is an open-ended cone shaped "target" whose inner surface toward the end of the bulb is coated with material which fluoresces with a bright green glow when struck by electrons. This fluorescent surface is clearly visible when looking toward the glass end of the bulb. Between the base of the tube and the target structure is a triode amplifier.

The construction at the target end of this type of electron-ray tube is shown at the left in Fig. 74–7. The cathode which serves the triode amplifier extends through a central opening in the conical target. This extension of the cathode supplies the electrons which cause a fluorescent glow on the target. Between one side of the extended cathode and the inner surface of the target is a thin, flat "ray-control electrode" which is attached to and is electrically a continuation of the triode plate. Over the exposed end of the cathode is a light baffle, a small disc that conceals the red glow of the heater.

When voltage on the ray control electrode changes in relation to voltage on the target, the fluorescent pattern on the target is varied as shown by Fig. 74–8, where we are looking at the target through the glass

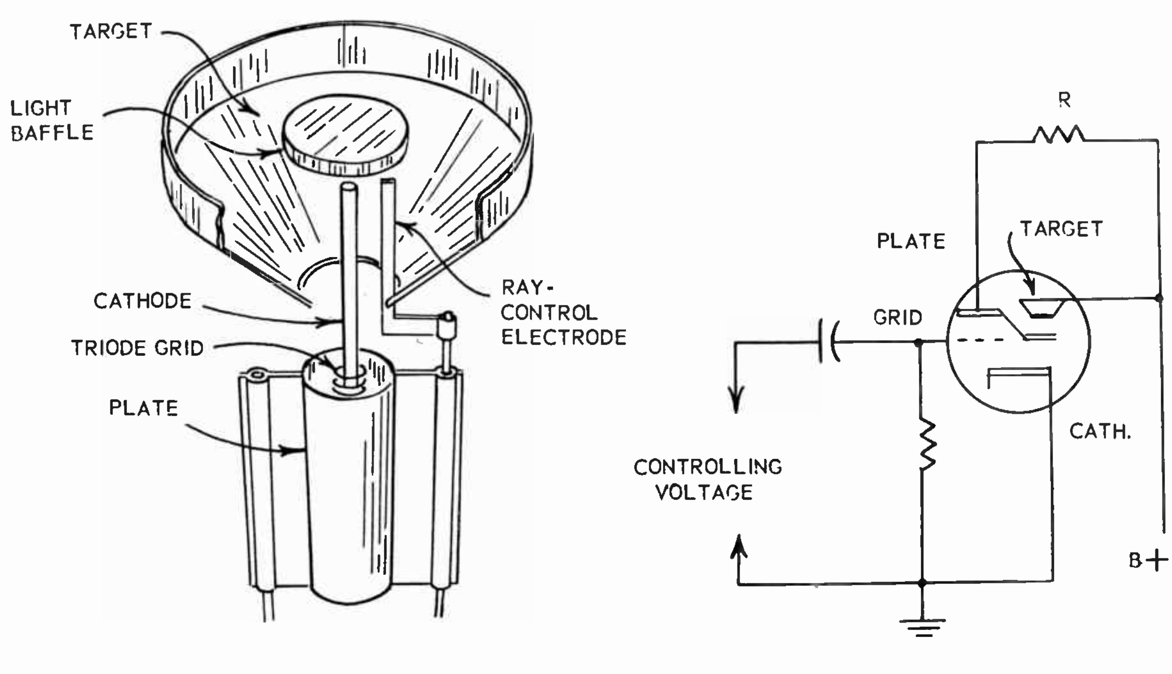


Fig. 74–7. The target end of an electron-ray tube and the circuit for such a tube.

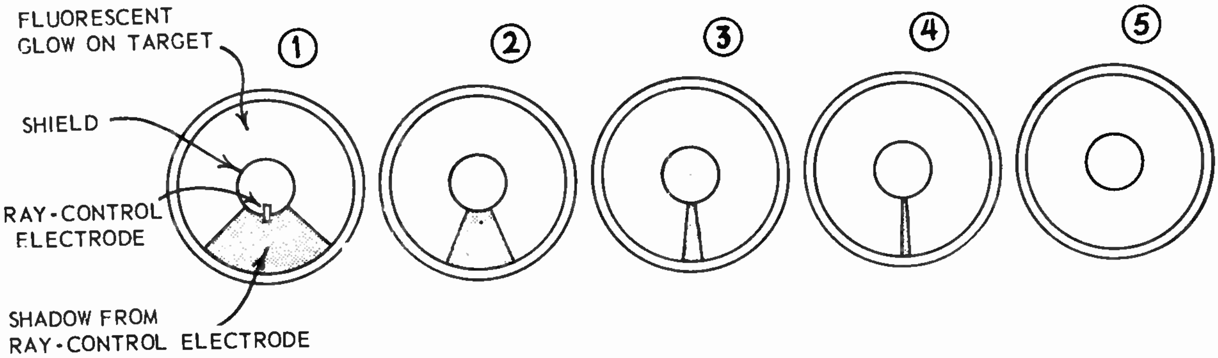


Fig. 74–8. How the shadow of the electron-ray tube indicates balance of a bridge.

end of the bulb. In diagram 1 the ray-control electrode is at a positive voltage equal to only about one-tenth of the target voltage. Then, with respect to the target, the ray-control electrode is strongly negative. The high positive voltage on the target is attracting electrons from the cathode, but the relatively negative ray-control electrode is keeping electrons away from about one-fourth of the target area, or throughout an angle of about 90 degrees. The relatively negative ray-control electrode is casting an electron shadow on the target.

When there is an increase of positive voltage on the ray-control electrode this voltage comes closer to the target voltage, and effectively becomes less negative with respect to the target. Then more of the electrons from the cathode can get around the ray-control electrode, and the shadow becomes narrower, as at 2. The higher the positive voltage on the ray-control electrode, or the less negative it is made with reference to the target, the narrower becomes the shadow angle, as at 3 and 4. Finally, as at 5, there is no shadow at all when voltage on the ray-control electrode becomes equal to a slightly less than half the target voltage. The entire target area now glows bright green. This all-over glow continues no matter how much more positive the ray-control electrode may be made.

How the ray-control voltage and the shadow angle on the target are governed may be explained with the help of Fig. 74–9, which shows connections for using the electron-ray tube as the indicator for an a-c bridge. The target, connected directly to the B plus supply, is at higher positive voltage than any other electrode. Voltage for the triode plate and the ray-control electrode comes through resistance R, which usually is one megohm. Depending on how much current for the plate and ray-control electrode is being taken through this high resistance, it will drop the voltage for the plate and ray-control electrode to some value between approximately one-tenth and one-half of the target voltage. If there is relatively large triode plate current, the positive voltage remaining for the plate and ray-control electrode will be dropped to a low value. With small triode plate current the ray-control voltage will remain comparatively high. Thus we find that voltage on the ray-control electrode is varied by triode plate current.

Triode plate current is regulated by voltage on the triode grid, and since this plate current determines the voltage on the ray-control electrode, it turns out that the ray-control voltage really is varied by changes

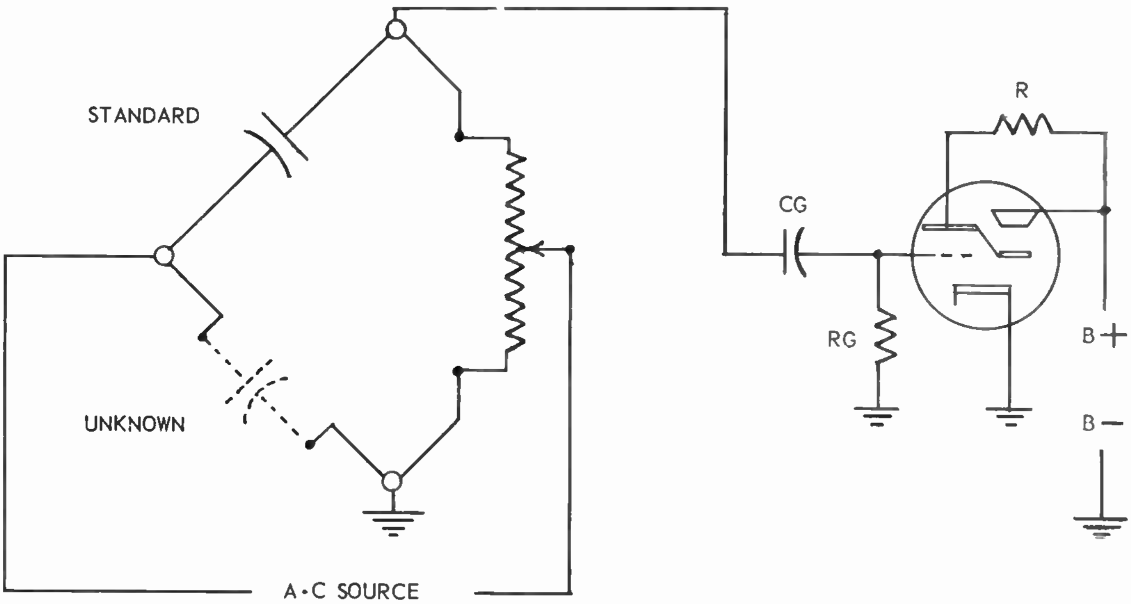


Fig. 74–9. An electron-ray tube connected as a bridge indicator.

of voltage on the triode grid. When the grid is at zero voltage with reference to the cathode there will be a large plate current. This means minimum voltage on the plate and ray-control electrode, and a wide shadow angle. When the grid is highly negative there will be but little plate current. This means high voltage on the plate and ray-control electrode, and either a narrow shadow angle or no shadow at all.

With one type of electron-ray tube, called a sharp cutoff type, the shadow angles of Fig. 74–8 correspond to the following voltages on the triode grid. Diagram 1, zero grid. Diagram 2, grid is 1 volt negative. Diagram 3, grid 2 volts negative. Diagram 4, grid 3 volts negative. Diagram 5, grid at 4 volts negative or at any greater negative voltage.

When this electron-ray tube is connected as the null indicator in an a-c bridge circuit, such as that of Fig. 74–9, alternating voltage from one side of the bridge goes to grid capacitor Cg, and from the other side of the bridge goes through the ground connections to the tube cathode. Capacitor Cg and grid resistor Rg provide grid-leak bias whose negative d-c value is approximately the same as the peak a-c voltage from the bridge.

With the bridge balanced there is zero a-c voltage going to the indicator circuit, and there is zero grid bias on the electron-ray tube. Then the shadow angle becomes of maximum width, as at 1 in Fig. 74–8. When the bridge is not balanced there will be a greater or less a-c voltage from the bridge to the indicator, and grid bias voltage on the electron-ray tube will become correspondingly negative. This unbalanced condition will be indicated by some shadow angle less than maximum, or else by the absence of any shadow at all.

Fig. 74–10. A capacitor tester employing a bridge circuit and providing tests for leakage and power factor.

We have examined a typical method of using an electron-ray tube as the null indicator for an a-c bridge. The circuit might be modified in various ways without affecting the principles of operation. We might use some other type of electron-ray tube, possibly one having two or more than two ray-control electrodes and no self-contained amplifier. This would call for an external amplifier. As we come to other parts of our work in which electron-ray tubes are useful we shall take up their applications, but for the present we may leave these interesting devices and get back to the measurement of capacitances and the testing of capacitors in general.

CAPACITOR TESTERS. Bridges designed for measuring capacitances during service operations usually provide in addition a test for leakage and also for what is called the power factor of electrolytic capacitors. Such an instrument is illustrated by Fig. 74–10. The target end of an electron-ray indicator tube is back of a panel opening which shows near the top of the picture. Lower down is a rectifier tube for the self-contained power supply that furnishes d-c voltages for the electron-ray tube.

Leakage tests are not made with a bridge circuit, but by applying to the capacitor a direct voltage obtained from a divider system on the power supply. Leakage current, if any, may be measured with a sensitive meter, or by applying a resulting voltage to the electron-ray tube, or by using a small neon lamp that flashes intermittently on slight leakages or glows steadily on large ones. The leakage resistance of a capacitor is the resistance of the dielectric in parallel with resistance through and over the surface of the protective insulation that houses the capacitor. It is the resistance offered to a continuous electron flow or to a direct current from one terminal to the other of the capacitor.

The leakage resistance of paper, mica, and ceramic capacitors in good condition always is well in excess of 100 megohms. Leakage current through such resistance is too small to be readily measurable when applying any voltage less than would cause breakdown of the capacitor.

The normal leakage resistance of an electrolytic capacitor in good condition is much lower than that of paper, mica, and ceramic units, and will allow a current that is quite easily measurable. As a result, the sensitivity of leakage indicators for testing electrolytics must be such that no leakage will be shown when the testing voltage applied to a good capacitor does not exceed the rated working voltage of the unit. An electrolytic capacitor which has been idle for even a short time may show high leakage when first tested. But after the d-c testing voltage has been applied for five to ten minutes the internal resistance will rise and the leakage indication should disappear. If leakage still appears, the unit should be replaced. Of course, it is necessary to apply the d-c testing voltage in correct polarity, or there will be a continued large leakage.

When preparing to test any capacitor of more than a few thousandths of a microfarad capacitance you always should short circuit its terminals on each other before disconnecting it from a receiver circuit. After any capacitor has been checked for leakage, always discharge the capacitor by shorting its terminals as soon as disconnected from the tester. These precautions will avoid unpleasant shocks and possible damage to equipment.

In the capacitance bridge circuits shown by Figs. 74–5 and 74–9 the unknown capacitance is compared with a known or standard capacitance by adjusting the ratio arms. We have assumed pure capacitive reactances in both positions. But because the plates and internal connections of all capacitors consist of conductors, both the unknown and the standard capacitor must have internal resistance as well as capacitive reactance.

This resistance of the conductors inside a capacitor must not be confused with the resistance through the dielectric and the insulation, which is the leakage resistance. Electrons should not pass through the path of leakage resistance, but the internal conductor resistance is in the path of all electrons flowing into and out of the capacitor. The conductor resistance opposes electron flow and causes dissipation of energy, just as does any other resistance through which a current must pass. The capacitor really consists of capacitance or capacitive reactance having in series with it the internal conductor resistance.

In paper, mica, and ceramic capacitors which are in good condition the internal conductor resistance is so small as to be negligible so far as routine service tests are concerned. But the internal resistance of electrolytic capacitors may be as much as several hundred ohms, which acts just like a separate resistor connected in series with a perfect capacitor.

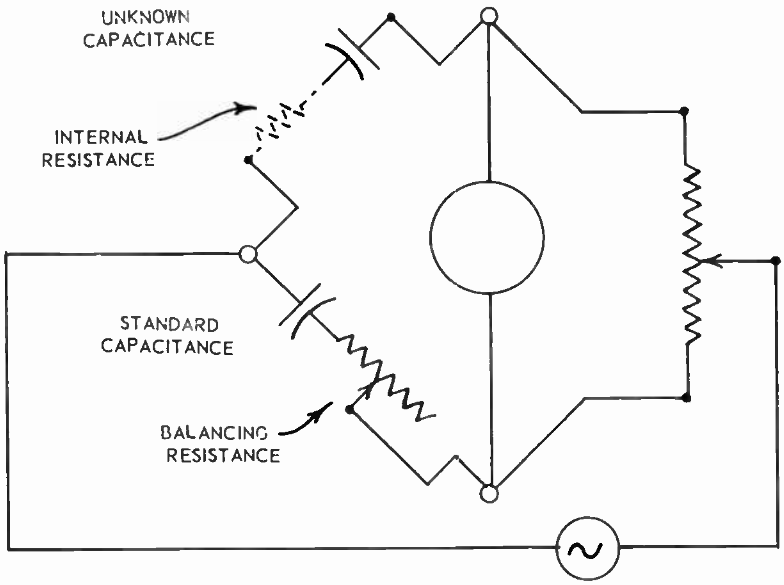


Fig. 74–11. A resistive balance element in the standard arm of a capacitance bridge.

Unless the internal resistance of the tested electrolytic capacitor can be balanced by an equal resistance in series with the standard arm of the bridge it will be impossible to balance the bridge. Even though there were to be a balance between the two capacitive reactances, there could be no null indication unless there were a simultaneous balance between the two resistances. To allow making a resistive balance we may connect an adjustable resistance in series with the standard capacitor, as in Fig. 74-11.

The balancing resistance of capacitor testers usually is marked as a measurement of power factor, because the amount of internal resistance in the capacitor being tested is a measure of power that is dissipated. Technically, the power factor of a capacitor is the cosine of the electrical angle by which the current leads the voltage. When the power factor is normally small, as in good capacitors, it is practically equal to what is called the dissipation factor. The dissipation factor is simply the ratio of the internal resistance in ohms to the capacitive reactance in ohms. Since reactance varies with applied frequency, the dissipation factor will vary with the frequency of bridge supply voltage. When the power factor adjustment on a capacitor tester is graduated in percentages, these are the percentages of the capacitive reactance which are represented by the internal resistance of the tested capacitor.

The greater the power factor or dissipation factor the more power will be turned into heat by the capacitor. Also, a high power factor lengthens the time constant of the circuit in which the capacitor is used, and thus reduces the effective capacitance below the actual capacitance of the tested capacitor.

When a capacitance bridge is equipped with a power factor adjustment, the first step is to leave this adjustment turned off or short circuited while balancing the bridge for capacitance to obtain the best possible null indication. Then the power factor adjustment is set where it gives the best possible null indication.

These two steps balance the bridge first for capacitance, then for resistance. It may be necessary to work back and forth between the capacitance adjustment and power factor adjustment to get the sharpest null indication.

The internal resistance which is involved in the power factor or dissipation factor cannot be measured with an ohmmeter, but only with a bridge. The ohmmeter applies direct voltage, which can force no measurable current through the capacitor dielectric and which, therefore, can force no measureable current through the internal resistance in series with the dielectric of the capacitor.

A few of the many variations in practical bridge circuits are illustrated by Fig. 74–12. Three capacitance ranges are provided in diagram 1 by using three standard capacitors at S, and cutting in one or the

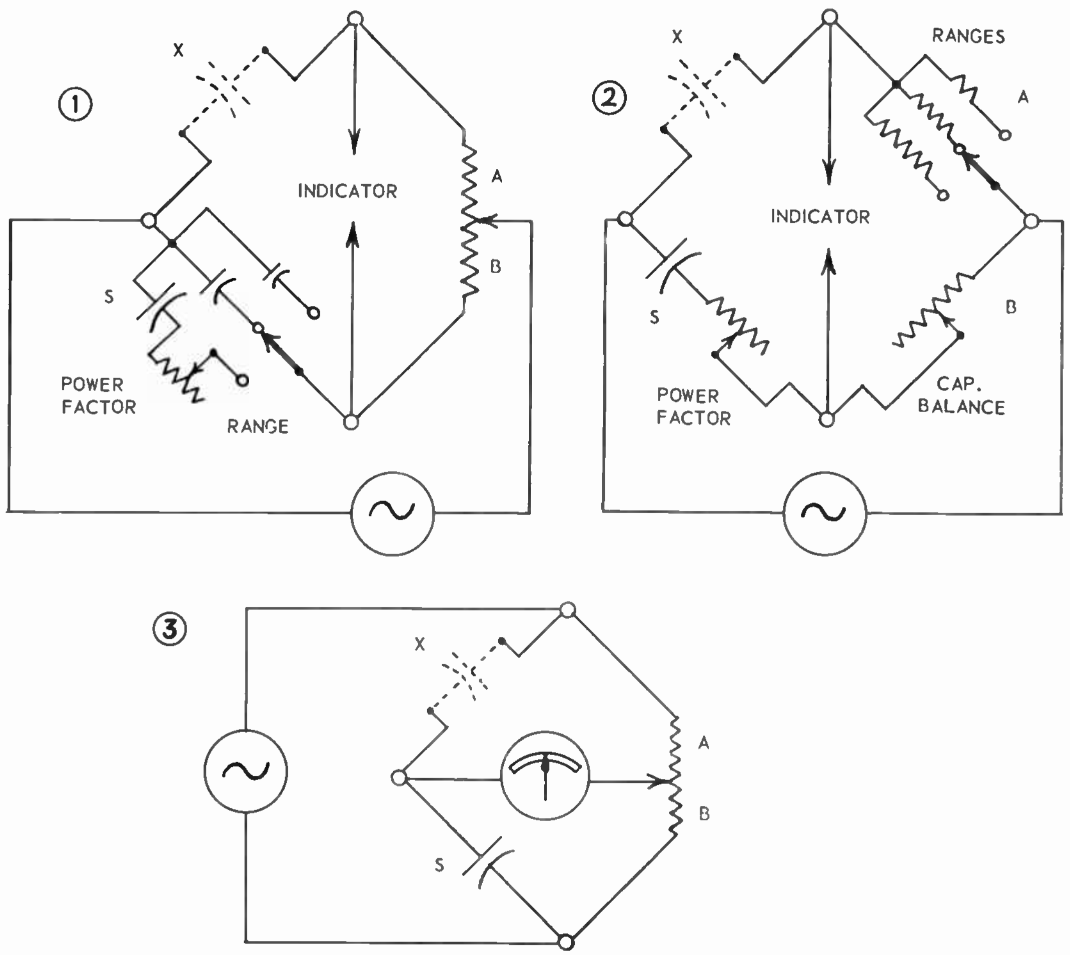


Fig. 74–12. Capacitance bridges with range selectors, 1 and 2, and interchanged connections of the power supply and indicator for a bridge, 3.

other by means of a rotary switch. A power factor adjustment is in series only with the largest standard capacitor, since power factor measurement or resistive balance would be needed only for electrolytic capacitors which come in the larger capacitances.

In diagram 2 the ratio arms, A and B, consist of separate adjustable resistors rather than a potentiometer. The continuously adjustable arm B is used for balancing the bridge for capacitance, and it would be graduated in micro-microfarads and microfarads. Arm A consists of several different resistances which may be cut in one at a time by a rotary selector switch to provide several ranges. With this arrangement the capacitance scale on arm B will have nearly uniform graduations from one end to the other, while with a potentiometer as in diagram 1 the scale will be crowded at one end and open at the other end.

Up to this point all of our diagrams have shown the a-c source connected to a point between the ratio arms, and the indicator connected across the ratio arms. Any of the bridges may have the power supply and indicator connections interchanged, as shown at 3 in Fig. 74–12. A bridge may be designed to work in just the same way with these connections as with those shown in earlier circuits.

INDUCTANCE BRIDGES. Although inductance is one of the most important properties of all tuned circuits, as well as of many filtering and decoupling circuits, this property seldom is measured during service work. It is usual practice to check tuning coils, couplers, peaking coils, chokes, and transformer windings only for open circuits or excessive resistance by using a continuity tester or ohmmeter. We assume that the value of inductance must have been correct to begin with, and that it is unlikely to change. Yet measurements of inductance may be decidedly helpful when winding a tuning coil or a high-frequency choke, when selecting a choke which is to be effective within certain frequency limits or in a power supply filter, when checking transformer ratios, testing for shorted turns, and for determination of Q-factors.

The quickest and easiest way to measure inductance and Q-factor is with a bridge. We might use a type of bridge which compares an unknown inductance with a standard inductance, as at the right in Fig. 74–5. The objection to this method is that a standard inductor of high accuracy, high Q-factor, and low distributed capacitance costs about as much as a complete service type capacitance tester. Fortunately, precision capacitors of equivalent excellence cost much less, and for this reason many inductance bridges have been designed which compare the inductive reactance of a tested inductor with the capacitive reactance of a standard capacitor.

Fig. 74–13 shows the inside of a service type inductance bridge using capacitors as standards of reactance. At the top of the picture is a range selector switch that controls two arms of the bridge. In the center is a variable resistance unit forming a third arm. The inductor to be measured is the fourth arm of the bridge. Down below may be seen a rectifier microammeter for the indicator and an adjustable resistor which allows making a resistive balance and also measures the Q-factor of the inductor being tested. This instrument requires an external a-c power supply.

The conductors in all coils and all windings have some resistance. These elements also have energy losses which are the equivalent of resistance. Therefore, it is necessary to balance an inductance bridge for resistance or dissipation of the inductor as well as for its inductance if a sharp null indication is to be obtained.

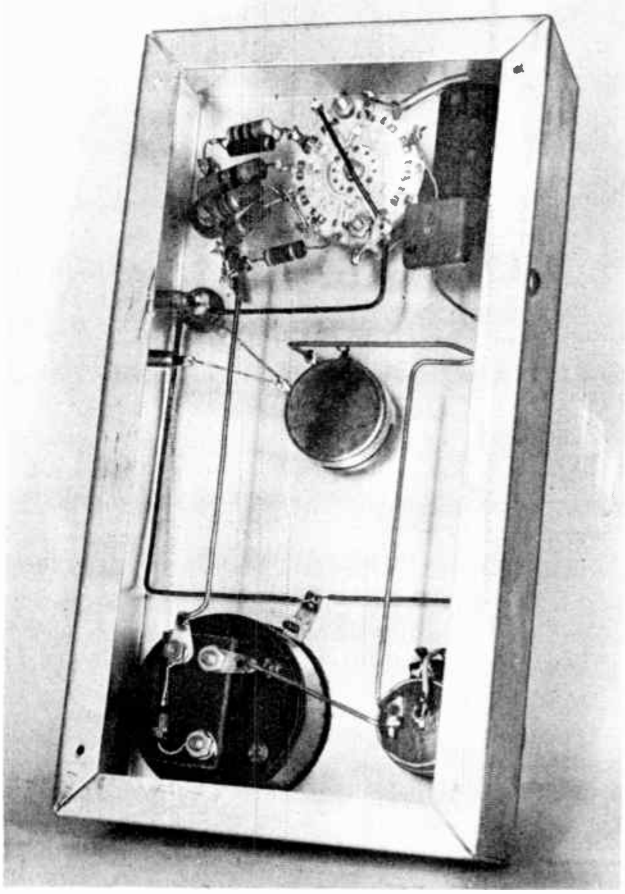


Fig. 74—13. An inductance bridge with an a-c microammeter for the indicator.

The circuit for a commonly used type of inductance bridge is shown by Fig. 74—14. In one arm we have the unknown inductance, X. In a second arm, A, is a variable resistor whose dial may be graduated in units of inductance. A third arm, R, consists of several resistors which are selected by the range switch. In the fourth arm are standard capacitors, S, also selected by the range switch, and in parallel with this arm is the resistive balance element from whose setting may be determined the Q-factor of the tested inductor.

With the arms thus arranged we have a Maxwell bridge, in which connection of the resistive-balance element allows checking inductors whose Q-factor is 10 or more. By placing the resistive balance element in series with the standard capacitors we would have a Hay bridge, suited for checking inductors having Q-factors greater than 10. By interchanging the positions of the elements in arms A and S while leaving the

X and R arms as shown we would have an Owen bridge. With still another arrangement, called the Schering bridge, there would be both capacitance and resistance in all arms except that for the unknown. These few variations are mentioned chiefly to bring out the fact that a large book might be written on the subject of bridges.

In all of these bridges the inductance of the measured element is proportional to the product of resistances and capacitance in the other three arms. In the Maxwell bridge of Fig. 74–14, and also in the Hay and Owen bridges, the formula for unknown inductance is simple.

Microhenrys = A in ohms x R in ohms x S in microfarads

It may seem strange that we don't have to consider the frequency applied to the bridge when measuring inductance. This is because any increase of frequency causes the inductive reactance of the measured unit to go up, while the capacitive reactance of the standard goes down. Any lower applied frequency drops the inductive reactance and raises the capacitive reactance. The product of the two reactances remains the same no matter how the applied frequency is changed. The Owen bridge also is independent of frequency, and the Hay bridge is practically so when the Q-factor of the measured inductor is more than 10.

Inductive reactance, and corresponding inductance, are determined by adjusting resistor A or the inductance dial of Fig. 74–14 to balance the bridge as sharply as possible. Then the resistive balance or Q-factor dial is adjusted to balance the bridge for resistance or dissipation of the tested inductor. The Q-factor of an inductor is the ratio of its inductive reactance to its resistance or equivalent energy loss, both in ohms. Consequently, the resistive balance dial may be graduated to read Q-factors.

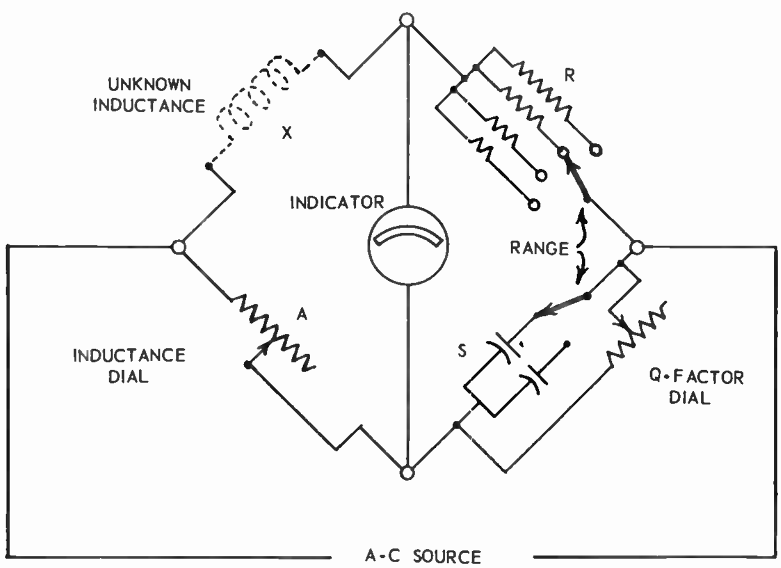


Fig. 74–14. Circuit connections for an inductance bridge.

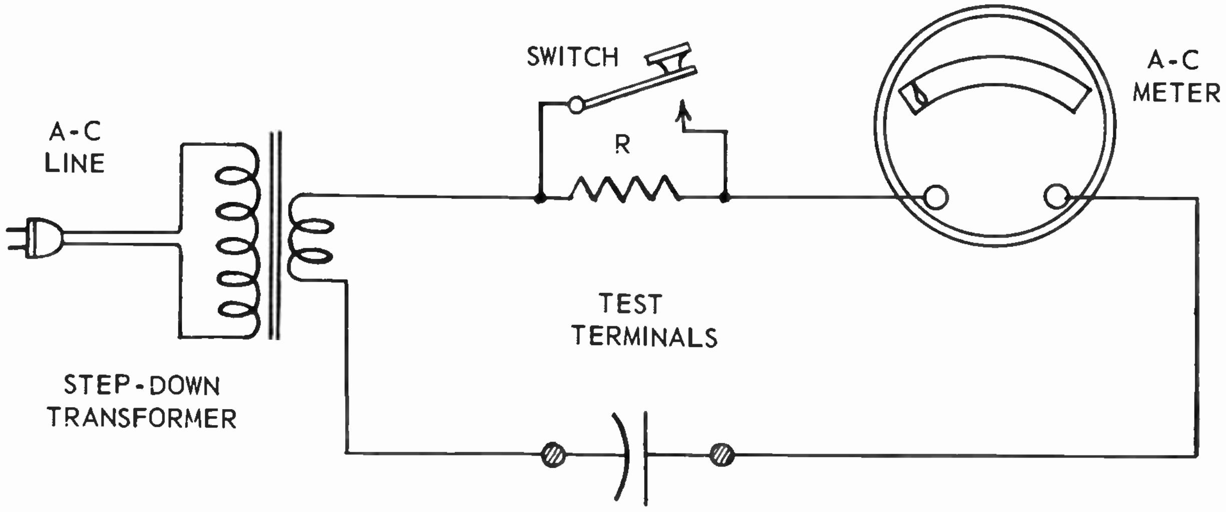


Fig. 74–15. Capacitance tester circuit employing an a-c current meter.

OTHER METHODS OF MEASUREMENT. All of the methods described in another lesson for measuring resistance by means of a d-c voltmeter and milliammeter, or with a d-c voltmeter of known internal resistance, may be adapted for measurement of capacitance and inductance by using a-c meters. Resistance in ohms, as measured by means of d-c instruments, is opposition to flow of direct current. Reactance in ohms is opposition to flow of alternating current. If we employ a-c voltmeters and milliammeters instead of d-c instruments it is possible to measure inductive or capacitive reactances in ohms.

Inductance is directly proportional to inductive reactance. If our instruments can be made to measure the inductive reactance of a coil or winding it is an easy matter to translate this measurement into equivalent inductance in henrys, millihenrys, or microhenrys. This may be done with simple formulas or by suitably graduating a dial scale for the testing instrument.

Capacitance is inversely proportional to capacitive reactance. Any measured value of capacitive reactance may be changed to the corresponding value of capacitance in microfarads or micro-microfarads, either by using formulas or by suitably graduating a meter dial scale.

An a-c milliammeter or a-c voltmeter may be used for measurement of capacitance with a circuit as simple as shown by Fig. 74–15. The capacitor to be tested is placed in series with a source of alternating current, the meter, and a protective resistance of a value which allows full scale deflection of the meter with the test terminals short circuited on each other. The press button switch is normally open when the capacitor is connected between the test terminals. If the meter reads full scale the capacitor is internally shorted, and if it reads zero the capacitor is internally open. If neither of these indications occur, the switch is pressed closed and the meter reading is observed.

You might graduate the meter scale in units of capacitance by testing a number of capacitors of known

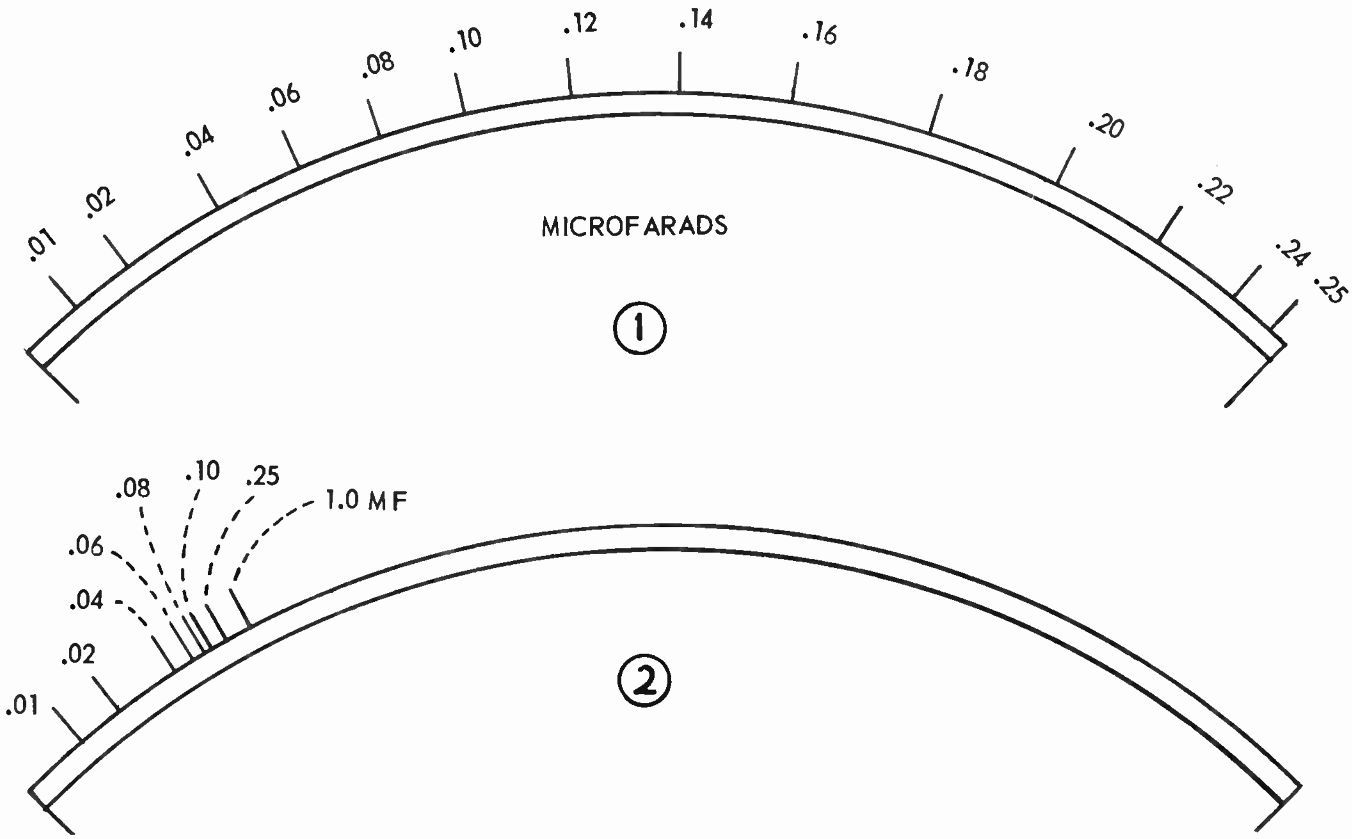


Fig. 74–16. Resistance of the testing meter determines whether capacitance scales are open or crowded.

values and marking their capacitances on the scale at the corresponding pointer deflections. The pointer would deflect proportionately to current in the testing circuit and the meter. For any given supply frequency the current would depend on the applied a-c voltage and on the total opposition to current flow. This total opposition would consist of capacitive reactance in the measured capacitor plus resistance in the meter.

As an example, assume that the supply frequency is 60 cycles, that 10 volts is the applied a-c potential difference, and that you check capacitors having values from 0.01 to 0.25 microfarad. When using a certain meter the scale might turn out to be easily read, as at 1 of Fig. 74–16. The graduations are quite uniformly spaced from end to end. When using a different meter the scale markings might start out from the low end as at 2. They are easily readable at the low end, but all the higher readings crowd together. Even though you check a capacitance as large as 1.0 microfarad, the reading would not be very much higher than for 0.1 microfarad.

The difference between the two scales is due wholly to difference between the internal resistances of the two meters. Resistance of the meter used in making scale 1 was about 5,300 ohms, and of the meter for scale 2 about 65,000 ohms. Why this makes such a difference between scales is explained as follows.

Whether a meter goes by the name of voltmeter or milliammeter, it is really a current meter because its pointer deflects proportionately to current. A voltmeter has relatively high resistance, while a milliammeter has less resistance. Otherwise they are alike. The current in either meter, and resulting pointer deflection, always are proportional to applied voltage and to the total of resistance and reactance in the test circuit.

If the meter resistance is small in comparison with the capacitive reactance of the tested capacitor, the graduations for capacitance will be quite uniform. This is because the small amount of meter resistance has little effect on total circuit resistance. This total is determined almost entirely by the relatively large reactance of the capacitor tested.

On the other hand, if meter resistance is large in comparison with the capacitive reactance, the circuit current will be determined almost entirely by the relatively large meter resistance. When the tested capacitances are great enough to have small reactances, the circuit current and the meter reading will remain almost unchanged.

Here is an example of what happens. The reactance of 0.5 mf at 60 cycles is about 5,300 ohms, and of 1.0 mf it is about 2,650 ohms. If meter resistance is 65,000 ohms the total circuit opposition when testing 0.5 mf is 70,300 ohms, and with 10 volts applied the current is about 0.1423 ma. When testing 1.0 mf the total opposition is 67,650 ohms and current is 0.1478 ma. The difference between currents is less than 4 per cent for a 100 per cent increase of capacitance tested.

The crowded scale at 2 in Fig. 74–16 could be spread out over the dial by applying a higher voltage, but the high end still would be badly crowded. In order to measure large capacitances, which have small reactances, the thing to do is use a meter of very low resistance. With an a-c meter having internal resistance of only about 100 ohms it becomes easily possible to read capacitances from 2 to 40 mf on a well distributed scale, but all smaller capacitances would be crowded into the first few degrees of the scale.

Measurements of inductance with a-c current meters involve the same general principles as measurements of capacitance. A very sensitive meter is needed for checking the small reactances of small inductances, while a meter of higher internal resistance will check the larger inductances such as power chokes and transformer windings.

TELEVISION

LESSON NO. 75

MEASUREMENTS WITH RESONANT CIRCUITS

Training Your Mind...

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

COMMERCIAL
TRADES
INSTITUTE

Training Your Mind
C.T.I.
Perfecting Your Skill

Chicago, Ill.

LESSON NO. 75

MEASUREMENTS WITH RESONANT CIRCUITS

Bridges and other test circuits which have been described are entirely adequate for measuring all the larger values of capacitance and inductance. But bridges ordinarily available for service work seldom will measure with acceptable accuracy any capacitances much smaller than 15 micro-microfarads nor any inductances smaller than about 10 microhenrys, such as commonly found in television circuits.

It is, however, not at all difficult to measure the small capacitances and inductances by connecting them into a resonant circuit, as in Fig. 75-1. Here we have a coil and a capacitor containing the two properties, inductance and capacitance, which produce resonance. Connected to the resonant circuit is a signal generator which provides radio-frequency voltages, and also an a-c vacuum tube voltmeter which will show a peak of voltage or a maximum voltage when the condition of resonance exists.

Resonance exists when there are certain frequencies, capacitances, and inductances. If we know any two of these factors with reasonable accuracy it is easy to determine the third one by using simple formulas. That is, knowing frequency and inductance allows determining capacitance, knowing frequency and capacitance allows determining inductance, and knowing both inductance and capacitance would allow computing the frequency of resonance.

Inductances from less than one microhenry to more than 1,200 microhenrys can be measured by using known frequencies from any ordinary signal generator and a single fixed capacitor, which, in a mica type of one per cent accuracy, costs very little.

Capacitances could be measured by using a known frequency with a known inductance, but we won't do it this way because you are unlikely to have available any inductor of sufficient precision for such work. Our method will be to measure one capacitance with the help of another capacitance, or else to use two known frequencies.

With known frequencies from the signal generator we may do many other useful things. We can determine the approximate operating limits of any tuned coupler or transformer before it is installed in a receiver. We can measure the resonant frequencies of antennas and stubs. Using also the vacuum tube voltmeter we can determine the Q-factor of tuned circuits.

Before talking about practical methods of making these measurements we must consider some of the errors most likely to occur, and how they may be reduced. Errors result chiefly from stray capacitances

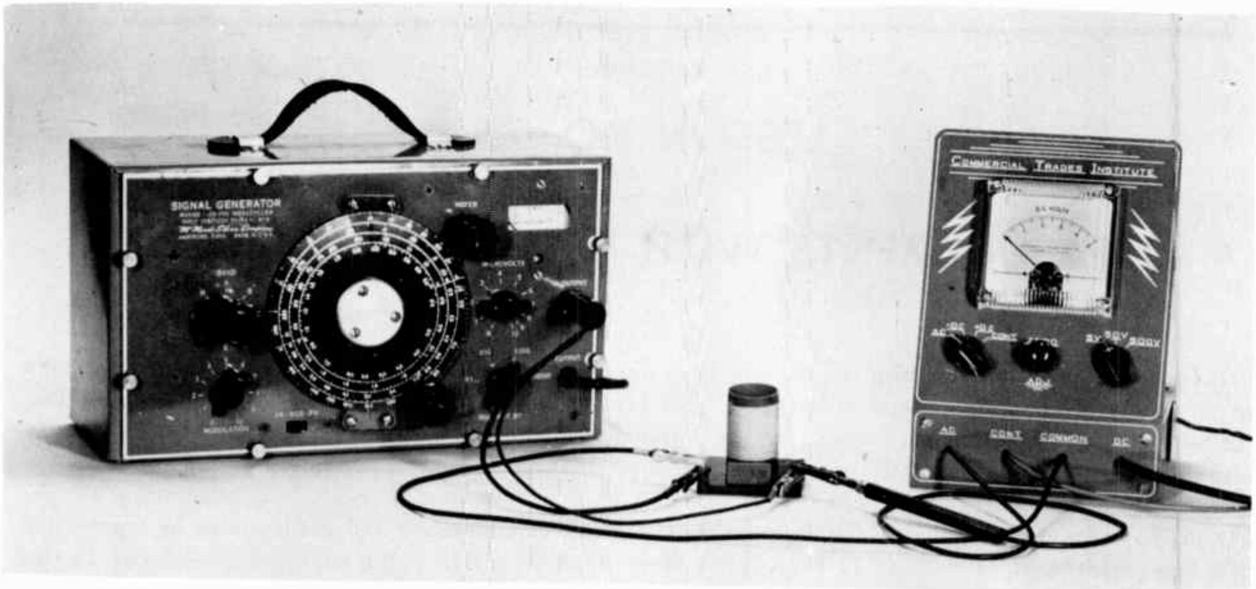


Fig. 75-1. Using a signal generator and a vacuum tube voltmeter for measuring the inductance of a coil.

and inductances which add themselves to the resonant circuit. The effect of stray capacitances is lessened by grounding one side of the resonant circuit, as at the left in Fig. 75-2. If the capacitor is a variable type, ground its rotor side.

Make all connections as short as possible to begin with, and never move them as a test progresses, for

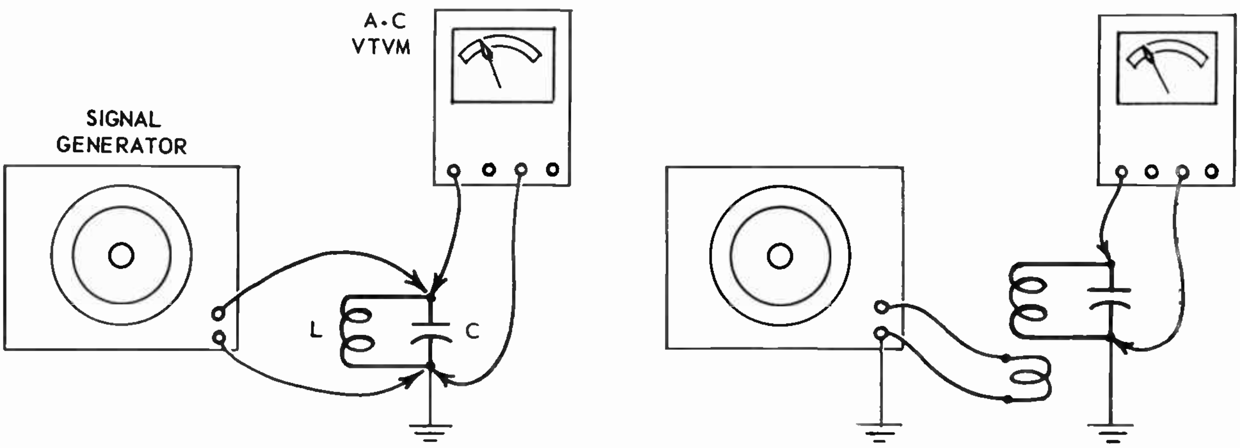


Fig. 75-2. Direct coupling and inductive coupling to a tested circuit.

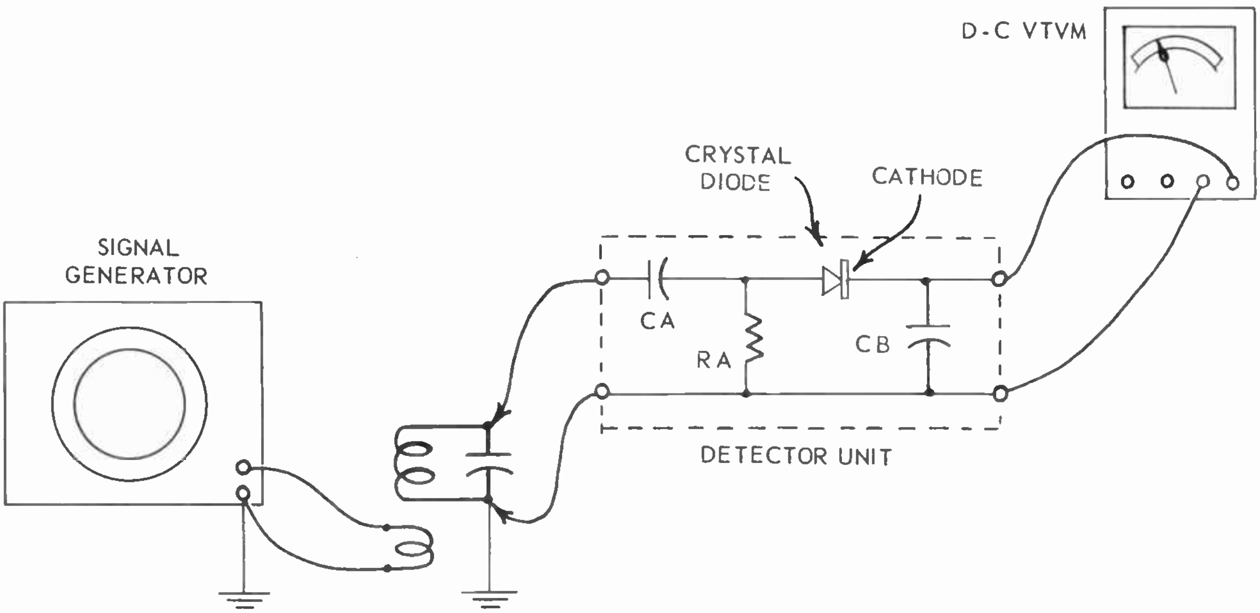


Fig. 75-3. Using a high-frequency detector unit between the tested circuit and the d-c input to the VTVM.

this would alter the stray capacitance and comparative readings would be meaningless. Keep odd pieces of metal and insulating materials away from the coil which is in the resonant circuit.

When testing any capacitor and inductor which are connected for series resonance in normal use, temporarily reconnect them for parallel resonance. The resonant frequency and all other factors are the same for both parallel and series resonance, but capacitance and inductance of the leads to the signal generator and resonance indicator have much less effect on a parallel resonant circuit.

40 The high side and ground leads of the signal generator must be connected directly to the resonant circuit if the generator is capable of delivering only small r-f power. But if the generator output can be made fairly high, always introduce the r-f energy into the resonant circuit by inductive coupling, as at the right in Fig. 75-2. This coupling improves the accuracy of all measurements. The coupling coil which is connected to the signal generator leads need have no more than 10 to 15 turns of any size wire, wound to a diameter of a half-inch to one inch. Such a coupling coil works well for all frequencies from 100 kc to more than 10 mc.

Place the coupling coil just close enough to the coil in the resonant circuit to obtain readable deflections on the indicator. Don't change the relative positions of the two coils during a test. If you do this or anything else which changes the degree of coupling, and the mutual inductance, start all over again.

If the signal generator gives only a weak output, it is possible to have higher peak readings on the VTVM by connecting a crystal detector between the resonant circuit and the positive d-c probe or terminal of the VTVM, as in Fig. 75-3. The rectifier may be a 1N34 germanium crystal diode or any equivalent type. Capacitor Ca should be of about 0.005 mf, Cb may be about 0.01 mf, and resistor Ra about one megohm.

HARMONIC FREQUENCIES. All signal generators produce the frequency for which they are tuned. This is the fundamental frequency. Practically all of them produce at the same time a frequency which is twice the fundamental, called the second harmonic, and also a frequency three times the fundamental, which is the third harmonic, and so on. Nearly every generator will deliver rather strong second harmonics, many will furnish a strong third harmonic, and some go much higher.

When using a signal generator as a source of known frequencies it is important to work with funda-

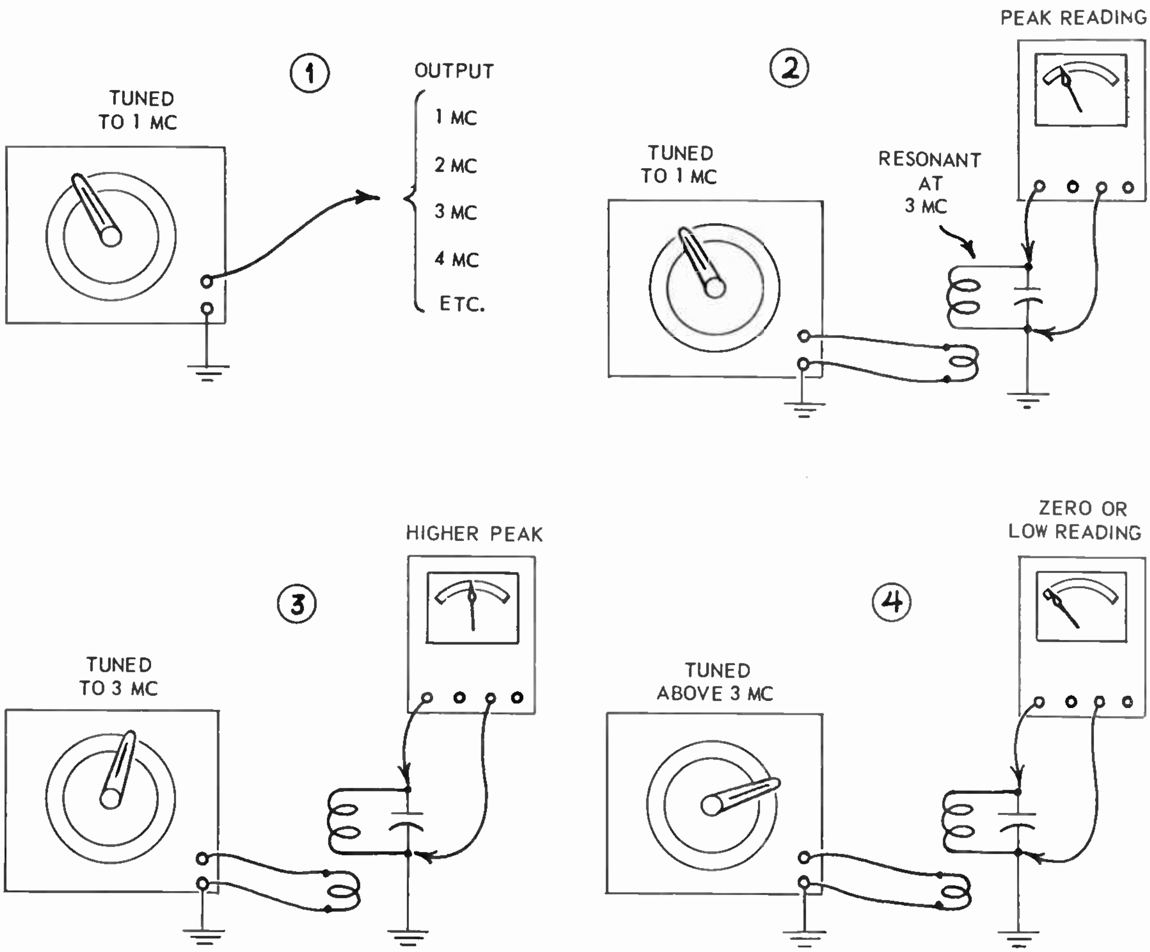


Fig. 75-4. Some of the effects caused by harmonic frequencies from a signal generator.

mentals rather than harmonics, or, if you do work with a harmonic, you must know which one it is. With a generator tuned to a fundamental frequency, as marked on the tuning dial, it is reasonably certain that the output of the generator contains also a second harmonic, probably a third harmonic, and maybe a great number of higher harmonics in considerable strength - all at the same time.

How to tell the difference between a fundamental frequency and a harmonic is illustrated by Fig. 75-4. At 1 a signal generator is represented as being tuned for 1,000 kc or 1 mc. The output will contain the 1-mc fundamental and harmonics of 2 mc, 3 mc, 4 mc, and probably still higher multiples of the fundamental. At 2 the generator is coupled to a circuit which is resonant at 3 mc. The circuit is excited by the third harmonic from the generator, and the VTVM shows a peak reading when the generator tuning control is brought to a scale marking of 1 mc.

In diagram 3 the generator is tuned to 3 mc. This strongly excites the circuit which is resonant at 3 mc, and the VTVM will show a peak of voltage. At 4 the generator is tuned above 3 mc. The reading of the VTVM will return to a lower voltage for all frequencies higher than 3 mc, because neither the fundamental nor any harmonic then can excite with any strength, the circuit which is resonant at 3 mc.

The resonant circuit that is being tested, however, will also show an indication of resonance at all harmonic frequencies of its fundamental. When the generator is tuned to 6 mc, the meter will show a weak indication since the resonant circuit is being shock excited at its second harmonic frequency.

You can be assured that the resonant circuit will produce maximum voltage when the correct frequency is supplied from the generator and this is the only logical and practical method to use. Find the frequency from the generator that will produce maximum voltage across the tuned circuit, and you can safely assume that not only is the generator operating at the fundamental frequency, but that the resonant circuit is being excited at its fundamental frequency.

CAPACITANCE MEASUREMENTS. The method of capacitance measurement illustrated by Fig. 75-5 requires an adjustable or variable capacitor with a dial marked in values of adjusted capacitance, or graduated in any way that allows determining this capacitance. Connect this "standard" capacitor into a parallel resonant circuit as at 1, using any convenient coil for the inductor. The inductance need not be known. Then proceed as follows.

1. Adjust the standard to near its maximum capacitance, say to 300 mmf, and carefully note this capacitance value.

2. Tune the signal generator for resonance, as indicated by a peak reading of the VTVM. It is not necessary to make a note of the resonant frequency.

3. Connect the unknown capacitance in parallel with the standard, without changing the generator frequency. If the unknown is a variable type, connect its rotor to the rotor of the standard. See step 2.

4. Reduce the standard capacitance to again obtain an indication of resonance. Note the new value of standard capacitance.

5. The unknown capacitance is equal to the difference between the higher and lower capacitance

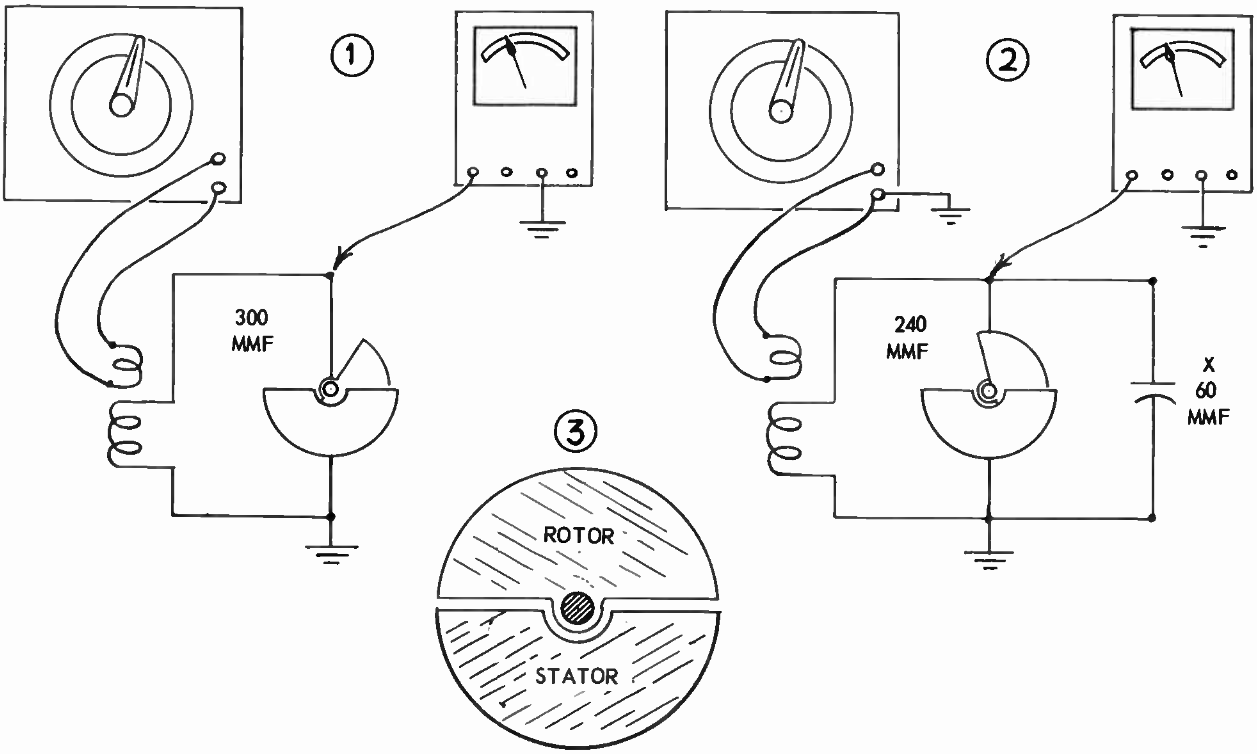


Fig. 75-5. Measuring an unknown capacitance with a calibrated variable capacitor.

settings of the standard. In the illustration the two settings of the standard are 300 mmf and 240 mmf. The unknown must be 60 mmf.

The standard capacitor may be any commercial type having semi-circular plates, as shown at 3 in Fig. 75-5. This is called a straight line capacitance type. You can safely assume that maximum and minimum capacitances are as specified by the manufacturer, and that intermediate values are proportional to how far the rotor is turned. The precision of such a standard is not too good, but it should suffice for all service measurements.

A slightly different method makes use of the following steps.

1. Connect the unknown capacitance across the inductor, without the standard, and tune the generator for a resonance indication on the VTVM.

2. Disconnect the unknown capacitance and connect the standard variable unit in its place.

3. Without changing the generator frequency, adjust the standard capacitor to obtain a reasonant reading.

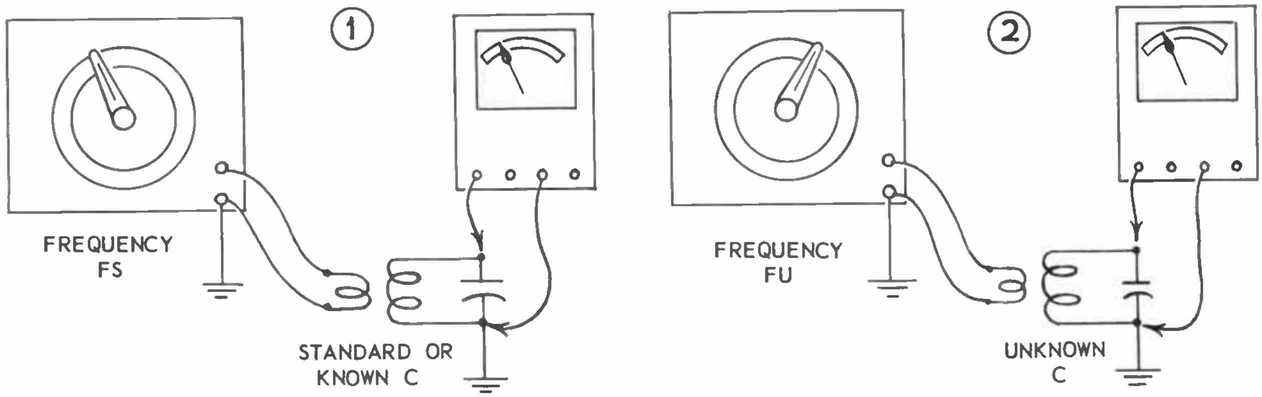


Fig. 75-6. Measuring capacitance by using two frequencies from a signal generator.

4. The unknown capacitance is equal to the adjusted value of the standard.

A third method does not require a calibrated adjustable standard capacitor, but only a fixed capacitor of known value and any convenient inductor, whose value need not be known. The procedure is as follows.

1. Connect the fixed capacitor of known value, your standard, into the resonant circuit as at 1 in Fig. 75-6.

2. Tune the signal generator for resonance. Note the frequency as closely as possible and call this frequency Fs.

3. Remove the standard capacitor and connect the unknown in its place, as at 2. The unknown may be a variable or fixed capacitor of any kind.

4. Retune the signal generator for resonance. Call this frequency Fu.

5. Square the number of kilocycles or megacycles of the first frequency, Fs, from step 2 above.

6. Multiply the square of Fs by the capacitance, in mf or mmf, of your standard capacitor.

7. Divide this product by the square of frequency Fu, from step 4.

8. The result of the division is the unknown capacitance in mf or mmf, depending on which of these units was used in step 6.

The actual unknown capacitance will be somewhat less than the value determined by any method of measurement, because stray capacitances add themselves to the measured capacitance. With short test connections well spaced and carefully arranged the difference may be as small as 10 to 15 mmf, but usually it will be on the order of 25 to 50 mmf.

INDUCTANCE MEASUREMENTS. The effective or apparent inductance of a tuning coil, a choke, or any other inductor may be measured as illustrated by Fig. 75-7. The inductor is connected in a parallel resonant circuit with a known capacitance, and the signal generator is tuned for a resonance indication. Then the apparent inductance is found from either of the following formulas. In one formula the capacitance is in microfarads and the frequency is in kilocycles, while in the other formula the capacitance is in micro-microfarads and the frequency is in megacycles.

10

$$\text{Microhenrys} = \frac{25\ 330}{\text{mf} \times \text{kc}^2} \qquad \text{Microhenrys} = \frac{25\ 330}{\text{mmf} \times \text{mc}^2}$$

Diagram 1 shows the inductor connected to a fixed capacitor whose capacitance is known with precision. The frequency of the signal generator is varied to obtain resonance. Then the generator frequency and the fixed capacitance are used in one of the formulas.

Diagram 2 shows the inductor connected to a variable calibrated capacitor, such as the kind mentioned earlier. The generator is set for any frequency and left there while the variable capacitor is tuned to resonance. Then the values of frequency and capacitance are used in either of the formulas. Since your signal generator usually is more precisely calibrated for frequency than an ordinary variable capacitor can be calibrated for capacitance, the method of diagram 1 ordinarily is preferable.

When varying the generator frequency be sure to tune to the frequency which gives the strongest resonance indication. Should you happen to use the generator's second harmonic for resonance of the tuned circuit and read the fundamental frequency from the generator tuning dial, the computed inductance will be four times as high as the actual inductance. This is because the generator dial is set at one half the tuned circuit frequency. Working on a third harmonic will give a computed inductance nine times too high.

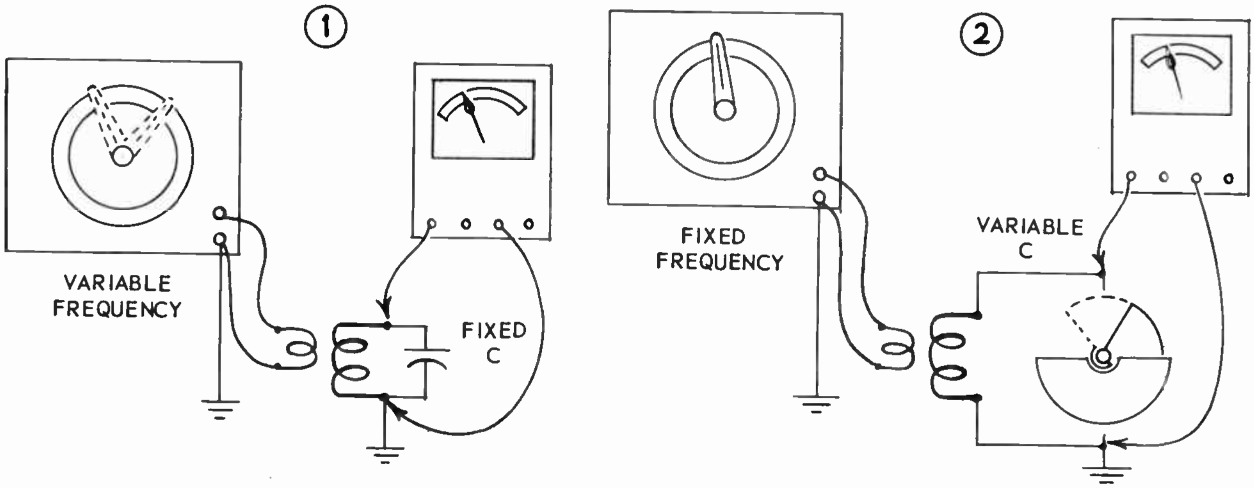


Fig. 75-7. Capacitance measurement by substituting a calibrated variable capacitor for the unknown capacitance.

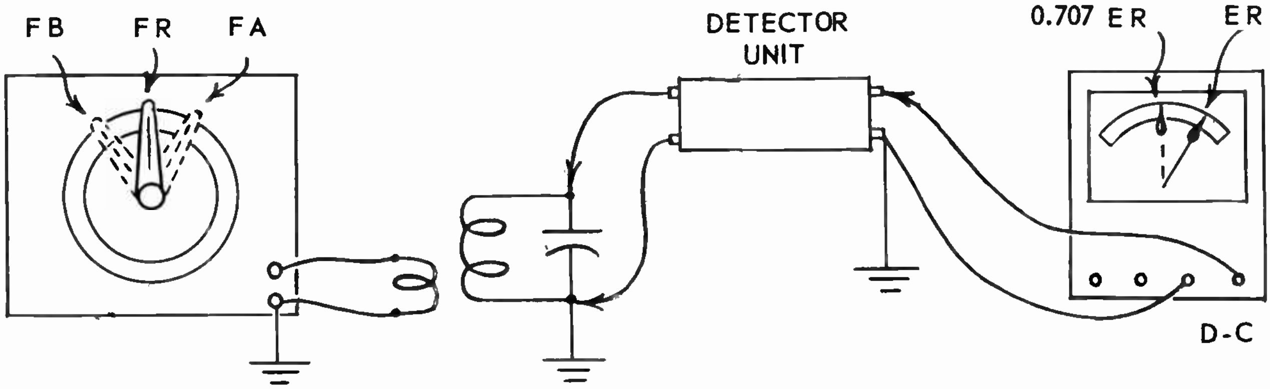


Fig. 75-8. Measuring the Q-factor of a tuned resonant circuit.

Inductance computed from a single measurement, as described, always will be greater than the true inductance. This is why we spoke of effective or apparent inductance. Your measurements are affected by distributed and stray capacitances, which affect the tuning in just the same way as additional inductance. By using the method of diagram 2, where the capacitance is adjustable, it is easy to determine true inductance. This is the way to do it.

Use a frequency low enough to allow setting the variable capacitance rather high on the first test. Call this capacitance Cf, because it corresponds to the fundamental frequency. Now, without altering the generator frequency, reduce the variable capacitance to get a second indication of resonance, which means that you have tuned the resonant circuit to respond to the second harmonic of the generator frequency. Call this lesser capacitance Ch, because it corresponds to the harmonic. Then proceed as follows.

1. Multiply capacitance Ch by 4.

2. Subtract this product from the first capacitance, Cf.

3. Divide the difference by 3. The result of this division is the value of distributed and stray capacitances, in the same unit for measuring the tuning capacitances.

4. Add the computed distributed and stray capacitance (from step 3) to the first tuning capacitance, Cf. Use the sum of these capacitances in either of the formulas for inductance. Now the formula will give the true inductance.

Q-FACTOR MEASUREMENT. The Q-factor of an entire tuned circuit, including the inductor, the capacitor, and their connections, may be measured as illustrated by Fig. 75-8. The signal generator is coupled to the resonant circuit in the usual way, and r-f voltage across the resonant circuit is measured with the vacuum tube voltmeter. Unless the signal generator is capable of delivering an unusually high output voltage it will be necessary to use the detector circuit of Fig. 75-3 in order to obtain the necessary high voltage readings on the indicator meter.

Measurement is carried out as follows.

1. Tune the signal generator to obtain a peak resonant voltage on the VTVM. Call the resonant frequency Fr. The value must be noted as precisely as possible.

2. Read the voltage at resonance as closely as possible. Multiply this voltage by 0.707 and note where this decreased voltage will be located on the meter scale.

3. Tune the generator to the frequency above resonance at which the meter reading drops to the voltage of 0.707 times the resonant voltage. Read this frequency as accurately as possible and call it Fa.

4. Tune the generator to the frequency below resonance at which the meter reading again becomes 0.707 times the resonant voltage. Call this frequency Fb.

5. Subtract the lower frequency, Fb, from the higher frequency, Fa.

6. Divide the resonant frequency by the difference between low and high frequencies as found in step 5. The result of the division is the Q-factor of the resonant circuit.

As an example, a particular coil-capacitor circuit was found to be resonant at 3.300 mc (Fr). The higher frequency at which VTVM voltage dropped to 0.707 of the resonant voltage was 3.326 mc (Fa), and the lower frequency for the same voltage was 3.255 mc (Fb). The difference between high and low frequencies is 0.071 mc. Dividing 0.071 mc into the resonant frequency of 3.300 mc gives 46.5 as the Q-factor of the tested circuit.

V *RESONANT FREQUENCY MEASUREMENT*. The frequency for which any combination of inductance and capacitance is resonant may be determined by energizing the resonant circuit from the signal genera-

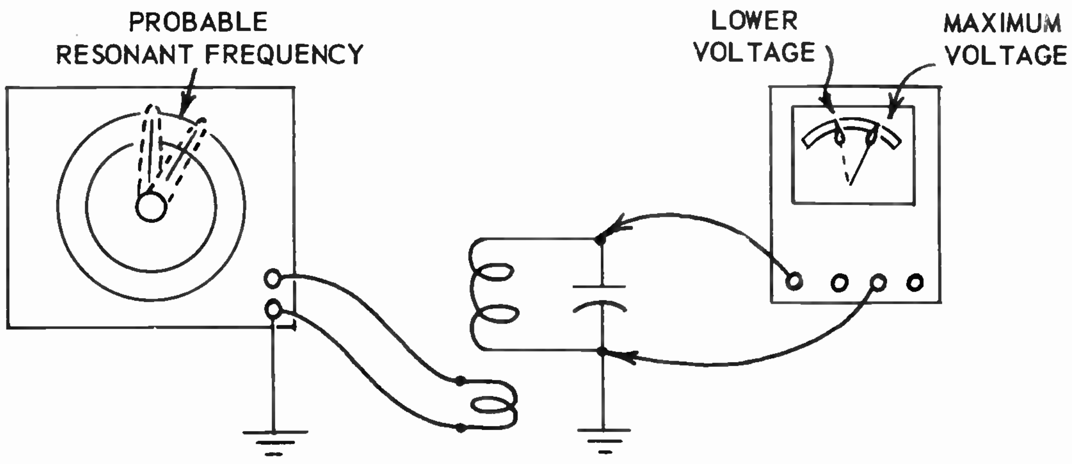


Fig. 75-9. Determining the resonant frequency of a circuit having low Q-factor.

tor and noting the frequency which causes maximum voltage on the VTVM. The indication of resonance will be much sharper and more accurate with inductive coupling to the resonant circuit, as at the right in Fig. 75-2, rather than with direct coupling as at the left.

If the resonant circuit has a low Q-factor the resonant peak may be so broad as to make identification of the true resonant frequency quite difficult. Then we may resort to the method illustrated by Fig. 75-9, which is as follows.

1. Tune the signal generator to obtain the highest possible reading on the VTVM.

2. Retune the generator to any slightly higher frequency which causes the VTVM to read a lower voltage which is easily and precisely identified on the meter scale. Note this higher frequency as closely as possible.

3. Tune the generator back through resonance and to a lower frequency at which the VTVM reads exactly the same lower voltage as in step 2. Carefully note this frequency.

4. It may be assumed that the resonant frequency is midway between the higher and lower frequencies that give equal voltages.

The lower the Q-factor of the tested circuit, and the broader its frequency response, the better this method works out. For high-Q circuits it is not so accurate as determination of resonant frequency from the relatively sharp peak voltage which then appears on the VTVM.

Any adjustable coupler, transformer, or trap unit may be tuned quite closely to the frequency at which it is to operate when installed. It is simply necessary to couple the signal generator output to the inductor of the tunable element, connect the VTVM, set the generator to the desired frequency, and adjust the tunable element for resonance. This is a common practice in service work involving replacement of some unit which is to be resonant at an operating frequency.

Some slight readjustment will be needed after the unit is installed, because stray capacitances, tube capacitances, and inductances of wiring connections then will affect the tuned circuit. The final setting will, however, be made in much less time than without the pretuning adjustment.

2 *NATURAL FREQUENCIES.* The natural frequency of an inductor is the frequency at which the inductance and the distributed capacitance are "self-resonant" without any external capacitor connected to the inductor. Tuning coils of all kinds usually have natural frequencies so high that they cannot be measured with any ordinary signal generator. R-f chokes and some other inductors having many turns and large inductances often have enough distributed capacitance to be self-resonant, and to cause trouble, at frequencies normally existing in associated circuits. The distributed capacitance, in combination with the inductance, produces a parallel resonant circuit which is complete within the inductor.

The natural frequency or self-resonant frequency may be measured as at the left in Fig. 75-10. The two terminals of the inductor are connected together through a resistor of about 5 ohms, or they may be short circuited on each other. Then the resonant frequency is determined by tuning the signal generator in the usual way. R-f chokes designed for use in standard broadcast radio receivers often are self-resonant at television carrier frequencies, and sometimes at television intermediate frequencies.

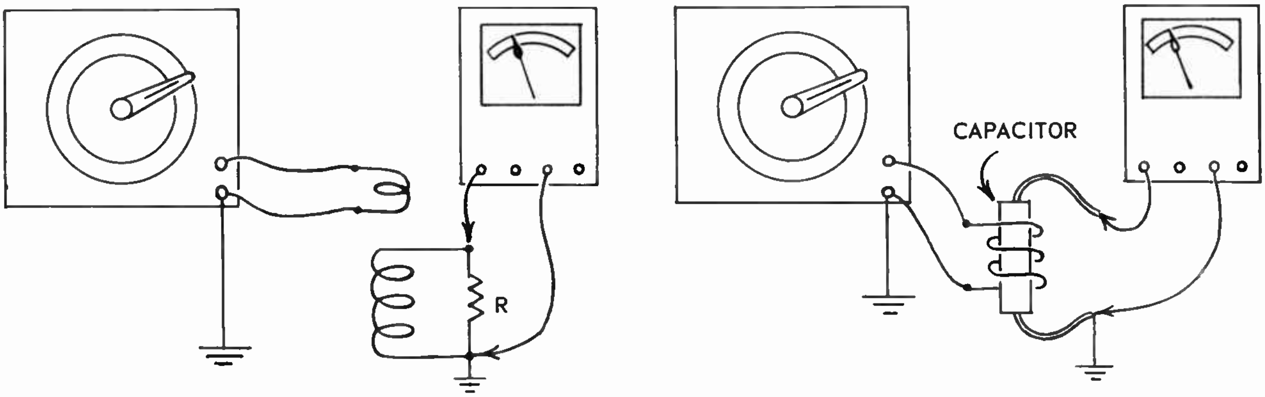


Fig. 75-10. Measuring the self-resonant frequency of an inductor (left) and of a capacitor (right).

Paper capacitors which are not of non-inductive construction often are self-resonant at television carrier and intermediate frequencies. This is because the many turns of foil which form the capacitor plates have enough inductance to resonant with the capacitance of the unit at some of the frequencies mentioned. It is best to make the test coupling from the signal generator by means of a coil which may be slipped over the capacitor, as at the right in Fig. 75-10. Then the leads from the VTVM are connected to the pigtails or other terminals of the capacitor. The self-resonant frequency is the frequency to which the generator is tuned for maximum reading of the VTVM, assuming, of course, that you are not working on a harmonic.

HETERODYNE FREQUENCY METERS. Most of the frequency meters which will measure the frequency of r-f carrier voltages, oscillator voltages, or i-f voltages, originally were designed for use around amateur transmitting equipment. Signal strengths in such equipment are higher than available in television and radio receivers, and the meters require more r-f energy than can be obtained from most receiver circuits.

There are a few types of frequency meters which require but little power from a measured circuit, and consequently are useful in our servicing work. Probably the best known and most commonly used instrument of this class is the heterodyne frequency meter, sometimes called a heterodyne oscillator or heterodyne detector.

The principle is illustrated by Fig. 75-11. An r-f voltage taken from the source of unknown frequency is coupled through a blocking capacitor to the input of any kind of detector, which is used as a mixer. To the mixer input is coupled also a voltage of known frequency, usually secured from an accurately calibrated oscillator which may be tuned through a wide range of frequencies. The unknown and known frequencies beat together in the mixer, just as the carrier and r-f oscillator frequencies beat together in the mixer of a superheterodyne receiver.

The output of the mixer contains a frequency equal to the difference between the known and unknown. As you vary the known frequency to bring it closer and closer to the unknown, their difference beat fre-

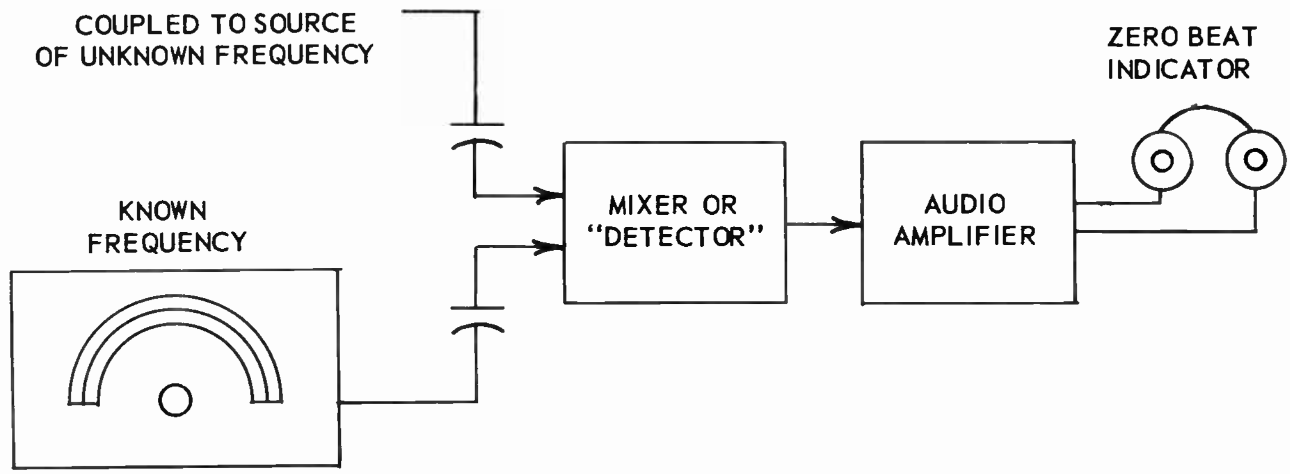


Fig. 75-11. The principal parts of a heterodyne frequency meter.

quency from the mixer output becomes ever lower. Finally, when the known frequency is of exactly the same value as the unknown, the beat frequency becomes zero and we have the condition called zero beat.

The output of the mixer is connected to any device which will indicate the condition of zero beat, and equality of the known and unknown frequencies. The zero beat indicator often is a headphone. When the known and unknown frequencies get close enough together to cause beats at an audio frequency there is a high-pitched note from the phone. The pitch becomes lower as the frequencies are brought still closer. At zero beat the sound from the phone drops to zero. Then the unknown frequency is equal to the adjustable known frequency, which is read from the tuning dial of the oscillator.

Instead of a headphone the zero beat indicator may be an a-c meter, whose reading becomes zero when the two frequencies are equal. It is possible also to use an electron-ray tube in much the same way as for the null indicator on an a-c bridge. Commercial forms of heterodyne frequency meters include the variably tuned calibrated oscillator, the mixer, an audio amplifier for the mixer output, and either some form of visual indicator or else terminals to which may be connected a headphone. There are also input terminals to which may be connected the leads from the source of unknown frequency. Some signal generators include a mixer and audio amplifier which, with the regular oscillator of the generator, allow using the instrument as a heterodyne frequency meter as well as a signal generator.

Frequency meters are employed in either of two ways. First, they may be used to measure the radio frequency at which any circuit actually is operating. This circuit may be in a receiver or in any other kind of equipment. For making such measurements the unknown frequency from the circuit being checked is coupled to the input of the frequency meter. Then the oscillator of the meter is tuned for zero beat, and the unknown frequency is read directly from the dial of the frequency meter.

The second way of using the frequency meter allows aligning or adjusting any oscillator which is to operate at some certain radio frequency. The circuit to be aligned is coupled to the input of the heterodyne frequency meter. The oscillator of the meter is tuned to the frequency at which the circuit is to operate. Then the trimmers or other tuning adjustments of the circuit are aligned to obtain zero beat. The circuit then is operating at the frequency indicated on the dial of the frequency meter.

Diagram 1 of Fig. 75-12 shows a connection for checking the frequency of a radio carrier (from a broadcasting station). The meter input lead is clipped to the high side of the signal grid circuit of a converter tube, and the meter is tuned for zero beat. The capacitance of the meter connection will detune the grid circuit to some extent.

At 2 the frequency meter is being used for checking or adjusting an r-f oscillator frequency, which here is from the oscillator circuit of the converter tube. Usually it is necessary only to lay the free end of of the coupling wire near the oscillator coil. If necessary, the pickup may be increased by attaching to the end of the coupling wire a small-diameter coil of five to ten turns, with one end of the coil left open.

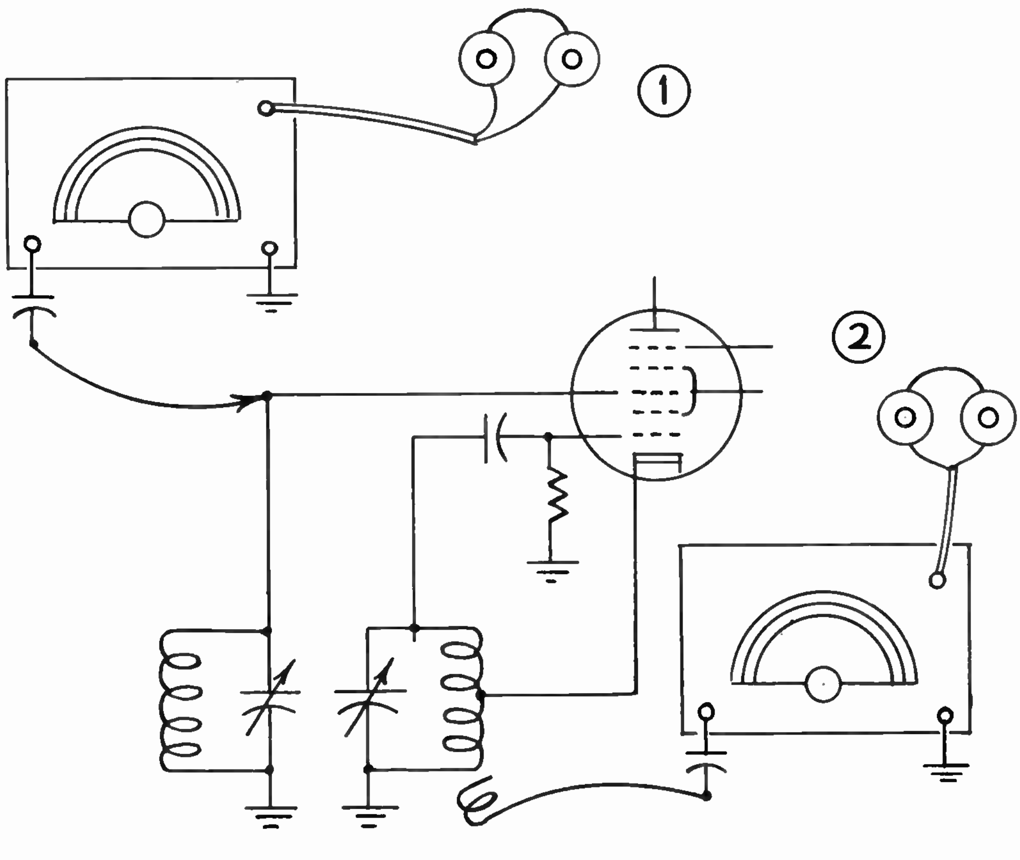


Fig. 75-12. Some methods of using a heterodyne frequency meter.

A small pickup coil or search coil on the end of the input lead may be used for obtaining r-f voltage from any circuit in any device which is operating at high frequency.

The strongest beat frequencies and the most distinct indication of zero beat will be secured when no harmonic frequencies are involved, when the frequency meter is tuned to the same frequency as that from the source of an unknown frequency. If the meter dial is tuned to one-half or to one-third or to any other simple fraction of the unknown frequency the corresponding harmonic of the meter frequency will beat with the unknown to cause relatively weak indications. If the source of unknown frequency is producing harmonics of its own fundamental frequency, these unknown harmonics will cause beats when the frequency meter is tuned to twice or three times or to any other whole number of times the unknown fundamental. The surest way to avoid harmonic effects is to reduce the sensitivity of the frequency meter until zero beat is indicated at only one setting of the dial rather than at several settings.

An experimental or temporary heterodyne frequency meter may be made up as shown by Fig. 75-13. The source of known frequency is any available r-f signal generator. The mixer is the audio detector of any standard broadcast radio receiver. The amplifier and zero beat indicator are the a-f amplifier and loud speaker of the receiver.

The low sides (ground or B-minus) of the signal generator and the source of unknown frequency are connected to chassis ground or B-minus of the receiver whose detector is to be used. The high side lead from the signal generator is connected through a blocking capacitor of about 0.001 mf to the plate or plates of the diode detector. No connections in the receiver need be opened or otherwise altered. The

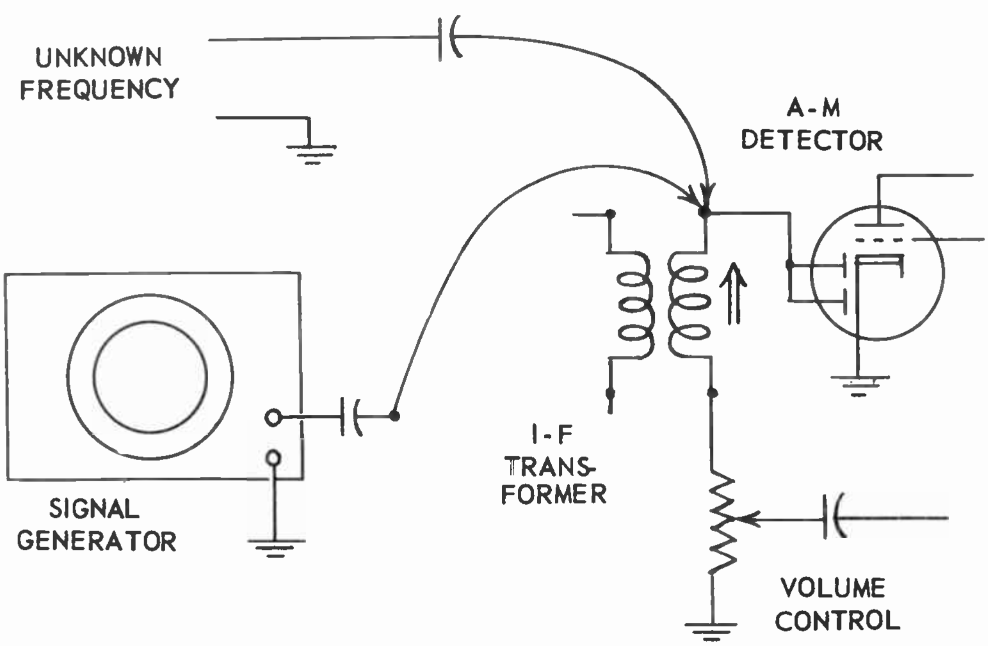


Fig. 75-13. A signal generator and the detector of a radio receiver used as a heterodyne frequency meter.

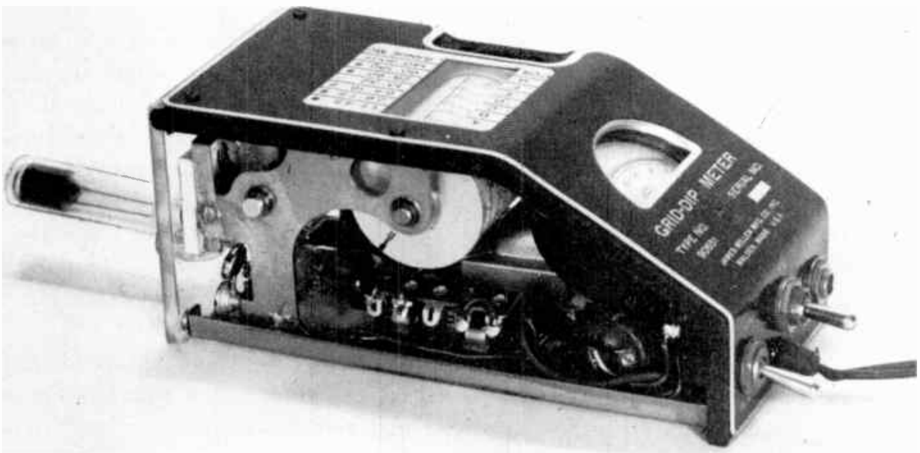


Fig. 75-14. A grid dip meter which is small enough to be held by one hand while tuned with the thumb of the same hand.

pickup lead from the source of unknown frequency is coupled through another 0.001 mf capacitor to the same detector diode plates.

Turn on the receiver and tune it to a point on its dial where no broadcast station can be heard with the volume control at maximum. Leave the volume control at maximum and slowly tune the signal generator until a high-pitched note is heard from the loud speaker of the receiver. Carefully tune the generator until the sound is zero, in between two regions of high pitch. Reduce the receiver volume control until there is a recognizable zero beat at only one generator frequency. Now the unknown frequency may be read directly from the tuning dial of the signal generator. This scheme works well all the way from the lowest generator frequency, usually around 100 kc, up to 20 or 30 mc or to even higher frequencies.

GRID DIP METERS. One of the most useful measuring instruments for any service operations involving radio frequencies is the grid dip meter, often called a grid dip oscillator. This type of meter is essentially a calibrated r-f oscillator in a circuit containing a milliammeter which indicates resonance. Fig. 75-14 is a picture of a grid dip meter with part of the housing removed to show some of the internal construction.

The action is most easily explained by reference to the circuit diagram at 1 in Fig. 75-15 which represents a typical grid dip meter with the exception of the B-power supply, which usually is built into the instrument. The oscillator here is a Colpitts type, although other types may be used. Tuning is by means of the variable two-section capacitor Ca-Cb.

The externally mounted tuning coil has plug terminals that fit into jack openings on the instrument housing. One such coil may be seen in Fig. 75-14. With any one coil plugged in, the oscillator may be tuned through a frequency range of approximately two or two and one-half to one. By using as many as ten different coils to provide ten ranges it is possible to tune from below the lowest intermediate frequency used in standard broadcast receivers to well above the highest carrier frequency in the very-high frequency television channels.

In the grid circuit of the oscillator, between grid resistor Rg and ground, is a milliammeter that indicates grid current. Depending on the frequency range being used and on the frequency to which the oscillator is tuned within that range, the pointer of the milliammeter normally stands somewhere between one-quarter and three-quarters of full scale. This milliammeter is the resonance indicator.

The grid dip meter is intended primarily for measurement of the resonant frequency of circuits or parts which are not connected into any live circuit, or which, if connected, are not carrying voltages or currents while measurements are made. For such purposes the external coil of the instrument is placed near

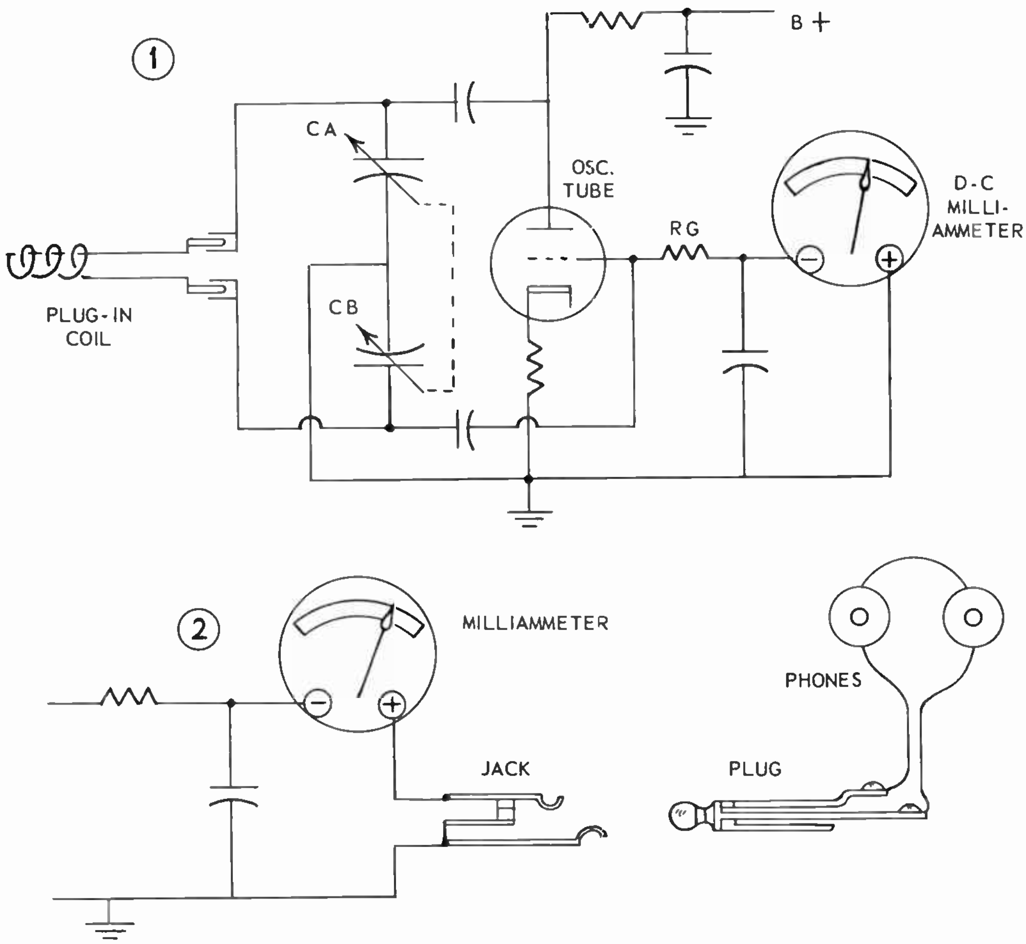


Fig. 75-15. Internal connections of one type of grid dip meter.

enough to the measured circuit or part to allow inductive coupling. Then the meter is tuned through a range of frequencies which is thought to include the resonant frequency of the circuit or part being tested.

a Some energy is taken from the oscillator circuit by the tested circuit at all frequencies. But when the grid dip oscillator is tuned to the frequency at which the coupled circuit or part is resonant the absorbed energy increases. This causes the oscillator plate current to increase and the grid current to decrease sharply. This decrease of grid current is indicated by the milliammeter, whose pointer drops or "dips" at resonance. This indicates that the resonant frequency of the tested circuit or part is the same as shown by the tuning dial of the grid dip meter.

When the instrument is tuned from one end to the other of a range of frequencies, the oscillator grid current with most designs increases or decreases at a nearly constant rate. But as the tuning goes through the resonant frequency of the coupled circuit or part, there is a sharp drop of the meter pointer, followed by a rise on the other side of resonance.

The grid dip meter may be used also for alignment or adjustment of any tunable coupled circuit, whether the circuit or part containing it is connected into a receiver or has been isolated and removed. The instrument is tuned to the frequency at which the coupled circuit is to be made resonant. The milliammeter pointer then will dip when the coupled circuit is aligned or tuned to the frequency at which the grid dip meter has been set. If the coil mounted on the instrument cannot be brought close enough to the tested parts for inductive coupling, one turn of insulated wire may be wound around the meter coil and the other end of this wire placed close to or clipped to the circuit being aligned or otherwise tested.

Tests such as described may be made on transformers, single winding couplers, traps, and all other tuned circuits. It is possible also to measure natural or self-resonant frequencies of chokes or other inductors, also of paper capacitors, and even of resistors, which sometimes are resonant at very-high frequencies.

Capacitances and inductances may be measured by methods described earlier in this lesson. The grid dip oscillator takes the place of the signal generator, and resonance is indicated by a dip of the milliammeter pointer instead of by the VTVM. Otherwise the procedures are unchanged.

The grid dip meter actually is a signal generator, because there is strong radiation from the externally mounted coil. There is no way of regulating or attenuating the strength of radiated signals, nor are these signals confined by shielding such as used in regular signal generators. As a result, when the grid dip meter is tuned to a broadcast radio frequency there will be pickup by all nearby radio receivers. If tuned to a television carrier frequency or intermediate frequency there will be pickup by all television sets in the close vicinity.

One of the more useful applications of the grid dip meter operated as a signal generator is the measurement of Q-factor of any tuned circuit in which r-f voltages may be induced by coupling to the coil of the meter. The procedure is the same as when using a regular signal generator, with a vacuum tube voltmeter employed for measuring maximum induced voltage and 0.707 times the maximum.

Most grid dip meters have in their oscillator grid circuits, in series with the milliammeter, a jack into which may be inserted a plug connected to headphones. This modification of the circuit is shown by diagram 2 of Fig. 75-15. The plug consists of a central conductor ending in a metal ball, with insulating tubing between this condnctor and an outer metallic sleeve. One headphone lead is connected to the central conductor and the other to the outer sleeve.

When the plng is pnshed into the jack one of the jack springs is raised, thus separating the jack contacts and opening the original circuit. The end of the raised spring now contacts the ball of the plug, while the sleeve of the plug makes contact with the lower jack spring. Thus the oscillator grid circuit is completed throngh the phones.

With headphones connected, the grid dip meter may be used as a heterodyne frequency meter. Beat notes will be heard from the phones, and zero beat will be indicated by the sound dropping to zero when the grid dip oscillator is tuned to the resonant or operating frequency of any live circuit to which the meter coil is coupled. The instrument is used exactly like any other heterodyne frequency meter. For more distinct indications of resonance the leads from the plug in the meter jack may be connected to an audio amplifier and loud speaker instead of to the phones.

The grid dip meter is especially useful for measuring resonant frequencies of any style of dipole antenna, and for cutting resonant sections of transmission line to effective wavelengths. To check antenna frequencies the transmission line is temporarily disconnected and the inner ends of the antenna conductors are joined with a very short wire, as at 1 in Fig. 75-16. With the coil of the grid dip meter brought close to this wire connection the resonant frequency of the antenna is that of the meter when a dip occurs.

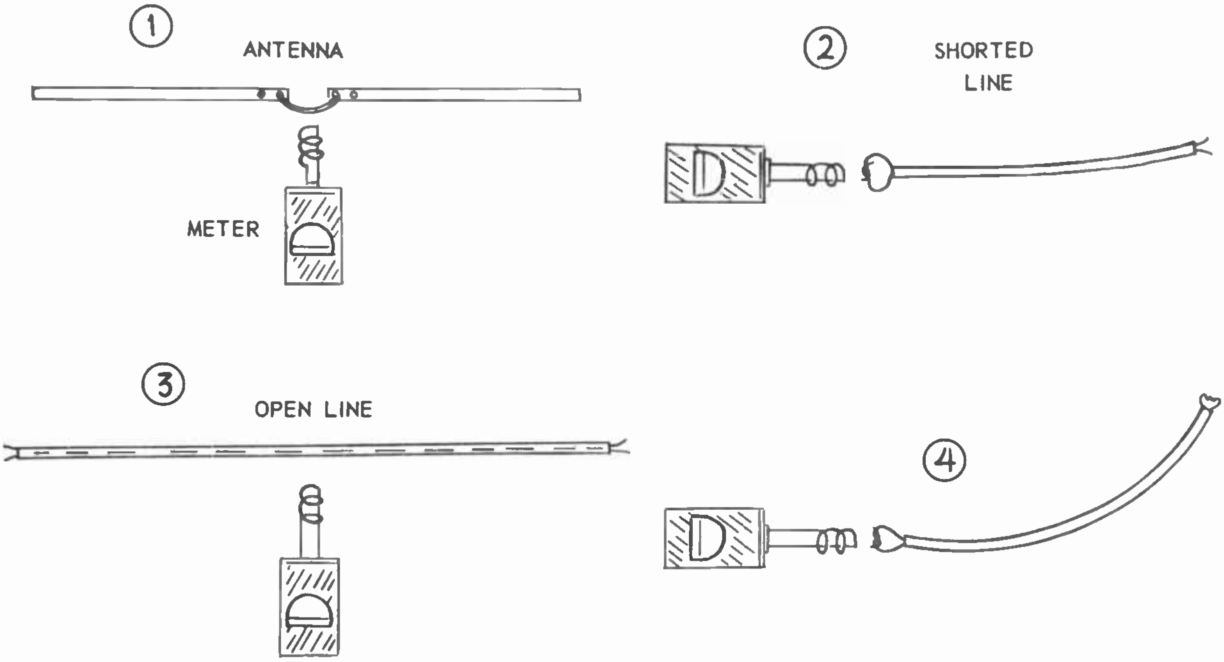


Fig. 75-16. Using the grid dip meter to measure resonance frequency of an antenna, and effective lengths of shorted or open resonant stubs.

There will be a resonance indication at the antenna fundamental frequency, where the antenna is effectively a quarter-wavelength long. There will be other but weaker indications at three times, five times, and other odd multiples of the fundamental frequency. At these higher frequencies the antenna is operating harmonically, at three quarter-wavelengths, five quarter-wavelengths, and so on. In order that these measurements may indicate true performance the antenna should be located and oriented for regular reception.

To check a shorted section of transmission line or a shorted stub couple the coil of the grid dip meter to the shorted conductors at one end of the line, as at 2 in Fig. 75-16. The resonant frequency corresponds to a quarter-wavelength. In other words, to cut a shorted stub to an effective quarter-wavelength, short the two conductors at one end and cut off small pieces of the line until it is resonant at this test.

For an open section or open stub couple the coil of the meter to the center of the line, as at 3, when cutting the line to an effective half-wavelength. Resonance is indicated at a frequency corresponding to a half-wavelength.

If you want to cut a quarter-wave open section or stub, temporarily short the two conductors on one end of the line and measure the resonant frequency just as shown by diagram 2. Then separate the shorted conductors and you have a quarter-wave open section. This method works out because in all these measurements you are trying to obtain a certain length of line, as measured in feet or inches, regardless of whether this length will appear as a parallel resonant circuit, a series resonant circuit, a capacitive reactance, or an inductive reactance.

If a section of line is to be shorted at both ends, couple the grid dip meter to either one of the shorted ends, as shown by diagram 4.

All of these tests may be made on either twin conductor or coaxial line, and on shielded or unshielded types. It is not necessary to consider the velocity factor of the kind of line being used, since you are measuring actual or effective lengths as based directly on frequency or wavelength.

To avoid harmonic responses always continue testing at lower and lower frequencies until reaching the lowest one at which resonance is indicated. There will be other resonances at odd multiples of this lowest frequency, which is the fundamental frequency at which the line is either a quarter-wavelength or a half-wavelength long.

76 → 80

COMMERCIAL TRADES INSTITUTE

KNOWLEDGE + AMBITION = SUCCESS

LEARN . . .

ACT . . .

ENJOY . . .

Training Your Mind

C. T. I.

Perfecting Your Skill

"The heights by great men reached and kept
Were not attained in sudden flight,
But they, while their companions slept,
Were climbing upward in the night."

TELEVISION

LESSON NO. 76

DECOUPLING AND DEGENERATION

Training Your Mind
C. T. I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind

C.T.I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON NO. 76

DECOUPLING AND DEGENERATION

Every tube may be good, there may be no opens, shorts, or accidental grounds, there may be no defective resistors, capacitors, or inductors – and still there may be plenty of trouble. Many of your longest hunts for mysterious causes of queer troubles will end with the discovery that a frequency from one part of a receiver is wandering into other parts where it doesn't belong, or that some circuit which should work only at frequencies supplied to it is producing an entirely different frequency on its own account.

Troubles of this class are numerous enough in broadcast sound receivers, whose operating frequencies include r-f carrier, r-f oscillator, intermediate, audio, and a-c power. The troubles are multiplied in television receivers where we have all the different frequencies shown by Fig. 76-1. With either kind of receiver, matters are made worse by the presence of direct currents for plates, screens, and grid biasing, because these currents like to pick up an a-c component from one section and carry it into other sections.

To avoid these troubles, or to cure them when they exist, we must be able to do three things.

1. Keep all frequencies within the circuits where they belong, and out of other parts and circuits where they may cause trouble.

2. Prevent signals from traveling in directions opposite to those shown by arrows in Fig. 76-1. Such backward travel of signals is called feedback.

3. Prevent the production of frequencies which never should exist.

All three objects may be accomplished by applying effective methods of decoupling, by observing correct practices in grounding, by using shielding where necessary, and by correct dressing of wires and parts. We shall discuss these preventatives or cures in order.

DECOUPLING. To decouple one circuit from other circuits means just the opposite of coupling, it means that couplings are prevented. To understand how couplings may be prevented we first should consider how they are made. When we couple one circuit to another, alternating voltages or currents in the one circuit cause voltages or currents at the same frequency to appear in the other circuit.

There are many ways of coupling two circuits. If you don't recall how it is done, look back at the lesson which deals with coupling one circuit to another. In one general class of couplings we find that a resistance, a capacitive reactance, and inductive reactance, or any kind of impedance is included in both cir-

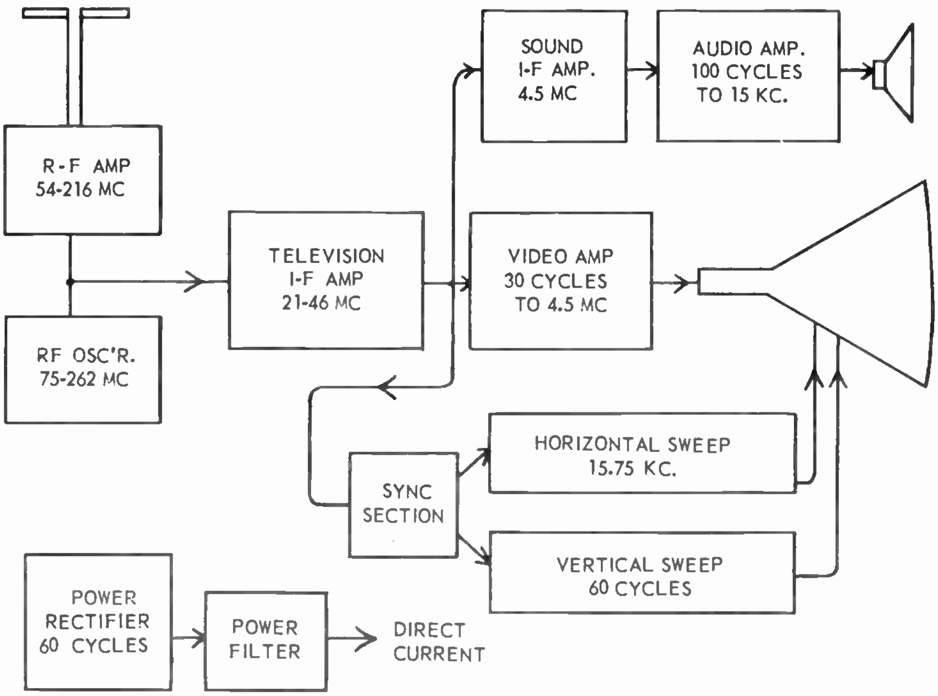


Fig. 76-1. The eight ranges of frequencies that must be kept separated in a television receiver.

cuits, or is common to both circuits. In another general class, called inductive or mutual inductive couplings, we find that a magnetic field produced by one circuit cuts through conductors which are in the other circuit.

In the simple amplifier of Fig. 76-2 there are couplings which are necessary and others which would cause trouble. There is a necessary inductive coupling from transformer primary La to secondary Lb, for inducing signal emf's in the grid circuit of amplifier A. There is a necessary resistance coupling by means of resistor Ro, which is in the plate circuit of amplifier A and in the grid circuit of amplifier B. Signal voltages produced across this resistor by plate current of amplifier A are put into the grid circuit of amplifier B by this resistance coupling.

Now let's look at some of the other couplings. Cathode resistors Rk and Rk are parts of the amplifier grid circuits, because the grid circuits are completed through these resistors, through ground, and winding Lb for amplifier A, or through grid resistor Rg for amplifier B. Resistors Rk carry the alternating plate current for each amplifier and also the alternating screen current for amplifier A. Since resistors Rk are in the grid circuits, plate circuits, and screen circuit, they couple these circuits together at each tube. As we shall learn, this kind of coupling sometimes is desired, but more often it is not wanted.

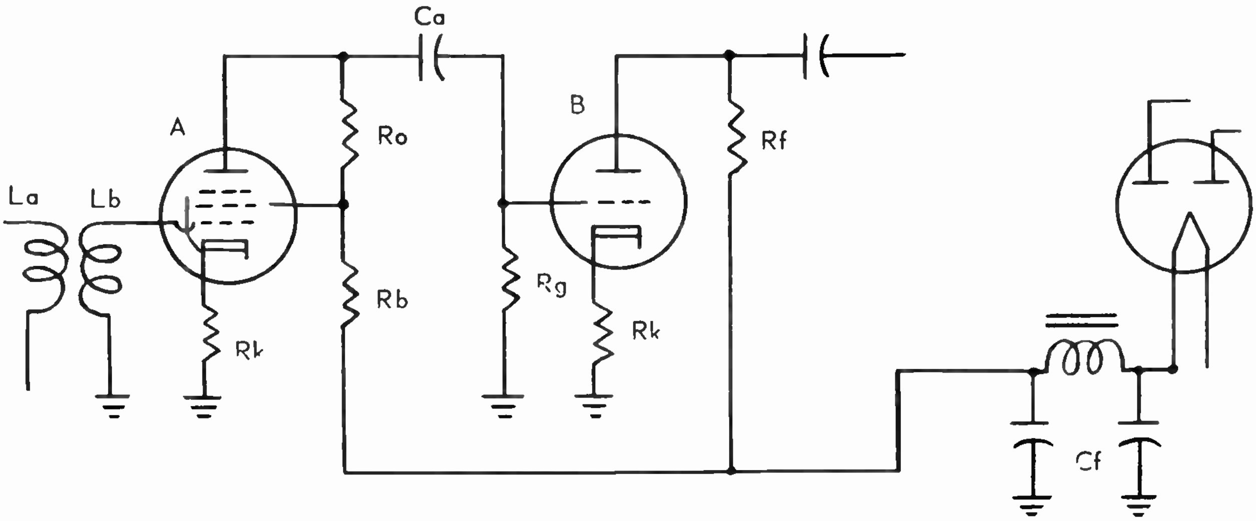


Fig. 76-2. Couplings which are necessary and others which are harmful in an amplifier.

Next, a coupling that never is wanted, the coupling through resistor Rb. Alternating plate voltage of amplifier A flows not only in resistor Ro but also in voltage dropping resistor Rb. Resistor Rb is in the screen circuit of this tube, it carries the screen current. Therefore, resistor Rb couples the plate circuit to the screen of amplifier A, and alternating voltages from the plate circuit will appear on the screen.

There is another unwanted coupling in our amplifier system. It exists in the B power supply. Alternating plate and screen currents of amplifier A can complete their path to the cathode of that tube only by going through one of the power filter capacitors at Cf, or through both filter capacitors and the choke, thence through ground to the cathode. Alternating plate current in amplifier B can return to the cathode of that tube only through the filter of the power supply, and ground. Thus we find that the capacitive reactance of the filter capacitors and the inductive reactance of the filter choke are in the signal circuits of both amplifiers, and provide capacitive and direct inductive couplings from one amplifier to the other.

In order to have effective decoupling we must be able to do two things.

1. Reduce the resistance or the reactance in the path of an alternating current to a value so small as to prevent harmful coupling with other circuits that go through the same path.

2. Allow direct currents or voltages for plates, screens, and grid biasing to pass to or through various circuits, while alternating currents are kept out of the d-c paths in which resistances or reactances might provide coupling.

These methods of preventing couplings require the use of resistors, capacitors, or inductors, either by themselves or in various combinations. The effects of each of these elements, by themselves, on high-frequency currents, low-frequency currents, and direct currents, are illustrated by Fig. 76-3.

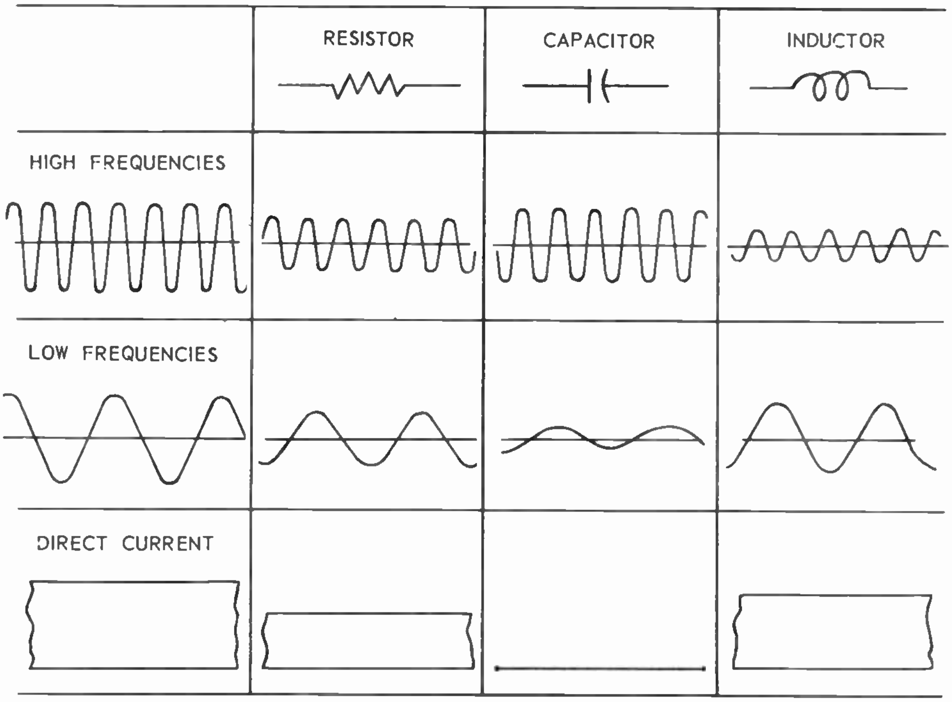


Fig. 76-3. The effects of resistors, capacitors, and inductors on high and low frequencies and on direct currents.

5

8

A resistor reduces all three kinds of currents in the same manner, the amount of reduction depending on the value of resistance. A capacitor of any given capacitance reduces high-frequency currents relatively little, reduces low-frequency currents much more, and completely stops direct currents. An inductor of given inductance reduces high-frequency currents a great deal, reduces low-frequency currents much less, and has no effect on direct currents other than a slight reduction due to ohmic resistance of the inductor. To what extent the alternating currents are reduced by capacitors or inductors depends on the capacitive or inductive reactances at the frequencies involved.

In Fig. 76-1 we have used capacitors for decoupling in the amplifier system that needed it. Capacitor C1 decouples the grid circuit of amplifier A, from the plate and screen circuits of this tube. This capacitor provides a path of low reactance from the low end of the grid circuit directly to the tube cathode, and it completes the grid circuit so far as alternating signal currents are concerned without going through cathode resistor Rk.

Capacitor C2 decouples the plate and screen circuits of amplifier A. This capacitor provides a low reactance path from the screen to ground, thence through resistor Rk to the tube cathode. Alternating voltages in the screen circuit then return to the cathode without having to go through resistor Rb which formerly provided a coupling. Alternating plate current also finds a low reactance path through capacitor C2 and ground to the cathode circuit without going through resistor Rb. If capacitive reactance of C2 is small enough it will provide negligible coupling for plate and screen currents, and the two tube elements will be effectively decoupled.

Capacitor C3 prevents coupling between amplifier A and amplifier B or other tube circuits through the power supply impedances which are common to all the circuits. Alternating plate and screen currents which are not returned to the cathode of amplifier A through capacitor C2 find a low reactance path through C3 to ground and the cathode circuit without having to go through any parts of the power supply.

Capacitor C4 performs the same function in the grid circuit of amplifier B that capacitor C1 performs for amplifier A. Note that C4 is connected below grid resistor Rg. Were this capacitor connected above the resistor it would decouple the grid of amplifier B from the plate of amplifier A and there would be no transfer of signals.

Capacitor C5 prevents coupling the plate circuit of amplifier B to other circuits through power supply impedances, acting for this amplifier stage in the same manner as capacitor C3 for the preceding stage.

Fig. 76-5 shows different positions for decoupling capacitors. Cathode resistors Rk here are paralleled by capacitor C6 on the first amplifier and by capacitor C8 on the second amplifier. These capacitors provide low-reactance paths for all alternating currents around their respective cathode resistors, and thus reduce or prevent all couplings which might occur through these resistors.

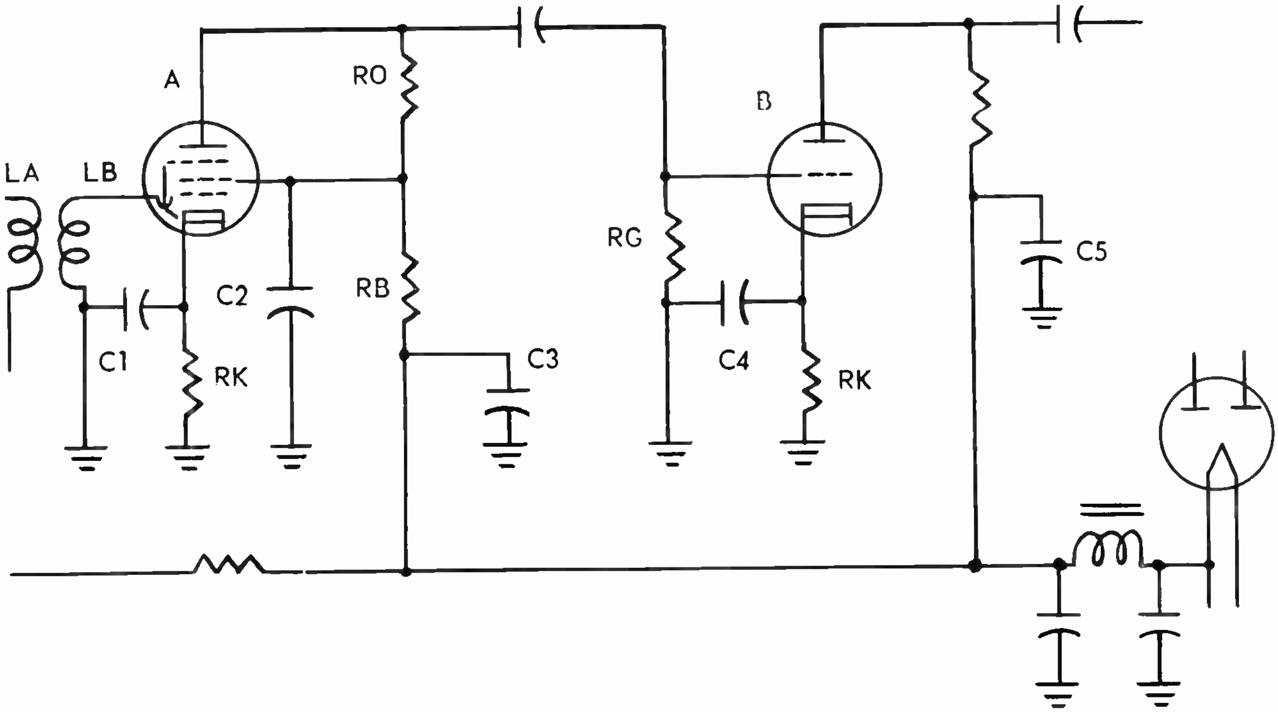


Fig. 76-4. Capacitors used for reduction or prevention of unwanted couplings.

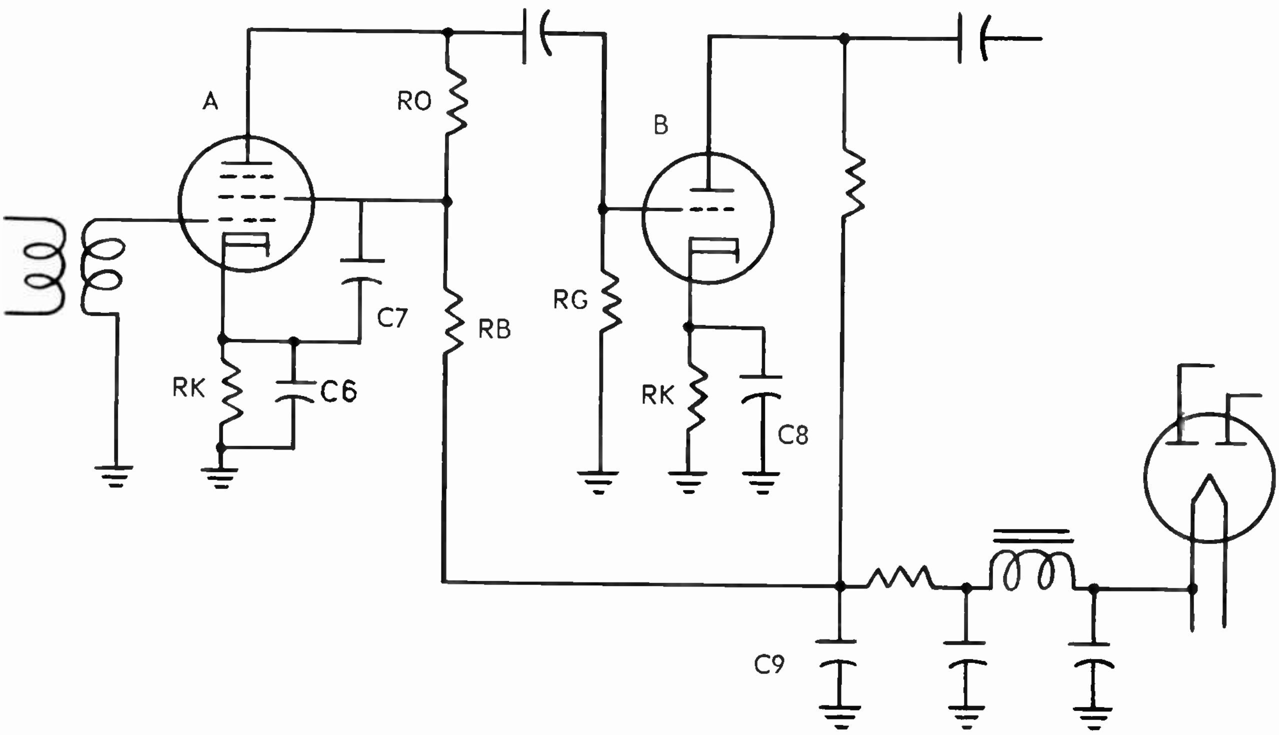


Fig. 76-5. Other connections for the decoupling capacitors in the amplifier.

Capacitor C7 provides a low-reactance path to the cathode of amplifier A from the screen and from the lower end of the plate load resistor. This arrangement quite completely decouples the plate and screen circuits, since their currents no longer have to pass together through cathode resistor Rk in getting to the cathode. It is especially important to decouple the screen of any pentode, because the screen is so close to the cathode inside the tube that any alternating voltages on the screen act almost like alternating voltages on the grid.

Capacitor C9 of Fig. 76-5 is used instead of capacitors C3 and C5 of Fig. 76-4, it is intended to prevent coupling of one amplifier stage to the other through impedance of the power supply. Individual decoupling capacitors, as C3 and C5, connected close to the plate and screen circuits of each stage are more effective than a single capacitor near the power supply.

Although the decoupling capacitors do provide paths of low reactance through which most of the alternating currents should pass, these paths would be of no avail were there another paralleled path having resistance or impedance of even fewer ohms than the reactance of the capacitor. Then most of the alternating currents would take the easier paths and we would have little or no decoupling effect.

Fig. 76-6 shows the resistances or impedances that are in parallel with each of our decoupling capacitors in the two preceding diagrams. The little diagrams in this new figure are merely small parts of the

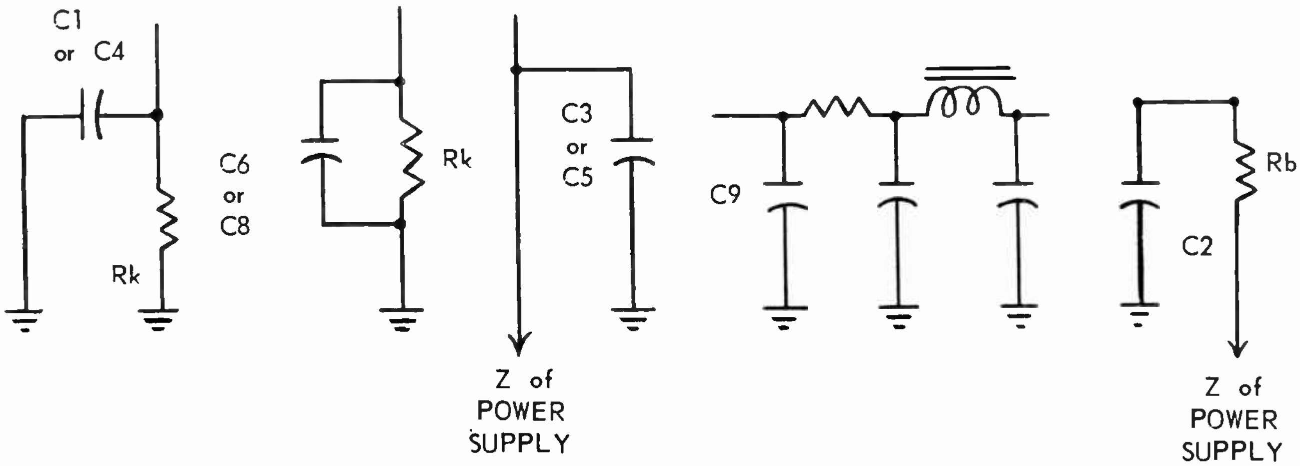


Fig. 76-6. The resistances or impedances in which coupling may exist.

larger diagrams. Capacitors C1, C4, C6, and C8 are paralleled by resistors Rk. In order to have effective decoupling, the reactances of the capacitors must have much less opposition to current flow than the resistances which are paralleled.

Capacitors C3, C5, and C9 parallel the impedance of the power supply, and their reactances in ohms must be much less than the power supply impedance in ohms if we are to have effective decoupling. Capacitor C2 is in parallel with the combined resistance of resistor Rb and impedance of the power supply. Resistor Rb may be called a decoupling resistor, because it acts in conjunction with capacitor C2 to cause decoupling.

7 The fraction of total alternating current that continues to flow through the parallel resistance or impedance, rather than through the cpaacitor, depends on the ratio of capacitive reactance to resistance or impedance. As an example, if the number of ohms reactance of the capacitor is 1/5 of the number of ohms of resistance or impedance, then 1/5 of the total alternating current will go through the resistance or impedance and 4/5 through the capacitor. With reduction of capacitive reactance to 1/10 of the resistance or impedance, 1/10 of the current would go through the resistance or impedance and 9/10 through the capacitor.

For reasonably effective decoupling it is considered that no more than 1/5 to 1/10 of the total alternating current should continue on through the resistance or impedance wherein coupling might occur. Capacitive reactance is computed for the lowest frequency at which there is to be satisfactory decoupling.

At 1 in Fig. 76-7 is represented an audio amplifier whose lowest operating frequency is to be 100 cycles. There is a 1,500-ohm cathode bias resistor with an 0.1 mf decoupling capacitor. The reactance of this capacitance at 100 cycles is about 15,900 ohms; it is so much greater than the resistance that we have practically no decoupling at all. It would take 10 mf to do this job on a 10 to 1 basis.

At 2 we have the same resistor and the same decoupling capacitor, but the lowest frequency at which we want good decoupling in this intermediate-frequency amplifier for an f-m broadcast set is 10 megacycles. The capacitive reactance at 10 mc is only about 0.16 ohm, so the capacitor is far larger than needed. On a 10 to 1 basis we would need a decoupling capacitor of about 0.0001 mf.

At 3 we have a television sound i-f amplifier in which there is to be decoupling on a 10 to 1 basis at 4.5 mc. The cathode resistor is 500 ohms and the decoupling capacitor is 0.00068 mf or 680 mmf. Capacitive reactance is about 52 ohms.

Apparently it takes a lot of figuring with reactance formulas to determine the required decoupling capacitor for a job, but with the help of the alignment chart in Fig. 76-8 almost any practical problem can be solved in less than a minute. There are three scales. The one at the left covers capacitive reactances from 0.5 ohm to 30,000 ohms. The one in the center covers frequencies from 30 cycles per second to 300 megacycles. The right-hand scale is for capacitances from 0.0001 mf, or 100 mmf, all the way to 200 mf.

The chart is used by laying a ruler or any other straight edge across the three scales, with the edge exactly on whatever two values you know. Then the corresponding unknown value may be read where the

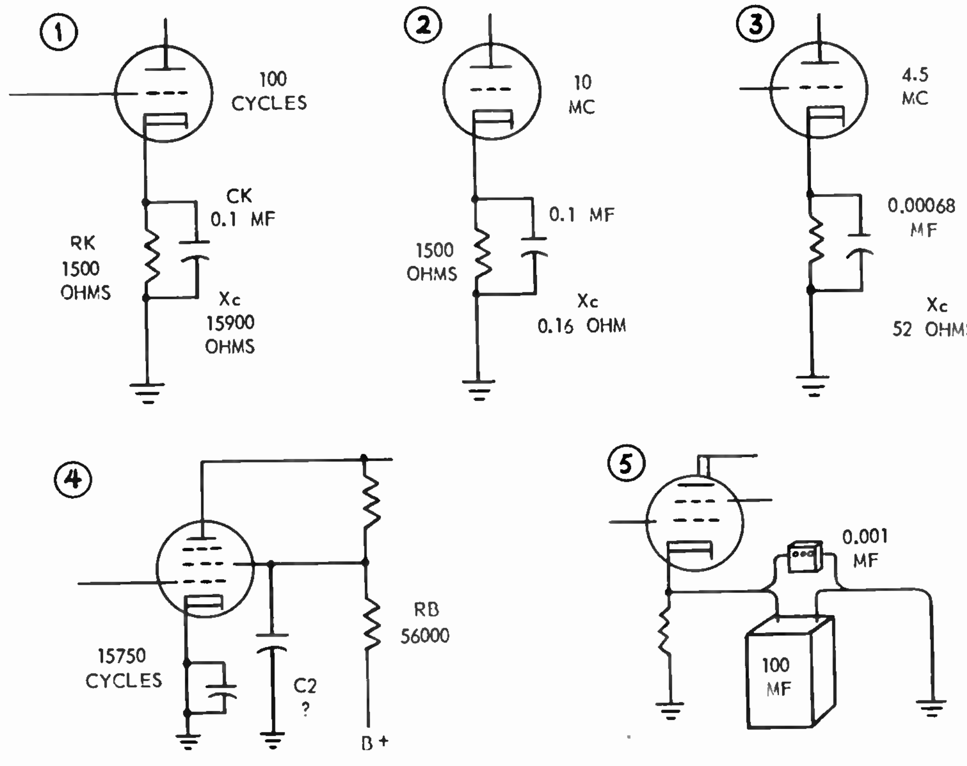


Fig. 76-7. Relations between reactance and resistance in some decoupling circuits.

straight edge crosses the third scale. If you know reactance and frequency, you can read the capacitance which will have this reactance. Knowing frequency and capacitance, you can read the resulting reactance. Knowing reactance and capacitance, you can read the frequency.

Readings taken from an alignment chart are not so accurate as those derived from a formula, but they are amply accurate for all service problems. Also, you are far less likely to make serious errors with the chart than when working a formula.

Let's consider this problem. At 4 in Fig. 76-7 is shown amplifier A from Fig. 76-4. We want capacitor C2 to provide 10 to 1 decoupling at a frequency of 15,750 cycles. The resistance involved is 56,000 ohms at Rb, plus whatever resistance or reactance exists in the common B power supply. We shall assume a well designed power supply, with negligible impedance compared with the resistance at Rb.

Capacitive reactance of C2 must not exceed 1/10 the resistance of Rb, or must not exceed 5,600 ohms. Lay the straight edge on the left-hand scale of Fig. 76-8 at what you estimate to be 5,600 ohms, and on the center scale at what you estimate to be 15,750 cycles (or 15.75 kilocycles). The straight edge crosses the right-hand scale at a little below 0.002 mf, so a capacitor of 0.002 mf will do the job, with something to spare. The reactance of 0.002 mf at 15.75 kc is about 5,053 ohms.

When computing decoupling capacitors for plate and screen circuits you will find that the higher the decoupling resistance the less capacitance is needed for effective decoupling. But the greater the decoupling resistance the higher must be the B voltage from the power supply to overcome the increased drop and leave enough at the plate and screen.

Doubtless you have realized before this that the decoupling capacitors now under discussion are the same identical units which we called bypass capacitors in other lessons. These capacitors do bypass alternating currents around resistors and impedances, so "bypass" is a correct name. But the purpose of the bypassing is to provide decoupling, so "decoupling capacitor" is another correct name.

You will hear it said that bypass or decoupling capacitors for any frequency above the audio range should be mica, ceramic, or non-inductive paper types. This is true, but there are some circuits where the decoupling capacitance must be so great as to be unavailable in anything except an electrolytic capacitor, and where there are frequencies higher than the audio range. The objection to the electrolytic, and also to most paper capacitors, is that they possess considerable inductance, and offer much inductive reactance which would provide coupling for the high frequencies.

To get around this problem we connect in parallel with the electrolytic unit a small capacitance mica or ceramic capacitor, as at 5 in Fig. 76-7. The capacitance of the mica or ceramic capacitor is only large enough for decoupling at the higher frequencies in the circuit, or for decoupling the inductive reactance of the large capacitor.

R-F CHOKES FOR DECOUPLING. Decoupling resistors used in connection with decoupling capacitors are satisfactory where the currents are small or where the resistances may be small. For large currents and high resistances the voltage drops in decoupling resistors would be excessive, and too much power in watts would be changed into heat. Then we resort to the use of r-f chokes, which may have high inductive reactance at radio frequencies and yet may have very small resistance and proportionately small voltage drop and heating.

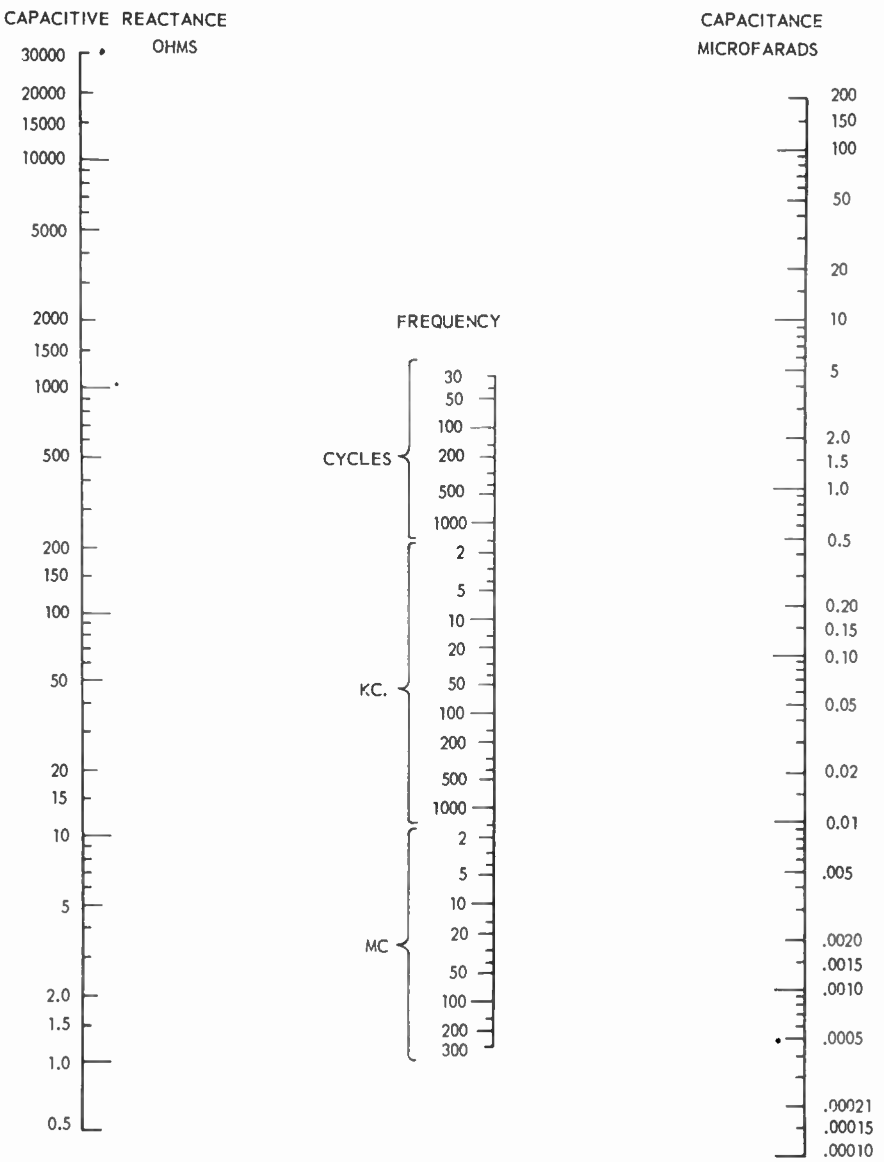


Fig. 76-8. A chart for determining the related values of capacitive reactance, frequency, and capacitance

Fig. 76-9 is a picture of a number of r-f chokes. All of these are of "air-core" construction, which means only that they have no iron in their cores. Some chokes which provide large inductances do have powdered iron cores. Very few r-f chokes have inductances greater than 10 millihenrys or 10,000 microhenrys, while the smallest inductances commonly used are on the order of 0.5 microhenry.

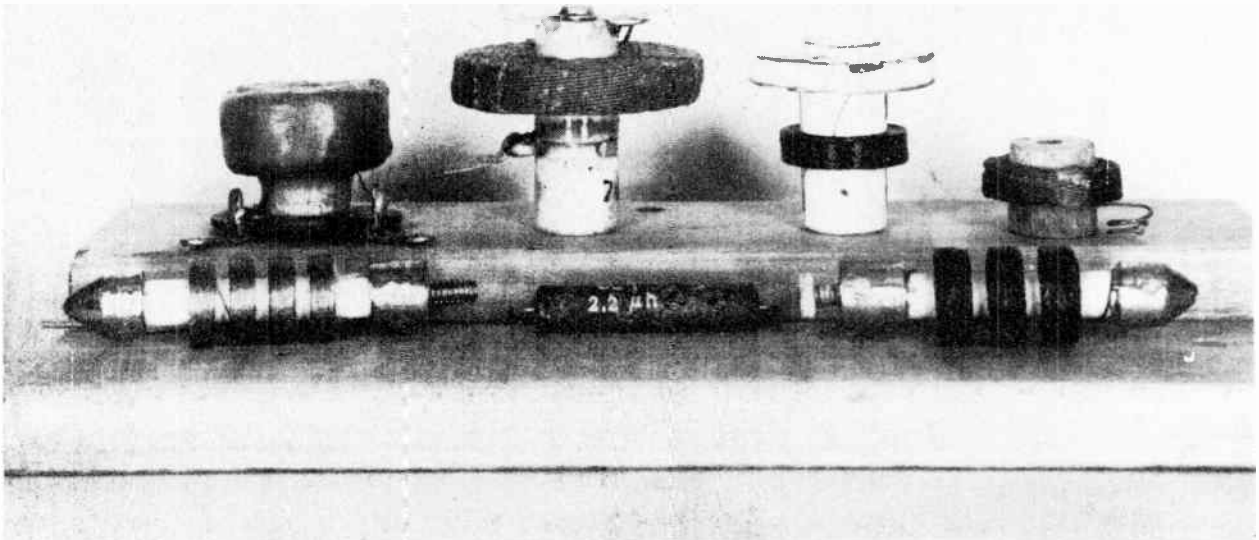


Fig. 76-9. Various styles of radio-frequency chokes.

An r-f choke must have the least practicable distributed capacitance, not because the choke may be self-resonant at some frequency within the normal operating range, but because this capacitance will allow high-frequency currents to pass through it without being opposed by the inductive reactance. A choke always is connected in series with the path wherein high-frequency currents are to be opposed, and when so connected the distributed capacitance is in parallel with the inductance to form an effective bypass around the inductance.

Distributed capacitance is lessened by using duolateral or honeycomb windings for the choke coils when there are a great number of turns. Side by side pie windings with spaces between them are used for the same reason. Chokes made with only a few turns of wire should be space wound to lessen the distributed capacitance.

A decoupling choke may be used in a grid circuit, a plate-screen circuit, or in both places as shown at 1 in Fig. 76-10. The inductive reactance of choke La opposes flow of alternating or pulsating grid currents to ground and forces them to return to the cathode through the low capacitive reactance of capacitor Ca.

The large inductive reactance of choke Lb prevents alternating plate and screen currents of the tube from going to the B power supply lines, and forces these currents to return to the tube cathode through the small capacitive reactance of capacitor Cb. At the same time this choke keeps alternating currents which may be components of the direct B plus current from getting into the plate and screen circuits of the tube.

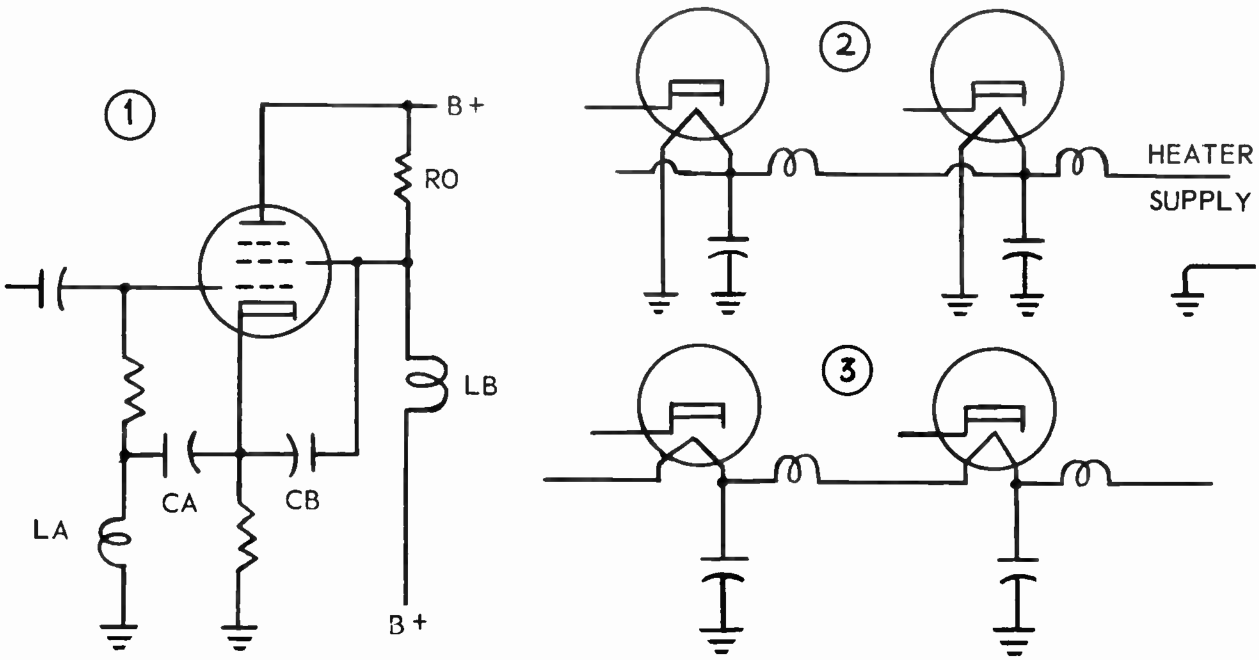


Fig. 76-10. R-f chokes employed for decoupling.

6 There is another advantage in using a choke rather than a resistor for decoupling. Every variation of plate current in a resistor causes a variation of voltage at the tube plate, because the drop of voltage in a resistor changes with every change of current. The ohmic resistance of a choke is negligible, and voltage drop across the choke remains practically constant for all normal plate current changes. Therefore, we have much better voltage regulation in the plate circuit, and energy of the alternating plate current or signal current is used almost entirely for producing alternating or signal voltages in the regular load resistor, Ro of Fig. 76-10.

If chokes are used in the grid circuit and also in the plate circuit of the same tube, the two chokes should not be of the same type or construction. Should two similar chokes be self-resonant at any operating frequency there would be a tuned parallel resonant circuit (choke inductance and distributed capacitance) in both circuits. This would form a "tuned-plate tuned-grid" oscillator, and the tube undoubtedly would oscillate at the self-resonant frequency. When chokes are mounted near each other their axes should be at right angles or otherwise positioned for least inductive coupling between them. Otherwise all the chokes, or all but one of them, should be shielded.

R-f chokes may be used for decoupling of heaters or filaments in the several stages of the r-f and i-f amplifier sections of television receivers, as shown at 2 and 3 of Fig. 76-10. Unless this is done it is quite possible for the common heater or filament supply lines to carry high-frequency voltages between the stages. Diagram 2 applies to parallel heaters. There is an r-f choke between each heater and those on either side. There is a bypass or decoupling capacitor from each heater to ground. Diagram 3 shows the equivalent arrangement for series heaters. Sometimes the capacitors are omitted and only the chokes are used, while in other cases there may be quite elaborate arrangements of parallel and series resonant circuits to oppose certain frequencies while passing other frequencies.

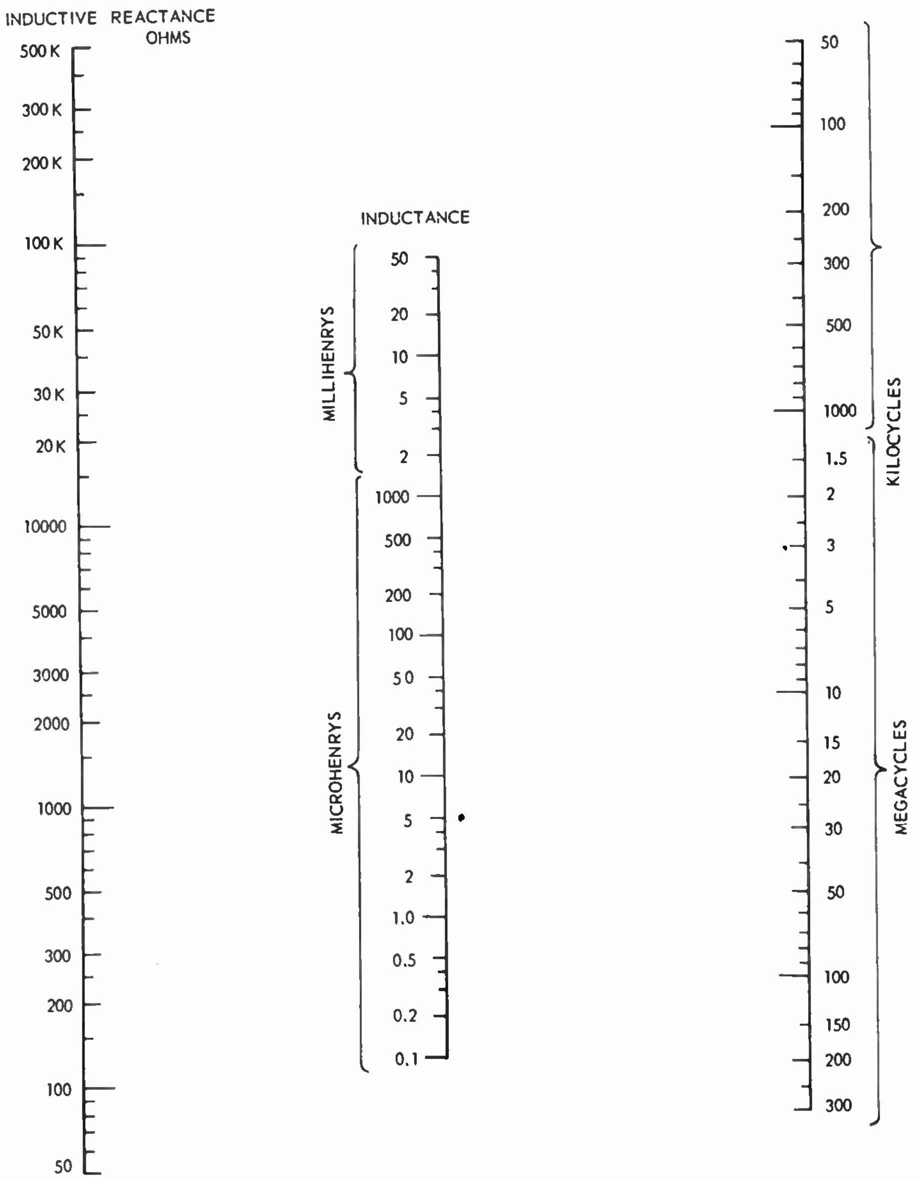


Fig. 76-11. A chart for determining related values of inductive reactance, inductance and frequency.

Heater or filament chokes most often consist of 15 to 20 turns of enameled wire wound to a diameter of about 1/4 inch. The wire commonly is of 20 to 24 gage size. The inductance of such a choke is about one microhenry.

Fig. 76-11 is an alignment chart showing relations between inductive reactance (left-hand scale), inductance (center scale), and frequency (right-hand scale). This chart, like the one of Fig. 76-8, is used by laying a straight edge on any two known values and reading the third unknown value where the straight edge crosses the third scale. The chart is designed especially for solving problems relating to r-f chokes. Maximum inductance is 50 millihenrys (50,000 microhenrys) and minimum frequency is 50 kilocycles.

The range of this chart may be extended for larger inductances and lower frequencies in either of two ways. First: If you read the megacycle scale as cycles and the microhenry scale as henrys, the reactances may be read without change. Second: If you read the kilocycle scale as cycles and the millihenry scale as henrys, the reactances again may be read without change. For instance, the reactance in ohms of 100 henrys inductance at 50 cycles is the same as for 100 microhenrys at 50 megacycles.

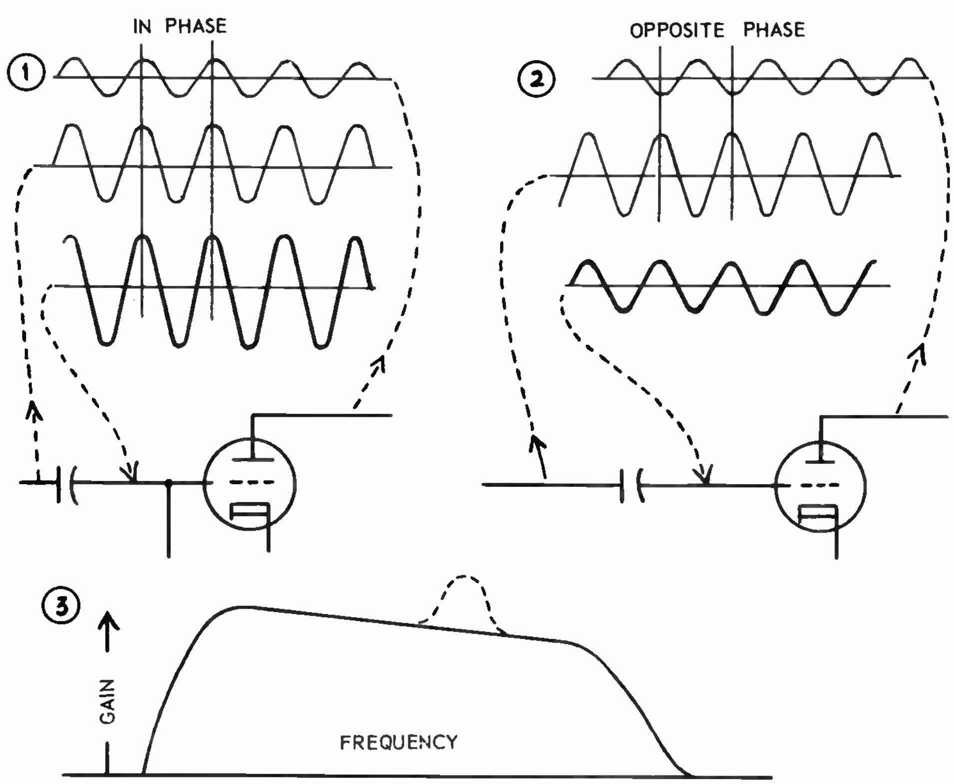


Fig. 76-12. Effects on grid voltage of feedbacks that are in phase or of opposite phase.

To see how decoupling is affected by using a choke, assume that capacitor Cb at 1 of Fig. 76-10 is of 0.001 mf capacitance, that choke Lb is of 1 millihenry inductance, and that the operating frequency is 5 megacycles. By using Fig. 76-8 we learn that the capacitive reactance of Cb is about 32 ohms or maybe a little less. By using Fig. 76-11 we find that the inductive reactance is slightly more than 31,000 ohms in choke Lb. Here we have a decoupling ratio of approximately 1,000 to 1. From the inductance chart you can learn that a choke for a decoupling ratio of 10 to 1 (about 320 ohms inductive reactance) would require only about 10 microhenrys inductance.

4 The effectiveness of decoupling when using either a resistor or a choke usually is checked by using an a-c vacuum tube voltmeter. With the voltmeter connected across the decoupling capacitor, the strongest possible signal is applied to the grid of the tube. If the meter reading remains at zero on a low voltage scale there is good decoupling, because the capacitive reactance of the capacitor is too small to give a measureble voltage drop with the existing alternating current. The next step is to measure the a-c voltage across the decoupling choke or across a resistor used primarily for decoupling. Here there should be little or no a-c voltage if the decoupling is good, because this path then should carry negligible alternating current.

FEEDBACK AND OSCILLATION. Decoupling is employed to prevent undesired alternating voltages of any kind from getting into an amplifier stage or any other stage where these voltages don't belong. Decoupling helps prevent signal voltages from passing between stages other than through intentional couplings, and it may be used to prevent signal voltages passing from the plate to the grid of the same tube when this is not desired.

If voltage coupled back or fed back from the plate circuit is in phase with voltage in the grid circuit, as at 1 in Fig. 76-12, the alternating voltage on the grid will be reinforced. The strengthened grid voltage is amplified in the tube, there is stronger alternating voltage in the plate circuit, and an increased feedback. This action is desirable in an oscillator, but not in an amplifier. Within the smallest fraction of a second the amplifier would become an oscillator, and any incoming signal would lose all control. Oscillation would continue at a frequency determined only by the values of inductance and capacitance in the tube circuits.

2 Any in-phase feedback that reinforces voltages normally on the grid of a tube is called a regenerative
3 feedback. Regeneration is defined as the process in which part of the a-c power from the output circuit of an amplifier acts on the input circuit to increase the amplification. Regeneration easily changes to oscillation. While you are aligning a television i-f amplifier the picture tube screen suddenly may become intensely bright. That means you have tuned some circuit to a frequency where feedback causes oscillation. The only way to stop the oscillation, and save the picture tube, is to turn off the power in a hurry.

If feedback voltage from the output of an amplifier is in phase opposition to the grid voltage, as at 2 in Fig. 76-12, the alternating grid voltage is weakened. This is the process called degeneration. As we shall learn later, degeneration is useful in some kinds of amplifiers.

When there is an in-phase or regenerative feedback too weak to cause oscillation it still may cause excessive amplification at one or more frequencies for which the tube circuits are resonant or only slightly off resonance. This action is called "peaking." Instead of the frequency response of an amplifier remaining fairly constant over a range of frequencies, as shown by the full line curve of diagram 3, there may be a sharp rise at a certain frequency, as shown by the broken line. One symptom of peaking is a television picture or pattern covered with small black horizontal streaks or specks a fraction of an inch long.

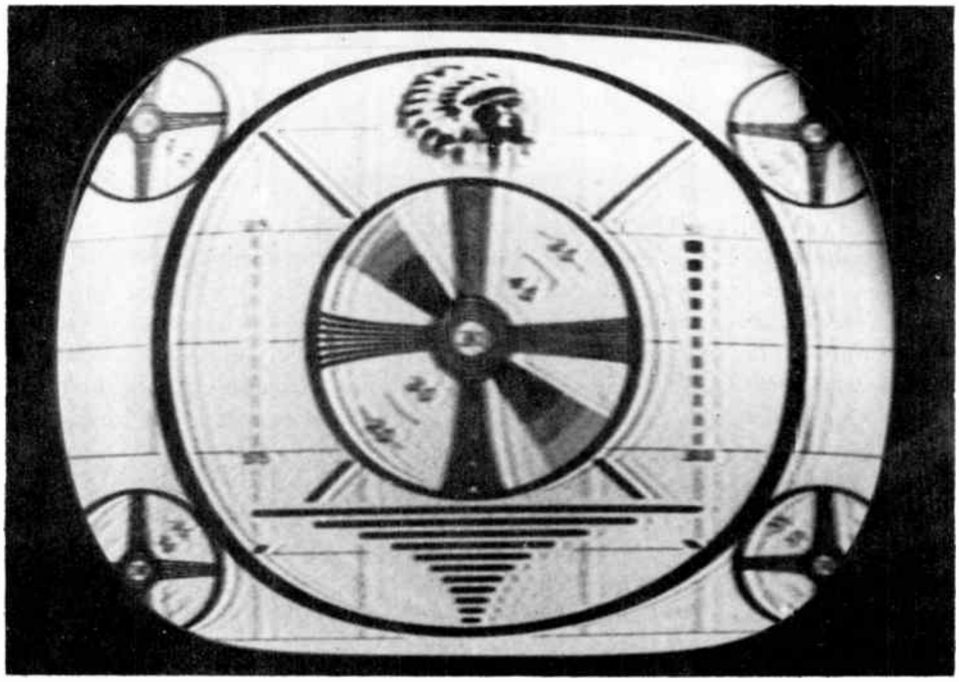


Fig. 76-13. Multiple images due to short time oscillations.

Oscillation may occur in circuits which do not contain a tube and which are not intentionally tuned to any frequency, but which have inductances that act with distributed and stray capacitances to make the circuit self-resonant at some frequency. When a fairly strong emf is induced in such a circuit there may be oscillation. The oscillation will decrease in strength and die out unless there is repeated induction of emf from some external source. The frequency of the inducing field need not be that for which the circuit is self-resonant, it may be of almost any frequency or there may be pulses at irregular intervals. This would be called shock excitation.

2 Oscillation which quickly dies away, but then recurrs, may be called "ringing". It may exist in video amplifier couplings and in other video amplifier inductors. When such trouble occurs in a video amplifier, the effect as seen on a picture tube is a repetition of that portion of the picture or pattern normally reproduced by a video frequency equal to that of resonance. There may be several images at about equal spacings and of decreasing intensity, somewhat as pictured by Fig. 76-13.

A kind of self-resonance different from anything heretofore considered may cause what is called parasitic oscillation at frequencies in the television carrier range. At these very-high frequencies it is possible for the reactance of tuning inductors to become so great that they act like r-f chokes or almost as open circuits. The reactance of tuning capacitors becomes so small that the capacitors act as short circuits. Then resonance at the very-high frequency is due to the inductance of wires leading to tube elements and to stray capacitance in wiring and parts.

The part which supports the parasitic oscillation may be located with a heterodyne frequency meter and a grid dip meter. First, turn off the tube which follows the suspected stage by disconnecting its heater or filament, but leave the tube in its socket. Turn on the set and use the heterodyne frequency meter to determine the frequency of oscillation. Then turn off the set and couple the grid dip meter successively to all tuning coils, peaking coils, small chokes, capacitors, resistors, and wiring connections while tuning the meter slowly back and forth across the frequency of oscillation.

Parasitics usually may be stopped by connecting a fixed resistor of 50 to 100 ohms in series with the tube grid, directly at the grid lug of the socket. An r-f choke of a few microhenrys inductance may be used instead of a resistor.

DEGENERATION. In early radio receivers the tubes were so few, and their mutual conductances so low, that anything which would increase the amplification was hailed with delight. In nearly all detectors and even in some r-f amplifiers, regeneration was pushed up to a point just short of oscillation in an effort to reinforce the weak signals. Nowadays we seldom employ regeneration intentionally, but degeneration or negative feedback helps solve many problems.

Degeneration is used in practically every audio amplifier of the high-fidelity class. It is used in video amplifiers, in r-f and i-f amplifiers where there is low-frequency modulation, in automatic frequency control systems, and in measuring instruments whose indications must remain nearly independent of changes in the characteristics of tubes and other circuit elements.

Degeneration, correctly used, will greatly extend the range of frequencies in which an amplifier has satisfactorily uniform response or gain, or it will equalize the gain at low frequencies, or at high frequencies, or at both ends of the range. Degeneration will reduce the strength of harmonic frequencies, help reduce phase shift between input and output of an amplifier, and it may be used to lesson the noise originating in tubes and resistors. Furthermore, degeneration lessens the effect of line voltage variation on amplification, and it may lessen the bad effect of changes in load impedance on fidelity of sound reproduction.

The only objection to degeneration is the reduction of amplification below that which could be obtained without a negative feedback. But the loss of gain may be easily compensated for by tubes of high transconductance or by an added stage of amplification, while the advantages are retained. The many applications of degeneration will be explained as the need arises. Just now we shall examine some of the elementary principles.

One method of obtaining a degenerative feedback utilizes the cathode biasing resistor <u>Rk</u> as at <u>1</u> in Fig. 76-11. This resistor is part of the plate circuit impedance through which a-c signal currents must flow, and it is also part of the grid circuit. Total impedance of the plate circuit is made up of the plate resistance in the tube, the load resistance at <u>Ro</u>, the capacitive reactance of the decoupling capacitor at <u>Xc</u>, and the resistance of cathode resistor <u>Rk</u>. We are assuming that reactance at <u>Xc</u> is small compared to impedance through the B power supply. A-c plate voltage which appears across resistor <u>Rk</u> is applied to the grid of the tube, because <u>Rk</u> is between the grid and the cathode.

As we have learned before, a-c voltages in a plate circuit are in opposite phase to a-c voltages on the grid and cathode resistor of the same tube. The a-c voltage across resistor <u>Rk</u> is in phase with the grid voltage since a positive signal at the grid causes more electrons to flow through resistor <u>Rk</u> and the tube,

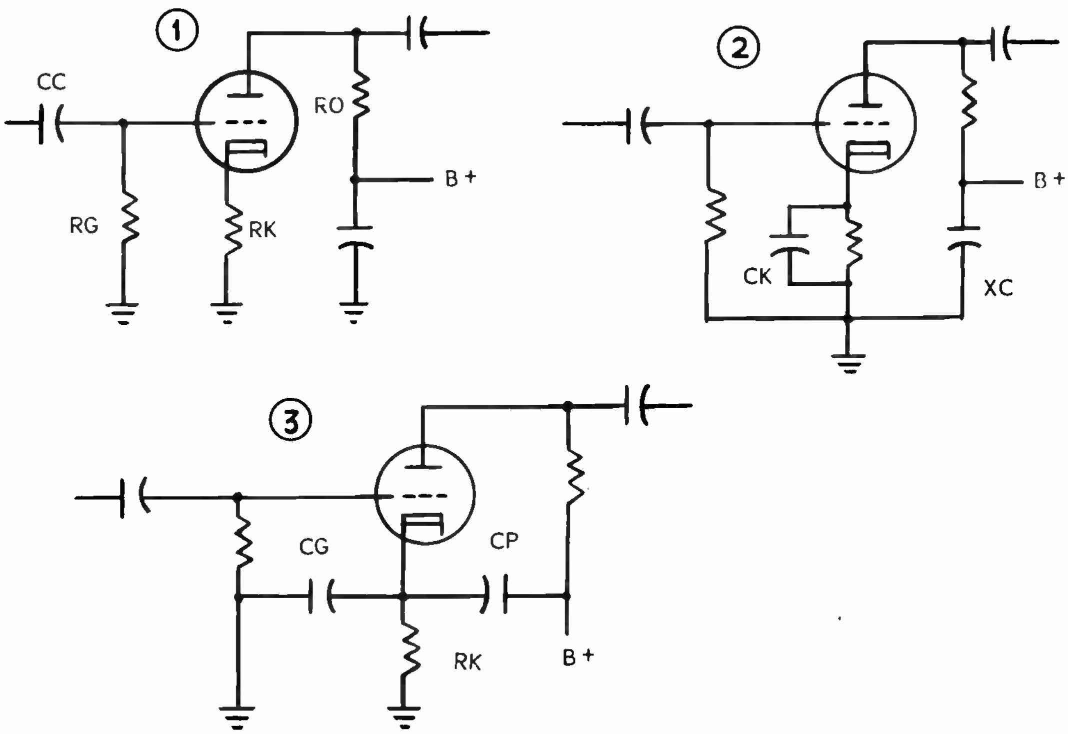


Fig. 76-14. Degeneration may be secured from feedback voltages in a cathode resistor.

resulting in a higher voltage drop across resistor Rk. With a higher or more positive voltage across resistor Rk, the bias applied to the grid of the tube becomes more negative, therefore this voltage acts as a degenerative feedback voltage. Actually the effective signal voltage is the difference between the input signal at the grid and the voltage developed across resistor Rk by that signal. It is the reduced grid voltage that now is amplified by the tube. Consequently, the signal output from the plate circuit will be lower than without the negative feedback. To make the output equal to that without the feedback we may either apply more signal through coupling capacitor Cc, use a tube of higher transconductance, or operate the present tube in a way which increases its transconductance.

The greater the resistance at Rk of diagram 1 the stronger will be the feedback voltage and the greater the degeneration, while less resistance will reduce the degeneration. This is all very well, but we cannot change the cathode resistor and thereby change the grid bias in order to obtain a desired amount of feedback. If, however, we place a capacitor across the bias resistor, as in diagram 2, the feedback and the grid bias become quite independent of each other. Grid bias still is determined by the direct current flowing in the resistor, and is unaffected by the capacitor because the capacitor carries only alternating signal current. The impedance to alternating signal current now is the impedance of the paralleled capacitance and resistance, and may be decreased as much as we like by using greater capacitance having less reactance. It is this impedance that determines the amount of feedback.

With no capacitor across the cathode resistor there will be maximum feedback and degeneration. With capacitance so great as to have negligible reactance compared to the resistance of Rk there will be negligible feedback, because the a-c voltage drop across such a small reactance will be very small.

A capacitor connected as at Cg in diagram 3 of Fig. 76-14 has exactly the same effect as one at Ck of diagram 2, because either of these capacitors is connected from cathode to ground, and across resistor Rk. A capacitor connected as at Cp of diagram 3 will keep most of the alternating plate current out of resistor Rk, if the capacitor is large enough, and will reduce the feedback while still decoupling the plate from the B supply circuits.

An entirely different method of obtaining a degenerative feedback is illustrated by Fig. 76-15. Alternating voltage from the plate is taken through the blocking capacitor Cb and the feedback resistor Rf directly to the grid of the same tube. Since alternating or signal voltage at a plate always is opposite in phase to alternating voltage at the grid the feedback here is negative or degenerative. If the reactance of capacitor Cb is small at all operating frequencies the amount of feedback is determined almost wholly by resistance at Rf, the less the resistance the stronger is the feedback.

If capacitance at Cb is small enough to provide relatively great reactance at low frequencies this increase of reactance lessens the degeneration at low frequencies, allowing the low-frequency response of the amplifier to be stronger than the response at the high-frequency end of the range.

Fig. 76-16 shows methods of obtaining the same type of feedback with somewhat different circuit arrangements. Referring to diagram 1, alternating feedback voltage from the plate of amplifier B goes to the grid of this same amplifier through feedback resistor Rf and interstage coupling capacitor Cc. Capacitor Cc acts as a blocking capacitor for d-c plate voltage, and takes the place of blocking capacitor Cb of Fig. 76-15.

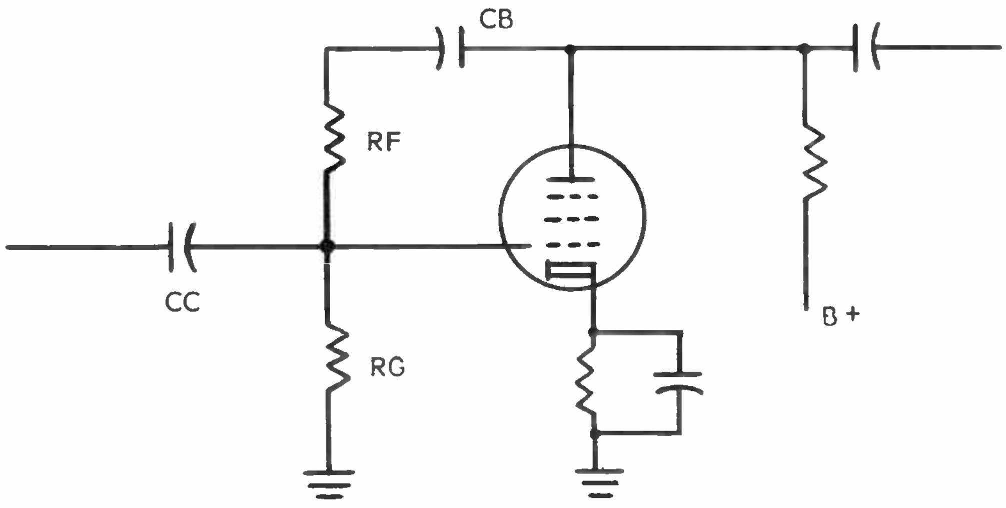


Fig. 76-15. A degenerative feedback connection from plate to grid.

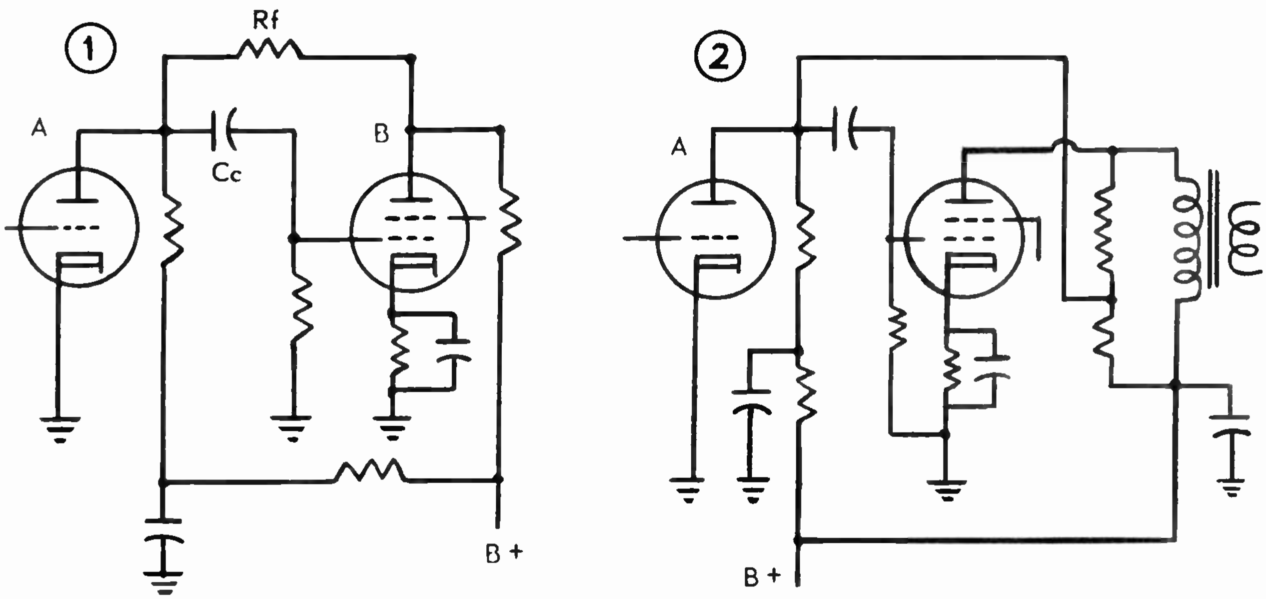


Fig. 76-16. Degenerative feedbacks from one plate circuit to a preceding plate connection.

There is no conflict of phase relations for the following reasons. Feedback voltage is in phase opposition to grid voltage on amplifier B, as it should be. Alternating plate voltage from amplifier A is practically in phase with grid voltage on B, so must be in phase opposition to plate voltage of B. Thus the feedback voltage here introduced ahead of capacitor Cc has the same degenerative phase relation to grid voltage on amplifier B as though the feedback went directly to the grid of B. If there is a difference between the plate voltages at the two tubes, this potential difference is across resistor Rf and this resistor carries direct current proportional to its resistance and the potential difference.

In diagram 2 of Fig. 76-16 feedback voltage again is applied ahead of coupling and blocking capacitor Cc, but here is taken from a resistance voltage divider connected across the primary of an output transformer in the plate circuit of amplifier B. Relative resistances in the two parts of the voltage divider determine what fraction of the alternating plate voltage from amplifier B is used for feedback. Instead of the fixed resistance voltage divider it is possible to use a potentiometer with the feedback line connected to the slider. This provides an adjustable feedback for allowing any desired amount of degeneration.

Fig. 76-17 illustrates some methods of taking degenerative feedback voltage from the plate circuit of one amplifier and applying it to the cathode of the preceding amplifier. Alternating voltage or signal voltage at the cathode of an amplifier is in phase opposition with voltage at the plate of the same amplifier. Voltage at the plate of amplifier A is practically in phase with voltage at the grid of amplifier B. Therefore, if a feedback voltage applied to the cathode of amplifier A is in phase to alternating voltage on the grid of amplifier B, this feedback applied to the cathode still will be out of phase or degenerative as it reaches the grid of amplifier B.

COMMERCIAL TRADES INSTITUTE

TELEVISION

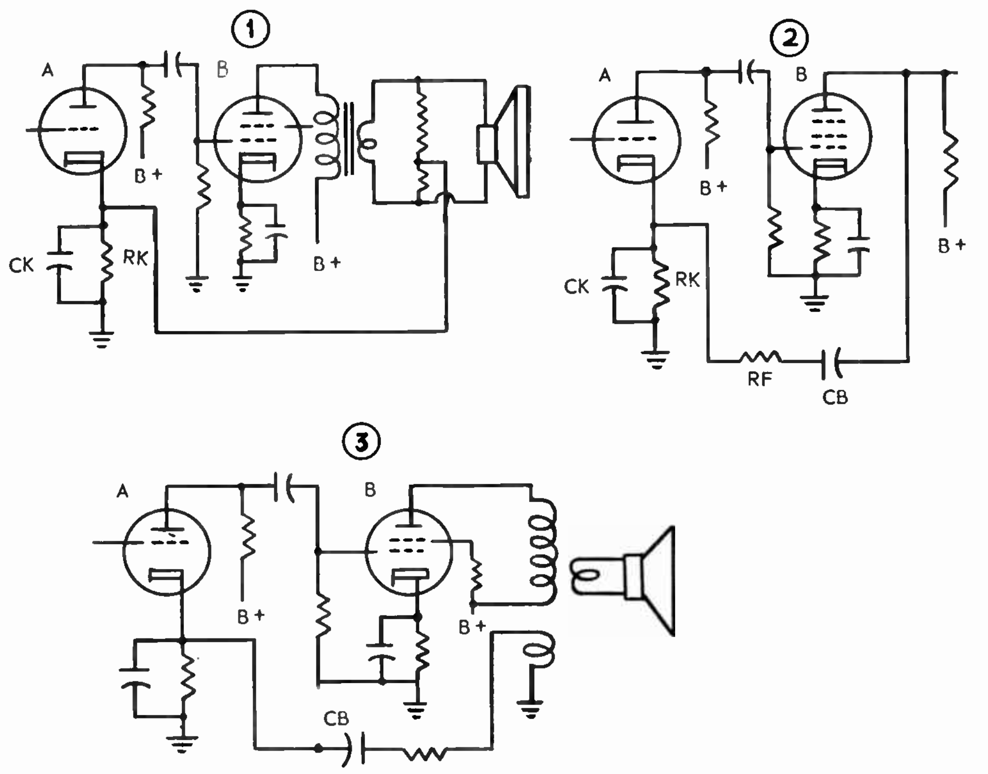


Fig. 76-17. Degenerative feedbacks to the cathodes of amplifier tubes.

In diagram 1 the feedback voltage is obtained from a voltage divider across the voice coil winding of the output transformer connected to the plate of amplifier B. If a connection of the voltage divider to the transformer turned out to be regenerative it would be necessary only to reverse the divider connections to make it degenerative.

In diagram 2 the feedback voltage is taken directly from the plate of amplifier B and is applied through blocking capacitor Cb and resistor Rf to the cathode of amplifier A. So far as phase relations are concerned this is equivalent to applying a feedback from the plate of B to the grid of the same tube, which makes the feedback degenerative. If capacitance at Cb is small enough to have considerable reactance at low frequencies it will cause an increase of gain at these frequencies, just as explained in connection with Fig. 76-15.

The high-frequency response or gain may be increased by suitable choice of capacitance for cathode bypass capacitor Ck. If capacitance at this point is such as to have fairly low reactance at the high-frequency end of the range it will bypass some of the feedback voltage at high frequencies. Reducing the degenerative feedback in this manner allows the gain to rise at the high frequencies. This method of raising the high-frequency gain is equally effective with any of the circuits shown by Fig. 76-17, for all of them have cathode bias and a cathode bypass capacitor on the tube to which the degenerative feedback is applied.

Diagram 3 of Fig. 76-17 shows a third winding or "tertiary" winding on the speaker output transformer. From this winding is obtained a degenerative feedback voltage applied to the cathode of amplifier A. Some speaker coupling transformers of this type provide a feedback voltage equal to 10 per cent of the plate signal voltage of the output amplifier. Others provide a 15 per cent feedback voltage. Capacitance at Cb may be made of such value as to lessen degeneration and raise the gain at low frequencies.

In our discussion of degeneration it has been assumed that feedback voltage is 180° out of phase or is in phase opposition to grid signal voltage, as shown by diagram 2 of Fig. 76-12. This will be true at the middle frequencies of a range but, as we learned when studying broad band amplifiers, there will be phase shifts which bring the grid voltage more nearly into phase with the feedback at both the lowest and highest frequencies.

With feedback directly from plate to grid of the same tube, as in Fig. 76-15, the phase angle between feedback and grid voltages will become less than 180°, but it never will be less than 90° whereat there would be zero degeneration. With other methods which take the feedback from one stage to a preceding stage the phase angle may become even less than 90°. Then the feedback changes from negative to positive and there will be regeneration and probable oscillation unless the feedback voltage is held to a low value.

Phase shift, and loss of the degenerative effect at lowest and highest frequencies, is lessened by all the practices that extend the frequency range of broad band amplifiers. Chief among these practices are the use of fairly small load resistances in amplifier plate circuits, large capacitances for interstage coupling, and highly effective decoupling of the plate and screen circuits.

OR GRADING.

TELEVISION

LESSON NO. 77

SHIELDING, GROUNDING, AND DRESSING

"Training Your Mind"

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind

C.T.I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON NO. 77

SHIELDING, GROUNDING, AND DRESSING

The decoupling methods which we have examined are effective in reducing or eliminating feedback couplings which would be due to resistances or impedances otherwise included in two or more circuits. They do not prevent couplings which are due to magnetic or electric fields which arise in one circuit and pass through other circuits. To prevent this latter kind of couplings we must place the parts or circuits in such relative positions that the field from one does not affect the other, or else we must shield either the parts producing the fields or the parts which may be affected by the fields. Shielding is most effective when placed around the parts which produce the fields, but when this cannot be done the shielding must go around the part to be protected.

3 Shielding for magnetic fields at power line frequencies and the lower audio frequencies must provide a path through which may flow the magnetic lines of force, as at 1 in Fig. 77-1. The magnetic lines concentrate in the shield metal and are diverted from the part to be protected. The shield metal for such applications should be made from high permeability iron or steel, which means the ability to carry magnetic lines very easily.

5 At radio frequencies the action is quite different, as illustrated at 2. The magnetic lines induce eddy currents in the shielding metal. These are small currents that whirl around the lines of force. The eddy currents produce magnetic fields of their own. The polarity of the eddy-current fields is opposite to that of the inducing magnetic field. The greater the conductivity of the shield metal the stronger are the eddy currents and their fields, and the more effective is the shielding. Any kind of metal may be used. Copper or aluminum are best, because of their high conductivity, but steel usually is employed because it costs less and does a good job. The higher the frequency of the magnetic fields the more effective is the shielding action.

Shielding for electric or electrostatic fields depends on still another action or principle, as illustrated at 3 in Fig. 77-1. If the polarity of the electric field is positive, the free negative electrons in the shield metal are pulled to the surface on which the field acts. The negative charge thus formed on the surface of the shield counteracts the positive charge that is causing the field. If the electric field is of negative polarity it drives free electrons away from the surface of the shield, and the surface remains with a positive charge. This positive charge on the shield counteracts the external negative field. The electric field lines of either polarity, end on the surface charges and cannot penetrate through the shield.

An electrostatic shield may be any kind of metal, of any thickness. The shield must be grounded to prevent the building up of a strong electric charge which, if retained, would radiate its own electric field

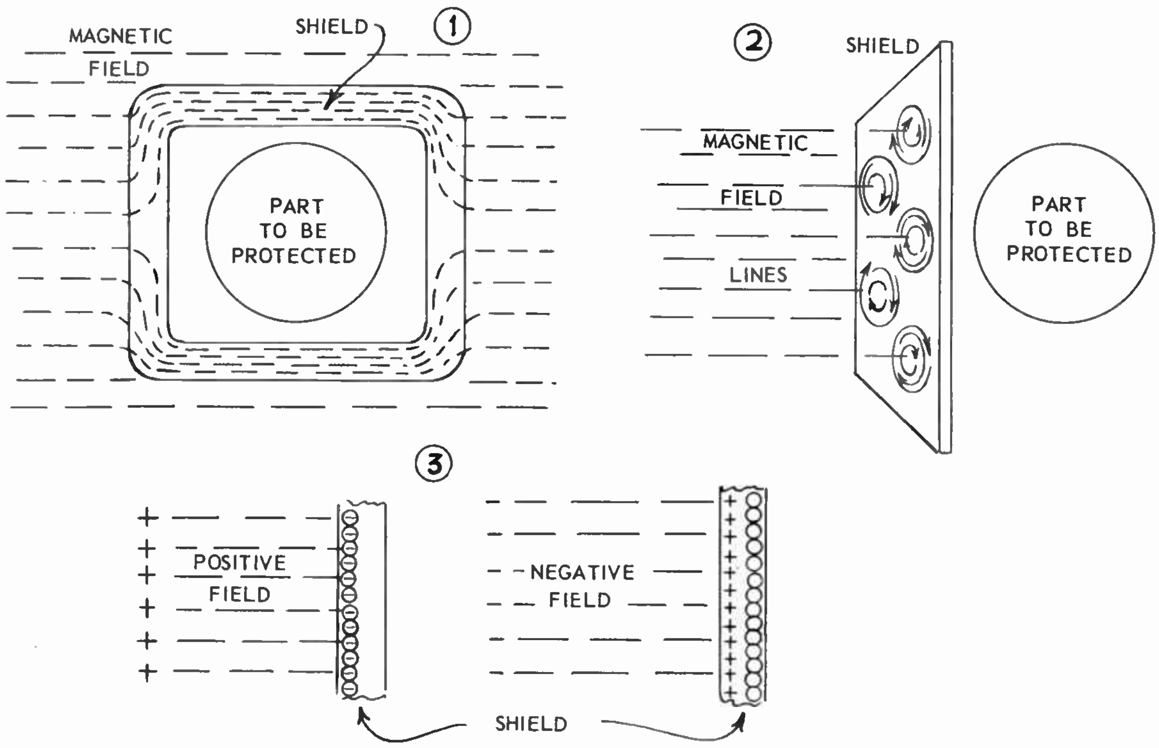


Fig. 77-1. The principles of magnetic and electrostatic shielding.

lines. Any grounded metallic shield is effective for electric fields and also for magnetic fields which alternate at radio frequencies.

SHIELDING PRACTICE. Shielding is used for tubes, all kinds of inductors, capacitors, individual wires and cables, complete amplifying stages, and power transformers. Fig. 77-2 is a picture of some typical shields. The two on the right are for glass-envelope tubes. Many glass tubes have an internal shield which is connected to one of the base pins. This pin, or rather the socket lug for it, is connected to ground in order to ground the internal shield. Such tubes may be used without additional shielding, but when employed in high-frequency high-gain circuits it is common practice to fit them with an external shield as well.

4 Some GT and GT/G glass-envelope tubes have a metal shell base which is connected to an internal shield and to one of the base pins. When an external shield is placed over the tube to make good contact with the base, this added shield is grounded through the connection to the socket lug for the grounding base pin. An external shield used with tubes having a base of insulating material must be well grounded.

Tubes having metal envelopes are shielded by the envelope. The envelope connects to one of the base pins, which must be grounded through the corresponding socket lug.

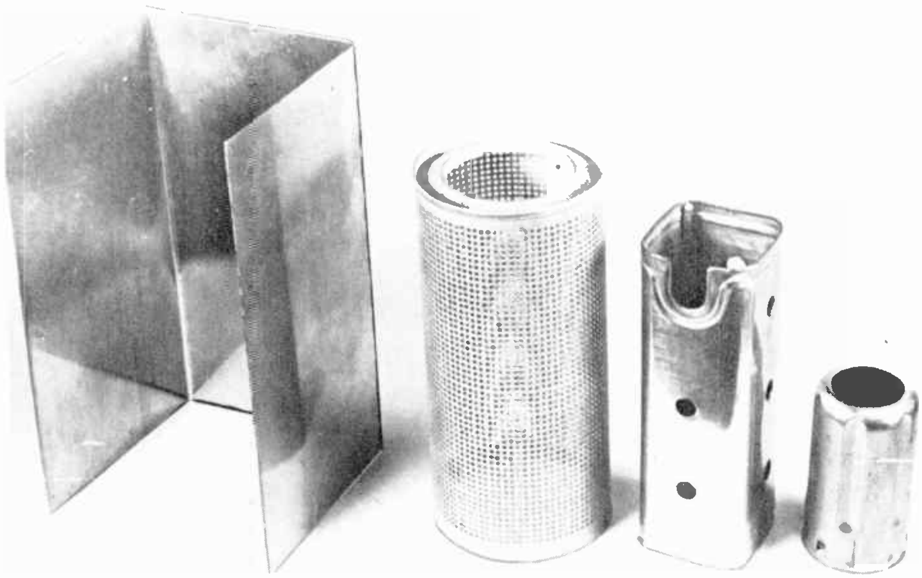


Fig. 77-2. Shields used for tubes and other parts of receivers.

Tuned transformers, single winding couplers, and other inductors in resonant circuits operating at radio frequencies or intermediate frequencies may require shielding. This is especially true where the windings are of large size and consequently have widely distributed fields of their own or have sufficient area to readily intercept other fields. The i-f transformers mounted on top of the broadcast receiver chassis of Fig. 77-3 are enclosed within metal shielding cans for this reason. The shields are grounded by the screws that hold them on the chassis.

Tuning inductors of small physical size often are used without shields. If a shield is used with any high-frequency inductor, large or small, there should be ample space between the inside of the shield and the winding if the Q-factor of the tuned circuit is not to be reduced by an excessive amount. Suitable clearance between inductors and a shield can is illustrated by Fig. 77-4, where the width of the shield is about three times the diameter of the windings and the height of the shield is more than double the distance from top to bottom of the windings. For television intermediate-frequency operation the shield diameter never should be less than twice the coil diameter, and the top and bottom of the shield should be above and below the coils by no less than one and one-half times the coil diameter.

When r-f chokes are used for decoupling, and there is shielding of the amplifier stage guarded by the chokes, the chokes always should be mounted inside of the stage shielding.

There is effective shielding between coils, other parts, or complete circuits, when one of them is mounted above the chassis deck and the other one below. R-f oscillator coils for standard broadcast and

Fig. 77-3. The i-f transformers are enclosed within shielding cans which ground to the chassis.

short-wave receivers often are placed under the chassis when r-f amplifier and i-f amplifier coils are mounted on top. The same principle is utilized in the mounting of many television i-f couplers.

The sections of ganged variable tuning capacitors usually have metal baffles between them, with the baffles attached to the grounded frame and rotor structure. The tuning capacitor in Fig. 77-3 is made this way. The grounded baffles act as shields to prevent coupling between r-f amplifiers and r-f oscillators which are tuned by the several sections of the capacitor.

In paper capacitors of inductive construction one of the terminals or pigtails is connected to the inside of the foil and the other one to the end of the last outer turn. When such a capacitor is used in a circuit where one of its terminals is connected to ground, this grounded terminal always should be the one on the outer foil, for then the foil acts as a fairly effective grounded shield for the remainder. In any other circuit the outer foil terminal should connect to the point of lowest a-c or r-f voltage, to the point which is toward the B power supply instead of toward a resonant circuit. The outer foil terminal usually is identified by a dark colored band, by an arrow, or by the words "outer foil".

When r-f oscillator coils, mixer coils, and r-f amplifier couplers are shielded or isolated from one another, and when there is effective decoupling, it is rarely necessary to shield complete amplifying stages from one another. If stage shielding is employed it may be carried out as shown by Fig. 77-5, where the shielding enclosures are indicated by the broken lines.

Within the shield containing one tube will be any kind of tuned grid circuit for that tube. A grid capa-

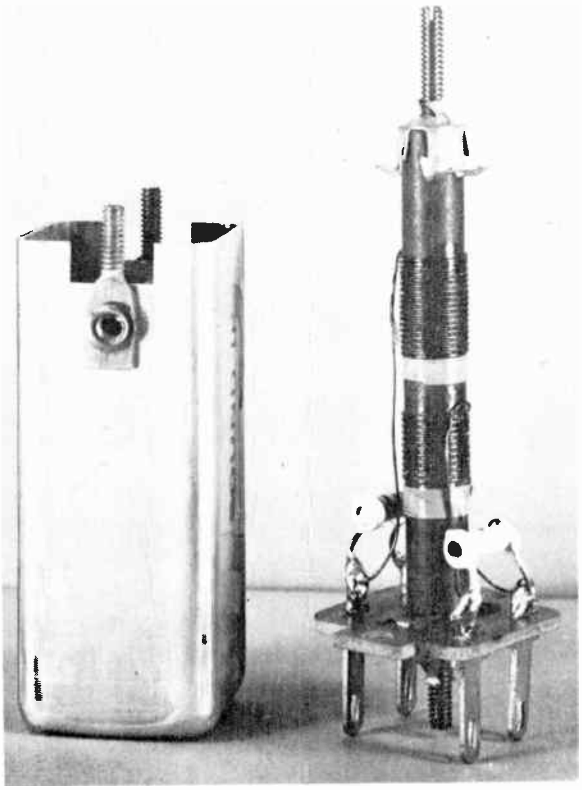


Fig. 77-4. A shield should be wider or of greater diameter, and also of greater height, than the coil windings which are protected.

citor and grid resistor or leak should be enclosed with the tube that they bias. As mentioned before, a decoupling choke should be enclosed with the tube whose plate and screen circuits are thus decoupled. This rule applies also to decoupling resistors. Cathode bias resistors and their bypass capacitors are with the tube that is biased.

High-frequency voltages from a-c power lines in a building may be coupled into amplifiers and other receiver circuits through the power transformer. The primary and secondary windings of the transformer consist of bodies of metal separated by insulation, which act as plates and dielectric for a capacitance which usually is several hundreds mmf.

This coupling is reduced or prevented by placing between the primary and other windings an electrostatic shield such as pictured by Fig. 77-6. Power transformers which are built with such a shield usually have it grounded to the core, on the assumption that the core will be grounded to the receiver chassis. Some transformers have a separate external lead from the shield, this lead being used for grounding. Unless the shield is grounded in one way or the other it is ineffective.

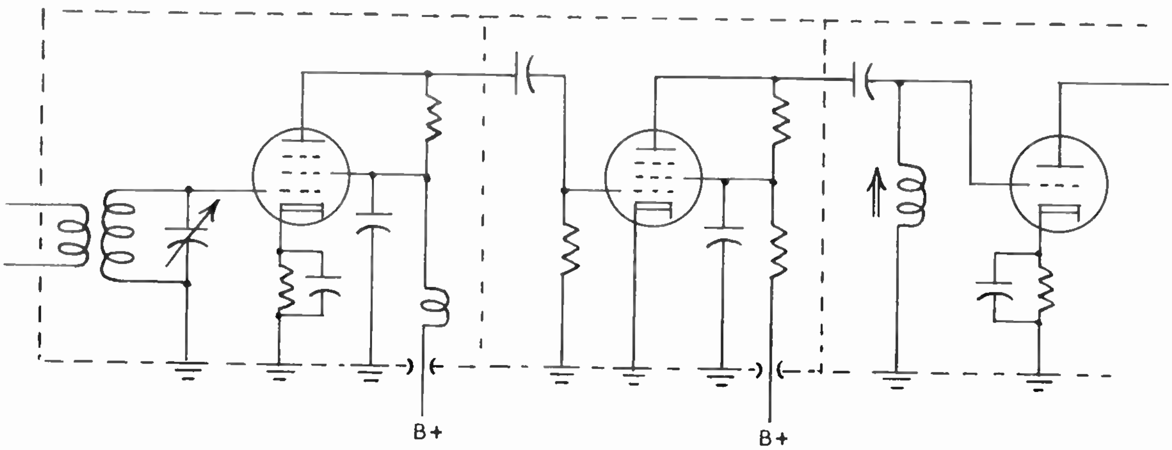


Fig. 77-5. All the parts of each amplifying stage may be enclosed within a single shielding cover.

Power transformers with electrostatic shields are used in receivers for protection against power-line interference of all kinds, including the voltage pulses resulting from opening any electric switches in the building wiring and from faulty operation of electrical appliances. These transformers are used in r-f signal generators to keep the generated high frequencies out of the power line, from where they would be radiated to any nearby radio or television apparatus.

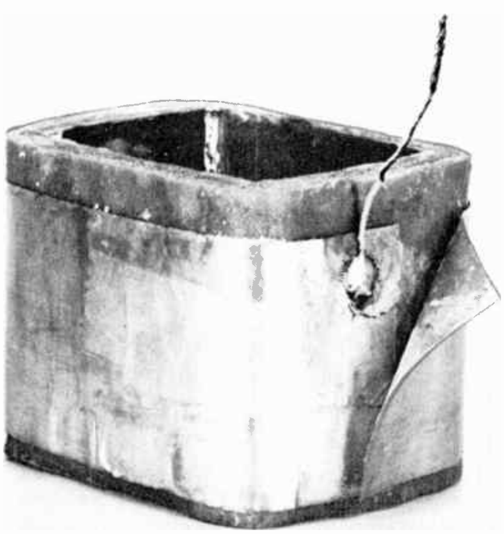


Fig. 77-6. An electrostatic shield around the outside of the primary winding in a power transformer.

When aligning or otherwise adjusting resonant circuits during service operations it is necessary to consider the effects of shields which are applied to tuning coils or tubes, or to parts of the receiver containing these parts.

When a shield is applied over a coil it always changes the inductance of the coil. Therefore, a resonant circuit which includes the coil will tune to a different frequency with the shield than without it. The greater the clearance between the shield and the coil winding the less is the effect on inductance and resonant frequency. The shield of Fig. 77-4 would change the coii inductance by about four per cent. A shield half as wide probably would make a change of 25 per cent.

A tube shield alters the resonant frequency of any tuned circuit connected to the tube. If a television r-f oscillator is aligned with the oscillator tube shield removed, and then the shield is replaced, the oscillation frequency usually will have been changed so much that a fine tuning control cannot correct it.

Shielding of any kind adds to the stray capacitance in the circuit for which the shielding is used. This is an important consideration in the design of high-frequency resonant circuits, such as those of television r-f and i-f amplifiers. It is a general rule that television and f-m receivers should be aligned with all shielding in place, if this is at all possible. Otherwise there is likely to be a considerable change in performance when shields are replaced.

GROUNDING. Men who have worked exclusively with broadcast radio receivers and with ordinary audio amplifier systems have learned that a ground connection anywhere on a chassis is equivalent to one anywhere else on the same chassis. At carrier and intermediate frequencies used for television and f-m broadcast, and even in high-fidelity audio amplifiers, this is no longer true. There is enough resistance in chassis metal to cause coupling at high frequencies.

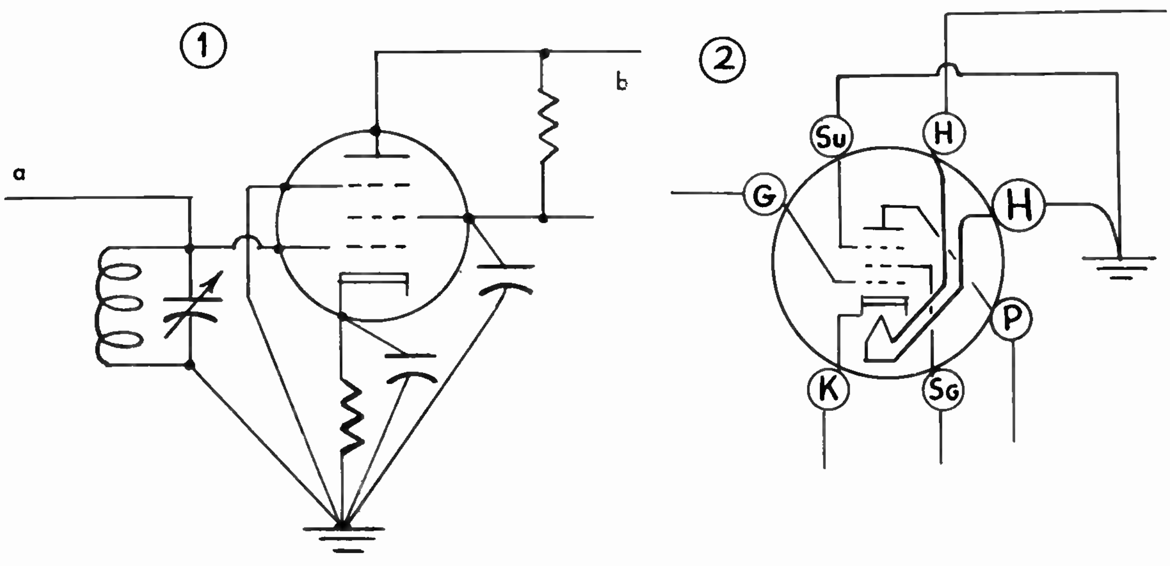


Fig. 77-7. Ground returns from all the elements of one tube should go to the same point on the chassis metal, and the ungrounded heater lead should be near tube pins for elements of least sensitivity.

COMMERCIAL TRADES INSTITUTE

TELEVISION

What probably is the most important single rule to be observed in the grounding of high-frequency circuits is illustrated by diagram 1 of Fig. 77-7. The rule is this: Connect the ground returns for all elements of any one tube to the same point. To begin with, the tube cathode or a cathode resistor is connected to some point on the chassis that is close to the tube. This same point is to be used for grounding all the other element returns to the cathode. In our illustration the grid return is through the tuning coil, or coil and tuning capacitor, whose low side is connected to our common ground. The suppressor is connected to the same ground point. So is the cathode bypass capacitor, and the screen decoupling capacitor.

If the bypass and decoupling capacitors have sufficiently small reactance, the paths for alternating signal currents of this one tube all come to the same (ground) point. None of these paths, except those for the incoming signal at a and the outgoing signal at b, extend beyond the parts shown by the diagrams.

Now look at diagram 2 of Fig. 77-7. It applies to a tube whose heater is in parallel with all other heaters. In any parallel heater system it is customary to carry one side of the heater supply circuit through chassis ground, and to run the other side through an insulated wire. The question is, which of the two heater terminals should be grounded?

With a tube whose base pins are arranged as in the diagram we should ground the pin that is next to the pin for the plate. Then the plate, which is one of the signal-sensitive elements of the tube, is unlikely to pick up a power line frequency from the heater pin and heater connection since the heater connection near the plate goes directly to ground. The other heater connection, which is on the hot or ungrounded side of the heater circuit, now is next to the suppressor terminal, and the suppressor is grounded. Which heater terminal connects to ground depends on the arrangement of base pins on the particular tube considered, but in any case it is merely a matter of using common sense. Get the grounded side of the heater toward a more sensitive element terminal, and the ungrounded side toward a less sensitive element.

Sockets for miniature tubes have a small metal ferrule at their center, in between the lugs for base pins. This center ferrule usually is connected to ground in order to lessen the capacitance and capacitive coupling between socket lugs and base pins of the tube. If the ferrule originally is grounded, keep it so when replacing the socket or making any wiring changes.

Ground connections are best when soldered directly to chassis metal. On many chassis there are numerous small projecting lugs put there for the express purpose of grounding. As mentioned before, always make the common ground connection for any one tube as close to that tube as possible. In high-frequency work never run a grounding wire or "ground bus" and connect the various grounded conductors to it.

Finally, when making any replacements or any wiring changes, always make all the ground connections exactly as originally placed. Any changes of grounding may upset the alignment, and are quite likely to allow harmful couplings.

POSITIONS OF CIRCUIT ELEMENTS. Careful consideration has been given by the manufacturer to the layout of all parts in high quality r-f amplifiers, i-f amplifiers, video amplifiers, and wide band audio amplifiers. Performance is likely to suffer if you make careless changes, especially in the arrangement of capacitors, inductors, and most resistors.

Tuned transformers or other tuned couplers should be as close as possible to the tubes of whose grid circuits they are a part, this is for the purpose of shortening the grid leads. Tuned oscillator circuits should be isolated as well as possible from all other circuits until they come together in converters or mixers. Peaking coils in video detector and video amplifier circuits should have short leads and should be kept as far from chassis metal as is consistent with lead length. R-f chokes should be mounted close to chassis metal unless the chokes are shielded, in which case they may go anywhere so long as their shields are grounded.

High-gain audio amplifier tubes must be kept out of low-frequency magnetic fields, or the fields kept away from such tubes. Low-frequency magnetic fields are found around power transformers, filter chokes in power supplies, and loud speaker field windings which are employed as filter chokes. The amplifiers will pick up power frequency hum or hum at twice the power frequency (with full-wave rectifiers) from fields of this kind. Care should be used also to keep high-gain audio amplifiers away from permanent magnet loud speakers, whose steady fields may divert the electron streams inside the tubes.

Interstage coupling capacitors should be kept well away from chassis metal to avoid having so much "capacitance to ground" as would bypass high-frequency signals. This applies to all high-frequency coupling capacitors, including those from the last video amplifier to the grid or cathode of the picture tube. A coupling capacitor should be placed close to the socket terminal for the grid of the following tube, with any extra length of pigtail or lead toward the preceding plate. Do not change the positions of coupling capacitors which are between tuned circuits unless you are ready to re-align the affected stages.

Audio-frequency coupling capacitors should be placed close to the chassis if they are fairly close to

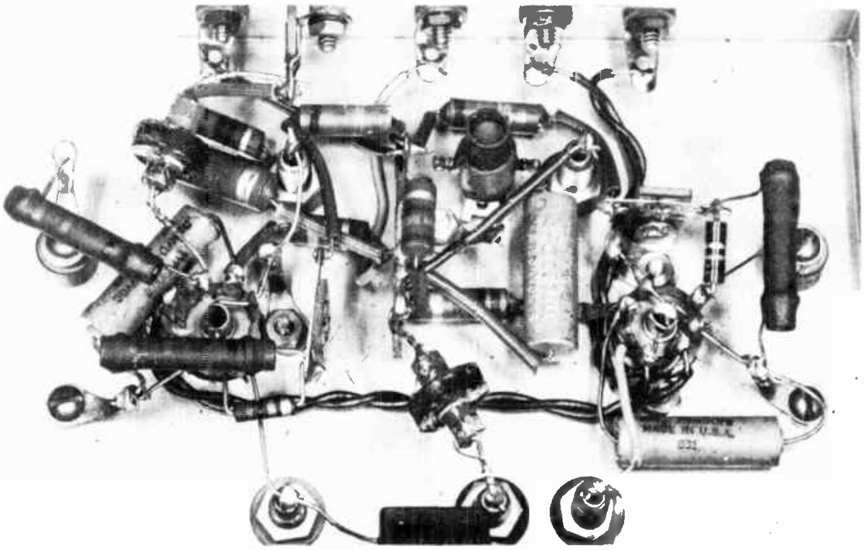


Fig. 77-8. The many resistors, capacitors, and inductors associated with an amplifying stage require careful layout. Here there is too much separation, and many parts which should be dressed close to chassis metal are far from the metal.

heater wiring or other leads carrying power-line frequency or power rectifier output frequency. This advice applies also to a capacitor which is between the slider of a volume control and following circuits. All audio-frequency capacitors are likely to pick up hum and put it into the amplifier circuits.

Bypass or decoupling capacitors for plates, screens, grids, and cathodes should be placed as close as possible to the socket lugs for these elements, with any extra length of lead or pigtail on the ground side of the capacitor or on the cathode side when bypassing is to the cathode of the tube. Even on the ground or cathode ends of these bypasses the leads should be short and direct. As an example, were a bypass to go between a plate or screen and the tube cathode, the capacitor should be mounted right on the socket lugs for these elements.

Coupling capacitors and also bypass or decoupling capacitors which become defective should be replaced with exact duplicates in order to preserve the original capacitances to ground and to other parts and the original inductances which exist to some extent in the capacitors. The replacement should at least match the original as to dimensions and construction, and, of course, as to capacitance value.

Grid resistors or grid leaks should go close to the tube sockets, with any extra length of lead in the ground or B-minus connection. Plate load resistors are treated similarly, placed close to the tube socket, with extra lead length on the B plus side.

If you make any alterations which require adding new parts such as switches or jacks, try to mount the added parts well away from all circuits carrying radio-frequency or intermediate-frequency currents.

When you remove tubes of the same type number from two or more sockets, mark each tube so that it may be replaced in the socket from which it came. The characteristics of two tubes of the same type number are not necessarily the same. Neither are their internal capacitances. Interchanging of tubes in r-f amplifier, r-f oscillator, i-f amplifier, or sync circuits will require realignment or readjustment. This does not mean that tubes never should be switched around. As an example, you might find that a weak video amplifier would work all right as an i-f amplifier, while the original i-f amplifier makes a better video amplifier. If you make the exchange it will be necessary to re-align the i-f amplifier.

DRESSING OF LEADS. The position of wiring connections between various circuit parts is called the "lead dress". We speak of dressing a lead close to or away from something else, or along some certain path. Correct dressing of leads has four principal purposes.

1. Reduction of stray capacitances. This may be accomplished by keeping certain leads clear of one another or clear of the chassis metal and of insulation having low dielectric constant.

2. Reduction of stray inductance. A lead less than an inch long, and of straight wire, has enough inductance to be resonant at television carrier frequencies when associated with stray capacitances or other capacitances. Inductance is reduced by using the shortest possible leads. Don't, however, change the lengths of leads in very-high or ultra-high frequency circuits, for the lengths may have been chosen to provide desired tuning inductances.

3. Reduction of couplings which are not wanted. Couplings are reduced or prevented by reduction of stray capacitance and of stray inductance, by following the practices mentioned in items 1 and 2.

4. Prevention of frequency mixups. That is, frequencies which should exist in only certain sections of the receiver should be kept out of other sections. Extra care is needed with circuits of high impedance. A circuit in which there is high impedance or high resistance ordinarily operates with high signal voltages. Such a circuit is more susceptible to pickup of unwanted frequencies than is a low impedance circuit.

Some of the practical applications of the principles will be explained, but in other cases you will have to guide yourself in accordance with the general rules.

Grid leads and plate leads should be kept well separated from each other, at least from the socket terminals as far as the decoupling capacitors, in order to prevent feedback couplings. Grid leads should be dressed away from the chassis metal, to avoid loss of signal strength through capacitance to the chassis. If the plate leads are long they may be dressed down fairly close to the chassis, since there are relatively strong signals in plate circuits. Both grid leads and plate leads should be short, but when a choice must be made on a line between stages it is best to keep the grid lead short and lengthen the plate lead.

Leads from any kind of antenna to any antenna coupler or to a tuner should be treated like grid leads. Don't forget that the lead to the signal grid of a converter is a grid lead.

Leads for r-f oscillators should be kept away from all other wiring, and from each other. Connections for amplifier and mixer circuits in a tuner should be as far as possible from r-f oscillator circuits in the

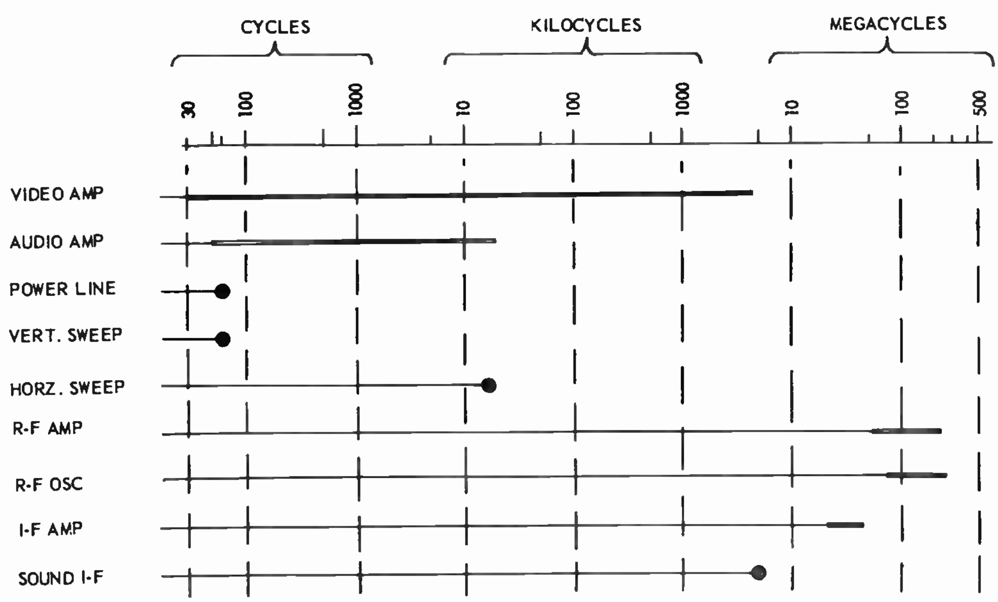


Fig. 77-9. There may be unwanted couplings and voltage transfer between any two circuits whose normal operating frequencies are in the same range.

tuner. Should it be necessary to replace any tuner wires that carry r-f or oscillator currents, always use the same length and same gage size as originally employed.

Screen leads should be dressed close to chassis metal. This is true also of leads which are the buses for automatic volume control or automatic gain control.

R-f and i-f circuits are not particularly susceptible to pickup of the much lower frequencies which exist in circuits carrying power line frequencies, audio frequencies, or vertical or horizontal sweep frequencies. The parts and wiring of these low-frequency circuits often are quite close to r-f and i-f circuits without causing trouble.

All the parts and wiring of audio amplifier circuits and of video amplifier circuits are capable of picking up voltages from other sections in which the frequencies extend through similar ranges. This means that there will be pickup from circuits operating at power line frequencies, at vertical sweep frequencies, and at horizontal sweep frequencies. Power line frequency will cause hum from the loud speaker or wide horizontal blackout bands across the picture tube. Vertical sweep frequency will cause a buzz from the speaker and black bars on the picture tube. Horizontal sweep frequency will cause a hissing sound from the speaker and black lines on the picture tube. Fig. 77-9 shows which frequencies may overlap or conflict.

The video amplifier circuits will pick up audio frequencies or audio modulation to cause "sound bars" across the face of the picture tube. Sensitive audio circuits can pick up the lower video frequencies.

We must not forget that the audio system of a receiver begins in the transformer or coupler that feeds the detector or demodulator, including both secondary and primary leads for a transformer, and continues all the way to the loud speaker. The video amplifier system must be considered as starting in the transformer or coupler that feeds the video detector, and it continues all the way to the grid or cathode of the picture tube.

There are many wires carrying power-line frequency. They include the wires from the power supply cord connection to the power transformer and to the on-off switch. They include also all the heater wires, and wires to a pilot lamp. Power-line frequency or twice this frequency is present in connections from the power transformer to the power rectifier, and from rectifier to filter. All these wires should be dressed close to the chassis metal, or run several inches from circuits sensitive to the low frequencies, or else be shielded wire.

Shielded wire often is used for leads between the volume control and other audio-frequency parts farther back in the chassis. It is used also between separate sections of large receivers. Shielded wire usually is kept in contact with chassis metal as much as possible. The shield covering should be solder-grounded to chassis metal at both ends, and at one or more intermediate points if the lead is long.

Excess length of any lead should be cut off before making connections, not coiled or pushed into corners. If the lead is not to be cut, the excess length should be dressed back close to the part at lower a-c or signal voltage, toward whichever end of the connection comes closer to ground, to a B plus line, or to a decoupling capacitor.

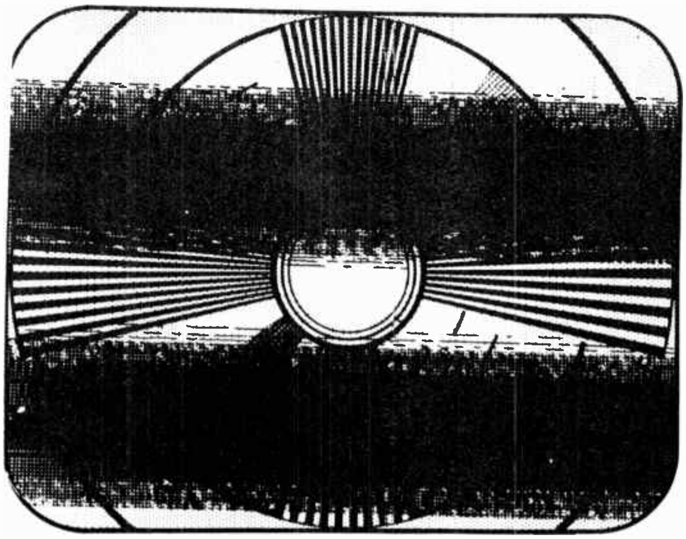


Fig. 77-10. How the pattern or picture is affected by a strong 120-cycle hum voltage.

1 *HUM.* The word "hum" originally referred only to the low pitched sound from a loud speaker caused by introduction into audio amplifier circuits of the power line frequency, or, with a full-wave power rectifier, of twice this frequency. Now we speak also of hum as affecting a television picture. The picture tube does not emit an audible hum, but patterns or pictures are affected by power line frequencies or double frequencies as shown by Fig. 77-10. Two black bands indicate a 120-cycle hum voltage, while one band would indicate 60 cycles.

A 60-cycle audible hum is disagreeable, but usually does not sound very loud. This is partly because ordinary audio amplifier systems have relatively low gain at 60 cycles, and partly because our ears are not too sensitive to a 60-cycle sound. A 120-cycle hum from an equal a-c voltage sounds louder and interferes to a greater extent with enjoyment of music and speech.

When a low-frequency hum voltage gets into a circuit that is carrying a higher frequency signal voltage, the signal rides the hum as shown by Fig. 77-11. Each voltage retains its own amplitude. Zero for hum alternations is on line a of the graph, while zero for the signal rises and falls with alternations of hum voltage. Were both these alternating voltages to be components of a direct voltage, zero for the direct voltage might be on line b.

2 Hum often originates in the low-voltage B power supply. The most common cause is insufficient filtering, which leaves too much ripple voltage. Doubtless you will recall that we examined a two-section power filter with the output of the first section furnishing B voltage for the output amplifier and output of the second section furnishing B voltage for a preceding amplifier. Any hum voltage fed to the preceding amplifier receives more amplification than hum voltage fed to the output amplifier, but since the B voltage to the first amplifier is better filtered there is an overall reduction of hum in the output.

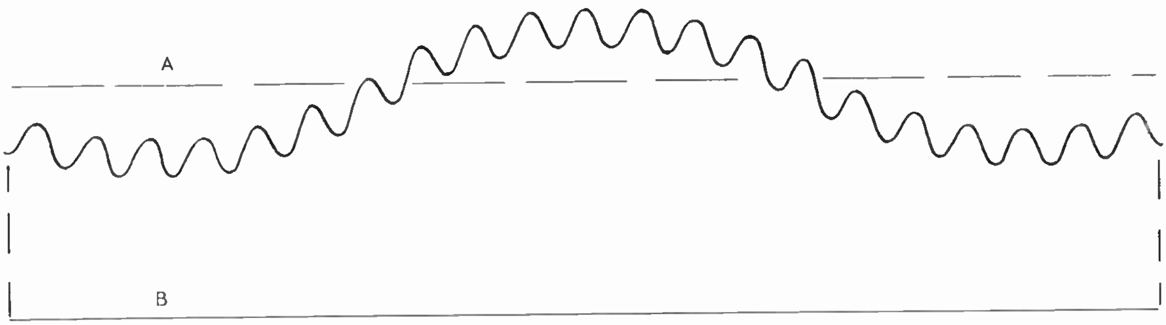


Fig. 77-11. A high-frequency signal voltage rides a low-frequency hum voltage.

It is easy to determine the amplifier stage most affected by hum voltage, or in which stage a hum originates. Commencing with the output stage and working back, short circuit each grid or each input to ground or to B-minus through a capacitor of one mf or more, or directly. If this shorting reduces the hum materially, the trouble originates in some earlier stage.

Hum will result from an open or disconnected power filter capacitor, or from high resistance in an old electrolytic capacitor which should be replaced. Long leads between the power transformer secondary and the rectifier plates may cause hum.

Hum may be due to the magnetic field from a power transformer or a power filter choke which gets into an iron-core audio output or speaker coupling transformer, or even into some air-core inductor. Turning the power transformer or choke into some different position may get rid of such hum by changing the direction of field lines.

A power transformer with core laminations loose enough to vibrate may cause hum by vibrating the r-f oscillator tube or any parts of the oscillator circuit, even the wiring connections. A really defective power transformer is almost certain to cause hum. The fault may be a few short circuited turns in a winding. Some cheap power transformers have such deficient coupling between primary and secondary as to allow spread of their magnetic fields over large areas.

Hum may result from amplifier tube shields that are not well grounded. Socket lugs for the pins connected to internal shields or to metal envelopes may have been left ungrounded. A high-gain amplifier tube may be exposed to an alternating magnetic field from the power system.

Ineffective decoupling is very likely to allow hum voltages to get into amplifier stages. Decoupling and bypass capacitor connections should be examined, and the capacitors themselves should be tested for capacitance, opens, shorts, and leakage if there is any reason to suspect the stage in which they are used.

Incorrect dressing of wiring connections is a common cause for hum pickup. As you will note from Fig. 77-9, low-frequency hum voltages are easily picked up by video and audio amplifiers. When examining the dress of wiring look for loose or defective soldered joints. They too can be the cause for hum.

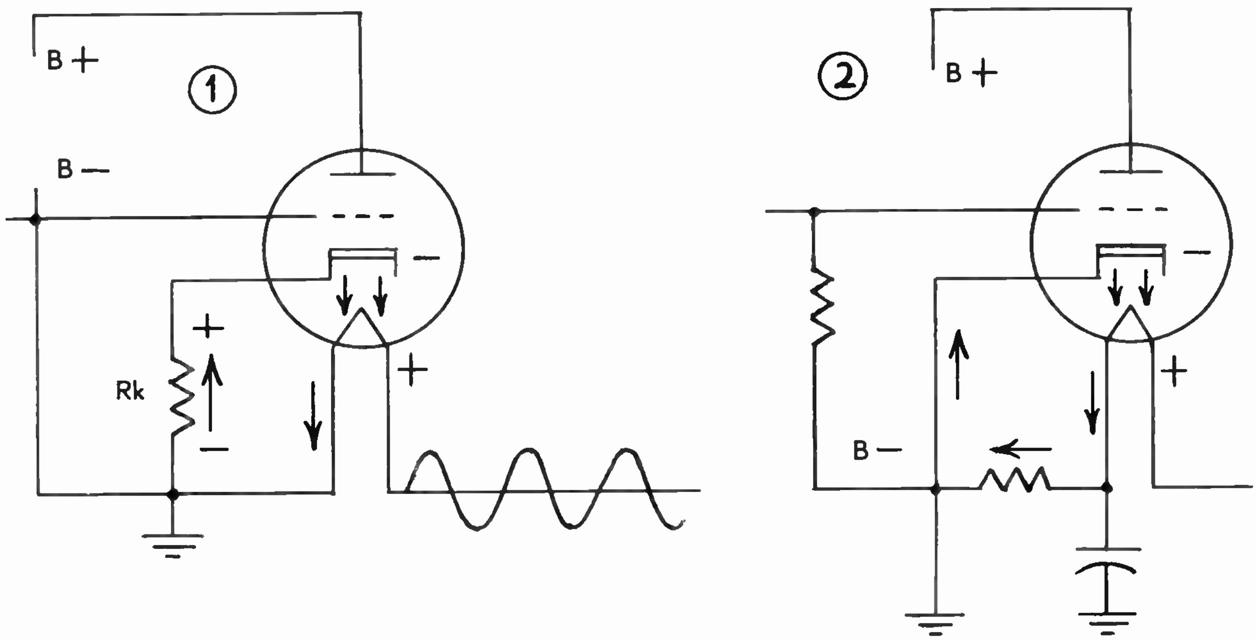


Fig. 77-12. Electron flows which may cause cathode-heater hum.

Degeneration applied to audio amplifiers will reduce the effect of hum voltages, but it will not eliminate hum which is due to poor filtering. Degeneration applied to the stage in which the hum appears to originate may be very effective. The hum voltage from the plate circuit is fed back in negative phase to the grid, and is reduced. Signal voltage coming to the grid from preceding stages is not directly affected by this hum feedback, and there is a net improvement.

It is possible for hum voltage to result from cathode-heater electron leakage within a tube, as shown at 1 in Fig. 77-12. Anything which makes a heater sufficiently positive with reference to the cathode will cause some of the emitted electrons to pass to and into the heater, just as they would pass into a positive element. The heater may become positive during positive alternations of a-c heater voltage. This can happen when peak values of the heater voltage exceed the cathode bias voltage. Then, during each positive alternation of heater voltage, there will be a pulse of current or electron flow from the heater through the bias resistor and back to the cathode. The accompanying pulses of voltage in the bias resistor are applied to the grid, and amplified by the tube affected and by following stages. The resulting hum voltage always will be at power-line frequency, since this is the frequency of heater voltage.

Cathode-heater hum produced as described may be reduced or eliminated by a large decoupling or by-pass capacitor across the bias resistor. The decoupling capacitance has to be very large, because the frequency of the hum voltage is only 60 cycles. Capacitors of 25 mf or more may be needed in high-gain audio amplifiers. In r-f and i-f stages the regular cathode decoupling capacitors usually are of small values, since only a little capacitance is needed at the high frequencies. These small paper, mica, or ceramic bypass capacitors for high-frequency decoupling may be supplemented by a large electrolytic

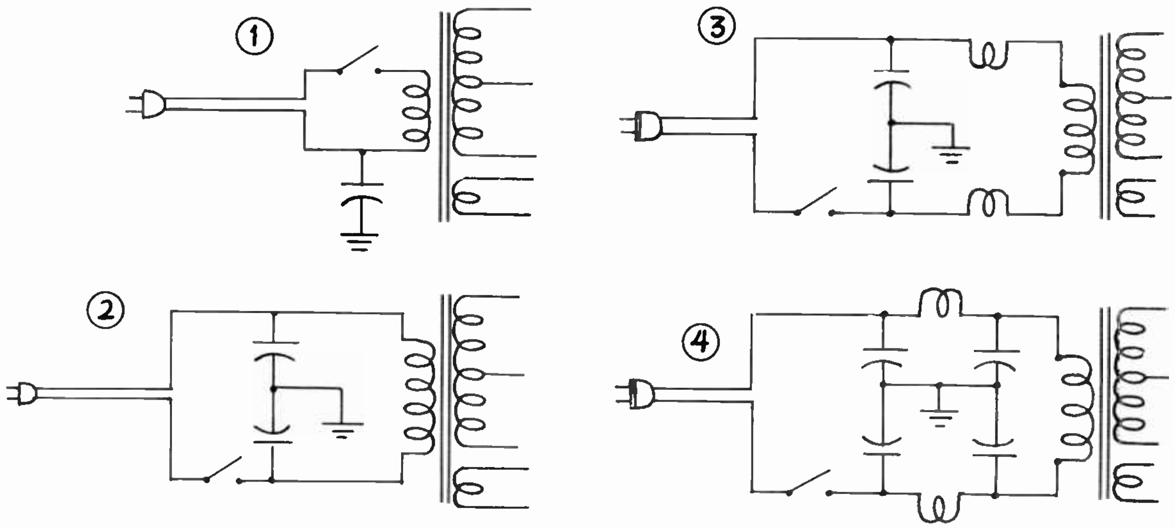


Fig. 77-13. Power line filters.

capacitor for prevention of cathode-heater hum which otherwise could originate in these stages. If a cathode connects directly to ground or to B-minus, with no biasing resistor, this kind of hum voltage will not develop.

There are other cases, as illustrated at 2 in Fig. 77-12, where there may be a large resistance or a large difference of potential between the heater and ground, and consequently between the heater and cathode. This condition may exist in transformerless receivers having series heaters. Cathode-heater leakage currents in this resistance would not ordinarily reach the grid of the tube in which is the leakage, but may be carried to other stages through the heater connections. The remedy is adequate bypassing or decoupling around the heater line resistance, or from the heaters to ground or to B-minus.

Cathode-heater hum is increased by operating tubes with excessively high heater voltages. Usually it is found that detectors and demodulators of all kinds are most likely to suffer from this kind of hum.

A kind of trouble called modulation hum appears only while a receiver is tuned to receive some carrier, and disappears between stations. The carrier is being modulated by a hum voltage. This hum may result from poor filtering of the power supply, or there may be coupling to the power line through the transformer. Such coupling is reduced by an electrostatic shield built into the transformer, and well grounded, also by using a line filter.

Line filter circuits are shown by Fig. 77-13. As at 1, the filter may be nothing more than a paper capacitor from either side of the power line to ground. There is better filtering with grounding capacitors on both sides of the line, as at 2, still better with both capacitors and r-f chokes as at 3, or with regular low-pass filters in both sides of the line, as at 4. The capacitors usually are of 0.01 mf value, and never of working voltage rating less than 300 for use on 110-120 volt power lines. The r-f chokes must be made of wire large enough to carry the a-c power current, and suitable insulated for connection to building power lines. It is always best to purchase line filters manufactured for this class of work.

Hum which appears only while tuned to a station may result from pickup in r-f oscillator circuits of any low-frequency voltages from B-power lines that extend into other sections of the receiver. This trouble is caused by ineffective decoupling.

CATHODE FOLLOWER. Earlier we spoke of the danger of coupling and signal pickup, and of possible regenerative feedback, into the conductors of a high-impedance circuit. In many cases it is worth while to reduce the impedance by using the arrangement of Fig. 77-14, which may be called a cathode follower, a cathode coupling, or an inverted amplifier. The signal input is to the grid, as usual. Here we may have an impedance as great as five megohms or more, by using a suitable grid resistor Rg and by biasing the tube sufficiently negative to prevent grid current. The output is taken from across cathode resistor Rk.

The output impedance is the parallel impedance of the cathode resistor and of the effective plate impedance of the tube. Since any parallel impedance or resistance always is less than either of its parts, the output impedance of the cathode follower always is less than the resistance of the cathode resistor, usually it is much less than half this resistance and quite often is as low as 50 ohms. Output impedance is decreased in three ways. First, by using a smaller cathode resistance at Rk. There is a limit to the reduction of this resistance, because oftentimes its primary purpose is for biasing the tube. Second, by using a tube with less plate resistance. Third, by using a tube with greater transconductance or mutual conductance, or one with a greater amplification factor.

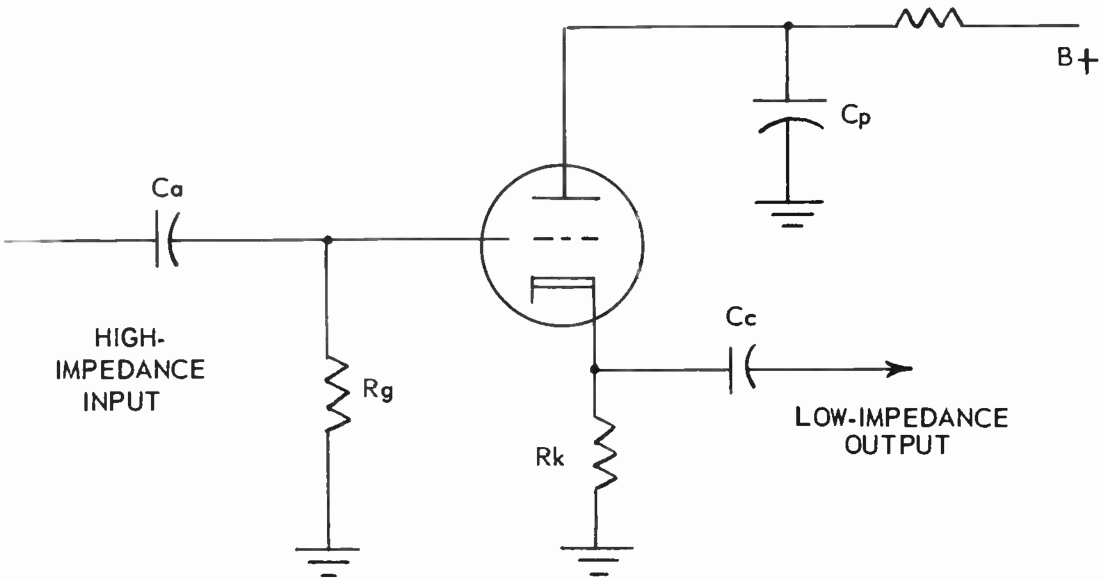


Fig. 77-14. Circuit connections for a cathode follower stage.

Note that the plate of the tube is connected in the usual way to the B plus supply, but, so far as signal currents and voltages are concerned, the plate is grounded through a large capacitance and low reactance at Cp.

Invariably there is a loss of signal voltage or a loss of gain between input and output of a cathode-follower. Rarely is the output signal greater than 0.90 of the input signal, and often it may be down around 0.60 to 0.70 of the input. This loss must be made up in other parts of the amplifying system. There is no inversion of signal voltage or polarity between input and output of the cathode follower.

In a connection made through a cathode follower it is possible to preserve uniformity of frequency response at very low and very high frequencies, since there is very little phase shift and the shunting capacitance is small. By omitting the output coupling capacitor, Cc of Fig. 77-14, and making a direct connection, it is possible to transmit the d-c component of a signal to a following stage.

Cathode coupling often is used to match the output of a video amplifier to a coaxial cable which runs to a remote picture tube unit, or to match the output of a high-fidelity high-gain audio amplifier to a remote loud speaker, or for matching to a cable which runs to any remote control unit. Wherever there is any tube with high-impedance input and a following volume or contrast control element of low resistance a cathode coupling may be used. Such couplings are common also in vacuum tube voltmeters, oscilloscopes, and other measuring and testing instruments whose input impedance should be maintained at a high level. Even in television sweep amplifiers we find cathode followers used to prevent feedback of pulse voltages to preceding circuits.

TELEVISION

LESSON NO. 78

NOISE AND INTERFERENCE

Training Your Mind

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

DO NOT TEAR – CUT ALONG THIS LINE AND SEND IN FOR GRADING.

Training Your Mind
C. T. I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON NO. 78

NOISE AND INTERFERENCE

The only variations of voltage that should affect the grid-cathode circuit of the television picture tube are those of the video signal carrying the picture. Any other voltage fluctuations will more or less completely distort the picture. The only voltage variations that should reach the sweep oscillators by way of the sync section are those derived from the sync pulses. Anything else may prevent synchronization. All this is a way of saying that the grid-cathode circuit and deflection system of the picture tube should be affected only by voltages of the composite television signal.

3 The number and kind of other voltage fluctuations which may reach the picture tube are astonishing, and they account for many of our more serious service problems. In a very general way we may classify all the unwanted voltages in accordance with their frequencies. If the undesired voltages occur simultaneously at scores of different frequencies with no regular pattern, and cover a very wide range of frequencies, they are classified as noise. If the trouble is due to some certain radio frequency or band of radio frequencies which follow a regular pattern, as in any kind of signal, it is classed as interference.

2 *NOISE.* Noise voltages may extend at the same time through frequencies from only a few cycles per second up to many megacycles. When such voltages get into the audio amplifier they may cause hissing, crackling, popping, snapping, and similar sounds from the loud speaker. When the same voltages get into the video amplifier they may cause bright flashes, small white streaks, black speckled marks, jittery pictures, and generally similar effects on the picture tube screen. Severe noise pulses may cause momentary loss of horizontal synchronization, vertical synchronization, or both.

Noise voltages are caused primarily by irregular changes of current or electron flow in circuits or parts which are inside the receiver, or which are outside and so connected or coupled to the receiver as to affect reception. Noise which originates outside the receiver nearly always is from electrical devices in the same building, or in the immediate neighborhood, or from devices connected to the same electric power service line as the affected receiver. A few of the things which may be suspected are as follows.

Any household appliance or convenience operated by an electric motor or vibrator, such as: Vacuum cleaners, mixers for foods or drinks, food grinders, dish washers, hair dryers, electric razors, sewing machines. Less commonly the noise comes from refrigerators, oil burners, and fans in heating or cooling systems.

Electric devices or machines in offices or factories, including: Anything driven by an a-c motor or any type of d-c motor. Elevators. Cash registers. Office machines which are motor driven. Electric welders. In some cases the noise comes from fluorescent lamps.

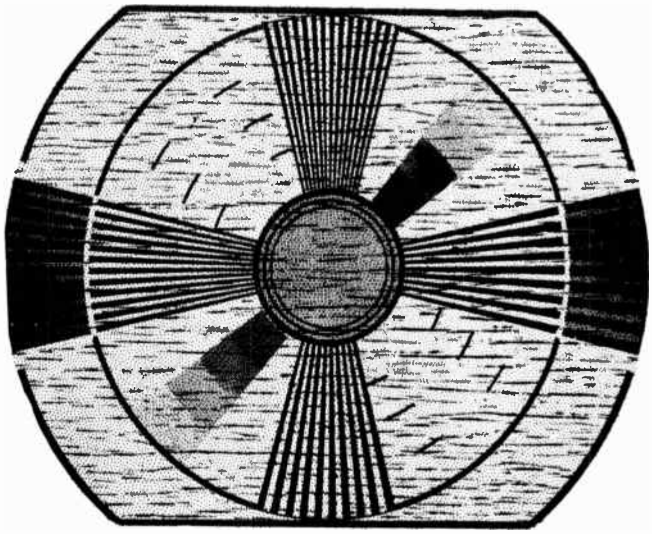


Fig. 78-1. One effect of noise voltages occuring at random frequencies.

A few of the noise sources found out of doors include: Busses and trucks on an adjacent street. Street and elevated cars and trains which run close to the building in which is the receiver. All types of flashing

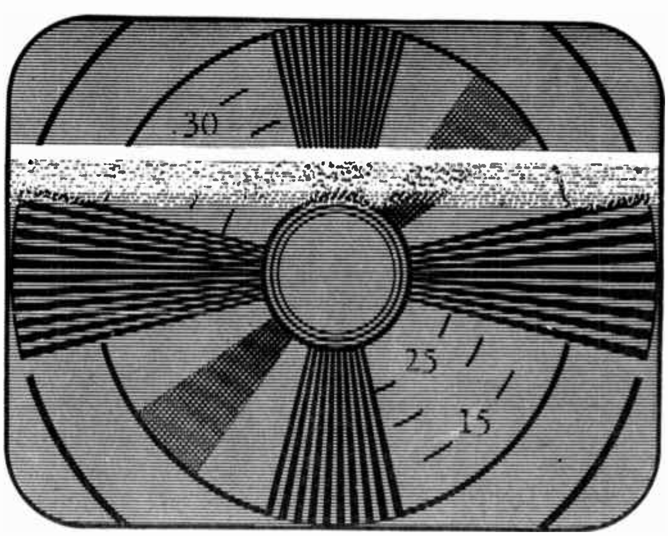


Fig. 78-2. Opening or closing a switch for nearby electrical apparatus may cause a bright flash on the picture tube screen.

signs. Automatic traffic signals. Some neon signs. Rarely the noise comes from private automobiles, since most of them have noise suppression for their own radio sets.

Picture flashes may be caused when any electrically operated device or machine is switched off or on. Fig. 78-2 shows a flash due to switching. Atmospheric static or lightning, which strongly affects standard broadcast radio sets, has but little effect on the very-high and ultra-high frequency circuits used in television and in f-m sound receivers.

3 Noise voltage fluctuations almost always have the form of amplitude modulation, not frequency modulation. Consequently, they may strongly affect the television picture, which is transmitted by amplitude modulation, and also the sync pulse portion of the television signal. These voltages may be picked up by the antenna, or they may directly affect the i-f or video amplifier circuits. The frequency range is so wide that all channels ordinarily are affected to about the same extent.

Noise voltages often enter a receiver through the power cord from a building circuit. Such trouble is reduced or eliminated by line filters and by power transformers having electrostatic shields. When noise effects are present you should try disconnecting the antenna or transmission line from the receiver. If the noise effects disappear, leaving a clear raster or a quiet loud speaker, the troublesome voltage probably is being picked up by the antenna or transmission line. If noise continues it probably originates in the receiver or else is entering from the power line. Noise filters may be used at any of the points shown by Fig. 78-3.

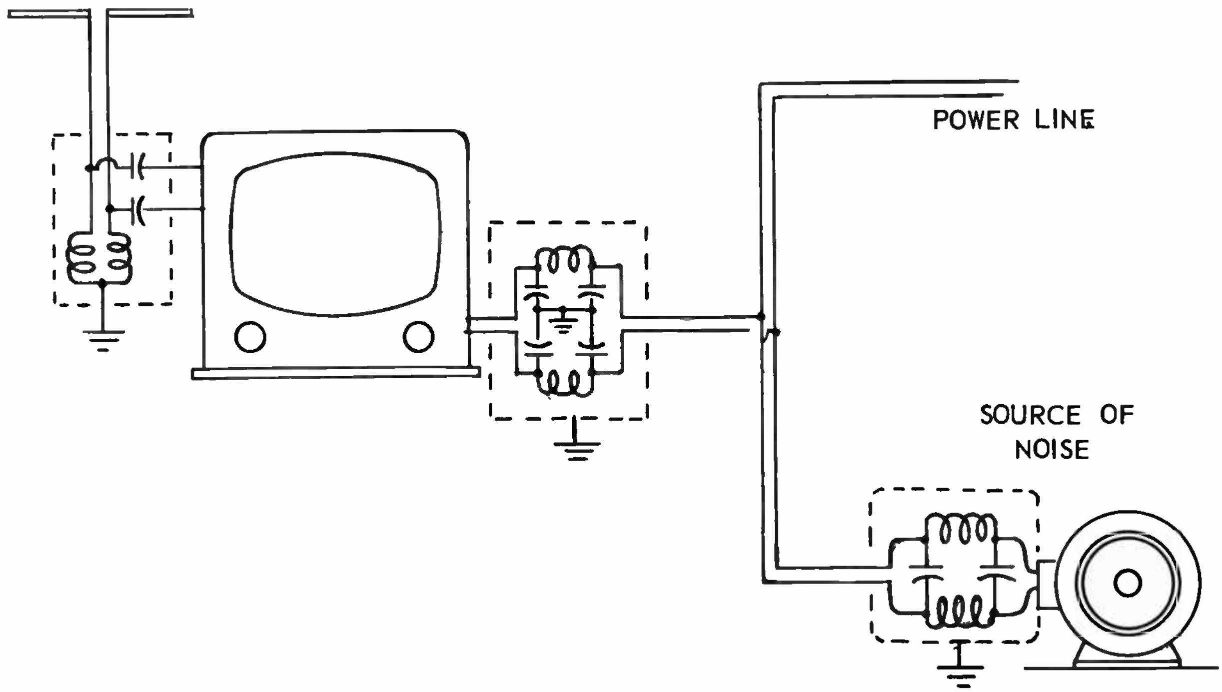


Fig. 78-3. Noise filters may be used at the transmission line, at the receiver power cord, or at the source of noise voltage.

COMMERCIAL TRADES INSTITUTE

TELEVISION

Noise which originates in any of the electrical devices previously listed is preferably stopped at its source rather than at the affected receiver. When filters and shielding are used at the receiver, the noise pulses still can come into the power lines and be radiated in the form of magnetic and electric fields which enter the receiver, or there may be field radiation from the offending device directly to the receiver.

Noise filters, usually called interference filters, are available in wide variety for installation at the source of trouble. Filters for large devices are designed for permanent mounting where the power wires enter the housing. Filters for small appliances are constructed to plug into the power receptacle or convenience outlet, and have openings that take the plug on the power cord for the appliance. Any filter must have sufficient current-carrying capacity for the apparatus to which connected.

Filter chokes have duolateral windings to reduce the distributed capacitance which would allow escape of high-frequency noise pulses. Inductances usually range from about 1/10 to more than 1/2 millihenry. The larger filters have bypass capacitors, and accordingly require a connection to ground.

Noise which originates within the receiver because of defective connections or parts may be due to any of a long list of faults, a few of which are as follows.

Any parts which are mounted so loosely as to vibrate when jarred or when reached by sound waves from the speaker. The noise is due to intermittent contacts of the loose parts with metal of the chassis. Look especially for shields which are loose or poorly grounded.

All soldered joints and screwed or riveted terminal connections are open to suspicion. Often there is looseness where two or more wire ends are soldered into the same lug. There may be cold solder or rosin joints, which may be corrected by melting the solder with a hot iron and letting it re-harden, or better by using a very little additional rosin-core solder while heating the joint. Screwed or riveted terminals may have corroded, and require cleaning or tightening. There may be defective insulation, although this usually would allow a dead short circuit. To locate the position of any of these faults, carefully and lightly tap, press, or pull on each suspected connection while the receiver is in operation. A change in the noise effects means that the trouble has been located.

Any adjustable control potentiometer or rheostat may cause noise, which usually becomes worse or better as the control is rotated. Resistance elements may have become rough along the surface contacted by the slider, or there may be looseness of part of a wire winding. The usual cause is overloading and overheating due to leakage, shorts, or grounds in other connected parts. Sometimes there is excessive direct current in a control unit whose primary purpose is to regulate an alternating current, with resulting noisy operation. The remedy is to connect a fixed capacitor of ample size, or low reactance, in series with the control unit.

1 The contacts in tuners, band switches, or in any other switches may be dirty or corroded. They may be cleaned with carbon tetrachloride on a small moderately stiff brush. Contacts which have become bent or which have insufficient pressure for any other reason sometimes may be corrected by careful pressing or very slight bending while the rotor is turned to some other position. If the contacts are scored, grooved, or badly worn the only correct remedy is replacement of the switch.

Variable tuning capacitors will cause noise voltages if there is dirt between the plates. Remove the dirt by turning the rotor all the way out and working a pipe cleaner in between each pair of plates of the stator and the rotor. Rotor and stator plates may touch against each other with the rotor in certain positions. The spacing usually may be corrected by turning the rotor into full mesh with the stator and then pushing between each pair of adjacent plates a stiff card which slides snugly into the spaces. Sometimes the entire stator section will have been pushed back or forth. With some constructions it is possible to loosen screws that hold the stator, then insert several spacing cards between rotor and stator plates, and tighten the stator support screws while the plates are held thus.

Effects similar to very strong noise voltages may be caused by flashovers or by corona discharges between any conductors in the high-voltage circuits for the picture tube anode. These troubles usually occur where there are sharp projections, as at joints where the solder has not been flowed smoothly and evenly into place. The most common location is at lugs on sockets for high-voltage rectifier tubes.

Even though none of the internal faults mentioned in preceding paragraphs are present there still will be thermal noise voltage produced in conductors and in the tubes there will be noise voltage due to shot effect.

Thermal noise results from irregular movements of free electrons in conductors, whether or not there is electron flow or current into one end and out of the other end. All the free electrons are constantly moving one way and another between the atoms. At any one instant more of the electrons which are not part of a through current may be moving toward one end than toward the other end of the conductor. This causes an instantaneous difference of potential or a voltage between the ends of the conductor, with the polarity of the voltage varying in accordance with the direction in which there happens to be a greater net movement of free electrons. These thermal voltages exist at all times, no matter what may be the material of the conductor, and regardless of whether there is current flowing all the way through the conductor.

Thermal noise voltages increase with higher resistance of the conductor, which means that fixed and adjustable resistors are the principal sources, and that high resistances tend to cause relatively high noise voltages. This thermal effect increases also with rise of temperature. A rise from 70° to 250° F would theoretically increase the thermal noise voltage by about one-third. Appreciable noise voltage may originate in filament-cathodes when the filament forms part of an amplifier circuit, this is because of the very high temperature of the filament. Thermal noise may be reduced to some extent by using resistors having wattage ratings great enough to insure operation at low temperature.

Thermal noise voltages are of a kind which would cause a continual hiss from a loud speaker or small flecks of light on a picture tube. Obviously, these effects will be apparent at all frequencies.

The shot effect in tubes is due to the fact that electron emission from a cathode is not at an absolutely uniform rate even when there are constant plate and screen voltages applied. Each electron jumps out of the cathode all by itself, and more of them may emerge at one instant than at another. If there is no space charge, and all emitted electrons go to the plate or screen as soon as emitted, the irregularity can cause considerable noise voltage. Many phototubes are operated in this manner.

Shot effect noise still is present, although greatly reduced, when the tube is operated with a space charge – as is the usual practice. Under this condition the tube noise is lessened by high mutual conductance or transconductance and by working the tube with plate current well under the maximum rating.

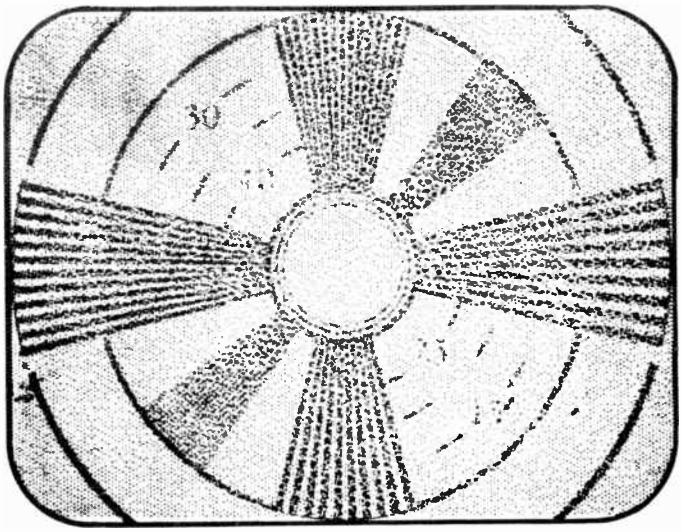


Fig. 78-4. A picture or pattern is filled with "snow" by thermal and tube noise whose strength exceeds that of the television signal.

7 Small screen current helps to reduce tube noise. Triodes tend to produce less tube noise than pentodes. If a given pentode tube is connected to operate as a triode, with plate and screen connected together, the noise voltage with the triode connection may be only one-half to one-fourth of that with the pentode connection. Cathode temperature lower than normal tends to increase the tube noise, because then there is less emission and a smaller space charge while a given plate current is drawn through the tube.

Tube noise may be especially troublesome in high-frequency broad band amplifiers wherein the grid circuit impedance is rather low. The low impedance reduces the signal voltage put onto the grid, and leaves the noise relatively larger compared to the signal.

Tube noise caused by irregularities of emission always is present to some extent in good tubes operated with suitable voltages on plate, screen, and grid. Should the plate voltage drop far enough below the screen voltage to allow secondary emission of electrons from the plate, or if there is secondary emission from any other cause, the tube noise will increase. Tubes which are not well evacuated, and which are "gassy", will produce strong noise voltages.

6 Tubes which are microphonic may also be the cause of noise voltages. A microphonic tube is one in which some of the elements may move in relation to one another when the tube is vibrated or jarred. The result from a loud speaker is an intermittent sound, at a pitch corresponding to the rate of element vibration, or a continued howl if the vibration is continual. The result on a picture tube may be bright flashes or else flickering or irregular displacement of the image. Television oscillator tubes, either r-f or sweep, cause the most trouble when they are microphonic. Microphonic audio amplifier tubes cause the characteristic sounds from the speaker.

The test for a microphonic tube is light tapping with the eraser on a pencil or with the end of a non-metallic alignment tool. When you tap the tube that is microphonic the trouble will become more apparent at the picture tube or speaker. This test should be made with the receiver in its cabinet, because vibrations which cause the microphonic effect often are due to sound waves in air partially confined within the cabinet.

Noise voltages which originate in the first stage of an amplifying system or which are introduced by any means into the first stage may cause serious difficulty, while noise of equal amplitude originating in or introduced into following stages is relatively unimportant. How this comes about is illustrated by Fig. 78-5.

We are assuming that signal amplitude reaching the grid of the first amplifier tube is four times as great as noise voltage amplitude coming to this same grid, as shown by the voltage waves between antenna and first grid. Were the amplification of this tube to be eight times, both the signal and the noise would be equally amplified and would appear in the plate circuit as shown by waves a and b.

Supposing now that additional noise voltage appearing between the output of the first amplifier and the input to the second amplifier is of the same amplitude as noise voltage going into the first amplifier. This added noise amplitude would be as represented at c. It does not add a great deal to the original amplified noise, and total noise input to the second amplifier would be of the amplitude shown at d.

Signal to noise ratio at a and b is 4.00 to 1.00. Adding the same noise amplitude that went into the first amplifier reduced the ratio only to 3.56 to 1. If another equal noise voltage were added between the

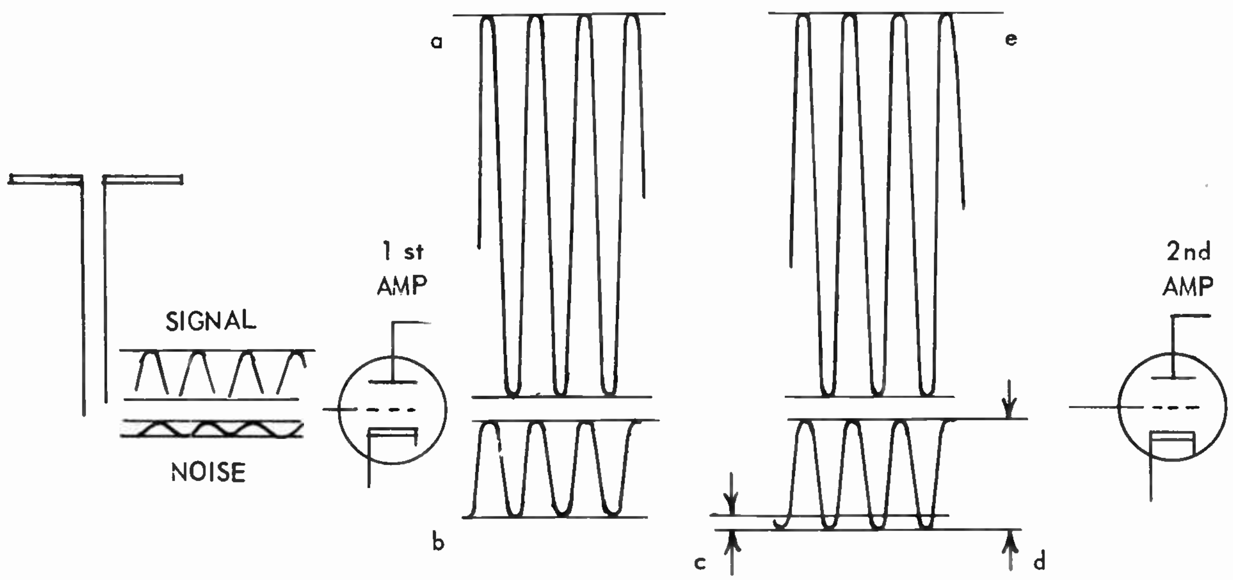


Fig. 78-5. A given noise voltage in an early amplifying stage may cause serious trouble, while the same noise amplitude in a later stage is much less important.

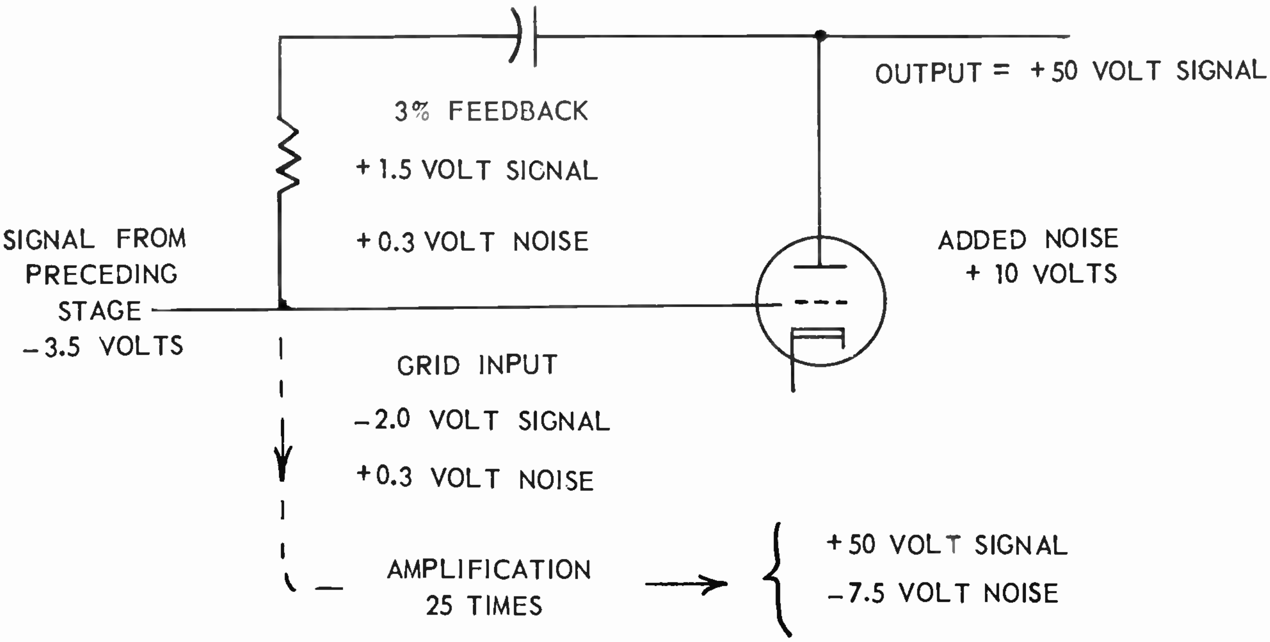


Fig. 78-6. How a degenerative feedback reduces the effect of noise originating in the plate circuit of the tube.

second amplifier and a third one, and were the amplification again 8 to 1 for both signal and noise, the ratio of signal to noise in the output of the second amplifier would drop to 3.51 to 1.00, which is a negligible change from the signal to noise ratio in the output of the first amplifier.

Noise which is introduced in the output of an amplifier may be greatly reduced by employing a degenerative feedback from plate to grid. Fig. 78-6 is an example in which certain definite voltage values are used merely to illustrate what may happen. Assume that in the plate output of the tube there is a 50-volt signal to which is added 10 volts of noise. We shall call the polarity or phase positive in order to identify inversions which occur.

With a 3 per cent degenerative feedback there is applied to the grid + 1.5 volts of signal and + 0.3 volt of noise. The signal from a preceding stage is - 3.5 volts, from which we subtract the + 1.5 volts of degenerative feedback to leave - 2.0 volts of signal on the grid. Amplification is assumed to be 25 times. There is inversion of polarities between grid and plate. Then the amplified output consists of + 50 volts of signal, as originally assumed, and of - 7.5 volts of noise. Polarity of this amplified noise voltage is negative, whereas the polarity of the original added noise is positive. Thus the noise voltage remaining for a following amplifier is reduced to + 2.5 volts.

We have assumed that no noise comes to our amplifier from the preceding stage. Noise that does actually come from that preceding stage will be amplified without any reduction due to the regeneration. However, noise produced in the output of the amplifier considered will be reduced by degenerative feedback from its plate to its grid.

5 *NOISE LIMITING FOR SYNC PULSES.* Strong pulses of noise voltage which add themselves to the composite television signal may act like sync pulses, and trigger the sweep oscillators at incorrect times. This will destroy synchronization while the noise continnes. In many television receivers there are means for limiting the strength of noise pulses. The noise limiting circuits usually are located somewhere between the video detector and signal input to the picture tube, or somewhere in the video amplifier. This location is chosen because signal takeoff for the sync section ordinarily is from some point in the video amplifier system.

Fig. 78-7 shows a noise clipper connected to the signal transfer line between two video amplifier tubes, to the output of an amplifier in which sync pulses are positive and picture signals negative. The clipper is a diode with its cathode grounded. The diode plate is connected to the signal line through capacitor C, and to ground through resistor R.

The positive sync pulses make the clipper plate positive and cause conduction through capacitor C, and the diode to ground, until the capacitor is charged to a voltage nearly equal to the sync pulse amplitude. Most of this charge is retained on the capacitor because capacitance at C and resistance at R are of such values as to provide a long time constant. The capacitor charge now makes the diode plate negative, and conduction drops to the small flow needed to maintain the capacitor charge. If there are noise voltage pulses more positive than the tips of the regular sync pulses the noise makes the clipper plate momentarily positive. Then there is conduction, and noise voltage passing on to the second amplifier is pulled down to a level only slightly above that of the sync pulse peaks.

Fig. 78-8 shows the circuit for a video amplifier operated as a noise limiter. Here the sync pulses are negative and the picture signals positive as they come from the video detector and go to the grid of the limiting amplifier. The amplifier is operated with very low plate voltage and a small negative bias, to al-

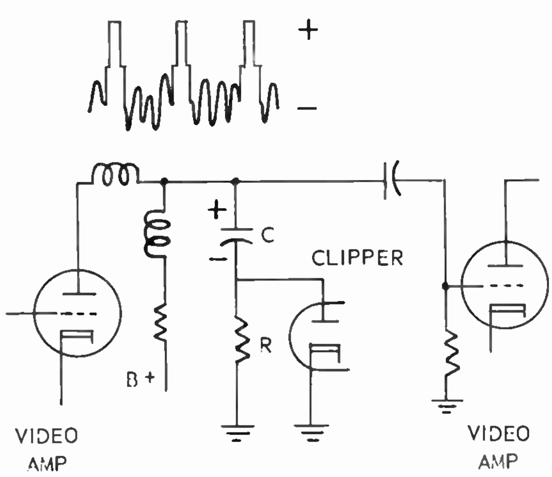


Fig. 78-7. A diode noise clipper between two video amplifiers.

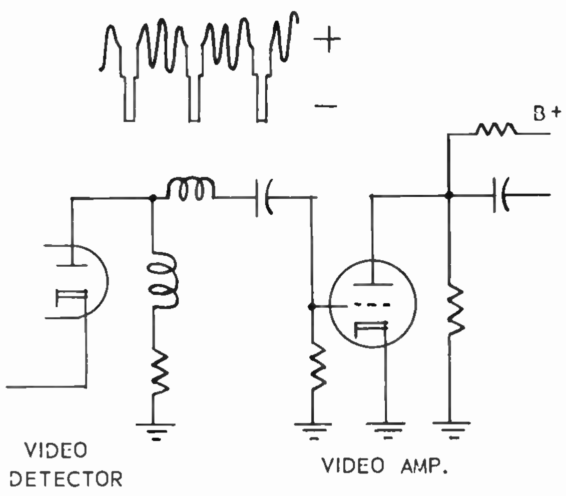


Fig. 78-8. A video amplifier operated as a noise limiter.

low about one-third of the normal plate current. Any strong noise voltage will cause a negative pulse of greater amplitude than the regular sync pulses. The increased negative voltage thus applied to the amplifier grid drives this tube beyond plate current cutoff. Thus the noise pulses are cut down to the level of the sync pulses.

Other and more elaborate methods may be used for limiting the effect of noise on the sync system. In some receivers the last i-f amplifier feeds not only the video detector in the usual manner, but also an additional i-f amplifier whose pass band is only about 0.1 mc, centered at the video intermediate frequency where the low-frequency sync pulses appear. This additional narrow-band amplifier then rejects all noise frequencies that are appreciably higher than the horizontal sync frequency. Its output goes to a separate diode detector from which the sync pulses pass to the sync section and the sweep amplifiers.

INTERFERENCE

You will recall that we originally classified interference as consisting of undesired voltages at a radio frequency and having some regular pattern, such as a steady alternating frequency or a signal frequency. We must remember that radio frequencies include all those from about 30 kilocycles up to hundreds of megacycles, and that interference may be anywhere in this wide range.

In the discussion to follow we shall explain almost every kind of r-f interference except the few kinds which we talked about in connection with traps used in i-f amplifier systems. These few include adjacent channel video and sound frequencies, and accompanying sound frequencies. The other kinds of r-f interference are most bothersome in localities where signal strength is low and receivers have to be operated with high gain, also where there are great numbers of receivers close together, and in areas close to radio, f-m, and television transmitters of any kind.

As a general rule the interference signals are picked up by the antenna or the transmission line. By the combined action of the r-f oscillator and the mixer of the receiver these interference signals then are converted to frequencies which may be amplified by the i-f amplifier of the receiver. If the interference cannot cause a frequency to which the i-f amplifier responds it cannot be classed as interference at all, for it will not get through this amplifier to the video detector and the sound system. Our first step in determining what signal frequencies can cause interference is to consider the response of the i-f amplifier system.

Just what frequencies can cause interference will depend on two factors. The first factor is the particular intermediate frequency employed in the receiver on which you are working. There will be a different set of interference frequencies for every different intermediate frequency which may be employed. For all following examples we shall assume that the receiver uses a video intermediate frequency of 25.75 mc and a sound intermediate of 21.25 mc. We could use the newer standard video intermediate of 45.75 mc and sound intermediate of 41.25 mc, but the ones first mentioned have been used in a large number of receivers and they allow certain interferences which are either impossible or else become relatively unimportant with the newer standards. Should you wish to do so, it is not difficult to work out the possible interference frequencies for any other set of receiver intermediate frequencies.

The second factor which affects the frequencies which may cause interference is the range of frequencies covered by the channel to which the receiver is tuned. The interfering frequencies are different for every different channel. In our examples we shall assume that the receiver is tuned for channel 5. You may compute the interference frequencies for other channels without too much trouble.

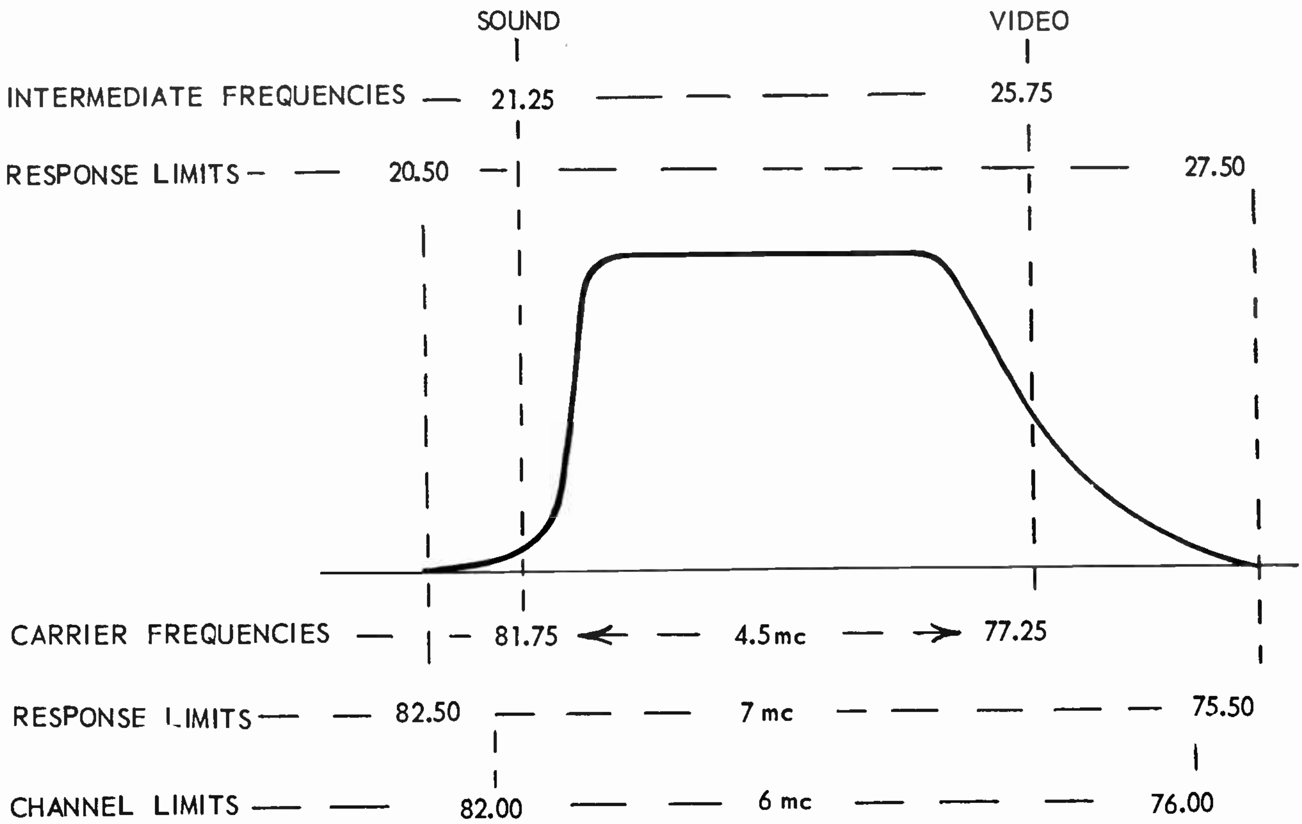


Fig. 78-9. The i-f amplifier response and the carrier-frequency response for channel 5 with a video intermediate frequency of 25.75 mc.

The frequency response of the i-f amplifier may be represented by the curve in Fig. 78-9. The response does not fall to zero until about 0.75 mc below the sound intermediate and about 1.75 mc above the video intermediate. Since the r-f oscillator frequency for our assumed intermediates must be 103 mc on channel 5, the response between the two zero points will be through carrier frequencies from about 0.50 mc below the channel limit on one side to about 0.50 mc above the channel limit on the other side. Thus we determine that any signal frequency between 75.50 mc and 82.50 mc may cause interference when our receiver is tuned to channel 5.

The first kind of interference to consider is that of image frequencies. Doubtless you remember that an image frequency is one that is just as far above the r-f oscillator frequency as the desired carrier frequency is below this oscillator frequency, and that the image frequency is higher than the desired carrier frequency by twice the intermediate frequency.

Fig. 78-10 illustrates some of the relations of image frequencies to our present investigation of interference. Diagram 1 shows the normal performance in the front end of a television receiver tuned to channel

5 and operating with our assumed intermediate frequencies. Subtracting the sound and video carrier frequencies from the r-f oscillator frequency gives the sound and video intermediate frequencies, because these intermediates are beat frequencies equal to the differences between oscillator and carrier frequencies.

Diagram 2 shows image frequencies which go from the antenna through the r-f amplifier to the mixer. The r-f oscillator frequency is, of course, unchanged. If you subtract this oscillator frequency from the image frequencies, the difference beat frequencies are exactly the same sound and video intermediate frequencies as obtained normally in diagram 1. If the r-f amplifier and antenna coupling are of tuned types they will be tuned to receive the regular carrier frequencies, and the images should be greatly attenuated. If, however, there is insufficient selectivity between antenna and mixer it will be possible for image frequencies to get through and cause interference, because they produce the intermediate frequencies to which the i-f amplifier is sensitive.

Diagram 3 of Fig. 78-10 shows image frequency limits as based on the response limits of our i-f amplifier system. The i-f response was shown in Fig. 78-9 as extending from 20.50 mc to 27.50 mc, and the

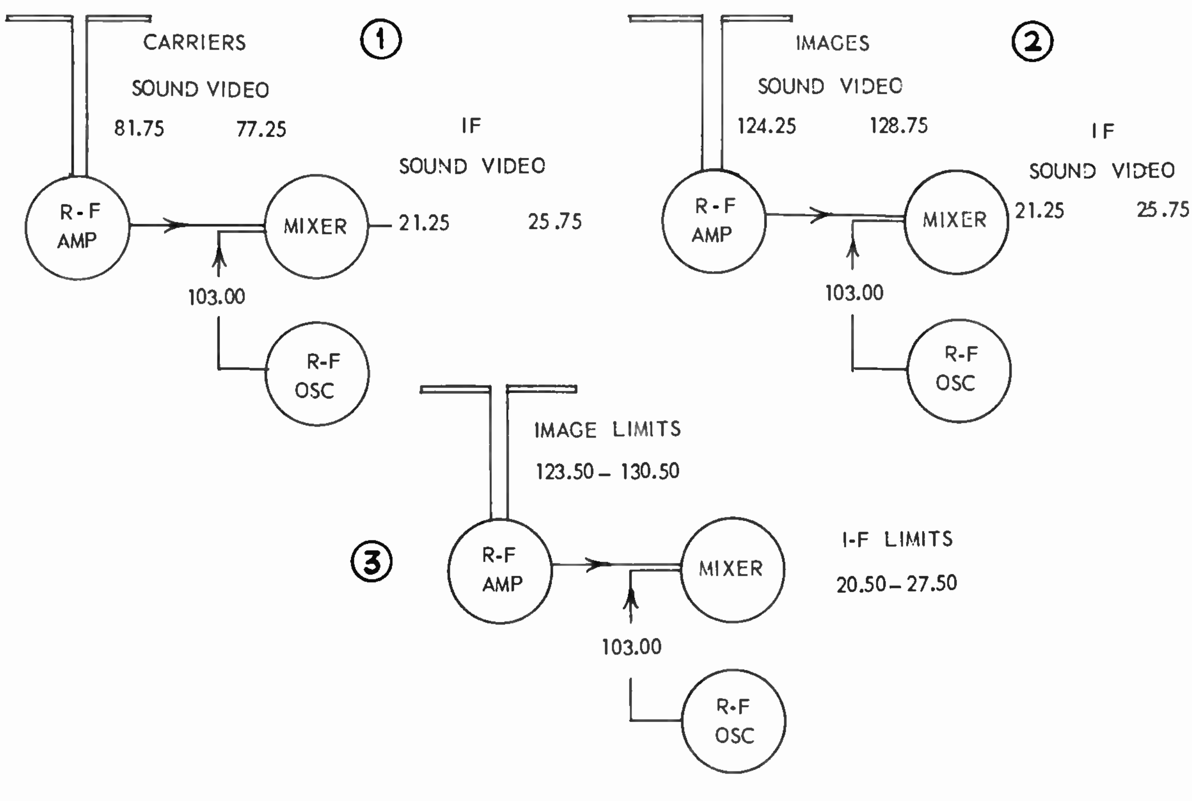


Fig. 78-10. Some examples of image frequencies.

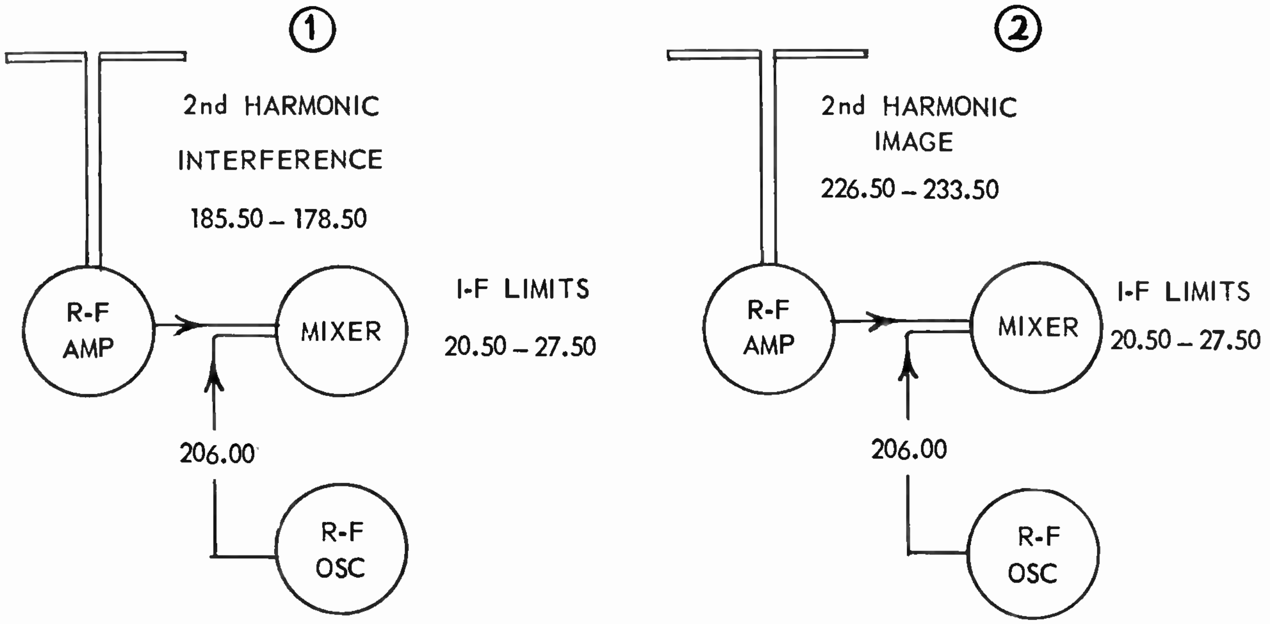


Fig. 78-11. How interference may result from oscillator second harmonic frequency.

image response limits will be proportionately extended as shown in Fig. 78-10. The image limits are equal to the sum of the r-f oscillator frequency plus the i-f amplifier limit frequencies.

Unfortunately our r-f oscillator cannot be depended upon to send to the mixer only the fundamental oscillator frequency. Nearly always it produces a second harmonic strong enough to beat with any other frequency reaching the mixer, and to thus form beat frequencies in the range for which the i-f amplifier is sensitive, provided the other frequencies are suitable for such action.

The first thing that may happen with the oscillator second harmonic is shown at 1 in Fig. 78-11. The second harmonic for our oscillator tuned to 103.00 mc for channel 5 is 206.00 mc. If you subtract from this second harmonic frequency the two frequencies marked "Second Harmonic Interference" the difference beat frequencies will be the limiting frequencies between which the i-f amplifier is sensitive. Consequently, signal frequencies within the range here shown as from 185.50 mc to 178.50 mc can cause interference which will pass through the i-f amplifier of the receiver.

Still another range of interference signals is based on image frequencies for the second harmonic oscillator frequency, as shown by diagram 2 of Fig. 78-11. Subtracting the oscillator second harmonic frequency of 206.00 mc from these image frequencies of 226.50 mc and 233.50 mc gives the difference beat frequencies that are the limits of sensitivity or of response for the i-f amplifier of our receiver.

If the receiver has a tuned antenna coupler or a tuned r-f amplifier with any reasonable selectivity there

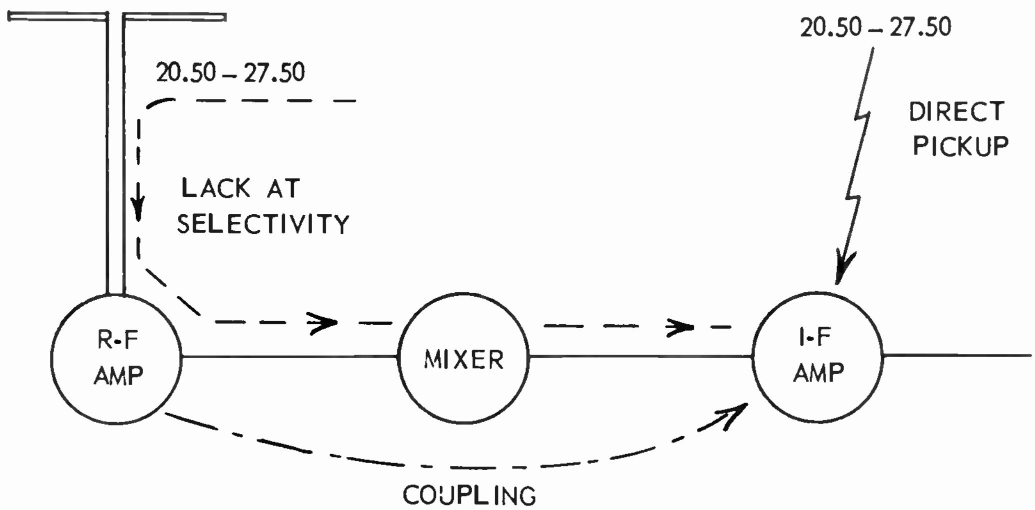


Fig. 78-12. Intermediate frequency interference may reach the amplifier in various ways.

should be little interference trouble from the signals shown in Fig. 78-11. The antenna coupler and r-f amplifier would be tuned for response in a normal carrier range of 32.50 mc to 75.50 mc, as shown in Fig. 78-9. But there are many receivers with untuned inputs, and as a result there may be and often actually is interference due to oscillator second harmonics and to images based on these harmonics.

(9) Although most of the interference signals get into the receiver by way of the antenna and transmission line it is quite possible to have pickup on circuit wiring in the i-f amplifier provided the interference frequency is within the response limits of this amplifier. In our present example these limits are 20.50 mc and 27.50 mc. Also, should strong interference at frequencies within these limits be picked up on the antenna or transmission line it could get through a non-selective r-f amplifier, or might be coupled from the antenna input circuits over into the i-f amplifier. These possibilities are indicated in Fig. 78-12.

The final possibility for entrance of interference signals to the receiver is direct pickup by the input circuits of the video amplifier. Since this amplifier is sensitive to frequencies all the way from around 30 cycles per second up to nearly 4.5 mc, any radio frequency below 4.5 mc might thus become interference. The interfering signal in this frequency range would have to be very strong.

There are simple tests for the last two classes of interference. If you remove the mixer tube or short its grid to ground and the interference effects continue, it is quite probable that pickup is in the i-f amplifier circuits. If you remove or disable the last i-f amplifier, the one preceding the video detector, and the effects of interference continue, there probably is pickup by the video amplifier circuits.

By looking back through the diagrams showing possible interference frequencies in Figs. 78-9 to 78-12 you will see that all of these frequencies, except that of the video amplifier, are determined in accordance with the frequency response or pass band of the i-f amplifier and the oscillator frequency for the channel considered. The oscillator frequency really depends on the i-f pass band, for it always is equal to the sum of the video intermediate frequency plus the video carrier frequency of the channel, or to the sum of the sound intermediate frequency and the sound carrier frequency of the channel considered.

If you know the intermediate frequencies used in a receiver, and look up the video and sound carrier frequencies for the channel on which there is interference, the possible interference frequencies may be determined as shown in the accompanying table. In the right-hand column are listed possible interference frequencies for a receiver using intermediates of 21.25 mc and 25.75 mc, and tuned to channel 5. These possible interference frequencies would be different for other intermediates and for other channels, but they could be computed as shown.

POSSIBLE INTERFERENCE FREQUENCIES

Band In Which Interference May Occur	How the Interference Ranges Are Computed	Ranges (mc) for Example Assumed
I-f pass band	From Sound intermediate - 0.75 mc To Video intermediate + 1.75 mc	20.50 to 27.50
Carrier frequencies	R-f osc'r freq. – i-f pass band	75.50 to 82.50
Image frequencies	R-f osc'r freq. + i-f pass band	123.50 to 130.50
Osc'r 2nd harmonic	(2 x osc'r freq.) – i-f pass band	178.50 to 185.50
Image, 2nd harmonic	(2 x osc'r freq.) + i-f pass band	226.50 to 233.50
Video amplifier	Same for all receivers and channels	0.03 to 4.5

The manner in which the interference ranges of our examples would be distributed in the r-f frequency spectrum is shown along the top bar of Fig. 78-13. Any of these frequencies might be those generated by the sources of interference, or they might be harmonics of the fundamental frequencies at which the sources are operating. That is, the interference frequency affecting our receiver could be a harmonic of the source frequency rather than its fundamental frequency.

If the receiver is being affected by a second harmonic, the fundamentals at which sources are operating are shown along the second bar of Fig. 78-13. If trouble results from a third harmonic, the fundamentals of the source would be as shown along the third bar. Along the bottom bar are combined all the frequencies which may be received as either fundamentals or second or third harmonics. From zero to 130 mc there are almost as many possible interference frequencies as there are frequencies on which interference is improbable or impossible.

SOURCES OF INTERFERENCE. Interfering signals at radio frequencies may come from any of a wide variety of sources, some of which are as follows.

Television broadcast transmitters operating on the channel to which the receiver is tuned or on other channels.

F-m broadcast transmitters giving services in either the entertainment or educational field.

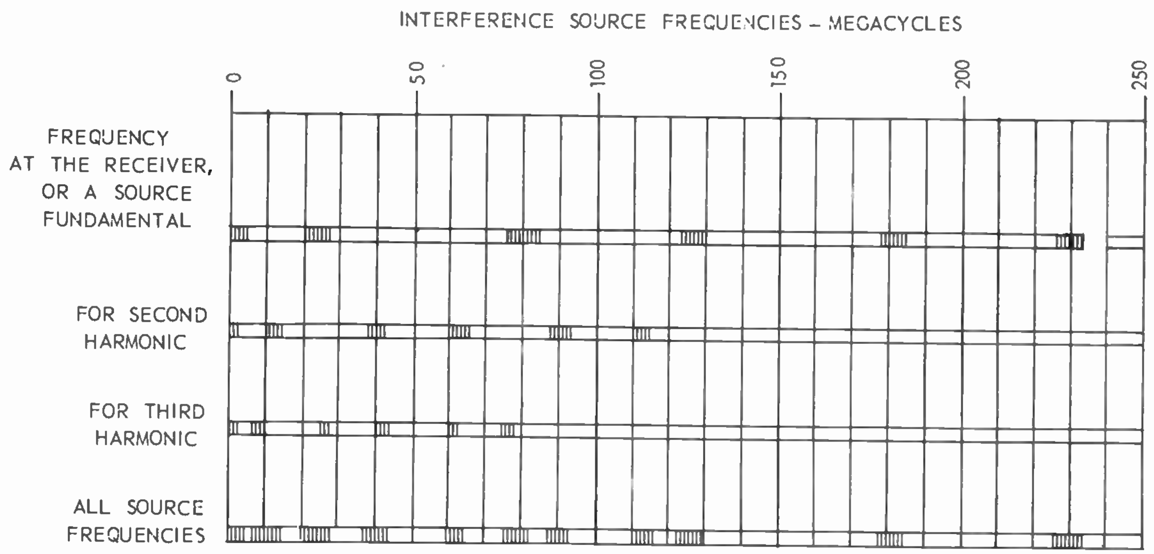


Fig. 78-13. Interference frequencies for receiver with video intermediate of 25.75 mc, tuned for channel 5. Bars show frequencies reaching the receiver on line for fundamentals, also fundamentals which may result in harmonics at the receiver.

Short-wave transmitters of all kinds. In this class are amateur radio stations and international short-wave broadcast stations.

Aviation services of many kinds; direction finding, landing beacons, airport control, civil aviation, and others.

Marine services such as direction finding, beacons, coastal harbor, ship to ship, fire, and others.

Police radio systems, state and local.

Newspapers; mobile transmitters, facsimile transmission, etc.

Weather reporting and forecasting.

Medical apparatus operating at radio frequencies.

Industrial devices operating at radio frequencies, as some types of heating and drying equipment, some sterilizing apparatus, etc.

There are many experimental and other more or less temporary services which may cause interference. The signals may be modulated with voice, music, code (dot and dash) or in any other manner, or they may be unmodulated. R-f signals which are being used for any kind of communication usually extend over a

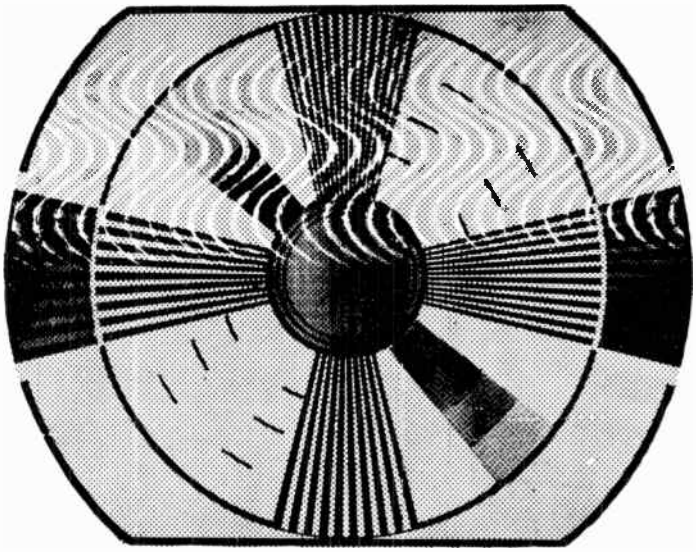


Fig. 78-14. Herringbone interference pattern from medical or industrial apparatus using radio frequencies at a 60-cycle pulse rate.

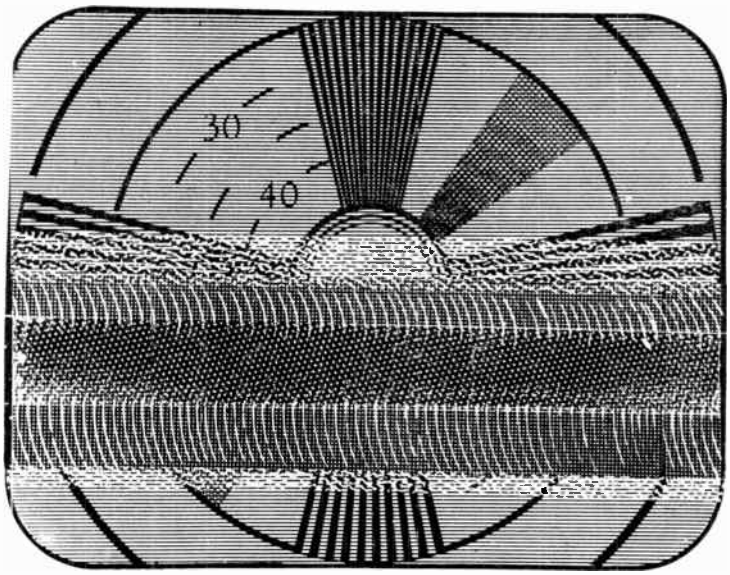


Fig. 78-15. Strong interference from medical diathermy apparatus.

a frequency band narrow enough to affect only one television channel, or two at the most. Interference from non-communication sources, such as those classed as medical and industrial, may extend over such a range of frequencies as to cause interference in many or all channels.

Industrial and medical apparatus which operates at radio frequencies causes interference patterns of the general type illustrated by Fig. 78-14. There is a more or less distinct herringbone extending horizontally all the way across the picture tube screen. The higher the interfering frequency the closer together will be the adjacent wavy lines of the pattern. Unless the interference frequency is continuous the pattern will consist of separated bands running horizontally. The example shown by the drawing would indicate that the interfering frequency comes on and off at a rate of 60 times per second, because there is one horizontal band during each vertical deflection period of 1/60 second, and the interference must be reaching a peak strength and then a minimum strength within each such period.

Two horizontal bands of the interference effect would indicate that the source is on and off, or goes through maximum and minimum, at a rate of 120 cycles per second, while three bands would indicate 180 cycle operation, and so on. If the interference is of low strength the pattern will be weak, but still will distort all or part of the picture. Extremely strong interference of this general class may cause almost complete blackout of parts of the picture, as illustrated by Fig. 78-15.

TELEVISION

LESSON NO. 79

REMEDIES FOR INTERFERENCE

Training Your Mind...

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind

C.T.I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON NO. 79

REMEDIES FOR INTERFERENCE

The best method for getting rid of the effects of any given kind of interference depends largely on the frequency at which the source of interference is operating. If we can determine this frequency, even approximately, our problem is greatly simplified. Oftentimes a knowledge of the probable frequency allows not only identifying the source of interference, but quite likely also its direction from the receiver.

To illustrate one method of determining an interference frequency we shall assume, for example, a frequency of 157,500 cycles or 157.5 kc per second. During one alternation of each interference cycle the grid-cathode voltage on the picture tube will be made less negative, and on the opposite alternation of interference it will be made more negative. The result on the screen of the picture tube will be one light-toned vertical line or bar followed by a dark-toned line or bar for each cycle of interference frequency. The effect is illustrated by Fig. 79-1.

We know that there are 15,750 horizontal picture lines per second, this being the horizontal scanning frequency. If there are 157,500 interference cycles per second it follows that during each horizontal scanning line there must be 10 interference cycles, since dividing 157,500 by 15,750 gives 10. Then, as in Fig. 79-1, there will appear along each horizontal line 10 light-toned sections with each followed by a dark-toned section of the line trace.

Every horizontal trace will consist of 10 pairs of light and dark sections. With the interference frequency an exact multiple of 15,750, as in our present example, all the horizontal traces will start on the same point of an interference cycle. Then all the horizontal traces will be alike, and the light and dark toned sections will be vertically in line to form vertical bars as illustrated.

The interference bars will be vertical on the picture tube screen only when the frequency interfering with a picture is some exact or integral multiple of 15,750. If division of the interference frequency gives some whole number plus a fraction left over, the light and dark bars will slope either to the left or to the right. This is because the fraction of an interference cycle left over at the end of each horizontal trace appears at the beginning of the next trace. All the interference cycles on this next trace are pushed a little to the right or the left. The same thing happens on a third horizontal trace, and on all following traces down to the bottom of the picture area. Then the interference pattern will slope either to the left, as in Fig. 79-2, or else to the right as in Fig. 79-3.

The interference frequency is computed by counting the pairs of interference bars, or else by counting either the light-toned bars or else the dark-toned bars, all the way across the pattern or picture. This

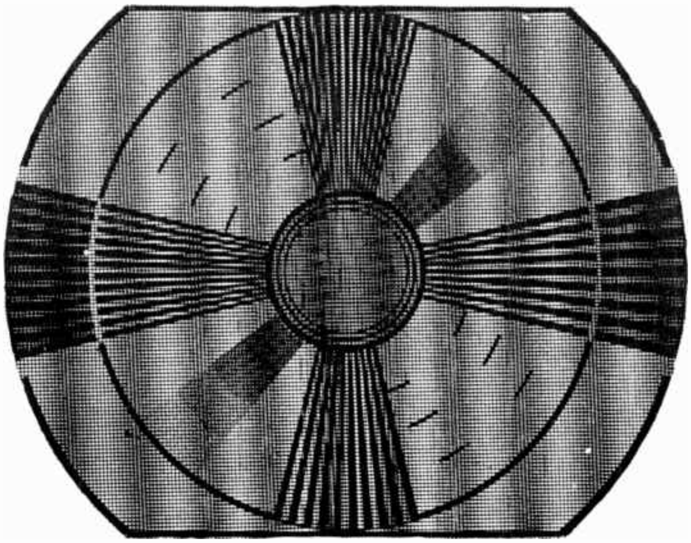


Fig. 79-1. Interference pattern from a frequency of 157,500 cycles per second.

number multiplied by 15,750 gives the approximate interference frequency. In Fig. 79-1 we have 10 pairs of interference bars which, multiplied by 15,750, gives 157,500 cycles per second for the interference. In Fig. 79-4 there are 20 pairs of interference bars from left to right across the pattern, and multiplying 20

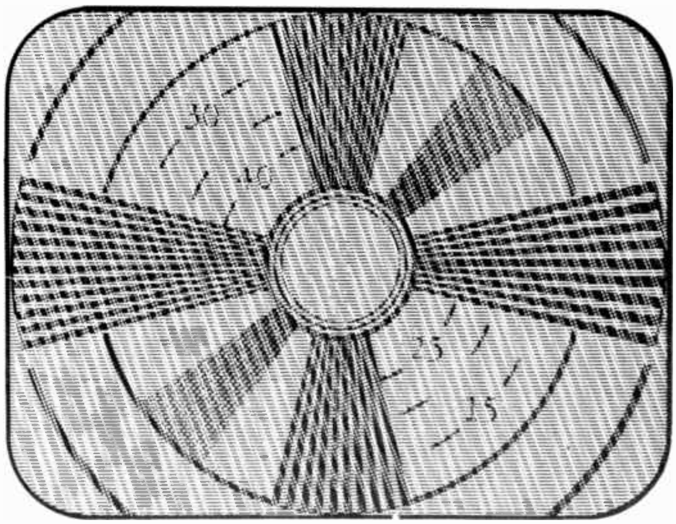


Fig. 79-2. A typical pattern of high-frequency interference.

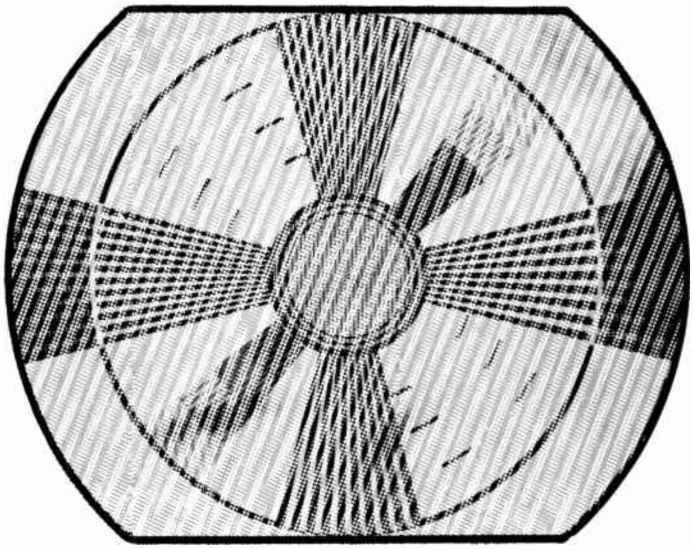


Fig. 79-3. The interference lines or bars may slope in either direction.

and 15,750 gives 315,000 cycles or 315 kc per second as the approximate interference frequency. Any other interference frequency may be determined similarly.

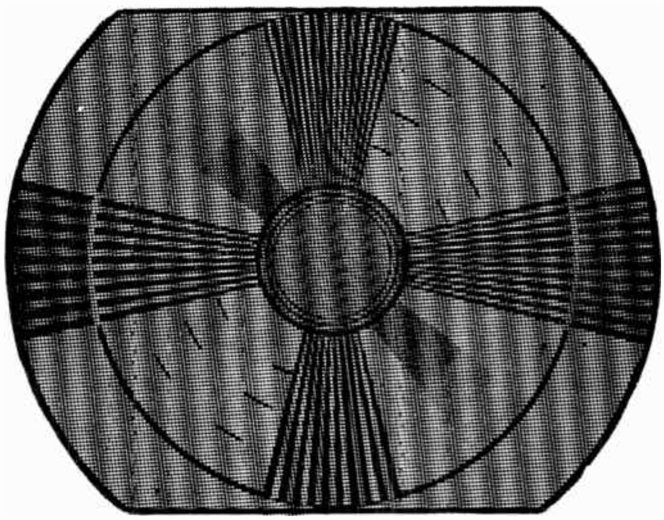


Fig. 79-4. Interference pattern from a frequency of 315 kilocycles.

It is often difficult to count the interference bars all the way across the width of the picture or pattern, because as a rule the bars shift their positions due to slight changes of interference frequency or of horizontal line frequency in the receiver. It is, however, relatively easy to measure the distance between centers of two adjacent light bars or two adjacent dark bars, then divide this separation distance into the width of the pattern or picture to determine the approximate number of interference cycles per horizontal line. This number of cycles then is multiplied by 15,750 to learn the approximate frequency in cycles per second of the interference.

The method of identifying an interference frequency by counting the number of cycles per horizontal line gives fairly accurate results when a television signal is being received along with the interference, and when this signal is holding the picture or pattern in horizontal synchronization. If you could cut off the television signal to leave only the interference, were the interference frequency not too far from an exact multiple of 15,750, and were the interference rather strong, it could synchronize the horizontal sweep oscillator at a frequency corresponding to a multiple of the interference frequency. Then the interference bars would remain vertical and steady, but the sweep rate would depend on the interference rather than being 15,750 times per second as from a television signal.

When the r-f interference frequency is constant or very nearly constant it is characterized by diagonal or vertical bars or lines which may be wide or very narrow. The bars may shift sideways slowly or rapidly, but they remain straight or have only moderate curvature.

If the interference frequency is shifting above and below some average value at a greater or less rate, but is remaining close to the same average, the effect may be a sort of herringbone pattern as illustrated by Fig. 79-5. The bars slope first one way and then the opposite way between top and bottom of the

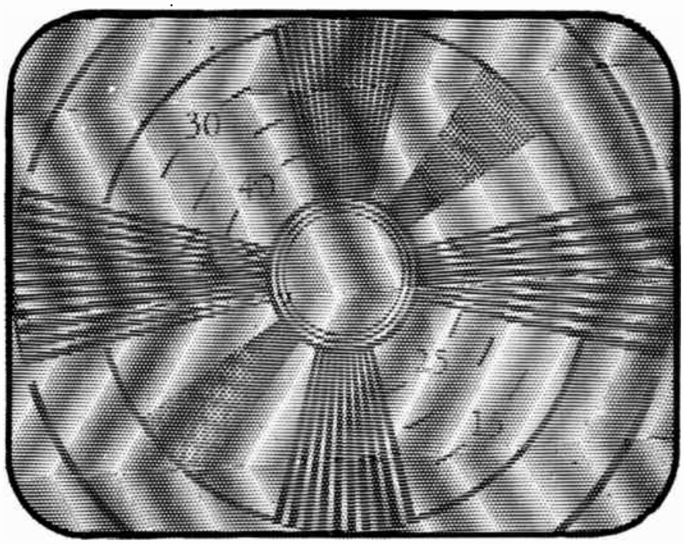


Fig. 79-5. Herringbone pattern as from low-frequency interference.

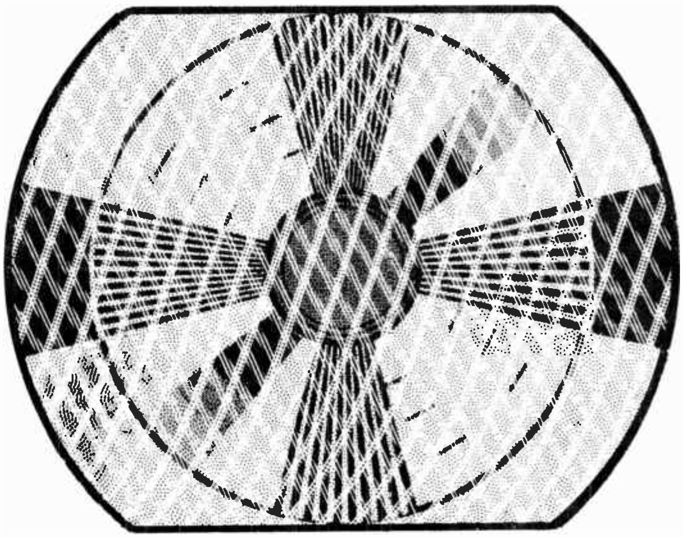


Fig. 79-6. A crosshatch interterence pattern.

picture or pattern area. Any pattern of zig-zag form from top to bottom, or any which consists of curves that bend back and forth from top to bottom, usually is referred to as a herringbone pattern. We looked at a narrow herringbone pattern when discussing industrial interference and interference from medical apparatus.

When the interference consists of two radio frequencies, or when there is rapid increase and decrease of the interference frequency, the effect may be described as a crosshatch pattern. Such a pattern will have lines or bars at two different slopes, and criss-crossing each other about as shown by Fig. 79-6.

Herringbone and crosshatch patterns may result from modulation of the r-f interference. Modulation also causes single sets of interference lines to weave back and forth, or to ripple in regular or irregular short curves along the length of the lines. The modulation may be of either the amplitude or frequency type, and may be carrying voice, code, or any other kind of signals. The pattern will change with every change of modulation strength or modulation frequency.

When interference bars run horizontally across the picture or pattern, rather than vertically or at a slant, the number of horizontal bars is related to the vertical scanning frequency of 60 cycles per second in the same way that vertical or slanting bars are related to the horizontal scanning frequency. That is, multiplying the number of light or dark horizontal bars by 60 gives the interference frequency in cycles per second.

Horizontal bars of the general type illustrated by Fig. 79-7 result from interference frequencies of 60 cycles or more, but less than 15,750 cycles, or they result from frequencies somewhere between the vertical and horizontal scanning rates. At any frequencies within this range one cycle of interference takes up

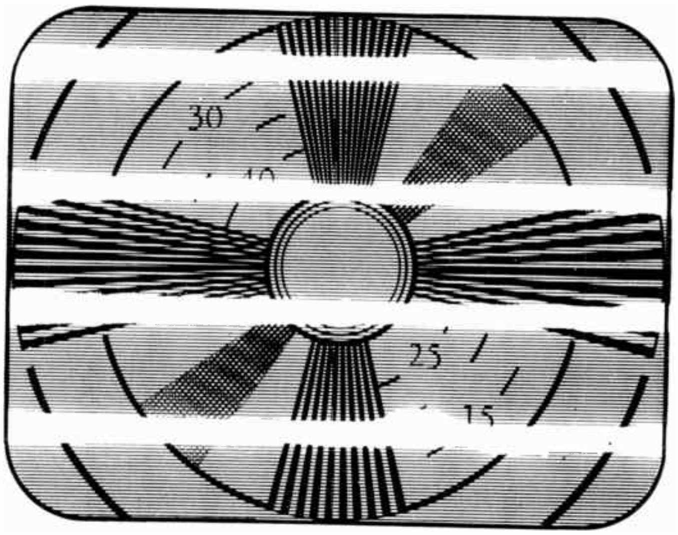


Fig. 79-7. Interference pattern produced by a frequency of about 255 cycles per second.

a time which is more than the period for one horizontal scanning line. Therefore, the effect of any such interference frequency will continue all the way across one or more horizontal scanning lines, and it must show up as horizontal bars.

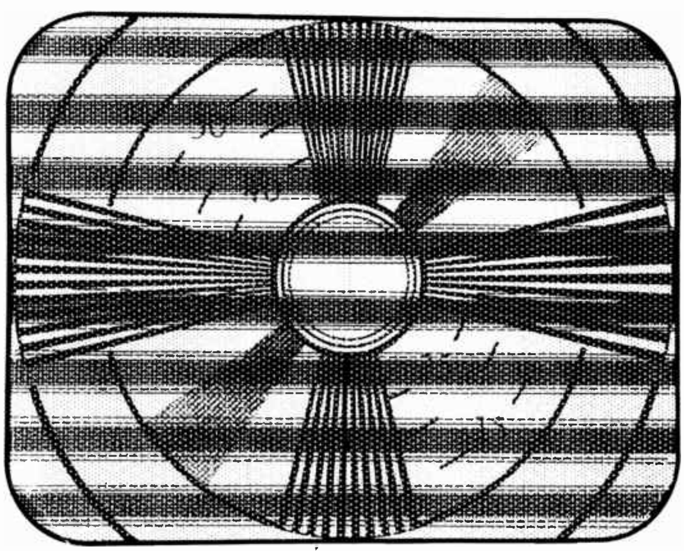


Fig. 79-8. Interference at about 480 cycles per second.

In order to have a clearly visible effect or to produce bars of enough width to be seen, each cycle of interference frequency must continue during at least two horizontal trace periods or must cover a height of two trace lines on the screen. To extend over two lines the frequency would have to be 7,875 cycles per second, which is one half the horizontal scanning frequency. The lower the interference frequency or the fewer its cycles per second the greater will be the number of horizontal scanning lines affected by each cycle, and the greater will be the width of the interference bars.

In Fig. 79-7 there are four light-toned bars alternating with four dark-toned bars, and there is part of another dark-toned bar. Each interference cycle is represented by one light bar and the adjacent dark bar. Here we have about 4¼ such cycles. Multiplying 4¼ by 60 gives the interference frequency as about 255 cycles per second for this particular example.

In Fig. 79-8 there are eight dark-toned bars, seven complete light-toned bars, and parts of other light toned bars at the top and bottom of the pattern. The number of interference cycles is apparently a little less than eight. Eight cycles multiplied by 60 would give an interference frequency of 480 cycles per second, so the actual interference frequency is slightly less than this value.

Interference bars that are horizontal usually are called sound bars, because they result from interference frequencies in the audio range or in the range of audible sound, at the video amplifier.

If the low-frequency (sound bar) interference is the modulation on a higher frequency which, by itself, would cause narrow vertical or diagonal bars, the effect on the picture tube screen will be somewhat as illustrated by Fig. 79-9. Here we have the same low-frequency horizontal bars as in Fig. 79-7, but in

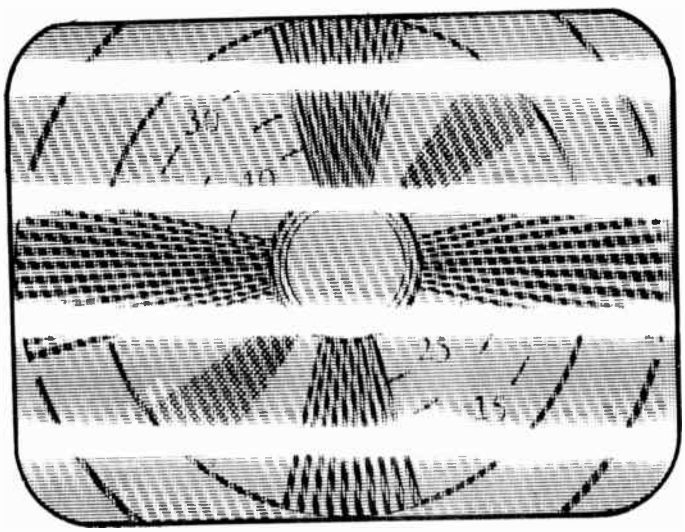


Fig. 79-9. Sound bars combined with a high-frequency interference pattern indicate a modulated radio frequency as the source.

addition there are narrow sloping bars which characterize radio-frequency interference. If the low-frequency modulation is quite strong the horizontal bars may be of herringbone form or possibly of the cross-hatch type.

Any low-frequency horizontal bars will remain stationary on the picture or pattern only when the low frequency is an integral multiple of 60, which is the vertical scanning rate. The multiple may be 120, 180, 240, or something like that. If the low frequency is a fractional multiple of 60 the horizontal bars will move up or down more or less rapidly on the screen when the picture or pattern is being held in vertical sync by a received television signal. If there is no signal, but only the low-frequency interference, the interference may synchronize the vertical sweep oscillator. Then the horizontal bars will remain stationary.

If the interference pattern seen on the picture tube indicates that there is modulation at an audio frequency it may be possible to listen to the modulation and thus identify the source of interference. In spite of the fact that television receiver sound systems are designed for reproduction of f-m signals most of these systems will have sufficient response to strong a-m signals to allow recognizing the words of station call letters and other voice announcements.

The sound takeoff of the television receiver is tuned for response at the sound intermediate frequency. The sound intermediate is originally formed from the sound carrier frequency. The r-f interference which is carrying audio modulation often may be heard quite clearly by adjusting a fine tuning control of the receiver to alter the r-f oscillator frequency enough to pass the interference through the receiver sound system. This usually is more successful on high-band channels than on low-band channels, because a fine tuning control can cause a wider shift of r-f oscillator frequency in the high-band channels. A much greater shift of frequency may be brought about by temporarily readjusting the alignment of the r-f oscillator for the channel in which interference is apparent. This latter method will, of course, require that the oscillator be returned to its correct alignment after the source of interference has been identified.

Without making any adjustments on the television receiver it often is possible to listen to announcements and other messages which are the modulation for high-frequency short-wave stations of all kinds. As a general rule these stations operate with amplitude modulation.

The source of f-m broadcast interference often may be identified by listening for the call letters and other announcements. F-m interference may affect television reception in channel 6, which is just below the f-m band in the frequency spectrum. Adjustment of the fine tuning control or temporary readjustment of the r-f oscillator frequency usually will bring in the f-m program quite distinctly.

The frequency of r-f interference may be determined by using an accurately calibrated r-f signal generator loosely coupled to the antenna input of the receiver. The coupling may be made from the high-side lead of the generator through a fixed capacitor of 10 to 25 mmf to an antenna terminal. There may be sufficient coupling with the output lead of the signal generator laid close to the transmission line where it connects to the antenna terminals of the receiver. When the signal generator is tuned to produce the same frequency as that of the interference there will occur a distinct change in the interference pattern. The interference frequency then is read from the tuning dial setting of the generator.

You must take care that the actual interference frequency is not a harmonic of the frequency to which the generator is tuned when the pattern is altered. After obtaining one indication, try tuning the signal generator to twice the frequency. If this produces the same result, try three times the original frequency. The actual frequency of interference is the frequency to which the generator may be tuned in order to produce the strongest change in the interference pattern on the picture tube screen. This identifies the frequency of interference coming to the antenna or transmission line of the receiver. We must not forget that this identified frequency may be a harmonic of the frequency at which the source of interference is operating.

As a final check it is advisable to tune the signal generator back and forth through the assumed interference frequency. If the same changes of interference pattern occur every time the generator output goes through the one frequency, you may assume that the changes really are due to the generator signal and not to something else which may reach the receiver input.

It is quite possible to have interference from some signal whose frequency is within the intermediate-frequency pass band of the receiver rather than at a carrier frequency. I-f interference may come from the antenna or transmission line through the tuner to the i-f amplifier, but it is just as likely or even more likely to be picked up in wiring and parts of the i-f amplifier. The effects as they appear on the picture tube are so nearly like those from carrier-frequency interference as to make it difficult to tell whether the trouble is from carrier or intermediate frequencies.

One of the easiest ways to determine which range of frequencies is causing the trouble is to change the adjustment of the fine tuning control. If the interference is at an intermediate frequency the fine tuning control will cause the pattern to shift and turn one way and the other, even with small changes of control

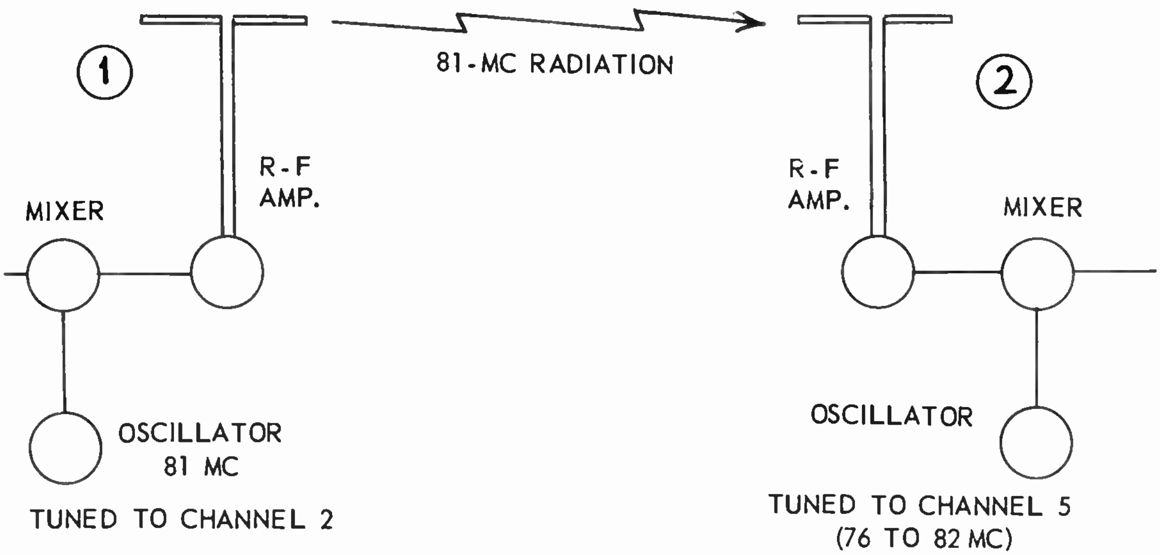


Fig. 79-10. The r-f oscillator of a television receiver may cause radiation and interference pickup by other nearby receivers.

adjustment in many cases. If the interference is at a carrier frequency, changes of the fine tuning control will have little or no effect on the interference pattern.

(10) Another check consists of tuning the receiver to different channels. If the interference effects are apparent on all channels it is quite probable that the interference is within the intermediate-frequency range of the receiver. Interference which appears on only one or two channels is at a carrier freqnency, or it is at a harmonic producing a carrier frequency from some lower frequency fundamental.

OSCILLATOR RADIATION. There are many television receivers in which voltage from their r-f oscillator goes not only to the mixer tube but also in the wrong direction through the r-f amplifier to the transmission line and antenna. Then there is radiation from the receiving antenna at the r-f oscillator frequency, and this radiation may be picked up by the antennas of any other nearby television receivers. If the frequency of the radiation from the first receiver is within the limits of a channel to which the second receiver happens to be tuned, there will be radio-frequency interference on the picture tube screen of the second receiver.

An example is illustrated by Fig. 79-10. Receiver 1 is tuned for reception on channel 2. We shall assume that its video intermediate frequency is 25.75 mc. The video carrier frequency for channel 2 is 55.25 mc. The r-f oscillator of this receiver must be operating at a frequency which is the sum of the video carrier and video intermediate, which is 81 mc. This 81-mc frequency is getting to the antenna of receiver 1, and from this antenna is radiated to the antenna of receiver 2. Receiver 2 is tuned for reception on Channel 5, whose frequency band is from 76 to 82 mc. This receiver is thus responsive to the 81-mc oscillator radiation from receiver 1. This radiation frequency will go through the r-f amplifier of receiver 2, and in the mixer will beat with the oscillator frequency of receiver 2. Since the resulting beat frequency is within the intermediate-frequency band pass of receiver 2 the beat will be amplified to produce r-f interference in this second receiver.

There are many other possible combinations of frequencies which may cause interference. Fig. 79-11 illustrates the combinations when the radiating receiver operates with a video intermediate frequency of 25.75 mc. At the top are represented the channels to which may be tuned the receiver that is interfered with. Next below are shown the r-f oscillator frequencies of the radiating receiver when tuned to certain channels. These channels include numbers 2, 3, 7, 8, and 9, with which there is interference respectively with another receiver tuned to channels 5, 6, 11, 12, and 13.

The offending receiver may radiate also at the second harmonic of its oscillator frequency. In this case there will be radiated interference when the receiver is tuned to channels 3, 4, or 5, as shown at the bottom of Fig. 79-11. The interference with another receiver then will be at the low end of channel 7, at the top of channel 8 and the bottom of channel 9, and near the middle of channel 12.

All our frequency computations are here based on the offending receiver having a video intermediate frequency of 25.75. mc. Any higher intermediate would cause a correspondingly high radiated frequency. If the video intermediate frequency were the newer standard of 45.75 mc none of the radiated frequencies would fall into another television channel. No change that could be made of the intermediate frequency of a receiver being interfered with would have any effect, because the radiated interference frequency would remain unchanged.

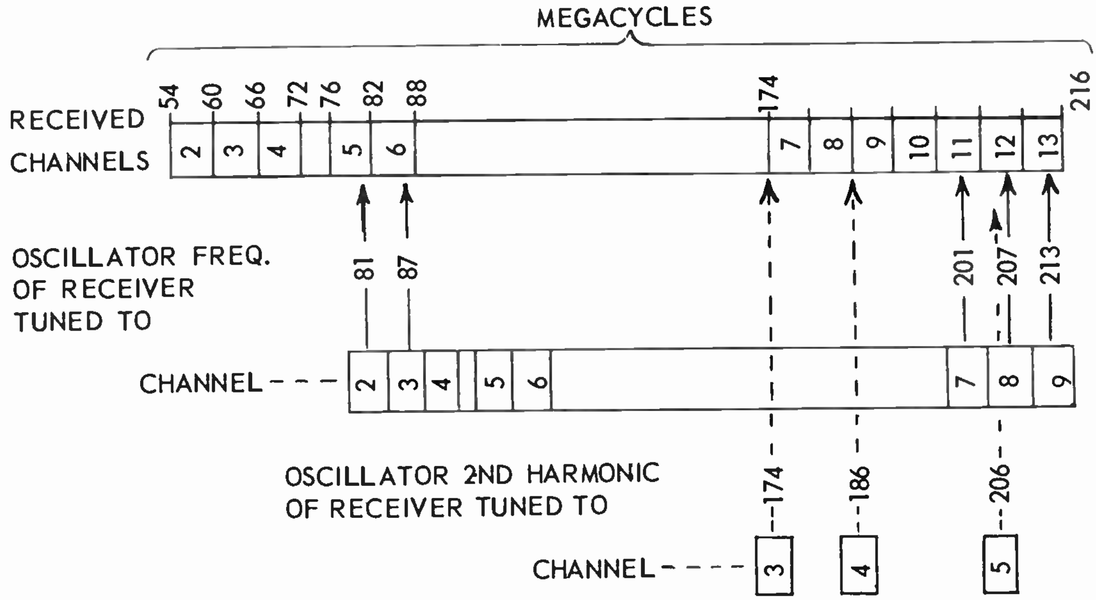


Fig. 79-11. Oscillator frequencies and their second harmonics which may be radiated from a receiver using a video intermediate of 25.75 mc when tuned to various channels.

Oscillator radiation may originate from f-m receivers and from any other high-frequency receivers as well as from television sets. The f-m broadcast band extends from 88 to 108 mc. When using an intermediate frequency of 10.7 mc the r-f oscillator frequencies of f-m broadcast receivers will be in the range between 98.1 mc and 118.1 mc, which does not extend into any of the television channel frequencies. But second harmonics of the higher f-m broadcast oscillator frequencies might cause interference in television channels 10 through 13.

4 Oscillator radiation is most troublesome where many receiving antennas are close together, also in localities where television signal field strength is rather weak and where receivers have to be operated at nearly maximum gain. Radiation is worse from receivers having antenna input to the grid of a triode r-f amplifier than when this amplifier is a grounded grid triode or a pentode. Oscillator voltage goes more easily from plate to grid in a triode than through a grounded grid triode or a pentode. Radiation is reduced by suitable design of tuner circuits. Recently designed tuners seldom allow oscillator radiation to exceed a strength of 10 microvolts, whereas some of the older types would allow radiation at several thousands of microvolts.

5 *DOUBLE CONVERSION INTERFERENCE.* Double conversion interference includes all the kinds which result when the r-f amplifier of the receiver interfered with is operating as a mixer or converter. Then there can be two frequency conversions, one in the f-f amplifier and another in the regular mixer or converter. This general class of interference may be called also r-f conversion interference, because the r-f amplifier is acting as a mixer or converter.

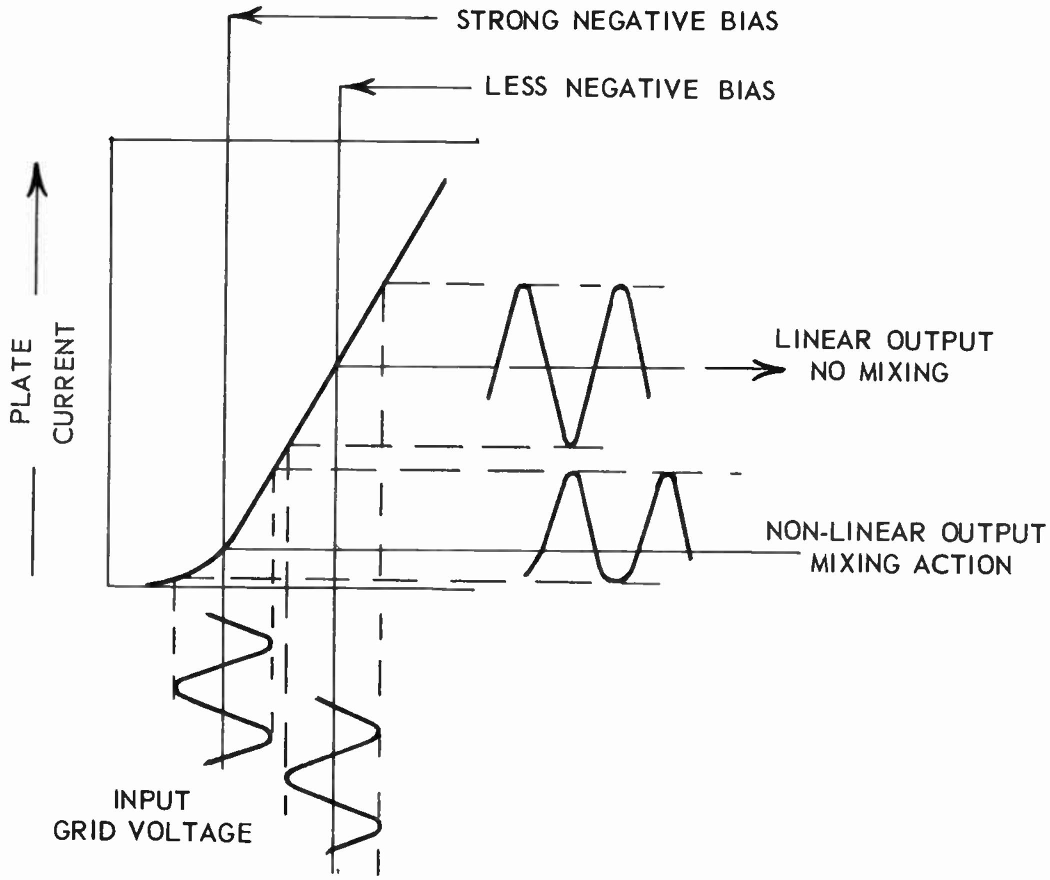


Fig. 79-12. An r-f amplifier may act also as a mixer when operated on the bend of its mutual characteristic.

A mixer, as doubtless you recall, is a rectifier, a detector, or any tube having a "non-linear" output in which variations of plate current and voltage do not follow the changes of applied grid voltage. A triode or pentode r-f amplifier tube will operate as a mixer when it is worked with such a combination of plate voltage and negative grid bias as to cause operation on the bend of the transfer characteristic.

Fig. 79-12 shows two alternating grid input voltages that are alike. One of these voltages swings back and forth around a strong negative bias voltage. One side of the alternating grid voltage is amplified more than the other side, and we have a non-linear output which can cause mixing action and a beat frequency whenever a second alternating voltage is applied to the grid. The other input grid voltage shown on the graph swings back and forth around a less negative bias. This brings the transfer action onto a straight portion of the characteristic curve. Then there is linear output, with both sides of the grid voltage equally amplified, and there will be no mixing action in case a second alternating voltage is applied to the grid.

The second alternating voltage on the r-f grid will be the r-f oscillator voltage when a small amount of this voltage goes the wrong way through the r-f tube and reaches its grid. This oscillator voltage then can beat in the r-f amplifier (working as a mixer) with any signal frequency coming from the antenna to the grid of the r-f amplifier. The output of the r-f amplifier then will contain sum and difference beat frequencies of the signal voltage and oscillator voltage. These frequencies go from the r-f amplifier to the regular mixer, and there they beat again with the r-f oscillator frequency. If the result of this second mixing or second conversion is within the i-f pass band of the receiver there will be interference which is amplified and delivered to picture tube or to the loud speaker.

An example of how all this may happen is illustrated by Fig. 79-13. Our receiver is tuned for reception on channel 5. The video carrier of channel 5 is at 77.25 mc, our assumed video intermediate frequency is 25.75 mc, so the r-f oscillator must be operating at the sum of these two frequencies, or at 103.00 mc. Part of the voltage at this oscillator frequency is getting to the grid of the r-f amplifier and thus to the antenna input circuit of the receiver.

To the antenna of the receiver is coming also an interference frequency of 181.25 mc. This is the video carrier frequency of a transmitter operating on channel 8. The interference frequency and the r-f oscillator frequency beat together in the r-f amplifier (operating as a mixer) to produce a difference frequency of 78.25 mc.

Now the interference beat frequency of 78.25 mc goes from the r-f amplifier to the regular mixer tube, where it beats with the r-f oscillator frequency of 103.00 mc which is being fed to the mixer as usual. The difference between 103.00 mc and 78.25 mc is 24.75 mc. This difference beat frequency goes from the mixer to the i-f amplifier system of the receiver. The frequency response of the i-f amplifier is shown by the curve at the right in the figure. The interference frequency at 24.75 mc is just one mc below the video i-f point on the response, and the interference falls right on top of the response curve, to get maximum amplification.

This particular variety of double conversion interference usually is called inter-channel interference, because it is due to a signal in one channel when the receiver is tuned to a different channel. It is apparent from Fig. 79-13 that the interference beat frequency which finally reaches the video i-f amplifier of the receiver might be of any value falling within the limits of the response curve. Working backward from this i-f response to the antenna we would find that any interference frequency between 178.50 and 185.50 mc would be a possible cause of interference. Transmission in channel 8 extends from about 180.50 to 185.75 mc, so everything except the sound in channel 8 could come through the i-f amplifier of our receiver tuned to channel 5, assuming a video intermediate of 25.75 mc for the receiver.

It would appear also that the sound from channel 7 would interfere with the receiver tuned to channel 5. Were the receiver tuned to channel 6 there could be interference from all except the sound from channel 9 and from all except the sound in channel 10.

Were the video intermediate frequency of our receiver to be 45.75 mc instead of 25.75 mc there could be interchannel interference when tuned to channel 4 (coming from transmission on channels 7 or 8), when tuned to channel 5 (from channel 11), and when tuned to channel 6 (from channel 13).

One remedy for double conversion interference of any kind is evident from examination of Fig. 79-12.

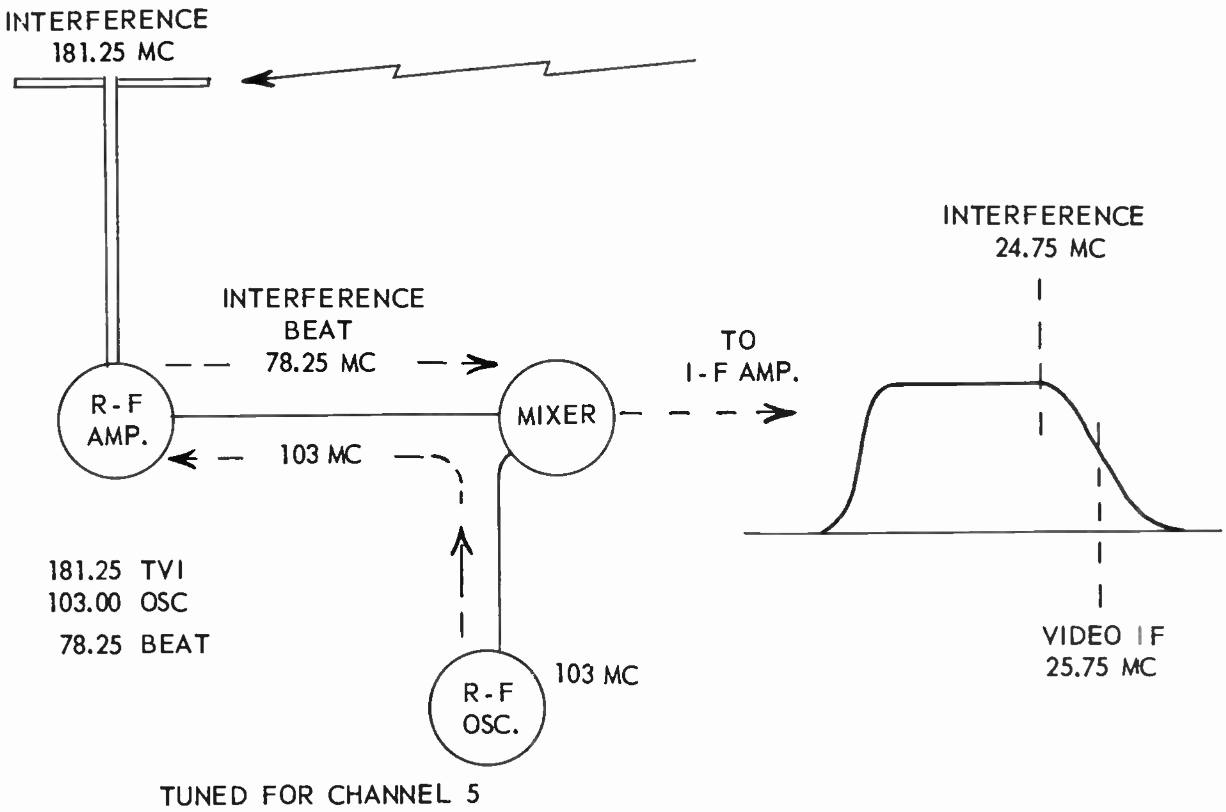


Fig. 79-13. An example of double conversion which results in interchannel interference.

If we operate the r-f amplifier on a straight portion of the transfer characteristic the output will be linear and there will be no mixing action in this tube. This could be accomplished by making the grid bias less negative or by increasing the plate voltage.

The r-f grid bias often is varied by the automatic gain control system. On strong signals the agc system makes the bias more negative, and there is likely to be double conversion and interchannel interference. On weak signals the agc system makes the bias less negative, and such interference becomes unlikely. When this kind of interference is very troublesome in a receiver having agc for the r-f amplifier, service technicians sometimes disconnect the r-f grid return from the agc bus and connect the return to a point of small fixed negative bias. If the r-f amplifier has cathode bias in addition to control from the agc system, the grid return may be connected to ground or to B-minus, whereupon the cathode bias usually insures operation on a straight portion of the characteristic. When such a change is made, it is quite likely that the mixer and the i-f amplifier may be overloaded when strong signals are amplified by the r-f amplifier.

OTHER TV CARRIER INTERFERENCES. Signals from transmission in an unwanted television channel may interfere with reception on another channel to which a receiver is tuned when the r-f oscillator of the receiver produces a strong second harmonic frequency. This is a form of interchannel interference, although it is not due to double conversion.

An example is illustrated by Fig. 79-14. The receiver is tuned for reception of channel 3, in which the video carrier frequency is 61.25 mc. With a video intermediate of 25.75 mc the r-f oscillator must operate at a fundamental frequency of 87.00 mc. The second harmonic from oscillator to mixer then would be 174.00 mc. The receiver antenna is picking up the sound carrier from channel 10, at 197.75 mc, and some of this carrier frequency is getting through to the mixer of the receiver. The difference between the carrier at 197-75 mc and the oscillator second harmonic at 174.00 mc is 23.75 mc, which is the beat frequency going from the mixer to the i-f amplifier. The interference beat of 23.75 mc falls on a part of the i-f response where there is maximum gain, and the interference will go to the picture tube.

This second harmonic interchannel interference can come from channels 8 or 9 when the receiver is tuned to channel 2, from channels 10 or 11 when tuned to channel 3, and from channels 12 or 13 when tuned to channel 4 - all this when the video intermediate of the receiver is 25.75 mc. With a video intermediate of 45.75 mc this kind of interference does not exist. With the higher intermediate the r-f oscillator frequency is increased by 20 mc, and the second harmonic by 40 mc. This, combined with the higher frequency of i-f response, throws the interference out of the response range.

Another kind of trouble, sometimes called co-channel interference, results from reception of signals

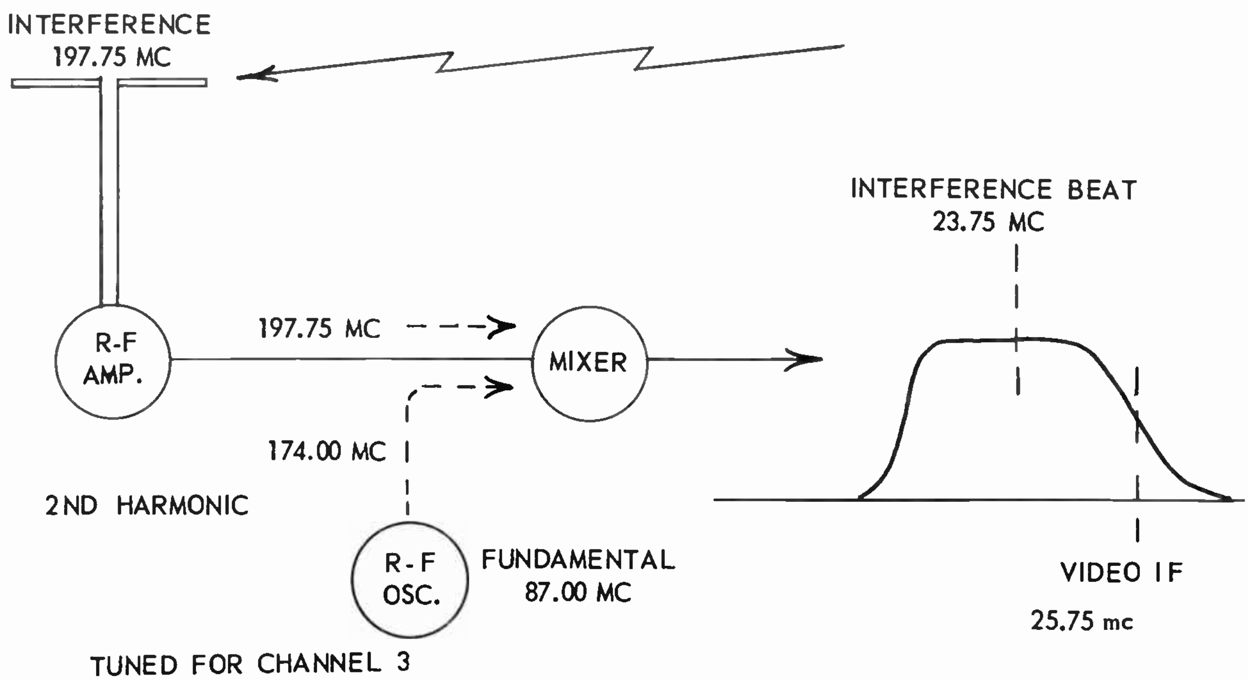


Fig. 79-14. Interchannel interference resulting from a second harmonic of the r-f oscillator frequency.

from two television transmitters on the same channel at the same time. This can happen when the receiving antenna is located between the two stations and is so far from each as to be in the outer fringe area of reception. The strongest parts of the carrier signals are the sync pulses, and as a consequence the interference effects are caused by horizontal or vertical sync pulses, or both kinds, acting together in the affected receiver.

6 If there is a slight difference between vertical sync frequencies of the two transmitters the effect will be a number of dark horizontal bars moving up or down on the picture tube screen. This may be called the Venetian blind effect. A slight difference between the two horizontal sync frequencies will cause one or more dark vertical bars to move from side to side on the picture tube screen. This may be called the windshield wiper effect.

Still another interference effect is due to what is called airplane flutter. This occurs when the transmitted signal that is being received on a direct line is also reflected from low flying planes passing over the receiving antenna. As a plane moves across the antenna position the phase of the reflected signal first adds to the total strength of received signal and then subtracts. The reproduced picture or pattern becomes alternately lighter and darker until the plane moves away. An effective automatic gain control often will almost entirely prevent this effect of airplane flutter.

It is possible for television image response to bring in interference from f-m broadcast stations when the video intermediate frequency of the television receiver is 25.75 mc or in this general range. To determine the possible image frequencies you would add the range of i-f response to the oscillator frequency for the channel to which tuned. It turns out that image responses which are within the f-m broadcast band may exist only when the television receiver is tuned to channels 2 or 3. With a television video intermediate frequency of 45.75 mc none of the image frequencies will be in the f-m broadcast band.

REMEDIES FOR INTERFERENCE. The first step in attempting to eliminate interference usually is to try orienting or turning the receiving antenna to some position which reduces or cuts out the interference while retaining desired signal pickup as much as possible. It may be necessary to substitute an antenna with a sharper directional pattern in order that orientation may be effective. Reflectors and directors usually help matters. Since the interference usually is weaker than desired signals it may help to substitute an antenna having greater gain. Then the receiver may be used with its own contrast or gain control at a setting low enough to cut off the interference while holding desired signals.

The effects of noise voltages reaching the antenna are lessened by using an antenna having greater gain, thus obtaining a better signal to noise ratio. Stacked antennas have marked ability to cut off noise impulses coming from above or below a horizontal line through the antenna position.

An unshielded transmission line may pick up any kind of interference, including all the r-f varieties. This pickup may be prevented by using coaxial line if the receiver is designed for 72- or 75-ohm input, or by using shielded twin conductor line for a 300-ohm balanced input. Horizontal runs of unshielded transmission line should be made as short as possible, and this type of line should be twisted about one full turn for every foot of run, both horizontal and vertical.

7 A test for transmission line pickup is made by temporarily disconnecting the line from the antenna while leaving it connected to the receiver. If the interference effects continue, all or most of the pickup is on the line rather than the antenna.

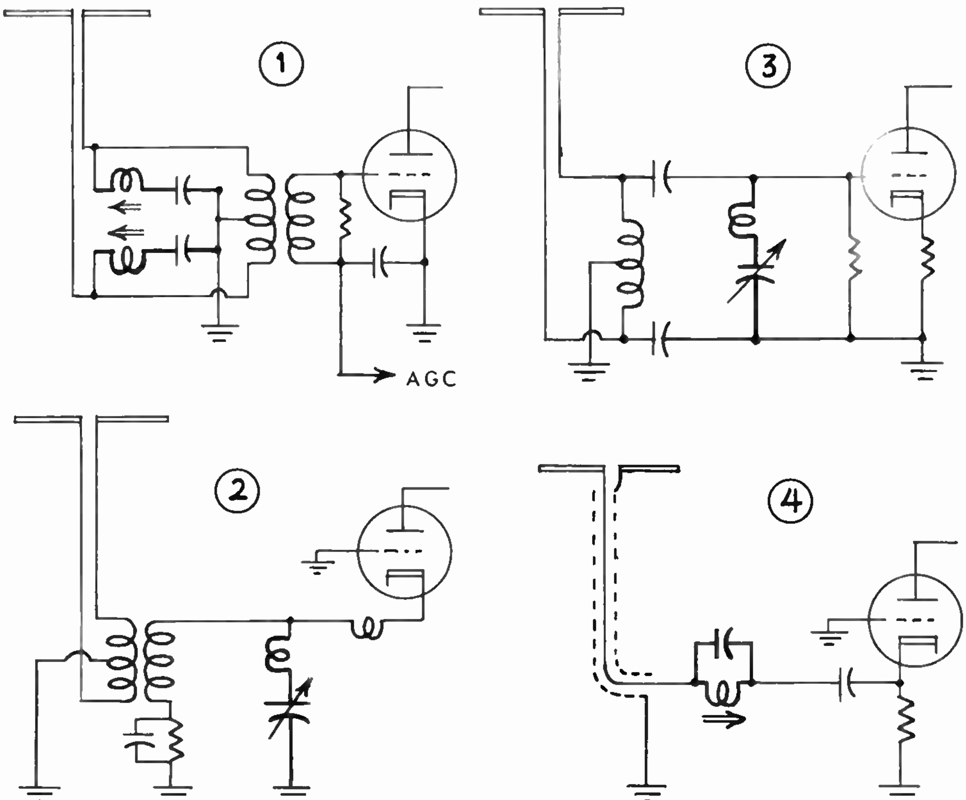


Fig. 79-15. Interference traps connected to antenna input circuits and r-f amplifier inputs.

A television booster or preamplifier connected between the transmission line and receiver may help to reduce interference. The booster increases the gain on the channel to which it and the receiver are tuned, and usually narrows the frequency response both above and below the range of channel frequencies.

ANTENNA TRAPS. Many receivers are equipped with adjustably tuned resonant trap circuits connected to some point in the antenna input for elimination of r-f interference frequencies. A few connections are illustrated by Fig. 79-15. In diagram 1 there are series resonant traps from each side of the transmission line to ground. At 2 there is a series resonant trap from the cathode of the r-f amplifier to ground. Diagram 3 shows a series resonant trap from the grid of the r-f amplifier tube to ground. Series resonant traps may be connected also from the mixer grid to ground,or between the two sides of a transmission line in order to short circuit the interference before it can go on to the r-f amplifier. In diagram 4 a parallel resonant trap is in series with the central conductor of a coaxial transmission line.

Some of the antenna circuit traps have fixed ceramic or mica capacitors used with inductors tuned by means of an adjustable core. Other traps are made with fixed inductance and an adjustable air-dielectric capacitor for tuning. Any of these traps ordinarily are mounted on the tuner or where the transmission line comes into the receiver. Fig. 79-16 is a picture of a double trap of the series resonant type which is constructed with two small ceramic fixed capacitors and two inductors having adjustable powdered iron cores in each.

Any one trap or pair of traps may be tuned through a frequency range having a ratio of two or three to one. If a trap is to be installed in a receiver not originally fitted with one, it is necessary to know the interference frequency or the range of interference frequencies in order to make a suitable selection. Frequency ranges of antenna traps commonly available from supply houses are as follows.

6 to 14 mc.	R-f short-wave interference.
10 to 30 mc.	Short-wave, or any interference at the i-f response.
13 to 28 mc.	Same as preceding type.
27 to 54 mc.	Short-wave, and some intermediate frequencies.
41 to 47 mc.	For the sound i-f at 41.25 and video i-f at 45.75 mc.
54 to 108 mc.	Low-band television carriers, also for f-m broadcast.
80 to 110 mc.	For f-m broadcast interference.
108 to 216 mc.	For television high-band interference.

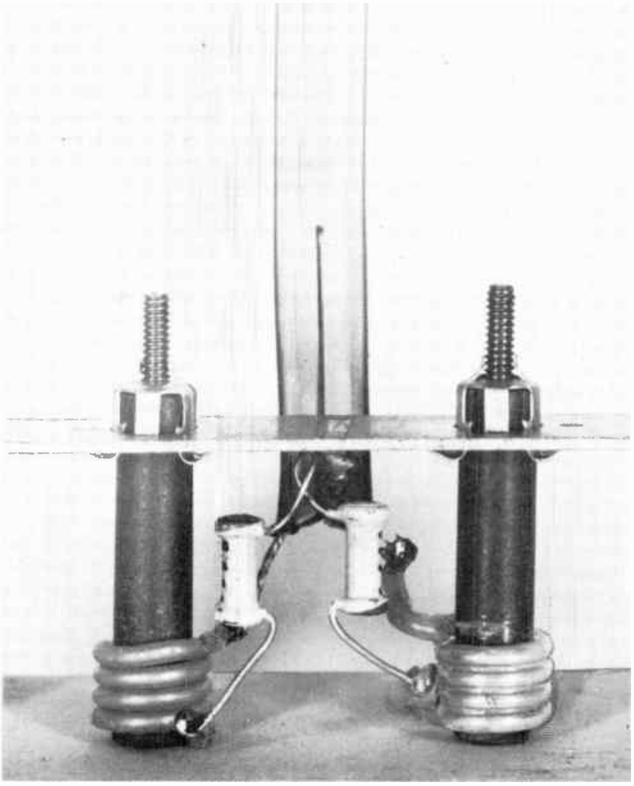


Fig. 79-16. A two-element antenna circuit trap of the series resonant type.

In our examination of television transmission frequencies which acted as interference it became evident that the interference usually originates from high-band frequencies, and affects receivers tuned to channels in the low band. To handle this general class of interference a trap tuned to a high-band interference frequency often is switched into the antenna input circuit as the tuner is changed to the low-band group of channels, then is switched off as the tuner is moved from the low-band to the high-band range. If this scheme is not used, and a trap is tuned to some high-band frequency, the trap must be switched off where-ever reception is wanted from the channel in which the trap is tuned.

Also available for installation between transmission line and antenna terminals of receivers are various high-pass filters. These filters consist of series capacitors and bypass inductors to ground, chosen of snch values as to pass quite freely all frequencies above 40 or 50 mc, while cutting off all lower frequencies. Thus the television carrier may go through the trap with but little attenuation, while r-f interference at lower frequencies is sharply reduced or eliminated.

In cases where interference is being picked up directly in the i-f amplifier, rather than coming through the antenna, a trap tunable to the interference frequency may be connected to the grid or plate circuit of the first or a following i-f amplifier tube. A trap of this kind would be similar to those used in i-f amplifiers for reduction of adjacent channel video or sound interference and accompanying channel sound, except that the added trap would be tuned to the particular interference frequency giving trouble. This frequency would have to be just outside the normal pass band of the amplifier system in order to avoid distortion of pictures or sound. The trap should be connected to the interstage coupler that is tuned closest to the trap frequency in order to have maximum rejection or absorption of the interference.

ADJUSTMENT OF ANTENNA TRAPS. An antenna trap may be adjusted for an approximately correct setting without the use of test instruments. At a time when the interference effect appears on the picture tube, adjust the fine tuning control or a continuously tunable channel selector to make the effect as bad as possible. If there is only a single trap element, adjust the trap core or capacitor for minimum interference effect, then return the fine tuning or the channel selector to their usual settings.

If there are two trap elements, as in Fig. 79-16, commence by adjusting the receiver tuning for maximum interference effect. Then turn both trap cores or both trap capacitors to the same or equal positions for a start. Next, turn the adjustment of either trap element a little ways in either direction, being sure to note which element is adjusted and which way it is adjusted. If this reduces the interference effect, continue turning the same adjustment in the same direction until there is minimum interference. If this first adjustment makes the interference worse, turn the core or capacitor in the opposite direction until there is minimum effect.

Now adjust the other trap element one way or the other to further reduce the interference effect if possible. Leave this second element as now set, go back to the first element, and try its adjustment to see whether any further improvement can be made. It is necessary to work back and forth between the two trap elements until both are adjusted for maximum reduction of the interference.

An antenna trap or traps may be adjusted by using an r-f signal generator which will tune to the interference frequency, and a vacuum tube voltmeter as an interference indicator. Proceed as follows.

1. Disconnect the transmission line at the receiver and connect the output of the r-f generator to the

antenna terminals of the set. The output impedance of the generator leads should match the input impedance of the receiver.

2. Connect the high side lead of the VTVM through a fixed capacitor of 0.01 mf or more to the high side of the video detector load resistor, or to any plate or any grid in the video amplifier, or to the picture tube control grid or cathode, depending on which of these latter elements is used for signal input. If the meter is connected to a plate circuit be sure that the blocking capacitor has a high enough voltage rating. Set the VTVM function control for measurement of a-c voltages.

3. Tune the signal generator to the interference frequency. Use audio modulation on the generator output, and set the output attenuator to deliver a strong signal.

4. Set the receiver channel selector to the channel in which interference appears.

5. Set the contrast control for normal reception, or else to the maximum position which would not cause distortion of a picture or pattern.

6. Adjust the fine tuning control as for normal reception.

7. Adjust a single trap element for minimum reading on the VTVM. If there are two trap elements, adjust them alternately (as previously explained) until both are set for minimum reading on the VTVM.

(9) If an antenna trap is fitted to the receiver, but is not required for interference reduction, do not leave the trap at a random adjustment. This might cut the response in one or more channels. Adjust the trap core for minimum inductance or the trap capacitor for minimum capacitance. This should throw the resonant frequency of a high-band trap out of the television range, or the resonant frequency of a low-band trap into the range between low and high television bands. Always check reception on all active channels before assuming that a trap will not cut a desired response.

STUBS FOR INTERFERENCE TRAPS. Resonant line stubs cut from pieces of transmission line often are used as traps for interference at the high-band television frequencies. A stub may be cut to a half-wavelength and shorted, as at the left in Fig. 79-17, or cut to a quarter-wavelength and left open, as at the right. Either of these stubs acts as a tuned series resonant circuit. When connected to the antenna terminals of the receiver, as illustrated, the interference frequency is short circuited through the stub and and is kept out of the receiver circuits. This method is most often used for double conversion interchannel interference and for interference resulting from oscillator second harmonics, but may be used also for f-m and other high-frequency interferences.

When a half-wave shorted stub is cut for the carrier frequency of some high-band channel, for the purpose of reducing interference on a low-band channel, the stub will greatly weaken or completely prevent reception when the receiver is tuned to the high-band channel. This reduction of the high-band signal may be lessened while still reducing interference on the low band by connecting a quarter-watt carbon resistor between the free ends of the stub instead of connecting the conductors together for a dead short.

It is best to commence with a shorting resistor of about 150 ohms, then decrease the shorting resistance if the interference is not reduced to a satisfactory degree. It may be necessary to go down to around 50 ohms, or even as low as 20 ohms to cut out the interference. Always use the greatest shorting resistance that reduces the interference effect.

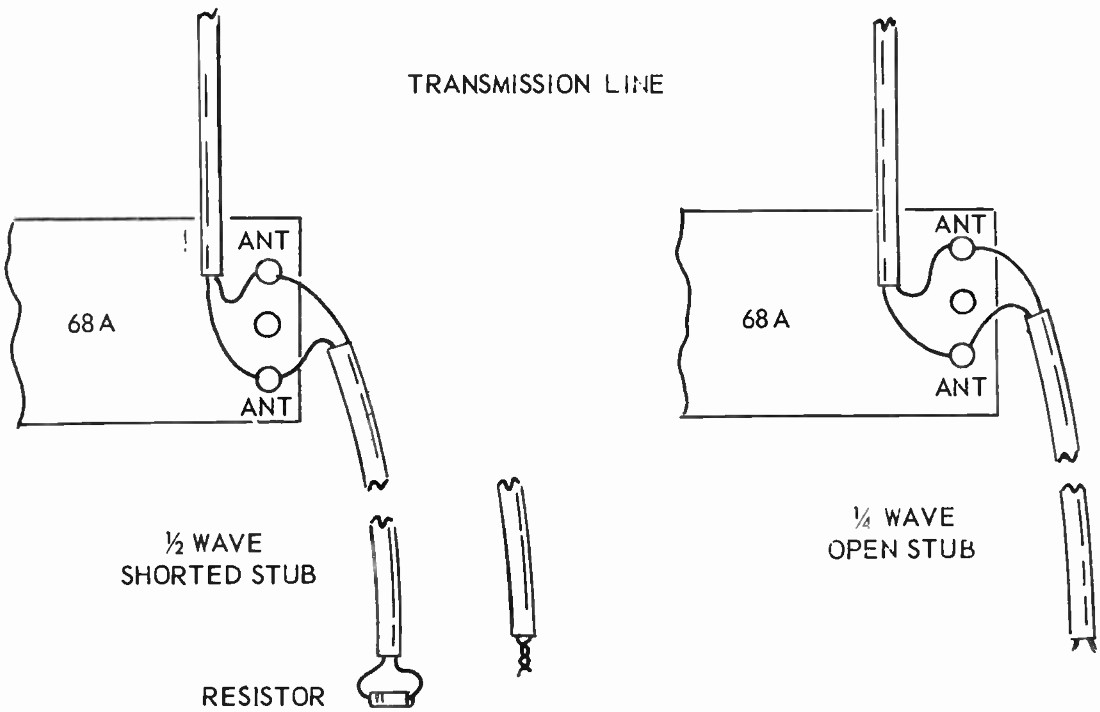


Fig. 79-17. Connections for resonant stubs used as interference traps.

If the interference is not at a frequency of a channel which it is desired to receive, the short should be made by twisting the line conductors together. This gives maximum attenuation of the interference frequency. If a direct short or a low-resistance short is used for a high-band frequency it will be necessary to connect the stub to the antenna terminals through a small knife-type double-pole switch. This switch can be closed for low-band reception and opened for reception on the high-band channel for which the stub is cut.

An interference stub may be made from 300-ohm twin conductor transmission line no matter what the type or impedance of the regular transmission line may be, and no matter what may be the input impedance of the receiver. The approximate length in inches for a half-wave shorted stub is found by dividing 4,900 by the frequency in megacycles. For a quarter-wave open stub the length is found by dividing 2,450 by the number of megacycles. Always commence with a piece of line a few inches longer than the computed amount, then cut off pieces about one-quarter inch long until there is maximum reduction of the interference. A shorting resistor is connected after the stub length has been determined by trial. Quarter-wave open stubs ordinarily are used only when the half-wave shorted stub would be inconveniently long, since the half-wave type will give better attenuation of the interference.

When installing an interference stub keep it well away from the chassis and from all other metal objects. Do not run the stub close to the transmission line. It is quite possible that an interference stub will upset the impedance match between receiver and transmission line to a greater or less extent.

TELEVISION

LESSON NO. 80

TELEVISION SOUND SYSTEMS

Training Your Mind...
C. T. I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind
C.T.I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON NO. 80

TELEVISION SOUND SYSTEMS

When first tracing the television signal through the receiver we progressed from the antenna to the picture tube grid-cathode circuit by way of the tuner, the i-f amplifier, and the video amplifier. Then we traced from the video amplifier through the sync and sweep sections to the deflection system for the picture tube. Now we shall go back and follow the sound portion of the signal as it travels the path from the sound takeoff to the loud speaker.

The television sound carrier is frequency modulated. The modulated carrier is changed to a frequency-modulated intermediate by the television tuner. Then, in a dual channel or split sound system shown at the top of Fig. 80-1, the i-f sound signal is taken from the tuner output or else from the first or second i-f stage to a separate i-f amplifier for sound. The output of this sound i-f amplifier goes to the sound demodulator, thence to the audio amplifier and loud speaker. In this dual channel sound system the sound intermediate frequency always is 4.5 mc lower than the video intermediate. For example, with a video intermediate of 25.75 mc the sound intermediate would be 21.25 mc.

1 The majority of recently designed television receivers do not use split sound, but employ the intercarrier sound system shown at the bottom of Fig. 80-1. The sound and video intermediate frequencies travel together from the tuner all the way through the television i-f amplifier to the video detector. This detector is a type of rectifier with non-linear output. Consequently, it acts as a mixer for the two intermediate frequencies and the detector output contains a beat frequency equal to the difference between the two intermediates. This difference always must be 4.5 mc, regardless of what intermediate frequencies may be used in the receiver. It always is 4.5 mc because the difference between video and sound carriers in all channels is 4.5 mc, and the difference between video and sound intermediates must likewise be 4.5 mc.

2 The 4.5-mc beat frequency in the detector output is frequency modulated with the sound or audio signals. It also is amplitude modulated with the video or picture signals and with the sync pulses. The sync frequencies are 15,750 cycles and 60 cycles per second. The video frequencies extend no higher than 4.0 mc. Therefore, to pick off the 4.5-mc sound-modulated signal we need only provide a takeoff circuit tuned quite sharply to 4.5 mc and thus leave behind the video and sync signals with their amplitude modulation.

The sharply tuned takeoff circuit is satisfactory because maximum frequency deviation for frequency modulation in television sound is only 25 kilocycles above and below the 4.5-mc center frequency, The sound signals will be carried within a frequency range between 4.475 mc and 4.525 mc, which is a deviation of 25 kc or 0.025 mc each way. To accomodate the full range of sound signals the frequency response of the take-off and of the following sound i-f amplifier need be only a little greater than this range. You will remember that maximum deviation for f-m broadcast signals is 75 kc.

Fig. 80-2 shows circuits for the portions of a dual channel sound system immediately following the mixer

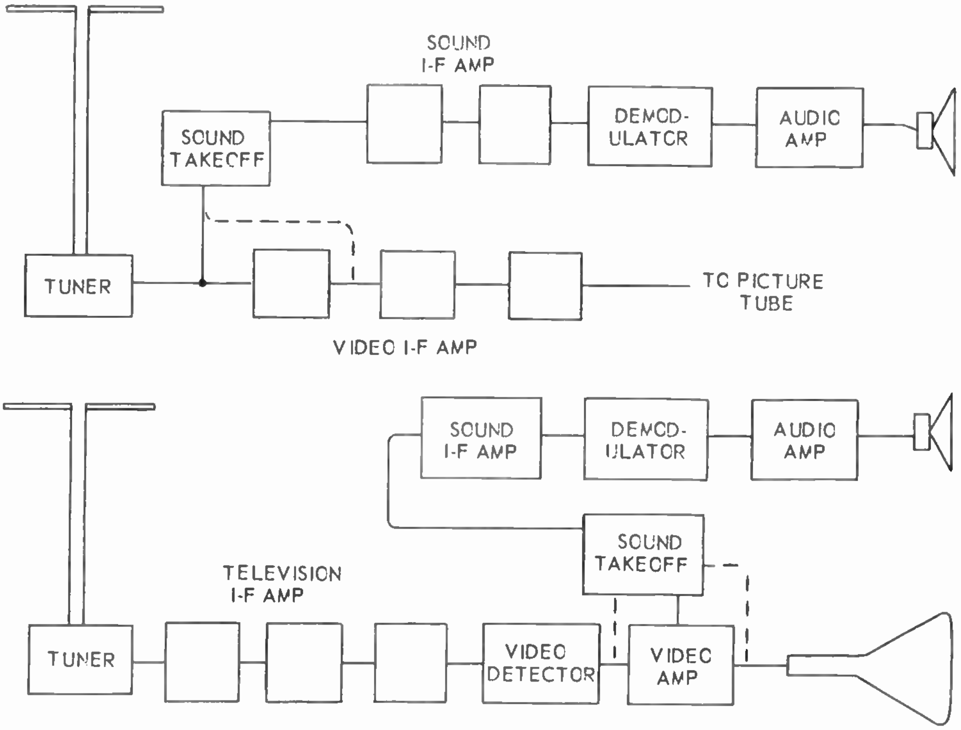


Fig. 80-1. The dual channel or split sound system (top) and the intercarrier sound system (bottom).

tube. Between the plate of the mixer and the grid of the first video i-f amplifier tube is a tuned impedance coupling coil. To this interstage coupling coil is coupled the sound takeoff coil. From a tap on the takeoff coil a lead goes to the grid of the first sound i-f amplifier. The interstage coupling coil is tuned by its own movable core to the frequency required for the video i-f passband. The takeoff coil is tuned by its separate movable core to the sound intermediate frequency, which might be 21.25 mc or 41.25 mc or whatever sound intermediate is used in the receiver. One style of double tuned coupler and takeoff is illustrated by Fig. 80-3. Many other equivalent circuits are in general use.

Except for the difference between operating frequencies, the television sound i-f amplifier with a dual channel system is like the i-f amplifier for an f-m broadcast receiver. There may be one or more amplifying stages and a limiter stage. The sound demodulator usually is a discriminator, although it could be a ratio detector. The output of the demodulator goes through a volume control to an audio amplifier. We are familiar with all these parts of the dual channel sound system from having studied them in connection with f-m broadcast receivers.

Fig. 80-4 shows typical circuit connections between a video detector and a sound ratio detector for an intercarrier sound system. The output of the video detector goes to the grid of a video amplifier. In the plate circuit of the video amplifier is one winding of a sound takeoff transformer. Both windings of this transformer are tuned by movable cores to a frequency of 4.5 mc. Thus the takeoff transformer picks from the video amplifier plate circuit the 4.5-mc beat frequency which is frequency-modulated with the sound signals of the television program.

The secondary winding of the intercarrier takeoff transformer is connected between the grid and cathode of a driver amplifier. The driver takes the place of the sound i-f amplifier section of a dual channed system. We need only one stage of driver amplification because the signal can be brought up to any desired strength in the television i-f amplifying stages which are between the tuner and the video detector. The f-m output signal from the driver tube goes through a transformer to the sound demodulator. The demodulator usually is a ratio detector in intercarrier sound systems, although it could be a discriminator were an extra tube to be used as a limiter or were the driver operated to have limiting action.

The sound takeoff transfomer acts as a tuned parallel resonant circuit in series with the output of the video amplifier. The high impedance of this circuit at 4.5 mc, combined with the energy extracted at this frequency from the amplifier plate circuit, prevents the sound signals from going to the picture tube. Thus the takeoff transformer acts much like a trap for sound signals. There are no accompanying sound traps in the i-f amplifier of a receiver using intercarrier sound, because the sound intermediate frequency must reach the video detector, where it beats with the video intermediate to produce the 4.5-mc f-m sound signal.

It is apparent that television receiver sound sections, to and including the demodulators, are not essen-

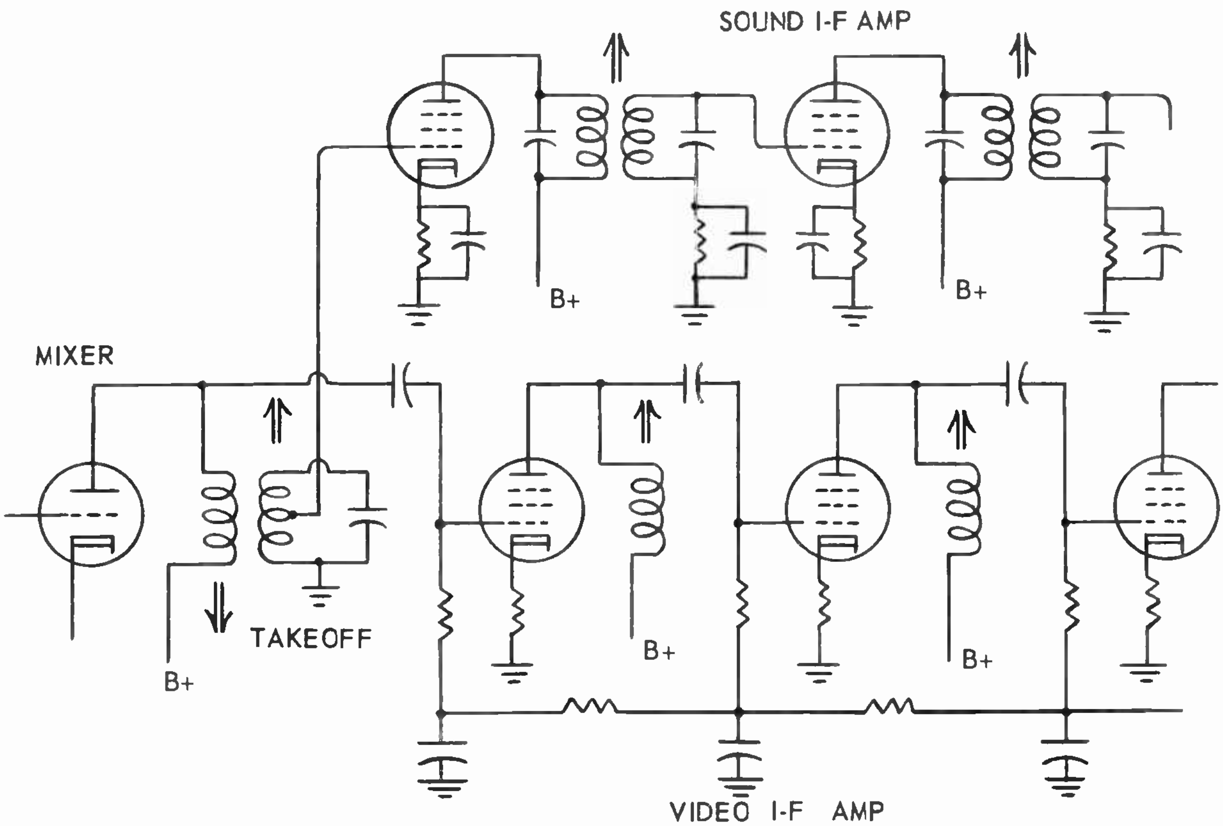


Fig. 80-2. Typical circuits which immediately follow the mixer tube for a dual channel or split sound system.

Fig. 80-3. A sound takeoff coupler used with a dual channel sound system.

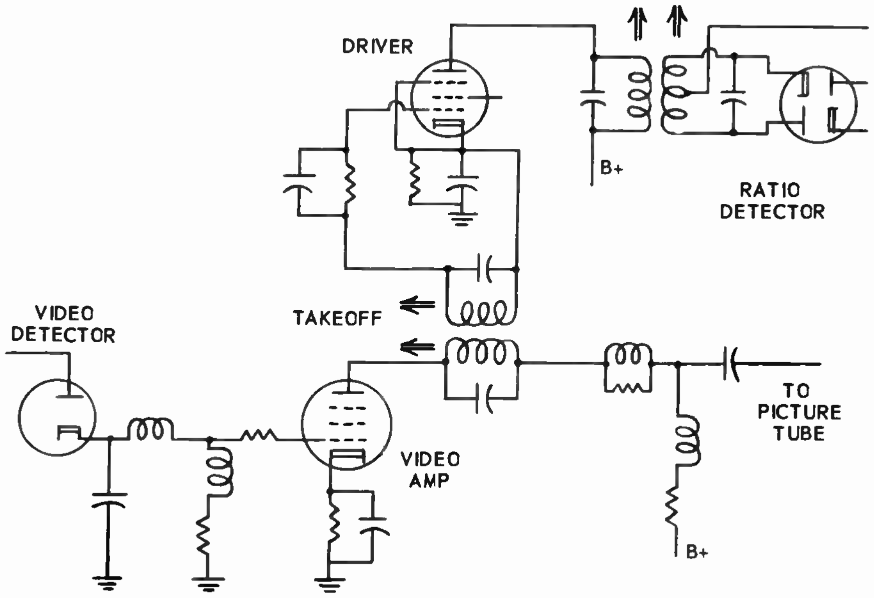


Fig. 80-4. Circuits between the video detector and sound demodulator for a typical intercarrier sound system.

tially different from i-f amplifiers and demodulators with which we became well acquainted when studying f-m broadcast receivers. The demodulator outputs are audio-frequency signals, no different from the audio signals obtained from any of the detectors used in a-m broadcast receivers. Therefore, we may proceed to an examination of audio amplifiers for television receivers of f-m or a-m sound receivers.

AUDIO AMPLIFIERS. The great majority of all television receiver audio amplifiers are resistance coupled types. An a-f resistance coupled amplifier is like a broad-band video amplifier without the broad-band features. That is, if we omit from a video amplifier the special features required for low-frequency and high-frequency compensation the remainder will be a resistance-coupled audio amplifier of usual design and construction. The elementary principles of all resistance-coupled amplifiers were covered in detail during our study of broad-band amplifiers.

3 Fig. 80-5 shows circuit connections for an audio amplifier of the general type used in many television receivers. At the left is a ratio detector whose output goes through the usual de-emphasis circuit used for f-m reception and to the volume control. The slider of the volume control potentiometer is connected to the grid of the a-f amplifier tube. This a-f tube would be classed as a voltage amplifier, since its purpose is to increase the a-f amplitude rather than add power to the audio signal.

4 Between the plate of the a-f amplifier and the grid of the output amplifier tube is an ordinary resistance coupling, with plate load resistor *Ro*, coupling or blocking capacitor *Cc*, and grid resistor *Rg*. The purpose of the output amplifier is to produce a moderate increase of audio signal amplitude, but chiefly to add power to the audio signal. The power is required for operation of the loud speaker which is coupled to the plate circuit of the output amplifier through an iron-cored *transformer*.

TONE CONTROLS. Tone control circuits are used to alter the apparent strength of certain audio frequencies as reproduced by a loud speaker. Tone controls usually are designed to add apparent strength to the

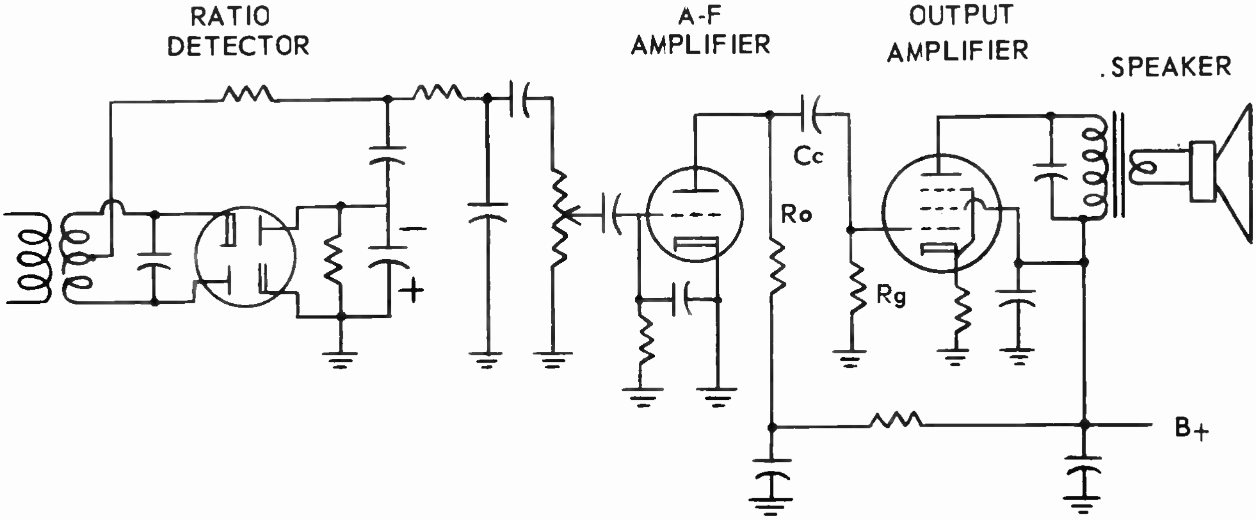


Fig. 80-5. Resistance coupled audio amplifier section such as used in many television receivers and f-m sound receivers.

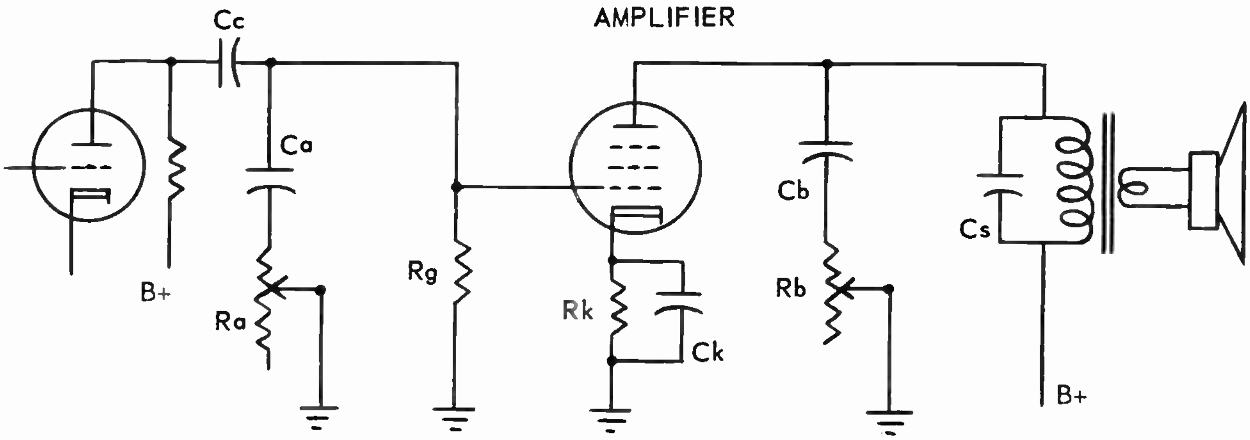


Fig. 80-6. Tone control elements used in an audio amplifier for emphasizing low or high frequencies.

low audio frequencies or bass notes, or they may be designed to emphasize the high audio frequencies or treble notes. Controls of one or both kinds are found on most f-m broadcast receivers and on a good many television and a-m sound receivers.

5 A tone control can add no sound frequencies that are not in the received signal, unless the added frequencies were to be in the nature of distortion. A tone control can reduce the audio signal amplitude at high frequencies, to leave the lows relatively stronger, or it can reduce the low-frequency amplitude to leave the highs relatively stronger. Also if the characteristics of the audio amplifier and loud speaker are such as to over-emphasize either the high or low frequencies, a tone control can help to flatten the response throughout the audio-frequency range.

Fig. 80-6 illustrates several methods of attenuating some frequencies and thus emphasizing other frequencies. The simplest and most commonly used tone control consists of a fixed capacitor *Ca* in series with an adjustable resistor *Ra*, connected from a signal coupling circuit to ground or B-minus. The reactance of the capacitor is large at low audio frequencies and is small at the high frequencies. The impedance of *Ca* and *Ra* together is equal to the sum of the capacitive reactance and the resistance, and may be varied by adjusting the resistor.

The impedance of the tone control elements is in parallel with grid resistor *Rg*. Consequently, varying the resistance at *Ra* varies the impedance of the entire circuit across which is developed the audio input voltage for the grid of the amplifier. This signal input impedance is being varied at the same time in accordance with frequency by the changing reactance of capacitor *Ca*.

Assuming that the audio signal voltage applied to the amplifier grid is proportional to impedance in the grid circuit, the action of this tone control is shown by Fig. 80-7. The graph is based on the use of 0.005 mf capacitance at *Ca*, of resistance adjustable from zero to one megohm at *Ra*, and of 0.470 megohm or 470 K ohms resistance at *Rg*. Each curve shows relative signal amplitude at frequencies from 100 to 12,800 cycles per second with the tone control resistor *Ra* set for a certain resistance.

With the tone control resistor set for one megohm there is very little control effect, the amplitude is very nearly the same at all frequencies. We are taking it for granted that the amplifying system has uniform gain at all frequencies when used without the tone control, which is not quite true, but simplifies the explanation of tone control effect.

As the tone control resistance is reduced there is greater and greater difference between the signal amplitudes at low and high audio frequencies. For example, with the control set at 250K ohms the amplitude at 1,600 cycles is down to about 35 on the relative scale but is down only to about 53 at 100 cycles. With the control resistance at zero the relative amplitude still is about 38 at 100 cycles, but is down around 4 at 1,600 cycles. This zero resistance curve shows the effect of capacitive reactance of capacitor *Ca* as this reactance varies with frequency. Note that the tone control always reduces the amplitude at all frequencies, but it provides greater reduction at high frequencies than at low frequencies. A tone control of the same type may be connected to a plate circuit, as shown by *Cb* and *Rb* of Fig. 80-6.

The cathode bypass capacitor *Ck* of Fig. 80-6 may be chosen of such value as to attenuate the low frequencies and emphasize the highs. The reactance of *Ck* is large at low frequencies, and it tends to force more of the signal current through cathode resistor *Rk*. The greater the signal current through *Rk* the greater is the degenerative feedback and the more the output is reduced. By using a rather small capacitance at *Ck* we may cause considerable degeneration at low frequencies, which leaves the highs emphasized. The reactance of *Ck* at the higher frequencies will drop low enough to bypass most of the signal current around *Rk*, so that there will be very little degeneration at these higher frequencies.

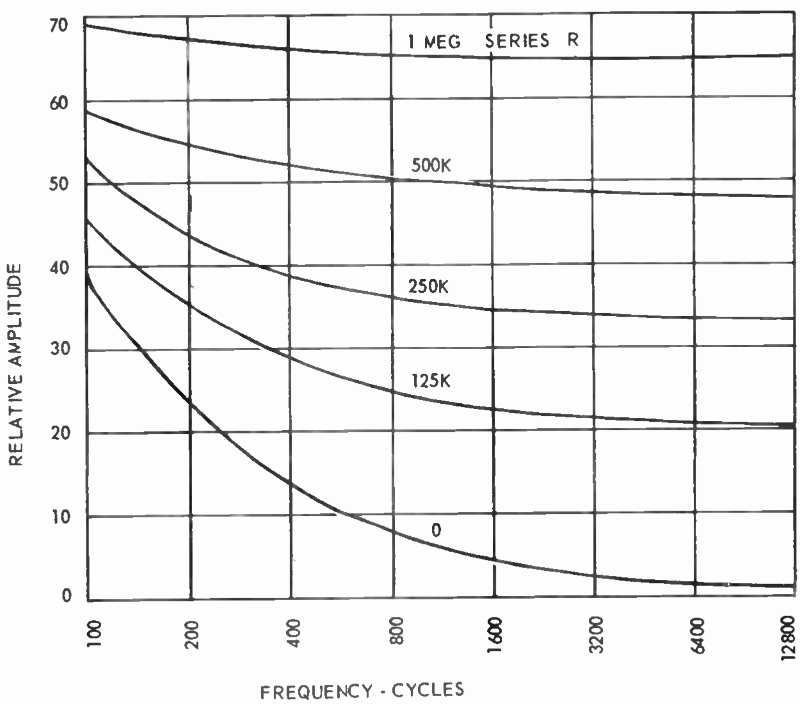


Fig. 80-7. How sound amplitudes at low and high frequencies are affected by adjustment of the most common type of tone control.

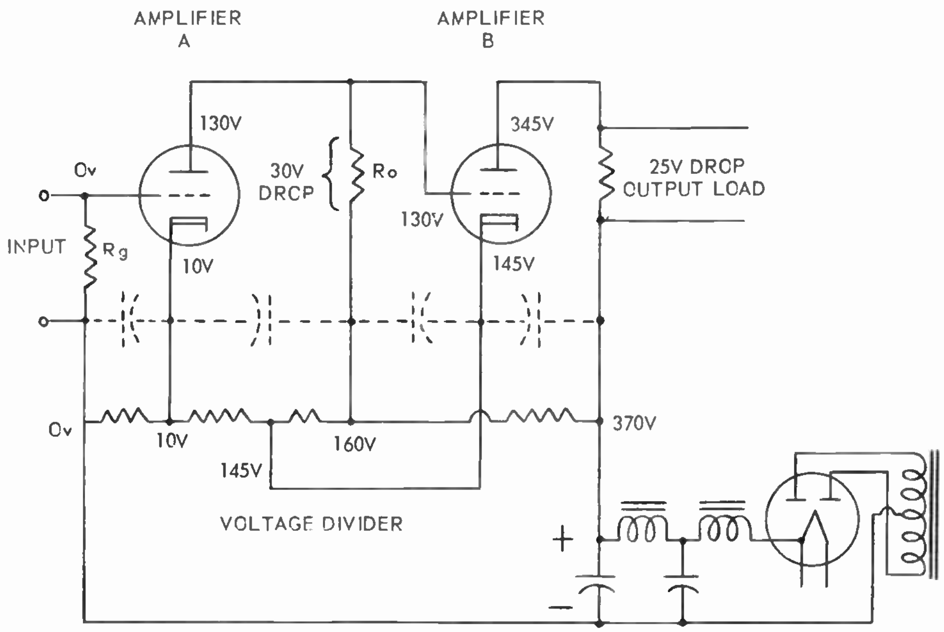


Fig. 80-8. Relations of tube voltages in a two-stage direct-coupled audio amplifier.

The capacitance of the coupling or blocking capacitor *Cc* of Fig. 80-6 affects the frequency response. When this capacitance is large its reactance will be small, even at low frequencies, and it will have little effect at any audio frequency. If the capacitance is made smaller its reactance will rise at the low frequencies. This attenuates the lows and emphasizes the highs. An adjustable tone control resistor may be connected in parallel with capacitor *Cc*, and an additional series blocking capacitor inserted between plate and grid. Increasing the resistance which is in parallel with the coupling capacitor then reduces the low frequency response and emphasizes the highs.

Tone controls are not always continuously adjustable. Many of them are operated with a selector switch that connects any of several tone control circuits to the amplifier. One control circuit and one switch position may give maximum low-frequency response, another may give fairly uniform response, and a third may give emphasis to the high frequencies.

DIRECT-COUPLED AMPLIFIERS. The coupling and blocking capacitor ordinarily used between the plate of one amplifier and the grid of a following amplifier tube may be omitted when using a direct coupled circuit such as illustrated in principle by Fig. 80-8. The voltages used in the example are chosen merely to illustrate the relations which are required. Actual voltages would depend on the type of tubes to be used.

The grid of amplifier *A* is connected through resistor *Rg* to zero voltage or B-minus on the voltage divider of the power supply system. The cathode of this amplifier is connected to a point on the divider which is 10 volts positive. Consequently, this amplifier works with a 10-volt negative grid bias, and there is no current in *Rg*.

The plate of amplifier *A* is connected through resistor *Ro* to 160 volts on the divider. We are assuming a 30-volt drop in *Ro*, which leaves 130 volts at the amplifier plate. Since this tube now is working with 130 volts on the plate and 10 volts on the cathode, it operates with the difference or with 120 plate volts.

The plate of amplifier *A* is conductively or directly connected to the grid of amplifier *B*, so the grid of amplifier *B* works at 130 volts positive. To provide 15 volts of negative bias for amplifier *B* we connect its cathode to 145 volts on the divider system, leaving the grid 15 volts negative with reference to the cathode.

We assume that amplifier *B* is to be operated with 200 plate volts, which requires that the plate be at 200 volts positive with reference to the cathode, or at 345 volts. Plate current of amplifier *B* goes through the output load resistance, in which we assume a drop of 25 volts. Adding this drop to the voltage at the amplifier plate gives a total of 370 volts required from the power supply.

6 To prevent undesirable resistance couplings in the voltage divider resistors it would be necessary to use large decoupling capacitors for all plate and grid returns, as shown by the symbols in broken lines. Although an amplifier laid out in this general manner, to have direct coupling from plate to grid, will handle very low audio frequencies it will not amplify sudden changes of direct current without distortion. This is because of the varying reactances of the decoupling capacitors and of the stray and tube capacitances as the direct current changes its value. At the instant of each change in direct current we have the equivalent of a change in an alternating current, and there are response delays due to time constants of the resistances and capacitances.

Were the rectifier power supply and voltage divider system to be replaced with batteries of suitable voltages, the very low internal resistances of the batteries would allow omitting the decoupling capacitors. Time constant delays then would practically disappear, and we would have what usually is called a direct-current amplifier system. The changes of output current in a resistance load then would follow changes of d-c input voltage almost instantly.

OUTPUT TUBES. Audio output tubes or audio power amplifiers in television, f-m, a-m, and combination receivers nearly always are beam power types or power pentodes. Both these types can provide power outputs up to two watts or even more, they have good sensitivity or good power amplification, and do not cause too much distortion of the audio tones.

In transformerless receivers of all kinds the power tube heaters may be rated and operated at 12.6 volts, or at 25, 35, or 50 volts. The tubes may be octal, lock-in, or miniature types. When the output tube is a miniature type it is common practice to have miniatures for the demodulator and a-f amplifier, although octal or lock-in output tubes often follow miniature tubes in the earlier stages.

The beam power tubes which have heater ratings of 25, 35, and 50 volts are designed to give satisfactory power outputs, up to around two watts, when used with plate voltages between 100 and 117. Some of the miniature power tubes in this class are designed for maximum plate voltages of 117 or sometimes up to 135. The octal and lock-in types permit higher plate voltages, and when so used will provide greater power outputs. The beam power and power pentode tubes designed for 6.3-volt heater operation require higher plate voltages for normal power outputs.

In a-m receivers and in combination fm-am sets a single output tube of any of the types mentioned is fed for a-m reception from a dual diode triode in which the diodes are the detector and the triode is the a-f amplifier.

In many television receivers a triple-diode triode tube is used as the demodulator and a-f voltage amplifier.

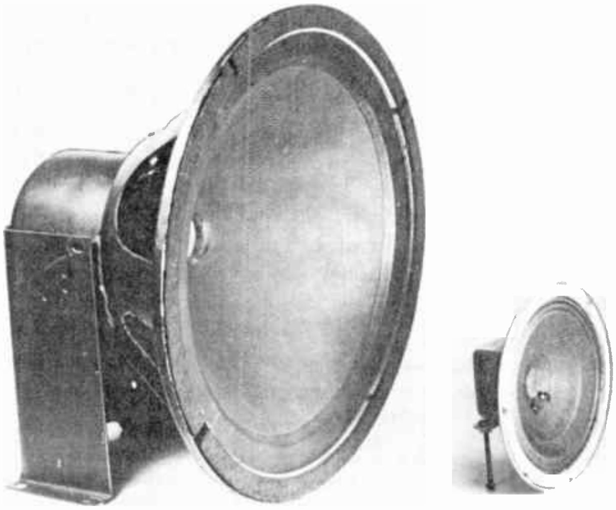


Fig. 80-9. Moving coil cone speakers of 12-inch and 5-inch sizes.

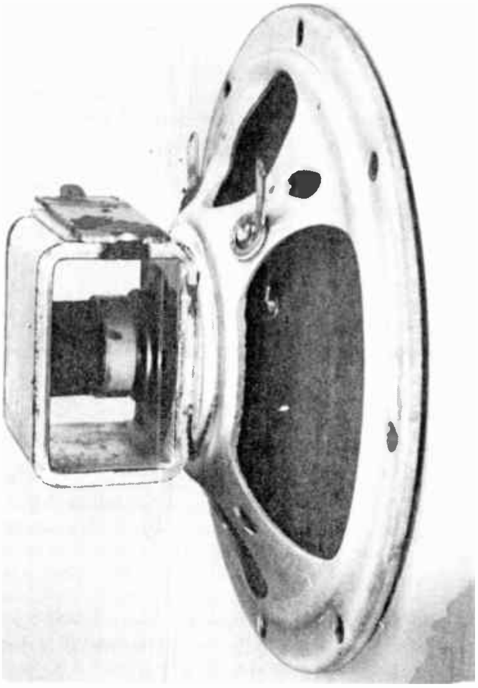


Fig. 80-10. A permanent-magnet or PM cone speaker.

The triode section is used for the amplifier function. The diode plate that operates with a separate cathode is used for one section of the demodulator. The other section of the demodulator makes use of the other two diode plates tied together. These two diode plates operate on the same cathode as the triode. This triple-diode triode tube is the same one used in combination fm-am sets for the three functions of a-m diode detector, f-m demodulator, and a-f voltage amplifier.

SPEAKERS. The speakers or loud speakers which are parts of all home radio and television receivers are of the moving coil type with a cone radiator. A 12-inch size and a 5-inch size of this type are shown by Fig. 80-9. Details of one of the smaller speakers, as viewed from its rear, appear in Fig. 80-10. Fig. 80-11 is a part sectional drawing showing all the principal parts of a small permanent magnet or "PM" cone speaker.

1 The part of the speaker that vibrates to produce sound waves in surrounding air is a cone of stiff fibre or paper. The conical form is adopted to obtain greatest possible rigidity. On a cylindrical extension of the center or apex of the cone is wound the voice coil. In this coil flow audio-frequency currents coming from the secondary of a speaker coupling transformer whose primary is in the plate circuit of the output tube or tubes of the receiver. Fig. 80-12 shows a voice coil, and part of the surrounding spider, attached to a cone. The ends of the coil are connected to two flexible leads which may be seen coming through the cone. These leads are connected to terminals for the coupling transformer secondary.

The voice coil is supported, by the cone and spider, in the exact center of a cylindrical air gap wherein is a very strong and concentrated magnetic field. In the type of speaker being examined the magnetic field for the gap is produced by a permanent magnet. The magnetic circuit passes from one pole of the magnet through the soft steel pot to the outside of the cylindrical air gap. From the other pole of the magnet the circuit is completed to the inside of the air gap through a soft steel cylindrical pole piece.

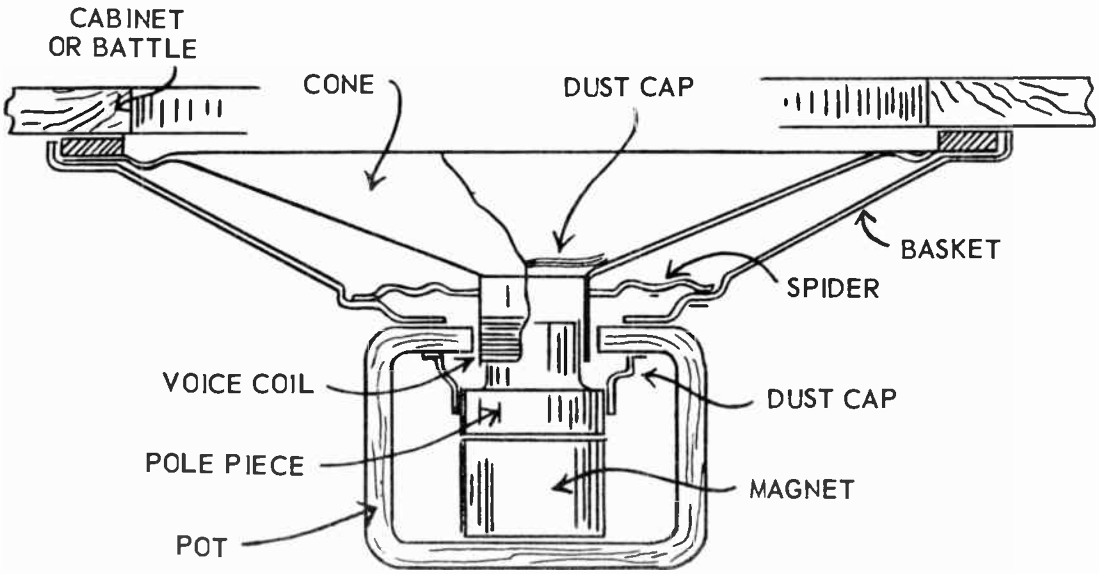


Fig. 80-11. Arrangement of parts in a typical small speaker of the PM type.

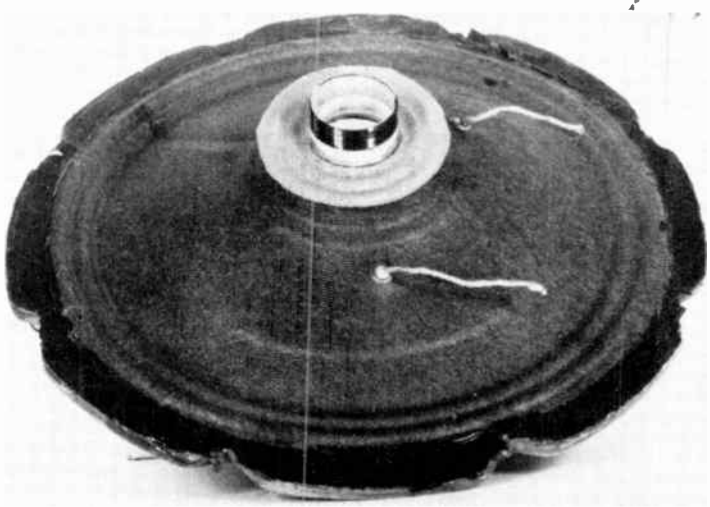


Fig. 80-12. The voice coil, the cone, and part of the spider from a small speaker.

Audio-frequency currents in the voice coil produce around this coil an electromagnetic field which varies in strength and polarity according to the audio signals. This field reacts with the permanent magnet field in the air gap to make the voice coil and attached cone vibrate at the audio frequencies of the signal. This vibration sets the surrounding air into motion, or produces sound waves, at these frequencies.

Dirt and dust are kept from the air gap by two dust caps. One, usually of thin felt, covers the center opening of the cone from the front. The other, of pressed steel, covers the pole piece and air gap from the rear of the speaker. This rear dust cap also supports and centers the pole piece. In small speakers one end of the permanent magnet may be cemented to the pot, and the pole piece cemented to the other end of the magnet. In larger units these parts usually are swedged into recesses or may be held with screws.

The outer edge of the cone is supported in the pressed steel basket. The center of the basket is welded, staked, or screwed to the pot, and the outer edge is attached to the cabinet or baffle which supports the speaker. The spider is a disc usually corrugated, made of paper, stiffened cloth, or thin plastic material. The outside of the spider is fastened to the basket and the inside is cemented to the extension of the cone.

Sound waves from the cone, or any other sound source, travel through air at a velocity of 1,127 feet per second when the temperature is 70° F. One complete sound wave consists of a region in which air particles are compressed, followed by another region in which the particles are more widely separated than in quiet air. Such a wave corresponds to the positive and negative alternations composing one cycle of the sound signal.

It is obvious that a forward motion of the speaker cone produces the portion of a sound wave which is a compression of the air. The same motion produces at the back of the cone the portion of a sound wave which is a separation. We call such a separation a rarefaction of the air, and say that waves consist of compressions and rarefactions. If the rarefaction produced at the back of the cone gets out to the edge of the cone and there meets the compression coming from the front, the two parts of the wave cancel or, at least, the sounds are badly distorted.

To prevent such cancellation or distortion of sound it is necessary to provide all around the outer edges of the speaker a baffle which will keep the parts of the waves separated until the front wave can move away and be heard by listeners. In most radio and television receivers the speaker is mounted in the cabinet, and the walls of the cabinet act as a baffle. If the cabinet is almost completely closed, sound waves from the back or enclosed side of the speaker cone are unlikely to interfere with waves from the exposed side. Otherwise, the greater the distance through unobstructed air from the back around to the front of the cone, the lower will be the sound frequencies reproduced without partial cancellation. This is because low-frequency sound waves are longer than high-frequency waves. A 100-cycle sound wave is 11.27 feet long, while a 5,000-cycle wave is only about 2.7 inches long.

Loud speaker frequency responses are typically of the forms illustrated by Fig. 80-13. That is, for equal input signal powers at all audio frequencies, the output sound powers will be about as shown. The two curves apply to two different speakers, both of excellent quality. The responses would not be nearly so smooth from one frequency to another as shown here, but would be full of short, sharp dips and rises all along their lengths. Always there will be a decided drop below 200 or 100 cycles, another drop above 2,000 or 3,000 cycles, with fairly uniform response through the middle frequencies.

At the lowest frequencies the cone remains nearly rigid, with the entire surface area vibrating like a solid piston. As frequency increases, the vibrations imparted to the center of the cone do not reach the outside

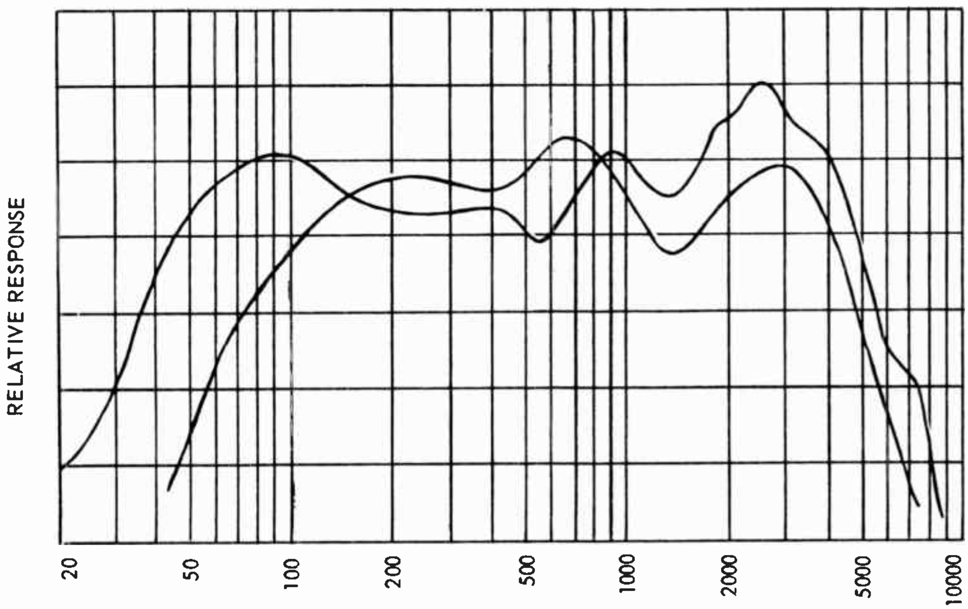


Fig. 80-13. Typical frequency responses of good quality cone speakers.

in full amplitude, and the cone acts as though it were of smaller diameter. At the highest frequencies the vibrations do not get far from the center, and the high-frequency limit is determined largely by the weight of the moving coil, the center of the cone, and by the stiffness of the spider. In general, the larger the diameter and the stiffer the cone the better will be the reproduction of lowest sound frequencies.

Except in a few channels, standard broadcast transmissions cannot carry audio frequencies exceeding 5,000 cycles per second without the modulation side bands getting into adjacent channels. Also, there is little or no transmission of frequencies below 100 cycles. There may be a somewhat wider frequency range in f-m broadcast transmission, also in television sound transmission.

The average sensitivity of human hearing is maximum around 2,000 to 3,000 cycles per second, and drops relatively little through frequencies down to about 500 cycles. There is an almost steady drop of ear sensitivity below 500 cycles and above about 4,000 cycles. If we could barely hear a sound at 2,000 cycles the intensity would have to be more than 50 times as great to barely hear a sound at around 100 cycles. As the volume control of a receiver is turned down, the low frequencies commence to disappear, while highs still are clearly audible. If notes of 1,000 cycles and 100 cycles have equal intensity at one volume level, and the intensity of both notes is dropped at the same rate, the 1,000 cycle note will remain after the lower note is gone. Doubtless you have noticed this effect when listening to radio; there is not a full range of musical tones when the volume is turned down unless the receiver has some special form of frequency response compensation.

One of the published ratings or characteristics of speakers is the maximum power in watts which they are designed to handle. This wattage rating must at least equal, and preferably should exceed, the maximum rated undistorted output of the power tube or tubes used in the receiver. These undistorted power outputs, in watts, are listed in tables of tube characteristics and average performances. The undistorted power rating does not mean complete absence of distortion, but a percentage distortion which is generally unobjectionable. A speaker of insufficient power handling ability will cause rasping and rattling sounds, and possibly burn-out of the voice coil or destructive vibration of the cone structure. Most receivers for home receivers have power ratings of 3-1/2 to 5 watts maximum, and sometimes less for small sets.

Speakers for home receivers usually are within the power handling limits of 2 to 12 watts and have diameters between 4 and 12 inches, or are of oval shapes such as 4 by 6, 5 by 7, and 6 by 9 inches. Approximate relations between maximum power in watts and diameters of the cones are in general as follows.

2 watts:	Diameter 4 inches.
3 watts:	Diameter 4 or 5 inches.
4 watts:	Diameter 4 to 7 inches.
5 watts:	Diameter 4-1/2 to 8 inches.
6 watts:	Diameter 5 to 10 inches.
8 watts:	Diameter 6 to 12 inches.
10 watts:	Diameter 7 to 12 inches.
12 watts:	Diameter 8 to 12 inches.

In the cataolgued or published ratings of permanent magnet speakers you nearly always will find the magnet weight in ounces as one of the items. The greater the weight of the magnet, with other design factors unchanged, the greater will be the speaker sensitivity and the longer the magnet will last. Magnet weights run all the way from around 3/4 ounce to 5 ounces and more for speakers with power ratings of 3 to 12 watts and cone diameters of 4 to 12 inches. A 3/4 ounce magnet might have a diameter of 11/16 inch and a length of about 7/16 inch in Alnico V material. A 1/8 ounce magnet might be 7/8 inch diameter and 11/16 inch long. Dimensions for given weights will vary with requirements of the speaker design.

SPEAKERS WITH WOUND FIELDS Some radio and television speakers do not have a permanent magnet to furnish the constant field for the air gap but employ instead a wire winding on a core which forms the pole piece. Otherwise the construction is like that of PM speakers. Rectified direct current from the low-voltage power supply of the receiver is passed through the speaker field coil. This coil ordinarily is used as a filter choke, sometimes as the only choke and again as a choke for a second section of the power filter.

The ohmic resistance of a wound field coil may be almost anything between 50 and 5,000 ohms. Resistances of 450, 1,000, and 1,500 ohms are common for sound radio receivers, while in television sets the field resistances more often are between 60 and 100 ohms. The magnetic strength depends, of course, on the number of ampere-turns of the winding, or on the product of d-c current in amperes and turns in the field coil. Catalog specifications for wound field speakers include the ohms of field winding resistance, the voltage required across the field for the necessary current, and, if used as a filter choke, the inductance in henries. Naturally the specifications include also the rated maximum power in watts, the cone diameter, and other details.

Speakers with wound fields instead of permanent magnet fields are called electrodynamic speakers or electromagnetic speakers. Fig. 80-14 shows how such a speaker may be connected when its field winding is used as a choke for the low-voltage B power filter.

The rectified direct current through the speaker field winding will have some ripple voltage. This ripple would vary the strength of the speaker field and cause slight vibration of the speaker cone at the power ripple frequency, which would be 120 cycles with a full-wave power rectifier and 60-cycle line power. To get rid of this effect most electromagnetic speakers are provided with a hum bucking coil connected as shown in the diagram.

The hum bucking coil consists of a few turns of wire placed on the outside of, or at one end of the field winding, so that this coil picks up some ripple-frequency voltage. The hum bucking coil is connected in series with the speaker voice coil and the secondary of the speaker coupling transformer or output transformer. The connections are made in such manner that voltage from the bucking coil is in opposite phase to the effect of ripple voltage or current in the field winding, and thus the power ripple effect is cancelled out so far as the speaker cone vibration is concerned.

Connections between the hum bucking coil, the voice coil, and speaker terminals for the secondary of the output transformer ordinarily are contained within the speaker structure, and are made to cancel hum effect no matter which way the transformer leads are connected to the speaker terminals.

9 *IMPEDANCE MATCHING.* It is highly desirable that the audio power output from the final tube or tubes in the receiver be transferred to the voice coil and cone of the speaker and thus to sound waves with the least possible energy loss or distortion. This is the function of the speaker coupling transformer. There is maximum power transfer between any source of energy and a load only when the internal resistances or impedances of source and load are equal.

The output impedance or plate resistance of the power tube always is several thousands of ohms, while the impedance of the voice coil on the speaker cone is in the range of three or four ohms up to maybe thirty or forty ohms. The speaker coupling transformer must allow the output tube to work into a load impedance (the transformer primary) of 2,000 to 10,000 ohms while the voice coil is fed from a source (the transformer secondary) whose impedance is very close to that of the coil itself.

The power tube "looks into" a load impedance comparable to its own internal impedance. The voice coil "looks into" its own impedance in the transformer secondary. Thus the unequal impedances of the tube and

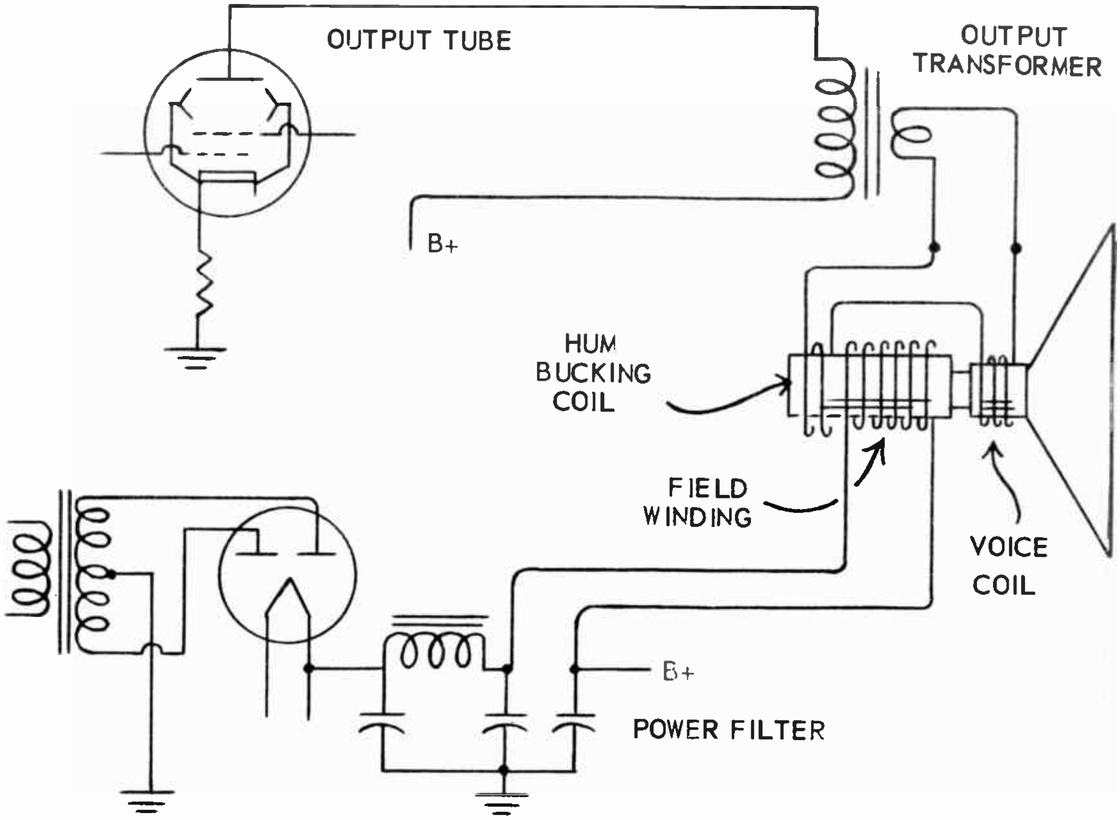


Fig. 80-14. Circuit connections for a speaker having a wound field and a hum bucking coil.

the voice coil are matched. Unequal impedances may be matched with various combinations of resistors, with tuned circuits, with quarter- and half-wave sections of transmission line, and numerous other ways, all of which introduce considerable losses of energy. An impedance matching transformer will do the job with very little attenuation. Unless there is effective matching of impedances the actual power from the output tube will be far less than the normal or rated power output for the voltages and currents with which the tube is operated.

Beam power tubes of types commonly used in transformerless receivers of all kinds have plate resistances of 10,000 to 15,000 ohms when operated with about 110 volts on their plates. These tubes usually are worked into plate load resistances or impedances of 2,000 to 2,500 ohms, whereupon total distortion due to formation of harmonics is around 10 per cent. The relatively small load resistances or impedances are permissible because of the relations which exist between plate voltages and currents in beam power and power pentode tubes. A power triode should be worked into a plate load resistance at least equal to its plate resistance, and preferably much greater.

Beam power and power pentode tubes of the 6.3-volt heater class have plate resistances usually around 20,000 to 70,000 ohms, and they are worked into load resistances or impedances of 5,000 to 8,000 ohms with total harmonic distortion of 8 to 12 per cent. Plate resistances and desirable load resistances for typical operating conditions are listed in the published data for all types of output or power tubes.

The impedance of the coupling transformer primary is very nearly proportional to the inductive reactance of this winding, because the ohmic resistance is so small in comparison with the reactance as to have little effect on impedance. Often the inductive reactance of the primary is made approximately equal to the recommended plate load resistance at the lowest audio frequency for which low-frequency and middle frequency amplifications are to be nearly equal. Greater impedance in the transformer primary will raise the low-frequency response. Standard types of speaker coupling transformers used for service replacements are available with primary impedances all the way from 1,500 ohms to 15,000 ohms and higher.

The RTMA standard voice coil impedance as measured at a frequency of 400 cycles per second is 3.2 ohms. This impedance, or an impedance of approximately 3 to 4 ohms, is found in the voice coils of speakers from the smallest sizes up to 8-inch cone diameter, and often in sizes up to 12-inch or ever greater diameter. Voice coils with 400-cycle impedances of 6 to 8ohms are common in speakers having cone diameters of 10 inches and more, and may be used in smaller sizes.

Voice coil impedance varies with frequency about as shown by Fig. 80-15. Impedance is minimum at 400 cycles. It increases with drop of frequency until reaching a value which may be four to six times more than the minimum at what is called the bass resonant frequency, then there is a rapid drop at still lower frequencies. The impedance increases rather slowly as frequency rises above 400 cycles. These variations of impedance are due largely to variations of the air load or to variations of opposition of the air to being set into motion by the cone.

When designing an impedance matching transformer the number of turns for the primary winding would be

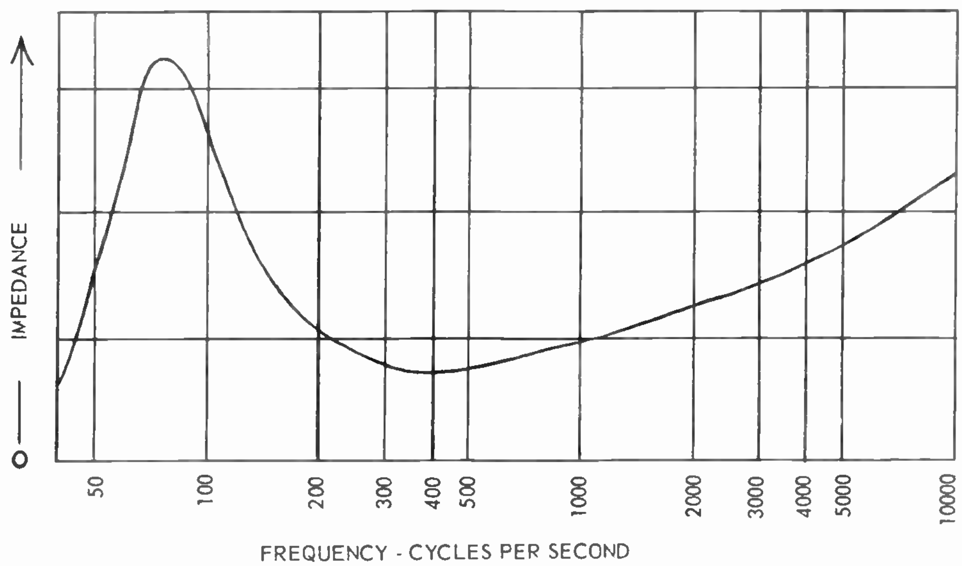


Fig. 80-15. The manner in which voice coil impedance varies with frequency.

selected to provide the desired plate load impedance for the output tube. The number of turns required would depend on many factors, such as size and kind of core material, on closeness of coupling between the windings, on the operating frequencies to be considered, and so on. Then the required ratio of primary to secondary impedances, for matching the tube load to the voice coil would be obtained by making the ratio of secondary to primary turns equal to the square root of the ratio of voice coil impedance to the tube load resistance. Knowing the number of primary turns and the two impedances, the number of secondary turns would be given, approximately, by this formula.

$$\text{Secondary turns} = \sqrt{\frac{\text{voice coil impedance} \times (\text{primary turns})^2}{\text{primary impedance}}}$$

The computed number of secondary turns will be approximate because of iron losses, copper losses, and other losses during energy transfer in the transformer. In matching for 2,500 ohms of plate load impedance to 3.2 ohms of voice coil impedance the number of secondary turns would be about 1/28 of the number of primary turns.

In selecting a replacement transformer the primary impedance never should be more than 10 per cent under the recommended load resistance for the output tube or tubes, although this impedance may be as much as 25 per cent greater than the recommended load resistance and still have excellent performance. If a given transformer were designed for some certain primary and secondary impedances, or for a voice coil of certain impedance, it could be successfully used on a voice coil of greater impedance but not on one of less impedance. The higher impedance voice coil would "reflect" into the primary an impedance greater than the rated primary impedance and the output tube would work into an impedance somewhat higher than the one recommended.

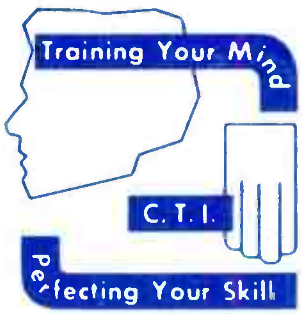
Sometimes you may use what is called a universal output transformer. The primary winding is tapped at various numbers of turns providing impedances for two or three values of recommended tube loads, and the secondary is tapped for two or more voice coil impedances. Should there be no voice coil tap within about 10 per cent of the rated voice coil impedance of the speaker, use a tap whose number of ohms is less than the rated voice coil impedance. This will increase the impedance reflected back into the plate circuit of the output tube.

81 → 85

COMMERCIAL TRADES INSTITUTE

KNOWLEDGE + AMBITION = SUCCESS

LEARN . . .



ACT . . .

"The heights by great men reached and kept
Were not attained in sudden flight,
But they, while their companions slept,
Were climbing upward in the night."

ENJOY . . .

COMMERCIAL TRADES INSTITUTE

CHICAGO, ILLINOIS

TELEVISION

LESSON NO. 81

AUDIO POWER AMPLIFIERS

Training Your Mind...

C.T.I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind

C.T.I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON NO. 81

AUDIO POWER AMPLIFIERS

Power is required to drive or operate the speakers which produce the audible sounds from AM., FM. receivers, Public Address systems and etc. Therefore a power amplifier stage must be used in conjunction with the speaker in these electronic devices. Any beam power, power pentode, or triode tube designed to deliver several watts of audio power is classed as a power amplifier tube.

One method of obtaining a large power output employs two or more power tubes connected in parallel. Typical circuits for two paralleled tubes are shown by Fig. 81-1. The a-f voltage amplifier is a triode, such as might follow any detector or demodulator. The output of the voltage amplifier is resistance coupled to the grids of two beam power tubes by means of load resistor Ro, coupling capacitor Cc, and grid resistor Rg. The plates of the two power tubes are connected together and to the primary of the speaker coupling transformer. The two cathodes are connected together, and to ground through cathode bias resistor Rk and bypass capacitor Ck. The power tube screens are connected together and to B plus.

With the same grid signal voltage and with the same d-c voltages on all the elements of each paralleled tube as would be used on a single tube, the audio output power is proportional to the number of power tubes in parallel; twice the power for two tubes, three times for three tubes, and so on. The effective transconductance of the stage is doubled when using two paralleled tubes. The effective plate resistance of the two tubes is half that of one similar tube. Consequently, the load resistance or impedance for two paralleled tubes may be half that required for a single similar tube.

With two paralleled tubes operated on the same d-c voltages as a single similar tube the total plate current is doubled, and so is the total screen current. This, of course, means a doubling of total cathode current. Since cathode current is doubled, the resistance for cathode bias must be only half that used on a single similar tube for any given bias voltage. The value required for a cathode bypass capacitor depends on frequency considerations, so is not altered when the tubes are paralleled.

Were the paralleled grids and plates connected directly to each other, one of the tubes might carry more of the total load than the other. Also, it is quite likely that oscillation would occur. These effects are prevented by connecting in series with each grid, right at the tube socket, a non-inductive resistor of about 100 ohms. Such resistors are marked a and a in Fig. 81-1. To further lessen possibility of unequal load division and oscillation, non-inductive resistors of 25 to 100 ohms may be connected to each plate, at the tube socket. These resistors are shown as b and b.

INTERSTAGE AUDIO TRANSFORMERS. The power amplifiers to be examined next will be the type

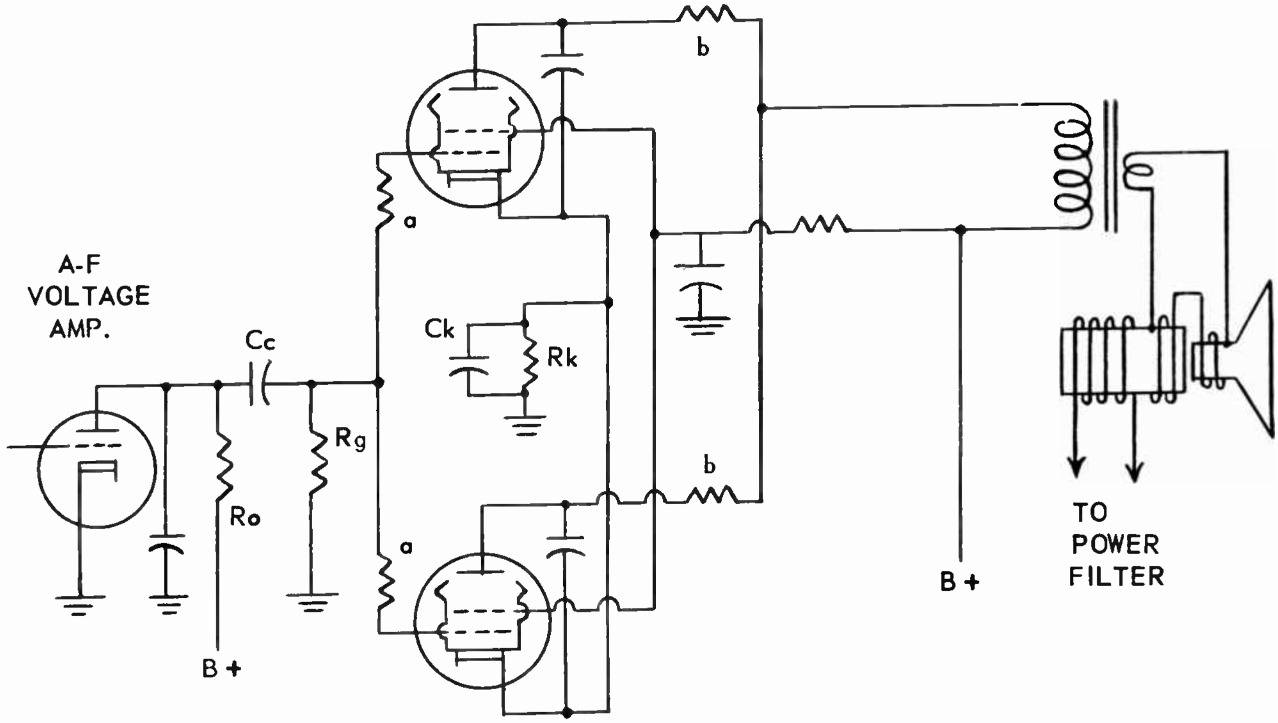


Fig. 81-1. Parallel power tubes in an audio amplifier.

using two power tubes in a push-pull connection, but before proceeding it will be well to learn something about audio-frequency transformer couplings which sometimes are found in such amplifiers. Audio transformers of the kind commonly used in receivers and many power amplifiers are rather small in size, as may be seen from Fig. 81-2 where a typical interstage transformer stands at the left of an ordinary power transformer.

Four classes of interstage audio transformers are shown in Fig. 81-3. At 1 we have a single plate to single grid transformer. At 2 there is a single plate to push-pull grids type. Diagram 3 shows a type classed as push-pull plates to push-pull grids. At 4 is a seldom employed type classed as push-pull plates to single grid. Where there are two tubes on either side of the transformer the winding for that side is center tapped.

There is no conductive connection between primary and secondary windings or between plate and grid windings of an audio transformer. Coupling is inductive, and no blocking or coupling capacitors are needed. In the primary winding or plate winding there is a direct plate current varying at audio frequency, or a direct current with an a-f component. Every increase and every decrease of primary current induces a corresponding emf in the secondary. This secondary emf is of one polarity when primary current increases, and is of opposite polarity when primary current decreases. Therefore, the a-f component of primary current induces an alternating a-f voltage in the secondary, and this a-f secondary voltage is applied to the grid or grids of following tubes. The direct plate current flows only in the primary. There

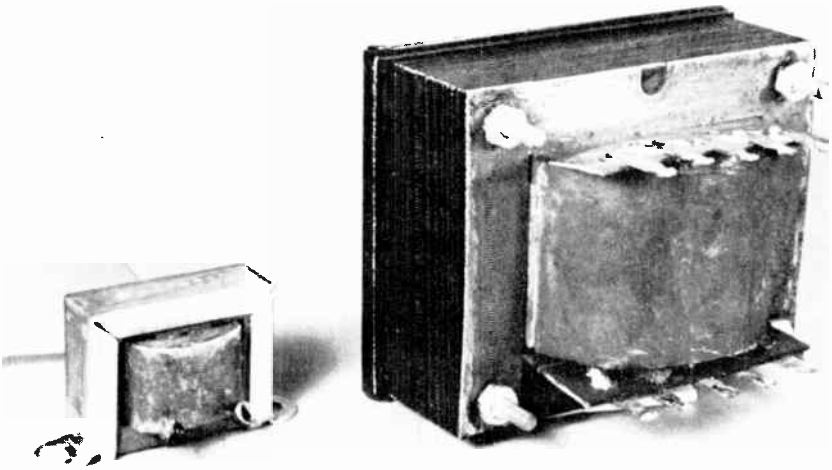


Fig. 81-2. An ordinary interstage audio transformer compared for size with a radio power transformer.

is no current in the secondary, but only a varying voltage, so long as the grid of the following tube remains negatively biased and so long as positive peaks of signal voltage do not exceed the negative bias voltage.

Ohmic resistance of the primary or plate winding in an interstage audio transformer is only a few hundred ohms, usually something between 200 and 1,000 ohms at most. For any given plate current the voltage drop in this small resistance is considerably less than in any plate load resistance used for resistance coupling. Then there is better regulation of d-c plate voltage than with resistance coupling. Also, it is possible to have any given voltage on the tube plate with a smaller voltage from the B supply.

An iron-cored audio-frequency transformer may have a step-up turns ratio and voltage ratio between primary and secondary. This increases the overall voltage gain of the amplifying stage, since the gain of the tube is multiplied by the effective step-up ratio of the transformer. The primary to secondary or plate to grid step-up ratio is approximately equal to the ratio of primary to secondary turns, although it always is somewhat less because of energy losses in the transformer.

Turns ratios and voltage ratios usually are somewhere between 1:1½ and 1:3½ when there is a step-up. The ratio is increased by using fewer primary turns, using more secondary turns, or both. Too few primary turns will so reduce the inductance and impedance of the primary as to severely cut the low-frequency response. Too many secondary turns so increases the distributed capacitance of the winding that there may be self-resonance between capacitance and winding inductance at some frequency in the audio range. Such resonance causes excessive peaking of the amplification at the resonant frequency, and there is undesirable distortion. These are the reasons for limiting the step-up ratio when good tone quality is to be preserved.

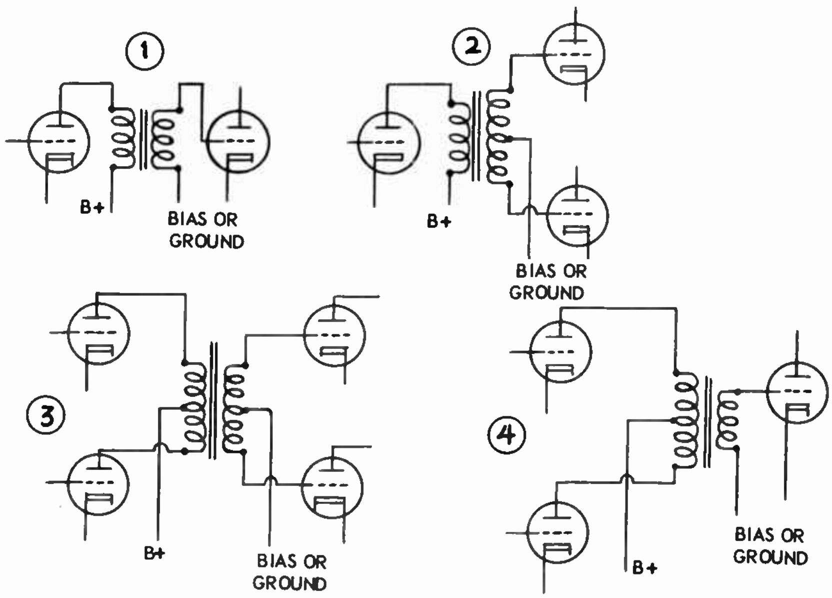


Fig. 81-3. Types and connections of interstage coupling transformers.

Primary inductance and resulting primary impedance must be great enough to provide a satisfactory load impedance for the connected tube. This tube most often is a triode for which desirable load impedances, and transformer primary impedances, range from 10,000 to 15,000 ohms. The ratio of impedances between primary and secondary is equal approximately to the square of the turns ratio. For example, with a primary to secondary turns ratio of 1 to 3 the impedance ratios will be as 1 to 9, because the square of 1 is 1 and the square of 3 is 9. Then, with a primary impedance of 10,000 ohms there would be a secondary impedance of about 90,000 ohms.

Few interstage transformers are designed to carry more than about 10 milliamperes of d-c plate current in the primary. This is satisfactory because tubes connected to the primaries are voltage amplifiers, and require only small plate currents. Excessive primary current will so increase the steady magnetism or magnetic flux as to "saturate" the core iron. Then the variations of flux which induce secondary emf's cannot be as great as they should be in following the audio signal. There is a reduction of inductance, of inductive coupling, and of signal transfer from plate circuit to following grid circuit.

Effective transfer of low-frequency signals requires maintaining a high impedance at the low frequencies. High impedance or high inductive reactance at low audio frequencies requires large inductance. Large inductance requires a primary winding of many turns on a large core of high permeability steel. Consequently, if an interstage transformer is to have good low-frequency performance it must be rather large and costly. For a limited frequency range, such as speech only, the transformer may be smaller in both core size and number of turns.

PUSH-PULL AMPLIFICATION. Fig. 81-4 shows, in an elementary way, the connections for a stage of push-pull amplification employing transformer coupling. There are many good reasons for using the push-pull method for power amplifiers. For one thing, it greatly reduces or entirely eliminates the production of second, fourth, and other "even order" harmonic frequencies which are troublesome in single-tube and in parallel connected power amplifiers. There also is nearly complete cancellation of hum effects due to poor filtering in the B supply. With some methods of operation it becomes possible to obtain from a push-pull stage more than twice the power output of a single-tube stage and more than the output from a parallel stage using similar tubes.

When an audio signal causes a change of plate current in the plate of the preceding a-f voltage amplifier and in the primary of the push-pull transformer, a corresponding emf is induced in the center-tapped secondary of the transformer. Opposite ends of this secondary are connected to the grids of the push-pull amplifier tubes. Then the grid of one tube is driven more positive (or less negative) while the grid of the other tube is driven more negative. There is inversion of voltage polarity between grids and plates, just as in all amplifiers, so the first plate becomes less positive (effectively more negative) while the other plate becomes more positive. Flow of plate current then is increasing in the tube whose grid is going positive and is decreasing in the tube whose grid is going negative. Now we have changing

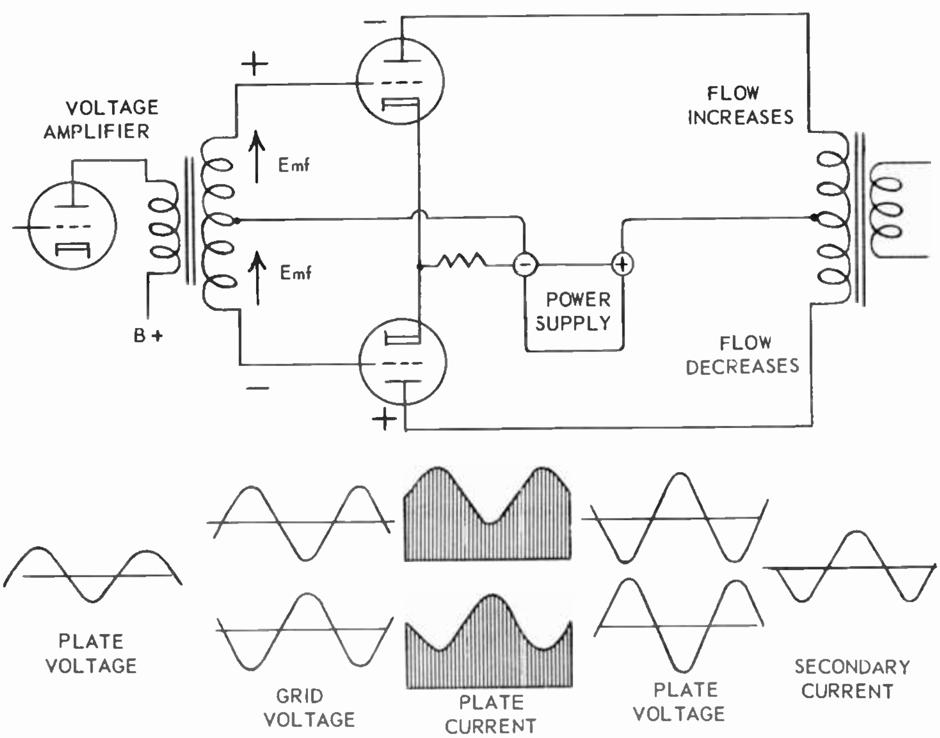


Fig. 81-4. Voltage and current variations in a push-pull audio amplifier.

electron flows in both halves of the center-tapped primary of the output transformer. The combined inductive effects of these primary current changes act together to induce a single emf and resulting current in the secondary of the output transformer.

The small graphs at the bottom of Fig. 81-4 show changes that are occurring at the same instants of time in grid voltages, plate currents, and plate voltages of the two push-pull tubes. Note especially the two plate currents, which are d-c currents with an alternating signal component. While one of these currents increases by any certain amount the other is decreasing by the same amount. As a result, total current for the two plates, taken through the B power supply, remains constant.

The coupling between plates of push-pull amplifier tubes and the speaker voice coil always is through an iron-cored output transformer with center-tapped primary. The coupling between the preceding a-f voltage amplifier and the push-pull grids of radio and television receiver amplifiers very seldom is by means of an interstage transformer, nearly always it is a resistance coupling.

6 The audio signal from the plate of the single a-f voltage amplifier must be made to furnish two signal voltages that are of opposite phase or are 180 degrees out of phase. These out-of-phase signal voltages are applied to the push-pull grids. The signal voltage for one of the push-pull grids may be taken from the plate of the a-f voltage amplifier. The opposite-phase signal voltage for the other push-pull grid is obtained from an inverter tube. You will find that the action of push-pull inverters of the types commonly used, is almost identical with that of inverters used with picture tubes of the electrostatic deflection type. In such picture tubes it is necessary to apply to opposite plates of a pair, two voltages which are of the same wave-form, of equal amplitudes, but opposite in phase. This is just what we need for the two grids of the push-pull tubes.

A widely used method of phase inversion is illustrated by the circuits of Fig. 81-5. From the plate of the a-f voltage amplifier tube the audio signal goes through capacitor Ca to the grid of push-pull amplifier A. This signal passes also through resistors Ra and Rb to ground, and through ground back to the cathode of the a-f voltage amplifier. Resistors Ra and Rb form a voltage divider. Resistance at Ra is much greater than at Rb, so that only a small part of the signal voltage appears across Rb.

The relatively small signal voltage across divider resistor Rb is applied through capacitor Cb to the grid of the inverter tube. The inverter amplifies this signal voltage, and from the inverter plate the amplified signal voltage goes through capacitor Cc to the grid of push-pull amplifier tube B.

The polarity or phase of signal voltage at the grid of the inverter is the same as at the plate of the a-f voltage amplifier and at the grid of push-pull tube A. Signal voltage is inverted between the grid and plate of the inverter tube. Therefore, the polarity or phase of the signal voltage applied to the grid of push-pull tube B is opposite to that of the signal voltage applied to the grid of push-pull tube A.

Amplitude of audio signal voltage applied to the grid of push-pull tube B must be equal to the amplitude of this signal voltage at the grid of push-pull tube A. If signal amplitude at the inverter grid were to be the same as from the a-f amplifier plate and at the grid of push-pull tube A, the amplification in the inverter tube would increase the signal amplitude and make it much greater at the grid of push-pull tube B than at the grid of A. This amplification in the inverter is counteracted by a proportional reduction of signal amplitude applied to the inverter grid.

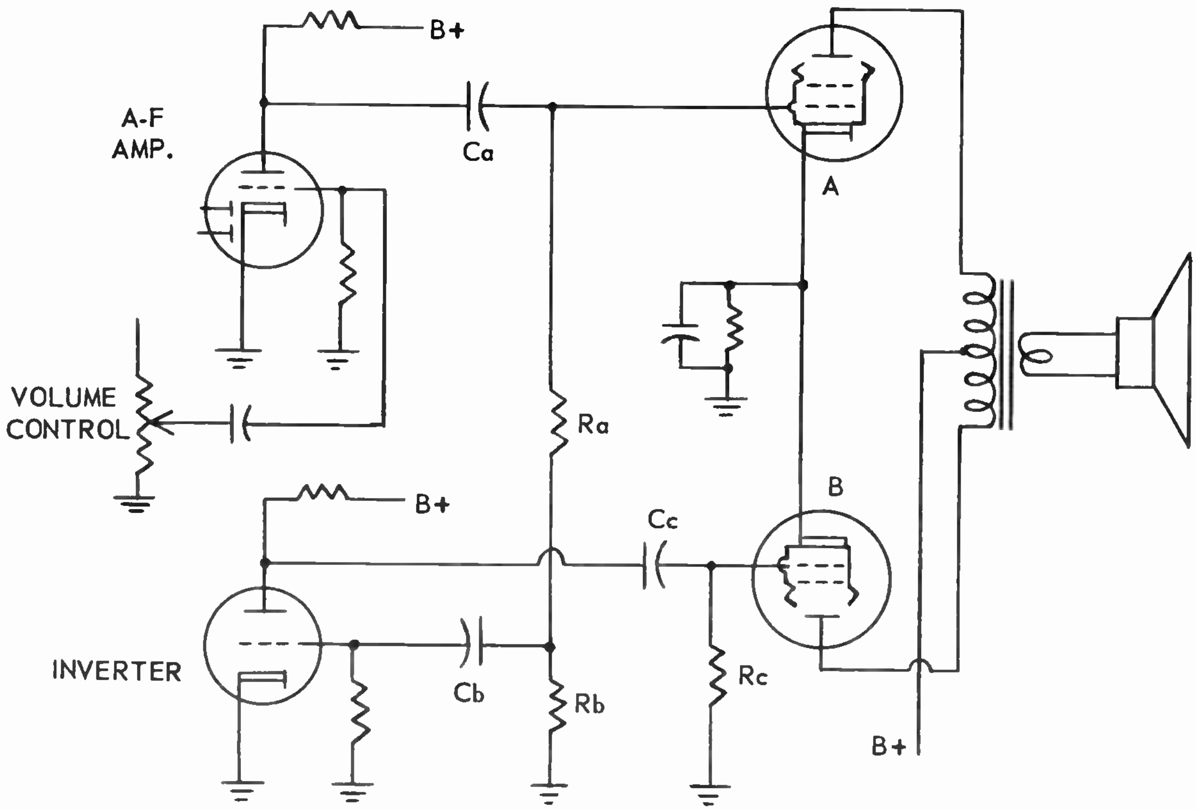


Fig. 81-5. Using an inverter tube and a voltage divider to furnish grid voltages to two push-pull amplifier tubes.

In a typical case the voltage amplification of the inverter might be 60 times. Then the signal amplitude at the inverter grid should be only 1/60 of that from the a-f amplifier plate and at the grid of push-pull tube A. The necessary reduction of signal amplitude is provided by the voltage divider resistors Ra and Rb. The resistance at Rb would be made 1/60 of the total resistance in Ra and Rb together.

The inverter circuits based on the voltage divider principle may be altered in many ways without changing the basic principle. A few variations are shown by Fig. 81-6. The a-f voltage amplifier and inverter functions are combined in a single twin-triode tube. The signal dropping voltage divider consists of resistors Ra and Rb, just as in the preceding diagram. The connection of grid resistor Rc has been changed from ground to the junction between Ra and Rb. With this new arrangement the total grid resistance for amplifier A consists of Ra and Rb in series. Total grid resistance for amplifier B consists of Rc and Rb in series. If Ra and Rc are made equal, which is convenient for construction, both push-pull amplifiers will have equal grid resistances to ground. With the arrangement of Fig. 81-5 this equality of grid resistances to ground would require making the resistance at Rc equal to the sum of the resistances at Ra and Rb.

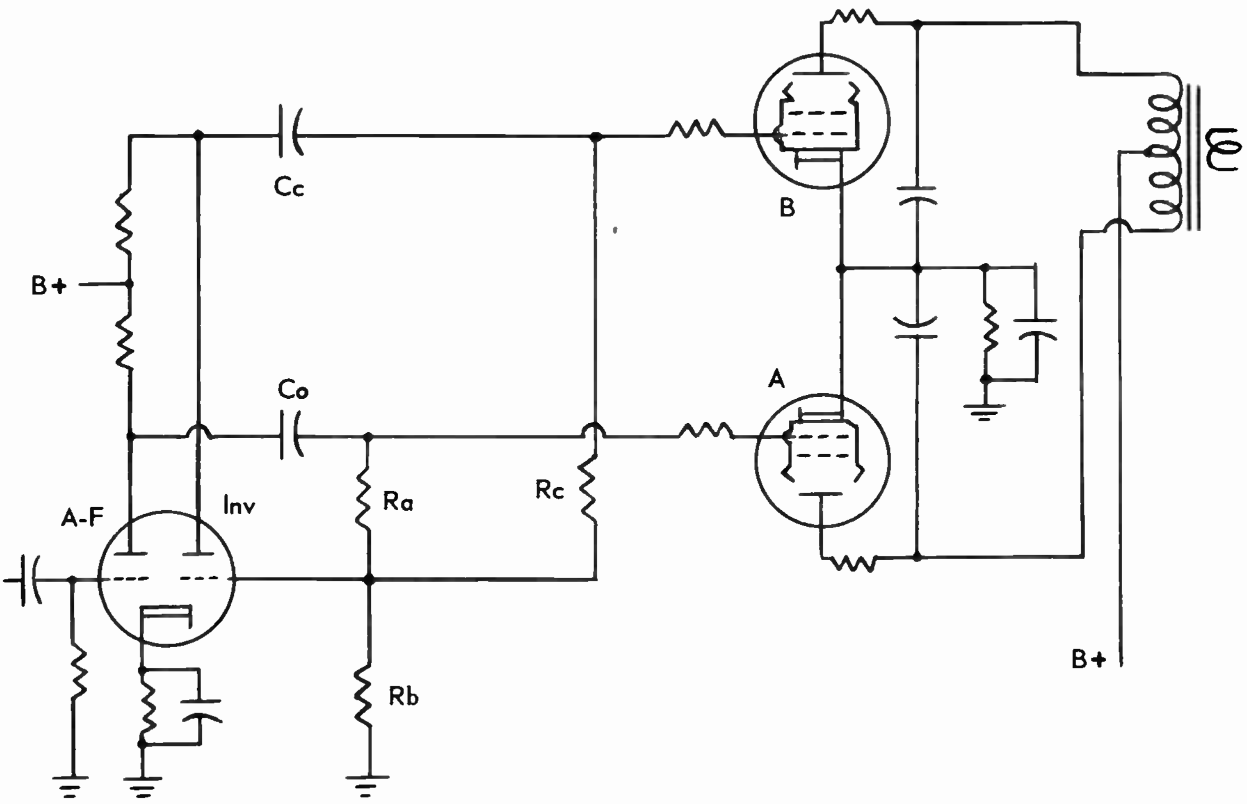


Fig. 81-6. The inverter and voltage divider principle may be utilized in various circuit combinations.

In Fig. 81-6 there are small non-inductive resistors in series with the grids and the plates of the two push-pull amplifiers. These resistors are used, as with the paralleled power amplifiers, to insure equal division of the load and to prevent oscillation.

An entirely different method of phase inversion is illustrated by Fig. 81-7. The grid of push-pull amplifier A is fed from the plate of the triode phase inverter tube. The grid of push-pull amplifier B is fed from the cathode of the phase inverter tube. When a signal voltage is applied to the grid of the triode phase inverter the polarity or phase of the signal is inverted at the plate of this tube, but it is not inverted at the cathode. This is a fact which we have encountered many times in various applications of triodes and pentodes. Here the result is application to the grids of the two push-pull amplifiers of audio signal voltages that are of opposite phase or polarity.

You should note that the plate-cathode circuit for audio signals in the phase inverter tube passes from the plate through resistor Ro and capacitor Cd to ground, thence through ground to cathode resistor Rk, through this resistor to the cathode, and from cathode to plate inside the tube. The load in the plate-cathode circuit consists chiefly of resistances at Ro and Rk, since capacitance at Cd is so great

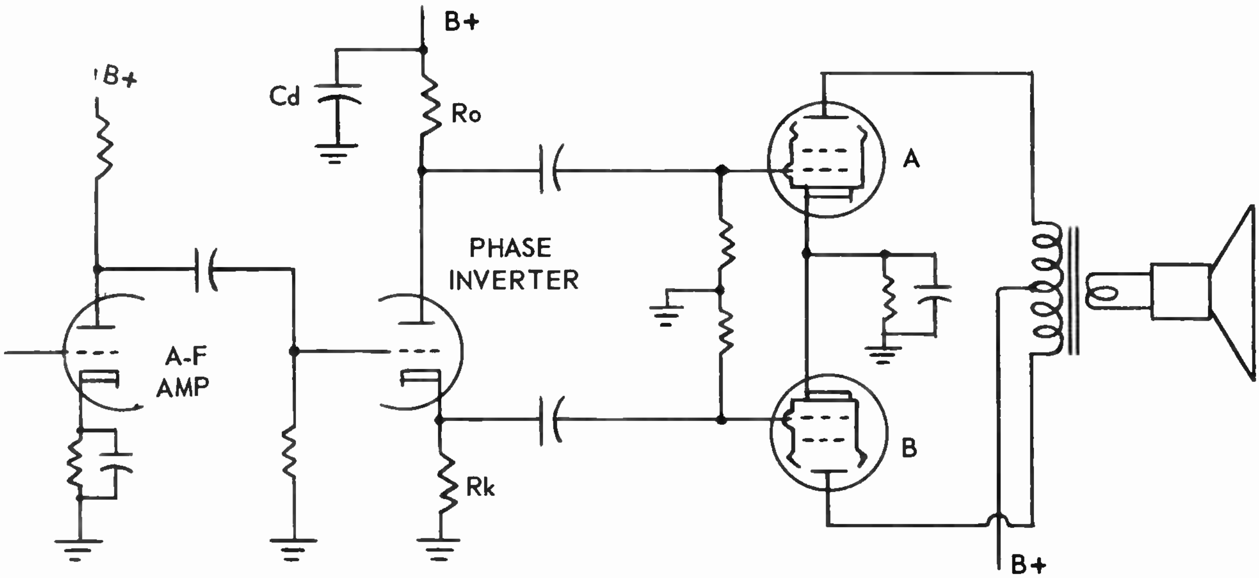


Fig. 81-7. A phase inverter delivering push-pull grid voltages from its plate and cathode.

as to have very little reactance at audio frequencies. Resistances at Ro and Rk are made equal. Then signal voltages of equal amplitudes are applied to the two push-pull grids, although these signal voltages are of opposite phase.

There is no voltage gain on either side of the phase inverter tube when the circuit is constructed as described. On the contrary there is a slight loss of signal amplitude, and signal voltage at the push-pull grids is slightly less than from the plate of the a-f voltage amplifier which precedes the inverter. The inverter tube merely replaces the interstage transformer shown in Fig. 81-4.

In Fig. 81-8 are shown a few of the many inverter circuits used in connection with diode detectors for a-m sound signals. In each diagram the two leads to the grids of the two push-pull amplifier tubes are marked A and B. In diagram 1 the combined detector and a-f voltage amplifier is a dual diode triode tube. The detector circuit is of the type with which we are familiar. Voltage divider resistors are marked Ra and Rb, as in earlier figures. The inverter tube is a dual diode triode type, but the two diode plates are grounded so that they take no part in the action. Oftentimes the triode section of a combination tube has the characteristics desired for some certain purpose, and the diodes are not needed or used. There is the incidental advantage that the same type of tube is used in two positions, reducing the number of types needed for servicing and also allowing an interchange of tubes in case the active diodes in one of them should give out while the triode section still functions.

Diagram 2 shows another dual diode triode tube for detector and a-f voltage amplifier with a triode as the inverter. Voltage divider resistors again are marked Ra and Rb.

In diagram 3 a dual diode triode tube is used as a combined diode detector and inverter instead of a

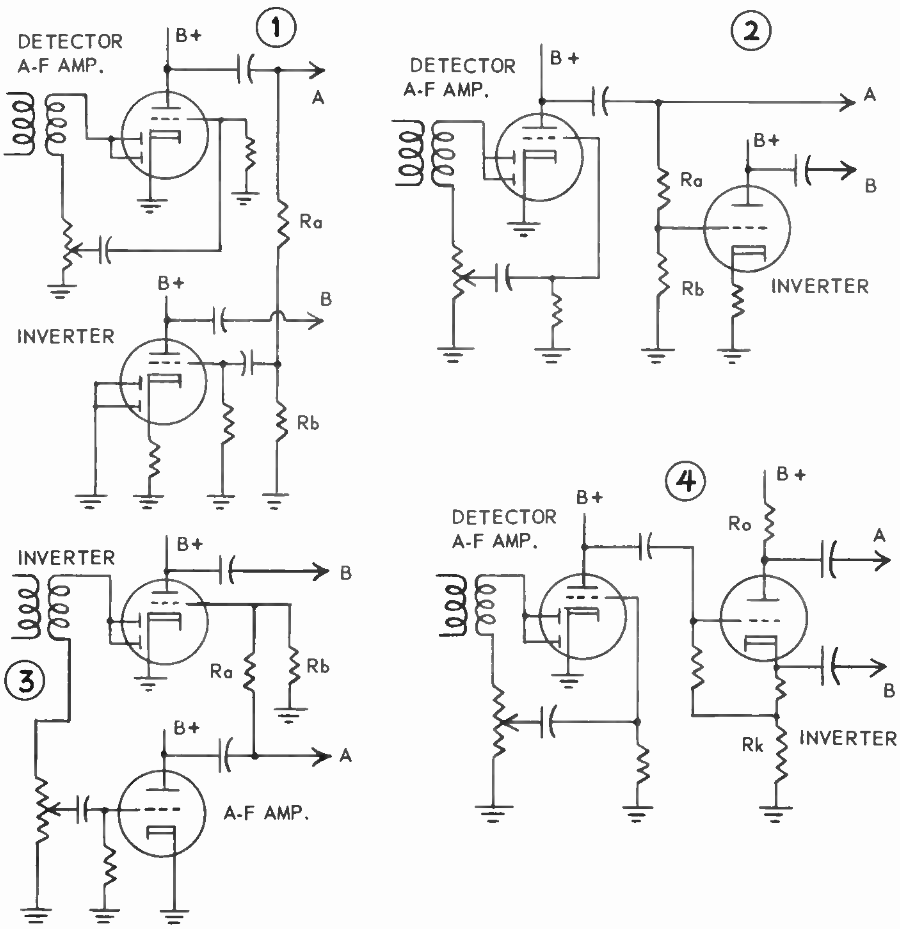


Fig. 81-8. Various ways of using push-pull inverters in connection with a-m diode detectors.

detector and a-f amplifier. Here the a-f voltage amplifier is a separate triode, although it might be a dual diode triode type with the diode plates unused. Sometimes the diode plates that are not used for detection are employed in an automatic gain control circuit, or for any other purpose. Voltage divider resistors are marked Ra and Rb, as in all other diagrams illustrating the use of a divider for the inverter signal.

Diagram 4 of Fig. 81-8 shows a dual diode triode tube as the detector and a-f amplifier. The inverter is a triode feeding one push-pull grid from its plate and the other push-pull grid from its cathode. Resistor Rk consists of two sections. The section next to the cathode provides a voltage drop giving a suitable negative grid bias for the tube. The section connected to ground adds enough extra cathode resistance for inverter action, so that total resistance at Rk is approximately equal to that at Ro or so that plate and cathode signal voltages are of suitable amplitudes for the push-pull grids.

Average plate current, or current taken from the B power supply, and also the screen current taken from this supply, will be twice as great for the two tubes in a push-pull amplifier as for one similar tube in a single-tube amplifier. This assumes, of course, that plate and screen voltages are the same in both cases. Push-pull operation will give twice the audio power output of a single similar tube only when the amplitude of the grid signal voltage is twice that on the single tube. Parallel operation of two power tubes allows twice the audio power output when the grid signal voltage is the same as that applied to a single tube. The need for greater input signal voltage is the only operating disadvantage of push-pull amplifiers as compared with parallel and single tube amplifiers for power.

Upon examining the connections to the primary winding of the output transformer used for push-pull amplifiers you will realize that plate currents for the two tubes flow in opposite directions in the two parts of the secondary which are on either side of the center tap. When there is no audio signal on the grids, these plate currents remain of constant value. The magnetizing effect of one plate current tends to produce a magnetic flux in one polarity, while the effect of the other plate current tends to produce a flux of opposite polarity. Actually, the two magnetizing effects cancel each other, and with no signal there is no flux in the transformer core. This removes the possibility of core saturation which could result from excessive d-c plate current, and allows using a relatively small and inexpensive output transformer, or allows using much greater plate currents with any size of transformer.

If there are ripple voltages or hum voltages from the B power supply these voltages cause equal and opposite magnetizing effects in the two halves of the output transformer primary, if the two tubes are well enough matched to take equal plate currents. In any case there is a decided reduction of hum effects. Plate current for the push-pull tubes may be taken from a point on the power supply filter which is closer to the rectifier, and at which there is more ripple voltage, than at takeoff points for other amplifier tubes. Any rapid fluctuations of supply line voltage which affect the power supply output voltage are wholly or partially compensated for by the same balancing magnetization effects that reduce hum voltage. If hum voltages are picked up in parts of the receiver ahead of the push-pull stage they will act like signal voltages and will not be reduced by any compensating action in the push-pull stage.

Since the two push-pull tubes take twice as much plate and screen current as would a single similar tube with equal d-c voltages on the elements, the cathode bias resistor for the two push-pull tubes need be of only about half the resistance that would be used for equal grid bias voltage on a similar single tube. Twice the cathode current in half the resistance gives the same voltage drop and same biasing voltage. Biasing resistances somewhat greater than half the value for a single similar tube are often used in practice, but excessive bias may cause plate current cutoff on strong signals.

Cathode resistor bypasses usually are of large capacitance, often of 10 to 20 mf. This large capacitance lessens the possibility of a degenerative feedback through the biasing resistance, and preserves maximum amplification. If the two push-pull tubes and other circuit parts were perfectly matched, the changes of plate current due to audio signals would be not only opposite but exactly equal for the two tubes. Then there would be no variation whatever in the current taken from the power supply and in the current through the biasing resistor on the tube cathodes. Under such ideal conditions there would be no need for a cathode bypass capacitor, for there would be no audio frequency currents to be bypassed around the biasing resistor. Sometimes a moderate amount of degeneration may be desired for improvement of tone quality. Then the cathode bypass capacitor may be omitted.

Recommended load resistances or impedances for two tubes in push-pull normally would be double the values recommended for a single similar tube operated with the same d-c voltages. Actually it is possible to use much smaller load resistances, possibly one and one-half times that for a single tube or even less. Such a reduction of load resistance or impedance allows obtaining more than twice as much audio power output as from a similar single power tube without excessive distortion.

REDUCTION OF HARMONICS. When an amplifier tube is operated with a bias more negative than normally used, or when a signal applied to the grid is stronger than usual, the results are as shown by Fig. 81-9. Positive alternations of the signal voltage receive linear amplification, meaning that changes of plate current follow the changes of signal voltage so far as waveform is concerned. Only a small portion of the negative alternations of grid voltage is amplified. The remainder makes the grid so negative that plate current drops to zero. There is plate current cutoff.

The plate current waveform and the waveform of an output signal voltage produced by changes of plate current will be badly distorted and quite unlike the grid signal. The distortion lies chiefly in the flattened negative loops of plate current. When an engineer looks at a plate current waveform of this nature he knows that there are strong second harmonic frequencies in addition to the fundamentals. The second harmonics do not exist in the grid signal voltage. They are formed in the plate circuit by working the tube beyond plate current cutoff.

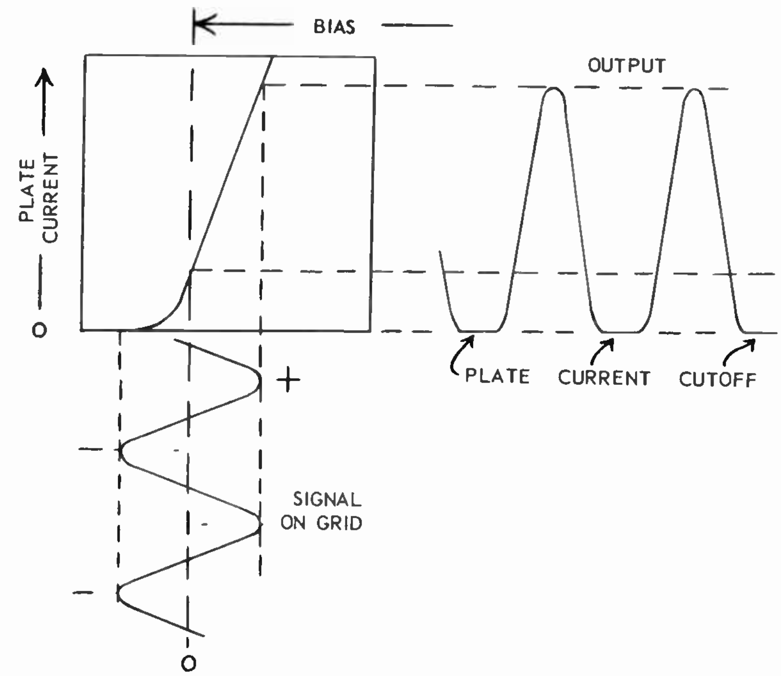


Fig. 81-9. Plate current cutoff may result from grid bias insufficiently negative or from grid signals which are too strong.

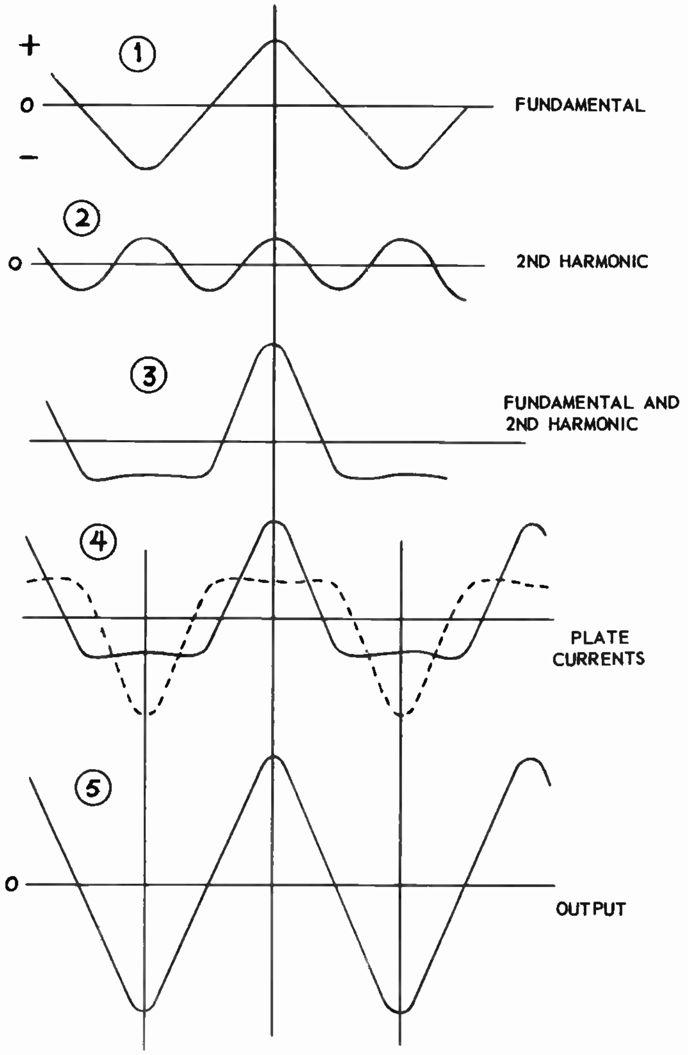


Fig. 81-10. How a second harmonic affects a combination waveform, and how push-pull operation removes the distortion.

How we know that the plate current waveform with flattened negative loops means the presence of a second harmonic frequency is shown by Fig. 81-10. At 1 is shown or represented a fundamental frequency, which may be a frequency of any number of cycles per second. At 2 is represented the second harmonic of the fundamental frequency. The harmonic frequency is twice that of the fundamental.

If you add algebraically the positive and negative values at many instants of time during the formation of fundamental and second harmonic frequencies the resulting waveform will be as at 3. When we add algebraically it means adding together the two values, when both are either positive or negative. To find resulting positive or negative values for the combined wave, it means subtracting one value from the other, when they are not of the same polarity.

Comparing the waveform at 3 in Fig. 81-10 with that of the output current in Fig. 81-9 shows that the two waveforms are essentially alike or, at least, that they have the same general characteristic. We have proved that the waveform at 3 contains a second harmonic, so we may conclude that the similar current waveform also contains a second harmonic.

Both tubes in a push-pull amplifier receive the same grid signal voltages, and both operate with the same grid bias and with equal plate and screen voltages. If one tube produces a second harmonic in its output the other tube will do the same. Signal variations of plate currents in the two push-pull tubes are of opposite polarity at any given instant, and they also are 180 degrees or a half-cycle out of phase with each other.

The out-of-phase variations of distorted plate currents are shown together at 4 in Fig. 81-10. The full-line curve represents the same current as shown at 3, which might be the current in either of the push-pull tubes. The broken-line curve shows simultaneous variations of distorted plate current in the other push-pull tube. The polarities are opposite and 180 degrees out of phase.

If you add algebraically the values in the two waveforms shown at 4 the combined values will give the waveform at 5. This is just the way in which values in the primary of the push-pull output transformer add their effects in the secondary of this transformer. The effects of second harmonic distortion have disappeared. We have now a fundamental frequency like the one at 1 of Fig. 81-10, except that the amplitude is much greater.

All even harmonics would be balanced out in the plate circuits of the push-pull tubes. That is, the effects of second, fourth, sixth, and other alternate harmonic frequencies would disappear.

The third, fifth, and other odd harmonics do not cancel out in the push-pull plate circuit. If you wish graphic proof of this statement draw a set of curves similar to those of Fig. 81-10, but substitute a third harmonic frequency for the second harmonic at 2. The combined waveform will be distorted. The distortions will be of the same form on both sides of the zero line, or there will be "mirror" distortion. When the distorted currents are placed in opposite polarity and 180 degrees out of phase, then combined as the final steps, the distortion will remain.

The fact that even harmonics caused by very strong grid signals are cancelled in the push-pull plate circuits allows push-pull amplifiers to be operated with very strong applied signals and without excessive distortion. Such operation will be considered in following paragraphs.

CLASSIFICATION OF AMPLIFIERS. There is a generally recognized system or code of letters and numbers for specifying the manner in which amplifiers are operated. The system is simple, for it uses only the capital letters A, B, and C, and the numbers 1 and 2 as subscripts to the letters. The three letters refer to biasing and its effects; whether or not the biasing allows plate current cutoff, and when there is cutoff, to the portion of the signal cycle during which it may occur. The numbers refer to grid current. If a letter or letters are followed by the subscript number 1 it means that the grid never becomes zero or positive and that there is no flow of grid current under any conditions. If the number is 2 it means that grid current does flow during some part of each signal cycle.

Class A operation is illustrated by Fig. 81-11. Grid bias is of such value as to bring the zero signal point to the approximate center of the straight portion of the characteristic curve. Signal voltage must be limited to an amplitude which brings the peaks onto neither the lower nor the upper bends of the characteristic; they must not approach plate current cutoff nor plate current saturation. Because there is no cutoff, plate current flows at all times. Since the positive peaks of signal voltage must not overcome the negative bias, but must leave the grid appreciably negative, there is no flow of grid current at any time.

Because there is no grid current with class A operation this method may be called class A1. The letter A was used alone before adoption of the subscript number system. Class A operation and class A1 operation now mean the same thing.

a When tubes are operated in this manner the waveform of output plate current and of voltage in the load is like the waveform of input signal voltage. Single output tubes, as distinguished from push-pull, may be operated only as class A or A1 when distortion is to be kept reasonably low. This applies to power triodes, power pentodes, and beam power tubes. All three types of power tubes are also operated class A or A1 in push-pull amplifiers where the object is to obtain more power than from a single tube and to have minimum distortion. Degeneration often is used with either single tubes or push-pull tubes in this operating class.

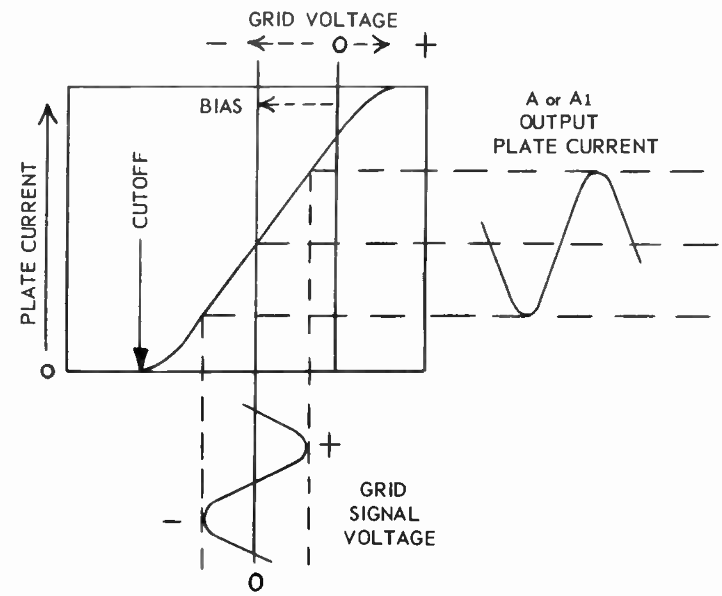


Fig. 81-11. Input and output signals with class A or A1 operation.

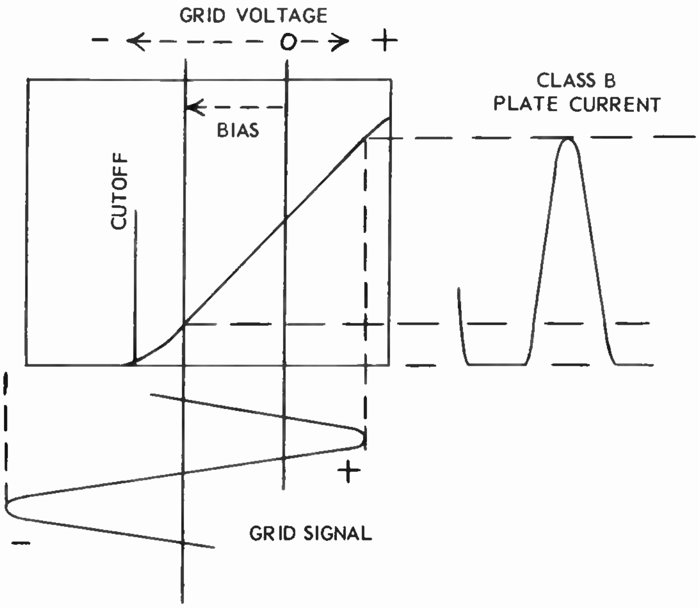


Fig. 81-12. There is plate current cutoff with class B operation.

Class A or A1 operation utilizes B power in the plate circuits less efficiently than with any other operating method. This is because the average plate current and the plate current at zero signal voltage is of rather high value, as you can see from Fig. 81-11. In other methods of operation you will see that average plate current is less, because there is cutoff during parts of the signal cycles, although audio power output may be greater.

The operating characteristics of push-pull power amplifiers as described earlier in this lesson will hold true when the amplifiers are operated class A or A1. With operation in other classes the plate currents in the two tubes do not remain equal and opposite at all times. Then there is not a constant current from the B-power supply and through cathode resistors on the tubes, and there are other differences in performance.

Were the signal voltage amplitude sufficiently increased, the positive peaks of the signal would exceed the negative bias and the grid would become positive during these signal peaks. Then, provided that the negative signal peaks did not cause plate current cutoff, we would have class A2 operation. The same thing could happen were the bias made somewhat less negative than for class A or A1 operation, for then the original signal amplitude could overcome the less negative bias to make the grid momentarily positive and to allow pulses of grid current.

Class B operation is illustrated by Fig. 81-12. Bias is sufficiently negative to bring plate current

(1) down almost to the cutoff point when there is no signal or zero signal voltage. The signal applied to the grid has much greater amplitude or swing than with class A operation. On the positive half-cycles of signal the amplitude more than equals the negative bias. Then, during a greater or less portion of these positive half cycles there is flow of grid current. Because there always is grid current it might seem as though this method of operation should be called B2, and probably it could be. However, there is no B1 class (with no grid current) and so there is no need for adding the subscript number.

During the greater portion of the negative half-cycles of grid signal the grid voltage is driven more negative than for plate current cutoff. Then the negative loops of output current are flattened off as shown in the figure. This flattening is like that in Fig. 81-9, and it indicates the production of second and other even harmonics in the plate circuit. As a consequence, class B operation is used only with power tubes in push-pull, which allows cancellation of the even harmonics. The push-pull tubes most often are triodes, though both power pentodes and beam power tubes may be operated class B.

Class B operation allows much greater audio output power than classes A, A1, or A2 with the same tubes and the same consumption of B power. Class B power efficiency is greater. Heat dissipated in the power tubes is less in relation to audio output power because plate current flows during only a little more time than in the positive half-cycles of input signal voltage. The B power supply must have good voltage regulation, because with plate current cutoffs there is not the exact balancing of the two plate currents as found when plate current flows during the entire signal cycle, and current from the power supply varies with signal amplitude.

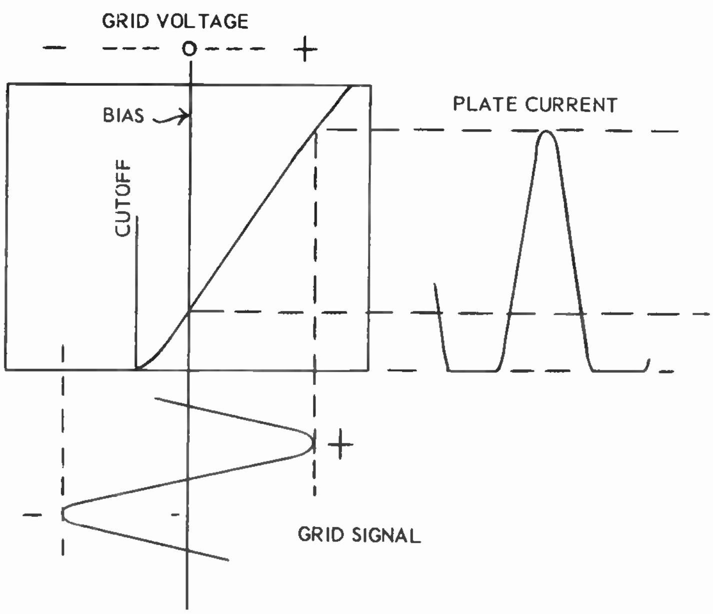


Fig. 81-13. Operation of a zero-bias class B amplifier tube.

There are a number of tubes especially suited for class B operation. They are designed to work with zero grid bias. Among these types are the 6AC5G single triode and the 6N7 and 1635 which are twin triodes with a single cathode. All these tubes are high-mu types, with high amplification factors and with closely spaced grid wires. The close spacing gives enough control of plate current so that this current is of only moderate value, often only 5 to 10 milliamperes, when the bias is zero and plate potential is several hundreds of volts.

Operation with zero grid bias is illustrated by Fig. 81-13. There is plate current cutoff during a large part of the negative half-cycle of input signal voltage. There is grid current during the entire positive half-cycle of signal voltage. In spite of this apparently wrong method of operation a zero grid class B push-pull stage will deliver 10 or more watts of audio power with total harmonic distortion of only 5 to 8 per cent.

With class B operation we have plate current flowing for only approximately half of each signal cycle, and with cutoff during nearly all of the other half. With class A operation the plate current flows during the entire signal cycle. It is not at all necessary to go to either extreme, we may work the tube or tubes in such manner as to have plate current during considerably more than the positive half-cycle of signal voltage, yet not during the whole time. This is called class AB operation; it is in between class A and class B.

One kind of class AB operation is illustrated by Fig. 81-14. The grid bias is more negative than for class A operation, possibly one fifth to one-third more than with the same tube operated class A or A1,

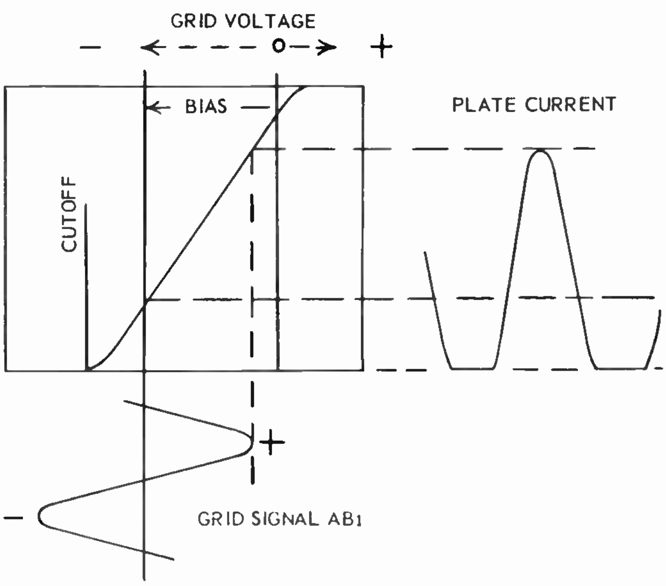


Fig. 81-14. With class AB1 operation there is plate current cutoff, but no grid current.

but it is less negative than for class B. There is plate current cutoff during part of each signal cycle, so class AB operation can be employed only for push-pull amplifiers where the even harmonics cancel. The tubes may be triodes, power pentodes, or beam power types. Smaller sizes of tubes can be used than with class A for equal power outputs, because the lessened time during which there is plate current reduces the heating and the necessary power dissipation ratings of the tubes. Each class AB push-pull tube may take average plate current only about two-thirds that for a single class A tube.

In Fig. 81-14 the positive peak amplitudes of signal voltage are less than the negative bias voltage. The grid remains negative at all times and there is no grid current. To indicate the absence of grid current we call this method of operation AB1. In Fig. 81-15 we still have the same type of grid biasing, in between that for types A and B, but the applied grid signal is strong enough that its positive peaks exceed the negative bias. Then the grid goes momentarily positive during these signal peaks and there are pulses of grid current. The presence of grid current is indicated by calling this class AB2 operation.

Here is something of importance. If you have any particular tubes working with a given bias and with given d-c plate and screen voltages, and apply a very weak signal to the grid of a class AB stage the operation actually will be class A. The weak signal will not cause either plate current cutoff or grid current. If signal strength increases to some extent the stage will operate class AB1, for the stronger negative peaks of the signal will cause plate current cutoff during small parts of each signal cycle. Should the applied signal become still stronger its positive peaks will rise enough to overcome the grid bias and allow grid current. Then the stage operates as class AB2. All this happens without change of element voltages, only because the signal changes in strength.

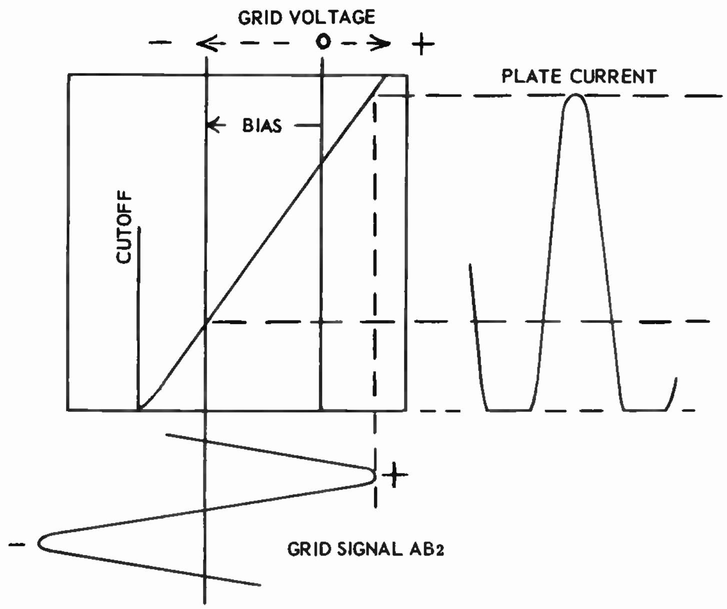


Fig. 81-15. With class AB2 operation there is plate current cutoff, also grid current.

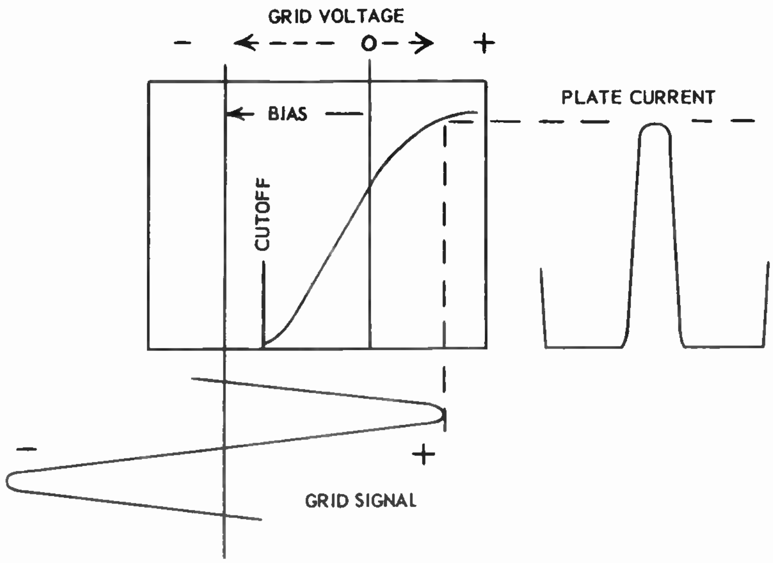


Fig. 81-16. The distorted output signal resulting from class C operation.

The audio power output for from a class AB1 push-pull stage can be greater than from a class A or A1 push-pull stage, and from a class AB2 stage it can be greater than from the class AB1 stage, always provided the input signal is strong enough to work the push-pull amplifier in the class considered. An AB1 push-pull stage can deliver one and one-half to two and one-half times as much audio power as a single similar tube operated class A or A1. Considering the same tubes operated with the same plate and screen voltages, a class AB2 push-pull stage can deliver two and one-half to four times as much audio power as a single class A amplifier. Don't forget, however, that these big powers call for strong input signals. The comparison means only that the push-pull stages mentioned are capable of handling the strong signals and of delivering high audio powers with distortions not exceeding and usually much less than with the single tube. It is not a matter of signal amplification, but only of signal and power handling ability.

10 Class C operation is illustrated by Fig. 81-16. The negative bias is greater than for plate current cut-off, and when there is no grid signal there is no plate current at all. The grid signal is so strong that its positive peaks drive the grid far positive, even to where there is plate current saturation and a flattening out of the characteristic curve. This flattens the tops of plate current alternations. Plate current flows only in pulses, between relatively long periods of cutoff. Class C operation causes such distortion that it cannot be used for audio amplifiers, even with push-pull stages. This method is used for r-f oscillators and for r-f amplifiers in the tuned circuits of some transmitters and for a few other special applications.

When audio amplifiers are operated in any of the classes allowing plate current cutoff, the current from the power supply and through the tube cathodes is not constant. When cathode bias is used with any of these amplifiers it is necessary to provide a large bypass capacitance to get the plate current variations around the biasing resistor and prevent these variations from adding themselves to the grid voltage. Capacitances of 50 mf or more are common. When fixed bias is used it is necessary to allow large bleeder current in the bias resistors on the power supply filter system; otherwise the bias voltage will fluctuate with variations of signal plate current. It is necessary also that the power supply have good voltage regulation. This may be brought about by using a choke input filter, by using chokes having low d-c or ohmic resistance, and a power transformer with a low-resistance plate winding.

DRIVER STAGES. When a push-pull stage is so operated that there is grid current the tube itself acts like a resistance in parallel with the input circuit. The greater the grid current the lower becomes this effective resistance, because more current always goes with less resistance in any kind of device. Power is required for the grid current flow within the tube, and also through the parts of the input circuit, just as power is required to maintain flow of plate current or screen current.

This grid circuit power must be furnished from the amplifier stage which precedes the push-pull stage. To reduce the required grip power everything in the input circuit must have the least possible resistance or impedance. This prohibits the use of resistance coupling and inverter circuits between a preceding amplifier and the push-pull stage in all except a few special applications in the AB2 classification.

The preceding tube, which must furnish power as well as signal voltage, is called a driver tube. Coupling to the push-pull stage is through a driver transformer. The secondary of this transformer is part of the push-pull grid circuit, and as such must have low impedance. The primary of this transformer is in the plate circuit of the driver tube, and must have enough impedance to form a reasonable load for this tube. This means that the driver transformer nearly always has a step-down turns ratio and voltage ratio from primary to secondary.

The driver tube itself should have the lowest possible plate resistance, so that transformer primary impedance won't have to be too great. Suitable driver tubes include power amplifier triodes, also power pentodes with their plates and screens connected together to act as a triode. Unless all these precautions are observed in the driver stage and transformer it is likely that there will be troublesome harmonic frequencies produced in the grid circuit of the push-pull stage.

TELEVISION

LESSON NO. 82

MEASUREMENTS IN AUDIO AMPLIFIERS

Training Your Mind...
C.T.I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind
C.T.I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON NO. 82

MEASUREMENTS IN AUDIO AMPLIFIERS

During service operations on receivers of all kinds and also on separate audio amplifiers it often is necessary, or at least desirable, to measure the audio power output. A rather simple method of doing this is illustrated by Fig. 82-1. The equipment is as follows. An r-f signal generator with audio-frequency modulation. The modulation frequency most often is approximately 400 cycles, so measured power output will be at this frequency. A vacuum tube voltmeter with an a-c range of zero to 5 or zero to 10 volts. An ohmmeter capable of accurately measuring resistances of 3 to 15 ohms. An adjustable non-inductive resistor, such as a rheostat or potentiometer, with maximum resistance of less than 20 ohms and a power rating at least equal to the maximum rated power output in watts of the power tube or tubes in the apparatus tested.

Connections are made as in Fig. 82-1. To one of the terminals at which the output transformer secondary is connected to the speaker voice coil, connect one lead from the VTVM and also one end of the adjustable resistor. Disconnect the other voice coil lead from the other secondary terminal, and connect to the transformer secondary the remaining lead from the VTVM. To this same secondary terminal tack solder or otherwise attach a short wire which may be connected alternately to either the voice coil or the remaining terminal of the adjustable resistor. Connect the r-f signal generator to the receiver antenna input in any way that would be suitable for alignment of the receiver. The remaining steps are:

1. With the transformer secondary connected to the voice coil, and to the VTVM, turn on the receiver and instruments, and let them warm up. Tune the receiver to a frequency where no broadcast station is heard, and with the volume control at a medium setting tune the signal generator for maximum reading on the VTVM. Do not alter the receiver and generator tuning dial settings during the remainder of the work.

2. Turn the volume control to maximum, note the VTVM reading, then reduce the volume for a reading of about 80 per cent of maximum, this to avoid distortion effects. Note the new VTVM reading very carefully.

3. Change the transformer connection from the voice coil to the adjustable resistor. If you don't want to do this with power on, pull the plug of the receiver power cord, but do not use the on-off switch, which usually is on the volume control.

4. With the receiver turned on and warm, adjust the resistor until the VTVM reads exactly the same voltage as in step 2.

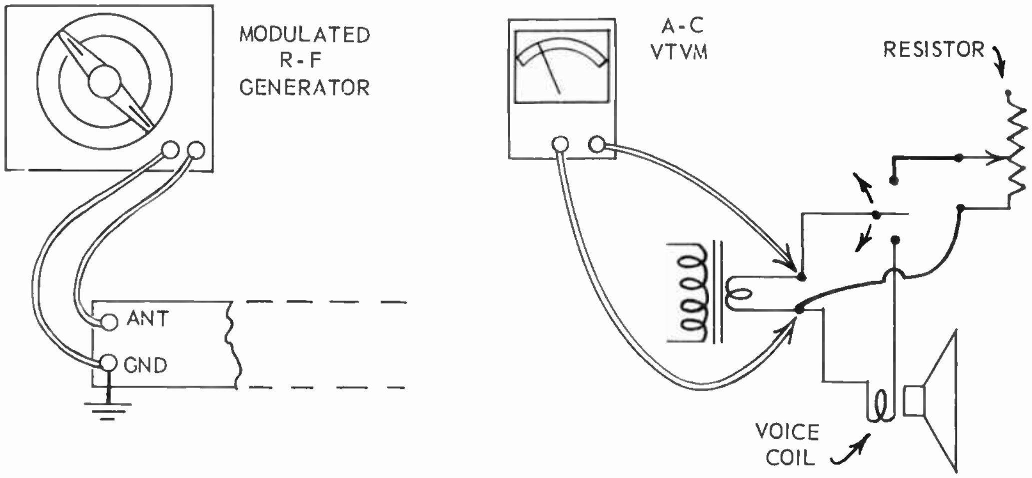


Fig. 82-1. Test connections for audio output power measurement with modulated signal generator, VTVM, and adjustable resistor.

5. Without touching its adjustment shaft, disconnect the resistor and measure its adjusted resistance with the ohmmeter. One side of the resistor may remain connected to the transformer, the voice coil, and the VTVM, while the other side is disconnected. The measured resistance in ohms is equal to the effective impedance of the voice coil at the audio frequency of generator modulation. Here we have, incidentally, a method of measuring voice coil impedance under operating conditions.

6. Square the number of volts measured in steps 2 and 4, which is the same number in both cases. Divide the result by the number of ohms measured in step 5. The quotient of this division is the power in watts being put into the voice coil.

It is advisable to make voltage and resistance measurements at several settings of the receiver volume control. The measured resistance, and voice coil impedance, should remain almost unchanged in all the tests. Otherwise you are using too high a volume setting and are getting severe distortion.

1 A vacuum tube voltmeter should be used in preference to an a-c rectifier voltmeter or an output meter such as used for output measurements during alignment of a sound receiver. These other meters have relatively low sensitivity, seldom more than 1,000 ohms per volt for a-c work, and their proportionately low voltage readings will cause computed power to be considerably less than the actual value. Also, some a-c voltmeters are accurate enough at power line frequencies but not at higher audio frequencies.

An audio-frequency signal generator may be used with connections shown in Fig. 82-2, instead of the modulated r-f signal generator. Connections for the VTVM and adjustable resistor are made as shown at the right in Fig. 82-1. The high-side lead from the audio generator may be connected to the slider of the volume control, at a, in the receiver or an audio amplifier. The volume control is set at its maximum point. Signal strength will be adjusted by the attenuator of the audio generator. If the generator is connected to

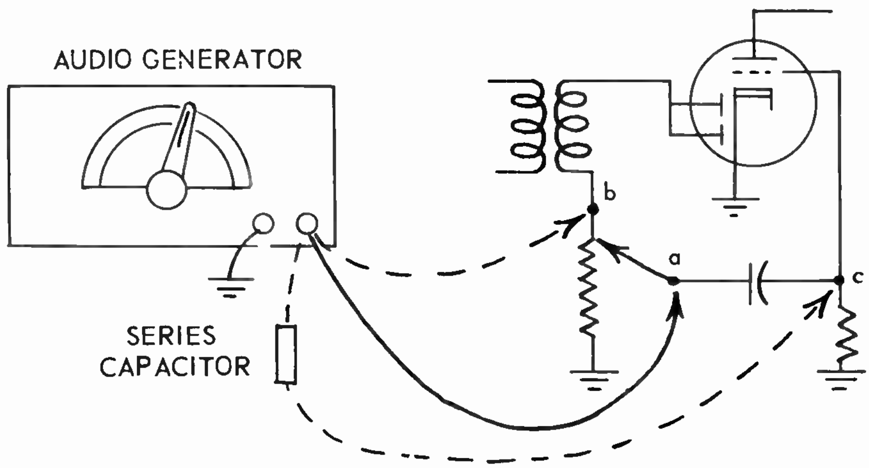


Fig. 82-2. Connections for measuring audio power output when using an audio-frequency signal generator.

the top of the volume control, at b, the signal strength to the a-f amplifier grid may be regulated either by the generator attenuator or the volume control. If the generator is connected directly to the a-f amplifier grid, as at point c, there must be a series capacitor in the lead. If there is no such capacitor built into the generator it is necessary to use a fixed paper type of about 0.01 or 0.02 mf connected externally as shown.

The audio power measurement is carried out with the same steps outlined for the modulated r-f generator method. The audio generator allows measurements at various frequencies. If the generator frequency

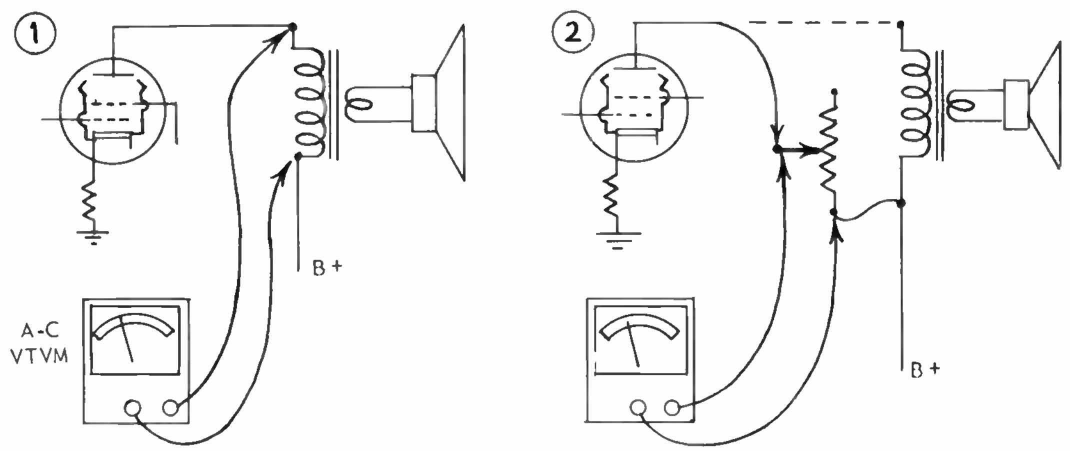


Fig. 82-3. Correct and incorrect methods of measuring audio power in the plate circuit of an amplifier tube.

is gradually reduced below 400 cycles while watching the reading of the VTVM you will find a voltage peak for some frequency adjustment. This is the bass resonant frequency of the speaker or voice coil.

Power output in the plate circuit rather than in the voice coil circuit may be measured as shown at 1 in Fig. 82-3 provided you know the actual impedance of the output transformer primary winding at the audio frequency to be used. This impedance sometimes is marked on the transformer, usually for 400 cycles or for 1,000 cycles. Measure the a-c voltage drop across the transformer primary, square the number of volts, and divide the result by the number of ohms impedance. This gives the power output in watts.

At first thought it might seem possible to measure plate circuit power output and transformer primary impedance with the same general method employed in Fig. 82-1, by temporarily disconnecting the primary and substituting an adjustable resistor, as at 2 in Fig. 82-3. This will not work out. When the resistor is adjusted for the same voltage drop as obtained across the transformer primary, the ohmic resistance of the resistor is much greater than the ohmic resistance of the primary. This greater resistance drops the plate voltage at the tube below its value when using the primary, and all the operating conditions are changed.

POWER OUTPUT FROM LOAD LINES. It is possible to compute the theoretical power output by drawing a load line on a family of plate characteristics, then measuring plate currents and voltages at operating points on the load line. To illustrate the method we shall use the plate characteristics of Fig. 82-4, which apply to one section of a 6SN7 twin triode voltage amplifier. A triode voltage amplifier rather than

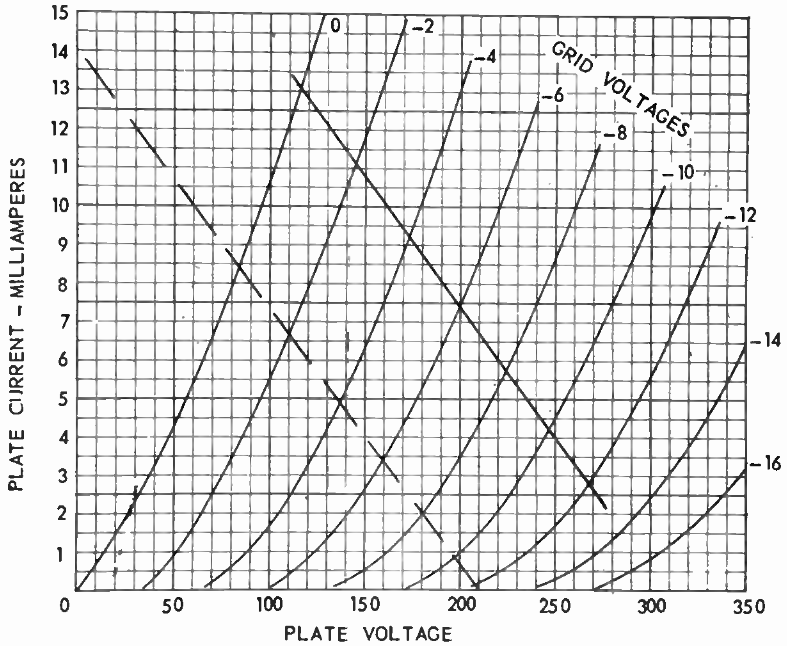


Fig. 82-4. How a load line may be used for determining power output of a triode operated as a class A or A_1 amplifier.

a triode power amplifier has been selected to show how power output may be computed for any kind of triode, whether or not it is intended for a power amplifier. The steps are as follows.

1. Select the plate voltage and negative grid bias to be used. These voltages must, of course, be within the maximum ratings for the tube. We shall use 200 volts at the plate and a negative grid bias of 6 volts. On the curve for the selected bias place a mark at the selected plate voltage. This allows reading the plate current when there is no signal; in this case it is very close to 7.5 ma.

2. Select a plate load resistance or impedance to be used. For our example we shall use 15,000 ohms.

3. Through the point of zero signal current determined in step 1 draw a load line for the selected load resistance or impedance. Methods of drawing load lines are explained in another lesson.

Here is a rather easy way to draw a load line. Divide the load ohms by 1,000 to determine the slope of the line in volts per milliampere. Dividing 15,0C0 by 1,000 gives 15 volts per milliampere. Now select any fairly high current shown on the plate current scale of the graph. We shall take 14 ma. Multiply this by the volts per milliampere, which, in our example, means 14 times 15, or 210 volts. Draw a line from the number of milliamperes (14) to the number of volts just determined. This is the broken line of Fig. 82-4. Any other line drawn parallel to this one is a load line for 15,000 ohms. Draw a parallel line through the point of zero signal current (found in step 1). This is the full line on Fig. 82-4, and is the load line with which we are to work.

4. Select the peak-to-peak audio signal voltage to be applied to the grid. We shall use 12 volts peak-to-peak, which will swing the grid 6 volts each way from the zero signal point. Mark the ends of the swing. In our example they will be at the intersection of the load line with the zero grid volts curve (6 volts less than the bias voltage) and with the 12-volt curve (6 volts more than the bias).

5. For the two points on the load line found in step 4 determine the two plate currents, and subtract one from the other to find the difference in milliamperes.

6. For the same two points determine the two plate voltages, and subtract to find the difference or change of voltage.

In our example the currents are 13.1 ma and 2.8 ma, approximately. The change of current is 10.3 ma. The voltages are approximately 116 and 268. The change of voltage is 152.

7. Use the changes of current and of voltage in this formula.

$$\text{Watts output} = \frac{\text{milliamperes change} \times \text{volts change}}{8000}$$

$$\text{Watts output} = \frac{10.3 \times 152}{8000} = \frac{1566}{8000} \text{ (approx.)} = 0.196 \text{ watt, approximately}$$

This output of 196 thousandths of a watt would be called 196 milliwatts output.

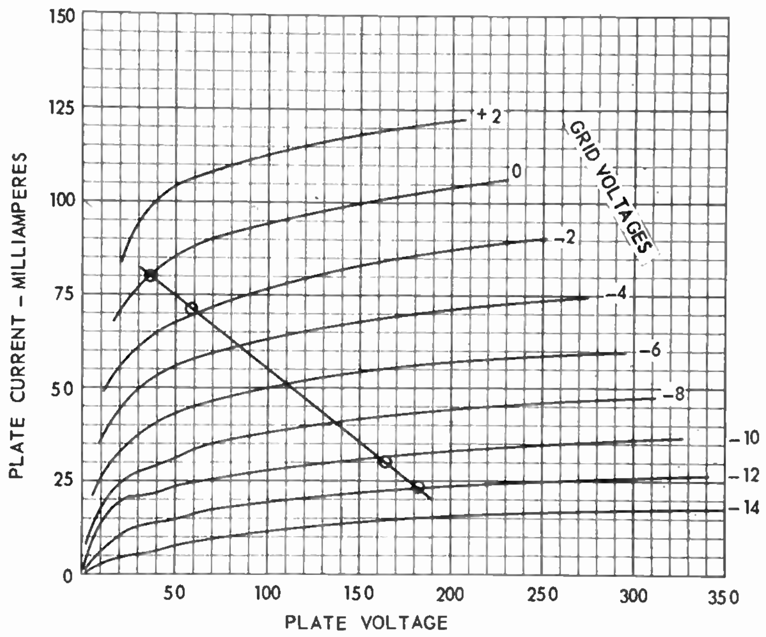


Fig. 82-5. How plate current values are taken from a load line for computing maximum power output of a single class A or A_1 beam power amplifier.

The power output computation just explained applies to a single triode tube operated as a class A or A_1 amplifier. It is very easy to compute the approximate maximum power output for two triodes used in a class A or A_1 push-pull amplifier by using a family of characteristics. For an example we shall use the characteristics of Fig. 82-4.

1. Select the plate voltage and negative grid bias to be used. We shall use 200 plate volts and 6 volts negative bias, as for the single tube in the preceding example.

2. Multiply the operating plate voltage by 0.6. The product of 200 and 0.6 gives 120 volts for our example.

3. On the curve for zero grid voltage read the plate current in milliamperes for the plate voltage found in step 2. This is the maximum permissible plate current, for if we go beyond zero grid bias there will be grid current and class A_2 operation.

4. Use the maximum plate current and the zero signal plate voltage in this formula.

$$\text{Maximum watts two tubes} = \frac{\text{maximum milliamperes x operating plate volts}}{5000}$$

On the zero grid curve we read the maximum current as 13.7 ma at 120 volts. Using this current and the operating plate volts (200) in the formula gives,

$$\text{Maximum watts} = \frac{13.7 \times 200}{5000} = \frac{2740}{5000} = 0.548 \text{ watt, for two push-pull tubes.}$$

The two push-pull tubes here would be the two sections of the 6SN7 twin triode.

Audio power output for a single power pentode or beam power tube with voltages giving class A or A_1 operation may be determined from a load line drawn on a family of plate characteristics for the tube. Be sure that the characteristics apply with the screen voltage actually used. There is a different set of plate characteristics for every different screen voltage on a pentode or beam power tube. For an example we shall use the characteristics of Fig. 82-5, which apply to a 35A5, 35B5, or 35L5 tube with 110 volts on the screen. This is the nominal voltage for both plate and screen in a transformerless receiver having no voltage doubler in the power supply.

1. Select the plate voltage, screen voltage, negative grid bias, and load resistance or impedance to be used. We shall assume 110 volts on the plate, 110 volts on the screen, a negative bias of 6 volts, and the recommended load for our type of tube, which is 2,500 ohms.

2. Draw a load line for the selected load ohms through the intersection of the selected plate voltage line with the curve for selected grid bias. This line is shown on Fig. 82-5.

3. Read one plate current where the load line crosses the curve for zero grid volts. This is 80 ma. Read another plate current where the load line crosses the curve for a grid voltage equal to twice the bias voltage. This would be the 12-volt negative grid curve. Plate current would be about 23 ma.

4. Subtract the smaller from the greater plate current and call the difference maximum Ip (plate current) change. Our maximum Ip change is 57 ma.

5. Read a third plate current on the load line at a grid voltage point equal to 0.3 times the bias voltage. In our example 0.3 times 6 (volts bias) gives 1.8 grid volts. We read a plate current of about 71 ma where we estimate the grid voltage of 1.8 to lie on the graph.

6. Read a fourth plate current at a grid voltage equal to 1.7 times the bias. Here we have 1.7 times 6 giving 10.2 grid volts. As nearly as can be estimated where this grid voltage would come along the load line, the plate current is approximately 30 ma.

7. Subtract the smaller plate current found in step 6 from the larger one found in step 5. Call this difference the optimum Ip change. It is the plate current change caused by an audio signal not so strong as to cause total harmonic distortion in excess of 10 per cent. Out optimum Ip change is 41 ma.

8. Multiply the optimum Ip change (from step 7) by 1.41.

$41 \times 1.41 = 57.81$ A value of 57.8 will be plenty close enough.

Add the maximum Ip change (from step 4) $57.8 + 57 = 114.8$

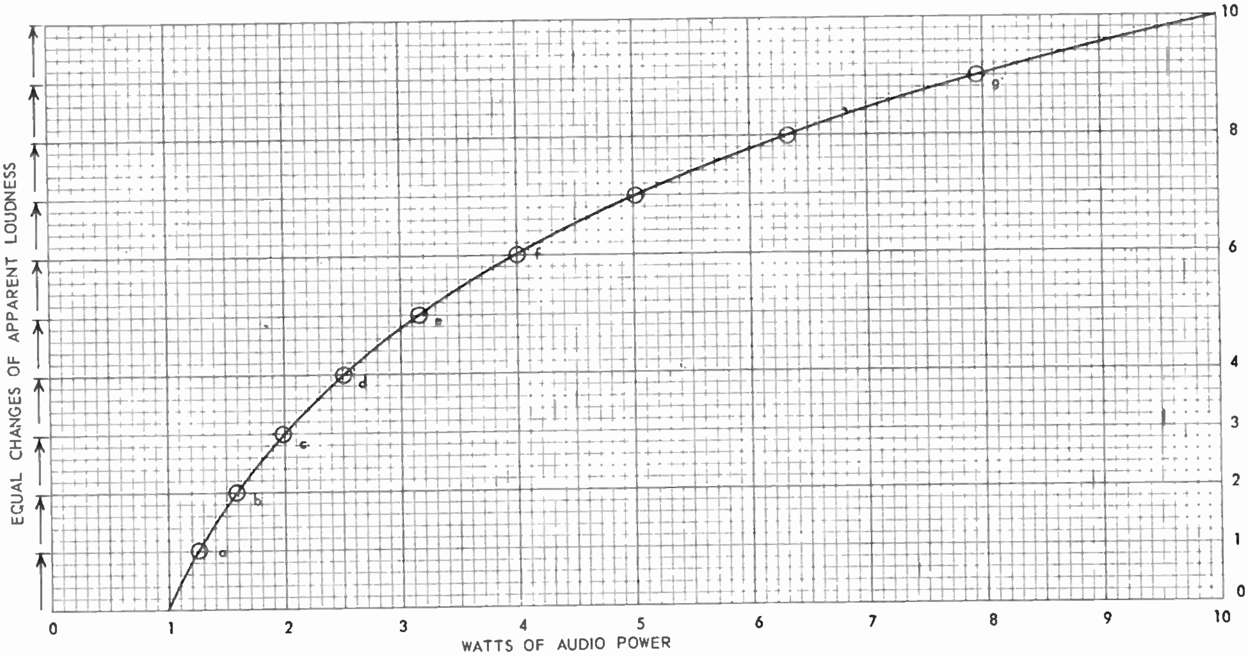


Fig. 82-6. How the average human ear judges the loudness resulting from increases of audio power in watts.

Divide this sum by 1000. $114.8 \div 1000 = 0.1148$

Square the result or quotient. $0.1148^2 = 0.0132$ approximately

9. Divide the number of ohms load resistance or impedance by 32.

$2500 \div 32 = 78.1$

10. Multiply together the final results or quotients of steps 8 and 9.

$0.0132 \times 78.1 = 1.03$ watts (approximate) audio power output.

If you wish to compute power outputs of push-pull power pentodes and beam power tubes, which usually are operated in classes other than A or A_1, the methods can be found in any radio engineering text.

DECIBELS. Our next subject relating to power amplifiers and to sound in general is a very useful way of specifying changes of either the power or the resulting sound volume. How the average human being judges the relative loudness of sounds when all are at the same frequency is illustrated by Fig. 82-6. The vertical scale represents equal increases in apparent loudness of sound, and the bottom scale is for watts of audio power.

We may commence with the loudness produced by 1 watt of audio power. Because this is our starting or reference point for sound volume, we shall call it zero on the loudness scale, meaning that, as yet, there has been no increase of the original loudness or volume. If audio power is gradually increased, our average listener won't notice the change of sound volume until audio power gets up to about 1.25 watt at a. If audio power is now held at a for awhile, to let the listener get used to the sound volume, and then is increased some more, there will be no perceptible change of sound volume until the power reaches about 1.6 watt, at b. If you are the listener, and base your impression of loudness on what happens at b, there won't be another noticeable increase of sound until the power gets up to about 2 watts, at c. It takes a little greater increase of audio watts every time there is a noticeable increase in sound volume.

Now imagine that we switch back and forth between audio powers of 1 watt and 2 watts. The average listener would judge the 2-watt sound to be twice as loud as the 1-watt sound, which is entirely natural and to be expected. But if the same average person listens for awhile to the 2-watt sound, the audio power would have to be raised to 4 watts, at f, in order to make the sound seem twice as loud. An increase of 1 watt doubled the apparent loudness in the first case, but it takes an increase of 2 watts to double it again.

In going from the 2-watt to the 4-watt sound there would be a barely noticeable increase of sound volume at d, with about 2.5 watts of power, another noticeable increase from d to e, where we have about 3.17 watts, and a third noticeable increase from e to f, where we arrive at approximately 4 watts of audio power. If you listen for awhile to the 4-watt sound, the power would have to be increased all the way to 8 watts,

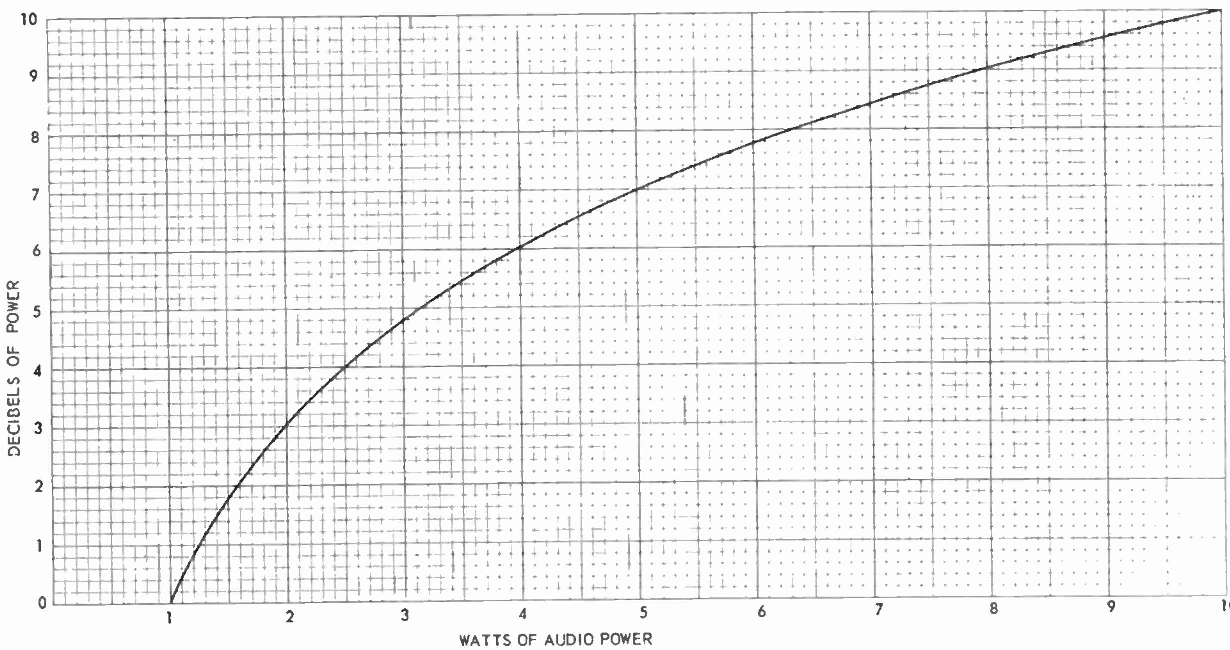


Fig. 82-7. The loudness scale may be graduated in decibels of power, relating decibels to audio power in watts.

at g, for the next apparent doubling of loudness. There would be perceptible increases at the two intermediate points marked on the curve.

There are many kinds of work during which it is more convenient to make measurements in units which are proportional to apparent changes of sound volume, than in watts of audio power. This is true with measurements of audio power outputs, with intensities of audible sounds, with either gains or losses of power in audio amplifiers, with the effects of frequency on gain or loss of either audio power or sound volume, and with almost everything relating to music, speech, and other sound effects.

Because the sensitivity of our ears becomes less and less with increases of actual ear-drum pressure from sound waves, the unit based on ear response avoids such immense numbers as, for example, in the statement that an overall gain is 3,162,000 times. We say the same thing by stating the gain is 65 decibels. We don't have to say that attenuation has dropped the power to 0.000000316 of its original value, we say that there is a loss of 65 decibels.

One decibel is the amount by which sound loudness or volume must be altered to cause a change barely noticeable by the average listener. Therefore, we may put a decibel scale on the graph of loudness versus audio power in watts and have Fig. 82-7. The curves on our two graphs are logarithmic. Substituting a logarithmic power scale for the linear scale of audio watts changes the curve to a straight line, and we have the more convenient graph at the left in Fig. 82-8.

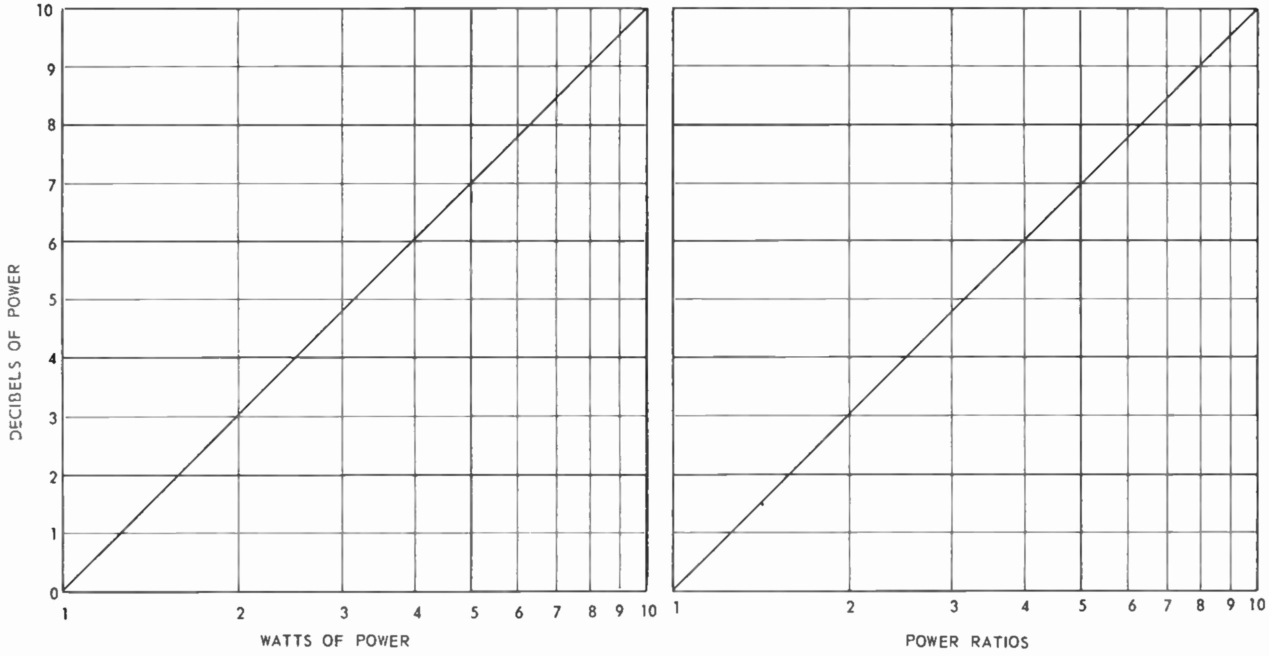


Fig. 82-8. Logarithmic power scales allow making straight line graphs for either watts or ratios of power.

In all these graphs the starting point, at zero decibels, has been taken as a power of 1 watt. The usefulness of this whole system may be greatly extended by substituting for the various numbers of watts the ratios of one power to another. That is, instead of thinking of the number 2 on the power scale as meaning 2 watts, we consider it to be 2 times any amount of power represented at 1, and we think of the number 3 as standing for 3 times whatever power may be represented at 1. On the right-hand graph of Fig. 82-8 the power scale numbers are called ratios. Everything else is exactly the same as on the graph at the left.

Now a power equal to any number of watts or to any fraction of a watt may be assigned to the figure 1 on the power scale. Then at 2 we have 2 times as much power, at 3 we have 3 times as much, and so on. The power in watts assigned to the number 1 becomes our reference power or reference level of power. This does not upset the significance of the decibel in the least. Every change of one decibel in audio power still causes a change in sound loudness just noticeable to the ear.

It is highly important to note that one decibel is not any constant amount of power. The graphs show that a decibel sometimes means only a small change of power, and again means a big change. We cannot say that one decibel equals so many watts or any certain fraction of a watt. Nor can we say that one watt equals so many decibels or any fraction of a decibel. The actual power in watts that corresponds to one decibel depends on two things, on the reference power or reference level that has been assumed, and on how far above or below that reference level this particular decibel happens to be.

When you say that an audio power or a sound volume is "up" a certain number of decibels, you are saying that there have been that number of noticeable changes since leaving the reference level. If the power or volume is "down" a certain number of decibels, there have been that number of perceptible decreases since leaving the reference level. There are simple means for translating numbers of decibels into equivalent watts of power, but only when we know the reference level – not otherwise.

The chart at the right in Fig. 82-8 may be extended to any degree in every direction, to show relations between decibels and powers which are fractions of the reference power and also relations between decibels and powers which are multiples of the reference power.

In Fig. 82-9 the power ratios have been extended down to 1/100000 or 0.00001 of the reference power, whatever it may be, and also up to 100,000 times the reference power. The center of this extended chart is at a ratio of 1 (really 1 to 1) and at zero decibels. This is the lower left-hand corner of the original chart in Fig. 82-8. The first "cycle" on the new chart goes up to a ratio of 10 and to plus 10 decibels, as at the upper right-hand corner of the original chart.

When the power ratio goes up by 10 times anywhere on the large chart we add 10 decibels, and when the power ratio drops by 1/10 anywhere on the chart we subtract 10 decibels or say that the power is 10 decibels farther down. Note that the total change of power in Fig. 82-9, from a ratio of 0.00001 to a multiple of 100,000, is an overall ratio of 10,000,000,000 to 1. The total change of decibels, from 50 down to 50 up, is 100 decibels.

(4) *DECIBEL METERS.* Many a-c voltmeters of the rectifier type and many vacuum tube voltmeters capable of measuring a-c voltages are designed for measurement of power gains and losses in decibels. The only difference between an a-c meter that does not measure decibels and one that does measure them is the addition to the latter type of a decibel scale on the dial. The meter itself may be the same in both cases.

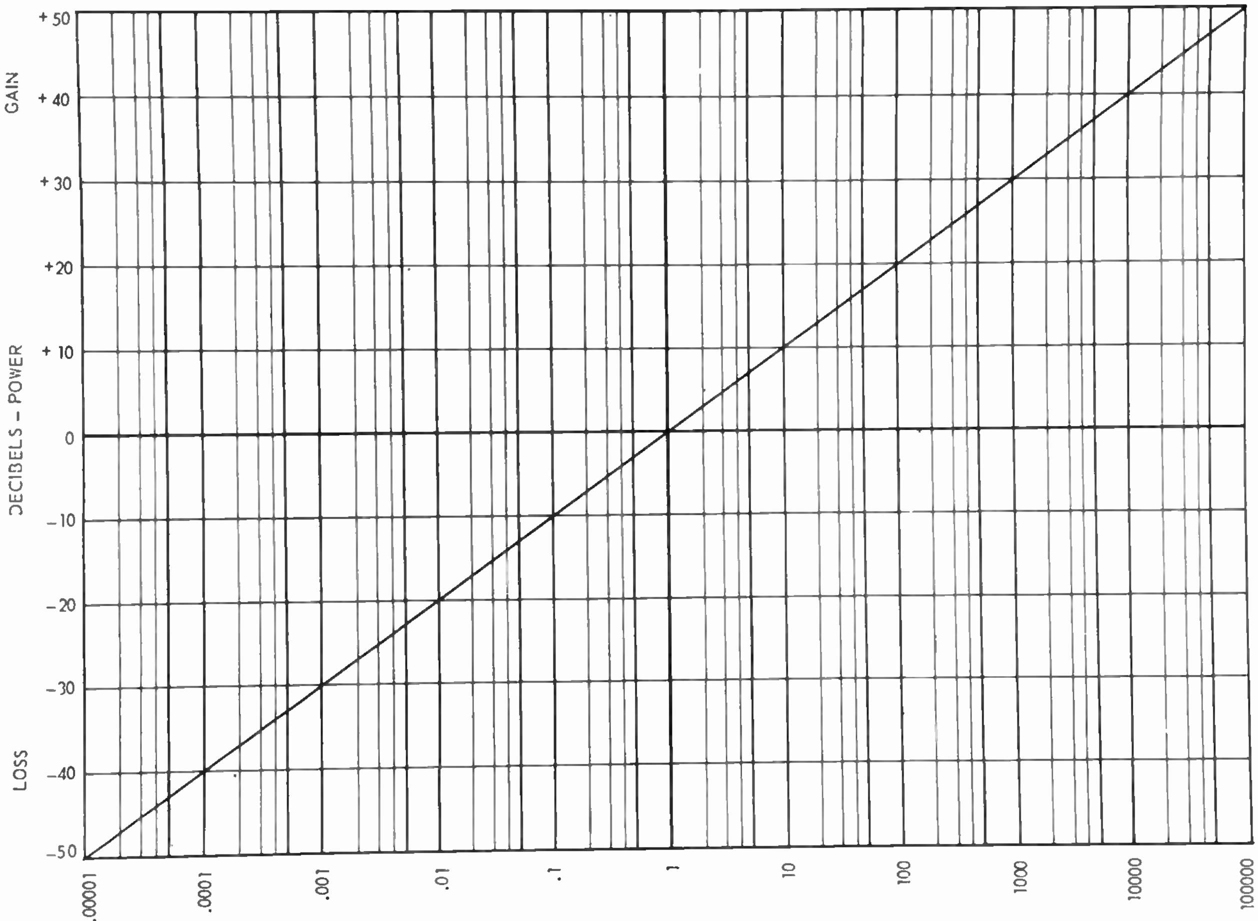


Fig. 82-9. A chart showing relations between decibels up or down and power ratios of gain or loss may be extended to any degree. A change of 10 times or of 1/10 in the ratio always means a change of 10 decibels.

(4) There are also special decibel meters, usually of the rectifier type but having greater internal resistances than those corresponding to 1,000 ohms per volt, which are common in a-c rectifier voltmeters.

Fig. 82-10 is a picture of the dial scales on a service type volt-ohm-milliammeter. From the top down the scales are for ohms, for d-c volts and milliamperes, for a-c volts, and for decibels. The decibel scale is immediately below the one for a maximum of 2.5 a-c volts. The left-hand end of the decibel scale is marked –4. A little ways up on this scale is zero, and at the right-hand end is +10 decibels. Under the decibel scale is written, "1 Milliwatt 0 Level 600 Ohm Line." Under the decibel scales of other a-c voltmeters you may find "6 Milliwatts 0 Level 500 Ohm Line" or "6 Milliwatts 600 Ohms" or possibly other combinations of milliwatts and ohms.

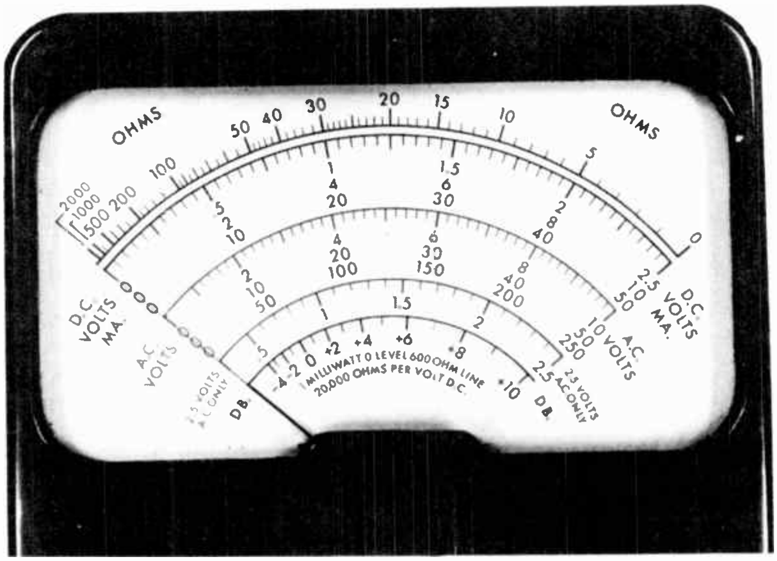


Fig. 82-10. The dial of a volt-ohm-milliammeter with decibel scale graduated for reference level of one milliwatt in 600 ohms.

The reason for these legends on the decibel scales of a-c voltmeters is that watts cannot be measured with a voltmeter unless you bring in the matter of resistance or impedance in ohms. When the meter pointer indicates a certain number of decibels it really is measuring a-c volts. If the a-c voltage being measured and indicated is that existing across a resistance or impedance specified on the meter, then the reading on the decibel scale holds good. Otherwise the decibel reading requires correction.

If you extend a line through the zero decibel mark on the scale of Fig. 82-10 it will cross the a-c 2.5 volts scale at very nearly 0.775 volt. This is because a drop of 0.775 a-c volt across a resistance or impedance of 600 ohms means that 1 milliwatt or 0.001 watt of a-c power is being used in that load. You can figure this out by putting the indicated a-c voltage and the specified ohms into the usual formula for watts, like this.

$$\text{Watts} = \frac{\text{volts}^2}{\text{ohms}} \qquad \frac{0.775^2}{600} = \frac{0.6}{600} = 0.001, \text{ or one milliwatt}$$

Fig. 82-11 shows the dial scales on a different volt-ohm-milliammeter. Underneath the bottom scale for decibels is written, "6 Milliwatts 0 Level 500-ohm Load". The zero point on the decibel scale is much higher up than in Fig. 82-10. A line drawn through the zero graduation cuts the a-c 2.5 volts scale at 1.73 or 1.732 volts. Putting volts and ohms into the formula for watts gives,

$$\frac{1.732^2}{500} = \frac{3}{500} = 0.006, \text{ or six milliwatts.}$$

Fig. 82-11. This dial scale for decibels is based on a reference level of 6 milliwatts in 500 ohms.

On some service type meters the decibel scale is not marked with the reference level in watts and the load in ohms for which the scale is calibrated. You can determine the reference level and load ohms by noting the number of a-c volts indicated when the pointer is at zero on the decibel scale, thus.

0.775 a-c volt.	1 milliwatt in 600 ohms.
1.732 a-c volts.	6 milliwatts in 500 ohms.
1.897 a-c volts.	6 milliwatts in 600 ohms.

Supposing that we had an a-c voltmeter with all the usual decibel scales on one dial. Without filling in all the graduations for decibels the combination dial would appear as in Fig. 82-12. At the top is an a-c volts scale, running from zero to 10 volts. Above this scale are marked several voltages which are of importance in relating the reference levels of the decibel scales, one to another.

To begin with, imagine that the meter pointer stands at 0.775 volt. This will also be zero decibels on the scale whose reference level is 1 milliwatt in 600 ohms, it will be –7 decibels on the scale for 6 milliwatts in 500 ohms, and will be –7.8 decibels on the scale for 6 milliwatts in 600 ohms.

Next, we shall imagine the pointer standing at 1.732 a-c volts. This is the zero point on the scale for 6 milliwatts in 500 ohms, it is +7 decibels for 1 milliwatt in 600 ohms, and is –0.8 decibels on the scale for 6 milliwatts in 600 ohms. Now go on imagining that the meter pointer stands at other values of a-c

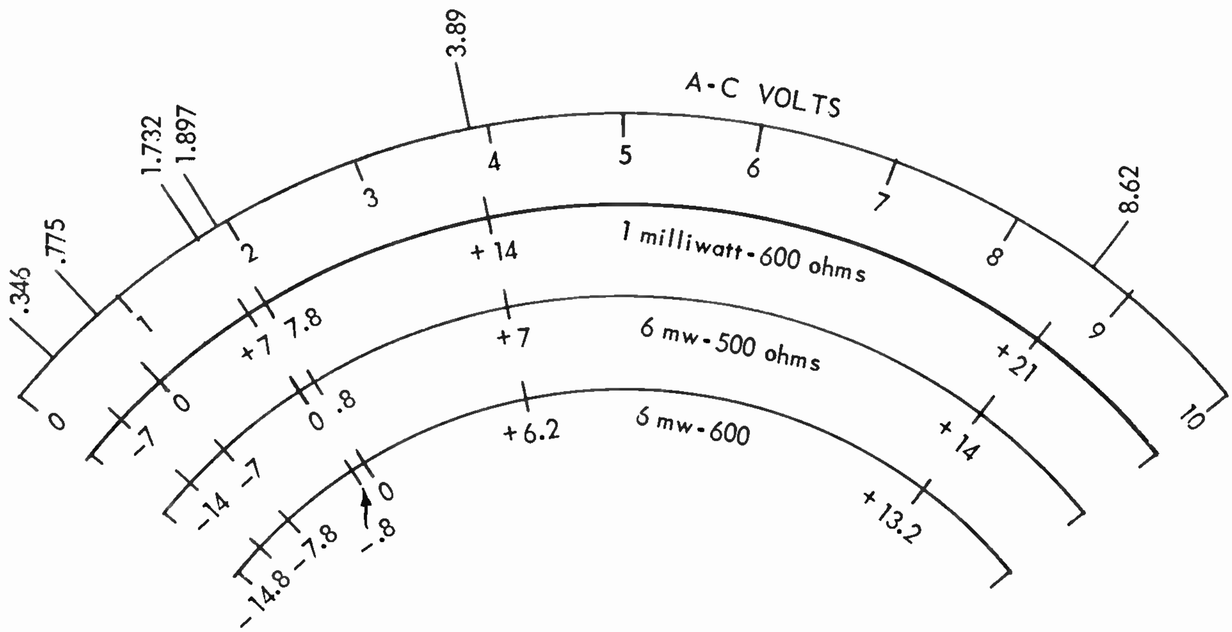


Fig. 82-12. Relations between readings of a-c voltage and of decibels on scales graduated for three different reference levels.

volts, and compare the numbers of decibels for the same a-c voltage on the three decibel scales. You will discover these facts.

a For any given voltage, decibels always are 7 greater on the 1 milliwatt 600 ohm scale than on the 6 milliwatt 500 ohm scale, and they are 7.8 higher than on the 6 milliwatt 600 ohm scale. Naturally then, for any given a-c voltage, the number of decibels on the 6 milliwatt 500 ohm scale is 7 less than on the 1 milliwatt 600 ohm scale. The whole thing sums up as shown by the accompanying table, where we are using the standard abbreviation for decibel, which is "db".

CONVERSION OF REFERENCE LEVELS

Original Reference Level (Scale on which readings taken)	To Convert To These Reference Levels		
	1 mw in 600 ohms	6 mw in 500 ohms	6 mw in 600 ohms
1 mw in 600 ohms	----------	Subtract 7 db	Subtract 7.8 db
6 mw in 500 ohms	Add 7 db	----------	Subtract 0.8 db
6 mw in 600 ohms	Add 7.8 db	Add 0.8 db	----------

Notice that additions and subtractions must be "algebraic". For instance, if you read –7.8 db on a 6 mw 600-ohm scale and convert to 1 mw in 600 ohms, it is necessary to add 7.8 db. When you add (+) 7.8 db to –7.8 db, the sum is 0 db. It is like having the thermometer go up 7.8° from 7.8° below zero, the final temperature is 0°.

It is fortunate that most meters of recent manufacture follow the uniform practice of having decibel scales with calibration or graduation based on the reference level of 1 milliwatt in 600 ohms. When some other reference level is not mentioned in discussion of amplifiers and of sound in general, it is fairly safe to assume this reference level.

The next problem with reference to reading decibels as measured by a-c voltmeters is that of converting a reading taken across some load other than that assumed for calibration into a corrected number of decibels. If you are using a meter calibrated for a 600 ohm load, and measure across the 4-ohm load of some speaker voice coil, the actual number of decibels will be about 21.78 more than the scale reading.

Unless you know or can measure with fair accuracy the actual resistance or impedance across which your meter is connected, it is useless to attempt reading decibels of power from the meter scale. When the actual load impedance or resistance is known it is easy to correct the meter reading and obtain the true number of decibels of power as based on the milliwatts of reference level for which the meter is calibrated.

How we determine the number of decibels to be added or subtracted from the reading on the decibel scale is shown by Fig. 82-13. The horizontal scales along the bottom of the graph refer to the ratio between the actual load resistance or impedance and the load for which the decibel scale is calibrated. There are four ratio scales. The first scale, A, goes from 1 to 10. The next one, B, goes from 10 to 100. Then scale C goes from 100 to 1,000, and scale D from 1,000 to 10,000. We could add any number of additional ratio scales, since each is 10 times the preceding one at each division of the graph. There are four vertical scales showing numbers of decibels to be added to or subtracted from the meter reading. Scale A extends from zero to 10 decibels, B goes from 10 to 20 decibels, C from 20 to 30 decibels, and D from 30 to 40 decibels. Each scale is formed by adding 10 to every number in the preceding scale, so we might add any number of decibel scales to accompany added ratio scales.

Decibel scale A is used with ratio scale A, scales B are used together, and so on. To illustrate using the chart let's go back to the problem of using the meter across a 4-ohm voice coil when the calibration is for 600 ohms load. The ratio of reference ohms to actual ohms is the ratio of 600 to 4, or is 150 to 1. On the ratio scales across the bottom of the chart we find the ratio of 150 on scale C and follow upward into the graph until coming to the diagonal line. From this intersection we follow to the left into the decibel scales. Our ratio is on the C scale so we read decibels on the C scale. On this scale we have followed into the section between 21 and 22 decibels, and from the intermediate graduations on the graph it can be estimated that the number of decibels should be about 21.78.

5 If the actual load resistance is less than the meter reference load, the number of decibels taken from the chart must be added to the meter reading. Why we add requires only a little thought. If you measure any certain voltage across a relatively small resistance or impedance (4 ohms in our example) there must be a lot of current in that resistance or impedance to cause such a voltage drop. At least there must be much more current than in a greater resistance or impedance (600 ohms reference level in our example). Large current means large power, so the power in our actual small load must be greater than in a larger

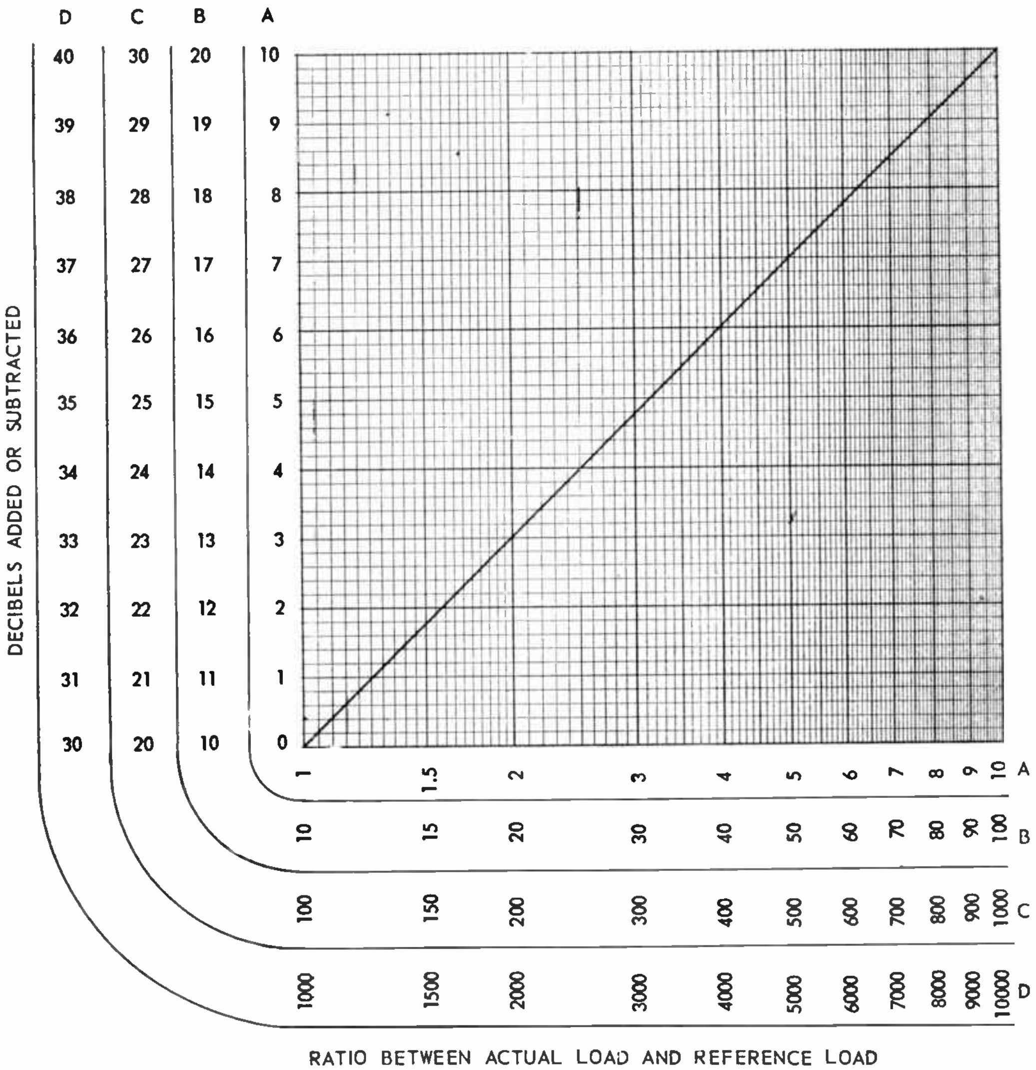


Fig. 82-13. This chart shows decibels to be added or subtracted with the actual load impedance or resistance is not the same as the number of ohms in the meter reference level.

load. Since actual power is greater than the power indicated by the meter we must add the decibels of power to the meter reading.

If you measure decibels (really a-c voltage) across a resistance or impedance greater than the reference value for which the meter is calibrated, then the number of decibels taken from the chart would be subtracted from the meter reading. Supposing you were to make a measurement across a plate load resistor of 12,000 ohms with a meter whose reference level is 600 ohms. Dividing 12,000 by 600 gives the resistance ratio as 20 to 1 or as 20. We find this ratio on scale B of Fig. 82-13. Following upward to the diagonal line, then across to scale B for decibels, we find 13 decibels. Because the actual load is greater than the reference level load this number of decibels must be subtracted from the meter reading. You might have a reading of 16 decibels. Subtracting 13 from 16 gives 3 decibels as the actual power based on the number of milliwatts for which the meter is calibrated.

Now we might go back to Fig. 82-8 and note that 3 decibels means a power ratio of very nearly 2 times. Assuming that the reference power level of the meter is 1 milliwatt or 0.001 watt, this ratio of 2 times would mean that the actual power is 2 milliwatts or 0.002 watt. For proof it would be necessary only to note that the meter pointer at 16 db points to about 4.89 volts on the a-c volts scale. Power in watts is equal to the square of the volts divided by the load ohms. The square of 4.89 is 23.91 or very nearly 24. Dividing 24 by 12,000 (actual load ohms) gives the power as 0.002 watt.

Even though you never make corrections for reference levels or for the resistance or impedance across which the meter actually is connected, the decibel scale still is useful in comparing powers. Every time the reading goes up by 3 db the actual power has doubled. You won't know the actual power in watts, but you will know that it has been doubled. If the increase between two measurements is 6 decibels, the power has doubled and doubled again, so has increased 4 times. If the increase is 9 db, or three 3's, the power has doubled three times and is raised 2 x 2 x 2 times, or 8 times. Conversely, every drop of 3 db indicates a halving of the power. If the reading goes down by 6 db (two 3's) the power has been halved and halved again, and is 1/4 the original value.

The final correction when reading decibels is that which compensates for the range being used. Most meters having decibel scales have several ranges for a-c voltage. The decibel scale applies directly only when the range selector is set for the range in which a-c voltages match the zero point on the decibel scale. This we discussed before, and made a list of a-c voltages corresponding to zero decibels for the various common reference levels. We may call this the reference range. Usually but not always the reference range is the lowest of all those provided. The correction is made as follows.

1. Divide the full-scale a-c volts of whatever range you use by the full-scale volts of the reference range. Call the result a ratio.

2. Look up the number of decibels corresponding to this ratio.

3. Double the number of decibels.

4. If the range being used is higher than the reference range, add the doubled number of decibels to the number indicated by the meter pointer on the decibel scale. If the range being used is lower than the reference range, subtract the doubled number of decibels.

If you compare the chart of Fig. 82-13 with the graph at the right in Fig. 82-8 you will see that the only difference is more and closer graduations in Fig. 82-13. This makes it possible to use this latter chart for determining the number of decibels corresponding to any power ratio, or the ratio corresponding to any number of decibels.

Here is an example in correcting for meter range being used. Assume that the reference range is from zero to 2.5 volts A.C. as on the meter of Fig. 82-10, and assume that you are making measurements on the 10 volt range of this meter. Dividing 10 by 2.5 gives the ratio as 4. Figs. 82-8 and 82-13 show that this ratio corresponds to 6 db. Doubling the db figure obtained from the chart gives you 12 db, which is now added to any reading that you now obtain on a decibel scale. Were you using the 50 volt range, the ratio would be 20, because 50 divided by 2½ equals 20. A ratio of 20 (Scale B of Fig. 82-13) corresponds to 13 db; doubling this would give you a figure of 26 db which is what you would add to the number of decibels read on the decibel scale of the meter.

6 *GAIN MEASUREMENTS.* In technical descriptions and catalog descriptions of audio amplifiers you nearly always find the overall gain specified as some number of decibels, possibly something like 64 db. 96 db, or 120 db. These are power gains. If you translate them to ratios the gains appear immense. This is because power in fractions of a watt applied to the grid of the first amplifier is exceedingly small, while power applied at the speaker may be many watts. The voltage gain, were input and output impedances equal, would be only the square root of the ratio of power gain. For example, a power gain in the ratio of one million to one would correspond to a voltage gain, between equal impedances, of one thousand to one.

The method of measuring the power gain of an amplifier is, in brief, as follows.

1. Measure the number of decibels at the amplifier input. Practically always this will be a minus number, a great many decibels below the reference level.

2. Measure the output power in decibels. This should be a plus number, above the reference level.

3. If it is necessary to change meter ranges between input and output measurements, make the necessary correction for this change.

4. The input and output resistances or impedances will be different. Determine the ratio of input to output resistance or impedance and make the necessary correction for this ratio.

5. The difference (algebraic) between corrected input and output decibels is the gain of the amplifier at the frequency used for testing.

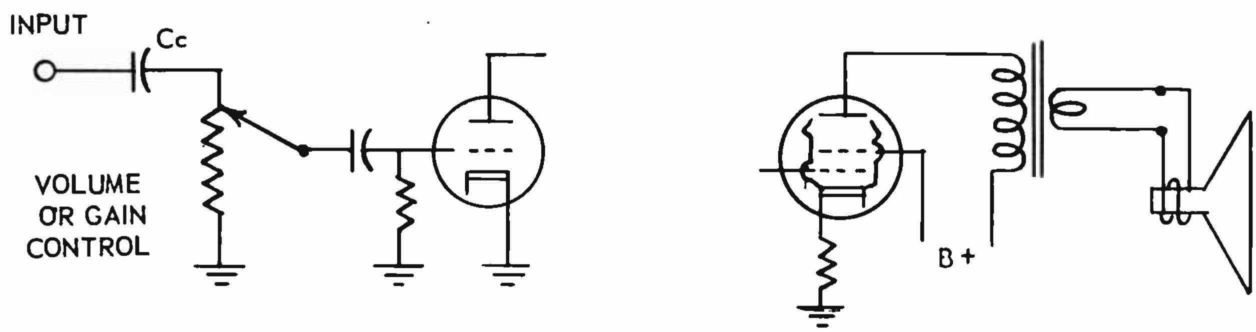


Fig. 82-14. It is necessary to carefully consider the input circuit when measuring amplifier gain.

In practice it is advisable to connect the decibel meter, usually your VTVM, to the output for a start. Vary the strength of input signal, from an audio generator, to note the setting at which output ceases to rise and begins to drop back. Use an input that gives an output somewhat less than the peak, in order to avoid distortion and excessive noise effect. The meter is not "frequency selective", and is indicating total output power, which includes the amplified tube and circuit noises. These could be as great or even greater than the signal output.

Assume that you are making measurements on the amplifier whose input and output stages are as shown by Fig. 82-14. First, make sure that the input signal really is applied across the resistance or impedance which you intend to use, or across one that can be measured. If you apply the signal to the regular input terminal this signal acts through the entire volume control resistance to ground, and the input resistance is practically that of the whole volume control, neglecting any reactance at Cc. But the volume control slider might be anywhere. The actual input to the grid would be only that through the volume control from the slider to ground, yet the generator still would work into the total resistance or impedance from the input terminal to ground. The slider must be set for maximum volume or gain if you assume this total input resistance or impedance.

It would be better to adjust the control slider for some certain resistance between it and ground by using an ohmmeter. Then connect the input lead from the generator to the slider, and you know the resistance actually used for input.

In order to keep the output power below a value at which there is severe overloading of the amplifier it is necessary to apply a very small input signal. This signal ordinarily is so small that it cannot be measured accurately on the decibel scale of the usual type of service meter. One way to get around this difficulty is to use a resistance voltage divider across the audio generator output and take some known fraction of the total voltage as an input for the amplifier. Then it becomes possible to measure total generator output with the meter, using points on the scale where reasonable accuracy is possible.

If 1/10 of the total generator voltage is applied to the amplifier output, and if the meter is used to measure decibels across the entire voltage divider, you must subtract 20 decibels from the meter reading to determine the decibels of input to the amplifier. For example, if the meter connected across the entire divider indicates +15 db, and 1/10 of the total generator voltage is being applied to the amplifier input, this input is –5 db. This is because subtracting 20 db from 15 db brings us down to 5 db below the zero level, or to –5 db.

If your gain measurements do not exactly match the manufacturer's claims for the amplifier, remember that broad tolerances are allowable.

TELEVISION

LESSON NO. 83

THE OSCILLOSCOPE

Training Your Mind...

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

COMMERCIAL
TRADES
INSTITUTE

Chicago, Ill.

LESSON NO. 83

THE OSCILLOSCOPE

When you stop to think about it, the greater part of all our work in télevision and radio servicing is concerned with the variations of voltages and currents called signals. We deal with little else than signal variations of voltages and current all the way from the antenna until sound comes from the speaker and pictures form on the picture tube. If all the signal strengths, polarities, and waveforms remain exactly right there is no trouble and no need for service. If any of them go wrong there is a faulty picture or distorted sound.

In earlier lessons we have looked at many diagrams representing the variations or alternations of voltages and currents which are signals and control pulses. Sometimes the signals and pulses were shown by photographs of traces formed on the screen of an oscilloscope. The oscilloscope, or oscillograph, is the only instrument which lets you see the voltage variations as they occur, and lets you watch the effect of every service adjustment as you alter its setting. The oscilloscope allows watching a sound signal all the way from the antenna to the voice coil of the speaker. It allows watching a television signal and everything that happens to it from the time the signal is formed at the antenna until it arrives at the grid circuit and the deflection system of the picture tube.

On the oscillograph pictured by Fig. 83-1 appears the face of a cathode-ray tube which, in every structural detail, is like a picture tube of the electrostatic deflection type. On the panel are controls for brightness, focusing, width, height, centering, frequency selection, and synchronizing. We have what amounts to a television set for looking at signals in action, rather than at scenery and human beings in action.

As at the left in Fig. 83-2, the oscilloscope will show something so simple as the alternating voltage wave from a 60-cycle power line, or, as at the right, it will show a wave having severe distortion caused by too little plate voltage on an audio amplifier tube. We may look at the horizontal sync pulses and the intervening active lines of a television signal, as at the left in Fig. 83-3. At the right the oscilloscope is displaying the frequency response of a television sound demodulator.

In a picture tube the electron beam is moved continually from side to side and vertically to cover a rectangular area with the luminous raster. Image lights and shadows are formed by varying the control grid voltage and beam intensity. This method of tracing is not used in the oscilloscope tube.

Traces are formed on the face or screen of the oscilloscope tube as shown by Fig. 83-4. The beam is swept continually in a horizontal direction, as in diagram 1, by applying a sawtooth voltage to the horizontal deflection plates. The sawtooth voltage sweeps the beam at moderate velocity from left to right,

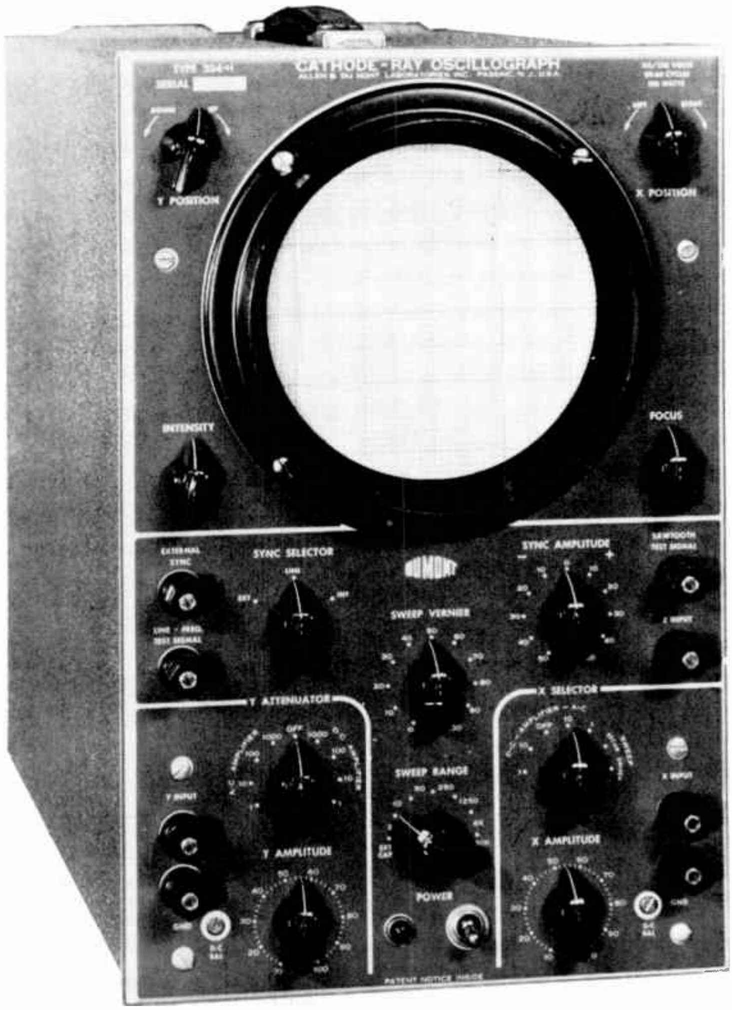


Fig. 83-1. The front panel controls and CRT face of a Du Mont oscillograph.

then returns the beam to the left at very high velocity. Thus we have a horizontal trace and retrace, just as in a picture tube.

8 Should there be no potential difference between the horizontal deflection plates while an alternating voltage is applied to the vertical plates, as in diagram 2, the beam will move up and down to form a vertical trace line on the screen.

In diagram 3 the beam is being swept horizontally by the sawtooth voltage, and during the early part of the horizontal sweep the upper vertical plate is made momentarily positive with reference to the lower

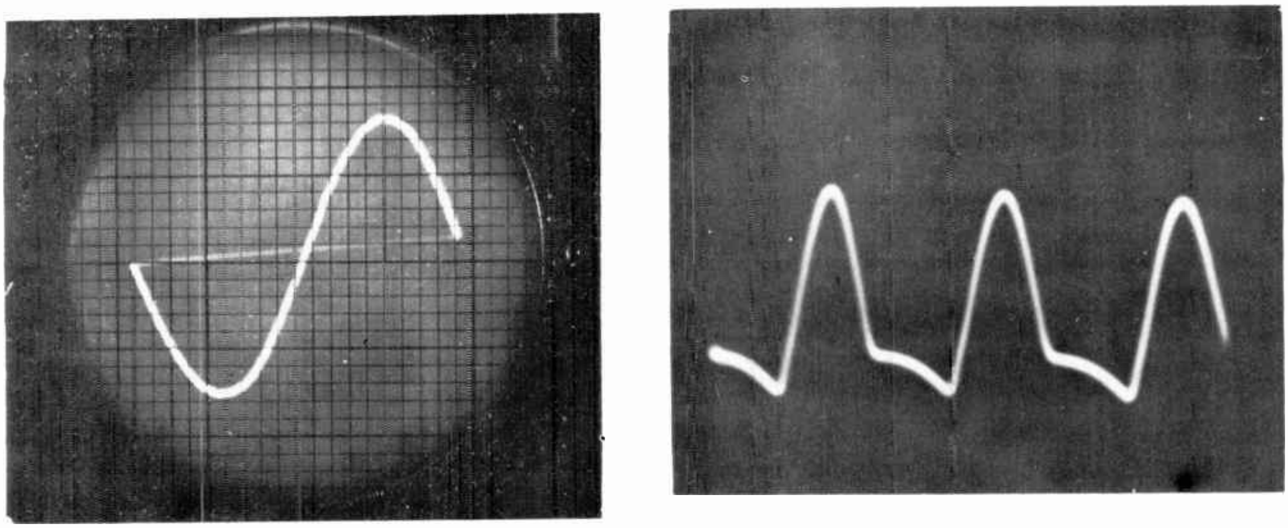


Fig. 83-2. A trace showing a single sine-wave cycle (left) and one showing distortion in an audio-frequency voltage (right).

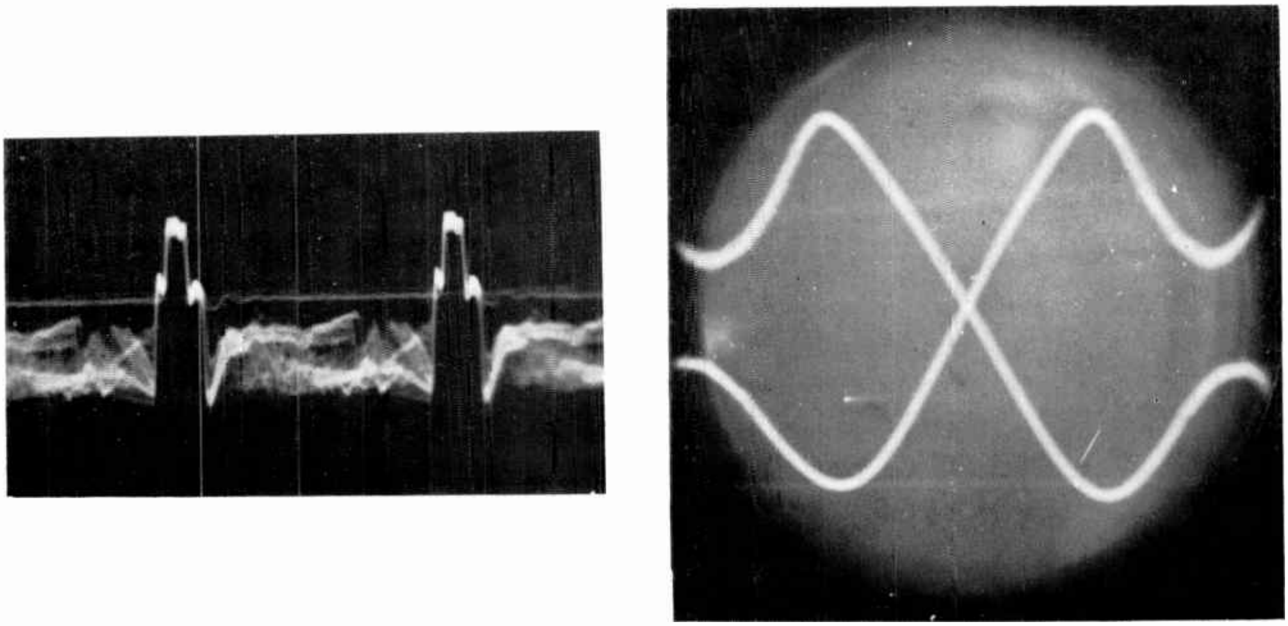


Fig. 83-3. Horizontal blanking and sync pulses of a television signal (left) and crossed S-curves from a ratio detector (right).

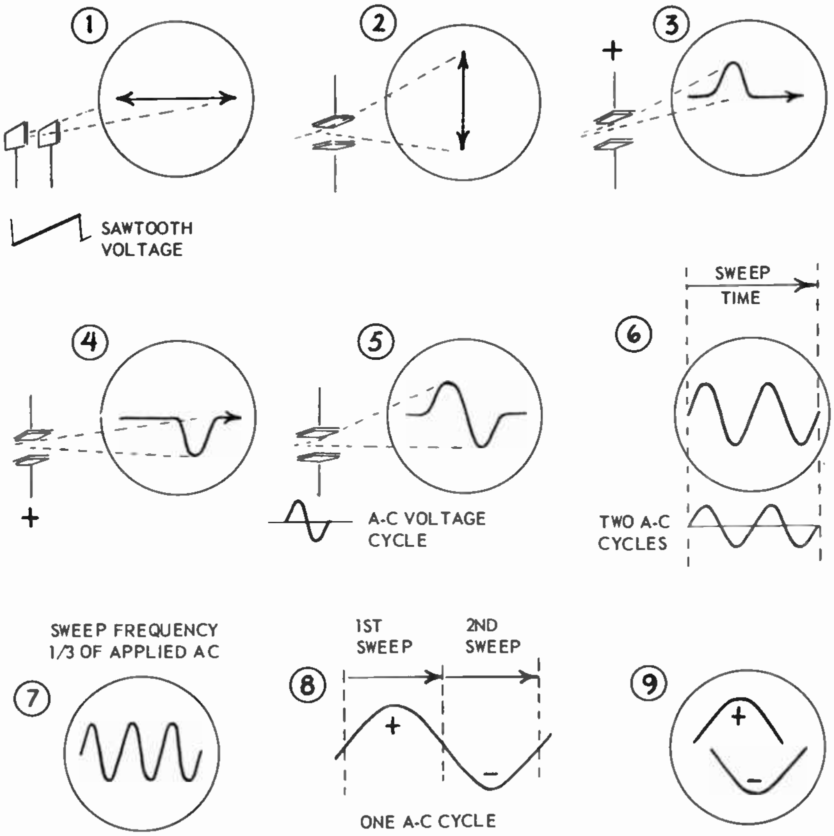


Fig. 83-4. How traces are formed on the screen of the oscilloscope tube.

plate. Then the beam moves upward and back again during its horizontal travel. If, as in diagram 4, the lower plate of the vertical pair is made momentarily positive with reference to the upper plate, the beam will move down and back again during its horizontal travel.

In diagram 5 the voltage applied to the vertical deflection plates during one horizontal sweep of the beam is alternating in polarity. The beam moves up, returns to its zero position, goes on down, and again returns to zero. Thus we have a true picture of the a-c voltage acting on the vertical plates.

Should there be exactly two a-c voltage cycles during the time of one horizontal sweep of the electron beam, as in diagram 6, the two cycles will appear on the screen of the oscilloscope tube. It is important that you realize just how this comes about. To have two cycles on the screen, the frequency of the applied a-c voltage must be double the horizontal sweep frequency. If the sweep frequency were to remain constant between diagrams 5 and 6 it would be necessary to double the frequency of the applied a-c voltage.

Ordinarily it is inconvenient or impossible to alter the frequency of applied voltage, for in examining any certain voltage we do not wish to alter it in any way. Then we may bring two cycles onto the oscilloscope screen by making the horizontal sweep period longer than before, by making this period as long as two cycles of the applied voltage. This means reducing the sweep frequency to one-half the value required for observation of one cycle, or making the sweep frequency half that of the applied a-c voltage. If the horizontal sweep frequency is made one-third that of the a-c voltage on the vertical deflection plates it will bring three a-c cycles onto the oscilloscope screen, as in diagram 7. Any other number of a-c cycles may be displayed on the screen by suitable adjustment of the sweep frequency.

2

Inasmuch as a reduction of sweep frequency brings more and more a-c cycles onto the oscilloscope screen, an increase of sweep frequency might be expected to bring a half-cycle or any other fraction of a cycle into view. This is correct, but the results seldom are of any practical use for the reason illustrated by diagram 8 of Fig. 83-4. Assuming that the frequency is twice that of the applied frequency, there will be one complete horizontal sweep of the beam during the time of a half-cycle, and a second complete horizontal sweep during the next half cycle. Due to persistence in the phosphor of the screen and in your eyes, the positive and negative half cycles will appear together. But they will appear as in diagram 9 or in some other relation to each other. A sweep frequency still higher would bring three, four, or more partial cycles onto the screen at the same time, and the resulting pattern would be confusing rather than helpful.

Fig. 83-5 shows the relations between the principal parts of a typical service type oscilloscope. At the upper right is the cathode-ray tube, whose name commonly is abbreviated to CRT. There is a low-voltage power supply for plates, screens, grid biasing, and heaters of the amplifier and oscillator tubes. For the elements and deflection plates of the CRT there is a high-voltage power supply. In the voltage divider system connected across the high-voltage supply are vertical and horizontal centering controls, a focus control, and an intensity control. The centering controls are of the same design used for electrostatic picture tubes. The slider of the focus control goes to the focusing anode or grid of the CRT. The intensity control takes the place of a brightness control for a picture tube, its slider going to the control grid of the CRT to vary the grid voltage, beam intensity, and brightness of the trace formed on the tube screen.

On the left-hand side of the diagram are three pairs of terminals for feeding various voltages into the oscilloscope circuits. The top pair is marked "Vertical Input", the next pair "External Sync", and the bottom pair "Horizontal Input". Each pair consists of one insulated terminal and one grounded terminal. Let's assume that a voltage to be observed is connected to the vertical input. In series with the insulated lead is a blocking capacitor which prevents any d-c component of the observed voltage from entering the scope circuits. Then comes an adjustable potentiometer marked "Vertical Gain Control". This control often is marked "Vertical Attenuator". It regulates the strength of observed voltage fed to the vertical amplifier.

The slider of the vertical gain control feeds into the vertical amplifier. This amplifier is much like the vertical amplifier for an electrostatic picture tube, but it amplifies the observed signal voltage rather than amplifying the output of a vertical sawtooth oscillator. The vertical amplifier ordinarily includes a phase inverter circuit for feeding two voltages of equal amplitude and opposite polarities to the vertical deflection plates of the CRT.

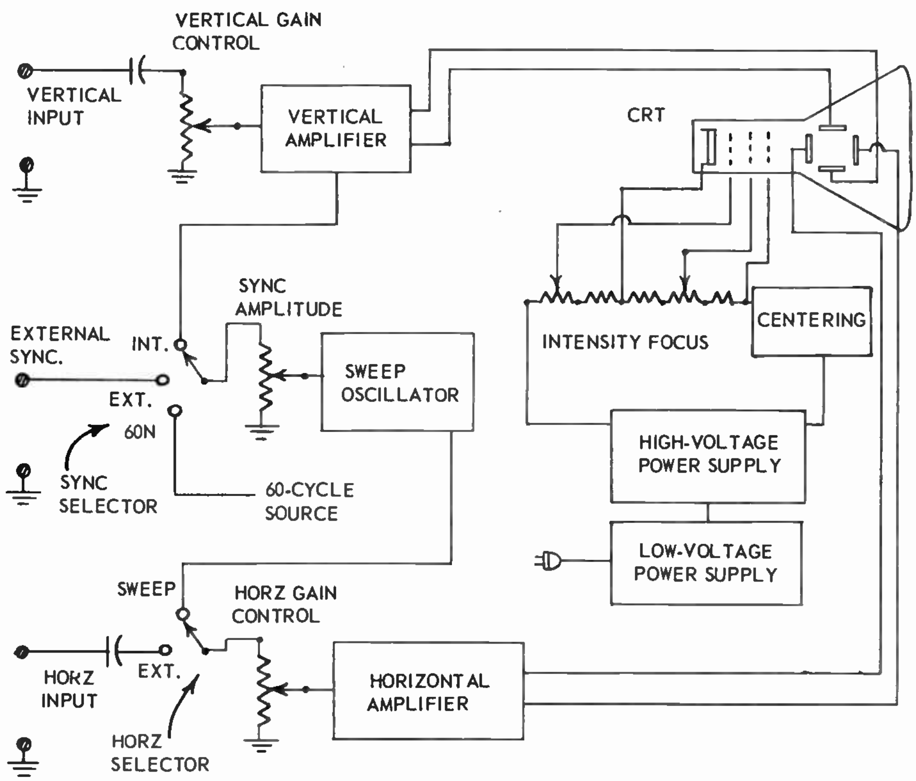


Fig. 83-5. Relations between the principal parts of a typical service oscilloscope.

The sweep oscillator produces the sawtooth voltage for horizontal deflection or sweep of the beam in the CRT. Not shown in the diagram, but to be discussed later, are adjustments for varying the sweep frequency of this oscillator over a very wide range. After these adjustments are set for approximately the desired sweep frequency, the oscillator must be held in exact synchronization with that desired frequency. A synchronizing voltage comes into the sweep oscillator through the "Sync Amplitude" control, which is simply a potentiometer for regulating the strength of sync voltage.

The sync amplitude control is fed from the "Sync Selector" switch. With this switch in the position shown, the sync voltage for the sweep oscillator is being taken from the vertical amplifier. A small portion of the amplified voltage to be observed is thus used for holding the sweep frequency at some exact fraction of the frequency of the observed voltage. The need for doing this was explained in connection with Fig. 83-4.

For some tests it is desirable to synchronize the sweep frequency with some voltage other than that being applied at the vertical input terminals, or it may be desirable to synchronize from a direct connection to the source of voltage being observed. Then the sync selector is set to connect the sweep oscillator to the external sync terminal. This terminal is connected to whatever voltage is to be used for synchronizing.

For still other tests it is necessary to synchronize the sweep frequency at exactly 60 cycles, which is accomplished by setting the sync selector on the 60-cycle position. The 60-cycle source commonly is the 6.3-volt or 12.6-volt heater circuit for the amplifier and oscillator tubes of the scope, or this sync voltage may be taken in any other way from the power transformer or power line. For this reason the 60-cycle sync position often is marked "Line".

The synchronized output of the sweep oscillator goes to the "Horizontal Selector" switch in the diagram of Fig. 83-5. With this switch in the "sweep" position, the sawtooth sweep voltage goes through the "Horizontal Gain" control to the input of the horizontal amplifier. This gain control acts for the horizontal amplifier as does the vertical gain control for the vertical amplifier. The horizontal amplifier usually is similar to the vertical amplifier, containing a phase inverter circuit for feeding two sawtooth voltages of opposite polarity to the horizontal deflection plates of the CRT.

There are many important applications of the oscilloscope in which a sawtooth horizontal sweep voltage cannot be used. Then the horizontal selector switch is set at its "ext." position, which connects the horizontal amplifier through its gain control and a blocking capacitor to the horizontal input terminals. Then any alternating voltage which is to control horizontal deflection of the CRT beam may be connected to the horizontal input terminals. The sync selector and the horizontal selector may be combined in a single switch.

Cathode-ray tubes for oscilloscopes usually have screen diameters of three, five, or seven inches, although smaller or larger sizes are used for instruments serving special purposes.

The tubes used in service oscilloscopes most often have phosphor number one (P-1), for the screen coating. This phosphor produces a green trace of medium persistence. It is easily visible in a normally lighted room, having a decided contrast with either daylight, incandescent light, or fluorescent light which reaches the screen in moderate intensity.

Among the special-purpose phosphors are the following. Number two, (P-2) giving a blue-green trace of long persistence, lasting five to 100 seconds to allow observation of very low frequencies and "transients" which occur between long intervals. Numbers five and eleven, both of which give a blue trace of short persistence well suited to high speed photography such as practiced with moving film. Number eleven may be used also for visual observation. Phosphor number seven produces a greenish-yellow trace with persistence so long that it may be observed for several minutes after the beam passes provided the surrounding illumination is not too bright. This phosphor is used similarly to number two, also for radar indicators.

Fig. 83-6 is a picture of the inside of an oscillograph. The cathode-ray tube, from its front face all the way back to the base, is enclosed within a grounded metal shield as a protection against stray electrostatic fields which would affect deflection of the electron beam. The CRT shield is made of high permeability iron or steel to provide a magnetic shield against electromagnetic fields from power transformers, filter chokes and other parts carrying power line freq. Such a shield must be completely demagnetized, and kept so. If you should bring a permanent magnet near an iron or steel shield, and thus

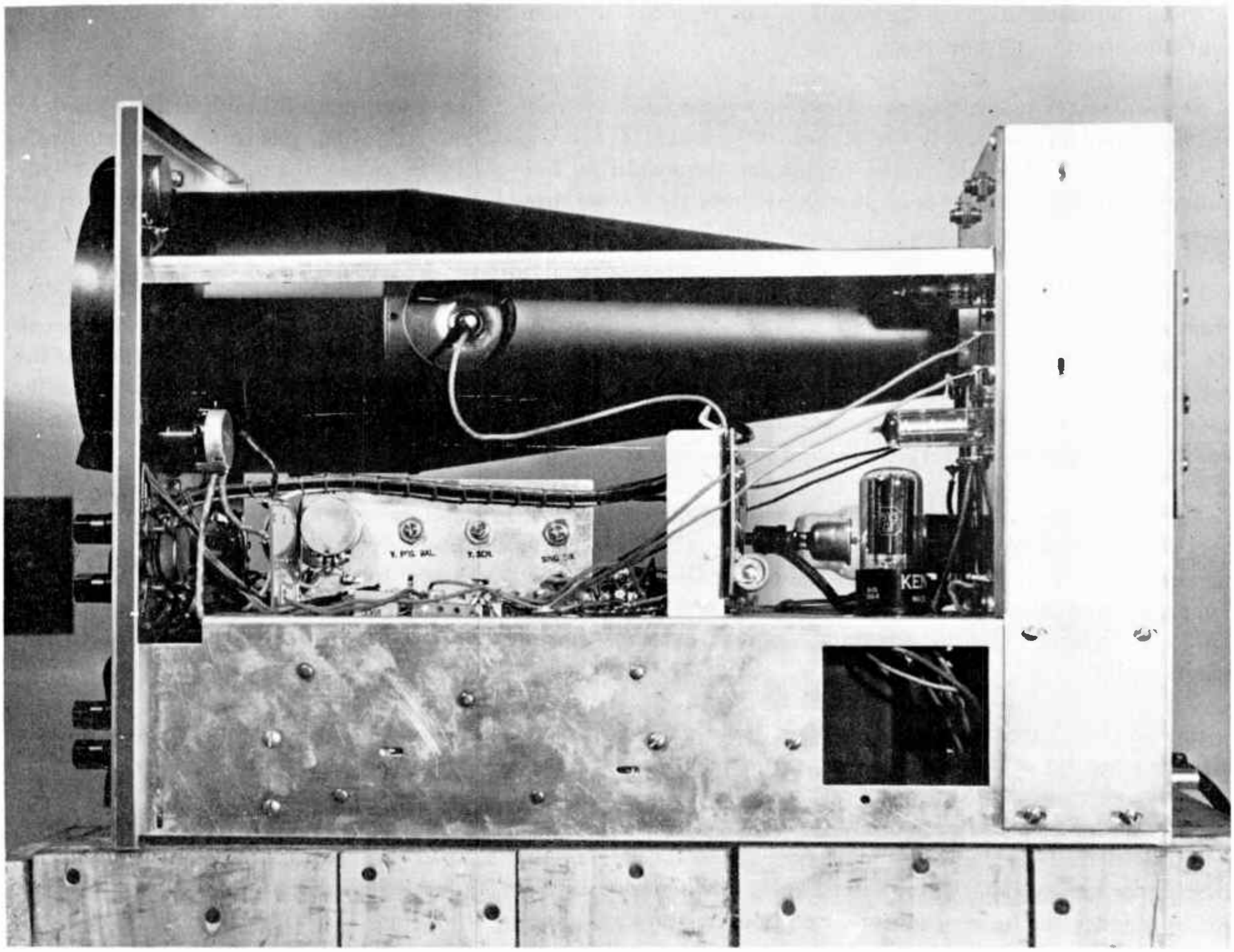


Fig. 83-6. A Du Mont oscillograph with its housing removed.

magnetize the shield metal, the traces formed thereafter will have a permanent "set" toward the right, left, top, or bottom of the screen.

SWEEP OSCILLATORS. The elementary circuit of a horizontal sweep oscillator such as used in a number of service oscilloscopes is shown at the top of Fig. 83-7. As you may readily see, the oscillator is a cathode coupled multivibrator which operates in exactly the same way as this type of multivibrator when employed in television sweep oscillator systems. The grid of the input triode A receives the sync voltage in the oscilloscope, and the sync pulses in the television receiver. The plate of A is coupled to the grid of triode B through capacitor Cc. The free running frequency of the oscillator may be adjusted within a moderate range by the "Vernier Frequency Control" marked Rf, which corresponds to the hold control in a television oscillator.

The sawtooth capacitor Cs charges from the B supply through resistors Rs, with the charge voltage gradually increasing on the capacitor to form the long slope of the sawtooth voltage wave which deflects the beam from left to right in the CRT. The sawtooth capacitor discharges through section B of the multivibrator, just as in a television sweep oscillator system, to form the sharp drop of voltage during retrace of the electron beam.

One method of providing the very great range of sweep frequencies required in the oscilloscope is shown at the bottom of Fig. 83-7. The multivibrator is the same as in the upper diagram. The grid of section A is fed from the sync selector switch of Fig. 83-5. The single sawtooth capacitor Cs and single coupling capacitor Cc of the simplified upper diagram are replaced by a group of six fixed capacitors lettered from a through f. Various combinations of these capacitors are brought into the oscillator circuit by moving the two rotors of the range selector switch. As shown in the diagram, the two contacts of the upper rotor connects to the upper terminals of capacitors a and b, while the lower terminals of these two capacitors are connected by the two contacts of the lower rotor of the switch. The two contacts on each rotor, and the two rotors, move back and forth together.

By tracing the connections, and comparing the upper and lower diagrams, you will see that range capaci-

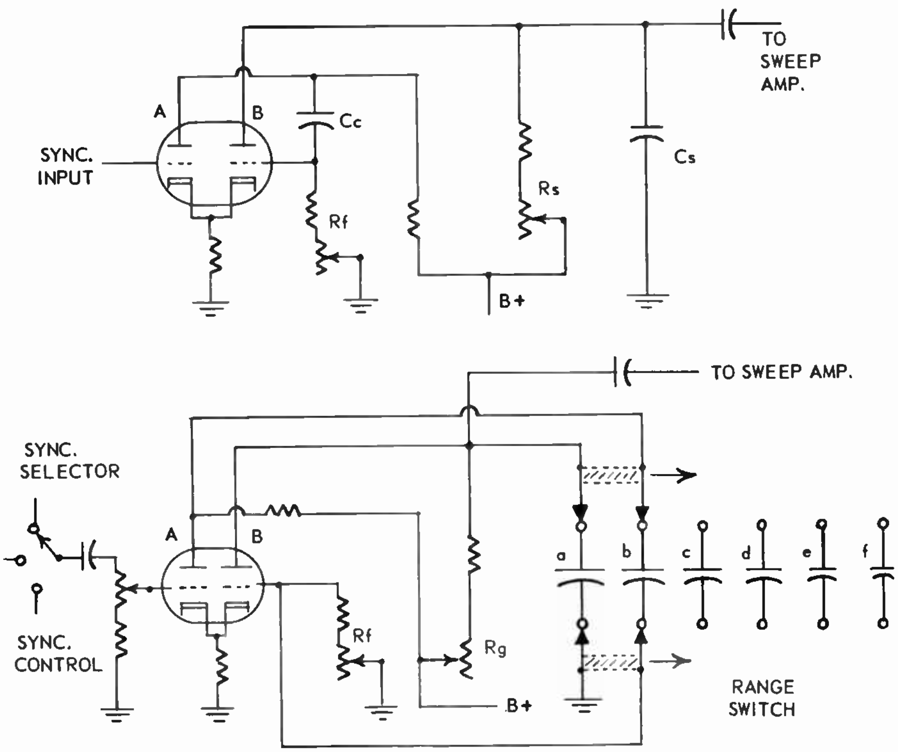


Fig. 83-7. Typical circuits for a multivibrator sweep oscillator.

tor a is now the sawtooth capacitor, being connected between the plate of triode B and ground. Capacitor b is now the coupling capacitor, being connected between the plate of triode A and the grid of triode B. The capacitances become progressively smaller from a to f. Typical values might be: a 0.5 mf. b 0.1 mf. c 0.02 mf. d 0.004 mf. e 750 mmf. f 100 mmf.

As the range selector switch is moved, its two rotors will next make contact with capacitors b and c, with b acting as the sawtooth capacitor and c as the coupling capacitor. Since the sawtooth capacitance now is only one-fifth as great as in the first switch position, the sweep frequency will be proportionately higher. In its next position the range switch will bring in capacitors c and d, then d and e, and finally e and f, to make the sweep frequency higher and higher. Capacitor f never becomes the sawtooth capacitor, it is used only as the coupling capacitor for the highest frequency.

With the range switch in any one position the sweep frequency may be varied within limits by adjustment of the vernier frequency control Rf. This control and the charging resistance Rs, are sometimes operated together, they consist of a dual potentiometer. With the range switch in its first position the vernier control might allow adjusting the sweep frequency from something like 15 cycles to 80 cycles, in the next position the range might be from 70 to 350 cycles, and so on until reaching a maximum frequency of possibly 30,000 cycles at the top of the highest range. In many oscilloscopes using this general type of sweep oscillator the maximum sweep frequency is 60,000 cycles or even higher.

The multivibrator sweep oscillator employs a vacuum tube, usually a twin-triode. In a large number of oscilloscopes the oscillator tube is a thyratron, which is a three element gaseous tube having a cathode, a control grid, and a plate or anode. During manufacture of the thyratron, the envelope or bulb is first thoroughly evacuated, as in the making of a vacuum tube, and then there is admitted to the bulb a very small quantity of some gas such as argon, helium, neon, xenon, or a mixture. This makes what is called a gaseous tube.

In any gaseous tube there occurs the process of ionization, which is explained as follows. The atoms of gas within the tube envelope are normally neutral, each one with enough negative electrons to balance its positive charge. When electrons emitted from the cathode are drawn to the positive plate at high velocity, these electrons strike the gas atoms with enough force to knock one or more negative electrons out of each atom suffering a collision. The additional negative electrons thus freed from the atoms proceed to the positive plate. The atoms which have lost one or more of their negative electrons are now positively charged, and are called positive ions. The positive ions go toward the relatively negative cathode, there to pick up negative electrons from the space charge which is most dense near the cathode. This means that the negative space charge is lessened. Since the rate of emission from the cathode has been heretofore limited chiefly by the repelling effect of the negative space charge, there is now a tremendous increase in the rate of emission. This means than an immense additional quantity of electrons can flow from cathode to plate, and the effective internal resistance or plate resistance of the tube drops to only a few ohms.

In actual operation of the gaseous triode or thyratron the grid is made negative with reference to the cathode. The negative grid holds back the electron flow from the cathode and thus prevents the start of ionization until plate voltage is made sufficiently positive to overcome the retarding effect of the grid. If the plate voltage is maintained at some certain positive value, as usually is the case, then the grid may be made less negative until it no longer can hold back the electrons. Then ionization com-

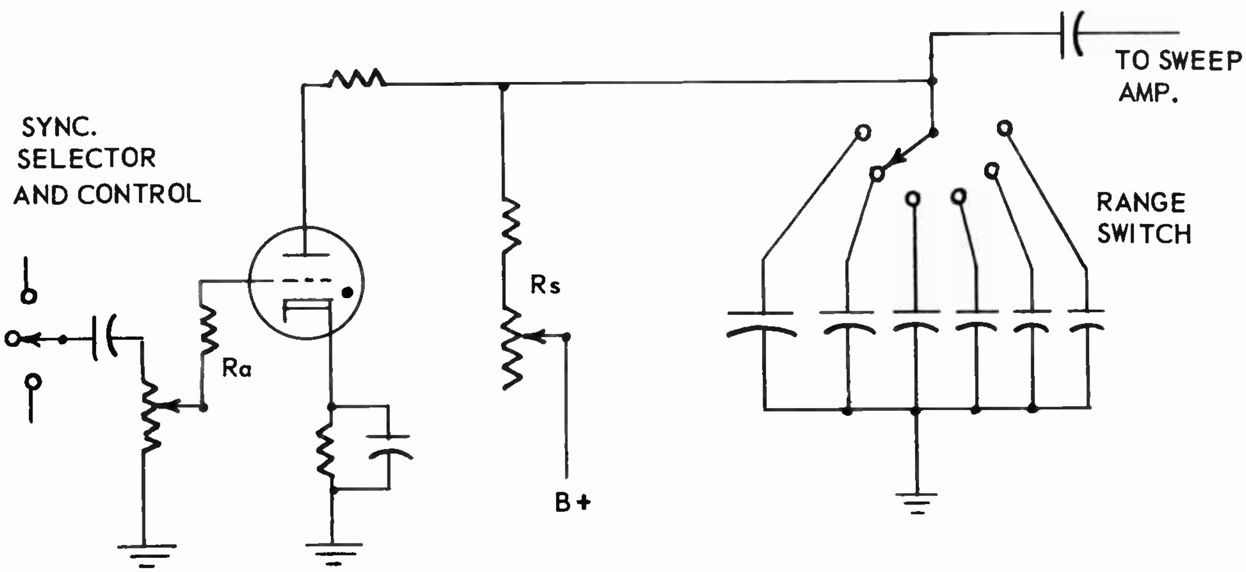


Fig. 83-8. Circuits for a gaseous tube sweep oscillator.

mences, and within about 10 microseconds at most the electron flow through the tube increases from practically zero to its maximum rate.

The circuit connections for a gaseous sweep oscillator, as shown by Fig. 83-8, are somewhat simpler than for the multivibrator because there is only a single triode and because there is no intertube coupling capacitor to be switched for various ranges of capacitance. The range switch connects any one of a group of sawtooth capacitors between the oscillator plate and ground. Adjustable resistor Rs is the vernier control for varying the sweep frequency through any one of the ranges. Reducing this resistance shortens the charging time constant for the sawtooth capacitor while increasing the plate voltage for the oscillator, both of which tend to increase the sweep frequency. This is the same function performed by resistors Rs in the multivibrator oscillator systems of Fig. 83-7.

The purpose of resistor Ra, in series with the oscillator grid, is to limit the flow of grid current during the operating period in which positive ions are being attracted to the negative grid. The value of resistance at this point is about 1,000 ohms for each peak volt of synchronizing voltage used on the grid. The black dot within the circle of the oscillator symbol indicates that this is a gaseous tube. Such a dot, which may be anywhere within the circle, always indicates a gaseous tube.

Fig. 83-9 shows average relations between positive plate voltage and negative grid voltage for the start of ionization in the two types of gaseous triodes most often employed as oscilloscope sweep oscillators. As an example in operation, supposing we have 70 volts on the plate of an 884 thyratron oscillator with a negative grid bias of 10 volts. The tube will not ionize or "break down" because, with 70 plate volts, the grid would have to be made about 8 volts negative for this to happen. If the synchronizing voltage applied to the oscillator grid were to have amplitude of slightly more than 2 volts the positive alternations of this sync voltage would bring the grid down to 8 volts or slightly less negative, and ionization would take place.

Supposing now that no sync voltage is being applied to the oscillator grid, and that we have available at B+ of Fig. 83-8 a supply voltage of 100. Whichever one of the sawtooth capacitors is connected through the range switch, now will charge through the resistance at Rs of that figure. The time rate of charge will depend on the sawtooth capacitance and the resistance at Rs. As the charge progresses, the capacitor voltage rise is forming the gradual upward slope of a sawtooth deflection voltage going to the sweep amplifier, exactly as in a television sweep circuit.

When the capacitor charge voltage rises to 90 volts there will be 90 volts on the oscillator plate. With our assumed grid bias of 10 volts negative there will be ionization or break down in the oscillator, as

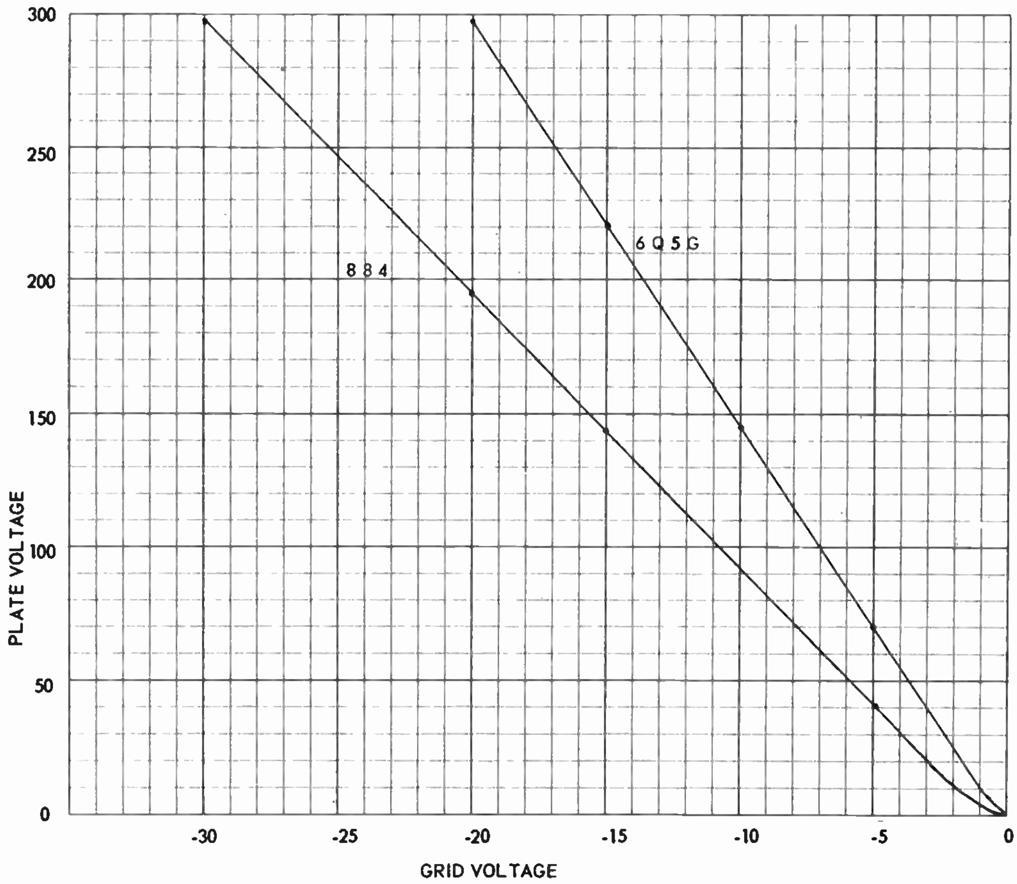


Fig. 83-9. Plate voltage versus grid voltage characteristics of two commonly used gaseous sweep oscillator tubes.

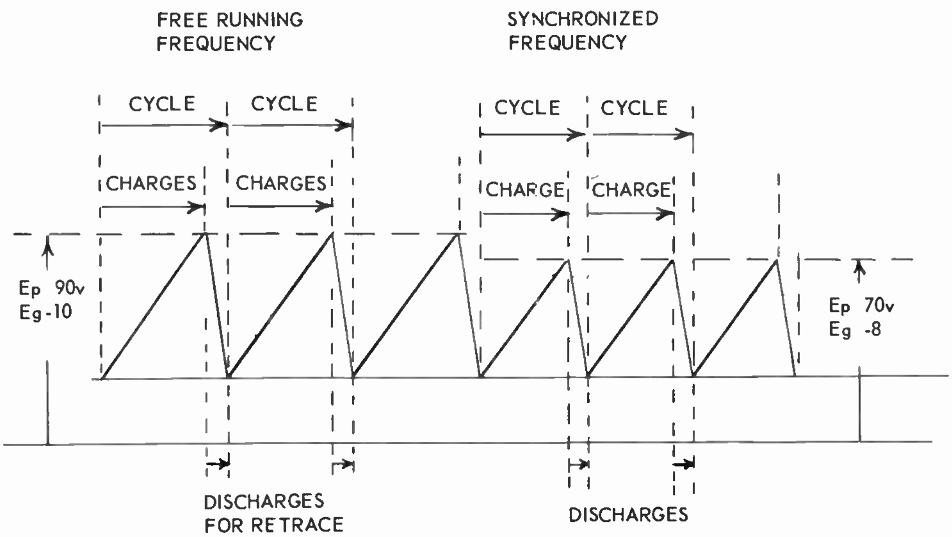


Fig. 83-10. Sawtooth waves at the free running frequency and after a synchronizing voltage is applied to a sweep oscillator.

shown by the relation between 90 plate volts and 10 grid volts in Fig. 83-9. The internal resistance of the gaseous oscillator drops almost instantaneously to a very low value, and the sawtooth capacitor discharges through the oscillator tube. The sudden drop of capacitor voltage forms the retrace portion of the sawtooth deflection voltage going to the sweep amplifier.

The action in the sweep circuit with no synchronizing voltage, as so far described, is illustrated at the left in Fig. 83-10. Each charge and discharge of the sawtooth capacitor completes one cycle of oscillation at the free running frequency, which is the sweep frequency with no sync voltage. What happens when the sync voltage is applied is shown at the right. Positive peaks of sync voltage drop the oscillator grid to 8 volts negative. Then ionization will commence every time the capacitor charge voltage and oscillator plate voltage rises to 70 volts. Essentially the same things happen with any other combination of B supply voltage, grid bias, and sync voltage amplitude.

5 The oscillator is synchronized by using the range switch and vernier frequency control to bring the free running frequency almost as high as the frequency of the sync voltage. It is the oscillator frequency that has to be changed because the frequency of the sync voltage cannot be altered, for this is the frequency of the voltage or current to be observed. Until the oscillator frequency is synchronized with the voltage or current to be observed, the trace on the screen of the CRT will be in rapid motion. As soon as the oscillator locks into sync the trace will stand still on the screen.

As the sawtooth capacitor nears discharge, its voltage and the voltage on oscillator plate drops so low, that electrons being emitted from the oscillator cathode are not drawn to the plate with enough velocity to continue the ionization action. Although this stops formation of new positive ions, a lot of them still remain inside the tube, mostly on the negative grid and all around it. This sheath of positive ions masks the negative charge of the grid itself and keeps it from opposing emission from the cathode. Because emission can continue, the tube remains conductive until nearly all these positive ions pick up electrons and become neutral atoms. This is the action called deionization. Until deionization is practically complete, the grid cannot stop electron emission and capacitor discharge, and a new charge cannot begin.

It is the time for deionization that limits the rate of charge and discharge of the sawtooth capacitor in a gaseous sweep oscillator circuit, and, as a consequence, limits the frequency of the sweep. Most gaseous oscillator circuits can reach sweep frequencies no higher than about 30,000 cycles. Careful design and construction will extend the limit to about 50,000 cycles.

POWER SUPPLY SYSTEMS. In the B power supply for an oscilloscope there must be a high-voltage

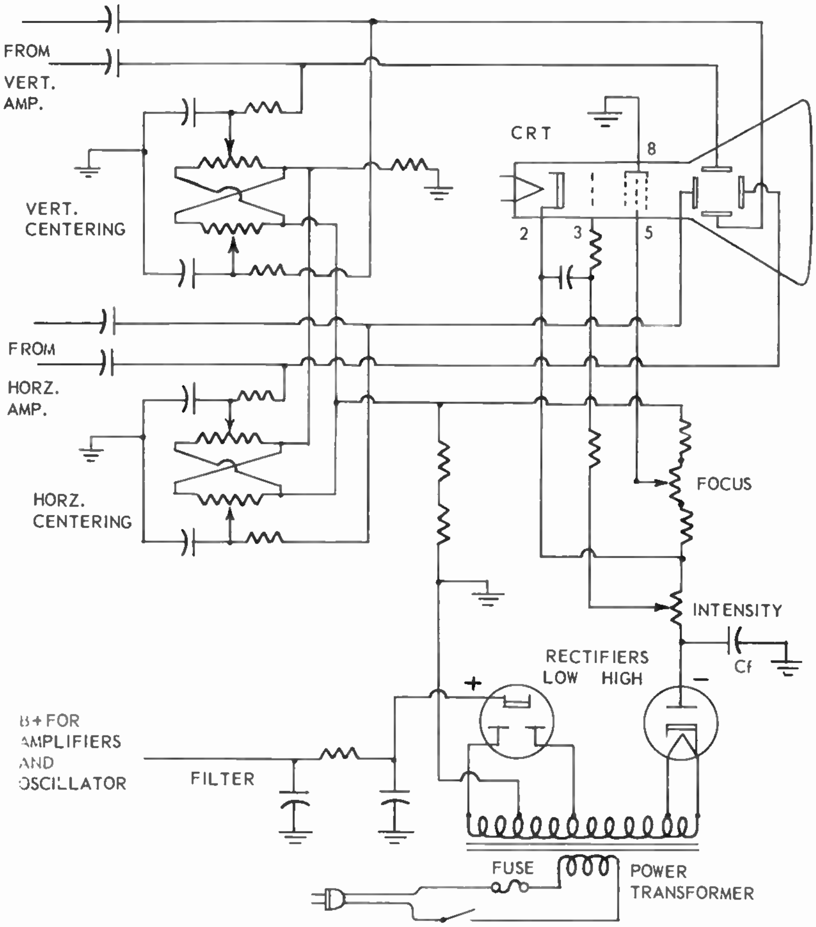


Fig. 83-11. Circuits of a high-voltage power supply such as used in oscilloscopes.

section for the deflecting plates and other elements of the CRT, also a low voltage section for plate, screen, and biasing circuits of the amplifier and oscillator tubes. Fig. 83-11 illustrates some of the features commonly found in power supplies of this type.

The low-voltage rectifier tube is a full-wave type connected to one end of the power transformer secondary in the same manner as any other full-wave rectifier, with about 300 to 350 a-c volts going to each plate from each side of the grounded center tap on this section of the winding. To provide high voltages for the CRT the transformer secondary is extended, and the far end or high-voltage end is connected to the cathode of the high-voltage rectifier.

The filter connected to the cathode of the low-voltage rectifier is shown as a resistance-capacitance type, but might be a choke-capacitance type. Voltage and current for amplifiers and oscillators is taken to the usual voltage divider and voltage dropping circuits from the output of this filter.

From the negative plate of the high-voltage rectifier a system of voltage dividing resistors extends to ground and also from the ground point back to the tap on the transformer secondary, thence through the secondary winding to the cathode of the high-voltage rectifier. The voltage divider resistance in combination with capacitor Cf from the rectifier plate to ground provide high-voltage filtering. The total divider resistance is on the order of one to two megohms. The single filter capacitor will have capacitance of 0.5 to 1.0 mf in most cases. This simple filtering is sufficient because the total high-voltage current ordinarily is only two or three milliamperes. Because of the small currents through them, the divider resistors may be of rather small wattage ratings in spite of their high resistances.

In the majority of oscilloscope high-voltage power circuits, and in the one of Fig. 83-11, the negative end of the circuit at the plate of the high-voltage rectifier is at the maximum potential with reference to ground. In the circuit illustrated the rectifier plate might be about 1,500 volts negative with reference to ground, which is at the end of the voltage divider system farthest from the rectifier.

Starting to trace from the plate of the high-voltage rectifier, the first divider resistance is a potentiometer for intensity control. The slider of this potentiometer goes to the control grid, element 3, of the CRT. Thus the control grid is made the most negative of all the elements. From just beyond the intensity control a connection goes to the cathode, element 2, of the CRT. This makes the cathode less negative than the control grid, or makes the grid more negative than the cathode.

Continuing along the voltage divider system, there is a fixed resistor and then another potentiometer which is the focusing control. The slider of this potentiometer is connected to the focusing electrode, number 5, of the CRT. This connection makes the focusing electrode less negative, and effectively more positive, than the cathode. The exact potential difference depends on the requirements of the particular CRT tube being used, but usually it will be between 200 and 600 volts.

After the focusing potentiometer in the voltage divider system there is another fixed resistor, then a line leading to the left, and down through two more fixed resistors to ground. The high-voltage anode, element 8, of the CRT is connected to ground. Consequently, this high-voltage anode is at a potential which is far less negative than that of the cathode, and is effectively far more positive than the cathode potential.

Because the high-voltage anode is grounded it is at the same potential as chassis metal, and you would get no shock by touching the chassis and the high-voltage anode terminal at the same time. But should you touch chassis metal and also a terminal for the CRT cathode, control grid, or focusing electrode at the same time, you would get a very severe shock which could be really dangerous and possibly fatal in case you have a weak heart. This is because all these other elements are at high voltages with respect to ground, although the voltages are negative to ground.

The deflecting plates of the CRT are connected to the high-voltage power supply system through the vertical and horizontal centering controls, just as with any other electrostatic deflection tube. If you trace the connections from the centering controls to the high-voltage divider system you will find that these connections come near the grounded end of the system. The centering controls are mounted on the panel of the oscilloscope, and because they are at potentials only a little negative with reference to chassis metal ground they are easy to insulate and safe to handle.

In the leads to the power transformer primary winding may be a cartridge fuse of 1 to 3 ampere capacity, or there may be fuses on both sides. The fuses will blow in case of shorted rectifier elements or other troubles causing overloads which otherwise might damage the internal parts of the oscilloscope. Often the transformer primary leads, or one of them, will go through a safety interlock switch that opens automatically when the instrument is removed from its cabinet.

Under no conditions should you touch any of the internal parts of an oscilloscope which is removed from its cabinet unless the power cord plug is pulled out of the power line receptacle. Do not depend on switches for disconnection. High voltages may appear at connections and parts which should be at low voltage to ground in case there has been a breakdown of insulation or a dead short or partial short through some capacitor.

Although the CRT voltages in an oscilloscope may not be so high as those for a picture tube in a television receiver, the scope may be more dangerous. This is because the high-voltage filter capacitor or capacitors in the scope are of rather large capacitance and they hold a great deal of energy. The high-voltage system of modern television receivers retains but little energy, and it is intentionally designed with such poor voltage regulation that any overload (such as the resistance of your body) instantly drops the voltage to a relatively low value. After an opened oscilloscope has been in operation, wait for a minute or two to allow complete discharge of the filter capacitance before touching anything. Better still, use a piece of insulated wire or a tool with insulated handle to short both terminals of each filter capacitor to ground, all this after the power is disconnected.

CONTROLS AND CONNECTIONS. The relative positions and the kinds of controls and binding posts or jacks on the panel of an oscilloscope will vary with the make and the model of instrument. The arrangement usually is somewhat as illustrated by Fig. 83-12. The face of the cathode-ray tube is at the top. Nearly always the intensity control is on the left and the focus control on the right of the tube face. Also, it is common practice to place the vertical centering and the vertical gain controls with the vertical input and ground terminals on the left side of the panel. The corresponding horizontal controls and terminals are on the right side.

The remaining controls and terminals may be almost anywhere. In the illustration there is a sweep vernier control for close adjustment of oscillator frequency in the upper center section of the panel, and

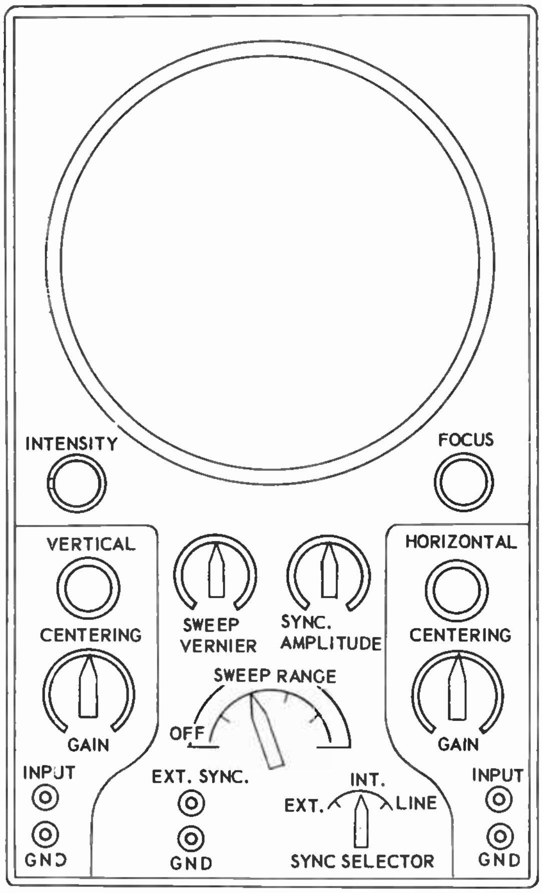


Fig. 83-12. Typical control arrangement on the front panel of an oscilloscope.

alongside is the control for adjusting the amplitude or strength of the sync signal voltage fed to the sweep oscillator. The range selector for sweep frequencies usually is somewhere in the center of the panel. There will be a sync selector switch with positions for external, internal, and line frequency voltages for synchronizing the sweep oscillator, also a terminal for connection of any external synchronizing voltage and often an additional grounding terminal.

As a rule there is no separate switch such as the one marked horizontal selector in Fig. 83-5, whose purpose is to connect the horizontal sweep amplifier to either the sweep oscillator or the horizontal input terminal. With the arrangement of Fig. 83-12 this purpose is served by turning the sweep range switch to its off position, which disconnects the oscillator output from the amplifier, then making a connection from the source of horizontal sweep voltage to the horizontal input terminal. This is a method widely used. On DuMont oscillographs the vertical controls and terminals are marked "Y-axis", and horizontal controls and terminals are marked "X-axis" controls and terminals.

The voltage to be observed on the oscilloscope must be brought from its source to the vertical input terminal through a shielded cable, with the cable shield connected to a ground terminal on the panel of the oscilloscope and, at the other end, to the ground side, the chassis, or to B-minus of the source. The principal reason for using a shielded cable is to prevent pickup of electric fields at power line frequency. These fields pervade all space where you are working. To note their effect, set the vertical gain of an operating oscilloscope fairly high and touch your finger to the vertical input terminal. On the screen will appear a 60-cycle trace with many harmonic and transient effects. Such an addition to any voltage which you want to observe leads to great confusion. A shielded and ground-connected cable should be used also for bringing any voltages from external sources to the horizontal input terminal and to the external sync terminal of the scope.

The internal connection from the vertical input terminal goes through a fixed capacitor to the vertical attenuator or gain control. This capacitor seldom is rated at more than 500 volts, and often at less. When either an a-c or d-c voltage connected to the vertical input exceeds about 300 volts it is essential to use an external series capacitor of ample voltage rating as a protective measure. This advice applies also to voltages applied to the horizontal input or to the external sync terminals.

As a general rule all the ground terminals on the oscilloscope panel connect to the instrument chassis and case, and thus are connected to one another. A ground connection at one terminal is the same as at any other terminal. When a ground terminal of the scope is connected to B-minus of a transformerless receiver having a hot chassis no other ground terminal of the scope may be connected to a water pipe

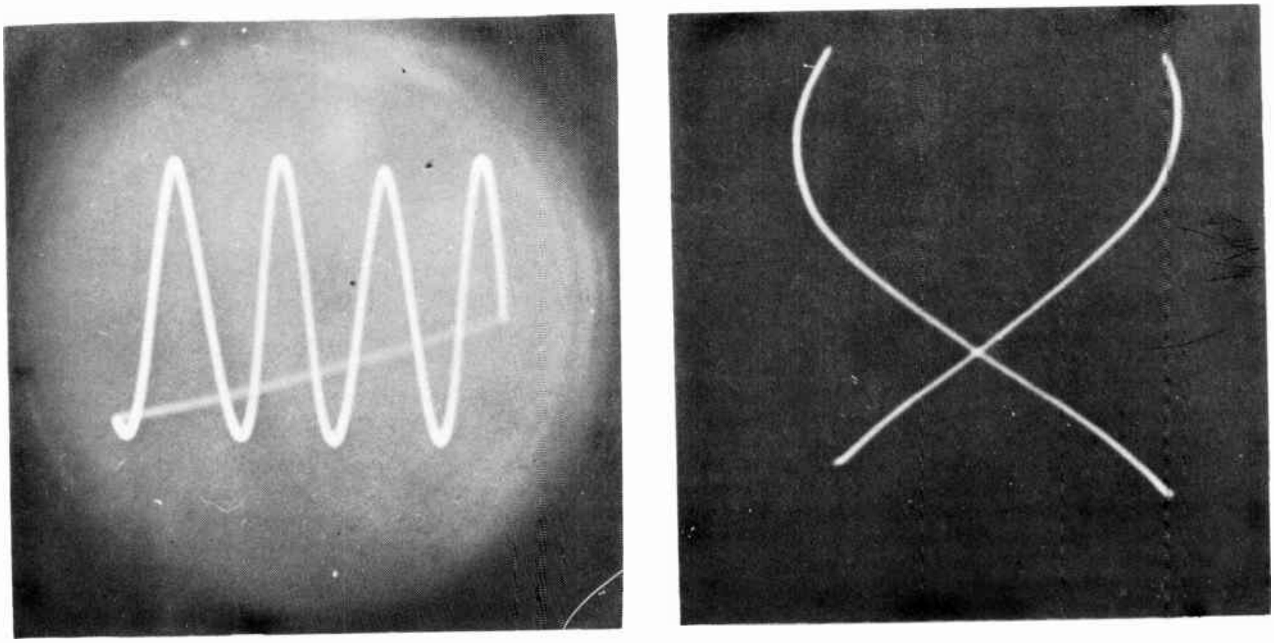


Fig. 83-13. Adjustment of the sweep frequency will bring several cycles onto the screen (left). Sweep frequency higher than vertical input frequency causes a split cycle to appear (right).

ground or to a service bench ground without danger of serious damage to the scope and possibly to the receiver. In other cases a water pipe ground usually helps to prevent pickup of fields at the power line frequency, since the housing of the scope then becomes an electrostatic shield.

OPERATING THE OSCILLOSCOPE. We shall assume that some voltage of unknown frequency is to be observed on the oscilloscope, and that the source of this voltage has been correctly connected to the vertical input. The steps for then placing the oscilloscope in operation and observing the waveform of the applied voltage are, in general, as follows.

1. Before turning on the power switch of the oscilloscope make these adjustments.

Turn the intensity control all the way down. In some instruments the "off-on" switch is attached to the intensity control, just as the "off-on" switch of a receiver may be connected to the volume control. In this case turn the intensity control only far enough to hear the switch snap to its "on" position.

Place both gain controls about one-third to one-half above their zero positions.

Place both centering controls at about mid-position.

Set the focus control at about mid-position.

Set the sweep range selector to the approximate frequency of the voltage to be observed, if you have any idea as to this frequency.

Place the sync selector switch at its internal position, which will allow synchronizing the sweep oscillator with the voltage applied to the vertical input. Other positions of this switch are used for tests to be explained later.

Set the sync amplitude control, which regulates the strength of synchronizing voltage, not more than one-fourth to one-third up from its zero position.

2. Turn on the power switch if it is separate from the intensity control, and wait a minute or more to see whether any kind of trace appears on the screen. If nothing appears, turn up the intensity control until a trace is seen. Never turn this control higher than necessary to produce a clearly visible trace. Too much intensity shortens the life of the CRT screen material. If the scope is left turned on between observations, turn down the intensity until the trace disappears.

3. Adjust the sweep vernier or fine frequency control to make the trace stand still on the screen or to have only slow movement toward either the right or left, if this is possible. If the trace continues to move rapidly sideways with all adjustments of the vernier control, try the sweep range selector at another position and again try adjusting the vernier control for a stationary or nearly stationary trace. Work back and forth between the range and vernier controls as necessary.

4. Usually it is desirable to have two or more complete cycles of observed voltage on the screen at the same time. Four cycles appear as at the left in Fig. 83-13. Part of the time of the fourth cycle is being taken up by the retrace time. With sweep frequency equal to vertical input frequency there will be

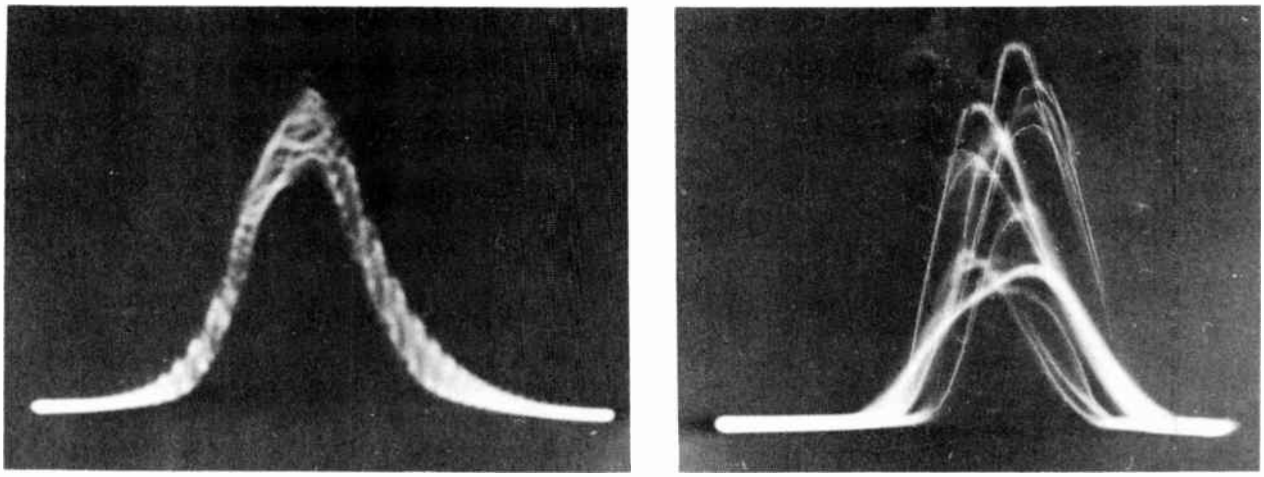


Fig. 83-14. A radio frequency on an audio-frequency trace, with sweep rate for the audio frequency (left). The effect of audio amplitude modulation on a radio-frequency wave (right).

one cycle, with the sweep at half the input frequency there will be two cycles, with the sweep at one-third the input frequency there will be three cycles, and so on. If the sweep frequency were made twice the vertical input frequency there would be a single split cycle, as pictured at the right, or with the halves in some other relation to each other.

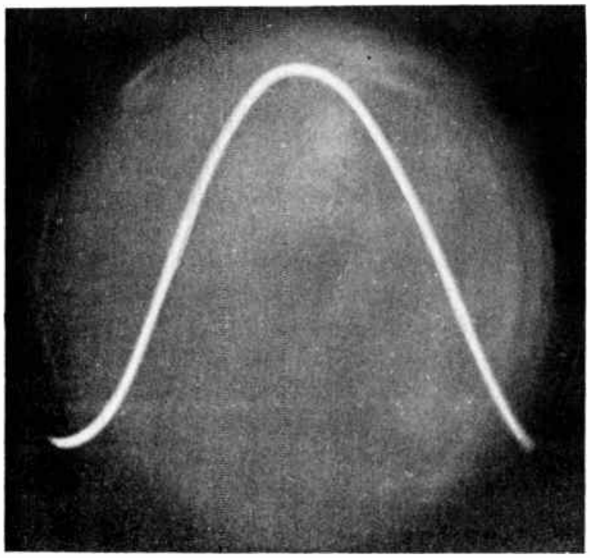
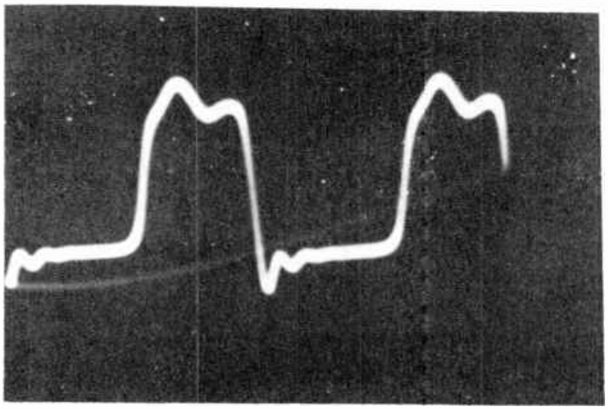


Fig. 83-15. A frequency response curve enlarged to nearly fill the screen area.



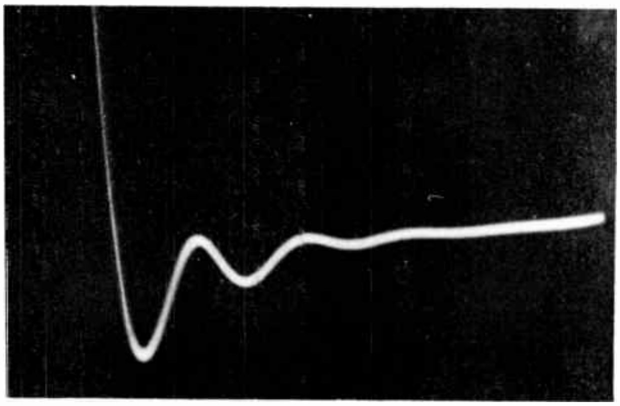


Fig. 83-16. Oscillation along the bottom of an audio-frequency trace (left) and an enlargement of the oscillation (right).

If the vertical input voltage contains more than one frequency, you can synchronize for any one, but not for more than one frequency at a time. At the left in Fig. 83-14 is shown a high radio frequency superimposed on an audio frequency trace, with the scope synchronized for the audio frequency. The little r-f waves will crawl back and forth along the trace. At the right is shown the effect of amplitude modulation at one frequency combined with a radio frequency trace. Either frequency could be synchronized.

5. Adjust the horizontal gain control and horizontal centering control to make the trace nearly fill the width of the screen while approximately centered between top and bottom.

6. Adjust the vertical gain control and vertical centering control to make the trace of the desired height for easy observation and for approximate centering between top and bottom. Fig. 83-15 is a frequency response trace enlarged and centered to nearly fill the screen area.

On most oscilloscopes the centering controls have enough range to move the trace all the way off the screen, either horizontally or vertically. Except at very high frequencies or with weak input voltages the vertical gain control will make the trace much larger than the screen area, so that only part of the trace may be seen. The horizontal gain control always will make the trace much wider than the screen.

These features are useful, as you may see from Fig. 83-16. At the left are two cycles of an audio voltage wave showing plate current saturation at the top and spurious oscillation below. Wishing to examine the oscillation more closely we may increase both gain controls and then center this one portion of the wave, with the result shown at the right.

7. Turn down the sync amplitude or lock-in control as far as possible while still holding a stationary trace. This will require readjustment of the frequency vernier as the sync amplitude is turned lower. Increasing the sync amplitude shortens the time of charge for the sawtooth capacitor on the sweep oscil-

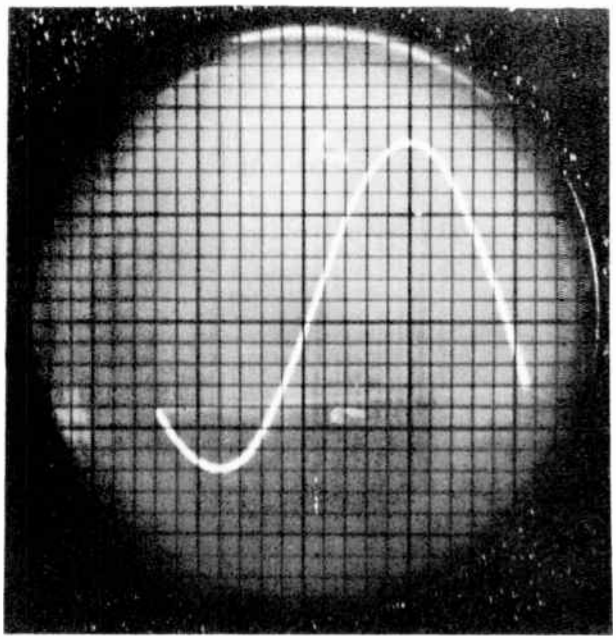


Fig. 83-17. This trace is drifting, due to too little sync amplitude.

lator, and thus increases the sweep frequency. Too much svnc amplitude will distort the sawtooth waveform and cause the trace to be crowded at one side. This fault you won't notice, but it will exist and will make observations unreliable unless sync amplitude is kept as low as possible.

A good procedure consists of first turning the sync amplitude control almost to zero. The trace then will drift toward one side or the other. Drift in one direction may be stopped by turning the frequency vernier one way or the other. Then the drift will reverse until the vernier is turned the other way. After getting the trace as nearly stationary as possible by adjusting the frequency vernier, lock the trace by turning up the sync amplitude control. The sync amplitude control will have no effect when the sweep frequency is higher than that of the observed voltage.

A trace of a single cycle in which there is drifting is pictured by Fig. 83-17. Incidentally, this illustration shows the use of a cross ruled graph scale in front of the CRT face. Such transparent ruled scales are permanently in place on some instruments, while with others the scale is detachable. The rulings allow measuring heights and widths of waves or parts of waves when estimating peak voltages and when making various comparative observations. The controls for gain and centering are used to bring the trace or any of its parts into any desired position with reference to the scale rulings.

7 8. The final adjustment is that of the focus control, to make the trace line as narrow and fine as possible and with the least possible fuzziness. The focusing usually requires readjustment every time the intensity control is altered. You will find also that other controls interact with one another, this being especially true of the controls for sweep frequency vernier, sync amplitude, and both vertical and horizontal gain -- as well as those for focus and intensity.

TELEVISION

LESSON NO. 84

TESTS WITH THE OSCILLOSCOPE

Training Your Mind

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind
C.T.I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON NO. 84

TESTS WITH THE OSCILLOSCOPE

When selecting an oscilloscope for service work the most important features or characteristics to be considered will include:

a. Sensitivity, or input voltage required to produce a trace of given size.

b. Frequency response, or maximum frequency of voltage on the vertical input for which gain in the vertical amplifier system remains reasonably constant or flat.

c. Impedance at vertical input terminals, which determines the loading or detuning effect on circuits to which the scope is connected.

d. Minimum and maximum sweep rates.

e. Size or face diameter of the cathode-ray tube.

Catalog descriptions specify all these characteristics for the various types and models of instruments.

The descriptions will list also a few or many of a variety of what may be called added conveniences. Among these extras are sync reversal, trace inversion, retrace blanking, intensity modulation, connections direct to CRT plates, single or driven sweeps, d-c amplifiers, and others. Each extra feature adds to the cost of the instrument, but each is worth its price if you need it or think you do. To complete our understanding of service oscilloscopes we shall first discuss the features possessed by all instruments, then some of the extras and what they are good for.

SENSITIVITY. Vertical sensitivity of an oscilloscope usually is specified as the fraction of an a-c volt, effective or r-m-s value, required to produce a trace one inch high on the screen when this voltage is applied at the vertical input terminals with gain or attenuator controls at their maximum settings. This measure of sensitivity is based on the deflection factor of the CRT and the gain in the vertical amplifier system, from input terminals to CRT plates.

Specifying oscilloscope sensitivity in volts per inch of deflection is a rather unfortunate and misleading choice for such a measure. It is misleading because the greater the fraction of a volt needed to produce a trace one inch high, the less will be the height for whatever a-c voltage actually is applied at the input. For instance, if one scope has vertical sensitivity of only 0.02 volt per inch and another has sensitivity of 0.05 volt per inch, the one with smaller sensitivity figure gives a bigger trace than the one with the larger sensitivity figure when the same voltage is applied to both.

If a-c voltages of opposite phase are applied to the opposite deflecting plates of a pair in the CRT itself, we may measure the result in r-m-s volts required for each inch of beam deflection. But we don't call this a measure of sensitivity in the CRT, rather we call it the deflection factor. The deflection factor voltage

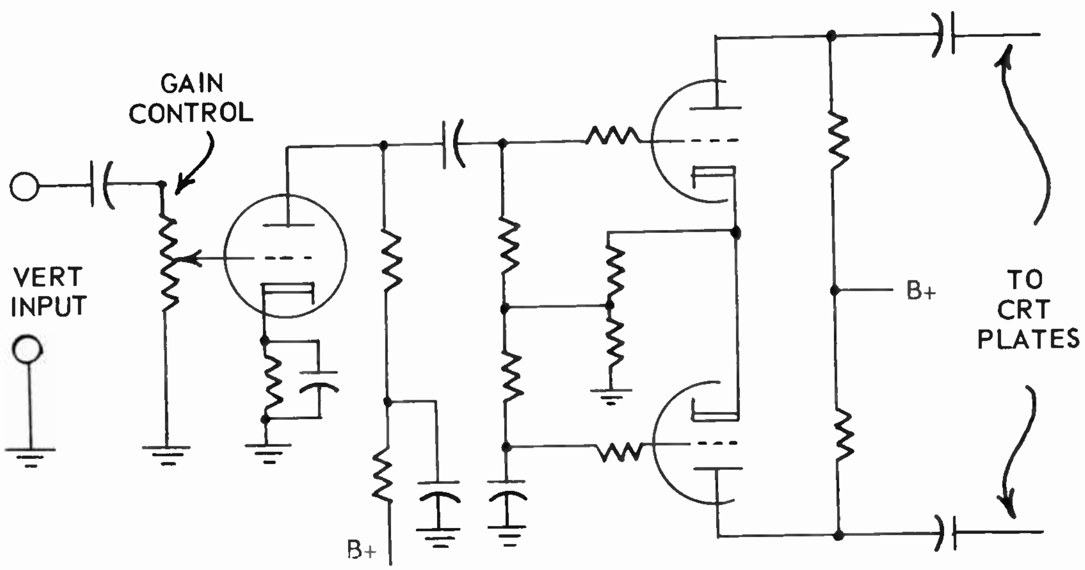


Fig. 84-1. An amplifier and gain control having flat response over a limited range of frequencies.

of a CRT is almost directly proportional to B voltage applied between the high-voltage anode and cathode, when voltage on the focusing electrode remains constant. That is, doubling the anode voltage will cut the deflection distance in half and will double the deflection factor, while halving the anode voltage will halve the deflection factor voltage.

The greater the anode voltage, within limits of tube rating, the more brilliant and sharper the trace may be made by correct focusing, but deflection distance is reduced. Lower anode voltage increases the deflection distance, but makes the trace fuzzier and less bright.

The deflection factor is not the same for both pairs of deflecting plates. The factor for the pair of plates toward the tube base is 10 or 15 per cent less than for the pair toward the tube face. The pair toward the base most often is used for vertical deflection, to give greater vertical sensitivity for the oscilloscope.

Since deflection factor voltage of the CRT tube is almost directly proportional to anode voltage, it is convenient to specify this factor as so many volts per inch of deflection for every kilovolt (1,000 volts) on the high-voltage anode. Cathode-ray tubes in oscilloscopes commonly have deflection factors ranging from about 22 to 38 volts per inch per kilovolt. Supposing that we have a tube in which this factor is 30. Deflection will be 1 inch for each 30 volts r-m-s across the deflecting plates of a pair when anode potential is 1,000 volts or 1 kv. Were the anode potential made 2,000 volts (2 kv) the factor would become 60, and it would take 60 r-m-s volts across the plates to deflect the beam 1 inch.

If our CRT did actually have a deflection factor of 30 volts per inch per kv, and were operated with 1,000 volts on its anode, what net voltage gain would be required in the amplifier system to provide sensitivity of 0.02 volt per inch for the scope? We simply divide the CRT deflection factor of 30 by the required sensitivity of 0.02 to find that the amplifier must provide voltage gain of 1,500 times, from input terminals to CRT plates.

For television servicing it is desirable to have vertical input sensitivity of at least 0.05 volt per inch, and sensitivities as great as 0.03 or 0.02 are preferred. Horizontal sensitivity is not so important. The sweep oscillator in any particular instrument will be designed to deliver plenty of sawtooth voltage to its horizontal amplifier. Any external voltage applied to the horizontal input terminals ordinarily will be amply strong to swing the beam far enough sideways. Horizontal sensitivity of about 0.5 volt per inch should be enough for service work.

FREQUENCY RESPONSE. The upper limit of frequency at which overall gain from input to CRT plates remains nearly constant depends on the design and construction of the amplifiers and the gain controls or attenuators. Sometimes the response is specified as being down so many decibels at some certain frequency, maybe 3 db down at 1 mc. If the 3 db refers to power, the gain drops to half its maximum at the specified frequency, if it refers to voltage the gain drops to 70 per cent of maximum at this frequency. The gain might be uniform or flat up to only a fourth or a third of the frequency for which it is down 3 db. or it might be flat to a greater percentage.

What kind of frequency response is needed depends on the kind of work for which the scope will be used. For alignment of television and f-m receivers any response found in any good make of service scope will be good enough at the high end, but it should be nearly flat down to frequencies of only a few cycles per second. If you wish to examine waveforms in the sync and sweep circuits of television receivers the vertical frequency response should be practically flat to a minimum of about 500 kc or 0.5 mc. Although the horizontal line frequency is only 15.75 kc, the pulses are in the form of square waves. A square wave contains harmonic frequencies of important strengths up to at least 10 times the fundamental, and if a square waveform is to be reproduced with good accuracy the response should be flat to 20 or more times the fundamental. For horizontal pulses this would require flat response up to 350 kc or more. It is understood that catalog listings of frequency response refer to sine wave inputs.

Horizontal frequency response need not be so good as vertical response. The horizontal response must handle sawtooth waves up to 15.75 kc, the horizontal line frequency for television. A sawtooth wave contains frequencies much higher than its fundamental because of the fast retrace. If the retrace time is 1/10 of the entire sawtooth cycle the effective retrace frequency is 10 times the fundamental, or is 157.5 kc. Then horizontal frequency response should be practically flat to 200 kc.

Fig. 84-1 shows fairly typical vertical amplifier circuits for an oscilloscope of limited frequency response, entirely satisfactory for television alignment, fairly so for observing vertical sync pulses, but not satisfactory for horizontal pulses. There is a simple potentiometer type gain control on the input, feeding the grid of a voltage amplifier. Push-pull amplifiers with resistance-capacitance phase inversion feed the plates of the CRT. Just as with any push-pull amplifier, power supply ripple voltages and second harmonics generated in the plate circuits tend to cancel out.

Fig. 84-2 shows vertical amplifier circuits with features allowing excellent frequency response. Tube *1* is operated as a cathode follower, with high impedance on the input side and low impedance in the cathode output containing the gain control potentiometer. Tube *2* is a voltage amplifier with a high-frequency compensating inductor or speaker in its plate circuit. Low-frequency compensation is provided by choice of values in the plate decoupling resistor and capacitor and by an adjustable grid resistance on the following tube. Tube *3* is a voltage amplifier and phase inverter with frequency compensation in its plate circuit. These frequency compensations are like those used in television video amplifiers. In other words, we have here a broad band amplifier system with flat response over a wide range and with minimum phase shift at high and low frequencies. Note also that cathode bias resistors are not bypassed, thus allowing degeneration to keep the response more nearly flat.

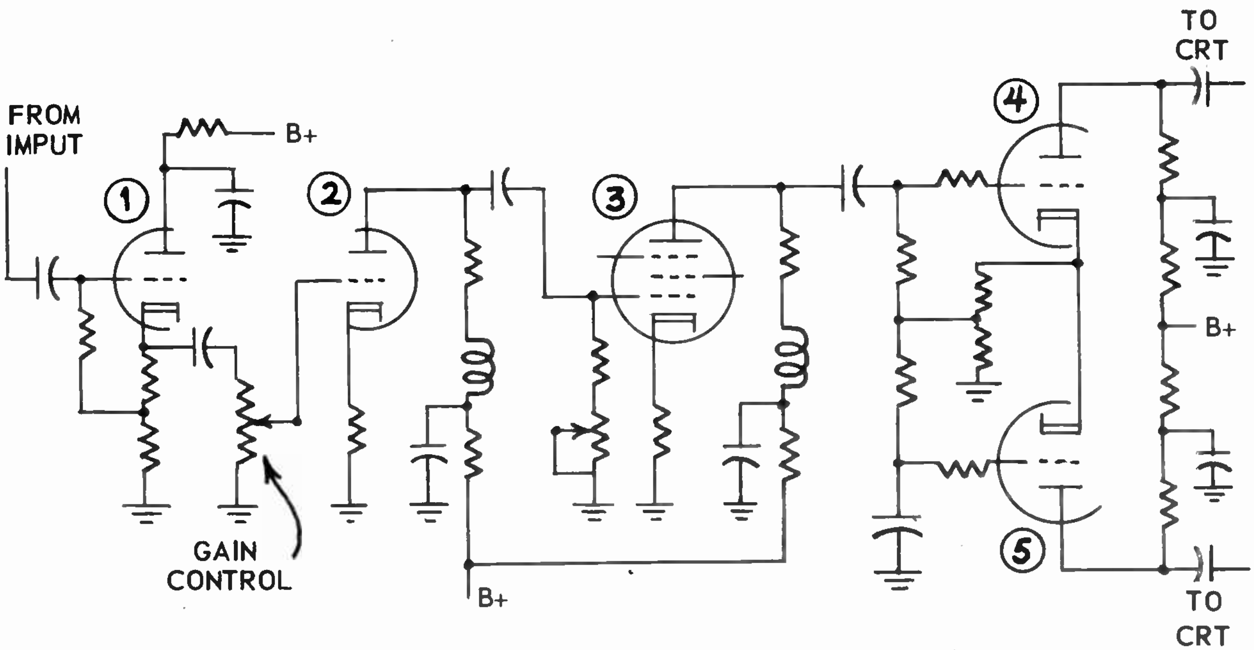


Fig. 84-2. A broad-band vertical amplifier system having an extended frequency response.

In actual practice the push-pull amplifiers of Fig. 84-2 might be voltage amplifiers, and they might feed additional push-pull beam power amplifier tubes whose outputs would go to the deflecting plates. This is because a broad band amplifier has relatively small gain per stage, and a number of stages are needed to provide the overall gain for good input sensitivity.

FREQUENCY DISCRIMINATION. When the observed voltage consists of both low and high frequencies, as in the case of a television sync signal, all frequencies should be uniformly amplified. This is why the flat portion of the response is more important than the limit of frequency at which waveforms are just visible. If all frequencies are not uniformly amplified there is frequency discrimination, and the traces do not follow the actual voltage waveforms.

There is serious discrimination between low and high frequencies in the ordinary potentiometer type of gain control constructed as at *1* in Fig. 84-3. For d-c voltages and for very low frequencies the fraction of input voltage applied to the grid of the tube is the ratio of resistance at *Rb* to total resistance *Ra*, or it is *Rb/Ra* just as with any resistance type voltage divider.

Actually, however, there are stray capacitances as represented in diagram 2. There is stray capacitance and grid-to-cathode capacitance to ground, at *Cb*, and capacitance from the grid terminal to the input terminal, at *Cc*. Here we have a capacitance voltage divider with total capacitance *Ca*. The capacitive reactances vary with frequency. Divider resistance *Ra* is shunted with varying reactance of *Ca*, resistance *Rb* is shunted with reactance of *Cb*, and resistance *Rc* is shunted with varying capacitance of *Cc*.

When you adjust the control slider for maximum gain, at the top of the resistance, the effect of the capacitances is only to reduce their reactance and the input impedance as frequency goes up. There is decreasing input impedance across whatever external circuit is furnishing the input voltage, and this circuit suffers increased loading. But the total voltage from the external circuit is applied to the grid of the amplifier tube at all frequencies, and there is not what we call frequency discrimination in the gain control.

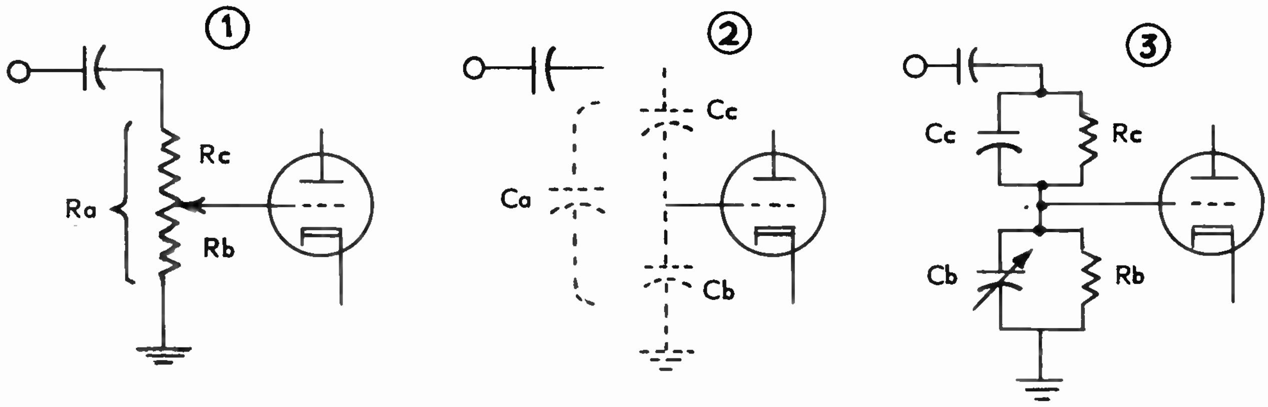


Fig. 84-3. Causes and cure for frequency discrimination in the gain control.

If the control is adjusted for anything less than maximum gain the percentages of input voltage delivered to the amplifier grid will vary with frequency unless you just happen to set the control at a point where there is balance between ratios of capacitive reactances and control resistances above and below the grid. In general, the lower you set the control in reducing voltage to the amplifier grid the worse is the discrimination between low and high frequencies.

As an example of what happens, assume that the total resistance of the control is 500,000 ohms, that it is set to deliver 10 per cent of the input voltage to the grid (so far as resistance ratio is concerned) and

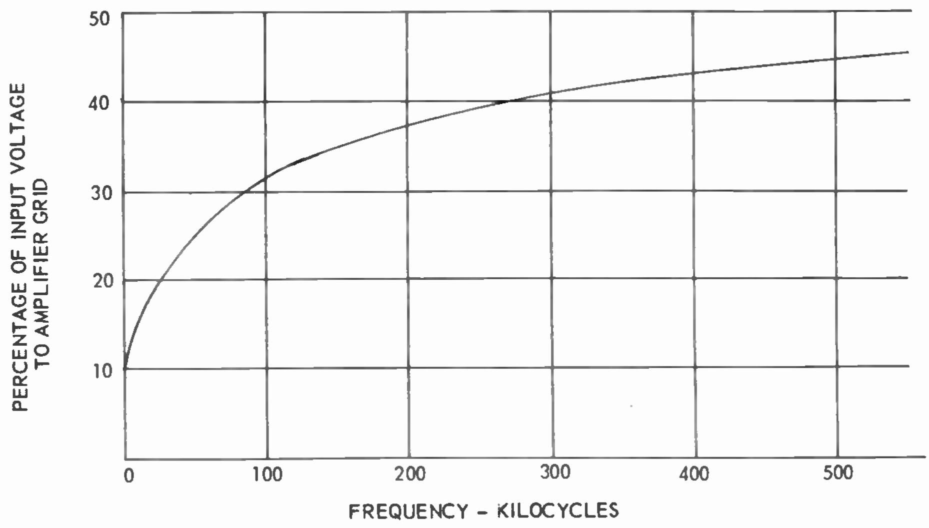


Fig. 84-4. How the observed frequency may affect the fraction of input voltage getting to the grid of the first amplifier tube.

that stray capacitances are 20 mmf on both sides of the grid. This would be an unlikely capacitance division, but it will serve for illustration. The actual percentages of input voltage on the amplifier grid will vary with frequency as shown by Fig. 84-4.

At 100 kilocycles the actual voltage to the amplifier grid has gone up from 10 to about 31 per cent of the input voltage, and at 500 kilocycles it is nearly 45 per cent instead of the assumed 10 per cent. At still higher frequencies the capacitive reactances would become so very small compared to the control resistances that the resistance ratio would have negligible effect. Voltage to the amplifier grid would depend almost entirely on the ratio of reactances which, in our example is 50-50 per cent. The curve would come continually nearer to 50 per cent.

Discrimination is worse with greater resistance in the gain control, and is lessened by using smaller resistance. But small resistance increases the loading of the circuit to which the oscilloscope is connected. For any particular percentage of input voltage delivered to the amplifier grid, frequency discrimination may be made unimportant by using the scheme shown at *3* in Fig. 84-3. There are two resistors, Rc and Rb, whose ratio would give the desired fraction of input voltage. These resistors are shunted by one fixed capacitor Cc and one adjustable trimmer capacitor Cb. If the ratio of capacitive reactances above and below the grid is made equal to or proportional to the ratio of resistances at any one frequency, this "proportionality" will hold for higher and lower frequencies and there will be little or no discrimination for this one fraction of input voltage delivered to the amplifier grid.

A practical way of using this general method of frequency compensation is shown by Fig. 84-5. A separate group of attenuating resistors and shunting capacitors is provided for each of several fractions of input voltage to be applied to the amplifier. A two-section selector switch makes connection from any one of these groups to the input terminal and to the amplifier grid. Switch positions on the diagram are marked *1, 10, 100,* and *1,000*. In position *1* the input terminal is connected directly to the amplifier grid. There is no

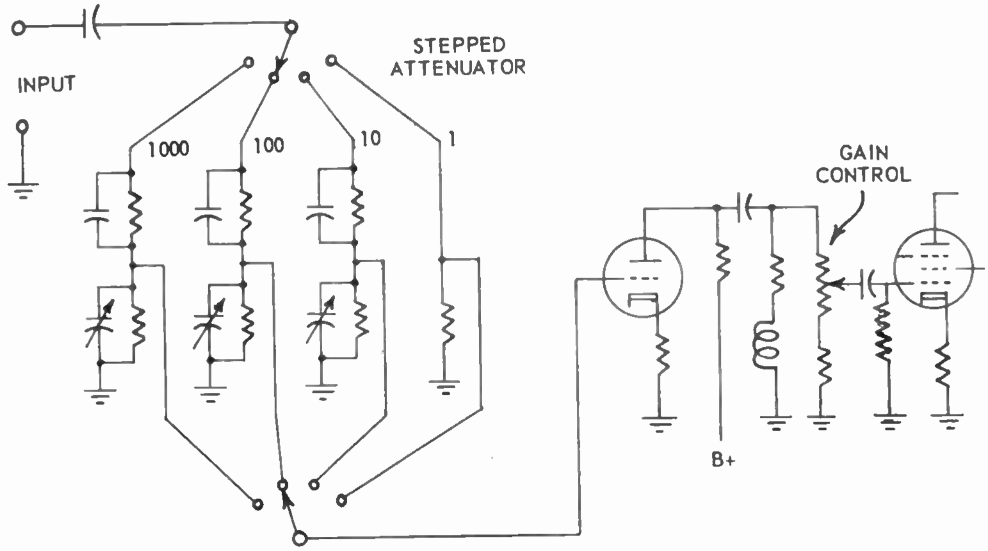


Fig. 84-5. Circuits for one style of stepped input attenuator with frequency compensation.

frequency compensation on this position, it is used only when maximum possible gain must be realized without regard to frequency discrimination.

The attenuating and compensating elements which are connected with the switch at *10* cause 1/10 of the input voltage to be applied to the amplifier. With the switch at *100* the amplifier gets 1/100 of the input, and at *1,000* it gets 1/1000. Between the limits of any two attenuator ratios the gain is regulated by a potentiometer-type gain control in the plate circuit of the first amplifier tube. With a potentiometer in this position its total resistance need be only a few thousand ohms, as compared with the half to one megohm commonly used for gain controls at the input. With the much smaller resistance of this gain control there is relatively little variation of resistance and capacitive reactance ratios with change of frequency. Still less gain control resistance may be employed when using a cathode follower system, as in Fig. 84-2.

4 *INPUT IMPEDANCE.* Input impedance of an oscilloscope is specified as a certain resistance shunted by a certain capacitance. The resistance usually is something between 1 and 5 megohms, and the capacitance between 20 and 50 or more mmf. As with any other testing instrument which is connected across measured circuits, the greater the resistance and the less the shunting capacitance the better are the results. Unfortunately, small shunting capacitance and high resistance are attained only at the expense of amplification or gain. If there is to be small input impedance combined with high gain, the amplifier must consist of a number of stages which brings up the cost of the instrument.

5 *SWEEP REQUIREMENTS.* A maximum sweep frequency of 20,000 to 30,000 cycles per second is ample for television servicing and for all usual sound receiver and audio amplifier servicing. In the case of television, no ordinary oscilloscope would have a sweep frequency high enough to allow accurately observing a single cycle at carrier frequencies, nor even at intermediate nor at the higher video frequencies. The next highest television frequency to be observed is that of one horizontal line, a frequency of 15,750 cycles per second. Nearly always you will want at least two line periods on the screen at one time, so the highest sweep frequency ordinarily used is 7,875 cycles per second.

It is possible to examine the principal features of one cycle of any voltage when the cycle takes up about 3/8 inch of horizontal space. On a 7-inch tube you can see about 17 such cycles, about 12 on a 5-inch tube, and 7 on a 3-inch tube. Frequency of an input voltage to be thus observed is equal approximately to the number of cycles on the screen times the number of cycles of sweep frequency. Multiplication will show that a sweep rate of 60,000 cycles with a 7-inch tube won't extend the observable input frequency much beyond the center of the standard broadcast carrier range, or to only about one megacycle. It would take a sweep of more than 250,000 cycles per second to get into the upper part of the video frequency range.

Other features of the sweep are more important, in our work, than maximum frequency. The sweep voltage at the plates of the CRT should be linear, meaning that the beam will travel across the screen at a constant speed. Supposing we were to have such extreme nonlinearity of sawtooth voltage as shown by the curve at the top of Fig. 84-6, and were applying a constant-frequency sine wave voltage at the input. Each cycle will take up the same time period as every other cycle, as shown along the bottom of the graph. But the beam will travel from left to right proportionately to the sawtooth voltage.

The uniform sine wave voltage would produce on the screen a distorted trace, as at the bottom of the figure. During the first cycle the sawtooth voltage would increase from *O* to *a*, and the beam would travel a proportionate distance to the right. During the second cycle the voltage would increase from *a* to *b*. This is a lesser increase then from *O* to *a*, and the beam would not travel so far. Every succeeding trace of one cycle would be shorter than the one before it, and you would not see a true picture of the input voltage. A number of sine wave cycles are used in order to clearly show the effect of nonlinearity. Any other voltage would be similarly distorted, although you might not realize what is happening.

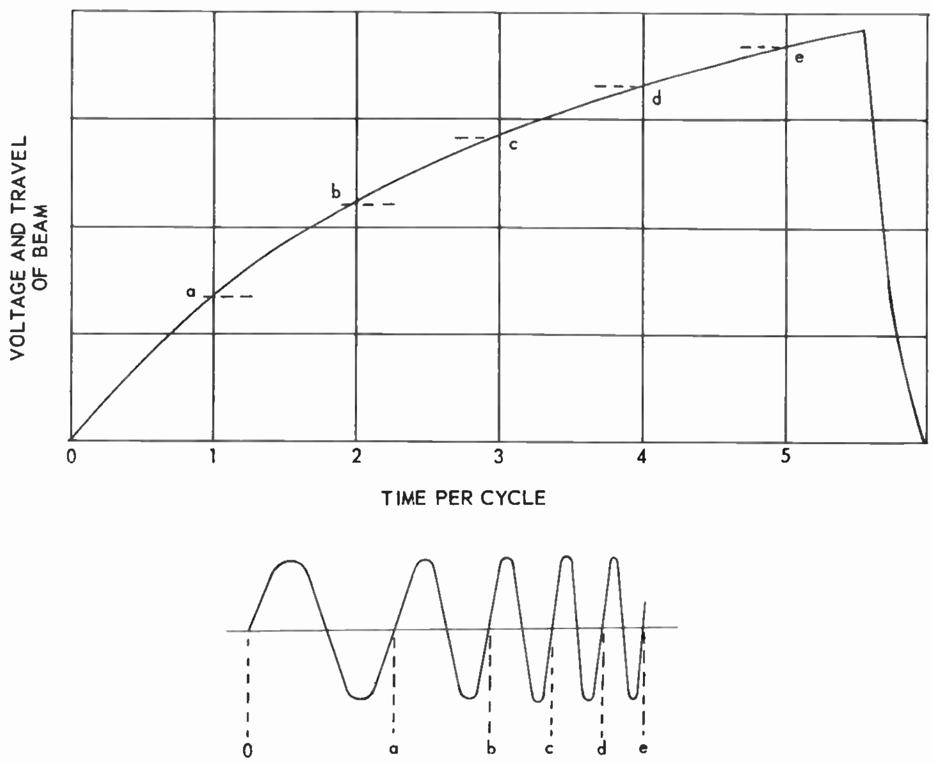


Fig. 84-6. The effect on trace form of non-linearity in the sawtooth deflecting voltage.

Any oscilloscope may be checked for sweep linearity by connecting to the vertical input a voltage of rather high frequency, say 100,000 cycles per second as obtainable from most r-f signal generators. The waveform of applied voltage is of no importance; its frequency will not vary appreciably during one trace. Adjust the scope internal sweep to get five or six cycles on the screen. If horizontal distances between all the peaks are equal the sweep is linear at this frequency, otherwise it is not linear. Slight non-linearity at the extreme left and right will do no great harm. Nonlinearity may result from action of the sweep oscillator, but more often it is the fault of the horizontal amplifier. The sweep may be an excellent sawtooth waveform at the oscillator output, yet be very bad at the CRT plates.

SIZE OF CRT. Face or screen diameters in popular service oscilloscopes are five inches and seven inches. Three-inch tubes, formerly used in nearly all scopes, still are found where small overall size of the instrument is of importance. The larger the screen diameter and area the easier it is to see fine details of traces such as those of television sync signals and various kinds of audio distortion. If the focus can be made sharp enough to trace a very thin, bright line there will be as much detail on a three-inch screen as on a seven-inch size, but it won't be so easy to see. Any screen diameter is satisfactory for alignment work, although the larger sizes always make for easier observation.

EXTERNAL SYNC. When first we discussed oscilloscope controls the sync selector switch was shown with three positions; internal sync, external sync, and line or 60-cycle sync. Operation with internal synchronization was explained at that time. The other two positions are used less often, but sometimes they are needed.

When you set the sync selector switch at "External" and connect any external source of voltage between the "External Sync" terminal and a ground terminal, this external voltage synchronizes the sweep oscillator. The external voltage is applied through the sync amplitude control potentiometer to the sweep oscillator grid. All the other controls work just the same as with the sync selector switch at "Internal". The range and vernier frequency controls must be adjusted for a sweep rate suited to the frequency of the observed signal. The sync amplitude control is used for holding the trace stationary. The controls for gain, centering, intensity, and focus are employed in the usual manner.

1 The external sync voltage may be the same as the voltage observed on the screen, which would require making the connections at *1* in Fig. 84-7, or it might be some other voltage used with the connections at *2*. External synchronization sometimes is used when the observed voltage contains many harmonic frequencies and has a waveform so irregular that internal synchronization from this waveform is difficult to hold without turning the sync amplitude control undesirably high. Then the sweep may be synchronized from some external source operating at the fundamental frequency of the observed voltage, or at some simple fraction of the fundamental for which the range and vernier frequency controls are adjusted. External synchronization may be used also when two separate voltages are applied to the vertical input to produce two simultaneous traces in order to observe their phase or amplitude relations. Then external synchronization at either of the observed frequencies may be used.

LINE OR 60-CYCLE SYNC. When the sync selector switch is turned to the position marked "Line" or "60-cycle" a voltage at power line frequency, taken from within the oscilloscope, is applied through the sync amplitude potentiometer to the grid of the sweep oscillator. The frequency range and vernier controls must be adjusted for line frequency or for 60 cycles. Then the sweep will be synchronized at this frequency while a voltage of any frequency at all is applied to the vertical input. All the other controls are used in the same way as with the sync selector at "Internal".

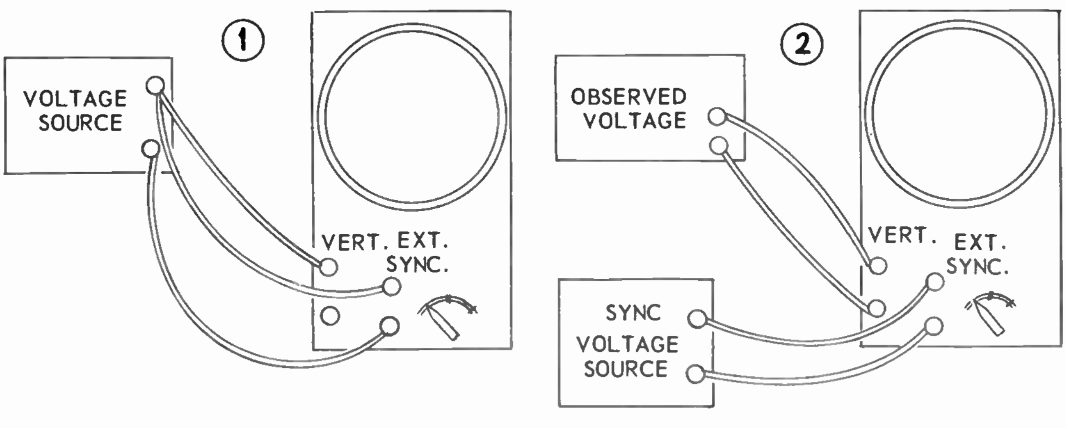


Fig. 84-7. Connections to the "External Sync" terminals of an oscilloscope.

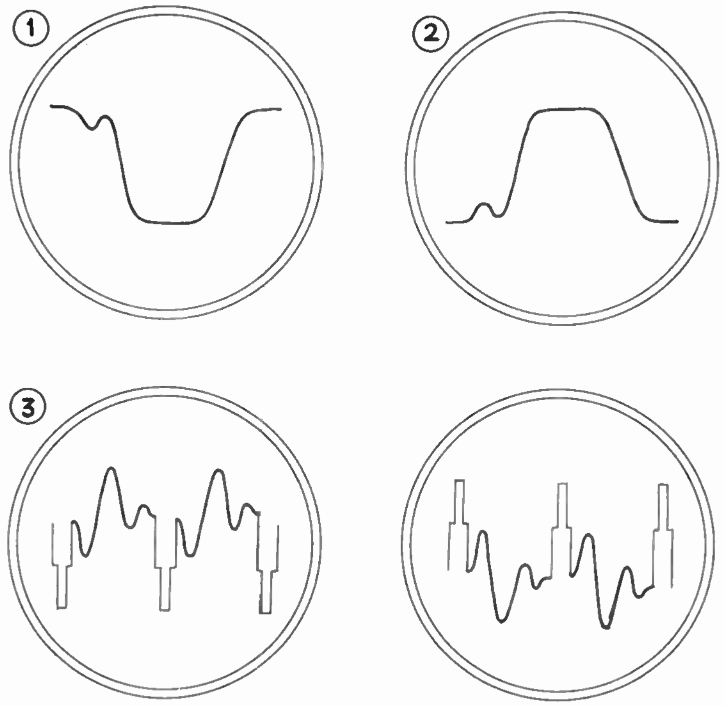


Fig. 84-8. Waveform traces may be inverted on the oscilloscope screen.

In some television oscilloscopes the 60-cycle position of the sync selector applies a sine wave at line frequency to the horizontal amplifier, through its gain control, and disconnects the sweep oscillator from this amplifier. Then the horizontal sweep is not a 60-cycle sawtooth wave but is a 60-cycle sine wave. This system may be used when employing a sweep generator for television or f-m receiver alignment, as will be explained in another lesson.

TRACE INVERSION. We get used to thinking of signal waveforms and other voltage waveforms as having a certain polarity. We think of a frequency response curve as showing gain increasing toward the top, and minimum at the bottom. But when you use an oscilloscope to inspect frequency responses or waveforms the polarity as it appears on the screen depends on the point in the receiver to which the vertical input in connected. On the grid side of an amplifier tube the polarity is in one direction, and on the plate side it is inverted.

Some oscilloscopes have a switch for inverting the trace on the screen, no matter what may be the polarity as taken from a receiver circuit. A frequencyresponse might appear as at *1* in Fig. 84-8. A trace inversion switch will turn it the other way up, as at *2*, and make the curve appear in this more familiar form. The curve and all the information it conveys are the same in both polarities. As at *3* a television signal might appear with sync pulses down. If you prefer looking at the signal with these pulses on the positive side, the waveform may be turned over with the trace inverting switch. Note that the curves and waveforms are not reversed between left and right, they are only inverted between bottom and top.

SYNC REVERSAL. When the voltage used for synchronizing the sweep oscillator is of sine wave form or of any form symmetrical above and below zero, as at *1* in Fig. 84-9, synchronization will be equally stable from either the positive or negative peaks. When observing any non-symmetrical voltage it is easier to obtain stable synchronization and a stationary trace when the more sharply peaked side of the voltage wave is of the polarity that triggers the sweep oscillator. An example is the television signal wave shown at *2*. The sync pulse side of the wave will trigger the oscilloscope sweep oscillator with certainty each time a pulse occurs. The picture signal side of the wave is too irregular for effective triggering.

The oscilloscope may have a sync reversing switch, usually marked positive for one position and negative for the other position. This switch is used by trying it in both positions, and using the position that allows a stationary trace with the lowest setting of the sync amplitude control. This will be the position that syncs the sweep oscillator from whichever side of the voltage wave has the sharpest and most distinct peaks.

The polarity of any voltage may be reversed between any source and any load by wiring a two-pole double-throw switch as shown at the bottom of Fig. 84-9. With the contacts closed toward the left of the switch, as in diagram *3*, the *A* side of the source is connected to the *C* side of the load or output, while *B* of the source is connected to *D* of the load. When the switch is thrown to its other position, as in diagram *4*, we have *A* of the source connected to *D* of the load, while *B* of the source is connected to *C* of the load.

7 When an oscilloscope vertical amplifier is of the push-pull type, sync reversal may be provided by taking the sync voltage for the sweep oscillator from either one or the other of the push-pull tubes. While the grid

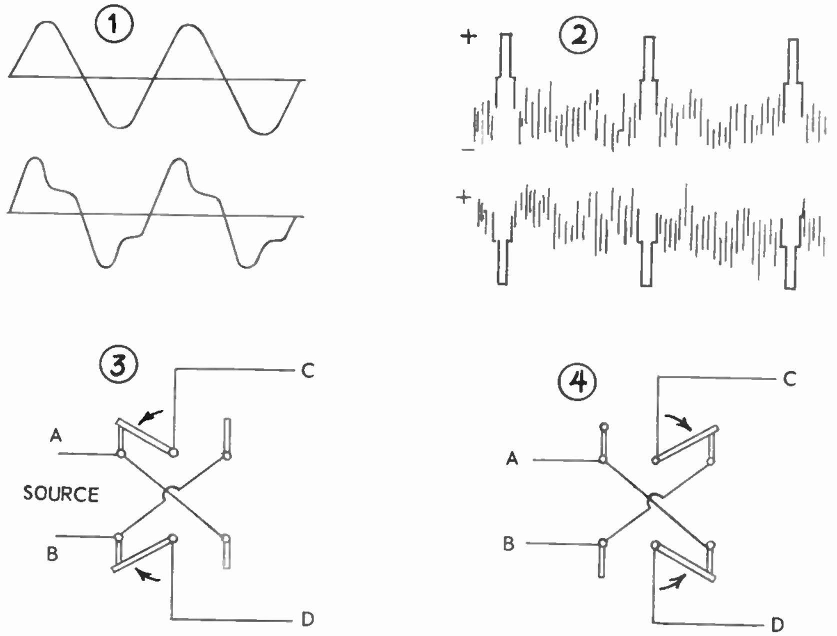


Fig. 84-9. The most stable synchronization is obtained from the sharper peaks of a voltage wave.

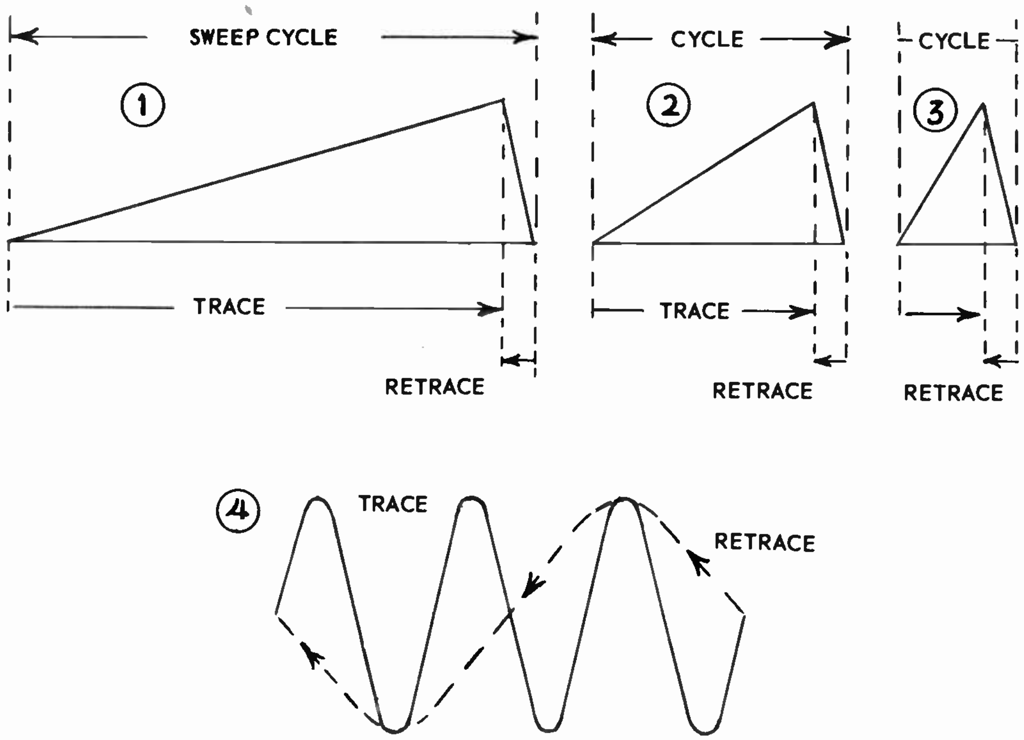


Fig. 84-10. Higher sweep frequencies increase the retrace time in relation to the time for forward traces.

of one of these tubes is going positive the grid of the other is going negative, and the plates likewise are of opposite polarity at any given time.

RETRACE EFFECTS. When using the internal sweep of the oscilloscope the beam in the CRT travels from left to right during the rise of sawtooth voltage from the sweep oscillator and retraces from right to left during drop of the sawtooth voltage. The retrace time is the time required for the sawtooth capacitor to discharge to the point where recharge again commences. This time for retrace remains fairly independent of change of sweep frequency.

When the sweep is at a low frequency, as represented at *1* in Fig. 84-10, the retrace time is a very small part of the total sawtooth and is a small part of each complete sweep cycle. At higher and higher sweep frequencies, as represented at *2* and *3*, the retrace time becomes a greater percentage of the total cycle and increases in relation to the periods of forward trace.

The frequency of any observed voltage remains constant during a forward trace and the following retrace. Assume, for an example, that there are four cycles of the observed voltage wave during each complete sweep period, and assume also that we have the condition of diagram *3* where the retrace time is about one-fourth of the total sweep cycle. Then, as shown by diagram *4*, there will be three cycles of the observed voltage during each forward trace, and one cycle during each retrace.

The beam travels faster during retrace than during forward trace, and this tends to make the retrace less bright. However, when using the highest sweep frequencies the beam is traveling so fast in both directions that the intensity has to be turned fairly high to observe the forward trace, and the retrace may become clearly visible to make a rather confusing pattern on the screen. The intensity control should be set low enough to leave the forward trace no brighter than absolutely necessary; then the retrace ordinarily will cause no difficulty.

a In some oscilloscopes there is a retrace blanking circuit. Most often a sawtooth wave taken from some plates in the horizontal system is passed through a series capacitor and shunting resistor to "differentiate" the wave and produce a sharp voltage pulse, just as in the differentiating filter leading to the horizontal sweep oscillator of a television receiver. The sawtooth voltage is selected to give a voltage pulse of negative polarity with reference to the CRT cathode, and this negative pulse is applied to the CRT grid to drive it beyond the beam cutoff voltage during retrace periods. In other cases an approximate sine wave is applied to the CRT grid in such phase relation to the sweep sawtooth that the positive alternation of the sine wave brightens the forward trace while the negative alternation dims the retrace.

INTENSITY MODULATION. During all ordinary operations with the oscilloscope the brightness of the trace is regulated only by the intensity control, which makes the grid of the CRT more or less negative with reference to the cathode. The brightness remains of a value determined by the setting of the intensity control. It is possible however, to vary the brightness at any desired frequency, if the oscilloscope is provided

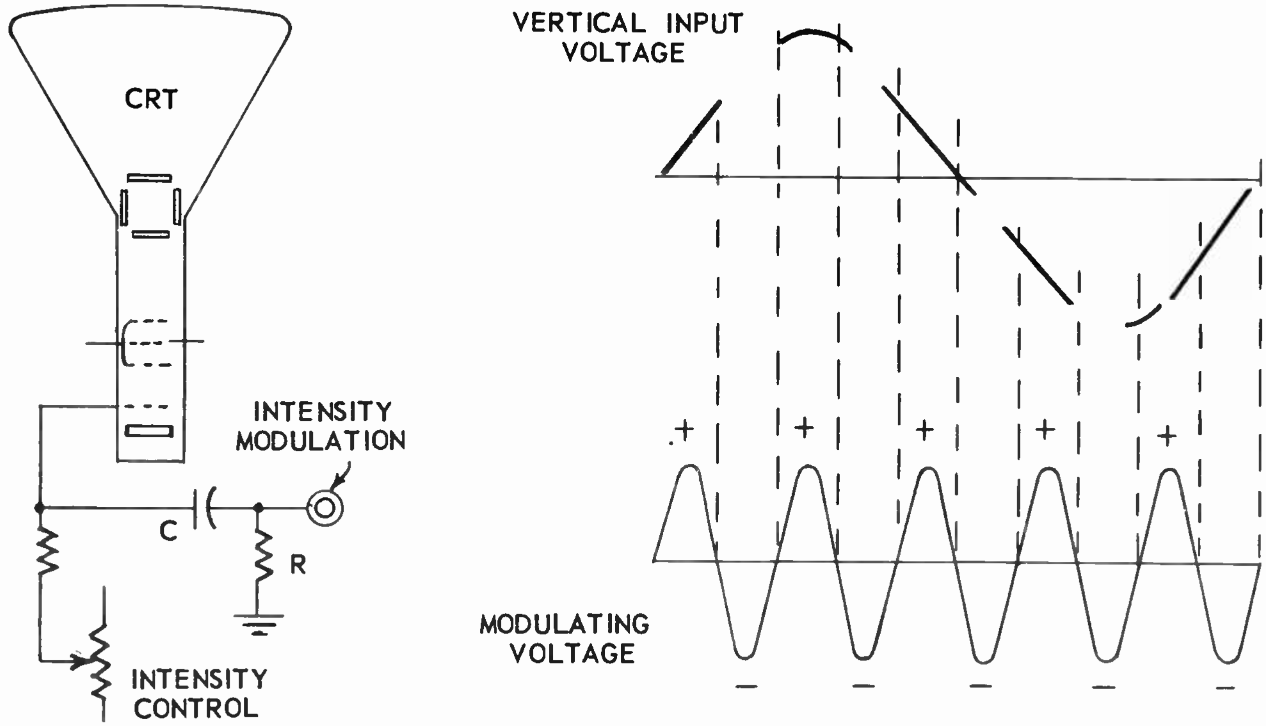


Fig. 84-11. An intensity modulation circuit, and the effect of such modulation.

with an "Intensity Modulation" terminal, and connections from this terminal to the grid as shown by Fig. 84-11. The internal connections usually consist of a series capacitor and a resistor to ground. Whatever alternating voltage is applied between the intensity modulation terminal and ground, is applied to the grid of the CRT.

At the right in Fig. 84-11 is illustrated the effect of an intensity modulating voltage on brightness of a trace. The trace is shown as consisting of a single cycle of any alternating voltage connected to the vertical input and ground terminals. The modulating voltage applied to the grid is of such frequency that five of its cycles occur within the period of one cycle of vertical input voltage, meaning that the frequency of the modulating voltage is illustrated as being five times the frequency of vertical input voltage.

Every time the modulating voltage goes through a positive alternation it makes the CRT grid less negative, and the portion of the trace thus affected becomes brighter. Every negative alternation of modulating voltage makes the CRT grid still more negative, and darkens or blanks the corresponding portion of the trace. The number of bright spots per cycle of vertical input voltage, also the number of dark spots, is equal to the frequency of intensity modulation voltage divided by the frequency of the voltage applied to the vertical input. This rule applies when the modulating voltage is some exact multiple of the vertical input voltage.

When division of the modulating frequency by the vertical input frequency gives some whole number plus one-half (1½, 2½, 3½, etc.) the number of bright spots or of dark spots will be twice the quotient of the division; For instance, with vertical input at 60 cycles and modulation at 150 cycles the division is 150/60 or

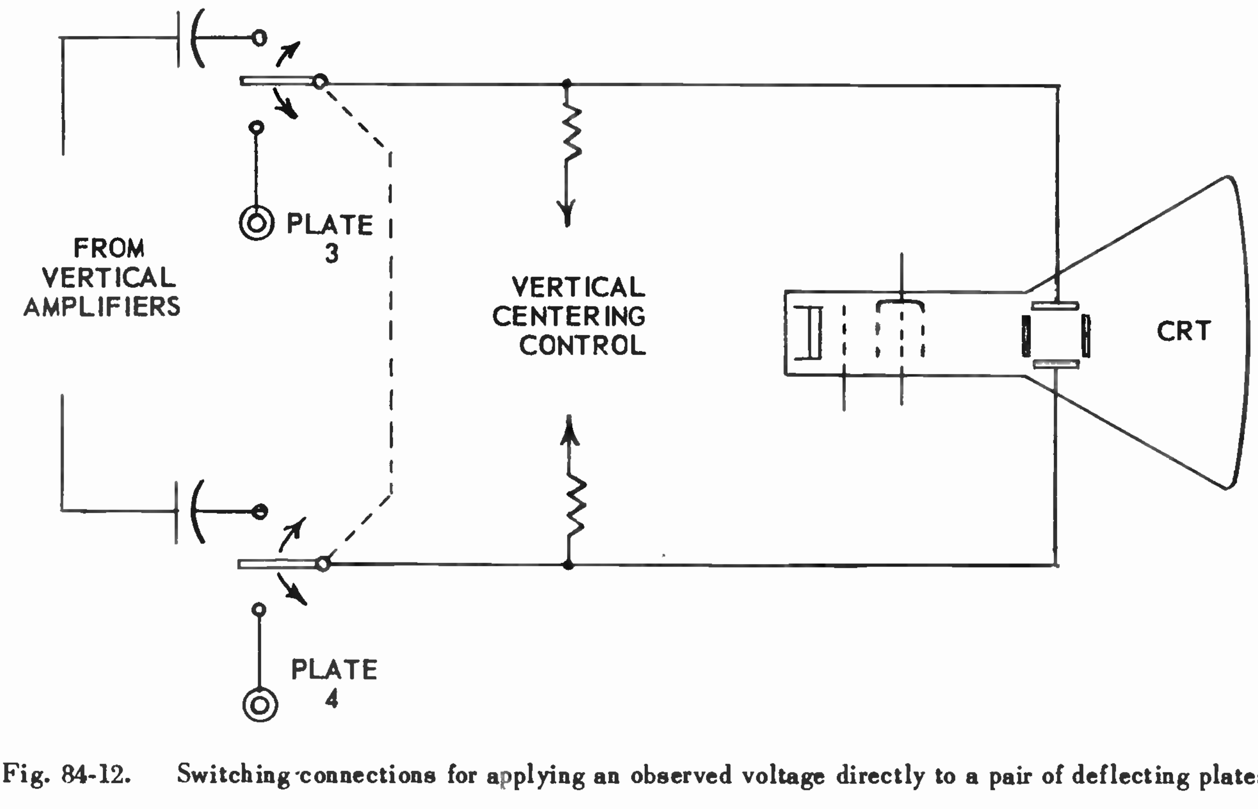


Fig. 84-12. Switching connections for applying an observed voltage directly to a pair of deflecting plates.

2½. Then there will be 2 times 2½ or 5 bright spots and 5 dark spots per vertical input cycle. This is because persistence of vision lets you see two superimposed traces at a time, and the extra half-cycle of modulating voltage puts in an extra spot.

Negative voltage on the CRT grid as determined by adjustment of the regular intensity control must always remain at least equal to the peaks of modulating voltage, otherwise the grid will be driven positive. Always commence with the intensity control at its lowest setting, and turn it up slowly until the alternate bright and dark spots appear. If the alternate spots appear with the intensity control at its minimum setting, the modulating voltage is too strong and must be reduced. Intensity modulation is not required for any of the usual operations in television and radio servicing.

DIRECT CONNECTIONS TO CRT PLATES. There may be provisions for making connections direct to the deflecting plates of the CRT instead of having these plates fed from the vertical and horizontal amplifiers. One method is illustrated by Fig. 84-12. A double-pole, double-throw switch is connected from its rotors to the two vertical deflecting plates. With the switch moved to one position the deflecting plates are connected as usual to the output of the vertical amplifier tubes. In the other position the plates are connected to two external terminals (shown as marked) for plates 3 and 4, which are the two employed for vertical deflection in many CRT tubes. Another double-pole, double-throw switch may be similarly connected for the horizontal deflecting plates, the horizontal amplifier tubes, and an additional pair of external terminals. Connections and switching for direct inputs often are at the rear of the oscilloscope, to avoid running long leads to the front panel.

With a direct connection, the distance the beam is moved by any given applied voltage is proportional to the deflection factor of the CRT, in volts per inch per kilovolt on the anode, and depends on the type of tube. With most oscilloscopes it takes anywhere from 15 to 30 r-m-s volts per inch of deflection. The leads which bring any external voltage to the plate terminals of the oscilloscope will usually connect right through to the centering controls and the high-voltage filter and voltage divider system of the scope. Consequently, it nearly always is essential to connect a blocking capacitor in series between the external lead and terminal. For low frequencies the blocking capacitance should be 0.25 mf or more, bnt may be proportionately smaller for high frequencies while still interposing small reactance. With most oscilloscopes the potential of the deflecting plates is close enough to ground or chassis potential so that the blocking capacitor need be rated at no more than 400 volts.

The CRT itself has an almost flat frequencyresponse to possibly 100 megacycles. With direct connection to the deflecting plates the limited frequency response of the amplifiers no longer has any effect and, in theory, the trace should follow the applied voltage with no frequency discrimination. Actually there is considerable capacitance and inductance in the internal parts of the high-voltage supply that remain connected to the plates, and there will be capacitance and inductance in any leads which bring the observed voltage to the plates. Consequently, the actual frequency response is far from unlimited in scopes of ordinary construction, although usually it is much higher with a direct connection, than through the amplifiers.

WAVEFORM OBSERVATIONS. When working on parts of a television receiver from the antenna through the video amplifier we are chiefly interested in frequency responses, or in the variation of gain with frequency, and are not interested in waveforms. From the video detector, through the sync and sweep sections, to the deflection system of the picture tube, we are chiefly interested in waveforms and in peak-to-peak voltages. Here it is necessary to employ the oscilloscope.

Waveforms in all circuits between the outputs of the sweep oscillators and the deflection elements may be observed with no received television signal. To have waveforms between the video detector and sweep

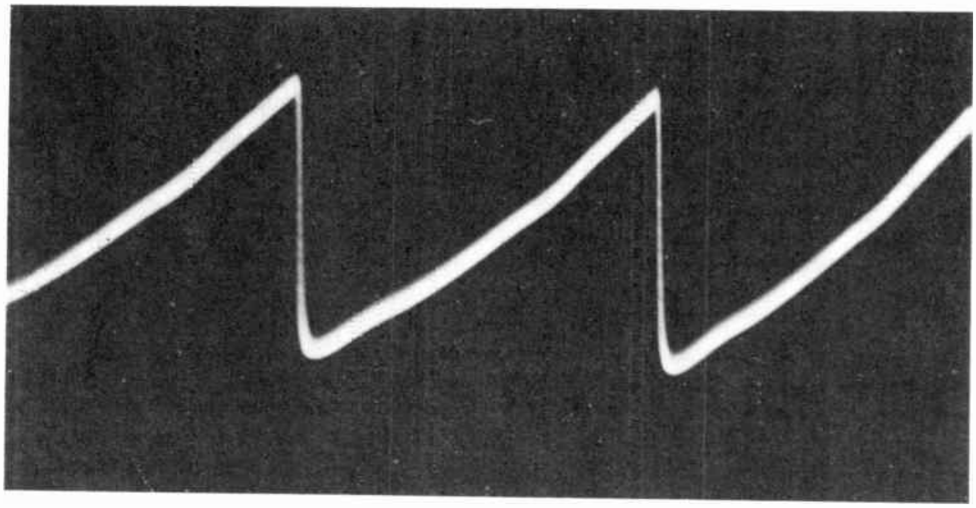


Fig. 84-13. With three cycles on the screen the center cycle may be examined in all its parts.

oscillators it is necessary to have a television signal, and it is just as well to have a received signal for all observations. With the oscilloscope connected, the receiver should be turned on and tuned to get the best possible picture. Sweep frequency should be made low enough to bring two or three cycles onto the screen at one time. This would mean two or three horizontal line periods, or two or three vertical field periods. This allows at least one of the periods or cycles to be observed without getting it too close to the edge of the screen. The center sawtooth cycle of Fig. 84-13 is easily examined.

A 30-cycle sweep will bring two vertical fields onto the screen, and a 20-cycle sweep will bring three fields. A sweep of 7,875 cycles will show two line periods, and 5,250 cycles will show three lines. The procedure is to adjust the sweep frequency controls for something near whichever of these frequencies you intend to use, then adjust the vernier frequency control and the sync amplitude control for a stationary trace.

The observed waveform from the detector should be very nearly that of the standard composite television signal. There is some slight variation between signals from different stations. The amplitude of this signal is altered by adjustment of the contrast control of the receiver, and if contrast is set too high the waveform will be distorted on the screen of the scope.

CURRENT WAVEFORMS. In a circuit containing appreciable inductance, capacitance, or both, the current waves and voltage waves will not be of the same form unless we have the condition of resonance. Consequently, it may be desirable or necessary to observe the current waveform as well as the voltage waveform. The amplifiers and the CRT itself of the scope are actuated by changes of voltage, not of current, so it is necessary to obtain a voltage waveform which follows the current.

This is done as illustrated by Fig. 84-14. Any lead or connections carrying the current to be observed is opened, then is reconnected with a resistor in series. The vertical input of the scope is connected across the resistor. The ground terminal of the scope must not be grounded, but connected only to the resistor. The

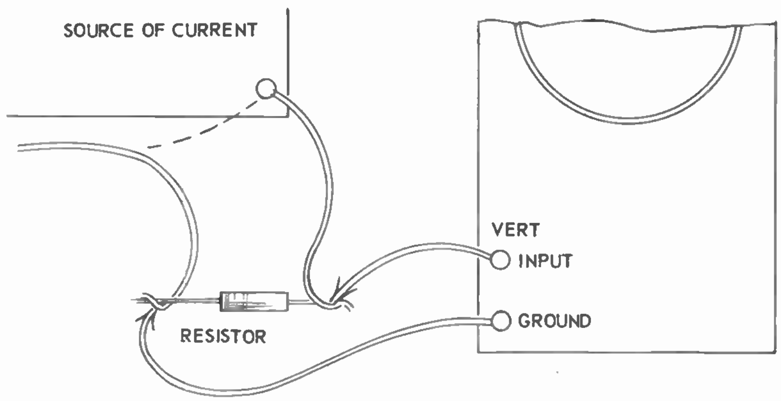


Fig. 84-14. Connections for observing a current waveform.

changes of voltage across the resistor, which are the same as changes of current, go to the scope and are shown by a trace on the screen. The resistor must be carbon or composition, and of the least resistance which will give a voltage drop measurable on the scope. The needed resistance becomes smaller with greater currents and with increased sensitivity of the scope. One to ten ohms often is enough.

PEAK VOLTAGES. Service manuals issued by television manufacturers often contain pictures of typical waveforms with notations of the peak-to-peak voltage of each waveform for a receiver in good operating condition. As a rule the waveforms are not symmetrical above and below zero, and because of this the peak values cannot be measured by either a moving coil rectifier voltmeter or a vacuum tube voltmeter. These

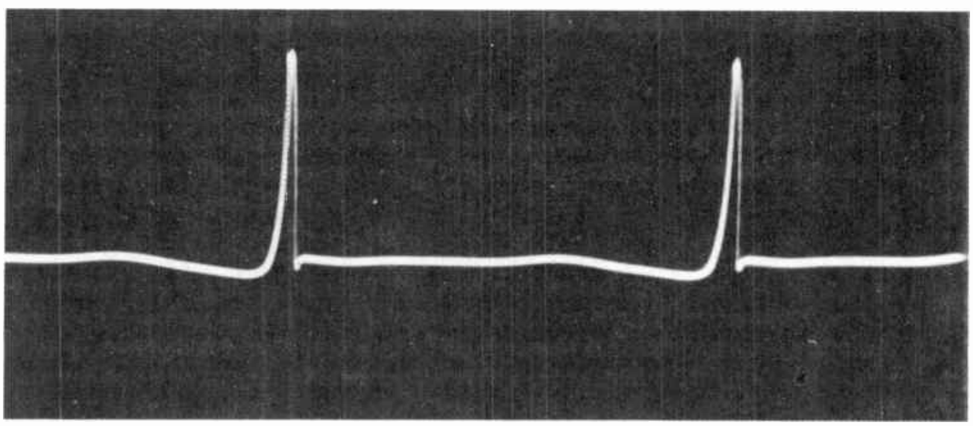


Fig. 84-15. There is no definite relation or factor between peak-to-peak and r-m-s voltages of a waveform such as this.

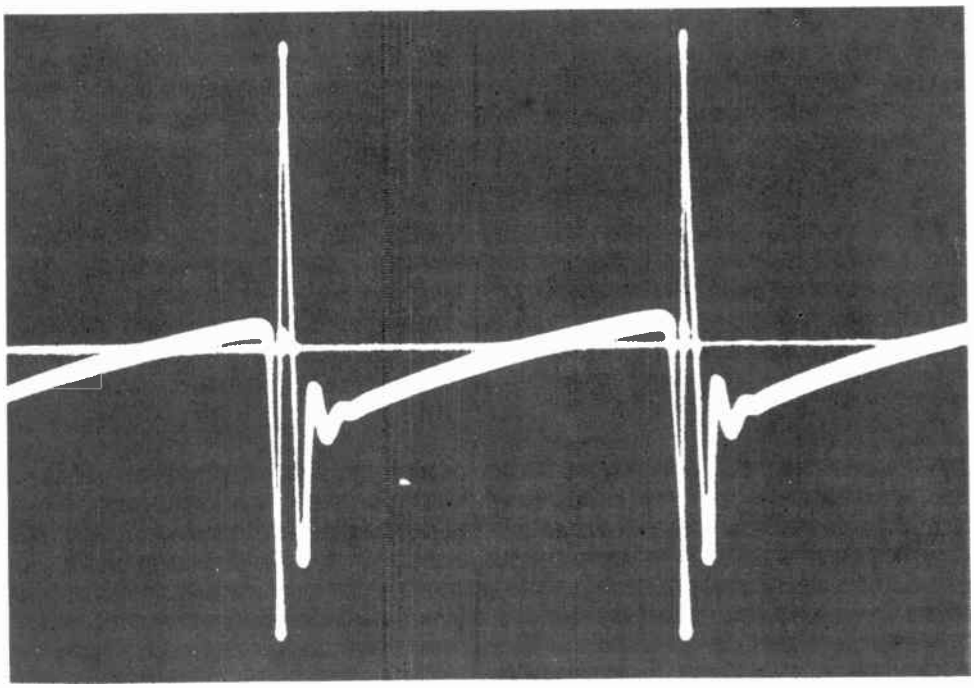


Fig. 84-16. The voltage or potential difference between any parts of this wave may be measured when necessary.

meters read effective or r-m-s values of alternating voltages. The peak value of a sine wave is equal to 1.414 times the r-m-s value, and the peak-to-peak value from maximum negative to maximum positive is equal to 2.828 times the r-m-s value. But now we are not concerned with sine waves. The peak-to-peak voltage of a wave such as shown in Fig. 84-15 would be more than 10 times its effective or r-m-s voltage as indicated by a meter.

Sometimes we wish to measure the rise or fall of voltage occuring in some one part of a wave. For instance, in looking for trouble in a television sweep oscillator circuit we might wish to know the charge voltage represented by the gradual upward slopes of the complex wave in Fig. 84-16. There would be no way of measuring it with a voltmeter, because the remainder of the wave would affect the reading.

Peak-to-peak voltages may be measured by comparison as shown in Fig. 84-17. As in diagram *1*, adjust the attenuator or gain control of the scope to make the trace of observed voltage of some height easily noted, such as exactly 1 inch, exactly 2 inches, or something like that. It is convenient to use a cross ruled graph scale. Disconnect the observed voltage from the vertical input cable, without altering the attenuator or gain control.

Next, as in diagram *2*, connect the vertical input cable to a source of adjustable sine wave voltage across which is connected an a-c voltmeter. Adjust this voltage to make the trace of the same heignt as be-

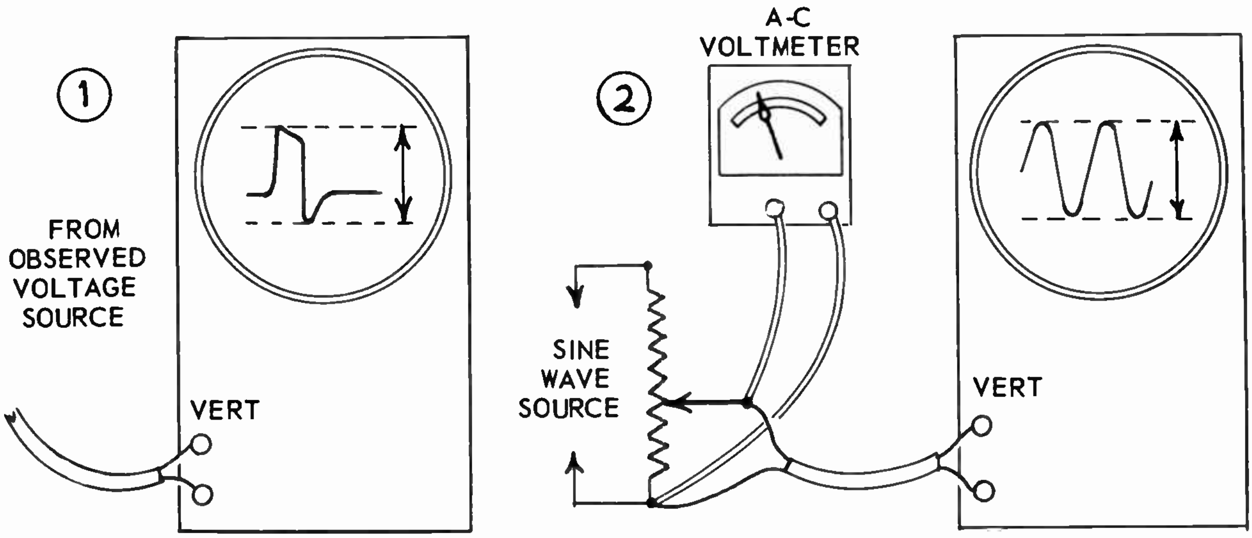


Fig. 84-17. Connections for measuring peak-to-peak and other voltages in a wave by comparison or by calibration of the gain control.

fore, still without touching the vertical gain control. Read the voltmeter and multiply the indicated voltage by 2.828 or 2.83. The product is the peak-to-peak voltage of the wave first observed.

Peak-to-peak voltages may be measured also by suitably calibrating the oscilloscope gain control. Make connections first as shown in diagram 2 of Fig. 84-17. Adjust the voltage to get a reading of 3.54 volts on the voltmeter. Adjust the vertical gain control of the scope to make the trace just 1 inch high. The scope now is calibrated for reading peak voltages in the ratio of 10 volts per inch of height, since 3.54 volts on the meter multiplied by 2.828 equals 10 volts peak-to-peak. The calibration is destroyed as soon as the gain control is moved.

Approximately correct peak-to-peak measurements may be made if you use the method just described to determine the volts per inch of trace height for many settings of the gain control and then draw a curve or graph relating heights to volts. When making any measurements of trace heights it is convenient to turn the horizontal gain control to zero and lower the intensity to prevent too bright a trace. This changes the trace to a thin vertical line which is easily measured.

If you have no means for adjusting the value of an external sine wave voltage it is possible to apply any known voltage to the vertical input and multiply this voltage by 2.828 to find the corresponding peak-to-peak value. Then the vertical gain control may be adjusted to give a trace height proportional to peak voltage. For example, assume that you take 6.3 a-c volts from a heater circuit or from a 6.3 volt terminal found on some oscilloscopes. Multiplying by 2.828 gives approximately 17.8 peak-to-peak volts. The vertical gain control then may be adjusted to make the trace 1.78 inches high. This calibrates the scope for 10 volts per inch of peak-to-peak volts – so long as the vertical gain control is not disturbed.

The methods of comparison and calibration for reading peak-to-peak volts hold good only through the range of frequencies in which the scope has flat vertical response. For higher frequencies it would be necessary to compare or calibrate with a sine wave voltage of nearly the observed frequency. Some r-f signal generators deliver fairly sinusoidal waveforms and some vacuum tube voltmeters have fairly constant frequency response up to something between 10 and 20 megacycles. If such a generator and voltmeter are used in the system of diagram 2 in Fig. 84-17 it is possible to make useful measurements well beyond the flat of the scope, since the vertical amplifier of the scope then is working at about the same frequency for the observed voltage and the comparison or calibration voltage.

DO NOT TEAR – CUT ALONG THIS LINE AND SEND IN FOR GRADING.

TELEVISION

LESSON NO. 85

TELEVISION ALIGNMENT

"Training Your Mind"

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind

C.T.I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON NO. 85

TELEVISION ALIGNMENT

1 The processes of aligning a television receiver consists of adjusting all the tuned amplifier circuits to resonance at certain frequencies. The frequencies to which the various circuits are aligned or tuned are those which allow necessary gains throughout the ranges of carrier frequencies, video intermediate frequencies, and sound intermediate frequencies. Since these are the three frequency ranges to be considered, it follows that our alignment operations may be grouped in three general classes.

First, to handle the carrier frequencies in the several channels, we align the tuner section which includes the r-f amplifier, the r-f oscillator, and the mixer. Second, to correctly handle the video intermediate frequencies we align or tune the video i-f amplifier coupling circuits when there is a split sound system, or align what we have called the television i-f amplifier circuits when there is intercarrier sound. Third, to handle the sound intermediate frequencies we align the sound takeoff, the sound i-f amplifier, and the demodulator circuits.

If a receiver is incorrectly aligned, nothing will help it more than doing a first class alignment job. If the receiver is correctly aligned to begin with, nothing can do much more harm than fooling with the alignment adjustments. If a receiver has been working satisfactorily until something suddenly goes wrong, there is very little chance that the trouble is misalignment.

Before touching the alignment of any receiver which has been operating, you should make the following checks. Examine the antenna, the transmission line, and all their connections. Tune the set to different channels; there may be trouble with the signal in one channel. Try adjusting all the controls on the front panel; possibly the operator doesn't know how to use them correctly. Examine all the tubes. If any glass-envelope tube shows no glow from its heater, or if any metal-envelope tube is not hot to the touch, replace that tube with one of the same type which you know is good.

2 Alignment really may be needed when any parts in any tuned circuit have been replaced, even though the new parts are exactly like the originals. Realignment may be required when a tube has been replaced in one of the tuned circuits. The set owner or an inexperienced serviceman may have turned some of the alignment adjustments. Realignment may be required due to changes of tube characteristics after the first few weeks of normal operation. Fixed capacitors and resistors change their values with age and long use, but in this case a realignment may hold good for only a short time because the faulty units are well along the road to becoming completely bad.

There are two general methods of alignment, both of which require introducing an r-f signal voltage at

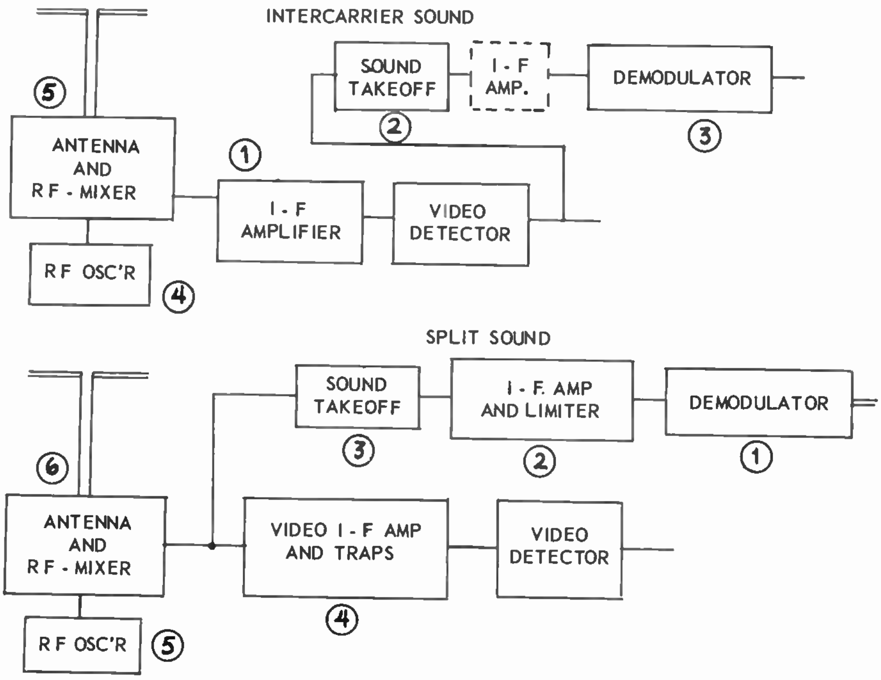


Fig. 85-1. The orders or sequences in which sections of television receivers may be aligned.

the input of the section to be worked on, and measurement of the output as the adjustments proceed. With the first method the signal source is an r-f signal generator and the output indicator is a vacuum tube voltmeter. With the second method, the signal source or voltage source consists of a sweep generator and marker generator, while the output indicator is an oscilloscope. We shall study both methods, but first there are a number of preliminary instructions and precautions to be observed no matter how you handle the job.

ORDER OF ALIGNMENT. If the entire receiver is to be aligned, the order in which the sections are worked on usually follows the numbering of Fig. 85-1. When there is an intercarrier sound system we usually commence with the i-f amplifier, then go to the sound takeoff and demodulator. Usually there is only a single amplifier tube in between, with its grid input tuned by the takeoff and its plate circuit by the demodulator transformer primary. The final steps would be in the tuner.

With a split or dual channel sound system the sound demodulator is aligned first, then the i-f amplifier and limiter are aligned to match the demodulator, and the takeoff, if tunable, is adjusted to suit the i-f amplifier. It is practicable also to commence with the i-f amplifier and limiter, then go to the takeoff, and next to the demodulator.

If there is no sound or poor sound, but a good picture, only the sound section would be aligned. Obviously the tuner and video i-f amplifier must be performing well. If there is a poor picture, but good sound, the video i-f amplifier should be given attention first. Next would come the tuner. The sound section should not be changed. If there is a poor picture and sound on only one channel, the tuner should be checked first, since both the video i-f and sound sections must be working satisfactorily to give good pictures and sound on the other channels.

TRANSFORMERLESS RECEIVERS. The ground terminals of signal generators, sweep generators, and oscilloscopes used during alignment ordinarily are connected together to equalize the potentials between all the instrument chassis and housings, and thus reduce any stray fields and undesired couplings. When working with power transformer type receivers the instrument ground connections can be made directly to the receiver chassis.

When working on transformerless receivers the chassis probably is hot, or may be made so if the plug of the power cord is inserted one way in the line receptacle. Should test instrument grounds be connected to a hot chassis it would be entirely possible to apply the full line voltage to the attenuators of most generators and of many oscilloscopes. Many attenuators are of low resistance, and the resulting current from the line would burn them out. Therefore, it is a safe rule never to make a direct conductive connection from the ground terminal of any instrument to a transformerless chassis, unless an isolation transformer is connected between the power line receptacle and the power cord plug of the receiver.

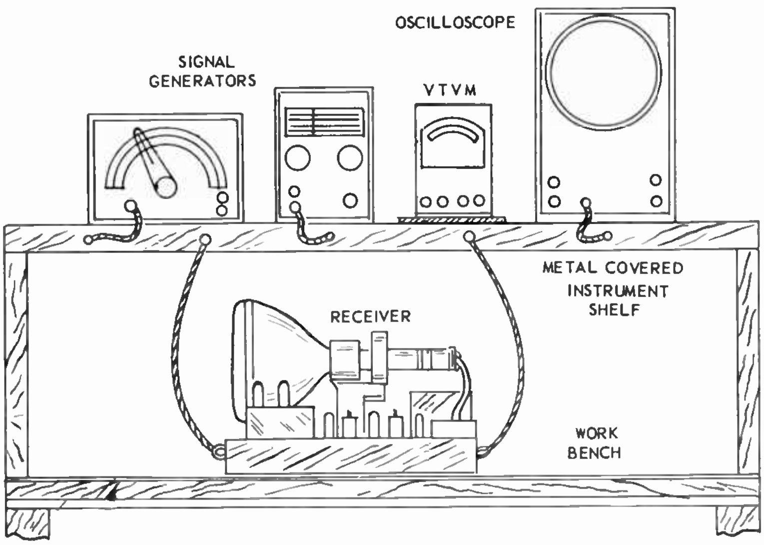


Fig. 85-2. Grounding and bonding connections for test instruments and receiver.

An isolation transformer is simply a power type transformer having a 1-to-1 voltage ratio and having its primary and secondary winding insulated from each other. The power rating of the transformer should be ample for any receiver likely to be handled, which would mean a minimum rating of 200 watts. If no such transformer is available, the next best thing is to make a grounding connection between instruments and receiver through a fixed capacitor of about 0.25 mf, rated at 250 volts or more. The reactance of such a capacitor at 60 cycles is about 10,700 ohms, which, at 117 a-c volts would limit the current to about 11 ma.

GROUNDING. When you connect any test instrument to any receiver other than a transformerless type one side of the signal circuit under test, and one side of the output circuit are completed through shielded test cables or leads. The other sides of these circuits ordinarily are completed through chassis metal of the receiver and instruments, or through the chassis grounds. Unless all the grounds are bonded together through conductors of low resistance the differences between r-f potentials in the various ground conductors will be enough to cause electrostatic fields which induce unwanted emf's in the circuits of the receiver. This may completely upset the normal performance and make your adjustments of little avail.

A method of bonding the grounds is illustrated by Fig. 85-2. All the test instruments are set on a metal-covered shelf above the work bench. Connections are made to the metal covering from at least one ground terminal of every instrument except the VTVM. The VTVM should be kept insulated from the metal covering, because in many tests its high-side and ground terminals both will be connected to points in the receiver circuits that are not grounded to the chassis. The metal covering of the instrument shelf should be connected to at least two well-separated places on the chassis of the receiver being handled. Oftentimes the work bench as well as the instrument shelf is covered with metal, but then something like a piece of fibreboard will have to be used over the bench when working on transformerless receivers. With such receivers the ground connections from the instrument shelf may be made to B-minus points in the receiver. The shelf covering is preferably non-magnetic, being of aluminum, brass, or copper in most cases.

Make all the ground connections before making the high-side connections. After the receiver and instruments are turned on and have had time to warm up, make any settings which produce some kind of steady output indication. Then touch the chassis of the receiver, the housings of all the instruments, and the metal coverings with your hand. If this causes any change of the output indication, the grounding is not sufficient or is connected to the wrong points on the receiver. Add more grounds or shift the original ones so that you do not affect the indications.

Any grounded shields which are on the tuner, any of the tubes, or any of the interstage couplers of the receiver should remain in place during alignment. Adjustments made without the shields will not hold good in normal operation.

SIGNAL INPUTS. Signal voltages from the generators used in alignment ordinarily are applied to the grid of one of the tubes in the receiver. To prevent short circuiting the grid bias voltage through the output circuit or attenuator of the generator it is necessary to use a fixed capacitor in series with the high-side lead to the grid. The capacitor should be a mica or ceramic type, usually of 20 to 100 mmf capacitance, although there is no objection to using larger sizes.

As shown by Fig. 85-3, the bias isolating capacitor may be fitted with an all-copper or bronze spring

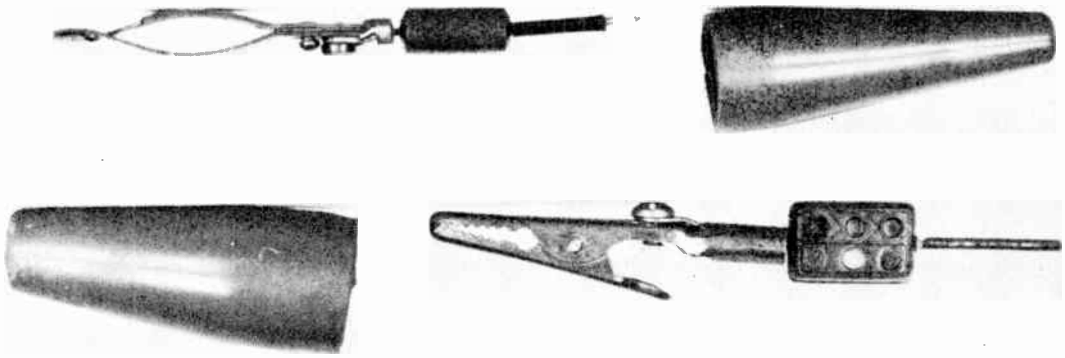


Fig. 85-3. Spring clips and rubber insulators are used with the fixed capacitors which prevent shorting the grid biases during alignment.

clip on one pigtail. The other pigtail is cut down to about a quarter inch, or is left somewhat longer and protected with a piece of spaghetti to leave about a quarter inch of bare metal exposed. A soft rubber insulating cover should be arranged to almost completely enclose the metal of the clip after the clip is attached to a grid connection, so that there is no possibility of shorting to nearby exposed con-

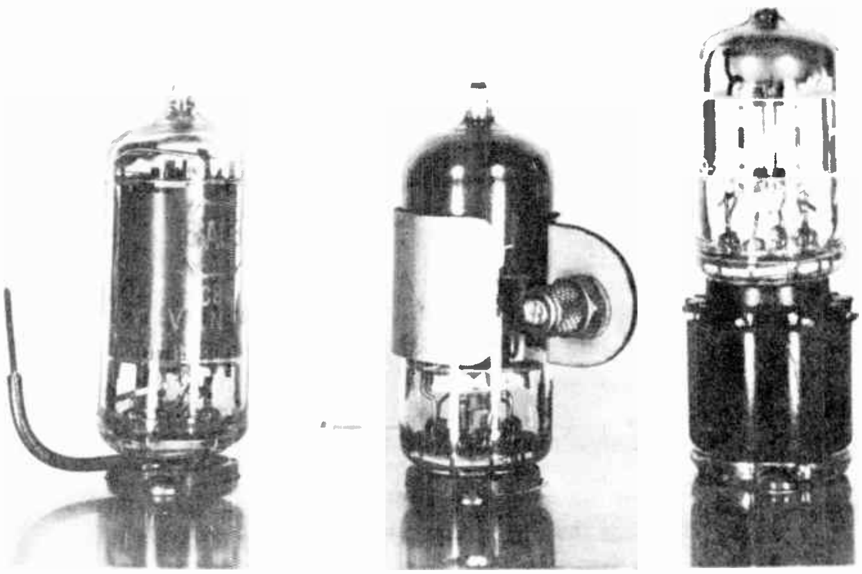


Fig. 85-4. Methods of feeding generator signal voltage to the elements of tubes from points on top of the chassis.

ductors. A spring clip on the end of the instrument cable then is attached to the exposed pigtail of the capacitor. This cable clip should be protected with another rubber insulator if there is any chance of touching conductors in the receiver. The insulators are removed from the clips in the picture.

In most receivers it is not difficult to make a connection at the lugs of tube sockets in the i-f and sound sections, although sockets in the tuner often are difficult or impossible to reach. When socket lugs are difficult or impossible to reach there are a number of ways in which you can connect or couple an input signal voltage to a grid or other element of any tube. Some of these methods are illustrated by Fig. 85-4.

The tube at the left has been removed from its socket and a short piece of wire attached to one of the base pins. Take about an inch of insulated solid wire, such as hookup wire, and bare one end for about a quarter inch and the other end for a little more. Turn the quarter-inch end into a single loop which slides tightly onto the selected base pin. With the tube replaced, the insulation on the wire will keep it away from chassis and socket metal. Bend the free end up for connection of the clip on the bias isolating capacitor.

On the tube at the center is a metal sleeve which slides down over the glass envelope to nearly surround the internal elements. The clip on the end of the signal generator cable is attached directly to the sleeve. There is no need for a bias isolating capacitor, since there is capacitive coupling from the sleeve to the tube elements, with the envelope and internal vacuum acting as dielectric. This capacitance is small, and it calls for a stronger signal voltage than when making connection through the usual capacitor. You can bend one of these coupling sleeves from any small, thin piece of aluminum or brass.

A close fitting tube shield may be used similarly to the metal sleeve. The shield must fit tightly enough or else it must be supported so that there is no contact with chassis metal.

The tube at the right in Fig. 85-4 has been removed from its socket, fitted into a test adapter, and the adapter put into the socket. For each element of the tube there is a small projecting pin on the side of the adapter, to which the bias isolating capacitor may be clipped. We looked at one of these adapters in another lesson.

A number of receivers are constructed with alignment test jacks or terminals which are connected underneath the chassis to points at which test signals should be applied, and to other points from which the output voltages should be taken. These test terminals are a great convenience for servicing.

Always keep the high-side and ground connections from the signal generator cable as close together as possible when making connections to receiver circuits. This prevents the signal voltages from going through any more of the chassis metal than absolutely necessary.

OVERRIDING OF AGC VOLTAGE. The actual frequency response of a video i-f amplifier or of an i-f amplifier and tuner together might be badly peaked, as at 1 in Fig. 85-5. Yet during alignment with a signal generator and any kind of output indicator the peaks might be absent, to leave an apparent response such as at 2. This will happen when the receiver has an automatic gain control system acting on tubes in the aligned circuits, and when you use a rather strong signal input from the generator. The agc system acts as it should act, and pulls down the peaks of amplification.

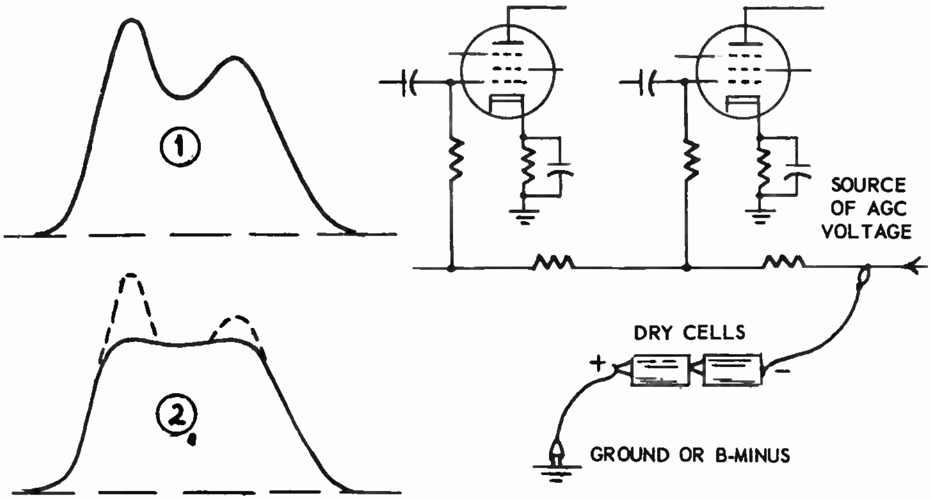


Fig. 85-5. Peaking may be disguised by action of the automatic gain control (left). Connections for overriding the agc voltage (right).

To prevent this action during alignment you may keep the signal generator output so low at all times that it does not cause the agc system to operate. If you are working with an oscilloscope for output indicator, and watching the response curve, the generator output must be kept low enough so that a slight increase or a slight decrease of output will not affect the form of the curve. These output changes will make the entire curve slightly higher or lower, but there must be no change in shape - no peaks should be added and none should disappear.

Without an oscilloscope it is difficult to determine whether or not the agc system is acting on the response, and no matter what kind of output indicator is used it is rather difficult to work with a very weak output from the signal generator. To eliminate these troubles, it is much easier to replace the variable agc bias voltage with a fixed bias voltage while aligning the receiver. This is called overriding the automatic gain control.

8 The method most often used for overriding the agc is illustrated at the right in Fig. 85-5. One or more dry cells are connected between the agc bus and ground or B-minus, with the negative side of the cell or cells clipped to the agc bus and the positive side to ground or B-minus. A 3-volt override, with two cells in series, usually is about right. In localities where television signals are weak, and agc voltage seldom rises very high, a single dry cell may be used. Then the signal generator output must be kept weak, to match the television signal strength. Where received signals are very strong you may use three dry cells, for a 4½-volt override.

If the amplifier tubes have cathode bias, as in Fig. 85-5, or have a fixed minimum bias, it may be sufficient to temporarily connect the agc bus to ground or B-minus and rely on the minimum bias while keeping the signal generator output very weak. If there is a separate agc rectifier or amplifier tube, none of whose sections perform any other function, it usually is advisable to remove this tube while substituting the battery bias during alignment.

(a) *PICTURE TUBE REMOVAL.* There are a number of receivers with which the picture tube must be disconnected from the chassis to allow removal of the chassis from its cabinet or to allow getting at the alignment adjustments. Since the VTVM, or an oscilloscope, will be used for alignment indications there is no particular need for having the picture tube in operation. With the picture tube removed you must carefully insulate the end of the high-voltage cable which normally connects to the anode, or else cut off the high voltage. The high voltage may be cut off in most sets by removing the high-voltage rectifier tube, or, with an r-f type of high-voltage power supply, by removing the oscillator tube. If the receiver is a transformerless type, removal of any tube will cut off the heater voltage to other tubes. With the picture tube disconnected, the two heater recesses in the socket for this tube may be connected together through a 10- or 12-ohm resistor of not less than 5-watt power rating.

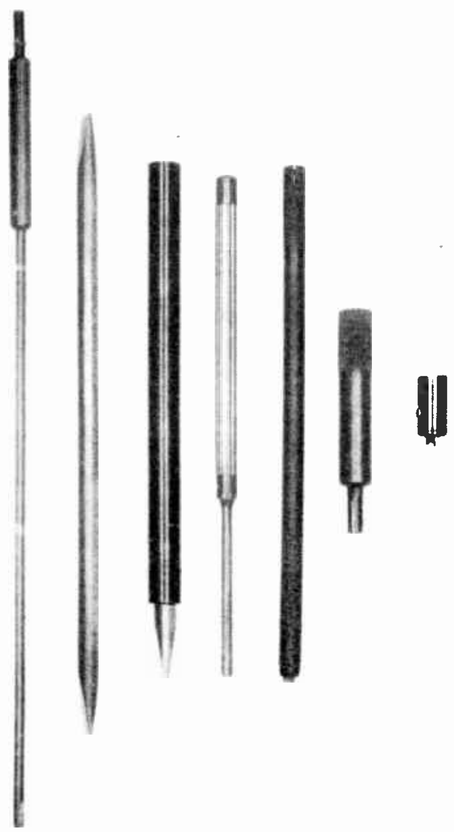


Fig. 85-6. Various types of screw drivers which may be used for alignment adjustments.

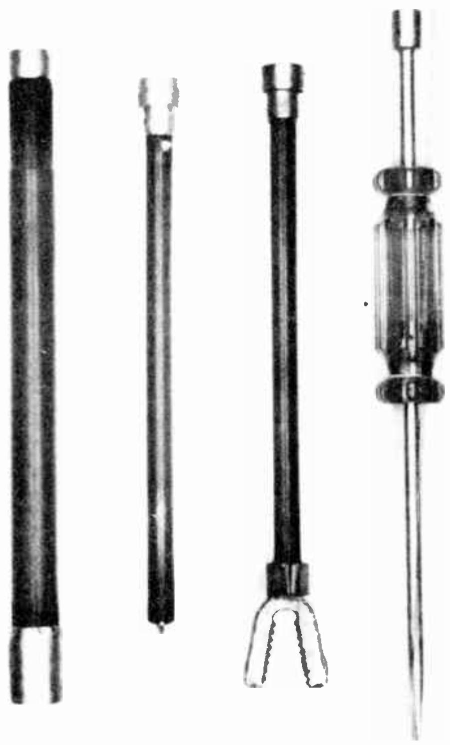


Fig. 85-7. Socket and alligator wrenches designed especially for alignment work.

If the speaker has a field coil connected into the low-voltage power supply filter, disconnecting the speaker will cut off all B voltage to amplifiers and other tubes. In this case it usually is possible to extend and reconnect the speaker leads. Otherwise you may temporarily substitute a fixed resistor of the same resistance as the d-c resistance of the speaker field coil. The substitute resistor should have a power rating of not less than 5 watts, and 10 watts will be better.

ALIGNMENT TOOLS. Movable cores in tuning inductors and trimmer capacitors used for tuning adjustments usually are provided with screw driver slots, although sometimes they may have threaded studs fitted with small hexagon heads which are turned when making adjustments. The screwdrivers used for slot adjustments or the socket wrenches used for hex head adjustments in the tuner section should have no metal of any kind in their construction. For adjustments in the i-f, sync, and sweep sections the screw drivers or wrenches may have metal tips, but the shafts and handles should be of insulating material. Small metal-bladed screw drivers often are used for adjustments in these latter sections of the receiver, but it is not good practice.

There is available from radio and television supply houses an exceedingly great variety of special alignment tools. A few of the screw drivers are illustrated by Fig. 85-6. The three at the left have no metal in their construction. They are made entirely of hard fibre or plastic materials. The four screw drivers at the right have small metal tips set into shafts of fibre or plastic material, with the exception of the center tool, which has a metal tip only at the top. The very long screw driver at the left is useful for reaching such things as r-f oscillator slugs in tuners which are set well back into the chassis. The short screw drivers at the right may be needed when the adjusters are in cramped spaces.

At the left in Fig. 85-7 is a tool having a socket wrench of the "small hex" size at the top and one of the "large hex" size at the bottom. The next tool has a small hex socket at the top and a screw driver tip at the bottom. Next is another tool with a small hex socket at the top, but this one has an alligator wrench at the bottom. The alligator will catch onto nuts or screw heads of various diameters. The tool at the right has a recessed screw driver at the top and an ordinary open tip at the bottom. There is a recessed screw driver tip also at the upper end of the tool third from the right in Fig. 85-6. The recessed tips do not readily slip off the slotted ends of adjustment studs which are in difficult places.

At the left in Fig. 85-8 is a prod having a point at the top and a metal hood at the bottom. Such tools are used for pushing wires out of the way or for catching wires that have to be lifted out of the way.

The two tools at the right in Fig. 85-8 are "tuning wands". A tuning wand consists of a piece of insulation holding a small piece of brass or bronze in one end and a small cylinder of powdered iron in the other end. Bringing the iron end of the wand near either end of the coil or inductor in a resonant circuit that is in operation adds to the inductance, therefore lowers the resonant frequency. Bringing the other, non-magnetic, end of the wand toward the inductor decreases the effective inductance and raises the resonant frequency.

5 The tuning wand is used to determine whether an adjustment should be altered to increase or decrease the inductance in an active tuned circuit before the adjustment actually is altered. If bringing the powdered iron end of the wand near the inductor gives more nearly the results you are after, the circuit needs more inductance and should be aligned accordingly. If the non-magnetic end improves the performance it means that the circuit has too much inductance, and the adjustment should be altered accordingly. If both ends make matters worse, the alignment already is correct.

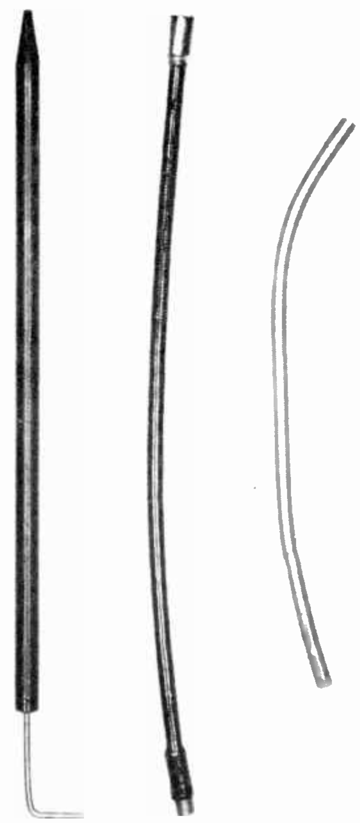


Fig. 85-8. A handy type of prod (left) and two tuning wands (right).

So far as change of resonant frequency is concerned, more inductance acts just like more capacitance, while less inductance acts the same as less capacitance. Then we may say that an improvement in performance with the iron end of the wand brought toward the coil means that the alignment adjustment should be changed to provide either more inductance or more capacitance. The appropriate alignment adjustment may be made with a movable core for inductance or a trimmer capacitor for capacitance, depending on how the circuit parts are constructed. An improvement in performance with the brass end of the wand brought toward the coil indicates the need for less inductance, by turning a slug farther out of its winding, or the need for less capacitance, by opening a trimmer capacitor.

Some wands have rigid shafts. The two illustrated have flexible shafts. You can make your own wand with a length of spaghetti for a flexible shaft, with a brass screw turned securely into one end, and with a small movable core from a discarded tuning inductor screwed into the other end. Insulate the exposed metals with two or three wrappings of Scotch tape. A ready-made wand sells for less than a dollar. You can make your own non-metallic screw drivers by filing to shape a fibre rod or a piece of some hard insulating material, like Bakelite. A small metal tip may be inserted by making a hack saw cut and wedging into it any small piece of steel.

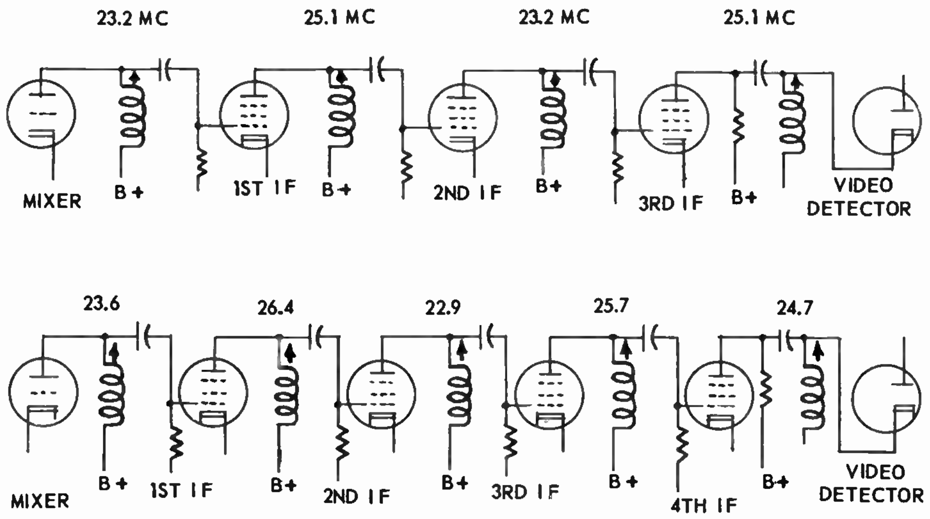


Fig. 85-9. Two orders of resonant frequencies used in interstage couplers of i-f amplifiers.

ALIGNING THE I-F AMPLIFIER. Now we may proceed to align the i-f amplifier system of a television receiver by using an r-f signal generator as the source of intermediate frequencies and a vacuum tube voltmeter as the output indicator. The frequencies used will vary with the make and model of receiver handled, but the general procedure is the same whether we have a stagger tuned amplifier, as represented at the top of Fig. 85-9, or one with all couplers tuned to different frequencies, as in the lower diagram. The general procedure is the same also whether there is intercarrier sound or dual sound.

The preliminary steps are as follows.

1. Override the agc voltage, if the receiver employs an agc system. Methods of doing this have already been explained.

2. It is a good idea, although not absolutely necessary, to disable the r-f oscillator during i-f alignment. The oscillator is not needed, since all the intermediate frequencies to be used will be furnished by the signal generator, not by beating of frequencies from the r-f oscillator and r-f amplifier. The presence of oscillator frequencies can only confuse the results.

If the r-f oscillator is a separate tube, and all heaters are in parallel, simply remove this tube from its socket. If the oscillator is combined with the mixer in a single tube, or if the receiver has series heaters, the oscillator cannot be removed. Then it may be disabled by connecting a capacitor of about 0.001 mf from the oscillator grid or plate to chassis ground or to B-minus.

3. Disconnect the antenna. This is expecially necessary if you do not disable the r-f oscillator.

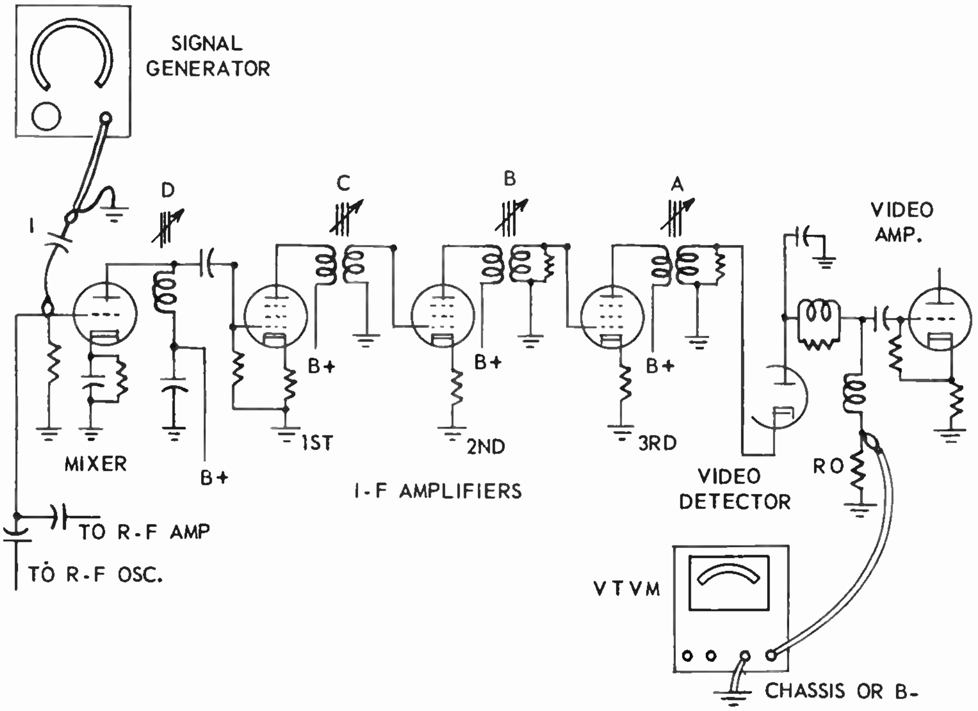


Fig. 85-10. Connections of the signal generator and the vacuum tube voltmeter for i-f amplifier alignment.

4. Set the channel selector to the highest channel that is not assigned to any station within your receiving range.

5. Set the contrast control at about mid-position. This control must not be set so high that pictures or patterns would be distorted during normal operation. A low setting is safer than a high one.

6. Turn the brightness control to minimum, or nearly there. This prevents electron beams so strong as might damage the picture tube screen.

Instrument connections are shown by Fig. 85-10. The high-side, or signal lead, in the cable from the signal generator is connected to the mixer grid through a bias isolating capacitor I, or without the capacitor to a ring slipped over the outside of the mixer tube. The ground lead of the generator cable is connected to chassis metal or to a B-minus point as close as convenient to the mixer grid connection. It may be more convenient to connect the signal generator to the antenna terminals of the receiver, using a fixed resistor of 100 to 150 ohms in series with each cable lead. As a rule enough signal strength at the intermediate frequencies will get through the r-f amplifier and mixer stages to serve our purposes.

If the signal generator allows audio modulation of its r-f output, turn off the modulation and use only unmodulated signal voltages. After connecting the generator, turn its attenuator or output control to minimum until we are ready to proceed.

The VTVM is connected across the video detector load resistor. Adjust or use the VTVM as for measuring d-c voltages, not for a-c voltages. Connect the d-c voltage lead, or the high-side lead, to the top of the detector load resistor directly, not with a series capacitor. Connect the ground lead of the VTVM to chassis ground, to a B-minus point, or directly to the low side of the load resistor, according to how the receiver is wired. If the low side of the detector load resistor is grounded, connect the VTVM ground lead to chassis ground. If the receiver is a transformerless type, connect the VTVM ground clip to a B-minus point. If the low side of the load resistor goes to some negative biasing voltage for the video amplifier, connect the ground side of the VTVM to the low side of the load resistor, and make sure that the housing of the VTVM is not grounded to your instrument shelf.

The VTVM function selector will later be set at positive or at negative d-c volts, whichever gives an up-scale reading. Always use the lowest voltage range or scale. As the work proceeds, the output of the signal generator must be reduced to keep the meter readings within this range. Never use a higher voltage range, for if this is needed, you are feeding in such a strong signal as to cause distortion in the amplifier.

In case the detector load resistor is connected to a biasing voltage for the following video amplifier, and you cannot conveniently connect the VTVM ground to the low end of the resistor, the meter will read up scale even with no signal input. Use the zero adjustment control of the VTVM to bring the reading down to zero, or to any easily noted scale division which will become the reference zero while making alignment adjustments.

With all the connections completed, turn on the receiver and the signal generator. Let them warm up for at least 15 minutes, and preferably longer, before commencing to make adjustments. This warmup period is absolutely necessary to allow stabilization of the generator frequency and the receiver resonant frequencies.

After the warmup period adjust the signal generator for the precise frequency at which the i-f coupler just ahead of the video detector is to be aligned. This would be coupler A of Fig. 85-10. Adjust the generator output or attenuator until there is any low reading on the VTVM. Now carefully adjust the movable core or trimmer capacitor of coupler A to get the highest possible reading on the meter. Turn the coupler adjustment back and forth through the position for maximum reading, to make certain that your final setting is the best possible.

If there is any other coupler to align for the same frequency, do not alter the signal generator frequency, but proceed to adjust or align this other coupler in the same manner, to get the maximum voltage reading on the VTVM. Reduce the output of the generator as required to keep the voltage readings well below the top of the meter scale.

Quite likely you now will have aligned couplers A and C, because in a three-stage stagger tuned i-f amplifier the alternate couplers usually are operated at the same frequency. If this is the case, the next step is to adjust the signal generator frequency precisely for the frequency at which coupler B is to operate, then align this coupler for maximum reading on the VTVM. The voltage indication will usually drop when you change generator frequency, but will come up again as you align coupler B. If there is another coupler to be aligned at the same frequency as B, do not alter the generator frequency but go ahead and align this other coupler for maximum VTVM reading.

After all the couplers have been aligned for their operating frequencies, start all over again with coupler A and with the generator set for the correct frequency. Go through the whole process the second time. The tuning of every coupler has some effect on all the other couplers, for they are coupled through the tubes, and every coupler has some effect on the overall response of the amplifier system. If you had an oscilloscope available, the next and final step would be to observe the overall response of the i-f amplifier system. We shall come to this later on.

The procedure just explained works out well and saves time when none of the stages are badly out of alignment. If there is a great deal of misalignment you may not be able to get enough signal to the last coupler with the generator coupled to the mixer. Then it is in order to follow the method illustrated by Fig. 85-11.

The VTVM is connected across the video detector load resistor Ro as usual, and remains so connected during the entire process of alignment. The high-side lead from the signal generator is first coupled to the grid of the final i-f amplifier, the one which just precedes the video detector. Then the coupler between this amplifier and the video detector is aligned for maximum reading of the meter at the correct frequency. This would be coupler A in Fig. 85-11.

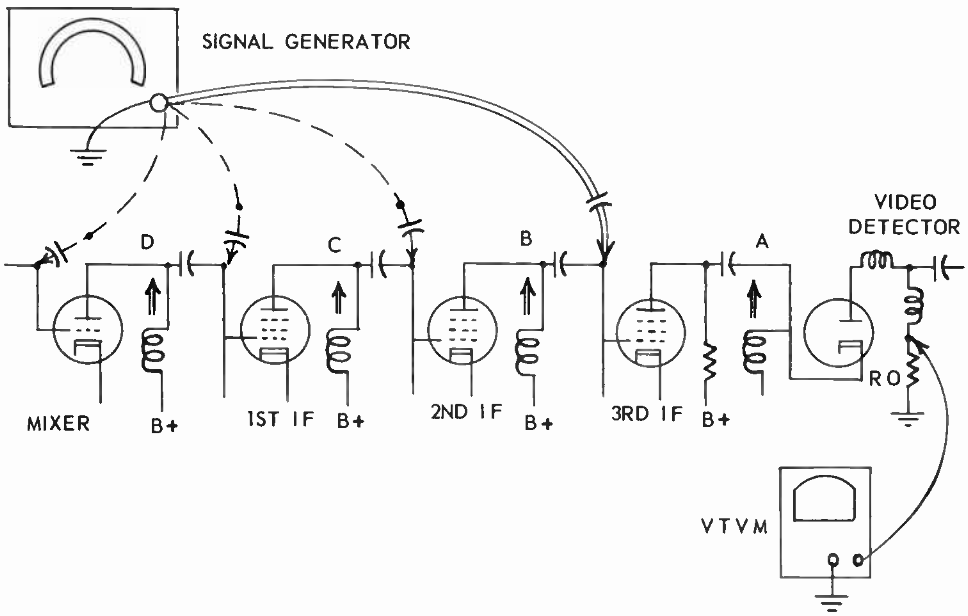


Fig. 85-11. When the amplifier is far out of adjustment, the signal generator connection is moved back stage-by-stage during alignment.

Now the high-side lead from the generator is moved back to the grid of next to the last i-f amplifier, as shown by one of the broken-line connections. Connections to the VTVM remain unchanged. Coupler B, between the last two i-f amplifiers, is aligned at its correct frequency for maximum meter reading. Then the generator input is moved back another stage, to the grid of the first i-f amplifier tube, while coupler C is aligned to the correct frequency and finally the generator connection is made to the grid of the mixer tube while aligning coupler D or the coupler between mixer and first i-f amplifier.

With this method of bringing in one more i-f stage for each succeeding step in alignment, you have to change the signal generator frequency at each step. This frequency always is adjusted for the first coupler following the generator connection, or the first coupler between the generator and the video detector. When using this method for coupler frequencies as shown at the bottom of Fig. 85-9 the generator would be tuned successively to frequencies of 24.7 mc, 25.7 mc, 22.9 mc, 26.4 mc, and 23.6 mc.

TRAP ALIGNMENT. In many of the receivers which employ a dual sound system, and in some others, there are various trap circuits on the interstage couplers of the video i-f amplifier. One possible arrangement is illustrated by Fig. 85-12. The upward pointing arrows indicate the coupler adjustments used for i-f alignment by methods already explained. The downward pointing arrows indicate trap adjustments for adjacent sound, adjacent video, and accompanying sound.

Note that the sound takeoff circuit is here considered as the equivalent of a trap for accompanying sound, since it takes out of the i-f amplifier input much of the energy at the sound intermediate frequency or at the accompanying sound frequency. This is the center frequency at which the f-m sound system operates in a receiver having dual or split sound.

Trap circuits usually are of high-Q design and construction to attenuate a narrow band of frequencies, possibly only 200 to 300 kilocycles. The attenuation may be 15 to 20 db, to reduce the trapped signal frequency to something like 1/30 to 1/100 of its untrapped strength. The range of frequencies through which an i-f trap may be tuned seldom is more than 2 to 3 mc, so each trap has to be designed for the particular frequency that it is to attenuate.

It is common practice to adjust or align all the traps before any of the interstage coupler alignment. Some technicians prefer adjusting each trap at the same time as the coupler with which it operates. Then, in Fig. 85-12, the accompanying sound trap on the cathode of the fourth i-f amplifier would be adjusted right after the coupler that immediately precedes the video detector. The next alignment would be of the coupler between the third and fourth amplifiers, where there is no trap. Then the coupler between second and third amplifiers would be aligned, and immediately afterward the adjacent video trap would be adjusted. So the work would proceed until aligning the coupler between mixer and first i-f amplifier, and the sound takeoff that is on this coupler.

Regardless of trap construction or purpose, the adjustment always is made to obtain maximum reduction or attenuation of the frequency to be trapped out as indicated by the meter. All the preliminary steps for trap adjustment are the same as for coupler alignment. The antenna is disconnected, the channel selector is set at the highest unassigned channel, the contrast is set at its normal position, and so on. The VTVM is connected across the video detector load resistor and the signal generator is coupled to the mixer grid of other tubes as in Figs. 85-10 and 85-11.

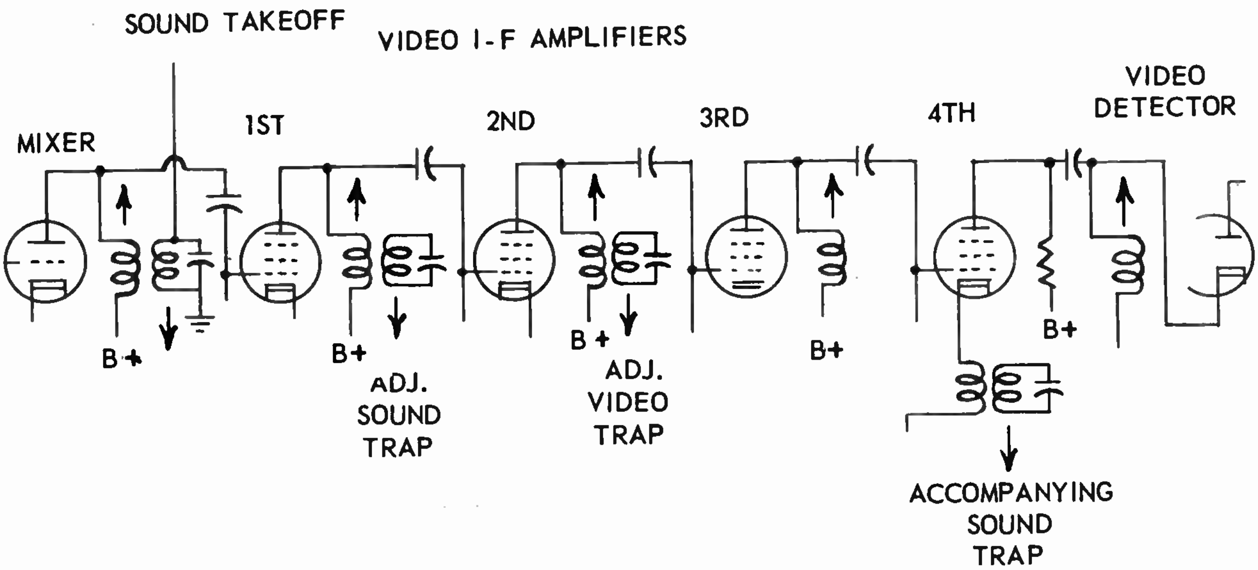


Fig. 85-12. Typical locations of traps in a video i-f amplifier system.

Tune the signal generator precisely to the frequency at which each trap is to be aligned, then adjust that trap for minimum reading on the VTVM. As the meter reading drops lower during the adjustment process, increase the output from the signal generator enough to get good distinct indications of minimum voltage. If two or more traps are to be set at the same frequency, first adjust the one farthest from the generator, or nearest the video detector. To have distinct voltage readings for setting this trap it usually is necessary to temporarily detune the other trap which is to work at the same frequency.

After adjusting all the traps, tune the signal generator very slowly through the entire range of frequencies in which the traps operate. Watch the VTVM reading, and at each low voltage note the frequency for which the generator then is set. This should be one of the trap frequencies.

Again check all the trap frequencies after the interstage couplers have been aligned. Every change of coupler resonant frequency affects the resonant frequency of the associated trap, and every change of trap frequency affects the resonant frequency of the coupler on which this trap acts.

I-F AMPLIFIER REGENERATION. During alignment of some receivers you will run into the problem of regenerative feedbacks that cause excessive peaking, or possible self-oscillation in some of the amplifier circuits. Excessive peaking will cause a picture or pattern to become covered with short black streaks about one picture line high and an inch or less long, horizontally across the screen.

Self-oscillation will cause the pointer of the VTVM to go violently off scale at the high end. The raster area of the picture tube will become brilliantly white. The oscillation cannot be stopped by reducing the signal generator output to zero. You must instantly turn off the power switch or pull the power cord plug to save the picture tube.

The most common cause for oscillation is getting two adjacent interstage couplers tuned to the same or nearly the same frequency while aligning them. Other causes are as follows;

The contrast control set too high.

The signal voltage from the generator is too strong. If the generator output control or attenuator won't bring the signal low enough, try using a smaller series capacitor between the generator high-side lead and the grid of the tube to which it connects. You may go down to 5 or 10 mmf.

Grounding and bonding of test instruments and receiver is insufficient, or the bonding cables may have to be connected to different places.

Signal generator leads are of excessive length, or the high-side lead is not shielded, or its shield is not connected to grounds at both ends.

VTVM leads are too long, are placed too close to receiver circuits, are not shielded, or the cable shield is not grounded at both ends. Try connecting a fixed capacitor of 0.001 to 0.005 mfd across the lead clips at the detector load resistor, or across this resistor, while proceeding with alignment. All this applies also when using an oscilloscope as output indicator.

Receiver wiring for grids, plates, or screens has been shifted to allow regenerative feedbacks, or such parts as coupling capacitors and peaking coils have been similarly dressed wrong.

A tuned cathode trap which is resonant in the i-f range may be the cause. If the trap is not in the cathode bias circuit, furnishing the resistance, short circuit the trap with a piece of wire. Otherwise short it with a fixed capacitor of about 0.001 mf while making the coupler alignment.

Decoupling or bypass capacitors in the i-f amplifier circuits may be defective.

Heater decoupling chokes have not been used where they are needed.

There are various temporary preventatives which may be applied to an i-f amplifier that oscillates. First, try overriding the agc with a stronger negative voltage, going to as much as 4½ or 6 volts (3 or 4 dry cells) while getting a preliminary alignment. Then go back to the lesser bias while completing the job.

Use the alignment method of Fig. 85-11 and detune all the interstage couplers between the signal generator and the coupler being aligned. Realign these other couplers as you come to them. Instead of detuning the couplers by turning their adjustable cores or trimmer capacitors, you may connect all the grids, ahead of the one being aligned, to ground through capacitors of 0.001 mf or through resistors of about 300 ohms each. Remove these detuning devices as you come to the couplers.

If the receiver persists in oscillating, even when correctly aligned, it would be possible to make certain alterations in the design provided you feel competent to undertake such work without too much danger of ruining the whole performance. However, oscillation is usually caused by faulty component parts, and should be located. It is necessary to locate the stage or stages in which the trouble originates. Try detuning each coupler by a small amount, then return its adjustment to the original setting. If one coupler is more effective than others in preventing oscillation, work on that coupler or on the grid circuit of the following tube. Another way is to temporarily connect an additional fixed resistor of 10,000 to 20,000 ohms in series with one i-f amplifier plate circuit after another, or 30,000 to 50,000 ohms in series with one screen circuit after another. If doing this to one tube will reduce the oscillation tendency, work on that tube or on the tubes input or output couplers. Some of the remedies which may be tried are as follows.

As at a and a in Fig. 85-13 temporarily connect a fixed resistor of 4,000 to 40,000 ohms across the coupler winding. The less the resistance in ohms, the more effectively oscillation will be suppressed, but the lower will be the gain of the stage. If the coupler has two windings, connect the resistor first on one side and then on the other, to determine which way prevents oscillation with least loss of gain. As at b a small size fixed resistor of 50 to 100 ohms may be connected in series with the grid, right at the grid prong of the socket. If there is a bypassed cathode resistor, try using a bypass of less capacitance or remove the bypass altogether to obtain a degenerative feedback when plate current becomes excessive at the approach of oscillation.

If the grid return is through a resistor, as at d, substitutes less resistance at this point. Less resistance will help suppress oscillation, but also will reduce the gain. It is possible, but not probable, that adding a heater decoupling choke as at e or both a choke and a capacitor to ground will help matters.

A cathode trap, as at f, often will prevent oscillation with little loss of gain, but it is rather difficult to install. The trap must be tuned to resonance at the frequency of oscillation. This requires measuring the resonant frequency by methods explained in other lessons. You may use a mica dielectric screw-adjusted trimmer capacitor for tuning, or the coil may be made space wound of number 16 or 18 enameled copper wire and either compressed or stretched to alter the inductance for tuning.

It is possible, but again not too probable, that placing a close fitting shield over any one or more of the i-f amplifier tubes will prevent oscillation. The shield must make good contact with chassis metal to provide effective grounding.

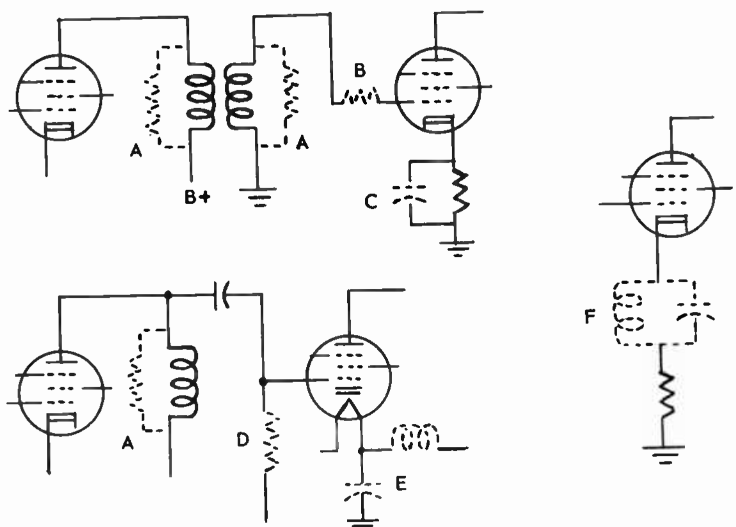
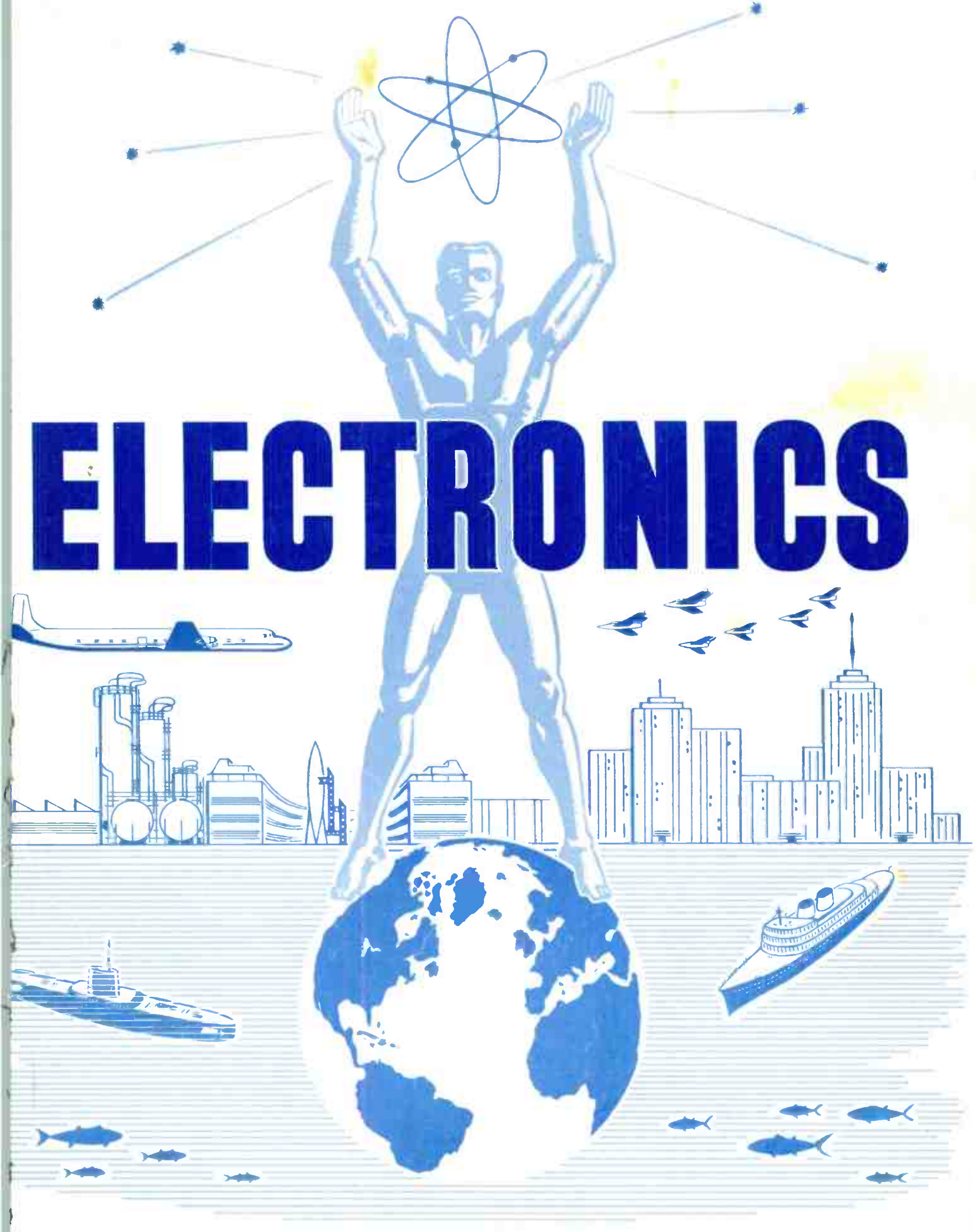


Fig. 85-13. Remedies for regenerative feedbacks and for oscillation in the i-f amplifier.

86-90

ELECTRONICS



COMMERCIAL TRADES INSTITUTE • CHICAGO 26, ILLINOIS

TELEVISION

LESSON NO. 86

SWEEP AND MARKER GENERATORS

"Training Your Mind"

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

COMMERCIAL TRADES INSTITUTE

Training Your Mind

C.T.I.

Perfecting Your Skill

Chicago, Ill.

LESSON NO. 86
SWEEP AND MARKER GENERATORS

We know that the ideal form of frequency response in a video i-f amplifier, as nearly as such a response can be obtained in practice, is about as shown at 1 in Fig. 86-1. During alignment of interstage couplers and traps it would be a great help to see the actual response curve, and watch the changes caused by altering various adjustments. An actual curve could be plotted by using the signal generator and vacuum tube voltmeter. You could tune the generator to many frequencies from somewhat below the sound intermediate to well above the video intermediate frequency, say at intervals of a quarter megacycle. The VTVM would indicate the voltage across the video detector load resistor at each of these frequencies. After marking all the corresponding voltages and frequencies on graph paper a curve could be drawn through these markings, which might appear as at 2.

This would be a slow process. After making one curve it would be necessary to plot many more in order to show the effects of alignment adjustments. Instead, we may use a sweep generator to continually vary the frequencies throughout any range required. Then the oscilloscope, connected across the video detector load resistor, will display the frequency response curve and show instantly the effect of every adjustment you make while bringing the curve to the desired shape. By using also a marker generator in connection with the sweep generator and oscilloscope it becomes possible to identify the frequency at any and every point along the curve being traced on the screen of the oscilloscope.

Before describing the steps in this method of alignment we must become acquainted with operation of the sweep generator and with the requirements for a marker generator. A sweep generator is similar to an r-f signal generator, but there are added circuits or parts which cause the frequency of the output voltage to continually and smoothly change back and forth through whatever range you select.

For instance, when examining a frequency response such as shown by Fig. 86-1, you could adjust the sweep generator to continually vary its output frequency from about 20 mc up to 28 mc, then back down to 20 mc, again up to 28 mc, and have this sweep continue indefinitely. The change from lowest to highest frequency and back to lowest occurs 60 times per second in practically all sweep generators. The sweep rate or sweeping frequency is 60 cycles per second. There never is need for varying this sweep rate, and it is not adjustable.

Output frequency from the sweep generator varies as shown at 1 in Fig. 86-2, where it is assumed that the sweep range is from 20 to 28 mc. During one period of 1/120 second (which is half of the 60-cycle period), the frequency increases from 20 mc to 28 mc. During the following 1/120 second (the second half of the 60-cycle period), the frequency decreases from 28 mc to 20 mc. This action continues as long as the sweep generator remains turned on.

COMMERCIAL TRADES INSTITUTE

TELEVISION

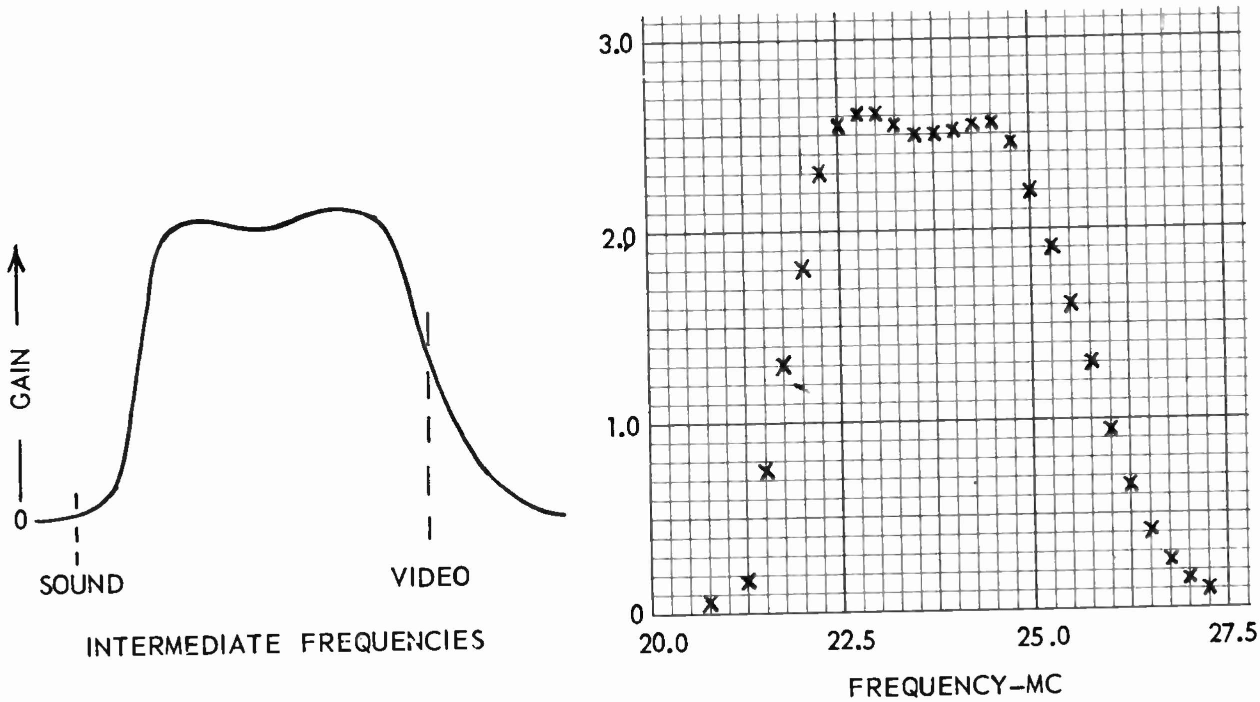


Fig. 86-1. A frequency response curve and how it might be plotted by means of a signal generator and vacuum tube voltmeter.

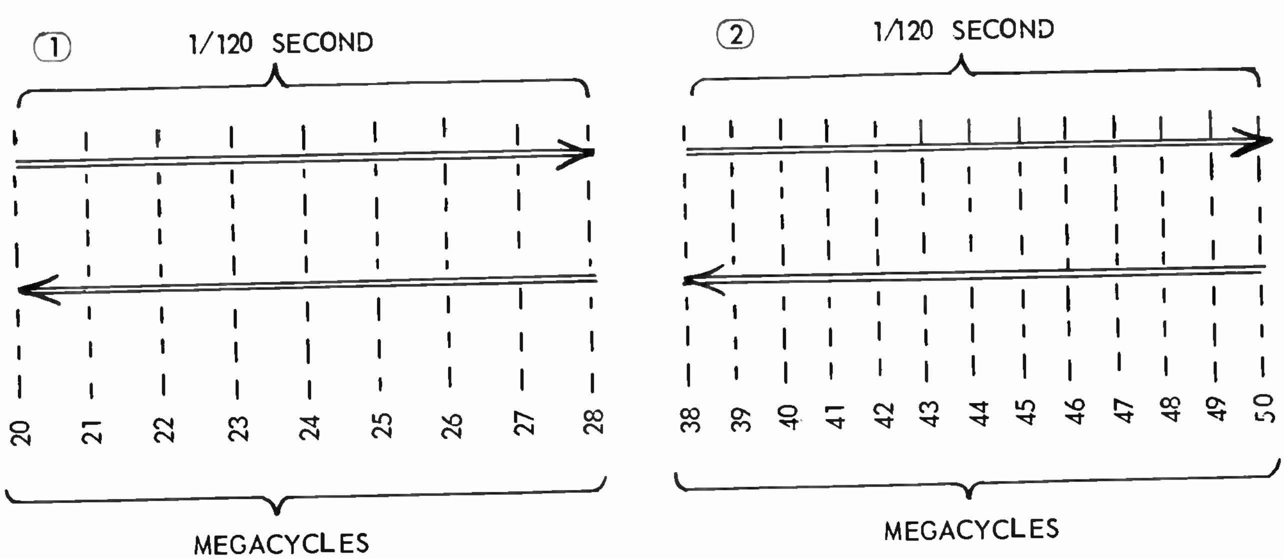


Fig. 86-2. The swept frequency varies from minimum to maximum and back again in two periods of 1/120 second each, or at a 60 cycle rate.

When output voltage from the sweep generator is coupled to the grid of the mixer or to the grid of any other tube, just as an ordinary signal generator would be coupled, the i-f amplifier is fed the continually varying frequencies. Amplification or gain at each instantaneous frequency depends on the tuning or alignment of the interstage couplers. Then the output of the i-f amplifier, as taken from the video detector load resistor, varies with the input frequency.

The sweep generator has adjustments that allow its output frequency to sweep back and forth through whatever range may be required. Were the adjustments set for sweeping back and forth between 38 and 50 mc the action would be as shown at 2 in Fig. 86-2. The sweep rate remains 60 cycles per second no matter what the range of swept frequencies.

The oscilloscope vertical input is connected across the video detector load resistor, and thus to the output of the i-f amplifier. This causes the electron beam in the CRT to be deflected vertically to a distance proportional to voltage gain of the i-f amplifier system at whatever frequency is coming from the sweep generator at a given instant. To observe the response it is necessary to deflect the beam in the CRT across the screen from left to right during a period of 1/120 second, and this must be exactly the same 1/120 second in which the sweep generator frequency is changing from 20 to 28 mc. The result, as at 1 in Fig. 86-3, is a trace whose height at various points from left to right across the screen is proportional to i-f amplifier gain at the various frequencies. This trace is a curve showing the frequency response of the i-f amplifier.

During the following 1/120 second, represented at 2 in Fig. 86-3, the oscilloscope beam will be deflected from right to left at the same speed with which it formerly traveled from left to right. During this same 1/120 second the frequency from the sweep generator is decreasing from 28 mc to 20 mc. Consequently, at any given point along either the forward or reverse trace the applied frequency will be the same. Since the frequency is the same, the gain of the amplifier and the height of the traced curve will be the same at every point no matter which direction the beam is moving. Due to persistence of vision and to persistence in the CRT phosphor, the forward and reverse traces appear as a single curve.

SWEEP GENERATOR REQUIREMENTS. The range through which the frequencies from the sweep generator may be varied usually is called the sweep width. We have spoken of one sweep width of 8 mc, from 20 to 28 mc, and of another width of 12 mc, from 38 to 50 mc as represented in Fig. 86-2. In a video i-f amplifier there may be more or less gain from about 1½ mc below the sound intermediate to 3 mc above the video intermediate. Therefore, it is necessary to provide for a minimum sweep width of 8 mc for i-f alignment. Response through the r-f amplifier and mixer is much wider, and for alignment of tuners it is necessary to have sweep width of at least 12 mc, and preferably 15 mc. If the sweep generator is to be used also for alignment of the f-m sound section of television receivers the sweep width should go down to at least 3 mc. To summarize, it is desirable to have the sweep width variable from 3 mc, or less, to about 15 mc.

Sweep width may be thought of as varying from below to above a center frequency. For alignment of video i-f amplifiers the center frequencies should be adjustable from at least 20 mc up to nearly 50 mc in order to accomodate both the present RTMA standards and the older standards. For tuner alignment the center frequency must be variable from below the carriers of channel 2 to above those for channel 13, or from about 50 mc up to around 220 mc. For alignment of intercarrier sound it is necessary to have a center frequency of 4.5 mc.

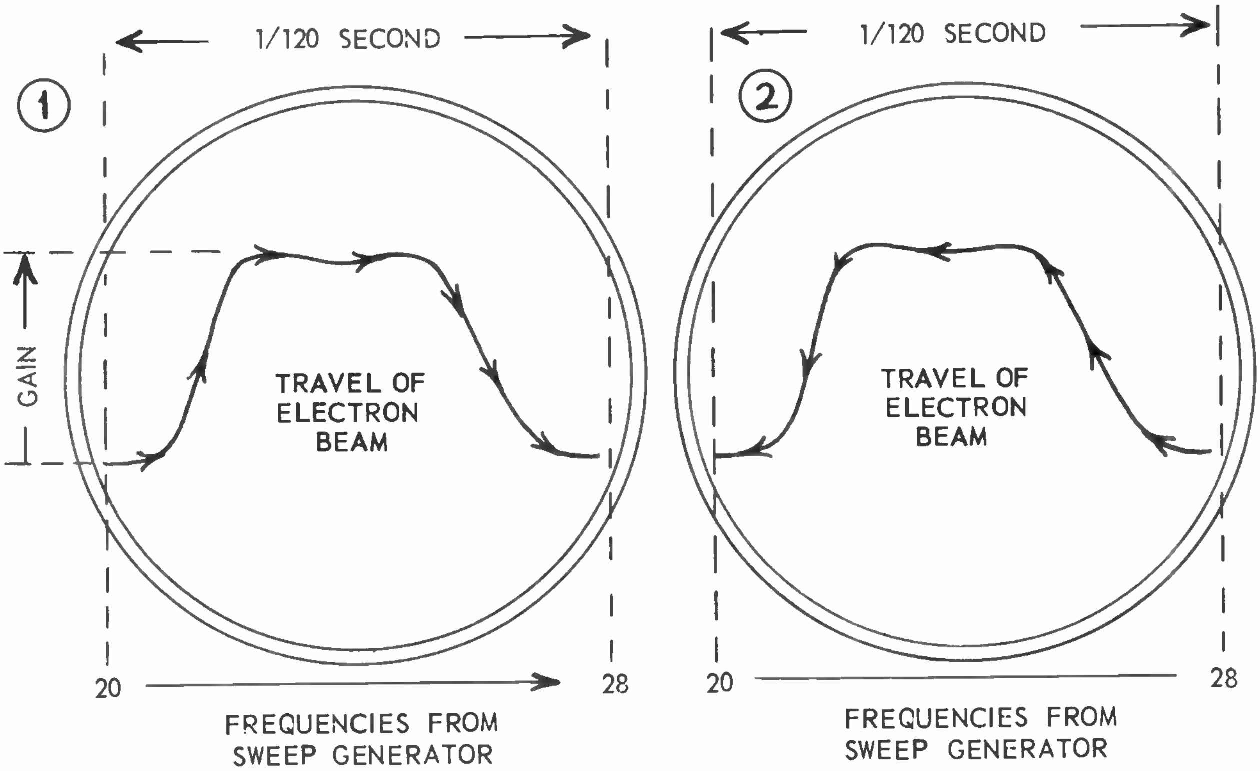


Fig. 86-3. The CRT beam is deflected equally by equal gains as frequency increases and then decreases.

The sweep generator should be capable of providing a maximum output of about 1/10 volt at all frequencies. This high output is needed when working through only a single stage of an i-f amplifier and when working through only the r-f amplifier or the r-f amplifier and mixer for tuner alignment. It is desirable to have an effective attenuator or output control. If, however, the sweep generator is as well shielded as it should be, and is used with a well shielded output cable, attenuation may be had from a small capacitance in the grid bias isolation capacitor used in series with the output. High quality generators usually include automatic line voltage regulation for maintaining a constant output during a series of tests.

Possibly the most important requirement for a sweep generator is an output voltage that is flat, or constant, throughout any sweep width that may be used. If there are even the slightest humps or dips of voltage at any of the swept frequencies these variations will be amplified by the receiver stages and will appear greatly magnified on the trace of the frequency response. If the alignment adjustments are changed to reduce or eliminate these variations coming from the sweep generator, the response will be decidedly incorrect during normal operation of the receiver. It is much better to do your alignment work with a vacuum tube voltmeter and single-frequency signal generator than with a poorly designed or constructed sweep generator and oscilloscope.

As a general rule the voltage output from a sweep generator will be flatter if all the frequencies are fundamentals rather than having some formed by harmonics. At least some sweep generators which employ

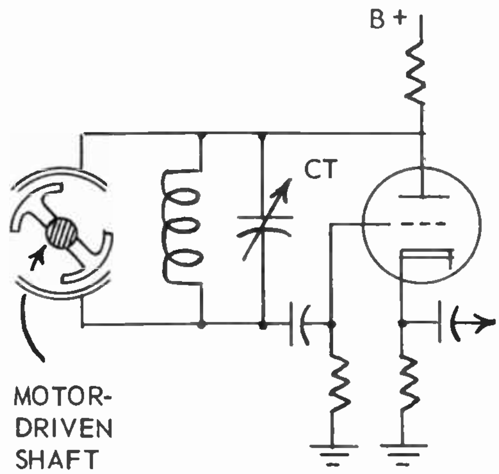


Fig. 86-4. Sweep by means of a revolving split stator capacitor.

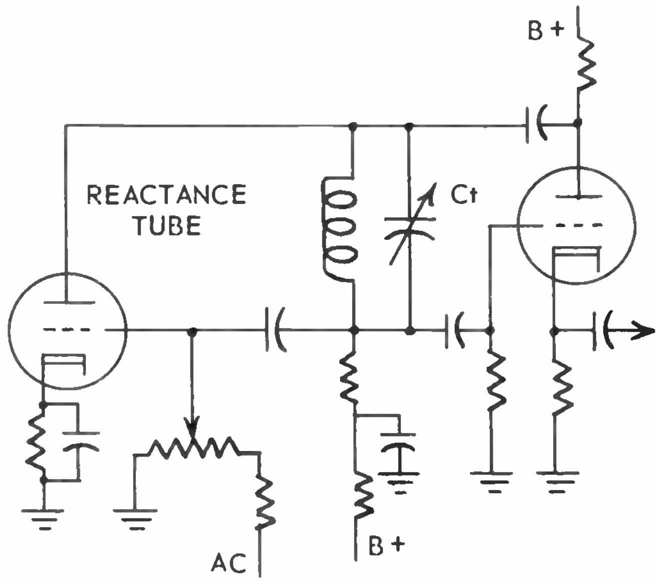


Fig. 86-5. Frequency sweep with a reactance tube.

harmonics furnish outputs that are too irregular within swept range to be useful for alignment work.

SWEEPING METHODS. There are many ways in which the output frequency of an oscillator may be varied at a rate of 60 cycles per second or at any other rate. In some of the earlier designs, whose elementary principle is shown by Fig. 86-4, there is a rotary capacitor in the tuned circuit. The capacitor is a split stator type with two stator sections and between them or interleaving with them a single rotor member which revolves. The rotor is mounted on, or driven by, the shaft of a synchronous motor. This is a type of motor that maintains its speed proportional to power line frequency regardless of ordinary changes in line voltage or in the driven load. Capacitance of the split stator unit goes through its minimum and maximum values twice during each revolution, and thus the frequency of the oscillator is continually varied between the highest and lowest values determined by adjustments of a tuning capacitor or a movable core in the tuning inductor.

A method used in many sweep generators employs a reactance tube, as shown in a general way by Fig. 86-5. The action of this reactance tube is like that of the reactance tube which is a part of some automatic frequency controls for television horizontal sweep oscillators, and which has been described in connection with those oscillators in another lesson. Referring to the present diagram, the grid of the reactance tube is fed with an a-c voltage at the 60-cycle power line frequency. This voltage may be taken from a heater circuit or from the primary or secondary side of the power transformer in the sweep generator.

How much the reactance tube varies the effective inductance of the oscillator tuned circuit depends on the amplitude of a-c voltage applied to the reactance grid. The greater this amplitude the greater is the

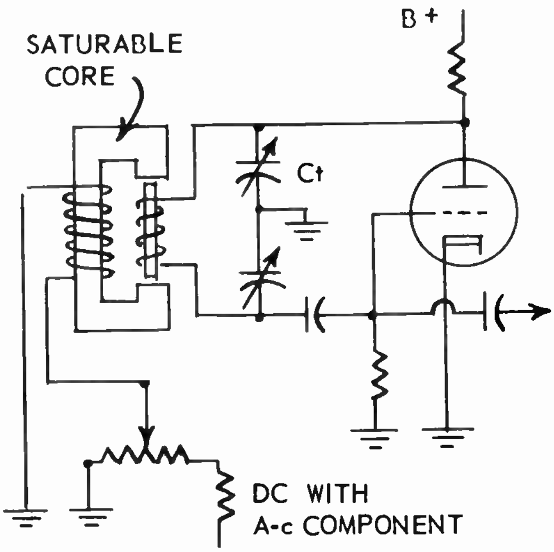


Fig. 86-6. Variable permeability method of frequency sweep.

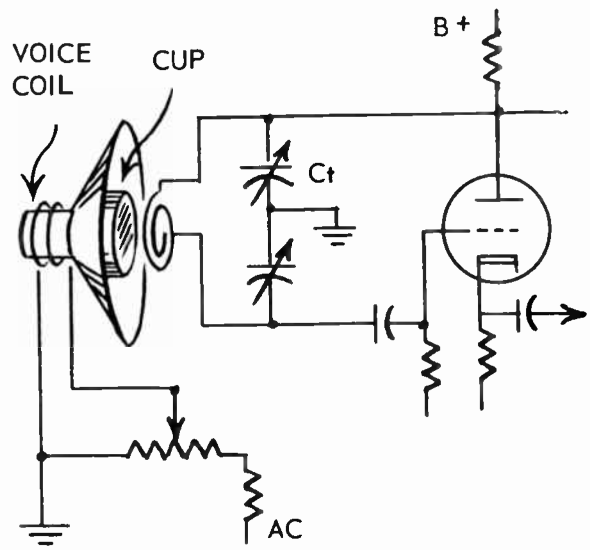


Fig. 86-7. Sweep by means of speaker and vibrating cup or disc.

variation of oscillator frequency, and the wider is the frequency sweep. The width of sweep is adjusted by a rheostat or potentiometer that alters the a-c voltage amplitude on the reactance grid. The oscillator may be tuned to some center frequency by capacitor Ct, whereupon the output frequency will sweep either way from this center value.

A variable permeability sweep is shown in principle by Fig. 86-6. The tuning coil in the oscillator circuit is wound on a small core of powdered iron. This small core forms part of the magnetic circuit of a much larger "saturable core". On the saturable core is a winding through which is passed a direct or unidirectional current having a large alternating component which causes the winding current to become alternately of maximum and minimum values, but to remain always in the same direction or polarity. The varying magnetic flux in the saturable core and the smaller powdered iron core varies the permeability of the smaller core so that it alternately has the properties of an iron core to produce large inductance, and very nearly an air core to produce small inductance.

The continually varying inductance in the core of the oscillator winding varies the oscillator frequency in relation to any center frequency which is determined by adjustment of a tuning capacitor or capacitors. The variation of inductance and of oscillator frequency depends on the strength of the varying direct current in the winding of the saturable core, and this current strength is controlled by a rheostat or potentiometer which becomes the control for sweep width. The a-c component of the varying direct current is taken from some internal source at the 60-cycle power line frequency.

Fig. 86-7 shows the principle of a frequency sweep with which the effective inductance of the oscil-

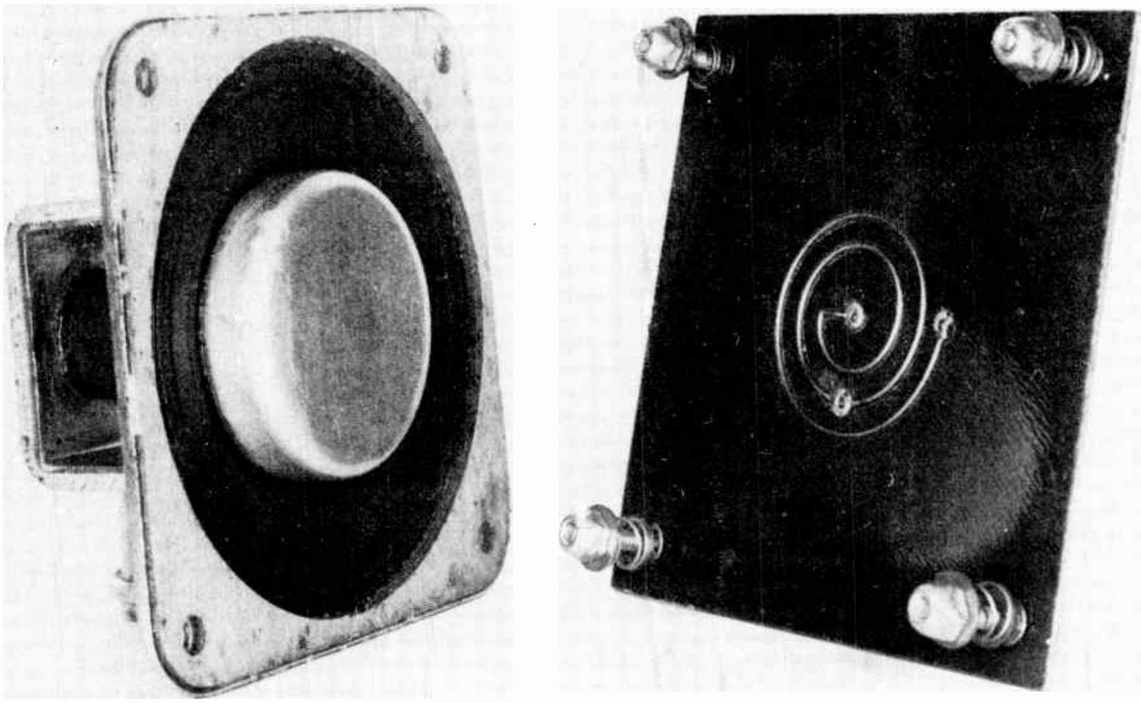


Fig. 86-8. The speaker unit and the oscillator inductor mounted on an insulating plate for frequency sweep.

lator tuning coil by a non-magnetic cup or disc that is alternately moved close to and farther from the turns of the coil. The principal parts used with this system are illustrated by Fig. 86-8.

The oscillator coil or inductor is a flat spiral of wire attached to an insulating plate that is mounted in a fixed position. The non-magnetic cup or disc is attached to the cone and thereby to the voice coil of a permanent-magnet speaker. Current at 60 cycles, taken from a heater circuit or other internal source, is passed through the voice coil. This causes the cone and the attached metal cup or disc to move toward and away from the oscillator inductor 60 times per second. Each time the metal approaches the coil there is a decrease of effective inductance and an increase of oscillator frequency. As the metal recedes there is increase of inductance and a decrease of frequency.

Oscillator frequency is varied with reference to a center frequency which is determined by an adjustable tuning capacitor. The distance through which the speaker cone and the non-magnetic cup vibrate is altered by changing the amplitude or strength of alternating current in the voice coil. This is done by a rheostat or potentiometer, which is the control for sweep width.

To provide a wide choice of center frequencies any type of swept oscillator may be used in connection with a separate beat oscillator and a mixer tube as shown in the simplified circuit of Fig. 86-9. The fre-

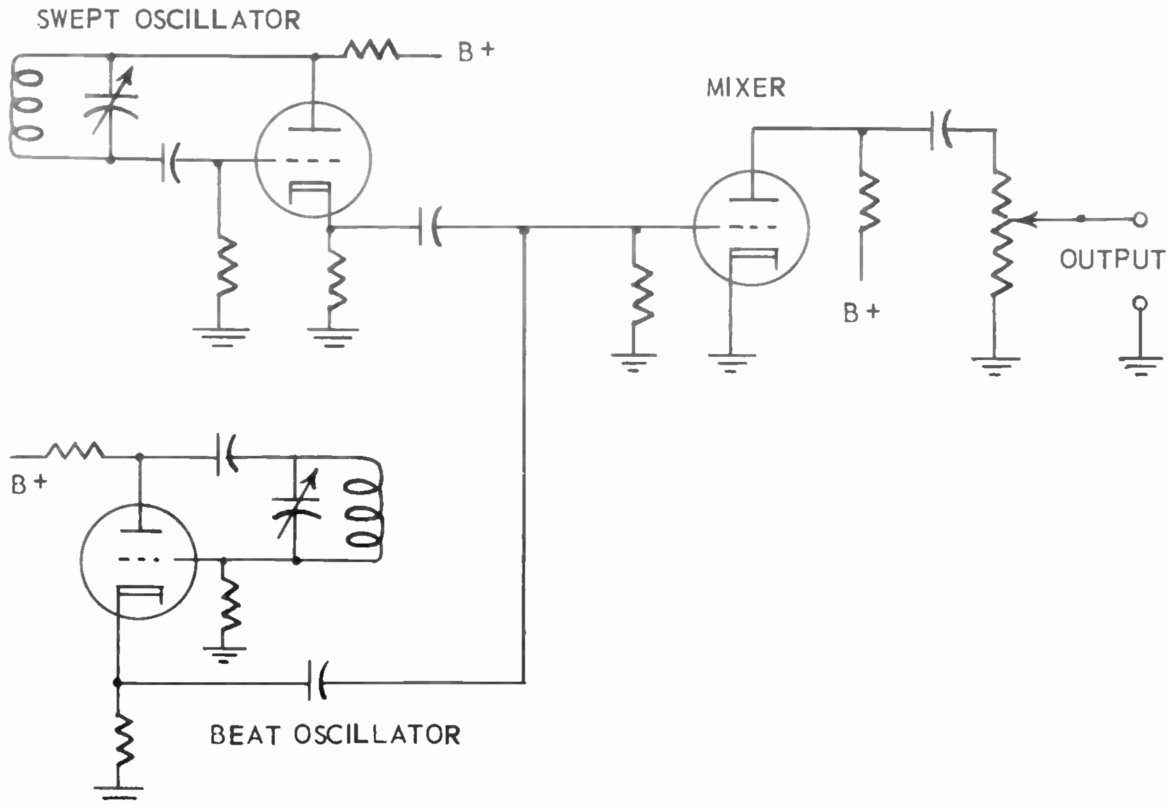


Fig. 86-9. Frequencies from the swept oscillator and from a beat oscillator produce a swept difference frequency in the mixer output.

quencies from the two oscillators are applied to the grid of the mixer tube, just as the two frequencies from an r-f amplifier and r-f oscillator are applied to the mixer grid in any superheterodyne receiver. The mixer output contains the sum and difference beat frequencies. The frequencies of the two oscillators are chosen of such values that sum frequencies from the mixer are higher than any television band, or it is desirable that they be chosen thus. Since the frequency coming from the swept oscillator is varying back and forth through the selected sweep width, the difference beat frequency from the mixer will be sweeping back and forth through this same width.

As an example in forming a number of ranges or bands of output frequencies we shall assume that the swept oscillator operates at a constant center frequency of 120 mc, and that the beat oscillator is provided with a set of tuning inductors that allow tuning through several ranges of frequency, all of them higher than the center frequency of the swept oscillator. The difference frequencies in the mixer output will be the differences between beat oscillator variable frequencies and the center frequency of the swept oscillator.

	Beat Oscillator Frequency Range	Swept Oscillator Center Frequency	Mixer Output (difference)
First i-f range	123 to 160 mc	120 mc	3 to 40 mc
Second i-f range	145 to 180 mc	120 mc	25 to 60 mc
Low-band carriers	172 to 230 mc	120 mc	52 to 110 mc
High-band carriers	255 to 340 mc	120 mc	135 to 220 mc

SYNCHRONIZED SWEEP. As you will have noticed when examining the various methods of sweeping the oscillator frequency it always is done by means of a voltage or current taken at power line frequency from the power transformer of the generator. Therefore, the sweeping frequency and the rate of change of this frequency will follow the power line frequency, and also the waveform obtained from the power transformer. If part of the same voltage or current that controls the frequency sweep is taken from the sweep generator and applied to the horizontal input of the oscilloscope, with the oscilloscope panel control set to use this input, the electron beam in the CRT will move at the same frequency and in accordance with the same waveform as the variations of frequency from the oscillator.

All this may be done and still the forward and return traces may appear separately, as in Fig. 86-10, instead of superimposed or as a single trace. There will be such offsetting of the forward and return traces

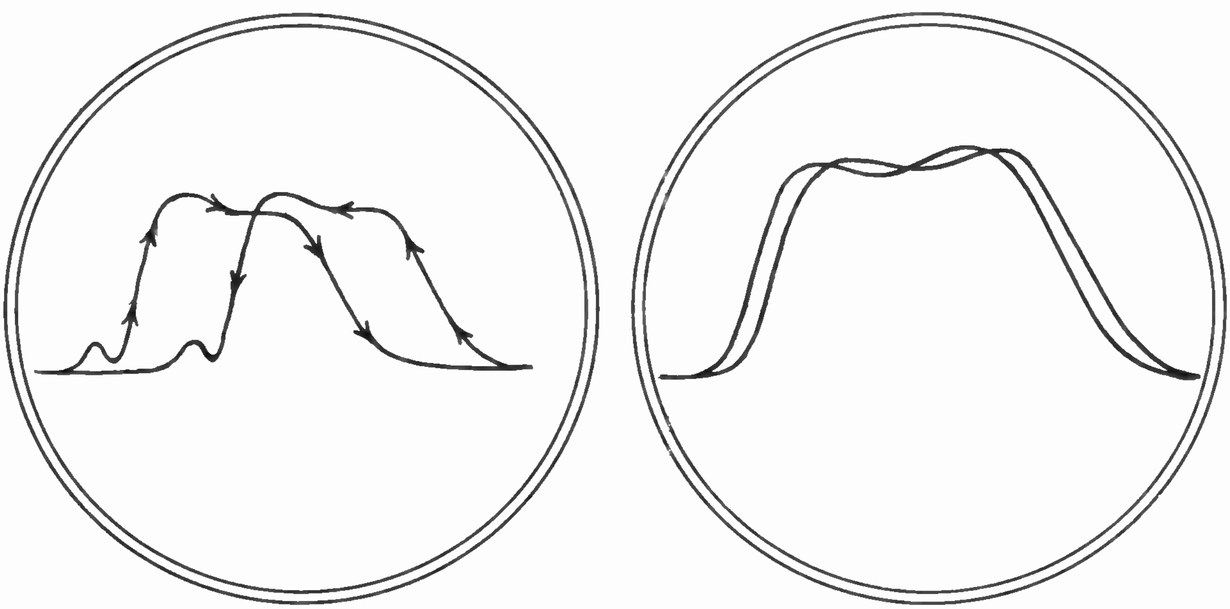


Fig. 86-10. These forward and return traces are not in phase or are not synchronized.

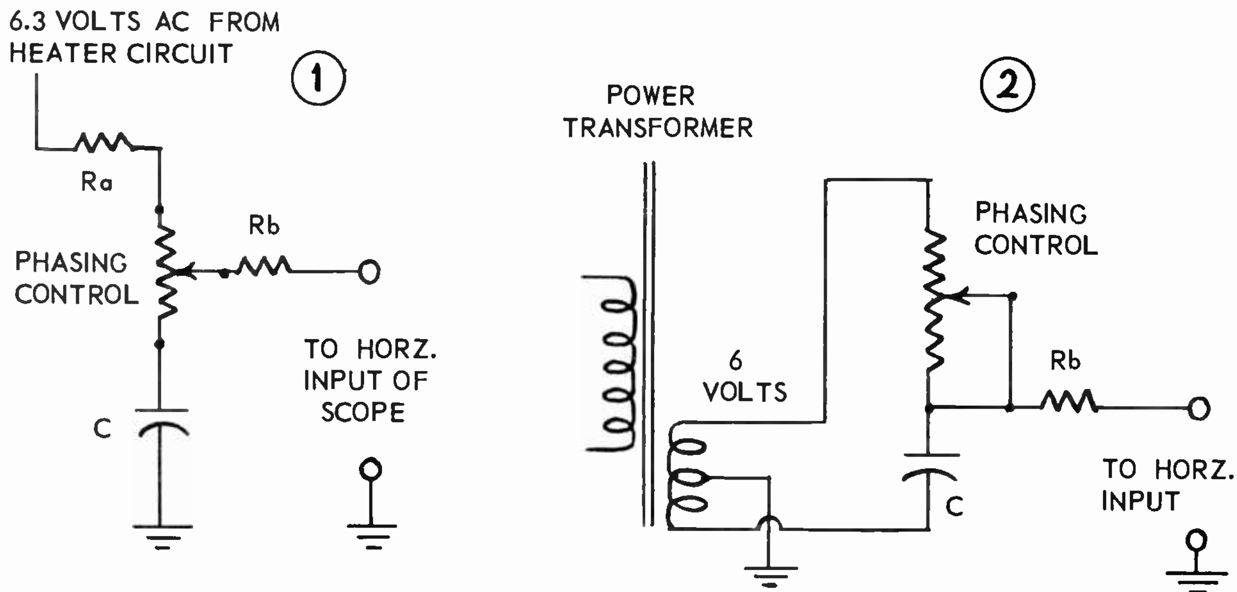


Fig. 86-11 Methods of phasing the horizontal deflection voltage for the oscilloscope.

when the horizontal deflection voltage from sweep generator to oscilloscope is not exactly in phase with the variations of frequency from the generator to the vertical input of the oscilloscope. Because of differences between inductances, capacitances, and resistances in circuits of the generator, oscillator, and circuits from which the horizontal deflecting voltage is obtained, it is not likely that deflection and frequency variations will be in phase.

To produce a single trace, it is necessary to bring the horizontal deflecting voltage into phase with the frequency sweep. This may be done with circuits such as shown by Fig. 86-11. In both diagrams there is a phasing control resistance in series with a capacitor C. The a-c voltage across the capacitor is approximately 90 degrees lagging with reference to current in the capacitor, while voltage across the resistance is in phase with current in the resistance and in the capacitor. By varying the resistance of the phasing control the voltage going to the horizontal input of the oscilloscope may be varied in phase or in its time relation to any other voltage or voltages taken from the same source or taken from within the sweep generator. The frequency sweep rate is controlled by such a voltage.

Resistor Ra, when used on a source of approximately 6 volts, usually is about 10,000 ohms. Resistor Rb may be anything from 25,000 to 100,000 ohms or more. Capacitance at C may be between 0.002 and 0.02 mf. The phasing control potentiometer will have total resistance of a half megohm to one megohm in most cases. The circuit at 1 will shift the voltage phase either ahead or behind a reference time. It is connected to a point in the generator such that the correction compensates for the original undesired phase shift. The arrangement at 2 employs a center tapped transformer winding that is not used for any other purpose. It will shift the phase either ahead of or behind another reference voltage originating from the same building power lines.

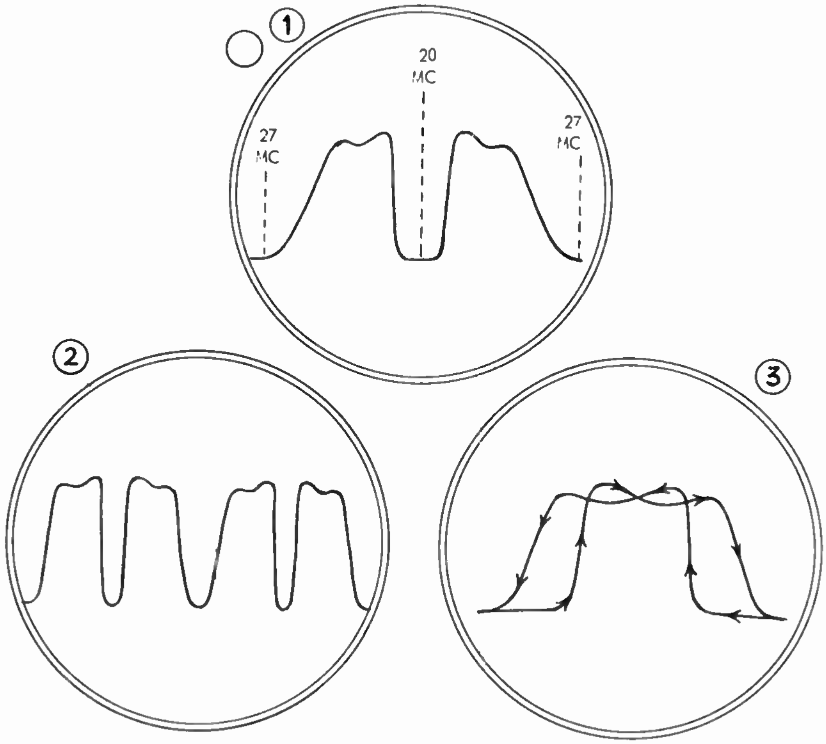


Fig. 86-12. Frequency response curves or traces produced by using the internal sweep of the oscilloscope set for 60 cycles (1) for 30 cycles (2) and for 120 cycles (3).

The phasing control usually is built into the sweep generator, and its output is fed to the horizontal input of the oscilloscope through an external cable. Some oscilloscopes have a phasing control built into them. Positions are available on the sync selector switch and possibly on the frequency range switch for bringing this control into operation and connecting it to the horizontal amplifier through connections inside the scope. A phasing control knob or pointer then appears on the front panel. The control shown at 2 in Fig. 86-11 could be built into the sweep generator, into the oscilloscope, or as a separate unit.

Any method of bringing the sweep of the CRT electron beam into phase with the variations of frequency is said to provide a synchronized sweep. Instead of employing a synchronized sweep voltage for the horizontal amplifier of the oscilloscope the internal sweep may be used. With the internal sweep adjusted to match the rate of frequency sweep, ordinarily 60 cycles per second, there will be two response curves on the screen of the scope, as at 1 in Fig. 86-12. One curve is formed while frequency is decreasing and the other is formed while frequency is increasing. This is because the CRT beam travels across the screen during 1/60 second, while there are changes of frequency, either up or down, during each 1/120 second.

The horizontal centering control of the scope may be used to center either of the response curves on the screen, which will throw the other one nearly off the screen. Then the horizontal gain control may be used to widen the curve that has been centered. The shape of a response curve thus produced will not be exactly the same as with a synchronized sine-wave sweep, because the sawtooth internal sweep is nearly linear while the sine wave is not linear. The difference in shape ordinarily is not enough to cause any difficulty in alignment.

If the internal sweep of the scope is adjusted for 30 cycles per second there will be four response curves on the screen, as in diagram 2. If the sweep is set for 120 cycles per second the two response curves will appear together, about as shown by diagram 3. Since one curve is formed with frequency going up and the other with frequency going down, they cannot be superimposed to form a single trace—the humps and dips won't match on the two curves.

THE MARKER GENERATOR. Our trace of frequency response would be of little practical use unless it were possible to know the exact frequency in megacycles occuring at each and every point along the curve. With a response curve such as shown on the oscilloscope at the upper right in Fig. 86-13 we might

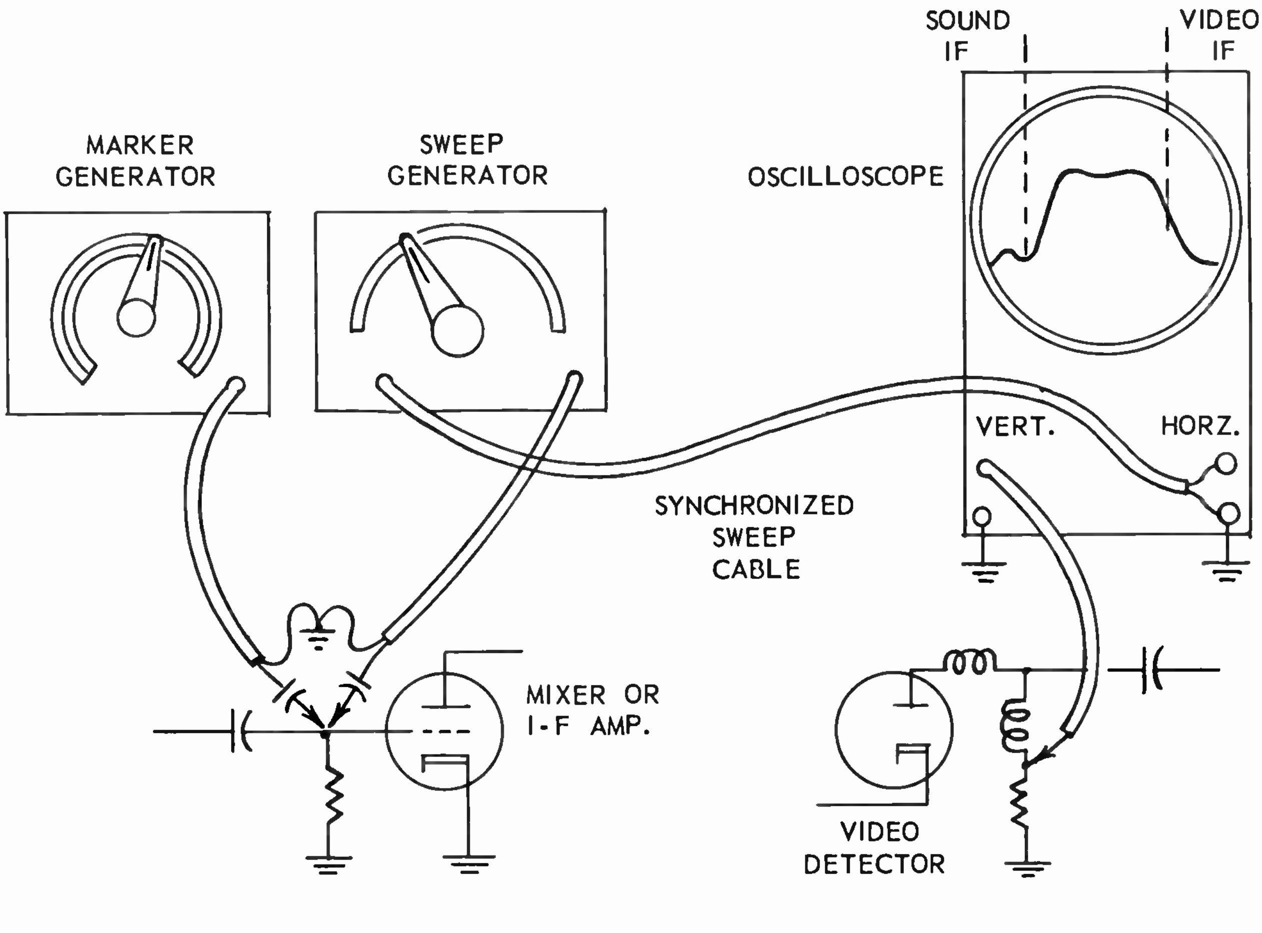
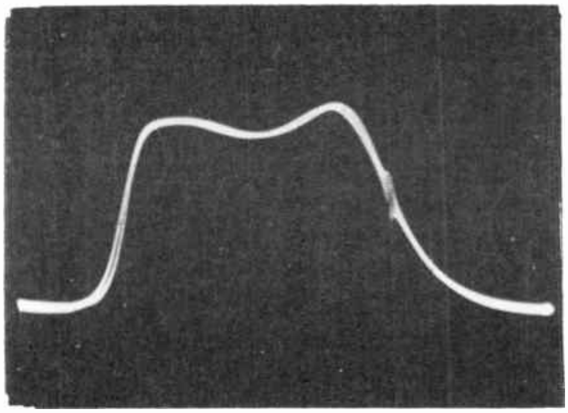


Fig. 86-13. Connections for the marker generator, the sweep generator, and the oscilloscope when response frequencies are to be identified.



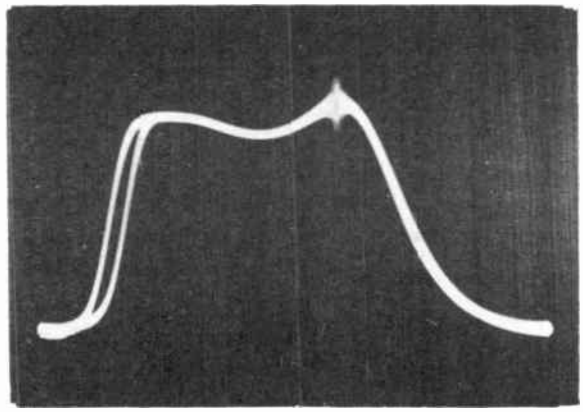


Fig. 86-14. Pips produced on the frequency response trace by action of the marker generator.

wish to know whether the dip caused by a trap really is at the sound intermediate frequency, and whether a point half way up on the high-frequency side of the curve really is at the video intermediate frequency.

The frequency at any point along the response curve may be identified by using a marker generator in connection with the sweep generator and oscilloscope. While the sweep generator is continually producing the whole band of frequencies within which there is amplifier gain, the marker produces only a single constant frequency. The outputs of the sweep generator and marker generator are fed together through isolating capacitors to the grid of the mixer tube or of any other amplifier tube.

If the frequency from the marker generator is adjusted to fall within the range of frequencies being produced by the sweep and amplified in the receiver, the sweep frequency will become the same as the marker frequency once during the forward trace and again during the return trace. For instance, with the sweep frequency varying between 20 and 28 mc, and the marker frequency at 24 mc, every time the sweep goes through its 24-mc value the two frequencies will be equal.

At the instant in which the frequencies from the two generators become equal the generator output voltages will add their amplitudes. The two voltages beat together to produce a large instantaneous amplitude which shows up on the trace as illustrated by Fig. 86-14. As the marker generator is tuned through the range of swept frequencies the point of extra amplitude, usually called a marker "pip", will move all the way across the trace or the response curve. The marker generator may be tuned to bring the pip to any desired position on the response curve. In the picture at the left the marker has been tuned to bring the pip about half way up on the high frequency side of the curve, where the video intermediate frequency should be. Then, merely by looking at the tuning dial of the marker generator, we may read the frequency at this point on the response curve and compare it with the video intermediate frequency at which the amplifier should operate.

In the picture at the right the marker pip has been moved to the peak which is nearest the video intermediate frequency on the response curve. The tuned frequency of the marker then is the frequency at which this peak is being formed. During the process of alignment we might work in another way. Say we know that the video intermediate frequency should be 25.75 mc. The marker generator is tuned to this frequency, and on the response will appear a pip at 25.75 mc. Then the alignment adjustments are altered to make this pip, and its frequency, come about half way up on the high-frequency slope of the response curve.

The sweep generator and marker generator often are constructed as a single instrument, which may be called a television alignment generator. Some instruments do not contain a marker oscillator circuit whose voltage combines with the sweep output, but have only a sharply tunable circuit which may be resonated at any frequency. When this circuit is coupled to the same point as the sweep output there is absorption of r-f energy at the marker frequency. Instead of a pip extending vertically through the response curve there appears a "dip", which is a short gap or possibly an actual downward dip in the curve at the frequency to which the marker circuit is tuned.

REQUIREMENTS FOR MARKER GENERATOR. The frequency ranges of the marker generator should be the same as those of the sweep generator, namely, 4.5 mc, then 20 to 50 mc, and about 50 to 220 mc or higher.

If used only for a marker, the generator requires only a low maximum output voltage. However, the same generator nearly always will be used for alignment of oscillators, for the peak responses of stagger tuned i-f stages, for work with the sound amplifiers, and various other uses where the maximum signal voltage output should be about the same as that from the sweep generator, or about 0.1 volt.

The attenuator or output voltage control should be capable of bringing the signal voltage down to only a few microvolts. If this cannot be done, much the same effect may be had by using a very small fixed capacitance between the high-side output lead and the grid or other tube element to which it is connected. The output cable always must be shielded and both ends of the shield must be grounded. It is highly important that the entire marker generator be well shielded by and within its housing.

Accuracy of frequency markings on the tuning dial is especially important when aligning tuner circuits. It should be possible to determine frequencies within 1/5 megacycle for accurate alignment. On channel 11, where the video carrier is at 199.25 mc, an accuracy of 1/5 megacycle allows an error no greater than 1/10 of one per cent. For i-f alignment it is desirable, but not quite as essential, to have high accuracy, for the following reason. Supposing that the video intermediate frequency is supposed to be 25.75 mc and the sound intermediate to be 21.25 mc. If you make an error and adjust the video for 26.00 mc and the sound for 21.50 mc there will be no perceptible difference in performance. With a dual sound system the sound intermediate frequency then would become 21.50 mc, and the ratio detector or discriminator would be adjusted to operate at this center frequency. The error of 0.25 mc is about 1 per cent of 25.75 mc.

The marker generator output frequency should show little or no drift after the warmup period. Automatic voltage regulation often is used in the generator power supply, and temperature compensating capacitors are in the oscillator circuit.

OSCILLATOR OUTPUT COUPLINGS. When the oscillator of a marker generator or of any other r-f signal generator is connected or coupled to receiver circuits, there is a tendency for inductances and capacitances in these circuits to alter the oscillator frequency. This, of course, will affect the frequency accuracy and is highly undesirable. In many signal generators you will find an amplifier stage between the output of the oscillator and the leads going to the receiver. This is called a buffer amplifier. Its purpose is to isolate the oscillator from the externally connected circuits and thereby prevent these circuits from

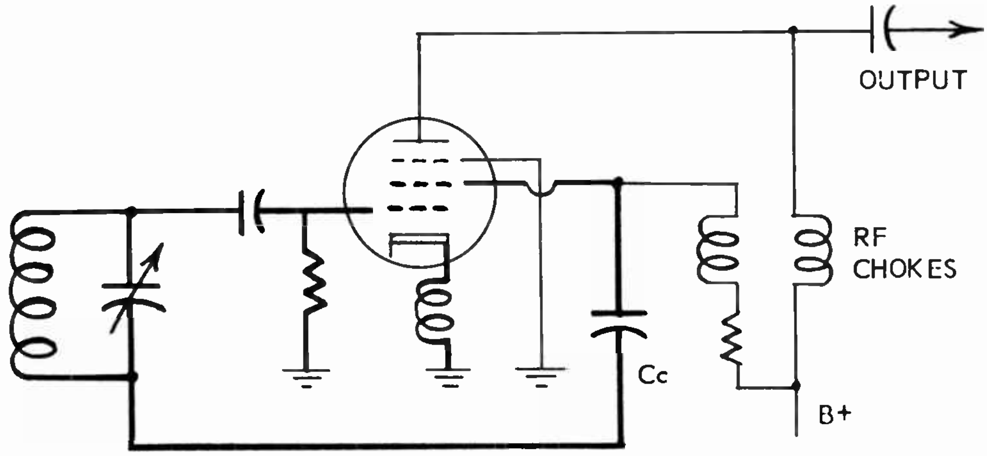


Fig. 86-15. Circuit for an electron coupled oscillator.

altering the oscillator frequency. A buffer amplifier ordinarily is of an untuned type.

Another method of isolating the oscillator circuits from externally connected circuits employs an electron coupled oscillator. This system is applied to a Colpitts oscillator in Fig. 86-15. It may be used with any triode oscillator circuit.

Although the tube is a pentode type the oscillator circuit of Fig. 86-15 really is operating as a triode type, as shown by heavy lines. The grid of the tube is used in the usual manner. The screen is used as a triode plate, and r-f voltage of the screen is coupled to one end of the tuned circuit through capacitor Cc.

B-voltage is applied to the screen and to the regular plate of the tube through r-f chokes to avoid grounding the signal voltages. The plate is operated at a higher voltage than the screen in order to attract electrons from the cathode to the plate. The rate of electron flow and r-f variations of this flow are controlled by the screen, acting as the oscillator plate, but the electron flow and r-f current variations continue through to the regular plate from which the accompanying r-f voltage goes to the output. The suppressor grid is grounded to provide further isolation between the oscillator circuit and the regular plate which is part of the output circuit.

Inductances and capacitances in an externally connected circuit have relatively little effect on the oscillator. This is because, in any screen grid tube, changes of voltage on the plate have much less effect on electron flow than have changes of voltage on the screen.

AUDIO MODULATION. The majority of r-f signal generators, including the marker variety, have provisions for amplitude modulation of the r-f output voltage at an audio frequency. The audio frequency for modulation most often is at approximately 400 cycles per second, although sometimes it is at 1,000 cycles, or may be variable. The r-f output voltage usually is modulated 30 per cent when this value is not ad-

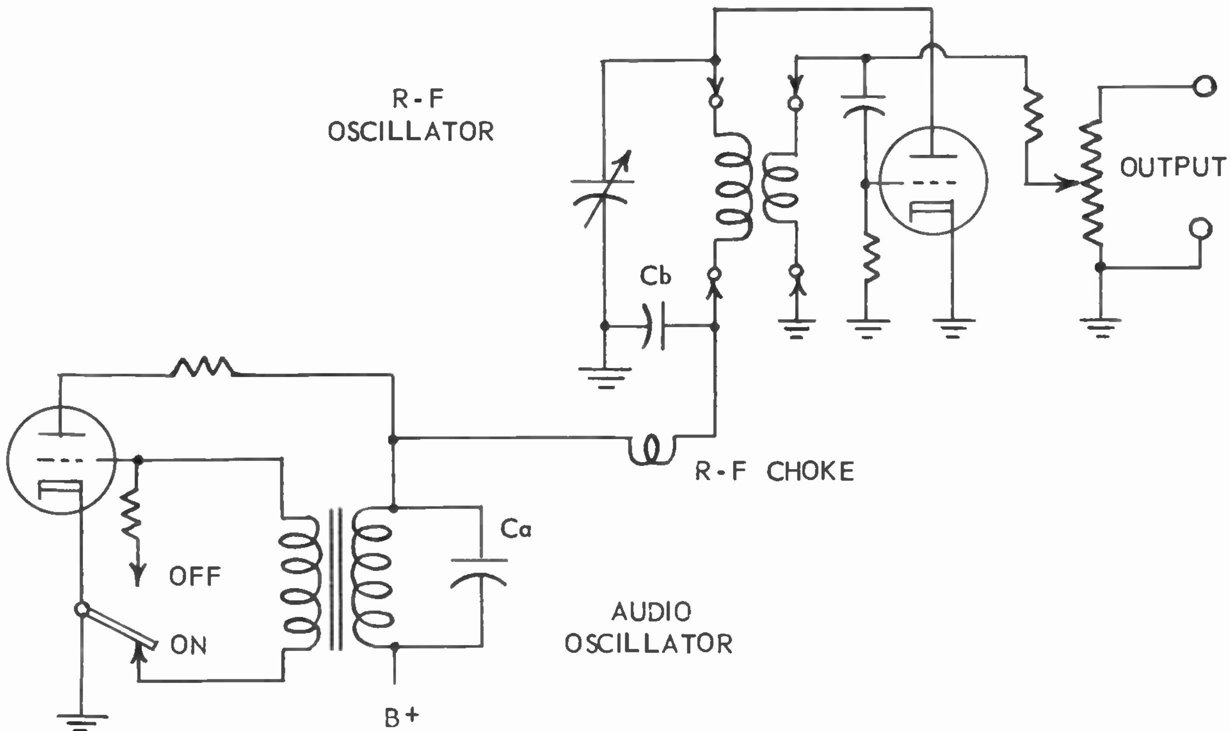


Fig. 86-16. One method of applying audio modulation to the output of an r-f oscillator.

justable. Some generators provide 30 per cent modulation and some other higher percentage, while still others allow adjustment of the modulation from zero to nearly 100 per cent or even for over-modulation.

One method of applying audio modulation to an r-f oscillator in a signal generator is shown by Fig. 86-16. The audio-frequency oscillator is a triode with feedback from plate to grid through an iron cored transformer. The plate winding of this transformer is tuned to the audio frequency by fixed capacitor Ca. With the switch in position for modulation on, as shown, the grid of the a-f oscillator is connected through the grid winding of the feedback transformer to ground and the tube cathode. With the switch in its "off" position, the lower end of the grid winding is open circuited and the grid of the tube is connected to ground through a resistor.

Plate current for the r-f oscillator goes from the plate of the r-f oscillator tube through the tuned plate winding, the r-f choke, and the a-f oscillator plate winding to B-plus. Consequently, the plate voltage on the r-f oscillator is varied at the audio rate in the audio feedback transformer, and the r-f output is varied or modulated at this rate or frequency. R-f currents for the r-f oscillator plate circuit are kept out of the audio circuit by the r-f choke, but pass to the variable tuning capacitor to complete the tuned circuit through a large capacitance at Cb.

Instead of applying the modulating voltage to the plate of the r-f oscillator it may be applied to the

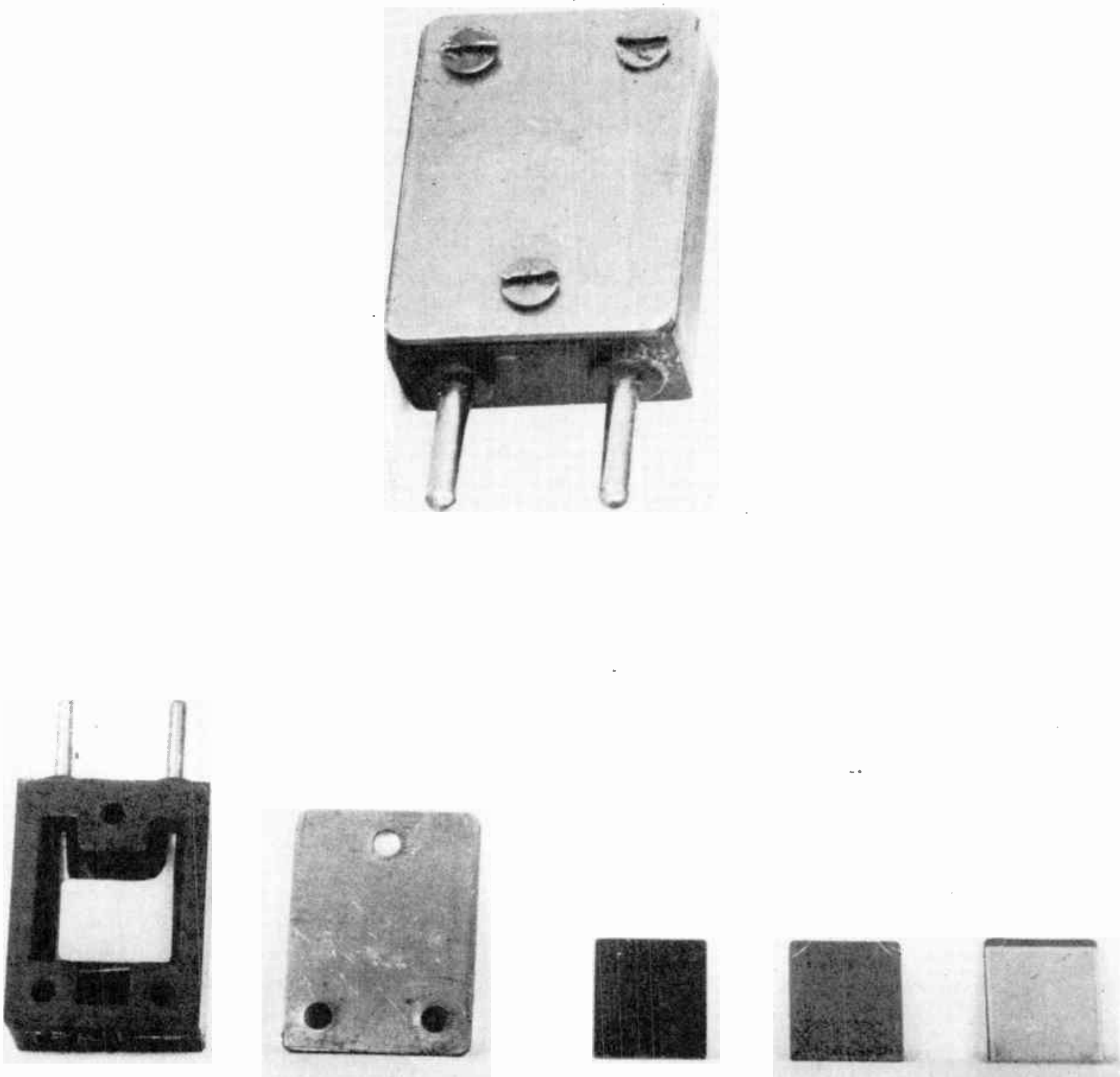


Fig. 86-17. A quartz crystal as it appears in one type of holder and when disassembled.

screen when this oscillator is a pentode. With electron coupled r-f oscillators the suppressor element may be coupled through a transformer winding or through a capacitor to the audio oscillator circuit instead of being grounded as in Fig. 86-15.

CRYSTAL FREQUENCY CONTROL. By far the most accurate and dependable means for obtaining a precise and constant frequency is to use a "piezo-electric" crystal, which is simply a small, thin piece of the mineral quartz which has been ground in a certain manner from the original large mother crystal, and made of some very exact thickness.

A quartz crystal is a highly elastic body. The scientific definition of this elasticity is not that the crystal will stretch like a rubber band, but that it will resist mechanical deformation and will quickly recover its original form or size after a stress is relieved. Any elastic body has some natural frequency of expansion and contraction or of vibration. If you snap a steel spring it will vibrate at a frequency depending on its material, dimensions, and shape. When a difference of potential is applied to the opposite flat faces of the quartz crystal the crystal will become slightly thicker or thinner, depending on the polarity of the applied potentials and on the manner in which the crystal has been formed or cut.

If the applied potential difference is alternating, and of the same frequency as the natural vibration frequency of the crystal, the crystal will vibrate vigorously at this frequency.

Now comes the strange part. When a quartz crystal becomes alternately thicker and thinner as it vibrates, the crystal produces electric potentials of its own. The potentials on opposite faces are alternating, of opposite polarities or both of zero amplitude at every instant, and they occur at precisely the natural frequency of the crystal.

When the crystal is connected in series with the grid circuit of a tube, and the B-power is turned on, the crystal is subjected to a small potential difference. This is enough to start the vibrations. The alternating potential developed by the crystal is amplified in the tube and part of the power from the plate circuit is fed back to the grid circuit and the crystal. This feedback need be only enough to make up for the mechanical power used by expansion and contraction of the crystal.

The crystal is the electrical equivalent of a series resonant circuit whose resonant frequency is the natural frequency of the crystal. The Q-factor of the crystal is much higher than that of any ordinary resonant circuit that could otherwise be constructed, which makes the oscillating frequency almost independent of the connected circuits. The crystal forces the oscillator to operate at the frequency for which the crystal is cut, or at some harmonic of this frequency.

Crystals are supported within some form of holder which allows easy connection to the oscillator circuits. One of the most common types is shown at the top of Fig. 86-17. Some have banana plug pins instead of the solid form. Holders may be fitted with solder lugs instead of pins. In some cases the holder has base pins like those of a standard octal tube, or may have a five-pin base.

The body of the crystal holder is of insulating material, as at the left in the lower row of pictures in Fig. 86-17. Next is the metal cover. At the right of the cover are two metal plates between which the crystal is supported, and at the far right is the quartz crystal itself. Some crystals do not require the metal plates for transfer of the alternating potentials, but have metallic films deposited on opposite faces of the quartz. Because the quartz is an excellent dielectric material the crystal in its holder acts like a capacitor at frequencies removed from resonance.

The thinner a quartz crystal the higher is its natural frequency, but the greater is the danger of fracture

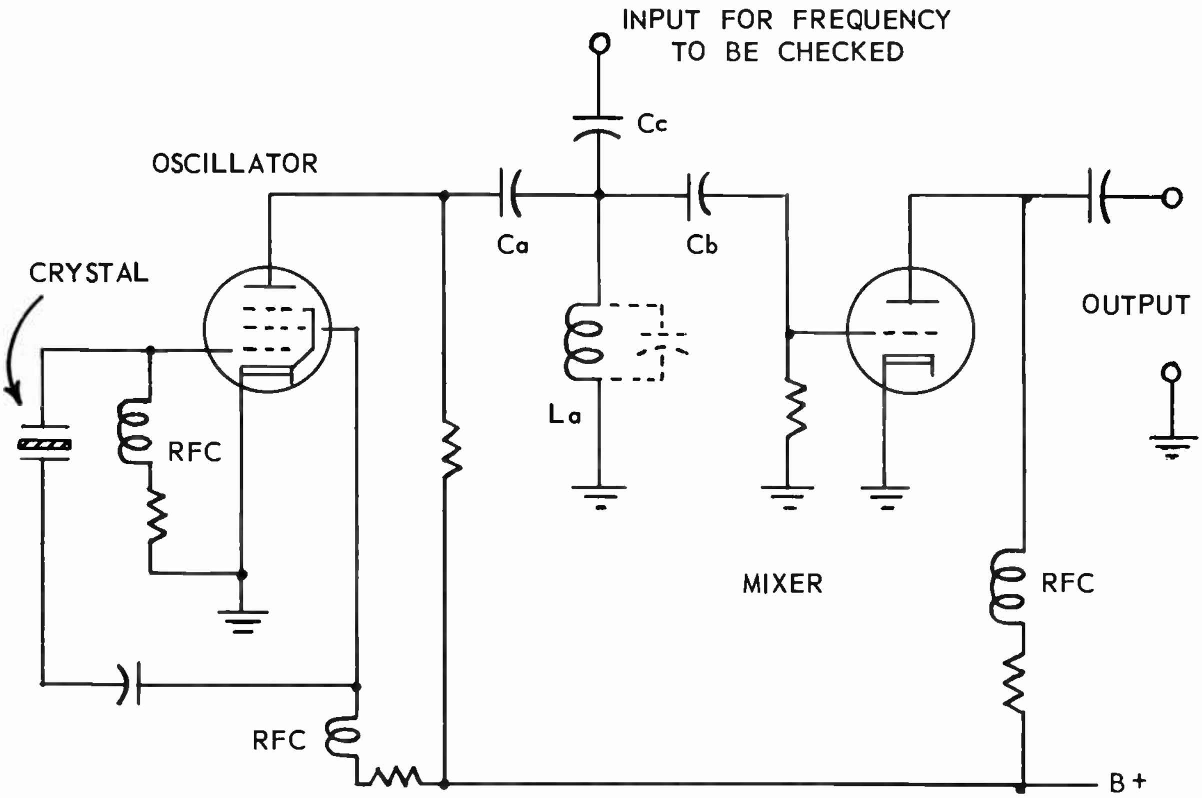


Fig. 86-18. A frequency calibrator employing a crystal controlled oscillator and a mixer circuit.

when high potentials are applied in an effort to make the crystal vibrate too violently. Crystals in general use for service type signal generators usually have natural or fundamental frequencies no higher than 10 mc. Since a crystal controlled oscillator may be designed to furnish harmonic frequencies up to the 100th and even higher, this is no handicap. The quartz can be cut and ground for fundamental frequencies as high as 50 mc, but it seldom is done.

The circuit most often used for crystal controlled service instruments is the Pierce oscillator circuit, of which one application is shown by Fig. 86-18. The entire instrument shown here is a crystal calibrator which is used for checking or calibrating the frequency being furnished by a marker generator or any other r-f signal generator.

The crystal controlled oscillator, at the left, is an electron coupled type. The crystal forms the resnant grid circuit which controls the operating frequency. The oscillator output goes from the plate of the pentode tube through capacitors Ca and Cb to the grid of the mixer tube at the right. The output or part of the output from the signal generator to be checked goes through capacitors Cc and Cb to the grid of the mixer. The output of the mixer may go to headphones or to an amplifier and speaker.

The crystal controlled oscillator will furnish its fundamental frequency and, with a suitable plate impedance at L_a and possibly some approximate tuning capacitance across this inductor, will furnish strong harmonics at least to the 50th and much higher with reasonably good construction in the circuits. Frequencies from the instrument being checked beat with the crystal oscillator fundamental and harmonics. When this external frequency is exactly the same as any of these crystal oscillator frequencies there will be zero beat in the mixer output. Then the reading of the tuning dial on the instrument being checked is compared with the oscillator (crystal) frequency or a harmonic to determine any error of calibration in the instrument being checked. Various plug-in crystals may be used to give harmonic frequencies in the required range.

Crystals may be marked with their fundamental frequency, but oftentimes the marking is of some harmonic frequency at which the particular crystal may be used or for which it originally was intended. Such markings are easily detected by checking for zero beat against such simple fractional frequencies as 1/2, 1/3, 1/4 or still smaller fractions of the marked frequency until arriving at the lowest frequency of oscillation. This is the fundamental of the crystal.

The fact that crystal oscillators furnish many harmonic frequencies allows checking and calibration in the television carrier ranges. For example, a crystal having a 25.75 mc fundamental will deliver a third harmonic which is the video carrier frequency of 77.25 mc for channel 5. A fundamental of 24.156 mc will have an eighth harmonic of 193.248 or practically 193.25 mc, which is the video carrier frequency for channel 10.

Most crystals of good quality such as used in service instruments have accuracy or plus or minus 0.02 per cent or 1/50 of one per cent at the standard operating temperature of 25° C. or 77° F. and at temperature variations down to about 40° F. and up to about 120° F. Operating temperature depends not only on that of the air and surrounding parts, but also on the amount of r-f current carried by the crystal. Plate voltage on crystal controlled oscillators should be kept as low as allows satisfactory performance or moderately strong outputs, this is to avoid overheating. There will be r-f current in the crystal just as there is in the grid circuit of any other self-biased oscillator.

Do not open the holder of a crystal to see how the oscillations originate. The least particle of dust, or the slight oiliness from a slight touch of the fingers may cause serious damage. An inactive crystal usually is the result of dirt, oil, or moisture getting inside the holder.

FOR GRADING: ____

TELEVISION

LESSON NO. 87

ALIGNMENT WITH THE OSCILLOSCOPE

Training Your Mind...

C.T.I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind
C.T.I.
Perfecting Your Skill

COMMERCIAL

TRADES

INSTITUTE

Chicago, Ill.

LESSON NO. 87

ALIGNMENT WITH THE OSCILLOSCOPE

When aligning an i-f amplifier or other sections of the receiver with an oscilloscope, sweep generator, and marker generator, the preliminary steps are the same as when using the vacuum tube voltmeter as output indicator. This refers to overriding an automatic gain control, setting the channel selector to an unassigned channel and disconnecting the antenna, adjusting the contrast control no higher than its normal operation position, and so on.

Next in order is connection of the test instruments to the receiver and to each other. For aligning the i-f amplifier the output of the sweep generator is connected to the grid of the mixer tube through a fixed capacitor of something between 100 and 1,000 mmf, or is connected directly to a sleeve slipped over the outside of this tube. Connect the output of the marker generator to the mixer grid through a fixed capacitor of 5 to 25 mmf, or to the sleeve on the mixer tube through 1,000 mmf or more. Do not connect the outputs of the two generators directly together but only through a capacitor, no matter how the outputs are connected to the mixer.

If the sweep generator provides synchronized sweep voltage, connect the leads from this voltage output to the horizontal input and the ground connection of the oscilloscope, and set the scope controls for utilizing this horizontal input voltage. Otherwise use the internal 60-cycle sweep of the scope or use the regular internal sweep with adjustments for operation at 60 cycles, as described earlier.

Connect the vertical input of the oscilloscope directly to the high side of the video detector load resistor. Make sure that the ground clips for the cable shields of all the instruments are connected to ground or to B-minus, depending on the type of receiver. Do not connect some of the cable grounds to chassis ground and others to B-minus, all must go to the same part of the receiver circuits.

Turn on the receiver and all the test instruments, and let them warm up for a few minutes. Some sort of trace should appear on the screen of the oscilloscope if the connections have been correctly made. The center frequency should be adjusted until the center of this hump is near the center of the screen. If the initial trace is only a straight horizontal line, adjust the center frequency of the sweep generator until a hump appears.

Try adjusting the attenuator or output control of the sweep generator, it should vary the amplitude of the curve. Adjusting the contrast control may also change the amplitude of the curve providing that it controls the gain of the video I.F. amplifiers. Next alter the sweep width control of the sweep generator. This should make the curve on the oscilloscope screen become wider or narrower in a horizontal direction without changing the height. As the curve is thus made narrower the horizontal base lines or zero gain

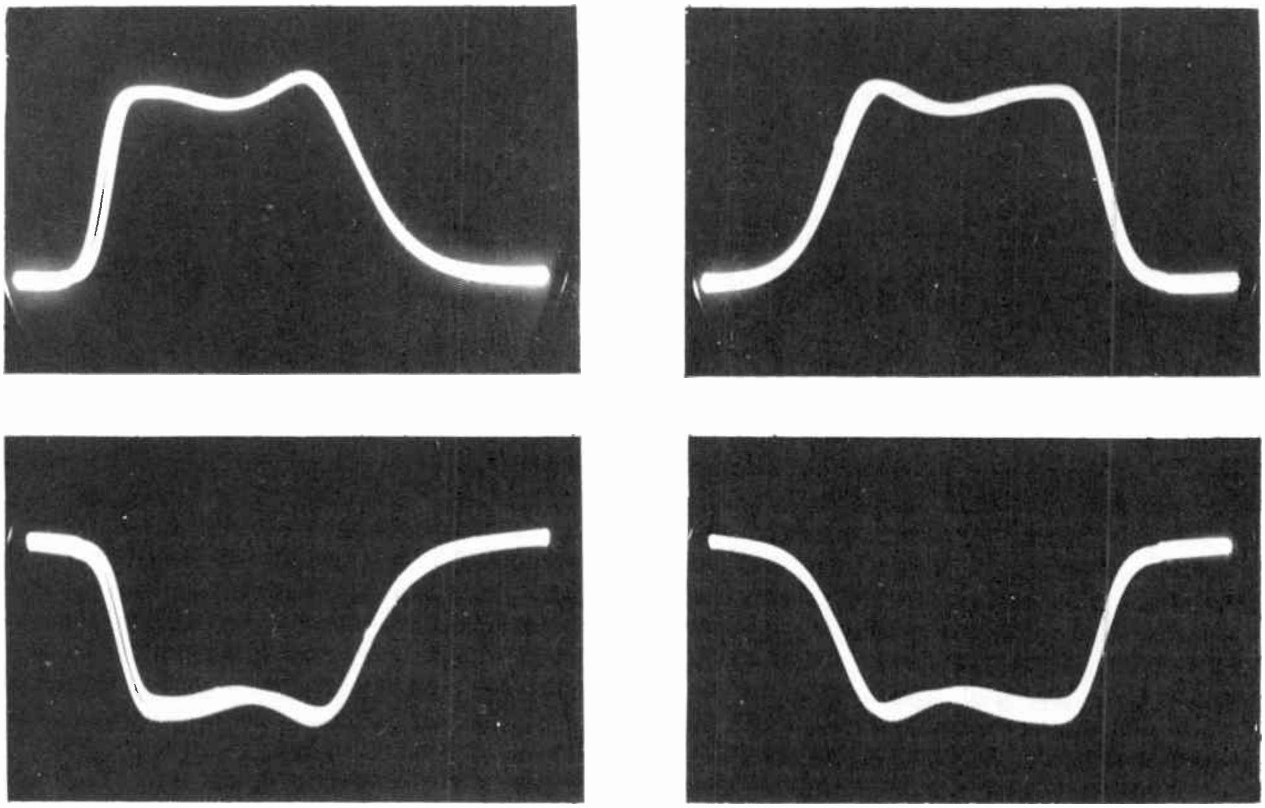


Fig. 87-1. Trace polarity may be positive or negative, and frequency may increase in either direction, depending on the instruments and how they are connected.

lines on both sides will become longer. If the sweep width is reduced enough, the horizontal base lines will disappear and the sides or skirts of the curve will come out to the edges of the screen.

Work back and forth between the various adjustments and controls which have been mentioned until you obtain a trace on the general order of those illustrated by Fig. 87-1.

The curve at 1 was formed with the oscilloscope connected to the top of the video detector load resistor. For the curve at 2 the scope was connected to the plate of the first video amplifier. The polarity has been inverted by the amplifier tube, between its grid and plate circuits, but the frequency response is unchanged. At both 1 and 2 the frequency increases from left to right, bringing the sound intermediate on the left and the video intermediate on the right side of the curve.

The curve pictured at 3 was formed with the oscilloscope connection moved back to the video detector load resistor, but with the synchronized sweep voltage changed to make it of opposite phase, in relation to the frequency sweep. Now we have the same polarity as at 1, but frequency increases from right to left.

This brings the video intermediate frequency at the left and the sound intermediate at the right. For curve number 4 the sweep voltage is of opposite phase, as for 3, but the oscilloscope was again moved ahead to the plate of the first video amplifier. Here the polarity is inverted, and frequency increases from right to left. The frequency response remains the same no matter what the polarity of the trace and no matter which direction the frequency increases. Any one of these curves can be used in the alignment of the video I.F. amplifiers.

Just how a frequency response curve appears on the screen of the oscilloscope depends on the relations of frequency sweep and horizontal deflection voltage, on whether your oscilloscope shows "positive up" or "positive down", and on the point in the receiver to which the scope is connected.

Now for some of the finer points in adjusting and connecting the test instruments. The stage being aligned must not be overloaded with excess signal strength from the sweep generator. Keep the generator output down where a slight change does not alter the form of the trace by changing the relative heights or peaks and valleys, but only changes the height. If the generator attenuator or output control won't bring the voltage low enough, use a smaller fixed capacitor from the high side lead to the tube in the receiver, or, if you are using a coupling sleeve, slide this sleeve up higher on the tube envelope. As you proceed with alignment, the response curve may tend to become higher. Keep it down by reducing the output of the sweep generator, not by turning down the gain control of the oscilloscope.

Adjust the sweep width control to bring the entire response curve onto the screen of the oscilloscope, with a little bit of the zero base line or horizontal extension showing on both sides. The number of megacycles of sweep width is of no importance, just so long as you can see the entire response with only enough base line to identify zero response. Do not alter the sweep width during a series of adjustments. This would change the apparent slopes of the curve and could be confusing.

If the sweep width control is set for too narrow a sweep or for two few megacycles of sweep the result will be to send some of the response off the screen of the scope, as at the left in Fig. 87-2. If the sweep

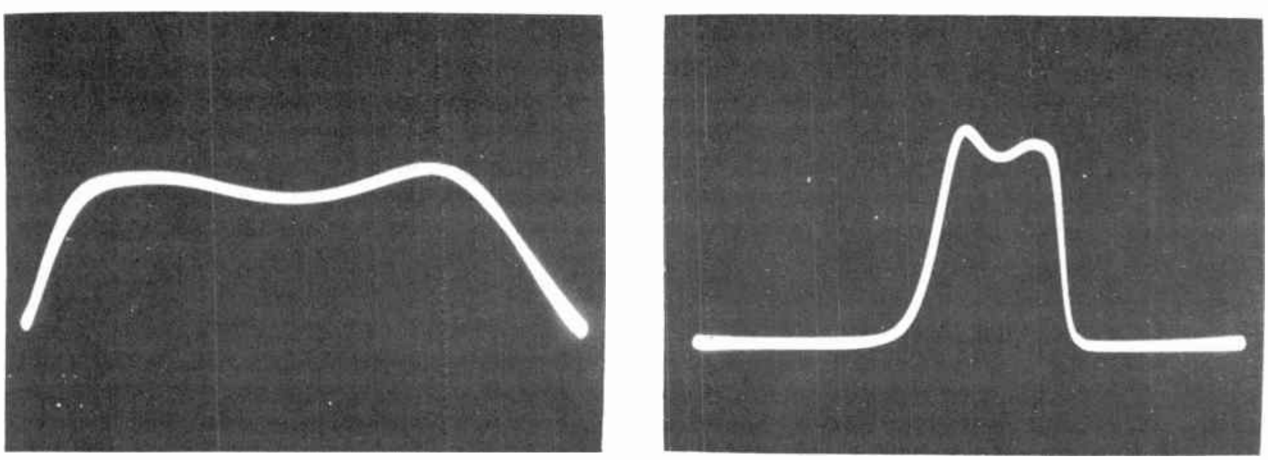


Fig. 87-2. The result of using a sweep which is too narrow (left) and of using an excessively wide sweep (right).

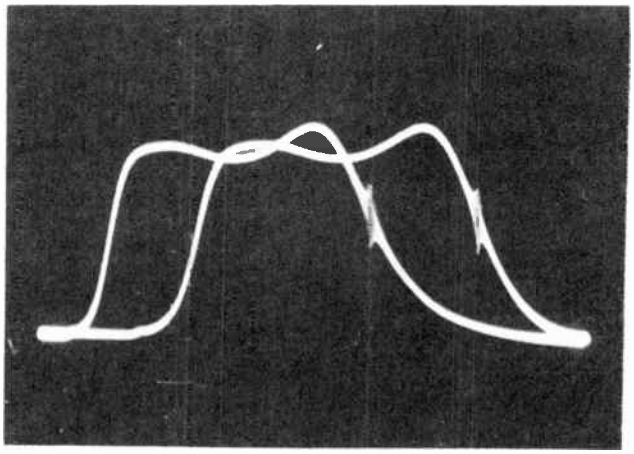


Fig. 87-3. The marker may be seen more distinctly by shifting the forward and return traces out of phase.

is made too wide, of too many megacycles, the response will appear too narrow and will have long base lines on both sides, as at the right. The frequency response is the same in both pictures. The one on the left was made with a sweep of about 5 megacycles and the one on the right with a sweep of a little more than 15 megacycles.

If the marker generator has provision for audio modulation be sure to keep this modulation turned off or to zero while producing marker pips on a response trace. For easy identification of pips the output of the marker generator must be kept low, and under no conditions should this output be great enough to distort the frequency response, or to materially lower the height of the curve. After tuning the marker to the desired frequency adjust its output so that the pip is just visible at whatever point of the trace it may be used.

The distinctness of a marker pip will increase when output of the sweep generator is decreased, and when the receiver contrast control is turned down. The marker pip will be large and distinct while you move it across the relatively flat top of the response trace where there is high amplifier gain, and will become less clear upon approaching points of zero gain. The steeper the slope of the side of the response on which the marker is placed, the more difficult it becomes to definitely spot the position. You may use the horizontal gain of the scope to temporarily widen the curve while examining a marker on one of the slopes. When using synchronized sweep, a marker on one of the slopes of the curve will become more distinct if you shift the phasing control to temporarily separate the sides of the curve. The offset forward and return traces, with a marker pip on each, appear as in Fig. 87-3.

When aligning, you adjust the interstage couplers to make the response curve of certain heights or shapes at various frequency points. For example, when working to bring the video intermediate frequency about half way up, or down, on the high-frequency slope, it is very helpful to maintain a marker pip at this frequency to show whether you are making matters better or worse.

Every change of adjustment on any one coupler affects not only one part of the response but all other

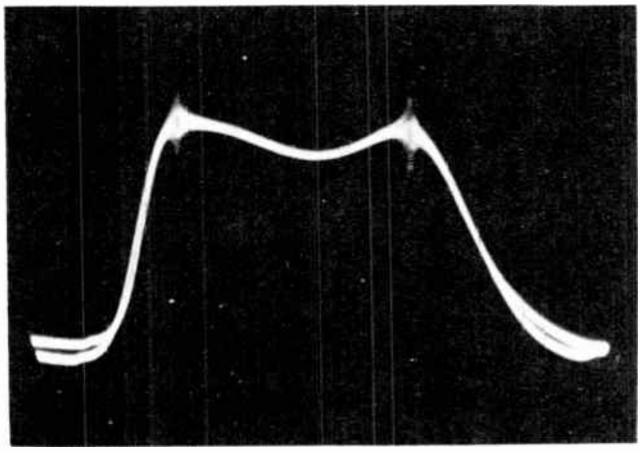


Fig. 87-4. Using two marker pips to identify two frequencies at the same time.

parts at the same time. Some one adjustment may have its greatest influence on one side, or the other side, or somewhere along the top of the response, but this adjustment will also have greater or less influence on every other part of the curve. In the case of the video intermediate frequency adjustment, while bringing this frequency where it belongs, you are more than likely to upset the response at the low-frequency end of the curve, so it is necessary to continually move the marker from one place to another in getting the entire response as you wish it to be.

It is convenient to use two markers to identify two different freqnencies at the same time, so that the effects of a given adjustment on both sides or ends of the response may be observed while attempting to improve one end. In Fig. 87-4, one marker is identifying the low-frequency peak, and another is identifying

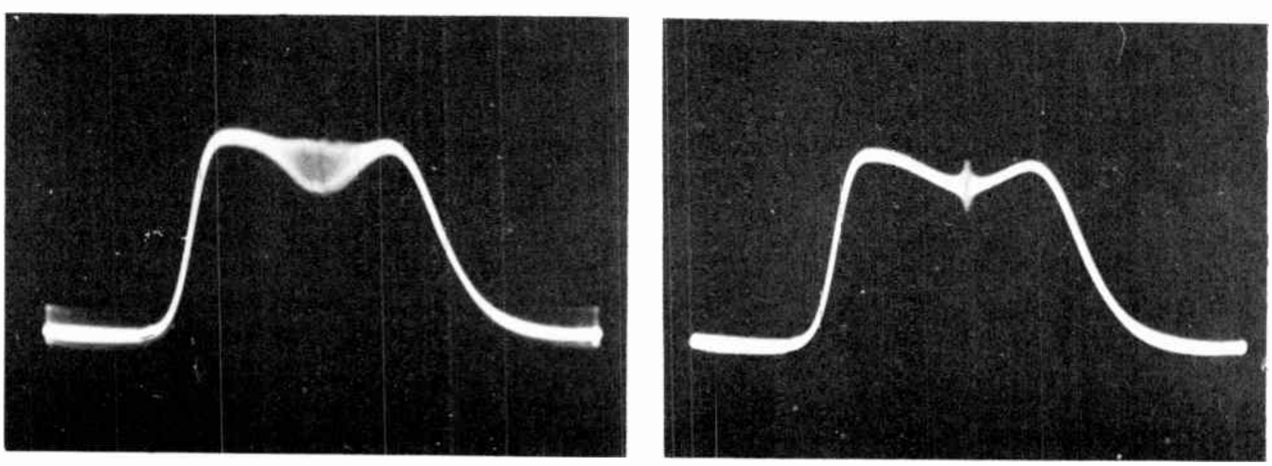


Fig. 87-5. A broad band oscilloscope will show fuzzy pips unless the vertical input is shunted with a capacitor.

the high-frequency peak. One marker might be set at the video intermediate frequency, and another at the sound intermediate, or one at the video intermediate, and the other as much lower at the desired pass band for the amplifier.

Some marker generators will provide two markers at the same time. Quite often one of these comes from a crystal controlled oscillator built into the marker generator, while the other is controlled by the variable tuning capacitor. More than two markers, or any number, might be used at the same time provided none of them are so strong as to pull the response curve out of shape.

The oscilloscope is most often connected across the video detector load resistor, as mentioned many times before. If the frequency response is weak, or if the scope is lacking in sensitivity, the vertical input may be connected to either the grid or the plate of any video amplifier tube, or it may be connected to the control grid or cathode signal input point of the picture tube. The response becomes progressively stronger in going toward the picture tube.

If you are using a broad band oscilloscope, with good gain at high frequencies, the marker pip probably will appear very wide and fuzzy, as at the left in Fig. 87-5. It is practically impossible to identify frequencies on the slopes of the trace with a pip of this kind. The pip may be sharpened, as at the right, by connecting a fixed capacitor across the receiver end of the vertical input cable of the oscilloscope. The connections are made as shown by Fig. 87-6. The capacitor pigtails may be fitted with spring clips or the attachments may be made in any other convenient manner. The bypassing capacitance may be almost anything from 500 mmf to 0.005 mf or even may be something like 0.05 mf if necessary to obtain a sharp pip.

FREQUENCY RESPONSE CURVES. Fig. 87-7 shows how the frequency responses of the separate stages of a video i-f amplifier combine their effects to form the final output at the load resistor of the video detector. The curves were taken originally from a video i-f amplifier having four tubes and five interstage couplers, each tuned to a different frequency.

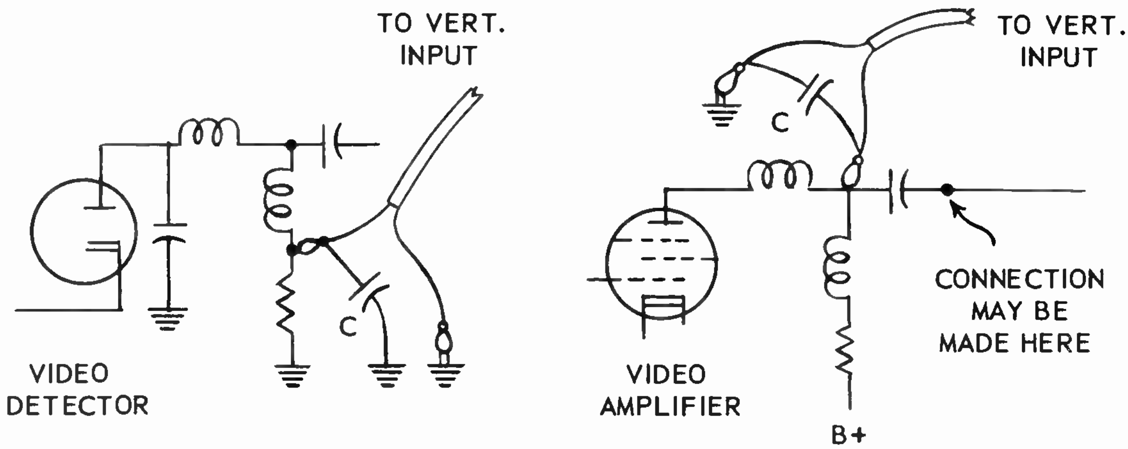


Fig. 87-6. How a shunting capacitor may be connected across the vertical input.

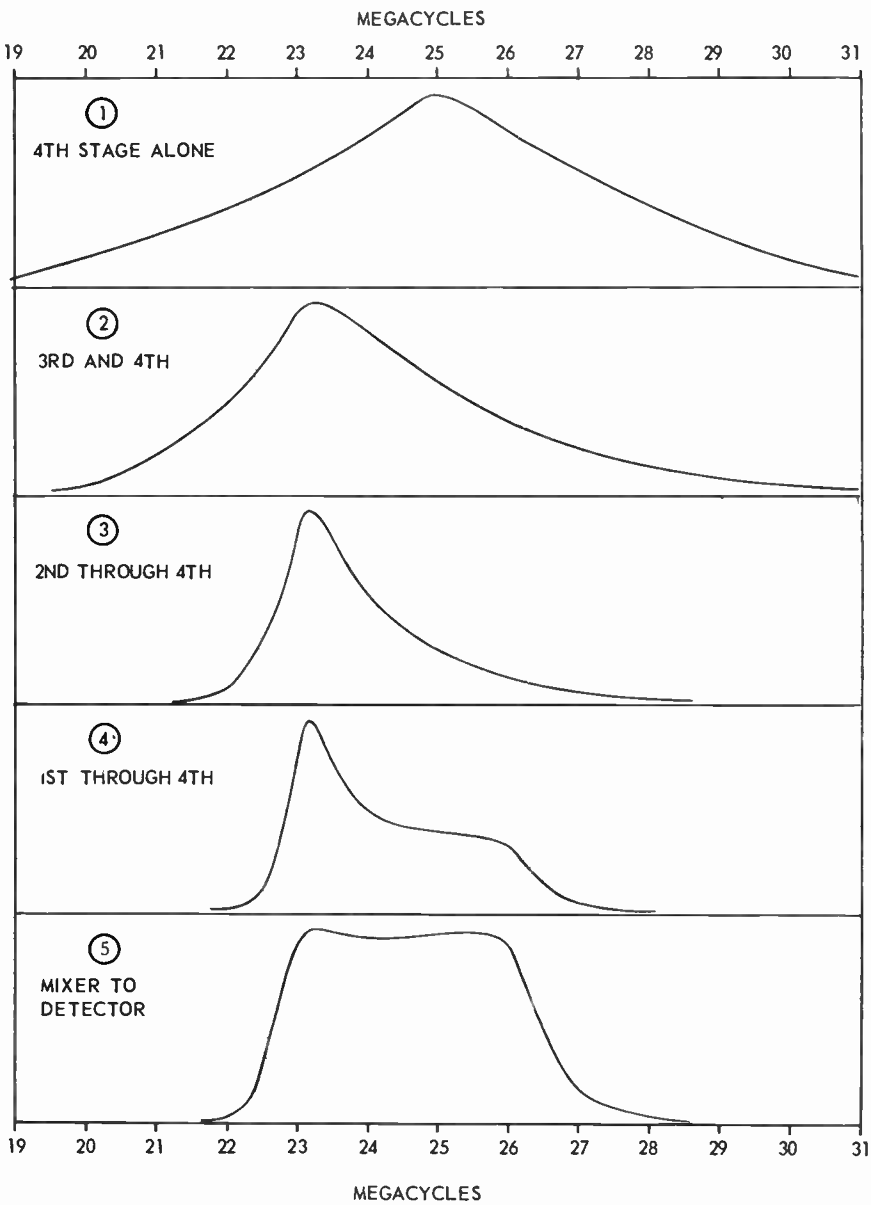


Fig. 87-7. Frequency responses as additional stages are brought into the i-f amplifier system between sweep generator and video detector.

Curve number 1, at the top, was made with the oscilloscope connected to the video detector load resistor, and with the sweep and marker generators coupled to the grid of the fourth or last i-f amplifier tube. Only this one amplifier and the coupler between it and the detector then were active, so the response is that of this one stage by itself.

For curve number 2 the generators were moved back to the grid of the preceding i-f amplifier, so that the last two stages were working together to produce the response shown. The oscilloscope remained on the detector load resistor for all the responses. For curve 3 the generators were moved farther back to bring in one more stage, this was repeated for curve 4, and for curve 5 the generators were coupled to the grid of the mixer tube to give the overall frequency response of the entire i-f amplifier system.

The final response or voltage gain at any frequency is the product of the gains at the same frequency in all the stages. That is, if gains of successive stages at a given frequency were something like 6, 8, 15, 12, and 4 the overall gain at this frequency would be 34,560, the product. Were the gain of any one stage to be zero at a certain frequency the overall gain would be zero at the same frequency, for when a gain drops to zero anywhere there is nothing to amplify in following stages. Although actual peak voltages for the curves of Fig. 87-7 differed widely, all have been reduced to the same height to clearly show the frequency characteristics.

The stages by themselves are broadly tuned, as is evident from curves 1 and 2 for a single stage and

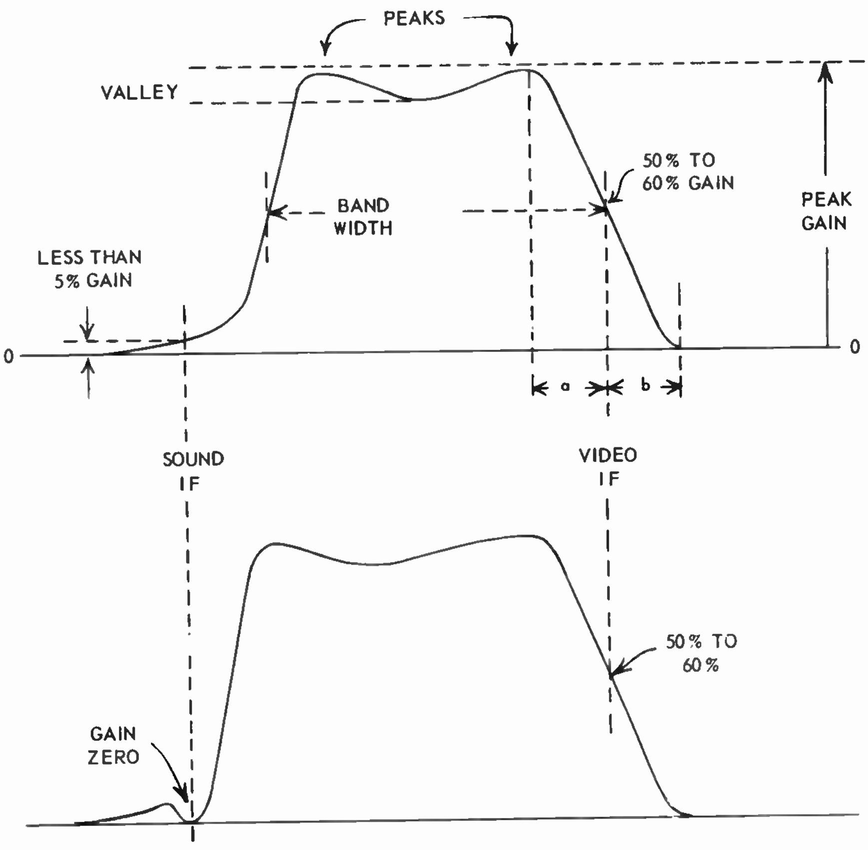


Fig. 87-8. The "check points" which are to be watched on the response curve while making i-f alignment adjustments.

for two stages in cascade. As more and more stages add their effects to the total, the response becomes narrower and the sides steeper. Were you to replace a weak amplifier tube in one stage with a good tube, the response would rise at and near the frequency to which the following coupler is tuned.

Fig. 87-8 shows most of the features you will strive for when making alignments of i-f amplifier systems. The upper curve applies to a system used with intercarrier sound. The desired characteristics are as follows.

Peak gain or maximum gain is represented by the height of the curve from zero response to the top of the higher peak, if there are peaks, or to the highest point on the curve in any case. The greater this gain, when all other requirements are satisfied, the better will be the performance.

It is absolutely essential that the video intermediate frequency be no lower than 50 per cent of maximum gain, and it should not be higher than 60 per cent. Whatever else happens to the response, this requirement must be met.

Response at the sound intermediate frequency, for intercarrier sound, must not exceed 5 per cent of the peak gain, and preferably is made 3 per cent or even less. With a higher sound response you will have difficulty in eliminating intercarrier buzz from the speaker.

Peaks occuring on the response curve must be nicely rounded off as shown. If a peak is quite high, and sharp, it indicates probable regeneration and the likelihood of oscillation.

Usually there will be two peaks. They need not be of exactly the same height, but keep the difference as small as possible.

The dip between peaks is called a valley. The bottom of a valley may drop to no less than 70 per cent of maximum gain, and even this much drop is rather bad. Keep the top of the curve as nearly flat as possible.

The band width of the amplifier is considered to be the number of megacycles horizontally across the response from half way down on one side to half way down on the other side. If this width is as much as 3.6 mc you are doing very well. It should be no less than 3 mc, for this would mean serious loss of high video frequencies and poor detail in the pictures. Although the fairly flat frequency response of the video amplifier (not the i-f amplifier) often goes to 4 mc, it is almost impossible to obtain an i-f band width so great as this without having rather elaborate types of interstage couplers.

The frequency differences marked a and b are measured from a video intermediate which is half way down the slope to the peak on one side and to zero response on the other side. Ideally these two differences would be 0.75 mc each, in order to have amplification of low video frequencies and of sync pulses which is equal to that for the higher video frequencies. The frequency difference at a will be quite satisfactory if it is as great as 0.7 or 0.8 mc. The difference at b almost always will extend well out, to at least a megacycle and maybe more. This is not too objectionable.

The lower diagram of Fig. 87-8 illustrates a frequency response which would be desirable when there is a dual or split sound system. The principal difference is in the amount of gain at the sound intermediate frequency. At the output from the video detector this sound response must be zero or so near zero that the gain cannot be measured. As a general rule the response at the sound intermediate is brought down to zero by traps for accompanying sound and by the trap action of the sound takeoff. Other than this one feature, everything said about the intercarrier type of i-f response applies with a dual or split sound system.

ALIGNMENT PROCEDURE. If you have available the manufacturer's instructions for alignment by all means follow them to the letter. Whether or not you have such instructions at hand, the first step is to set up the frequency response on the oscilloscope screen, by following earlier instructions. Use the marker generator to check the frequencies at all important parts of the response curve, in accordance with Fig. 87-8. If you know the video intermediate frequency at which the receiver is supposed to operate this will be a great help. Otherwise you will have to assume that a frequency about half way up on the high frequency side of the response is the video intermediate or close to it.

If there has been poor low frequency response, and especially if there has been difficulty in holding synchronization, the video intermediate coming from the tuner probably is too low down on the i-f response. Then you may assume that the video intermediate to be used for alignment (the one from the tuner) is somewhat higher in frequency than the frequency half way up on the existing response. For example, the frequency at the half way point might be something like 25.8 mc. With the troubles just mentioned it might be well to try aligning for a video intermediate of 26.1 or 26.2 mc. It is possible to have snappy looking pictures and good synchronization with a video intermediate that comes up to 70 or even 75 per cent of the peak gain, although this may cause difficulties with sound reproduction.

If the amplifier is badly out of adjustment it is an excellent idea to make the preliminary alignment by using the vacuum tube voltmeter and fixed frequencies from a signal generator rather than swept frequencies. This requires knowing the frequencies at which the couplers should be peaked.

It should be mentioned that many an experienced technician has set up his own series of frequencies to obtain passable results – but this is hard on the next man who attempts an alignment without knowing

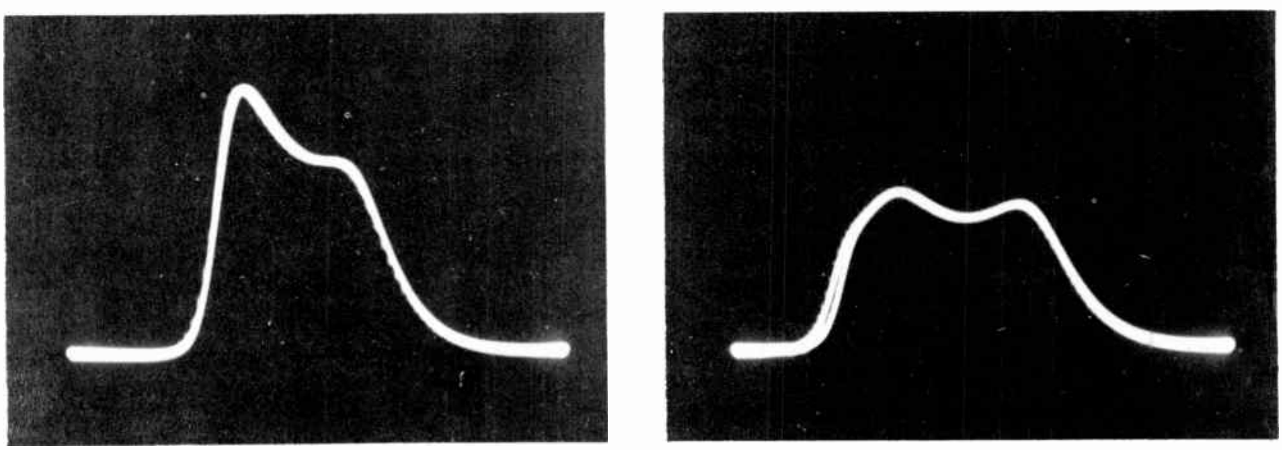
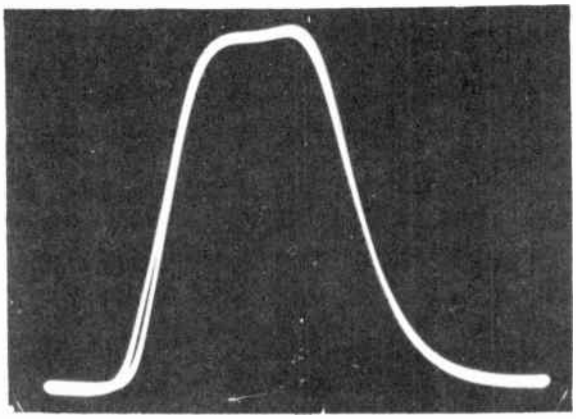


Fig. 87-9. Peaking and lack of band width (left) are corrected by alignment adjustments (right).



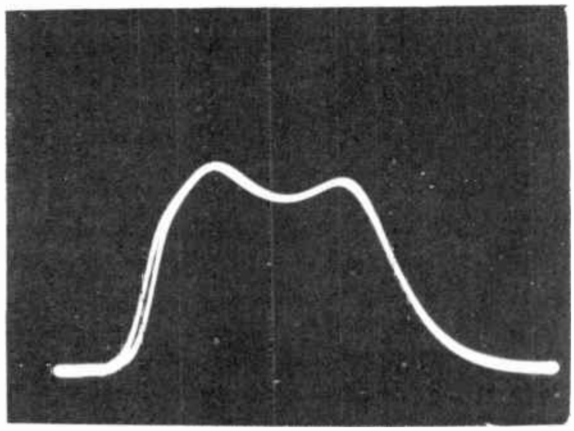


Fig. 87-10. High gain (left) must be sacrificed if sufficient band width (right) is to be obtained by alignment.

what has been done, and who attempts to use the correct frequencies for that receiver. Had there been a better set of frequencies for the couplers and tubes used in the set, they undoubtedly would have been used by its manufacturer.

Before commencing any adjustments it is desirable to determine which parts of the curve will be most affected by each coupler. This you can do by lightly touching the grids of each amplifier tube, or the leads connected to the grids. Whichever part of the curve drops the most will be dropped by increasing the inductance of the coupler preceding this grid or the one feeding into this grid. The same part of the curve will be raised by decreasing the coupler inductance.

If it is possible to reach the coils or coil forms of the couplers it is very easy to make more satisfactory preliminary checks by using one or more tuning wands. If the powdered iron end of a wand improves the shape of the response when this end is brought to or inserted a little ways into a coil form, making an adjustment for more inductance in that coil will do the same thing. If the non-magnetic end makes an improvement, reduce the inductance of the coupler.

With the receiver turned on its side to expose coupler coils which are underneath the chassis a tuning wand will stay in place when pushed a little ways into a coil form, to maintain a certain shape of the response at one place. Then a second wand may be tried in the other couplers to make a further improvement. Each coupler then is adjusted as its wand is removed. The response at the left in Fig. 87-9 has a bad peak at the low-frequency end and is so narrow as to cause an insufficient band pass with the video intermediate frequency half way up on the high-frequency side. Using the powdered iron end of a tuning wand, brings the response to the form at the right, with the peak dropped to where it is not harmful and with the band width increased by 0.8 mc on the low-frequency or sound end of the curve.

The object of many adjustments is to extend the response on the high-frequency side, to bring the video intermediate frequency where it belongs, or to extend the response on the low-frequency side to obtain more band width. At the same time you will wish to obtain, or retain all the peak gain that is possible.

No one adjustment will extend the frequency range while increasing the gain, it may do one or the other but not both.

The difference between the response curves of Fig. 87-10 was made with adjustment of one coupler. At the left there is high gain, but the band width is slightly less than 2.8 mc. At the right there is less gain, but the band width is almost 3.7 mc. When any one adjustment raises the response at either side it is practically certain to be narrowing the frequency band on the same side of the response, and when one adjustment extends the frequency it will lower the gain on the same side. To increase the band width and gain at the same time requires changing two or more adjustments.

If you obtain a response of the desired shape, but find that the entire curve is too low in frequency, it may be raised in frequency or moved horizontally with reference to frequency by slightly decreasing all the coupler inductances. Conversely, an entire response may be shifted to a lower frequency by slightly

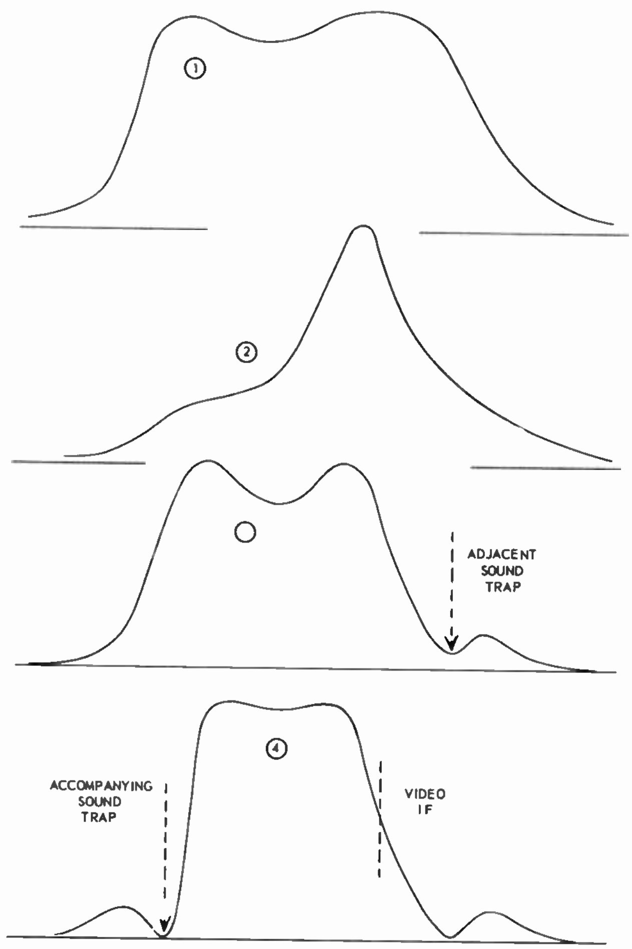


Fig. 87-11. The effects of traps on the shape of the final i-f response curve.

increasing all the inductances. It is difficult to do this without getting some parts of the response out of shape, but it is not too difficult to restore them.

No matter how good your alignment appears on the screen of the oscilloscope, never assume it to be really good until you have watched a transmitted picture or test pattern and listened to the accompanying sound with all the instruments and accessories disconnected from the receiver. After you have made a few alignments, the resulting pictures, patterns, and sound should be good, but your first attempts may not work out too well.

TRAP ALIGNMENT. Fig. 87-11 shows a series of frequency responses taken in the same general manner as those of Fig. 87-7, with the generators moved progressively back to the mixer grid while the oscilloscope remains on the video detector load resistor. The stages brought into the active circuit for curves 1 and 2 contribute peaking at certain frequencies. On the coupler brought in between curves 2 and 3 there is a trap tuned to the adjacent sound frequency, which is 1½ mc above the video intermediate frequency. This trap pits a decided dip in the curve.

On the coupler brought in between curves 3 and 4 is another trap tuned to the accompanying sound frequency, which places a second dip in the response. The effect of the adjacent sound trap persists, so that in the final frequency response for the entire i-f amplifier system there are dips at both ends of the curve.

Traps are aligned with the same instrument connections as for interstage couplers. The marker generator is tuned to the frequency of whatever trap is to be adjusted. The trap is then adjusted to reduce the response to zero or nearly so. The output of the marker generator must be increased to make the pip visible as long as possible. When the pip finally disappears as the trap is correctly adjusted, the tuning of the marker generator may be varied a little to each side of the trap frequency in making certain that the best adjustment has been made.

Traps intended for accompanying sound, adjacent sound, and adjacent video frequencies, may be used only for these particular frequencies, not for shaping the response in other ways. An accompanying sound trap usually will give a desirable steepness to the low-frequency side of the curve and will allow a wider band pass that could be had without the trap. As mentioned earlier, some traps are for the express purpose of shaping the response curve. As a rule these curve-shaping traps are intended to bring down the height of a peak which otherwise would be formed somewhere along the response with all couplers correctly aligned.

OVERALL CHECK. After you complete the alignment of the i-f amplifier system it is advisable to make an overall check from the antenna terminals through the tuner to the video detector. The connections are shown by Fig. 87-12, and explained as follows.

Sweep generator. With the antenna or transmission line disconnected, connect the high-side and ground leads for the swept frequency to the two antenna terminals of the receiver. Some sweep generators are provided with an output cable having on its free end a matching pad which matches the internal impedance of the generator to the 300-ohm internal or input impedance of the tuner. If there is no such pad, connect in series with each lead a fixed carbon resistor of 100 to 150 ohms, which ordinarily will make an acceptable match. Later adjust the center frequency and the sweep width to bring onto the oscilloscope the com-

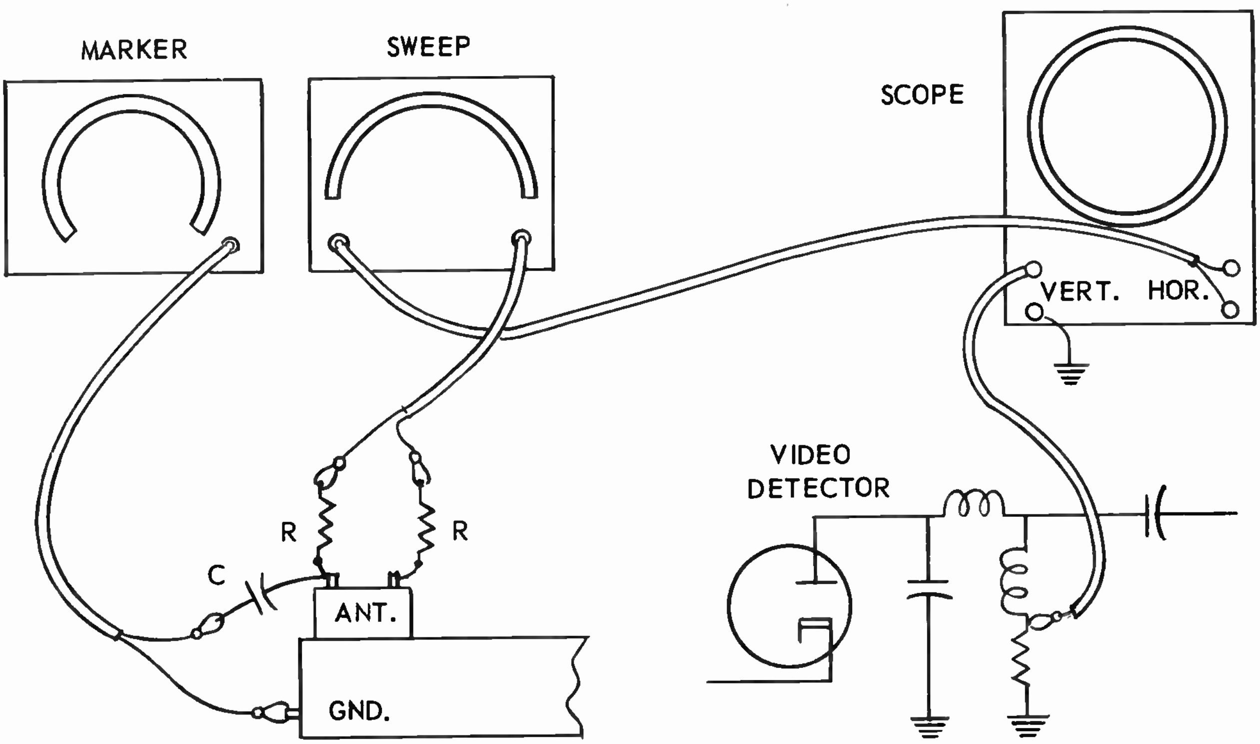


Fig. 87-12. Instrument connections for making an overall check from antenna terminals to video detector output.

plete response curve for each channel as it is checked.

Marker generator. Connect the high-side lead from the marker to either of the antenna terminals through a small fixed capacitor. This capacitance usually should be no more than 2 to 5 mmf. Connect the ground lead to chassis ground or B-minus.

Oscilloscope. Connect the vertical input of the scope to the video detector load, in the usual manner. Connect the horizontal input of the scope to the horizontal deflection voltage or to the synchronized sweep output of the sweep generator.

Fine tuning control. If the receiver has a fine tuning control set it to about mid-position or wherever it usually is set for normal reception. Do not again change this control during following observations.

Channel selector. Set the channel selector of the receiver to the channel which is to be checked first. Make sure that the center frequency of the sweep generator is set for the same channel, which usually will be done after everything is operating. Everytime you change the channel selector to the next channel, the center frequency of the sweep will have to be changed accordingly.

Turn on the receiver and all the instruments, and let them warm up for 15 minutes or more. Then adjust the sweep generator to bring a response curve onto the screen of the scope, and try out the marker generator to obtain a pip on the trace.

The frequency responses for the various channels will not have just the same shape as the frequency response of the i-f amplifier alone, because of the effect of circuits in the tuner. Fig. 87-13 shows a set of responses for twelve channels as observed on a receiver which was giving good performance on all the low-band channels except number 6, very poor performance on channel 7, fairly good on numbers 9 and 11, and entirely satisfactory on the other high-band channels. Of course, not all the channels were allocated. It happened that high-band channels 7, 9, and 11 were in use, and examination of the tuner showed that it has been misadjusted on these channels in an apparent attempt to obtain more gain without preserving the band width.

The dots on the response curves show locations of the video carrier frequencies and of the video intermediate frequencies during actual reception and during shop tests on the receiver. You will see that these frequencies are too far down on the channels giving trouble, and that the band widths are too narrow. The band width is narrow also on channel 6.

During an overall check some technicians introduce the marker frequency at the mixer grid, just as during an i-f alignment, instead of at the antenna terminals. If your marker generator will not produce accurate frequencies in the carrier ranges you may use marker pips at intermediate frequencies with the marker generator connected to the mixer grid. The pip will be moved back and forth across the trace by tuning of the marker generator, just as during i-f alignment. Thus the marker may be used for checking the position of video and sound intermediate frequencies, and other frequencies, also for checking band width while the swept frequencies are introduced at the antenna terminals.

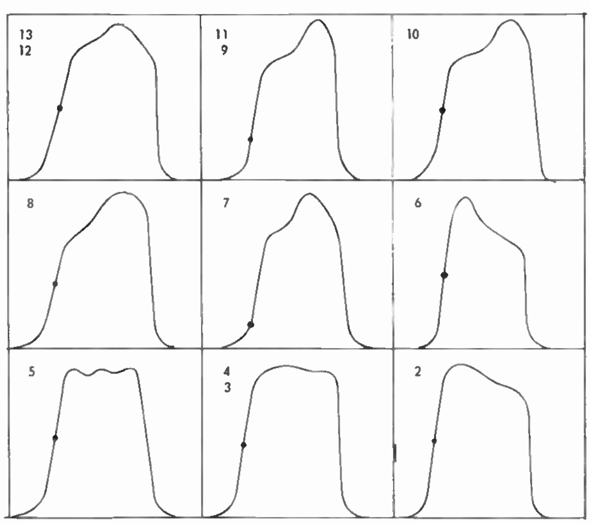


Fig. 87-13. The overall frequency response will not be the same on all channels, nor will it have exactly the same shape as the i-f response.

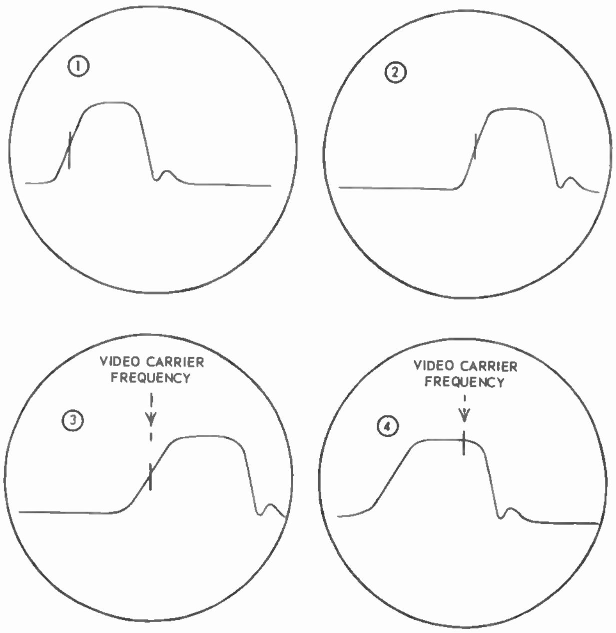


Fig. 87-14. Fine tuning adjustment will shift the response horizontally along the swept frequencies, and may or may not shift the marker pips.

When using an intermediate frequency into the mixer grid for marker pips the effect of varying the fine tuning control will be as shown at the top of Fig. 87-14. In diagram 1 the pip is correctly placed about half way down on the high-frequency side of the response. Varying the fine tuning control will move the response curve bodily to the right or left. In diagram 2 the response has been moved to the right. The position of the video intermediate pip on the response is the same as before. The marker pip and the response curve move together, and equally.

The lower diagrams illustrate the effects of varying the fine tuning control when the marker frequencies are introduced at the antenna terminals, as in Fig. 87-12, and when video carrier, sound carrier, and other carrier range frequencies are used for markers. In diagram 3 the fine tuning control is adjusted to bring the video carrier frequency about half way down on the high-frequency side of the response. In diagram 4 the fine tuning control has been readjusted to move the entire response curve to the left or to lower frequencies. But the video carrier frequency remains in its original position in relation to the total width of sweep. Adjustment of the fine tuning control has brought the video carrier frequency onto the top of the response curve.

Diagrams 3 and 4 show actual effects of fine tuning adjustment. The purpose of this control, or one of its chief purpose, is to bring the various transmitted frequencies to their correct positions on the response. Adjustment of the fine tuning has the effect of moving the video carrier and video intermediate frequencies higher or lower on the slope of the response. Diagrams 1 and 2 do not correctly show the action of the fine tuning control.

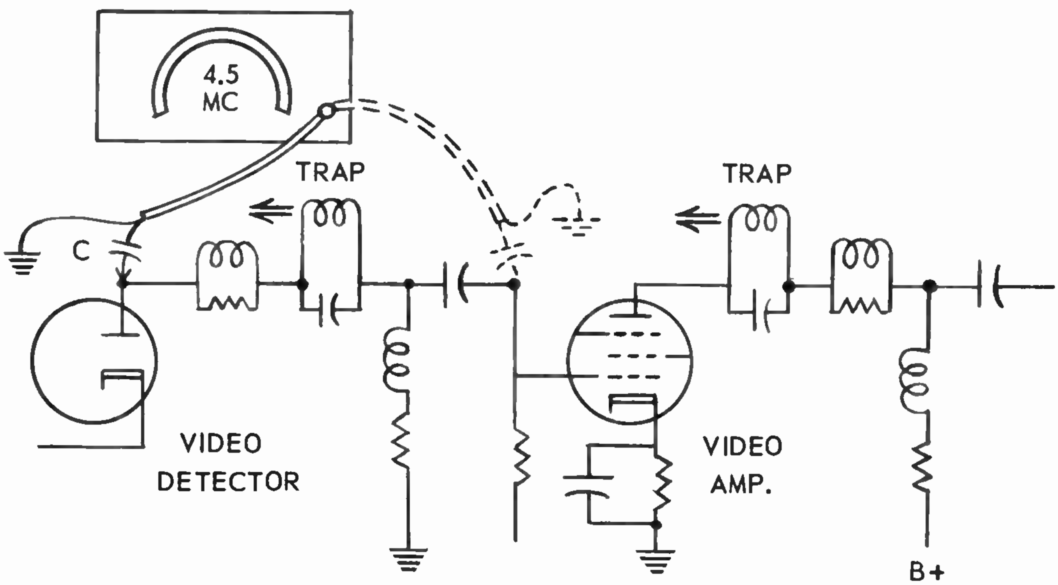


Fig. 87-15. Connections for adjusting a trap for the intercarrier beat frequency.

While observing the overall response from antenna to video detector, with swept frequencies applied to the antenna terminals, make no attempt to adjust the i-f couplers or traps for improvement of a response. Such adjustments on one channel would throw off the responses in other channels, would cause weak pictures, loss of sound, loss of synchronization, and possible regeneration or oscillation on other channels. Align the i-f amplifier stages only while swept frequencies are applied to the mixer or to one of the following amplifiers. Poor frequency responses on one or more channels during the overall check calls for alignment of the tuner circuits, which we shall consider a little later on.

INTERCARRIER TRAP ADJUSTMENT. In receivers which do not employ intercarrier sound there often is a trap for the intercarrier beat frequency of 4.5 mc located somewhere between the output of the video detector and the grid or cathode of the picture tube, whichever is used for signal input.

This trap may be adjusted or aligned with the oscilloscope vertical input connected to the signal input element of the picture tube and with the marker generator or other single frequency signal generator connected as shown by Fig. 87-15. If the intercarrier beat trap is between the video detector and the first video amplifier tube, the signal generator is connected through a capacitor to the output element, or terminal of the detector, which is shown as the plate in the diagram. If the trap follows a video amplifier tube the generator is connected to the grid of the amplifier immediately preceding the trap, as shown by broken lines in the diagram.

The signal generator is used with audio-frequency modulation, and the scope is used with its internal sweep adjusted to the frequency of audio modulation at the signal generator. The generator must be tuned very accurately to 4.5 mc. The trap then is adjusted for minimum height of the audio modulation trace on

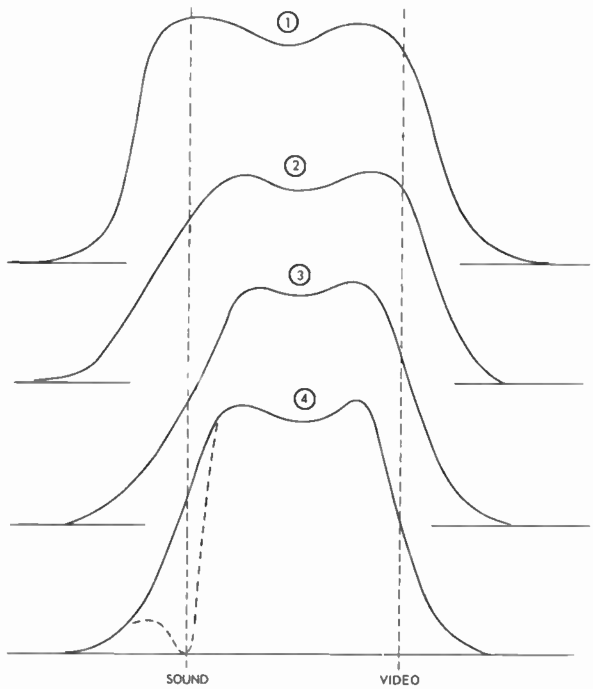


Fig. 87-16. Successive response curves for an i-f amplifier employing overcoupled interstage transformers.

the screen of the oscilloscope. This method usually is not as satisfactory nor so easy to carry out as another which we shall employ after learning to use a detector probe with the vacuum tube voltmeter.

OVERCOUPLED INTERSTAGE TRANSFORMERS. There are a few receivers in which part or all of the i-f stages are coupled by means of transformers having separate primary and secondary windings, each with its own tuning slug or trimmer capacitor. Most often these transformers are overcoupled. As you will recall, overcoupling causes the resonance curve to have two separate peaks rather than a single peak. The closer the coupling the farther apart the resonant peaks are moved. The peaks come together when the coupling is loose enough to be classed as undercoupled. All this takes place with both windings tuned to the same frequency.

Fig. 87-16 shows a set of alignment curves for an i-f amplifier in which all the transformers are overcoupled. These response curves are made like the ones of Figs. 87-7 and 87-11, by moving the connection of the sweep generator progressively back from the grid of the amplifier immediately preceding the video detector to the grid of the mixer tube. The double peaked response at 1 is that for the last i-f amplifier and the transformer which is between this amplifier and the video detector. The response at 2 is that from the two amplifiers preceding the detector, and the transformers which follow them. The response at 3 is with a third stage brought in, and the one at 4 is for the entire amplifier system from mixer grid to video detector output.

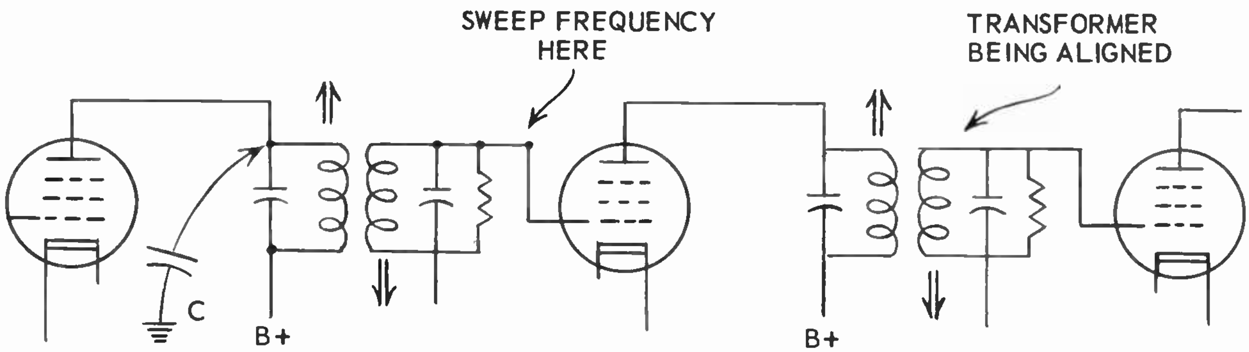


Fig. 87-17. Shunting a transformer primary while aligning an i-f amplifier having overcoupled transformers.

The positions of the sound and video intermediate frequencies on each of the responses are shown by the vertical lines. As more and more amplifier stages become active the band pass becomes narrower and the curves tend to have somewhat steeper sides. In the final response, number 4, the gain at the sound intermediate frequency has to be dropped by using one or more traps for accompanying sound, making the actual response at the low-frequency end as shown by the broken line portion of the curve rather than by the full line which would represent the response without traps.

I-f amplifiers having overcoupled transformers are aligned stage by stage with the help of response curves from manufacturers' instruction manuals, commencing with the sweep generator to the grid of the last i-f amplifier while aligning the last transformer, then working back to the mixer while making alignments to obtain successive curves such as the set shown by Fig. 87-16. A wide sweep will be needed when commencing the alignment.

The marker generator may be connected through a small capacitor to the same grid as the sweep, and moved back with the sweep, or the marker may be connected to the mixer grid and left there during the entire process provided the pips show up clearly enough. The marker generator is used without modulation.

The oscilloscope is connected in the usual manner to the video detector load resistor, and remains there during the entire job.

Should the responses show dips or peaks where there should be none, and if these are not readily removed by alignment adjustments, it will be necessary to shunt one of the transformer primaries to ground through a capacitor of 100 to 1,000 mmf as shown by Fig. 87-17 at C. This shunting capacitor is connected to the high side of the primary or plate winding of the transformer preceding the one to be aligned, or the transformer to whose secondary is connected to the sweep generator. This removes any effects of having this primary winding resonant at a frequency within the response range. The shunting capacitor is moved back stage by stage when the sweep generator connection is moved back. It is not needed on the grid of the mixer tube.

Each overcoupled transformer ordinarily has two adjustments, one for the primary and the other for the secondary winding. One adjustment may have its principal effect on the high-frequency peak and slope of the response, while the other adjuster affects the opposite side of the curve. One method of alignment consists of turning one adjuster all the way out while setting the other one to give the approximate desired form to one side of the response. Then the other adjuster is used to shape the other side. With another method a fixed resistor of 300 to 500 ohms is connected across the primary while the secondary winding is adjusted for peak response. Then this resistor is moved to the secondary while the primary is adjusted for a peak response. Finally the resistor is removed entirely and the two windings are slightly readjusted if necessary to obtain the desired shape of the curve.

TELEVISION

LESSON NO. 88

ALIGNMENT OF SOUND SYSTEMS

"Training Your Mind"

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind

C.T.I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON NO. 88

ALIGNMENT OF SOUND SYSTEMS

After the i-f amplifier has been correctly aligned we usually work next on the sound system rather than the tuner. This is because the tuner must furnish signals at intermediate frequencies for both video and sound. Were either the video system or the sound system incorrectly adjusted we might easily misalign the tuner in attempting to correct for faults in either of the other systems. But when both sound and video are correctly adjusted it will be safe to go ahead with alignment of the tuner.

Most of the individual steps in alignment of sound systems are carried out in the same general manner no matter what may be the design. There are, however, some rather important differences between the procedures for alignment of intercarrier sound systems and for dual or split sound systems.

In diagram *1* of Fig. 88-1 the rectangular blocks indicate the parts of a typical intercarrier sound system which are adjusted during alignment. First comes the i-f amplifier extending between the mixer and the video detector. For intercarrier sound we require in this i-f amplifier a frequency response which is somewhat different than for dual sound. This i-f response would, of course, be obtained during the preceding alignment of the i-f couplers, since you will know that the receiver uses intercarrier sound.

Next comes the sound takeoff, which usually follows a video amplifier tube but which may follow the video detector. The takeoff ordinarily feeds into a driver tube. Then comes the demodulator transformer, the demodulator tube, and the audio amplifier. Should the sound takeoff follow the video detector there usually is an additional sound i-f amplifying stage between takeoff and driver tube. In this stage will be a coupling which requires alignment.

Diagram *2* of Fig. 88-1 shows in the rectangular blocks the parts which require adjustment during alignment of a typical dual sound system. The sound takeoff may come immediately after the mixer or it may be on the output of any of the video I.F. amplifier tubes. The takeoff feeds into the first sound i-f amplifier tube. There may be one, two, three, or even four sound i-f amplifying stages. The last stage or possibly the last two stages may operate as limiters. In each amplifying stage is a coupling transformer which requires alignment. After the limiter comes the demodulator transformer, then the demodulator tube, and finally the audio amplifier.

The sound demodulator may be either a ratio detector or a discriminator. All the processes of alignment are the same for one as for the other. Although we shall be talking for the most part about the alignment of television sound systems, which operate with frequency modulation, the same general methods will apply to the i-f amplifiers and the demodulators of f-m sound receivers and combination am-fm sound receivers.

Alignment of any sound system may be carried out with either of two groups of testing instruments. All the work may be done with a constant frequency signal generator as the source and with a vacuum tube volt-

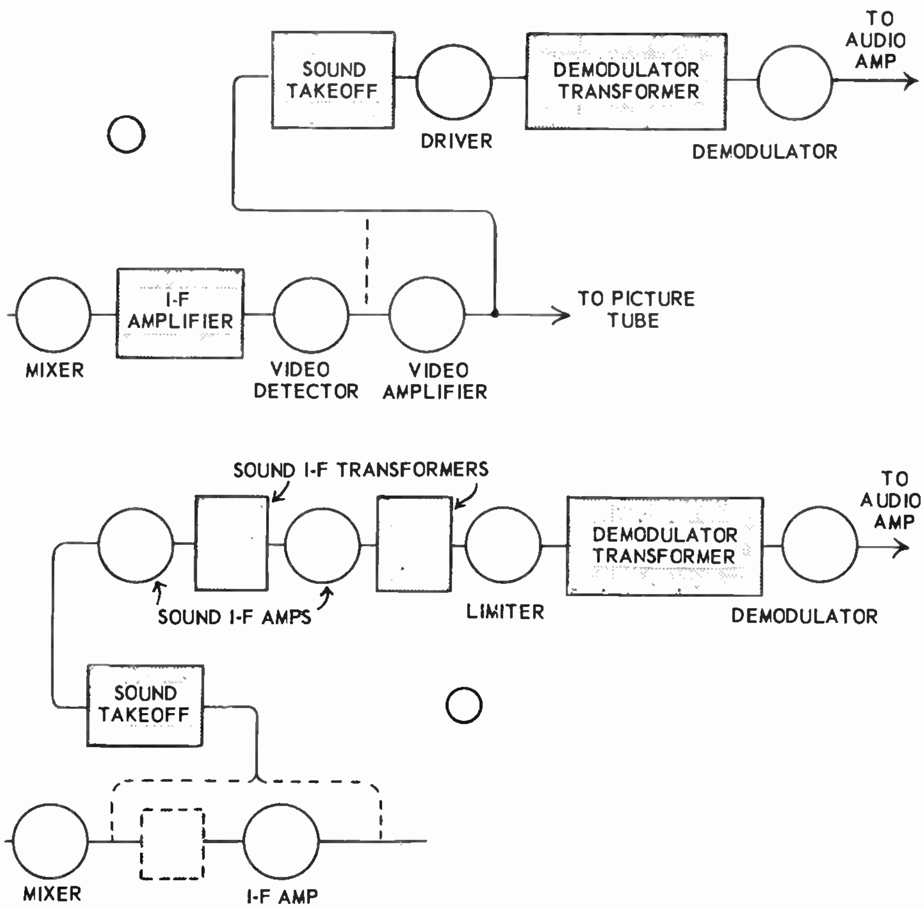


Fig. 88-1. The parts shown shaded are aligned with an intercarrier sound system (top) and with a dual sound system (bottom).

meter as the output indicator. Or all the work may be done with a sweep generator and marker generator as signal sources, and with an oscilloscope as output indicator. We shall deal first with oscilloscope methods, because this will make it easier to see what you are trying to accomplish when working with the VTVM.

INTERCARRIER SOUND. Regardless of the video and sound intermediate frequencies with which a receiver may operate, the center frequency for an intercarrier sound system always is exactly 4.5 mc. This applies to the takeoff coupler, the demodulator transformer, and any sound i-f couplers which may be used. The frequency must remain 4.5 mc because it is the difference between video and sound intermediates, and between video and sound carriers.

For alignment with either the VTVM or oscilloscope it is essential that the signal generator or marker generator be able to deliver a 4.5 mc signal with great precision. The total plus and minus frequency deviation for television sound is 50 kc or 0.050 mc. This total deviation is 1/90 or about 1.1% of 4.5 mc. If your signal is accurate only to within plus or minus 1% it could fall 45 kc above or below the true center of the sound deviation, and cause extremely bad misalignment. To remain with even 5 kc of the received center frequency for sound the test signal must be accurate within 0.1%. The practical solution is to use a signal from a 4.5 mc crystal (accurate to 0.02%) or one from a generator which has been calibrated from a crystal.

Now about the special type of frequency response required in the i-f amplifier when there is intercarrier sound. First we may look at the wrong kind of response, at *1* in Fig. 88-2. The sound intermediate frequency falls on a point in the response curve where there is considerable slope. Although frequency deviation for the sound signals extends over only 1/90 of the total distance between the intermediate frequencies, this deviation still is along the same slope. This is shown at *2* by a greatly enlarged portion of the response curve at the point of sound signals.

As the deviation goes to a lower frequency it extends to a point on the frequency response where the gain is much less, relatively, than when going to a higher frequency. This variation of gain or signal voltage which occurs with frequency modulation really is a variation of signal amplitude, and we have introduced a strong amplitude-modulated signal accompanying the frequency-modulated signal. Amplitude modulation so strong as shown could not be removed by a limiter nor by limiting action of a ratio detector, and would ruin the reproduction of f-m sound signals.

A desirable form of i-f amplifier frequency response is shown at *3*. The sound intermediate frequency falls on a part of the curve that is relatively flat. As a consequency there is very little variation of gain and very little amplitude modulation accompanying the frequency deviation, as shown at *4*.

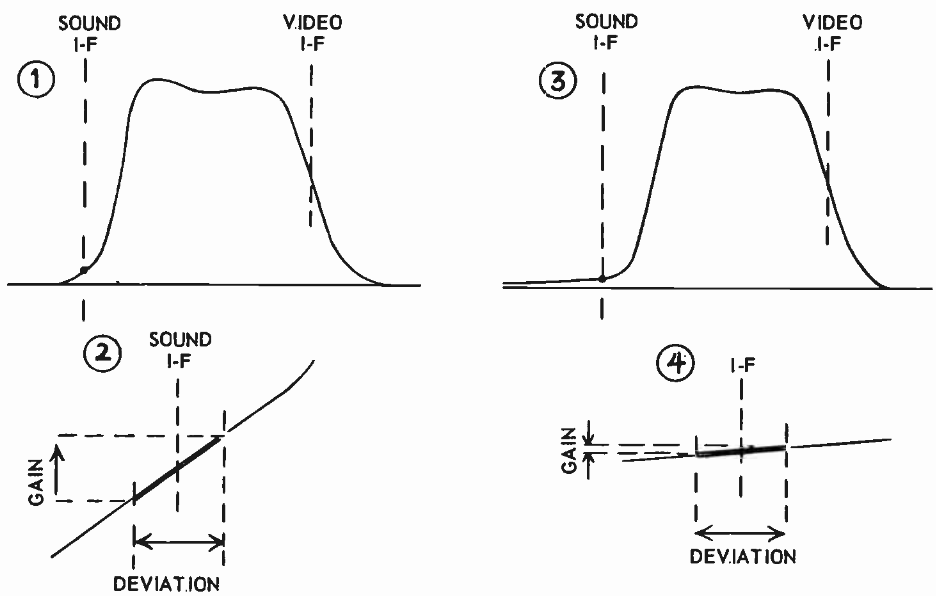


Fig. 88-2. With an intercarrier sound system the sound intermediate frequency should not be on a steeply sloped portion of the response curve.

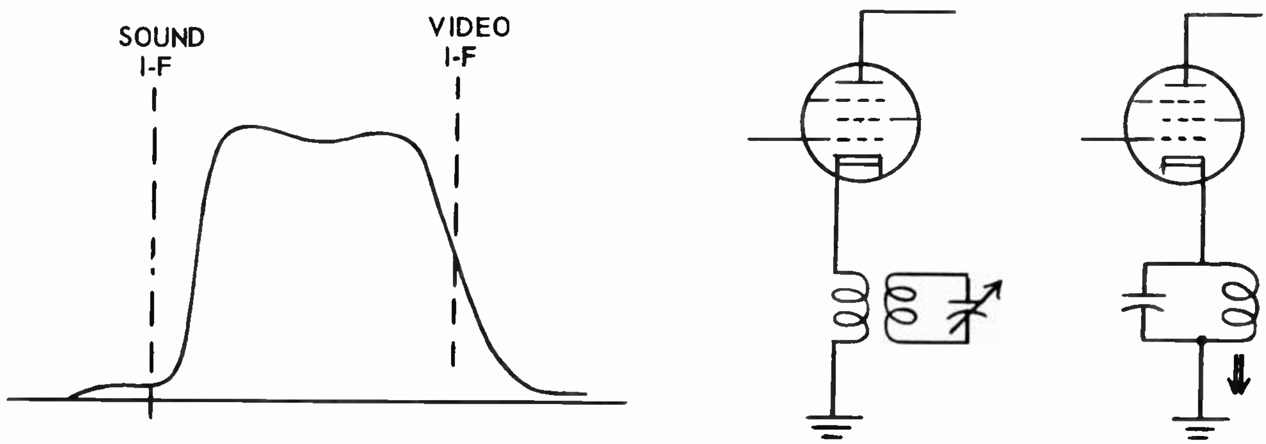


Fig. 88-3. A shelf or plateau on the video i-f response when there is an intercarrier sound system.

In a number of receivers the i-f response is flattened at and near the sound intermediate frequency by using a trap on some one of the i-f amplifiers. The result, illustrated at the left in Fig. 88-3, is a sound shelf or sound plateau on the response. The trap most often is in the cathode lead for the first, second, or third i-f amplifier, as shown at the right. Traps used for this purpose are of rather low-Q construction in order to have their effect extend over a fairly wide range of frequencies without dropping the response all the way to zero at any frequency. The trap is tuned to the sound intermediate frequency, or a little above or below this frequency until results are as desired.

Under certain conditions an intercarrier sound system may produce a type of noise called intercarrier buzz. It is not in the nature of hum, but is a sharp and clear cut, rather rasping noise at a low audio frequency. The three most common causes are incorrect setting of a fine tuning control, wrong alignment of the r-f oscillator for the channel where trouble occurs, and excessively high setting of the contrast control. When the sound takeoff follows a video amplifier the buzz may result from overloading of this amplifier due to low plate or screen voltage, wrong grid bias, or a worn out tube.

Intercarrier buzz may occur when the picture becomes nearly all white, during which time the signal voltage applied to the picture tube grid rises to maximum. The cause also may be over-modulation at the transmitter, or too much frequency deviation. Buzz may occur during momentary drops of carrier strength, when an agc system acts to increase amplification just as does an excessively high setting of the contrast control.

Incorrect adjustment of the secondary in the demodulator transformer will allow buzz and other noise whether the adjustment is out of the way in one direction or the other. Should the i-f amplifier system be aligned to bring the sound intermediate too high or onto a sloped portion of the frequency response there will be buzz. Another cause is any failure of a-m limiting action, as from a fault in the limiter stage preceding a discriminator, or, with a ratio detector, an open or defective large capacitor on the load side of the tube.

ALIGNMENT PRELIMINARIES. The steps which should precede alignment of a sound system are practically the same as those for video i-f alignment. The antenna should be disconnected, and it may be well to short circuit the antenna terminals on each other or to ground. The channel selector should be set to the highest channel in which there is no transmission for your locality.

Ordinarily there is no automatic volume control for a television sound system, although this type of control is common in f-m and in combination sound receivers. When there is automatic volume control it should be overridden in the same manner as automatic gain control for video i-f alignment. Since all signal voltages will be supplied by the test instruments it is desirable to disable the r-f oscillator during the alignment processes, then allow this oscillator to operate during a final check with a received signal.

SOUND I-F ALIGNMENT. As mentioned before, one or more sound i-f amplifier stages always are found with a dual sound system, and sometimes there is such a stage with intercarrier sound. The center frequency for a dual sound system always is the sound intermediate frequency for the particular receiver, while the center frequency for intercarrier sound always is 4.5 mc.

Frequency responses such as may be obtained from sound i-f amplifier systems are illustrated by Fig. 88-4. These responses are taken with the input signal applied to the grid of the tube ahead of the sound takeoff, and show the output of the entire i-f amplifier system which precedes the demodulator. The top of the response may be rounded as at *1* or double peaked as at *2*. It is not necessary to have a perfectly flat top extending through the range of deviation frequencies. Peaks or dips represent unequal gains. This produces some amplitude modulation. But unless there is severe peaking or sloping, the amount of amplitude modulation thus produced can be removed by limiting action.

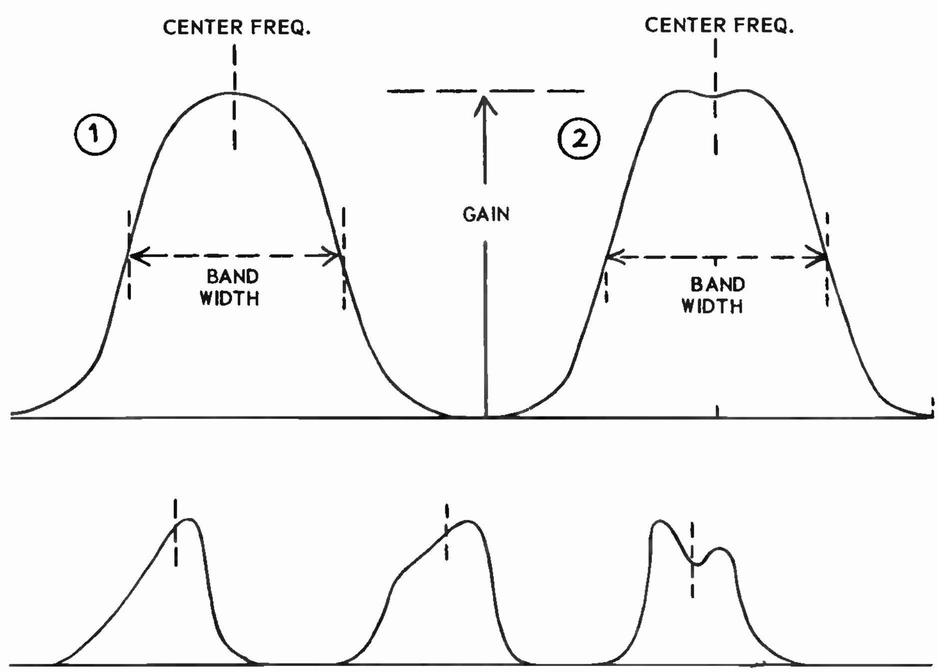


Fig. 88-4. At *1* and *2* are desirable responses for a sound i-f system. Some undesirable responses are shown below.

The band width between points about half way down on opposite sides of the response should be no less than about 200 kc or 0.2 mc under any conditions, and usually is made 300 to 400 kc or 0.3 to 0.4 mc. The center frequency must be at the center of the band width or very nearly there. Opposite sides of the response curve should be of nearly the same shape, or symmetrical. Double peaks should be of equal or nearly equal heights. When these requirements are satisfied, the greater the gain or the greater the height of the peak or peaks the better will be the performance. Several undesirable responses are shown at the bottom of the figure.

Instrument connections for alignment of a sound i-f system are illustrated by Fig. 88-5. The sweep generator and marker generator are coupled through small capacitors to the grid of whatever tube precedes the sound takeoff. This tube might be the mixer or it might be any of the video i-f amplifier tubes. The vertical input of the oscilloscope is connected to the high side of the limiter grid resistor. The limiter will be the tube which is just ahead of and which feeds into the demodulator transformer. The limiter grid resistor most often is on the low side of the coupler between the last sound i-f amplifier and the limiter, as shown in the diagram. No matter where the grid resistor is connected, the oscilloscope goes to its high side. If the low side of the grid resistor goes to ground, the oscilloscope input is connected to ground. If the low side of the grid resistor does not go to ground, but possibly to B-minus or to a bias voltage source, connect the ground terminal of the scope to the low side of the grid resistor, not to chassis ground.

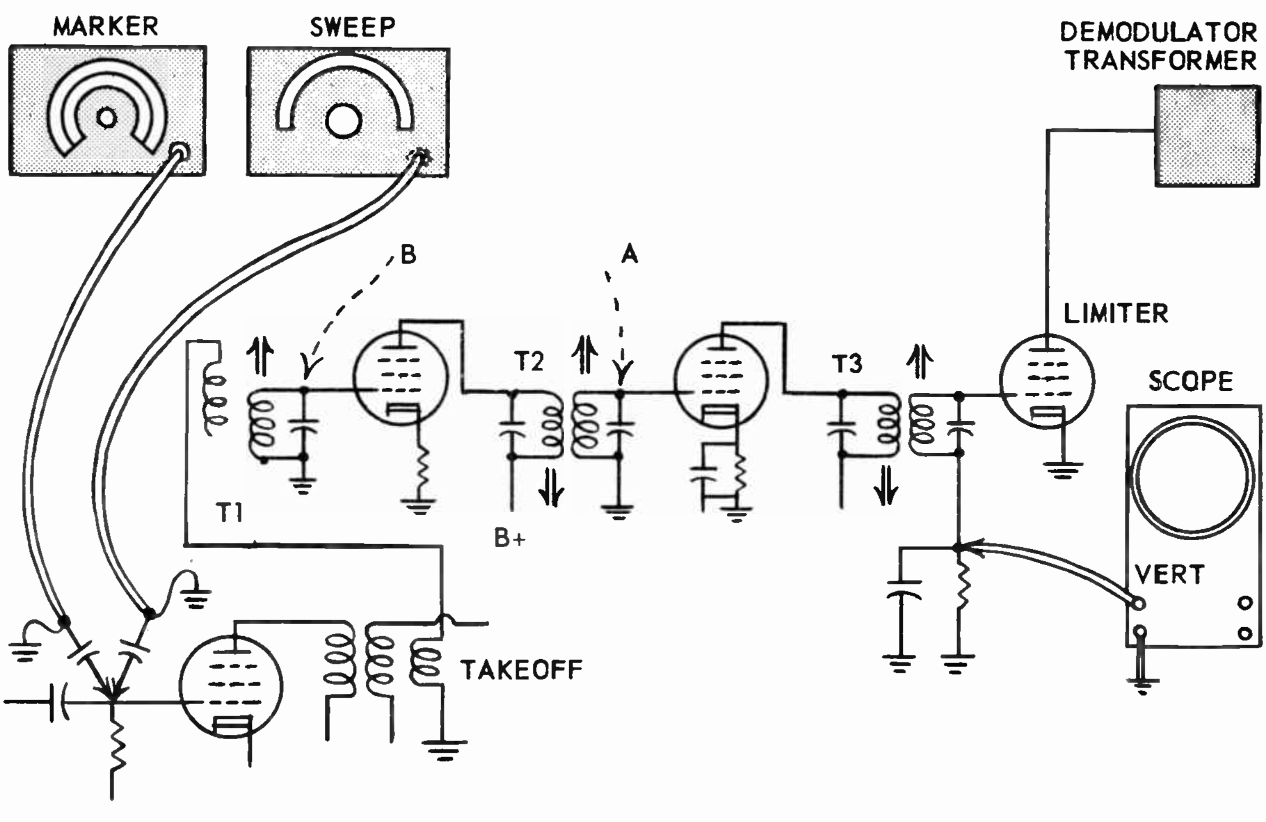


Fig. 88-5. Instrument connections for alignment of a sound takeoff and i-f amplifier system.

Unless the oscilloscope has vertical input impedance of at least one megohm, connect in series with the vertical input lead a fixed resistor of one-half to one megohm. This is necessary in order to avoid excessive alteration of limiter grid bias.

If there is originally only slight misalignment it now is possible to adjust the sound takeoff and the interstage couplers or transformers to obtain a frequency response such as shown by Fig. 88-4. It will be necessary to use a sweep width of one to three megacycles. The marker generator is used to identify the position of the center frequency on the response. The marker pip is also moved half way down on opposite sides of the response, and the corresponding frequencies are noted to check for required band width.

It is of greatest importance to keep the output of the sweep generator at the lowest value which allows a good response trace while the vertical gain of the oscilloscope is at or near its maximum setting. If the output control of the sweep won't bring the output voltage low enough, use a smaller series capacitor on the high-side lead. If this is not done there will be limiting action and the trace will not show the true frequency response. It is necessary also to keep the marker output low enough to cause little or no distortion of the response curve when the marker pip is moved across the response curve. This may require a series capacitor, on the marker lead, of only two or three mmf. As the response curve becomes higher during alignment, keep it on the scope screen by reducing the output of the sweep generator, do not reduce the vertical gain of the scope.

The takeoff or the interstage couplers may be so far out of adjustment that no usable response trace can be obtained with the generators coupled ahead of the takeoff. Then it is necessary to make the alignment stage by stage. Leave the oscilloscope on the limiter grid resistor, but move the connections of the generators to the grid of the tube preceding the last interstage coupler or transformer. In Fig. 88-5 this would be at *A*. Now proceed to align the last coupler, which would be transformer *T3* on the diagram. Since there will be only one stage of amplification between generators and scope the sweep output will have to be fairly high. Band width will be greater than for the entire sound i-f amplifier system.

Next move the generator lead connections back to the next preceding grid, point *B* on the diagram, and align coupler *T2*. Keep moving the generator leads back to the grids of preceding tubes, while aligning the coupler following each tube, until the generators are coupled to the grid of the tube ahead of the sound takeoff. Keep the trace on the screen of the oscilloscope by reducing the sweep output as you proceed, not by reducing the gain of the scope.

There is a possibility that the response trace may develope fuzzy peaks of excessive height or there may be a fuzzy horizontal streak. This indicates oscillation in the amplifiers. Quite likely the sweep output is too great, or the receiver contrast control is set too high, or some of the preliminary precautions have been neglected. In extreme cases of oscillation you may temporarily remove the amplifier tube which precedes the one to whose grid the generators are connected, or you may connect the grid of this preceding tube to ground through a capacitor of about 1,000 mmf.

DEMODULATOR REQUIREMENTS. The demodulator system consists of the demodulator transformer, the twin diode demodulator tube, and the load circuit which feeds the audio output. The audio output voltage from the demodulator system depends in the first place on the signal voltage applied to the demodulator transformer from the last i-f amplifier, the limiter, or the driver tube.

Relations between f-m input voltage and audio output voltage are shown by Fig. 88-6. We shall assume that the frequency response of the amplifier or driver stages are as shown at *1*. With a signal coming originally from the sweep generator this will be the voltage applied to the primary of the demodulator transformer as frequency sweeps below and above the center value. As frequency decreases there is conduction

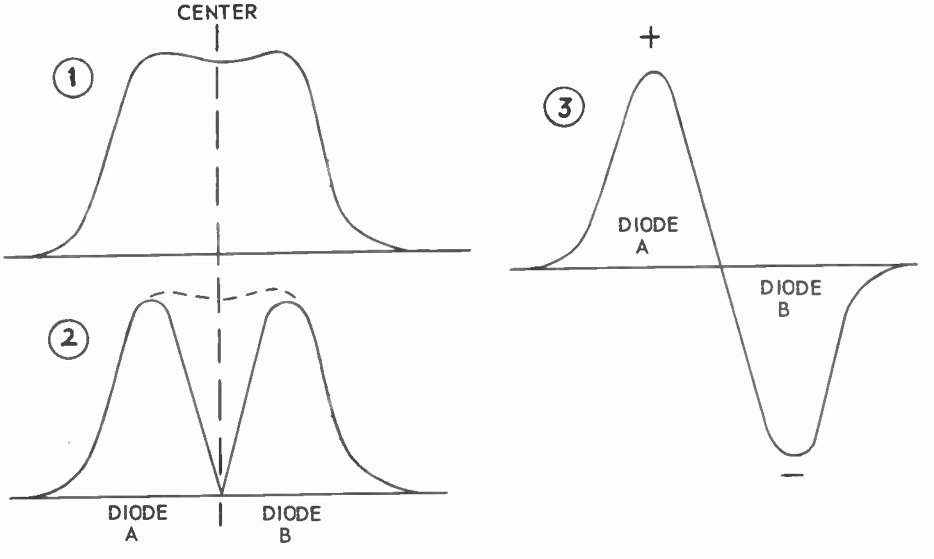
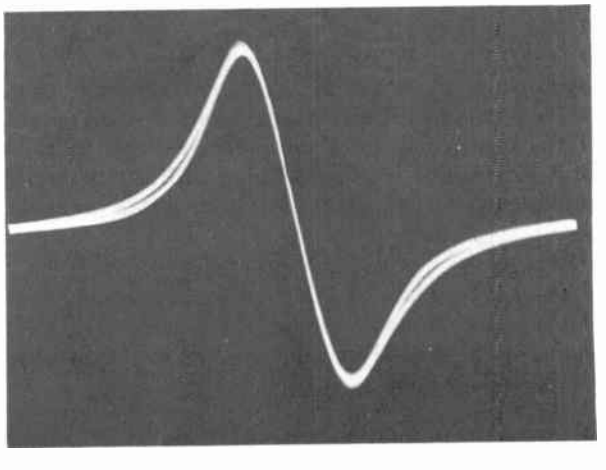


Fig. 88-6. The S-curve results from demodulator diode conductions in opposite polarities.

in one of the diodes of the demodulator, and as frequency increases there is conduction in the other diode. The diode conductions are represented at 2. The demodulator circuits are so arranged that conduction in one diode causes the audio output voltage to go positive, while conduction in the other diode causes the audio output to go negative. To show this correctly the conduction of one diode must be inverted, as at 3.



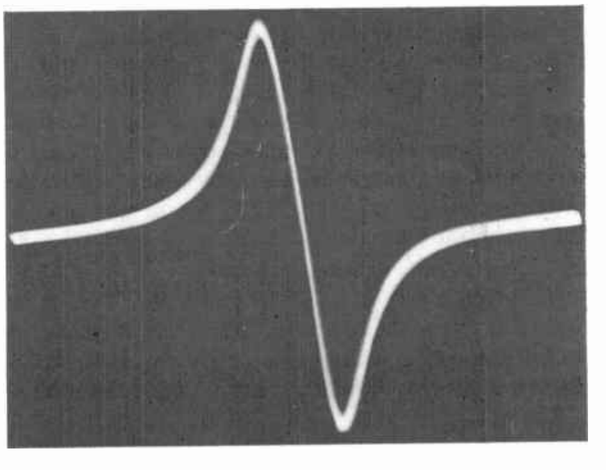


Fig. 88-7. The width of the S-curve is affected by the band pass of preceding amplifier stages and of the sound takeoff.

This is the form of audio output voltage produced when swept frequencies are applied to the tube ahead of the demodulator transformer.

All that we need do to actually watch the audio output voltage during alignment of the demodulator is connect the oscilloscope to the audio output point of the demodulator. Resulting traces will appear as in Fig. 88-7. The trace at the left is quite similar to diagram *3* of Fig. 88-6.

Supposing now that the band width of the frequency response from stages ahead of the demodulator has been made too narrow. Then, with the same frequency sweep, the audio output will be as shown at the right in Fig. 88-7. The positive and negative audio peaks cannot be so far apart in frequency because the narrow band width of preceding stages drops the gain to zero before there is a very great deviation in either direction.

Traces such as shown by Fig. 88-7 are called demodulator S-curves. When the characteristics of the S-curve meet certain requirements the demodulator has been correctly aligned, or, it has been aligned as well as possible when considering the frequency response of parts ahead of the demodulator.

What we wish to obtain in the S-curve is illustrated by Fig. 88-8 and explained as follows.

1. The center frequency is to be midway between the positive and negative peaks.

2. The straight or linear portion of the trace should extend as far as possible above and below the center frequency. Since maximum deviation for television sound is plus or minus 25 kc the straight portion must go this far above and below the center frequency if there is to be undistorted audio output. It would not be safe to depend on such close limits, so we ordinarily specify a minumum of about 75

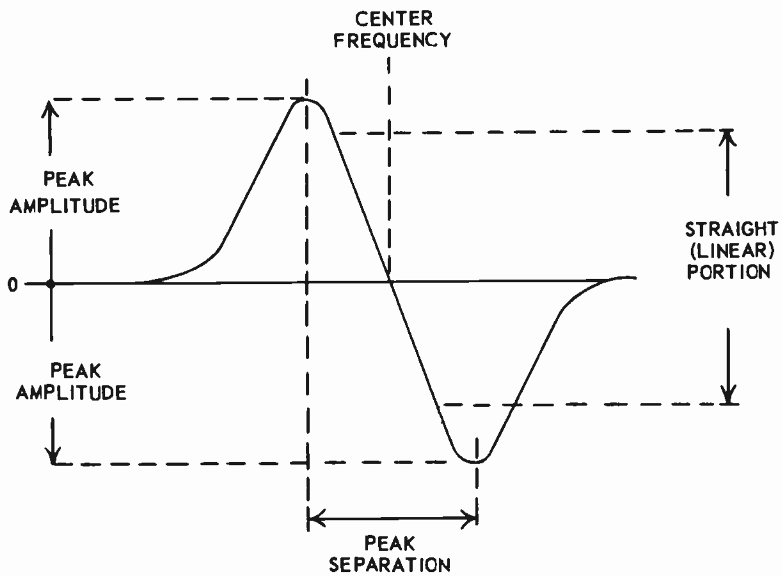


Fig. 88-8. Features of the demodulator S-curve which are to be checked during alignment.

kc on each side of the center. For an f-m sound receiver, with its greater maximum deviation, the straight portion of the S-curve should go to at least 150 kc each way from the center.

3. The positive and negative peaks are to have equal or very nearly equal amplitudes. Equal amplitudes, while satisfying other requirements, are possible when the two diodes of the demodulator have equal conductions, and when the output from preceding stages has a satisfactory form of frequency response. The two peaks represent conductions and resulting voltages from the two diodes.

4. The positive and negative sides of the S-curve should be of practically the same shape, or should be symetrical.

5. The peak amplitudes should be as great as possible. Maximum peak amplitudes will be possible only when primary and secondary of the demodulator transformer are correctly adjusted and when all other couplers as far back as the sound takeoff have been correctly aligned.

6. It is desirable to have fairly wide frequency separation between positive and negative peaks. Satisfactory separations in various television receivers run all the way from somewhat less than 150 kc to more than 600 kc. The possible peak separation depends on the frequency responses of parts of the demodulator system, as explained in connection with Fig. 88-6.

ALIGNMENT OF DEMODULATOR. Instrument connections for alignment of the demodulator transformer are shown by Fig. 88-9. The outputs of the sweep and marker generators are connected through capacitors to the grid of the tube preceding the demodulator transformer. This tube may be the driver for an intercarrier sound system or the limiter for a dual sound system.

The vertical input of the oscilloscope is connected to the audio output of the demodulator. Ordinarily the most convenient point for connection is at the high side of the volume control potentiometer, as shown by the full line cable. The scope may otherwise be connected to the input or the center of the de-emphasis filter, as shown by broken lines. It is not necessary to use either a capacitor or a resistor in series with the lead for the scope vertical input.

Unless your oscilloscope has fairly high vertical sensitivity the S-curve produced with the generators connected as shown may not have enough height without undesirably high output from the sweep generator. In this case move the generator connections back to the grid of some preceding tube in order to have the added amplification of such tubes. You may move the generator connections as far back as the grid of the tube which precedes the sound takeoff. In this case the S-curve will show the effects of any misalignment of the couplers which now come between the demodulator transformer and the tube to which the generators are connected. Unless these other couplers are correctly aligned it may be difficult to obtain a satisfactory S-curve from the demodulator output or, even though the S-curve appears to be good, there may be poor reproduction of sound during normal reception.

The marker generator usually is connected to the same point as the sweep generator, although the marker might remain connected as in Fig. 88-9 when the sweep connection is moved farther back. This tends to avoid distortion of the curve with an excessively strong marker pip. The marker generator is used without audio modulation, and is tuned as required to check frequencies along the S-curve.

With all test connections completed and with the generators, oscilloscope, and receiver warmed up, it is in order to bring an S-curve onto the screen of the scope. Commence by adjusting the sweep width for about 3 mc. Adjust the vertical gain of the scope to maximum and turn up the output control of the sweep generator about half way. Now adjust the center frequency of the sweep generator to bring any kind or shape of S-curve to the center of the screen. If the trace is too high, turn down the output control of the sweep generator but do not reduce the vertical gain of the oscilloscope. Reduce the sweep width to leave only the up

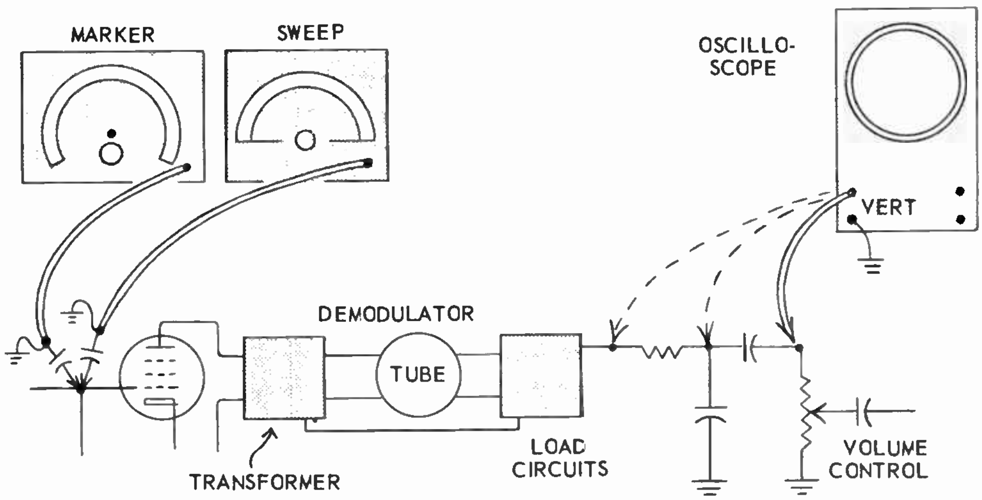


Fig. 88-9. Instrument connections for alignment of the demodulator transformer.

and down peaks, with very little of the horizontal lines remaining on each side. It will be necessary to re-adjust the sweep center frequency in order that the S-curve remains at the center of the screen.

Now try out the marker by tuning it very slowly back and forth through the center frequency and as far either way as causes the pip to disappear to the right or left. You will find that the marker pip shows up strongly at both peaks of the S-curve, but becomes weaker as it moves toward the center, and probably disappears entirely at the center frequency. This is because, at the center frequency, the trace is passing through the point of zero gain and there is no amplification of the marker beat. This disappearance of the marker pip is somewhat of a handicap, since it is essential to identify the frequency at the center of the S-curve, and make it equal to the center frequency of the applied signal.

One method of getting around the difficulty is to carefully check the frequencies at the highest points of the two peaks, and assume that the center of the curve is at a frequency half way between. Another way is to increase the output of the marker generator as much as possible as the pip comes toward the center of the curve, and note the point along the curve where the strong pip fades out. This is the point of zero gain on the S-curve, and should be at the center frequency.

The frequency at the point of zero gain on the S-curve may be identified quite closely if the marker generator provides audio modulation, as most generators do. Make the sweep width great enough to leave the outer sides of the S-curve nearly flat, as at the left in Fig. 88-10. Turn on the audio modulation and tune the marker generator back and forth through the center frequency. Except at the frequency of zero gain the audio modulation will show up on both sides of the S-curve about as at the right, as a series of small curves weaving one way and another. Turn the marker output rather high, to emphasize the audio modulation, then tune the marker to the frequency at which the modulation effect disappears from the ends of the S-curve. The marker now is tuned to the frequency of zero gain, which should be the center frequency. Use this method only for checking the center of the S-curve, turn off the audio modulation while identifying peak frequencies.

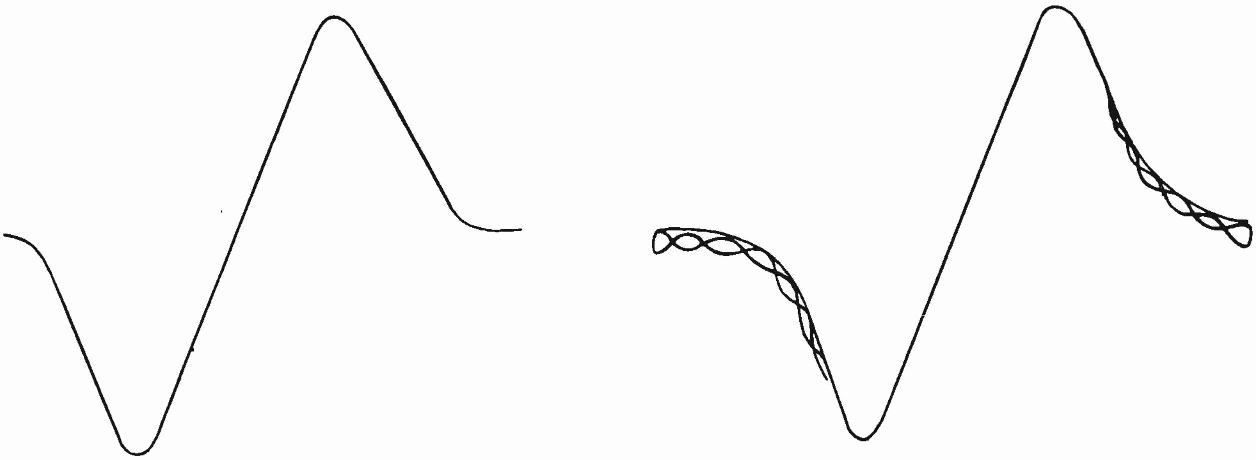


Fig. 88-10. Audio modulation on the marker generator shows up as wavy lines except at the center frequency for sound.

Every characteristic of the S-curve is altered to some extent by every adjustment of either the primary or secondary in the demodulator transformer, but in a general way the primary adjustment has greatest effect on the following.

Obtaining maximum peak amplitudes.

Making the shapes of the positive and negative parts symmetrical.

Obtaining the longest possible straight portion above and below the center.

Adjustment of the transformer secondary has the principal effect in bringing the center frequency midway between the positive and negative peaks.

The primary and secondary both require adjustment in making the two peaks of equal amplitude.

To bring the entire S-curve to a higher frequency range the cores of both primary and secondary must be turned farther out of their windings. First adjust the primary while attempting to hold maximum peak-to-peak amplitude, then adjust the secondary for equal peak amplitudes above and below the center. To bring the entire curve to a lower frequency range turn the cores farther into both primary and secondary windings.

The whole job of demodulator alignment usually boils down to alternate adjustments of primary and secondary until you obtain the desired overall result. Examples of what happens are shown by Fig. 88-11. The condition at the left was corrected by adjusting the primary to resonate at a slightly lower frequency. The condition at the right was corrected by adjusting the secondary to a slightly lower frequency.

We really are interested only in the two peaks of the S-curve and in the part of the trace which is between the peaks. It may be somewhat easier to observe the effect of adjustments by watching only these parts of the curve, as in Fig. 88-12. This is accomplished by reducing the sweep width while readjusting the sweep center frequency as may be necessary to keep the trace centered on the screen.

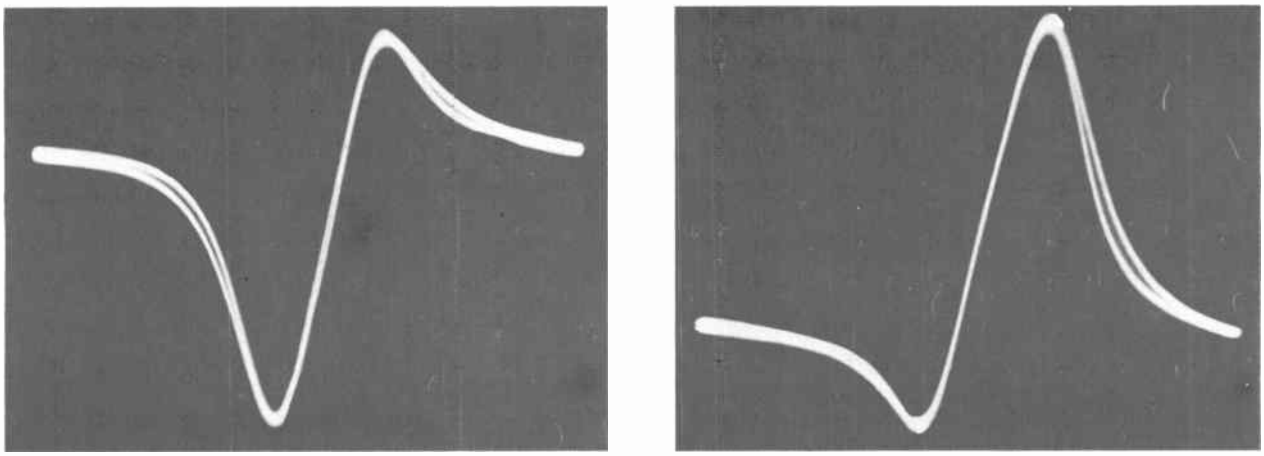


Fig. 88-11. Faulty S-curves which are corrected by suitable adjustments of primary and secondary in the demodulator transformer.

After completing alignment of the demodulator the results should be checked as follows. Turn off the oscilloscope and the sweep generator, but leave the marker turned on and connected as usual. Use the marker generator with audio modulation. Turn the volume control of the receiver to maximum. Now tune the marker generator back and forth through the center frequency. The audio tone should be heard from the speaker on both sides of the center frequency, but should fade almost to inaudibility at the exact center

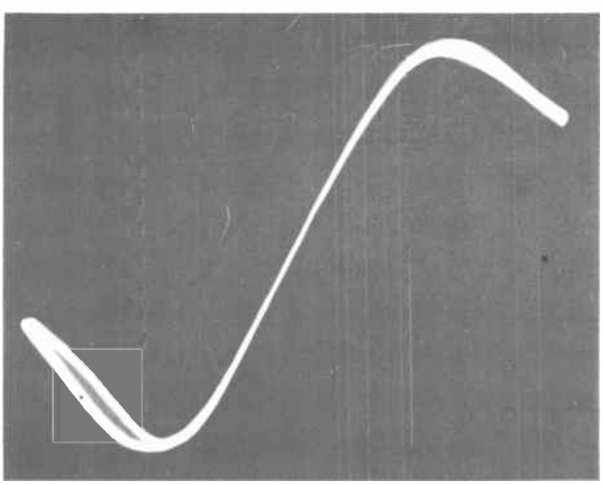


Fig. 88-12. Only the two peaks and the central portion of the S-curve are important for alignment.

frequency. If the audio tone does not nearly disappear at the center frequency, the secondary adjustment of the domodulator transformer needs slight readjustment to bring the sound to zero or a sharp minimum at the center frequency.

A final check for buzz with an intercarrier sound system is made as follows. Disconnect all the test instruments from the receiver, also place the automatic volume control, the r-f oscillator, and everything else as for normal reception. Connect the antenna and tune in a television station to get the picture and sound. Turn up the contrast control until a buzzing noise is heard. Now try making a very slight readjustment of the secondary in the demodulator transformer, keeping track of the setting from which you start. The secondary should be left at the setting for minimum buzz and maximum sound volume, or at the setting for the minimum buzz between much louder buzzing sound on either side.

It is quite obvious that either of these "final checks" might be applied to a receiver which you suspect of being slightly out of correct sound alignment. This could be done without having made any of the regular alignment adjustments, merely as a means for preliminary trouble shooting which might avoid the need for complete realignment of a sound system.

ORDER OF ALIGNMENT. With the methods of sound system alignment which have been described we first adjust the takeoff and any i-f couplers for the center frequency and desired band pass, then adjust the demodulator transformer. Some technicians prefer to commence the sound alignment by adjustment of the demodulator transformer. Then they adjust any i-f transformers and finish the work by adjusting the takeoff.

With this latter method the test instruments must be connected as in Fig. 88-9 while adjusting the de-

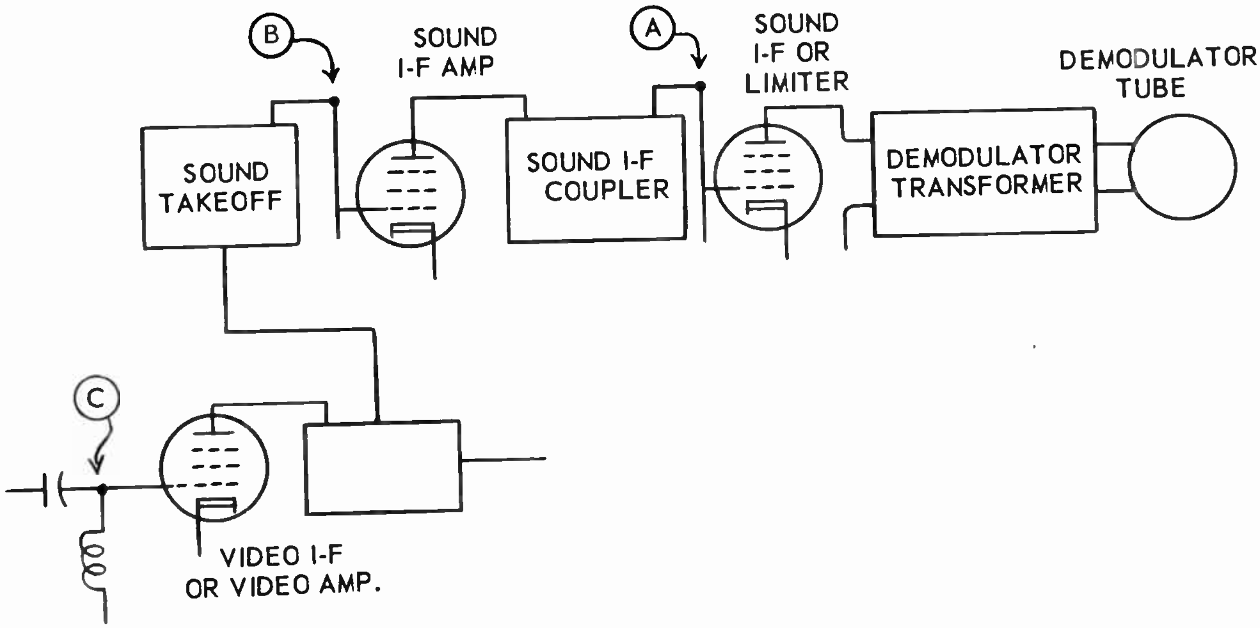


Fig. 88-13. Sweep and marker connections for alignment proceeding from the demodulator transformer back to the sound takeoff.

modulator transformer to obtain a correct S-curve. The generators cannot be connected farther back toward the sound takeoff, nor ahead of the takeoff, because none of the couplers from the takeoff to the last i-f transformer will have been adjusted as yet.

The oscilloscope vertical input is connected to the top of the volume control or the audio output of the demodulator to show an S-curve, and remains there during all adjustments. Points to which the sweep and marker generators are connected are shown by Fig. 88-13. For alignment of the demodulator transformer the generators are connected to point *A*, the grid of the sound i-f amplifier or the grid leak resistor of a limiter just ahead of the demodulator.

Then the generator connections are moved to *B*, the grid of a preceding sound i-f amplifier, while the sound i-f coupler following this tube is adjusted to produce maximum and equal peak amplitudes on the S-curve. If there is still another sound i-f amplifier tube and coupler the generators are moved back to the grid of this other amplifier while the coupler following the amplifier is adjusted for maximum and equal peaks on the S-curve. Finally the generators will be connected to *C*, the grid of the tube preceding the sound takeoff, while the takeoff coupler is adjusted for maximum and equal peaks on the S-curve. While the generators are connected at *C* it is advisable to try making slight changes of adjustment on any i-f couplers for the purpose of obtaining greatest possible peak amplitudes on the S-curve.

INTERNAL SWEEP OF OSCILLOSCOPE. The S-curve may be observed without employing a synchronized sweep voltage for the horizontal input of the oscilloscope, using instead the internal sweep of the scope adjusted for a sweep rate of 60 cycles.

With the sweep and marker generators connected and operated as usual, and with the oscilloscope vertical input on the audio output of the demodulator the selector of the oscilloscope is changed from the horizontal input position to the internal sweep position. When sweep width of the sweep generator is made several

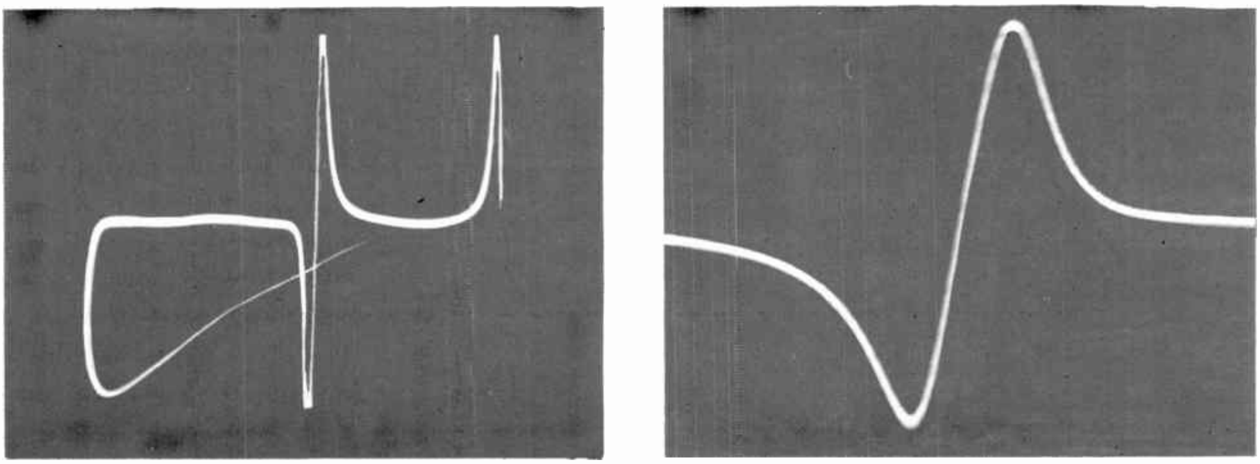


Fig. 88-14. Using the internal sweep of the oscilloscope to display an S-curve.

megacycles, and the internal sweep of the oscilloscope is adjusted for 60 cycles, a trace such as shown at the left in Fig. 88-14 will appear.

There are two complete S-curves on this trace. One of the S-curves is partly obscured by the return trace interval. One of the curves is produced during the 1/120 second in which sweep frequency is increasing, and the other is produced during the following 1/120 second in which frequency is decreasing. Thus the two curves are formed during the two periods of 1/120 second each, or in the 1/60 second in which the beam crosses the screen of the scope when the internal sweep is set for 60 cycles.

The next step is to increase the horizontal gain of the scope and operate the horizontal centering control as required to center the S-curve that is complete. The result is shown at the right in Fig. 88-14. This single S-curve may be used for all alignment work in just the same way as the forward and reverse superimposed curves obtained with a synchronized sweep voltage.

F-M RECEIVER ALIGNMENT. All of the methods which have been explained for alignment of television sound systems with sweep and marker generators and an oscilloscope apply also to alignment of the i-f amplifier stages and the demodulator transformer in f-m sound receivers. The center frequency for all f-m receivers designed during recent years is 10.7 mc. Band width for the i-f transformers should be about the same as for i-f transformers in television sound systems, or should be at least 200 kc.

In cases where the generator connections would be to the grid of the tube preceding the sound takeoff in a television set these connections would be made to the r-f signal grid of a converter tube or to the grid of a separate mixer tube in an f-m receiver. If there is difficulty in obtaining an S-curve sufficiently high on the oscilloscope its vertical input may be connected to the plate of the first audio amplifier tube, which tube often is combined with the demodulator. If a limiter tube has a cathode bias resistor it usually is satisfactory to connect the oscilloscope vertical input across this resistor instead of to the limiter grid resistor while aligning the i-f stages.

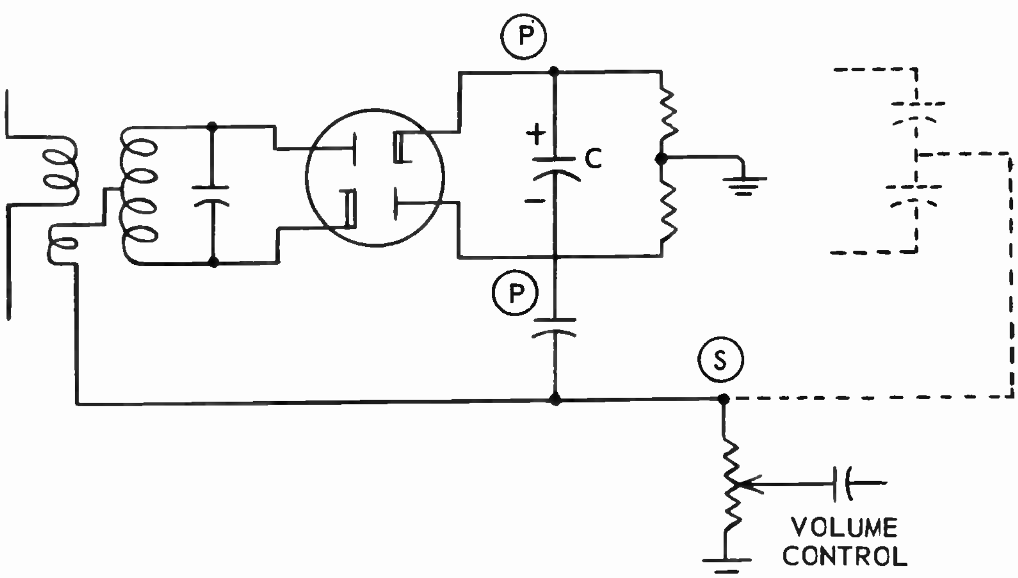


Fig. 88-15. Points at which the VTVM is connected during alignment of a ratio detector.

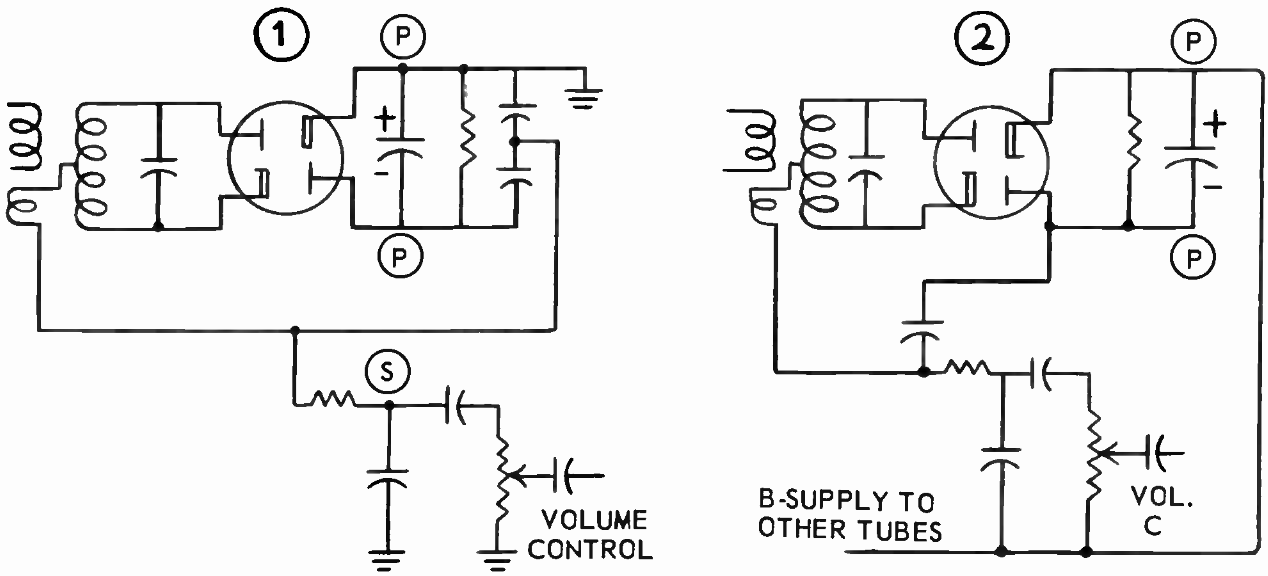


Fig. 88-16. Ratio detector circuit variations, and points at which the VTVM is connected during alignment.

SOUND ALIGNMENT WITH VTVM. The entire television sound system may be aligned with an accurate constant frequency signal generator as the source and with a vacuum tube voltmeter as the indicator, instead of with sweep and marker generators used in connection with an oscilloscope. Instead of a vacuum tube voltmeter it is possible to use a d-c moving coil meter having sensitivity of at least 20,000 ohms per volt.

The signal generator may be connected through a capacitor to any of the points shown in Fig. 88-5 for the sweep and marker generators. For alignment of the demodulator transformer only, the signal generator may be connected to point *A* in that figure. For alignment of the sound takeoff and all following couplers, including the demodulator transformer, the signal generator may be connected to the grid of the tube preceding the sound takeoff. The generator must be tuned precisely to the sound center frequency, and is used without audio modulation.

We shall deal first with alignment of sound systems in which the demodulator is a ratio detector. Connections for the VTVM vary with the manner in which the ratio detector is wired. Several common methods are shown by Figs. 88-15 to 88-17.

In Fig. 88-15 the large load capacitor *C* has across it two resistors whose junction is connected to ground. As shown by broken lines there may be also two small capacitors whose junction is connected to the audio output line which goes to the volume control. Note that there is no capacitor in the audio output line to the volume control.

At *1* in Fig. 88-16 there is a single undivided resistor across the large load capacitor, and there are also two small capacitors whose center junction is connected to the audio output line. Note that there is a capacitor between the audio output line and the volume control.

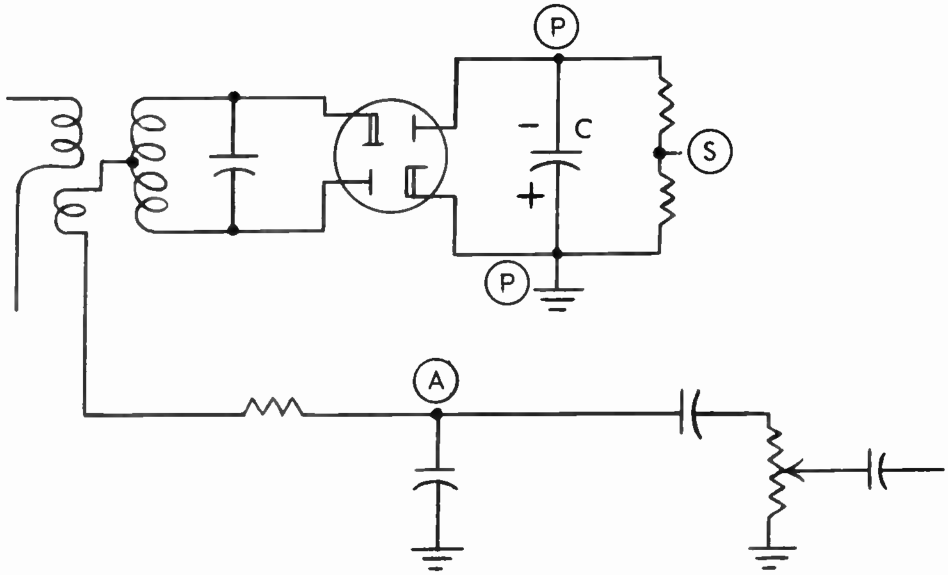


Fig. 88-17. A ratio detector with its load connected only to ground, and connection points for the VTVM during alignment.

At *2* in Fig. 88-16 is a variation sometimes found when the cathodes of some or all the tubes in the sound system are connected to plates and screens of i-f amplifiers and other tubes in a series B-supply arrangement. There is only a single undivided resistor across the large load capacitor.

In Fig. 88-17 the large load capacitor *C* has connected across it two resistors whose junction is not grounded. There is a capacitor between the audio output lead and the volume control. This arrangement, like the one at *2* in Fig. 88-16, may be used with a series B-supply system.

For alignment of the sound takeoff, any i-f amplifier stages, and also the primary of the demodulator transformer, the two leads from the VTVM may be connected to the two sides of the large load capacitor at the points marked *P* and *P* on all the ratio detector diagrams. In some of these circuits neither side of the large load capacitor is grounded. In this case, when connecting the VTVM across the load capacitor do not connect either meter lead to ground and make certain that the case of the VTVM is not grounded but is supported by insulation on the test bench or instrument shelf.

In Fig. 88-15 the VTVM could be connected between either one of the points marker *P* and ground, which would place the meter across either half of the resistor combination which is in parallel with the load capacitor.

With the circuit of Fig. 88-17 the VTVM may be connected in either of two ways. First, the meter may be connected from either of the points marked *P* to point *S* between the load resistors. Second, the meter may be connected between point *S* and point *A* on the audio output lead ahead of the capacitor in the line to the volume control. With either of these connections make sure that the case of the VTVM is not grounded.

With the VTVM connected in any of the ways described, adjust all slugs or trimmers of the sound takeoff. of any i-f couplers which may be used, and also the primary of the demodulator transformer for maximum reading on the meter.

Now we shall consider alignment of the secondary in the demodulator transformer when the demodulator is a ratio detector.

With a circuit such as shown by Fig. 88-15 connect the VTVM from point *S* to ground, or from the top of the volume control to ground. If there should be a capacitor in the line to the volume control, make the meter connection ahead of this capacitor, not to the high side of the volume control. The same kind of meter connection, from point *S* to ground, will serve for the circuit at *1* in Fig. 88-16.

The circuit at *2* in Fig. 88-16 does not allow using the VTVM for aligning the secondary of the demodulator transformer. After the primary has been aligned as described, remove all the instrument connections, tune in a station, turn up the contrast control, and adjust the secondary for minimum buzz. This always is the final check with an intercarrier sound system, no matter what the alignment method.

With the circuit of Fig. 88-17 connect the VTVM from point *S*, between the two load resistors, to ground.

When the voltmeter is connected in any of the ways described for secondary adjustment, align the secondary of the demodulation transformer for an exact zero reading which will occur between a positive peak and a negative peak as the adjustment is turned one way and the other. This is the zero point at the center of the S-curve. Make certain that turning the adjustment in one direction will cause a positive peak and then a decline of meter reading (one peak of the S-curve) and that turning the adjustment the opposite direction will cause a negative peak and then a decline (the other peak of the S-curve). Leave the adjustment for exact zero between the two peaks. If you should leave the adjustment for zero at the outside of either peak there will be no sound reproduction or very bad reproduction.

If the demodulator is a discriminator type, such as shown by Fig. 88-18, there are two methods of alignment. For both methods the signal generator is connected to the grid of the tube ahead of the sound takeoff. The first method is as follows.

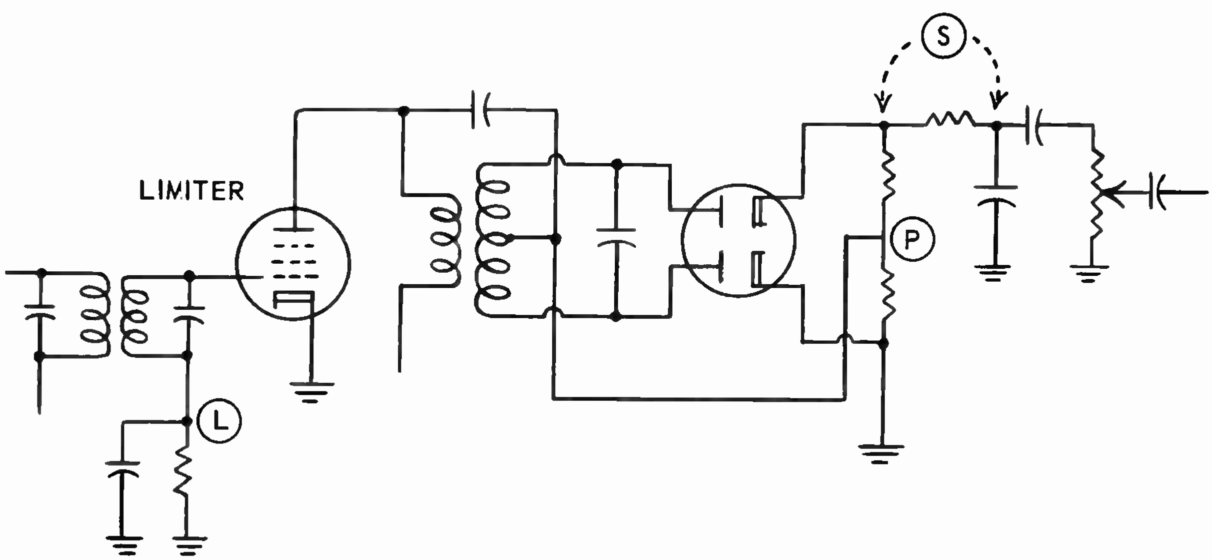


Fig. 88-18. Points on a discriminator circuit to which the VTVM is connected during alignment.

Connect the VTVM between ground and point *S*, which is the audio output at any point preceding a capacitor in the line to the volume control. If there is no series capacitor, connect to the high side of the volume control.

Detune the secondary of the demodulator transformer by turning its core out two or three full turns.

Adjust for maximum meter reading (a) the sound takeoff, (b) all sound i-f couplers or transformers, and (c) the primary of the demodulator transformer. The output of the signal generator must be reduced as necessary to avoid limiting action.

Now adjust the secondary of the demodulator transformer for exact zero between positive and negative peaks. Make sure that this zero is between the peaks.

The second method requires changes of the VTVM connections, as follows.

Connect the VTVM across a limiter grid resistor, point *L* to ground in Fig. 88-18, or across a limiter cathode resistor. Adjust the sound takeoff and all the sound i-f couplers or transformers for maximum reading of the meter.

Next, connect the VTVM between ground and the junction between the two discriminator load resistors, point *P*. Adjust the primary of the demodulator transformer for maximum meter reading.

Finally connect the VTVM between ground and the audio output, point *S*. Adjust the secondary of the demodulator transformer for exact zero reading between positive and negative peaks. Take the usual precautions to have the zero between two peaks.

After any type of demodulator has been aligned the peaks of the S-curve may be checked with the VTVM connected as for obtaining a zero reading when adjusting the transformer secondary.

First, tune the signal generator to the frequency at which there is a peak reading below the zero for center frequency. Note this frequency, also the voltage.

Second, tune the signal generator to the peak reading above the center frequency. Note this frequency and the voltage of the peak.

The two frequencies should be about equally below and above the center frequency, and the two peak voltages should be approximately equal. Otherwise there is a possibility that the transformer secondary is misaligned.

When working with an intercarrier sound system always make a final check for intercarrier buzz. Disconnect all the test instruments, connect the antenna, and tune in a station. Turn up the contrast control to produce a buzz, if this is possible. If there is a buzz, try making a slight readjustment of the demodulator transformer secondary to obtain minimum buzz.

TELEVISION

LESSON NO. 89

TELEVISION R-F OSCILLATOR ALIGNMENT

Training Your Mind... C.T.I. Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind
C.T.I.
Perfecting Your Skill

COMMERCIAL
TRADES
INSTITUTE

Chicago, Ill.

LESSON NO. 89

TELEVISION R-F OSCILLATOR ALIGNMENT

There are two principal operations in aligning tuners of the types in general use. First, we adjust the frequency of the r-f oscillator. This is one of the most common service operations on the majority of receivers, and is quite easily performed.

Second is adjustment of the r-f amplifier circuits. This part of the work consists of aligning the coupling between r-f amplifier and mixer, and sometimes also the coupling between antenna and r-f amplifier. In some tuners there are accessible movable cores or trimmer capacitors for adjustment of these r-f circuits. Even then the operation requires great care and good testing equipment. In other tuners the r-f adjustments are of types difficult to alter until you have had considerable actual practice. Fortunately, it is only in rare cases that there is any real need to realign the r-f circuits.

Both electrically and mechanically there are greater differences between various types of tuners than between any other parts of television receivers. In spite of these differences we may think of any tuner as consisting basically of the elements shown in Fig. 89-1. To begin with there is a coupling between the antenna or transmission line and the r-f amplifier. Sometimes this coupling is fixed for all channels. In other cases it is broadly tuned for the high band and tuned differently for the low band channels. Again there may be tuning adjustments for some channels but not for others, or there may be separate tuning adjustments for each and every channel. There may be one or two R.F. amplifier stages in the tuner.

At B and C the output of the r-f amplifier is coupled to the input of the mixer. This coupling always must be tuned in some manner or other to suit the frequencies of each channel. The coupler may be a transformer, with a tuning adjustment at B for the r-f amplifier plate circuit on each channel, and another adjustment at C for the mixer grid circuit, again on each channel. There may be adjustments only for certain channels. Instead of a two-winding transformer with two adjustments this coupler may be an impedance type, in which B and C are tuned at the same time. With either type of coupler there may be adjustments for each individual channel or for only some of the channels.

The r-f oscillator must be tuned, at D, for each channel. The oscillator frequency must be different for each channel to be received. When this oscillator frequency is fed into the mixer, along with carrier frequencies from the r-f amplifier, the resulting beat frequencies become intermediate frequencies in the output of the mixer.

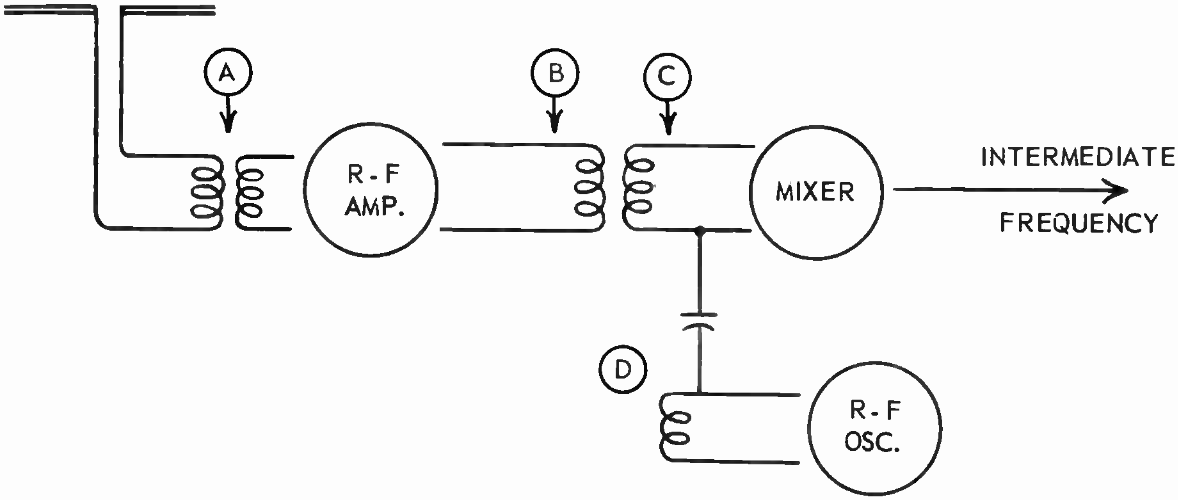


Fig. 89-1. The essential elements of a tuner.

EFFECTS OF OSCILLATOR FREQUENCY. Before undertaking the alignment of r-f oscillators it is desirable to understand how the oscillator frequency affects relations between the r-f amplifier and the i-f amplifier of the receiver. The frequency response of the r-f amplifier is very wide, often being somewhat as shown at 1 in Fig. 89-2. This particular response curve is marked with frequencies which would exist when the receiver is tuned for channel 5. For other channels the shape of the r-f response would be similar but the frequencies would be those for each channel.

We shall assume that our r-f oscillator is tuned to 103 mc, for reception on channel 5. This oscillator frequency will beat in the mixer with all the amplified carrier frequencies. The difference-frequency response is shown at 2. This is the i-f frequency band at the output of the mixer. Note that the i-f response is just as wide as the r-f response, the difference being only in the frequency values.

The i-f amplifier system is tuned or aligned to amplify a relatively narrow band of frequencies, as shown at 3. With the r-f oscillator correctly tuned the i-f band pass will have the frequency relation to the r-f band pass as shown. From zero gain to maximum and back to zero, every part of the i-f band pass falls on a portion of the r-f band pass where there is practically uniform gain. Then, for all the frequencies amplified in any degree by the i-f system, there will be almost constant r-f voltage input from the mixer plate to the i-f system. When this condition is realized in practice the output of the i-f amplifier to the video detector and sound system will be proportional to relative gains or to frequency response of the i-f amplifier.

Supposing now that the r-f oscillator is incorrectly tuned to a frequency of 105 mc, still with the selector set for channel 5. The channel frequencies still come through the r-f amplifier as at 1 in Fig. 89-2. But with the changed oscillator frequency the beat frequencies or intermediate frequencies in the mixer output are as shown by curve 4. At the lower peak the frequency now is the difference between 105 mc from the oscillator and 82 mc from the r-f amplifier, or is 23 mc. By subtraction we find that the upper peak is at 29 mc instead of the original 27 mc.

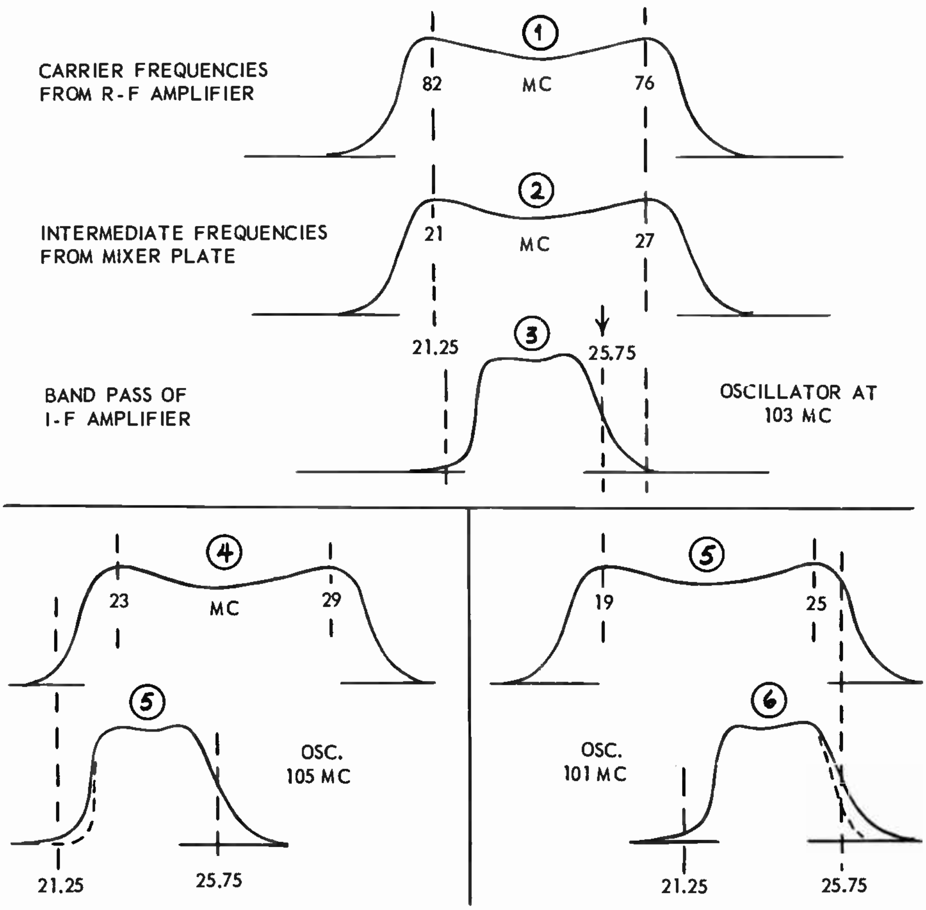


Fig. 89-2. Relations between r-f and i-f frequency responses as affected by adjustment of oscillator frequency.

Altering the frequency of the r-f oscillator has not changed the frequency response of the i-f amplifier system. Consequently, the i-f response has to match the changed range of intermediate frequencies. The relations between i-f and r-f band passes are as shown by curve 5. The entire i-f response no longer falls on the constant gain portion of the r-f curve. Rather the low-frequency end of the i-f response is well down on the r-f response, and in all probability there would be no reproduction of sound.

Were the r-f oscillator incorrectly tuned at 101 mc the frequency relations between the i-f and r-f responses would be as shown at 5 and 6. Now the video intermediate frequency point of the i-f response comes where gain has commenced to drop off on the r-f response, and there would be difficulty in getting good picture reproduction, although there would be plenty of sound output. With an r-f band pass any narrower than has been assumed there would be real trouble with any appreciable mistuning of the r-f oscillator.

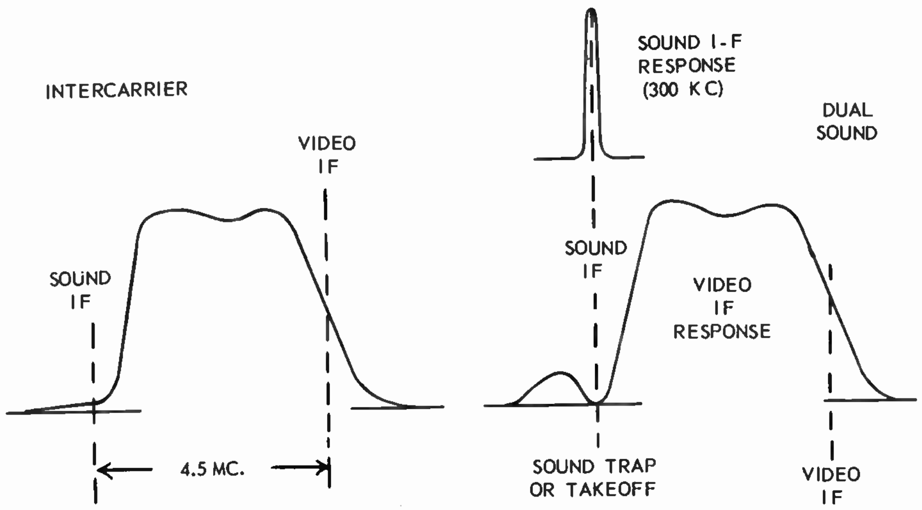


Fig. 89-3. I-f responses for intercarrier sound systems (left) and for dual sound systems (right).

With intercarrier sound, and an i-f amplifier system aligned for suitable frequency response, it usually is possible to make slight alterations of r-f oscillator frequency without getting into serious difficulties. The reason is illustrated at the left in Fig. 89-3. Note that the low-frequency end or sound end of the i-f response has a long and gentle slope at the low level of gain required for the sound intermediate frequency.

When the r-f oscillator frequency is altered to bring the video intermediate frequency higher or lower on the high-frequency slope the sound intermediate frequency is shifted to the left or right. If the sound intermediate is not moved to a point of excessively low gain, nor too high onto a sharply sloped part of the response the reproduction of sound will remain acceptable during a fairly wide change of video intermediate frequency on the response slope. There will be a variation of sound volume, but this usually can be compensated for by adjustment of the volume control.

Conditions with a dual sound system are represented at the right in Fig. 89-3. Up above the video i-f response is shown the sound i-f response for a maximum width of 300 kc. The peak of sound i-f response will fall at the same point on the video i-f response as a dip due to sound traps or to trap action of the sound takeoff. It is apparent that a very small shift of r-f oscillator frequency will move the sound intermediate frequency not only off the peak of sound response, but completely beyond the range of any sound response at all. At the same time the sound intermediate frequency will move away from the point of trap attenuation on the video i-f response, and sound bars probably will appear in the pictures.

With a dual or split sound system the r-f oscillator frequency must be exact within 0.05 mc or better to have acceptable sound reproduction. It is not possible to move the video i-f point up or down on the i-f response without losing the sound and, usually, without having sound bars in the pictures at the same time.

OSCILLATOR ADJUSTMENTS. Many tuners provide individual adjustments for oscillator frequency on each channel. A fairly typical arrangement is illustrated by Fig. 89-4. The separate oscillator inductors, one for each channel, are directly back of the front plate of the tuner housing, which is at the right in the picture. Inside each inductor is an adjustable core with a screw driver slot. The core slots are reached by inserting a screwdriver through openings in the front plate. The openings usually may be exposed while the chassis is in its cabinet by removing the dial plate or escutcheon on which are the channel numbers. This may also require removal of the tuning knob of the channel selector.

The construction illustrated is that of a tuner employing a rotary selector switch and stationary inductors. In some turret tuners with which the oscillator inductors rotate as the turret is turned from channel to channel, the inductor cores come successively into position behind a single opening in the tuner housing. The core which can be reached through this opening with the selector set for any given channel is in the oscillator inductor for that particular channel. There are, of course, many other arrangements of oscillator tuning inductors or trimmer capacitors. Some are accessible with the receiver in its cabinet, while others require removal of the chassis in order to make adjustments.

OSCILLATOR ALIGNMENT (Intercarrier Sound). It is possible to make oscillator alignments with or without testing instruments. The methods differ in some respects for receivers which employ intercarrier

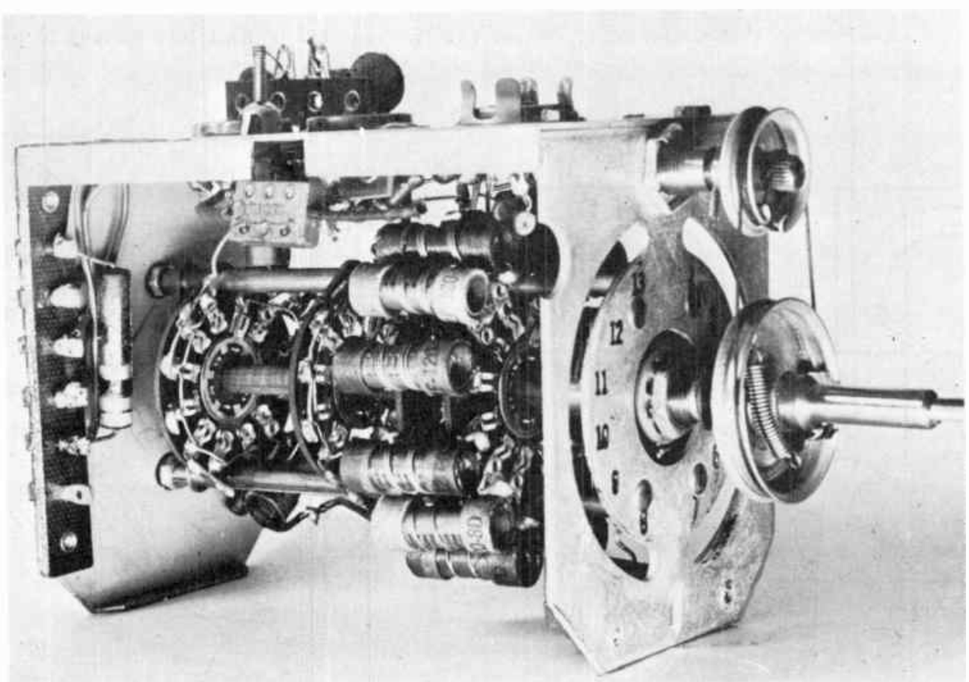


Fig. 89-4. Movable cores in the oscillator tuning coils are accessible through the front plate of this tuner.

sound and for those having dual or split sound systems. We shall deal first with methods requiring no instruments, as applied in the case of intercarrier sound.

Set the channel selector to the lowest or highest channel in which there is reception for your locality. Let the receiver warm up for at least 15 minutes, adjusting the operator's controls as required. If there is a fine tuning control set it at approximately mid-range and leave it there during the entire process of alignment. Adjust the oscillator slugs or trimmers only with a non-metallic tool. Make certain that you alter the oscillator tuning only for the channel to which the selector is set, and be sure to change the selector when going from one channel adjustment to another.

Adjust the oscillator tuning for the best possible detail in pictures or patterns, and with clear outlines of all objects. There must at the same time be ample sound output without having to turn the volume control to maximum. Do not try for the brightest possible picture, but for the best one. The brightest picture would mean that the video i-f point is too high up on the i-f response curve, and the sound i-f point doubtless would be too far down on the other side of the response. Repeat the operation for each channel in which there is transmission.

The next method employs sound bars on the picture or pattern as the indicator for correct adjustment. Carry out the preliminaries as previously described. Adjust the oscillator slug or trimmer in the direction that causes the picture to become brighter and causes alternate lighter and darker sound bars to appear horizontally on the picture tube screen. Now turn the oscillator adjustment in the opposite direction just far enough to make the sound bars disappear. In carrying out this process you have left the video i-f point as high as possible on the response while moving the sound i-f point low enough to prevent intercarrier buzz and other causes for unsatisfactory sound reproduction.

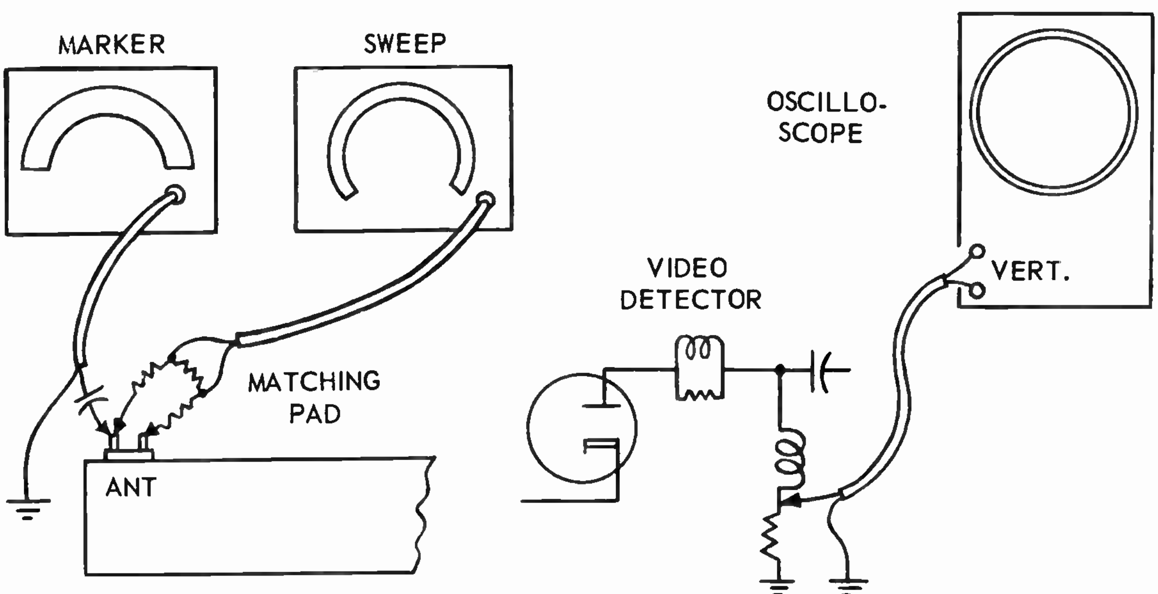


Fig. 89-5. Instrument connections for alignment of the r-f oscillator.

When the sweep generator, marker generator, and oscilloscope are used for aligning the r-f oscillator, connections are made as shown by Fig. 89-5. Since the generator signals will pass through the video i-f amplifier system on their way to the oscilloscope, this amplifier system should be in such condition as to provide a satisfactory frequency response. The steps in alignment are as follows.

1. It is desirable, but not absolutely necessary, to override the automatic gain control.

2. Disconnect the antenna or transmission line.

3. Connect the sweep generator to the antenna terminals through an impedance matching pad, as shown. We shall discuss such matching pads a little later. Otherwise the sweep generator high side may be connected through a 300-ohm carbon resistor to either antenna terminal, and the low side to chassis ground.

4. Connect the high side of the marker generator through a small capacitor to either antenna terminal, and connect the low side to chassis ground.

5. Connect the vertical input of the oscilloscope to the high side of the video detector load, and connect a ground lead from the oscilloscope to chassis ground or to B-minus.

6. Adjust the contrast control to its usual operating position.

7. If there is a fine tuning control, set it at approximately mid-position and leave it there during all following steps.

8. Turn on the receiver and all instruments, and let them warm up for at least 15 minutes.

9. Place the channel selector in position for the lowest channel, number 2, while adjusting the oscillator frequency for this channel. Then go to channel 3, and so on up to channel 13. Make sure that you alter the oscillator adjustment only for the channel to which the selector is set, and be sure to change the selector every time you go from adjusting one channel to the next channel.

10. Use only a non-metallic alignment tool for making adjustments.

11. For each channel proceed thus:

   a. Tune the center frequency of the sweep generator to get a response curve on the oscilloscope.

   b. Tune the marker generator to the video carrier frequency of the channel. The marker pip should appear about half-way down on the high-frequency slope of the response curve.

   c. Tune the marker to the sound carrier frequency for the same channel. For a receiver using an intercarrier sound system the pip should be no higher than 5 per cent of peak gain, and on the low-frequency slope of the response.

This same method may be used for a receiver employing dual sound. The only change is in the very ast step, where the sound carrier pip should be in the dip caused by a sound trap or by the sound takeoff.

COMMERCIAL TRADES INSTITUTE

TELEVISION

Now we may examine some methods of r-f oscillator alignment which are especially applicable where there is a dual sound system. Preliminary steps are much the same as for other alignment methods in which test instruments are employed. Briefly, they are as follows: Disconnect the antenna, set the contrast control to its usual position, place the fine tuning control at mid-position and leave it there, override the automatic gain control, be sure that the channel selector is set at the channel aligned. The sound i-f amplifier and demodulator positively must be in correct alignment.

The first method employs the speaker or else an output meter for the indicator. In addition to observing the preliminaries already mentioned, set the volume control at about mid-position.

A signal generator with a 60 or 400 cycle sweep should be used with this procedure and is connected to either antenna terminal and to ground. Tune the generator to the exact sound carrier frequency for each channel as the alignment proceeds.

When using the speaker as output indicator, adjust the r-f oscillator frequency on each channel for maximum sound loudness with the generator output as low as possible.

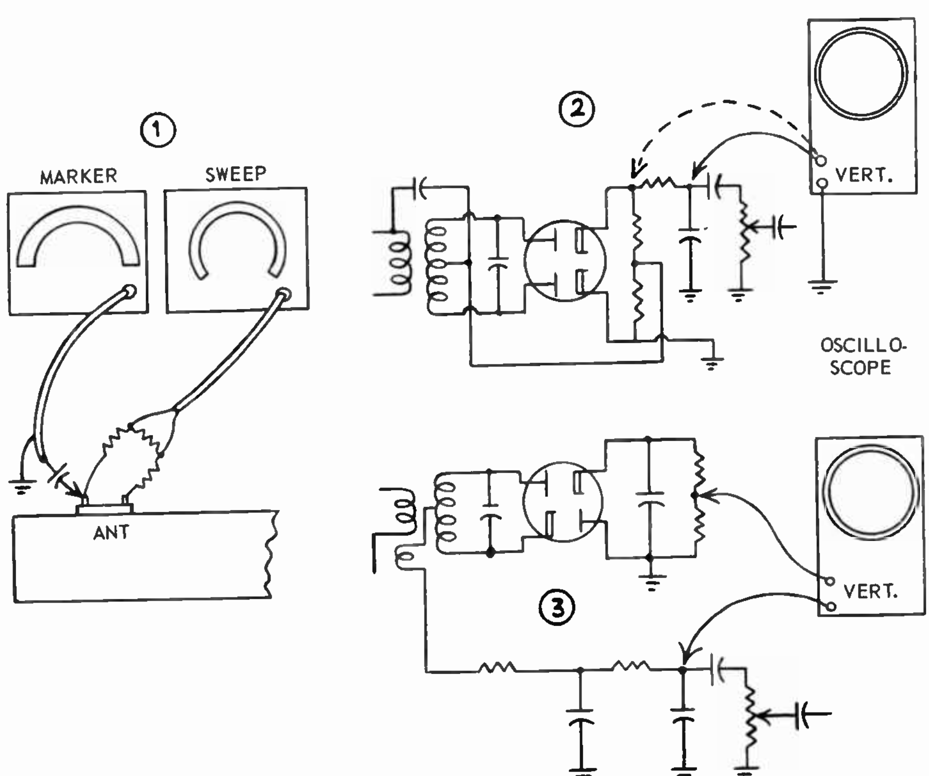


Fig. 89-6. Connections for oscillator alignment when using the sound demodulator output with the indicator.

Instead of using the speaker, an output meter may be connected across the voice coil or to the audio output tube in the same manner as for aligning a standard broadcast receiver. Adjust the oscillator frequencies for maximum meter reading with a low output from the signal generator.

Instead of using a signal generator it is possible to make fairly good adjustments by tuning in a television station on each active channel, then adjusting the oscillator frequencies for maximum sound volume or maximum reading of the output meter with the receiver volume control turned rather low.

The second method of r-f oscillator alignment where there is a dual sound system employs a vacuum tube voltmeter as output indicator. Preliminary steps are the same as previously mentioned, except that the receiver volume control may remain turned down or off.

The signal generator now used is a constant frequency type, connected to either antenna terminal and ground, and tuned to the exact sound carrier frequency for each channel aligned. This time the generator is used without modulation.

The VTVM may be connected to the audio output point of the demodulator, just as for alignment of the secondary in the demodulator transformer. On each channel adjust the r-f oscillator frequency for a zero meter reading which is sharply defined between positive and negative voltages or voltage peaks as the oscillator adjuster is turned one way and the other from its correct setting.

Instead of connecting the VTVM to the audio output of the demodulator, it may be connected across the limiter grid resistor, or it may be connected across the unbalanced output of the demodulator, exactly as when aligning the sound i-f amplifier and the primary of the demodulator transformer. Adjust the oscillator frequency on each channel for maximum meter reading while the output of the signal generator is very low.

The third method which may be used where there is a dual sound system employs an oscilloscope, a sweep generator, and a marker generator. Like the oscilloscope method described in connection with Fig. 89-5, this one may be used with either type of sound system, either dual or intercarrier. That earlier method is, however, especially well suited for a receiver having intercarrier sound, while the present one is best adapted where there is dual sound.

Instrument connections for the present method are shown by Fig. 89-6. As at 1 in the diagram, the marker generator is connected through a small capacitor to either antenna terminal, and to ground. The sweep generator is preferably connected to the two antenna terminals through a matching pad, but may be connected through a 300-ohm carbon resistor to either antenna terminal, and to ground.

The oscilloscope is connected to the audio output of the sound demodulator, just as when aligning the secondary of the demodulator transformer. With a discriminator the scope would be connected as at 2, with its vertical input to an audio output point on the discriminator circuit. With a ratio detector the vertical input and ground of the oscilloscope might be connected as at 3, or, depending on the particular kind of detector circuit, connected in any of the ways explained in connection with alignment of the secondary for ratio detector types of demodulators.

The channel selector of the receiver is turned to the channel for which the oscillator is first to be aligned, and the sweep generator is tuned and adjusted to bring an S-curve onto the screen of the oscilloscope. The marker generator is tuned to the exact sound carrier frequency for the channel. Then the r-f oscillator is tuned or aligned to bring the marker pip to the center of the central slope of the S-curve. The process is repeated for each channel.

MATCHING PADS. When the output of a sweep or other signal generator is connected to the antenna terminals of a receiver or tuner the impedances should be fairly well matched. The reason is the same as for matching of antennas, transmission lines, and receiver antenna inputs, it is to have maximum energy transfer while avoiding such wave reflections as would distort the frequency responses. Reasonably good impedance matching is necessary when aligning the r-f circuits of a tuner, and is highly desirable when other work is being done on the tuner.

Signal generators of all kinds nearly always are equipped with shielded output cables which are made of coaxial transmission line conductors. You will have to take it for granted that the cable and the output control or attenuator in the generator are matched well enough to provide satisfactory transfer of signal energy into the cable. The cables supplied with some signal generators are fitted with impedance matching pads at the receiver or tuner end, the match usually being for a receiver input impedance of 300 ohms. In this case it is necessary only to connect the "terminated" end of the cable to the antenna terminals of the tuner or receiver in order to have good matching between the impedances of the cable and the receiver.

If the generator output cable is not terminated with an impedance matching pad it is quite easy to make one with fixed carbon resistors connected as shown by Fig. 89-7. All three resistors may be of the smallest wattage size. All three may have resistance values expressed in the "preferred numbers" which are readily available from supply houses. Resistor Ra is shunted across the cable conductors. Resistors Rb are alike, and are in series with the two leads for the antenna terminals.

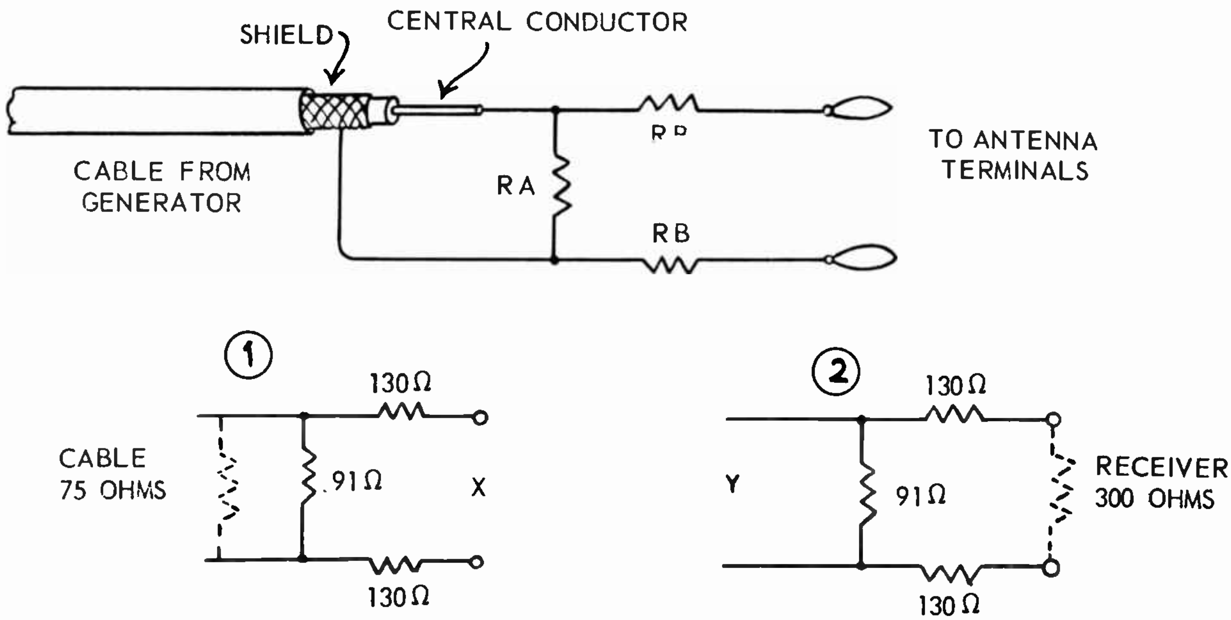


Fig. 89-7. The circuit commonly used for impedance matching pads between generators and antenna terminals.

In diagram 1 we are assuming that the cable impedance is 75 ohms, that resistor Ra is 91 ohms, while the two Rb resistors are 130 ohms each. The receiver antenna terminals are at X. The resistance or impedance across the receiver terminals is determined thus. The 75-ohm cable impedance and 91-ohm shunt resistance are in parallel with each other for a combined impedance of 41 ohms. This 41 ohms is in series between the two 130-ohm resistors to make a total resistance or impedance of 41 plus 130 plus 130, or 301 ohms as seen by the antenna terminals. This matches the 300-ohm input impedance of the receiver.

Impedance as seen by the cable is illustrated by diagram 2. The two 130-ohm resistors are in series with the 300-ohm input impedance of the receiver, making a total of 560 ohms. This 560 ohms is in parallel with the 91-ohm shunt resistor. The parallel resistance of 560 ohms and 91 ohms is 78.3 ohms, which is good match for the 75-ohm impedance of the cable.

Suitable resistor values for matching various line impedances to 300-ohm receiver input impedance are listed in the accompanying table for matching pads.

MATCHING PADS FOR 300-OHM RECEIVER INPUT IMPEDANCE

Cable Impedance	Shunt Resistor	Series Resistors	Resistance At Cable End	Resistance At Receiver End
50, 52 ohms	56 ohms	130 ohms	51 ohms	287 ohms
73, 75 ohms	91 ohms	130 ohms	78 ohms	301 ohms
95, 100, 105 ohms	120 ohms	120 ohms	98 ohms	295 ohms
150 ohms	220 ohms	100 ohms	153 ohms	289 ohms

Although a matching pad helps greatly when it comes to observations of true frequency response, the pad severely reduces the signal voltage reaching the receiver. A response trace which hardly can be observed on the screen of an oscilloscope with a pad in use will rise to many times the height without the pad, when there is no change in the vertical gain control of the oscilloscope. Unless the generators are capable of giving strong voltage outputs, and the oscilloscope is highly sensitive, it often is a choice between observing a good trace without the pad and observing nothing at all with the pad.

R-F ALIGNMENT. Adjustments in tuned circuits between antenna and r-f amplifier, and between this amplifier and the mixer, should not be altered unless you know exactly what you wish to accomplish and are sure that you know how to do it. Correct r-f alignment can do wonders to improve the performance of a television receiver, especially in fringe reception areas, but incorrect adjustments can do just as much toward ruining the performance.

The primary purpose of r-f alignment is to obtain maximum possible gain over a band width that includes both video and sound carrier frequencies, with something to spare on each side. It is worse than useless to obtain high gain over only a narrow band, for this will cause parts of the received signal to fall where there is little or no amplification. Unfortunately, band width and gain are mutually opposed, when either is increased the other will decrease.

There is a natural tendency for band width to increase and for gain to decrease at higher frequencies or at the higher channel numbers. This is because energy loss and high-frequency resistance increase as

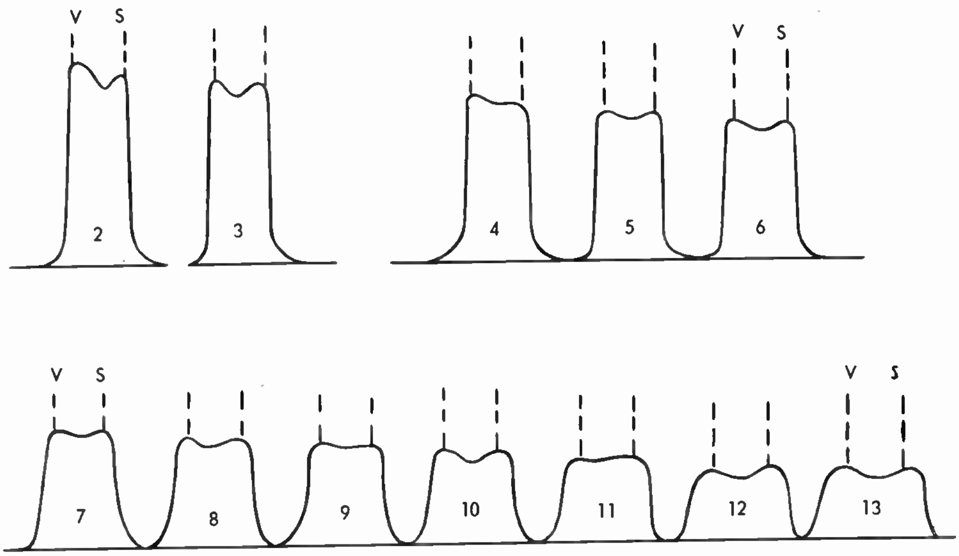


Fig. 89-8. Variations in gains and bandpasses of r-f responses for the various channels.

frequency goes up. The natural widening of frequency response is opposed in many tuners by using looser coupling between tuned circuits at the higher frequencies.

Even when there is band width compensation by change of coupling the relative gains and band widths in low-band and high-band channels are likely to be somewhat as shown by Fig. 89-8. The positions of the video carrier frequencies on the band passes are at V, and positions of sound carrier frequencies are at S. These two frequencies, which always have a constant separation, are on the outside of the r-f peaks for channel 2 and are well inside the peaks for channel 13.

The shapes of the response curves need not be alike for all channels. In fact the responses hardly ever would be alike. Fairly typical differences between channels are illustrated by Fig. 89-9. Quite often the higher of the two peaks will be on one side of the responses in the low band and on the opposite side for the high-band channels. One of the peaks may become higher and higher as channel frequencies increase or decrease. There may be a nearly flat-topped response for any one channel, with unequal peaks developing on channels either higher or lower. Of the several responses illustrated, only the ones for channels 2 and 10 are practically flat-topped, but receiver performance was entirely acceptable on all channels.

Principal requirements for the r-f response on all channels are illustrated by Fig. 89-10. The video carrier frequency should fall no lower on one side of the response than 75 per cent of maximum gain, and preferably is up on the peak or slightly inside the peak. The sound carrier frequency should fall no lower than 60 per cent of maximum gain on the other slope of the response, and preferably should be at or near the peak on that side of the curve.

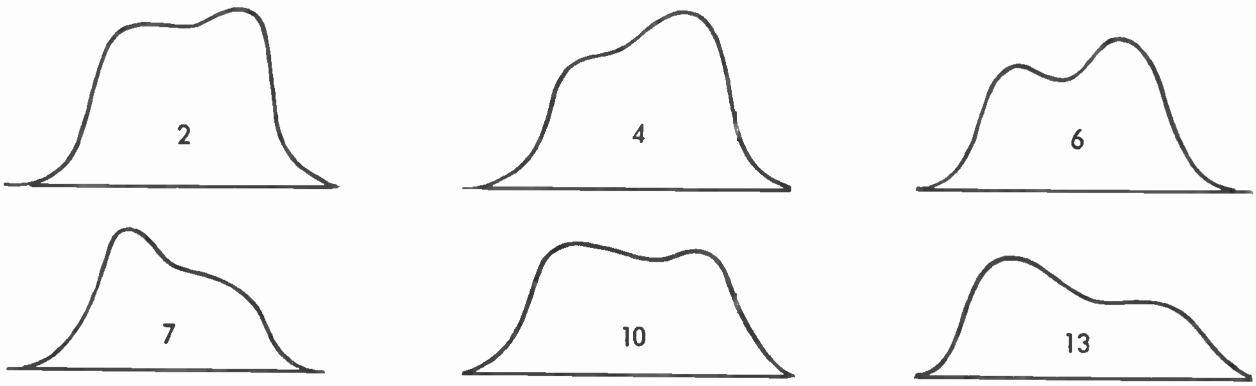


Fig. 89-9. R-f responses may show decided differences on various channels.

The top of the response curve should be as nearly flat as possible, but in making it flat for some channels you are likely to get excessive peaking on other channels. Consequently, it is necessary to make the best compromise adjustment that does not upset any one channel too much. It is, of course, desirable to have greatest possible gain so long as the other requirements are satisfied.

TUNER CIRCUITS. There are so many variations in structural and circuit details of various makes and models of tuners that it would take up the space of many lessons to give explicit instructions for r-f align-

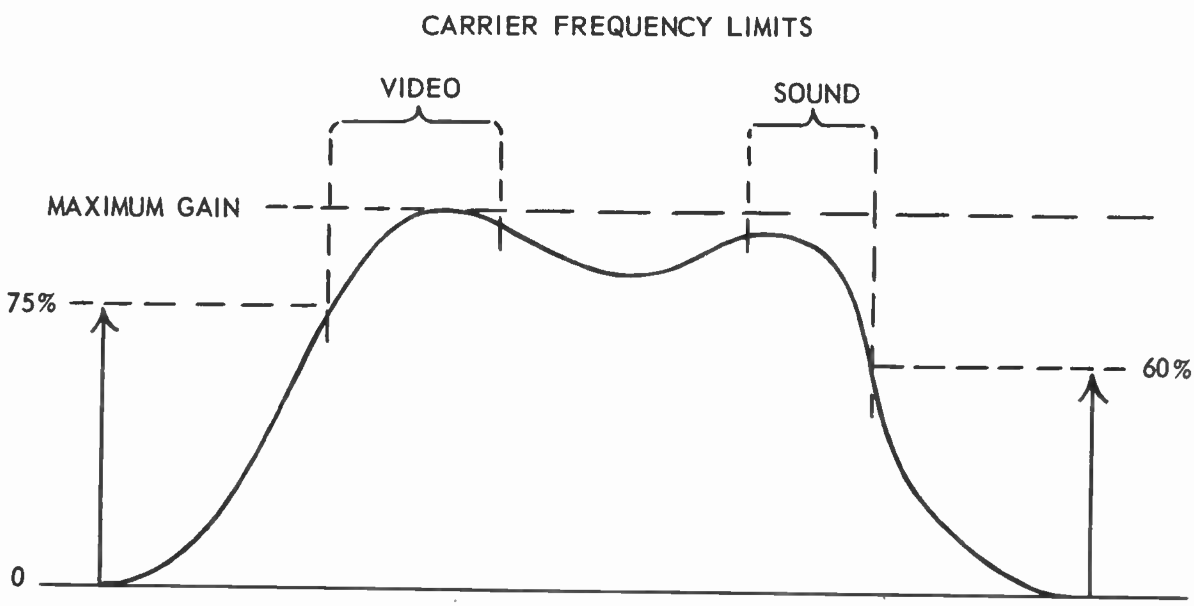


Fig. 89-10. Features of the r-f frequency response which are checked during alignment.

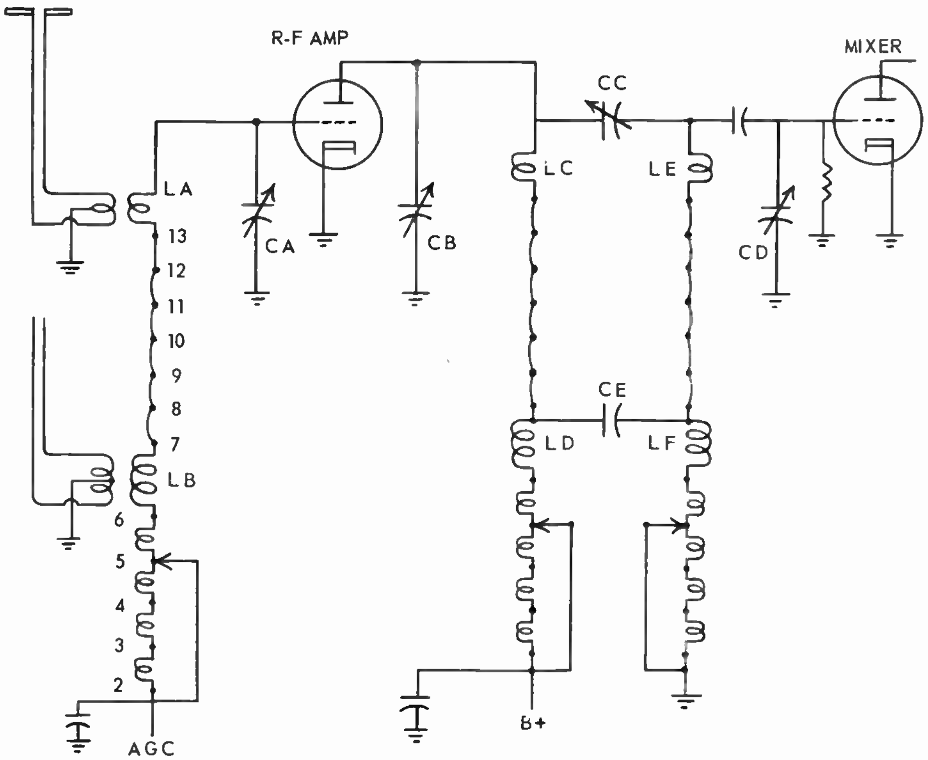


Fig. 89-11. Elements generally employed in tuners having series inductors for channel tuning.

ment of every separate unit. Even were this done we would have the problem of new designs which appear continually, every one at least slightly different from all the others. Construction details and electrical principles of many popular tuners have been explained in other lessons. To cover the methods of r-f alignment for all tuners, the present ones and those yet to come, we must base our work on certain general circuit principles which apply to all the varieties.

Always there is a mixer and either one or two r-f amplifiers. When there are two r-f amplifiers with a tuned coupling between them, this coupling will be of a type which could as well be used between a single r-f amplifier and the mixer. Thus it becomes possible to consider any r-f amplifier system as a single unit connected between the antenna and the mixer. The r-f amplifier tube or tubes, also the mixer tube, may be either a triode or a pentode. The number of elements in these tubes makes no difference so far as alignment is concerned.

A great number of tuners contain the circuits shown in simplified form by Fig. 89-11. Here we have a tuned coupling between the antenna and the grid-cathode circuit of the r-f amplifier, and another tuned coupling between the plate of the r-f amplifier and the mixer grid. The three coupling inductors shown in this diagram could be variably tuned for each channel by means of three rotary selector switch sections having on each rotor either a single contact tongue or a shorting segment.

So far as alignment methods are concerned, this diagram may represent also the class of tuners in which the coupling elements are sections of shorted resonant lines. In this case each of the three sets of inductors would consist of a pair of similar strings instead of the single strings shown. There would be two switch sections and two rotors for each pair of inductors. Provided the two similar adjustments on each section of each tuned line are altered in the same manner and by equal amounts, to preserve the electrical balance, alignment of resonant line tuners differs but little from procedures which will be outlined for handling single section inductors in each of the three positions.

In addition to the parts shown by Fig. 89-11 the tuner would include the r-f oscillator and its tuned circuit, but since we are dealing now only with alignment of the circuits associated with the r-f amplifier tube the oscillator is omitted from this and several following diagrams. The tuned circuit shown between antenna and r-f amplifier grid often is omitted, being replaced by an untuned coupling or by a choice of two couplings which are broadly resonant in the low band and in the high band.

Our circuit diagram illustrates an important feature which has come into increasing use during recent years, the use of trimmer capacitors Ca, Cb, and Cd. Capacitor Ca alters the tuning in the r-f amplifier grid circuit for all channels at the same time, either raising or lowering the resonant frequency on every channel. Capacitor Cb performs a similar function in the r-f amplifier plate circuit, and capacitor Cd does the same thing for the mixer grid circuit tuning.

These overall adjustments allow easy compensation for differences between interelectrode capacitances when tubes are replaced. They also allow bringing the frequency responses for all three circuits into such relations as to provide maximum possible gain, and they allow making desirable changes in overall band width.

Capacitors Cc and Ce provide coupling between the r-f amplifier plate and the mixer grid while acting also as blocking capacitors. Often you will find two or more such coupling capacitors, also inductive links and other coupling elements in this position. When a number of such elements are used the intention is to so alter the coupling as to maintain a fairly constant band width on all channels or, at least, on high-band and low-band channels. One of these coupling elements, as Cc in the diagram, often is adjustable. A weaker coupling narrows the band pass and usually increases the peak gain, while a closer coupling has opposite effects.

The circuits between r-f amplifier and mixer may be either undercoupled or overcoupled. With overcoupling the two circuits are separately tuned to the same frequency, usually to a frequency near the center of the band or slightly higher than the center. Then the two peaks move farther apart to increase the band pass as the coupling is made closer, and come closer together to narrow the band pass as the coupling is made looser.

When there is undercoupling one of the peaks will be affected chiefly by adjustments in the plate circuit of the r-f amplifier, and the other peak by adjustments in the mixer grid circuit. However, adjustments in either or both circuits will have some effect on both peaks and on the general form of the response curve.

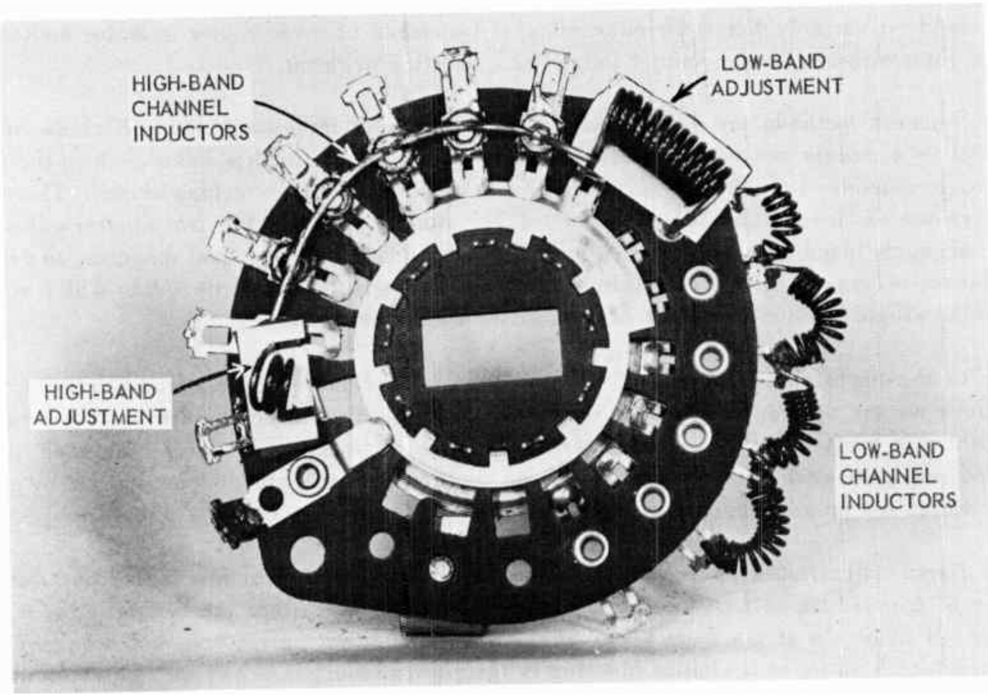


Fig. 89-12. One plate of a selector switch used for channel tuning.

In the circuit under discussion there are adjustable inductors for the high band on each of the three tuned circuits. These are inductors La for the grid circuit of the r-f amplifier, Lc for the plate circuit of this amplifier, and Le for the mixer grid circuit. There are other adjustable inductors for the low band on each circuit, these being Lb for the r-f grid, Ld for the r-f plate, and Lf for the mixer grid.

These adjustable inductors commonly consist of small self-supporting coils such as those you can see on the switch plate of Fig. 89-12. At one end of the inductor string is a coil of three turns forming the high band adjustment. Then come the sections of practically straight wire providing the necessary added inductance for tuning through high-band channels 13 to 7. Next is a coil of 13 or 14 turns forming the low-band adjustment. The remainder of the inductor string consists of four smaller coils which provide the additional inductance for tuning through low-band channels 6 to 2. The coil inductors are adjusted by either spreading the turns for less inductance or squeezing the turns back together for more inductance, all of which will be explained in more detail a little later.

Since the high-band adjustments, La, Lc, and Le of Fig. 89-11, are in circuit at all times they have some effect on all channels. But the inductances at these points are so very small they have negligible effect in comparison with the relatively large inductances in the low-band end of the string, while having a very considerable effect on all the high-band channels. The low-band adjustments, Lb, Ld and Lf, are in circuit only for the low-band channels. They affect the tuning of all channels in this band, but have no effect whatever on high-band tuning.

Later we shall learn how to alter the inductance between any two switch points, even where the inductance is provided by a nearly straight wire. When making such alterations you must keep in mind that every alteration between two high-band points will affect the tuning of every channel of lower number. For example, if you alter the inductance between switch points for channels 11 and 10 this will affect the tuning for channels 10, 9, 8, and 7 – because the altered inductance remains in the active circuit for all these other channels, but is cut out of the active circuit for channels 11, 12, and 13. The same thing is true when altering any of the inductances in the low-band end of the string, every such adjustment affects all the channels still lower down. It is for this reason that alignment always must commence at the high end of inductor strings of this general type, and proceed through to the low end in order.

Since any adjustment other than for the lowest channel in each band will affect other channels we often find that a compromise must be made by working back and forth from one end to the other in order to obtain the best average performance on all channels. If programs on one channel come through much weaker than those on other received channels, it may be advisable to make such adjustments as to favor this weak channel while somewhat reducing the response on channels where signal field strengths are greater in your locality.

The rule about beginning the alignment on the highest channel is subject to one modification. There are a number of tuners of the general type being discussed in which adjustments for channel 2 affect all other channels. That is, there is an "end inductance" for channel 2 which remains in circuit for all channels. You can check for this effect by making a slight change on channel 2 while observing the re-

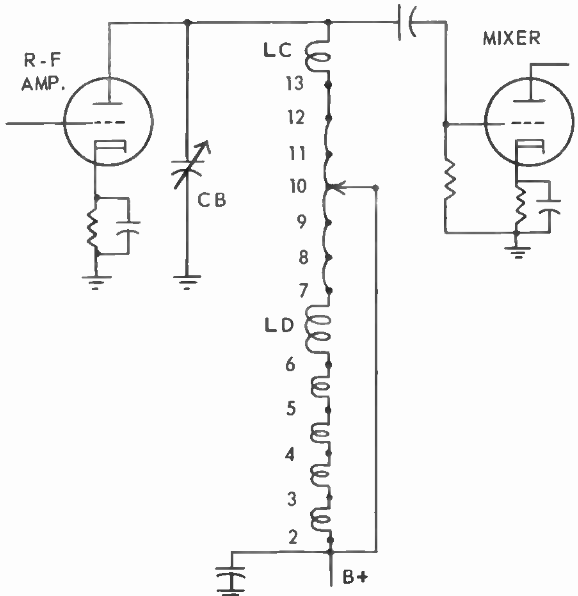


Fig. 89-13. A step-tuned impedance coupling between r-f plate and mixer grid.

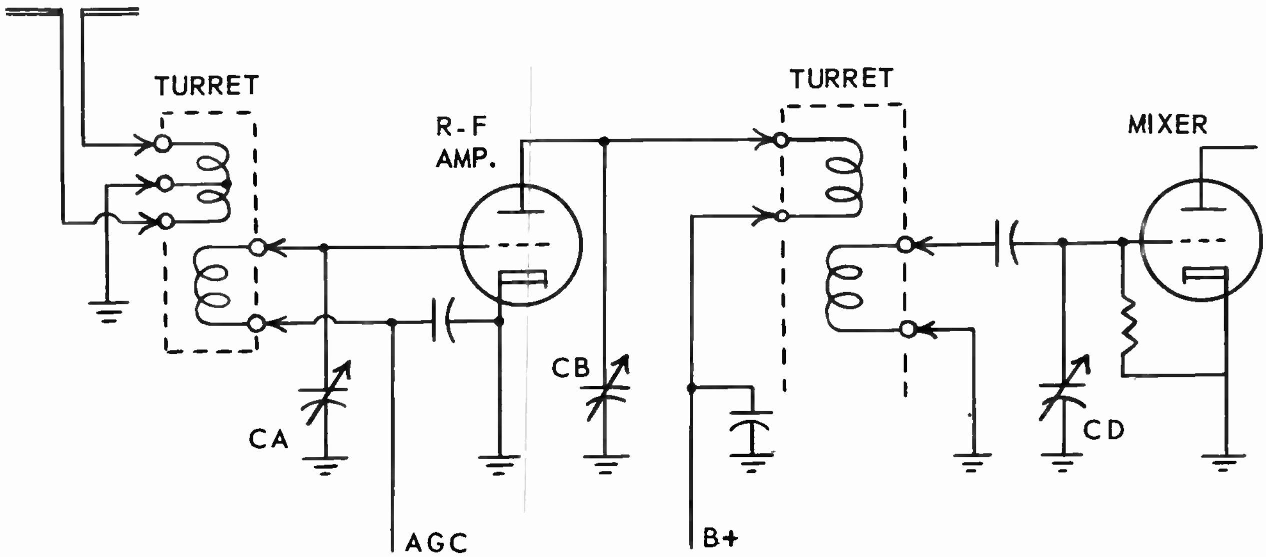


Fig. 89-14. The principal circuits in one style of turret tuner.

sponse on another channel. If your adjustment alters the response for the other channel, the work must commence with channel 2, then go to channels 6 to 3 in order, and finally to channels 13 to 7 in order.

It has been mentioned that the inductor string for the grid of the r-f amplifier often is omitted from the tuner. Further simplification is possible, as illustrated in Fig. 89-13, by using only a single inductor string as a tuned impedance coupling between the plate of the r-f amplifier and the grid of the mixer. Everything that has been said about band adjustments and individual channel adjustments for two or more strings applies where only one is used. Inductor Lc forms an adjustment for the entire high band, inductor Ld is for the entire low band, the small sections of inductance may be altered to affect individual channels, while trimmer capacitor Cb may be provided to bring the entire tuned coupler into alignment with television carrier frequencies.

Fig. 89-14 represents the elementary circuit for one style of turret tuner wherein an entirely different set of tuning inductors is brought into the active circuits by rotating the turret to the position for each channel. The important thing to note here is that again we have a tuned circuit for the grid of the r-f amplifier, another tuned circuit for the plate of this amplifier, and a third tuned circuit for the grid of the mixer, just as in Fig. 89-11. Also, we have the three overall trimmer capacitors Ca, Cb, and Cd, for the three tuned circuits.

In the type of turret tuner illustrated there is inductive coupling from antenna to r-f grid, and inductive coupling between r-f plate and mixer grid. In other turret tuners there might be capacitive couplings or combinations of inductive and capacitive couplings. Resonant frequencies for the coils of turret tuners may be altered by spreading or squeezing turns and by other means such as employed with any kind of tuner. Couplings and band width, also peak gain, may be altered by moving inductively coupled coils closer together for more width, or farther apart for less width and higher peaks.

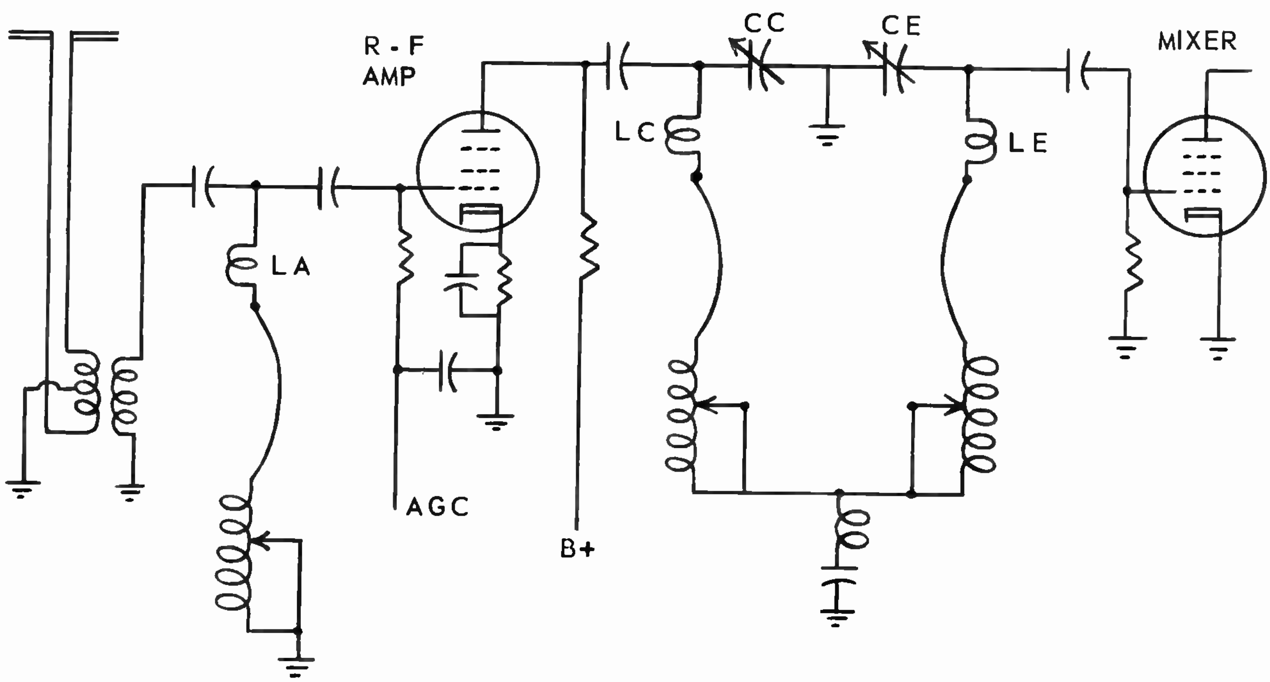


Fig. 89-15. The principal circuits for a tuner employing continuously adjustable inductors.

Fig. 89-15 represents the circuits for one kind of continuous tuner, with which the channel selector is moved not from one fixed position to another fixed position for each channel, but is turned much like the dial of a sound radio receiver until the desired channel is received with most satisfactory picture and sound. Contact sliders may move along a spiral inductor for all channels, or they may move on an inductor formed into part of a circle for the high-band channels and on a looped inductor for the low band channels -- as in some of the tuners using "printed circuits".

Once more we have the three tuned circuits, one for the r-f grid, a second for the r-f plate, and a third for the mixer grid. There are adjustable end inductors La, Lc, and Le which have their principal effect in alignment of the high band channels. Trimmer capacitors Cc and Ce allow adjustment of the degree of coupling between r-f plate and mixer grid, and thus allow altering the band width.

Fig. 89-16 represents a style of tuner in which the channels are selected by adjustment of variable tuning capacitors rather than by changing the values of inductance. Another difference between this tuner and others which have been shown is in the use of two r-f amplifier stages instead of a single stage. Between the two r-f amplifiers and also between the second r-f amplifier and the mixer we have what amounts to tuned impedance couplers, which transfer the signals from stage to stage in the same basic manner as does the tuned impedance coupler of Fig. 89-13, although the form of coupler is radically different. Inductors Lc and Le for the high-band channels may be altered by spreading or squeezing the coil turns. and this method is used also for the low-band inductors Le and Lf. Band switches for changing

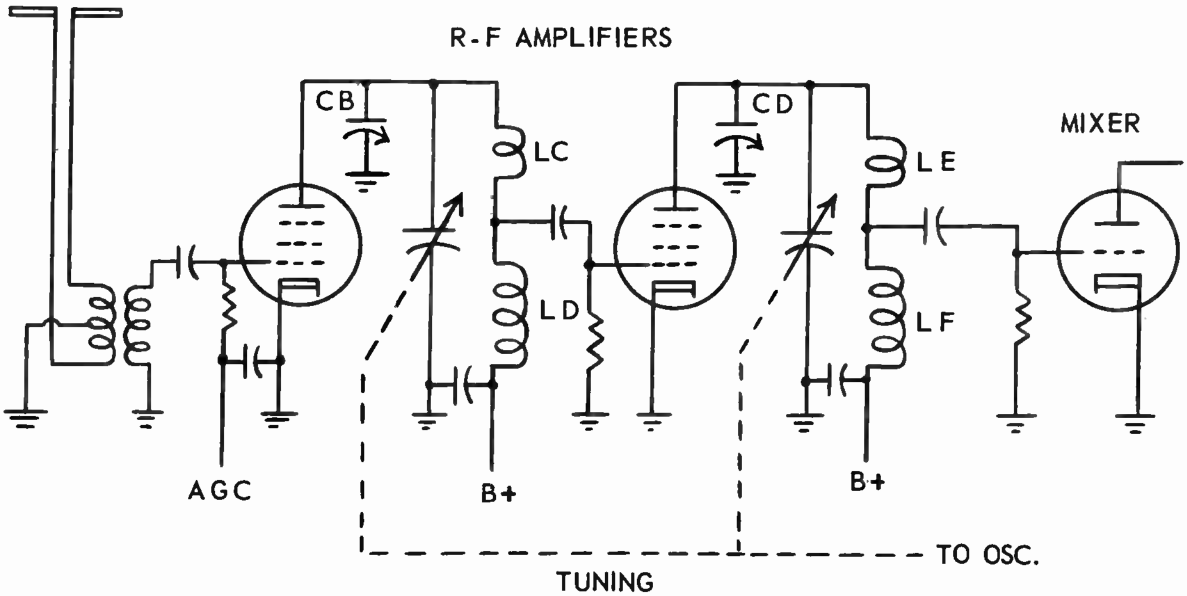


Fig. 89-16. The circuits for a tuner having two r-f stages and channel selection by variable capacitors.

the inductors are omitted from the diagram. There are small trimmer capacitors Cb and Cd which are in parallel with the variable tuning capacitors.

The types of tuners which have been illustrated by circuit diagrams are decidedly different one from the other in circuit arrangements, and would differ even more in their mechanical constructions and appearances. Yet in all of them we find essentially the same fundamental elements which may be aligned or tuned in much the same manner to produce the signal transfer, amplification, and band width which must be obtained from any tuner.

Now we may proceed to what might be called the mechanics of tuner alignment, the precautions to be observed, the manner in which test instruments are connected and used, and how it is possible to raise or lower the frequencies in circuits employing any of the usual forms of inductance and capacitance. When applying these methods you must learn to look upon the tuner itself as consisting of certain essential circuits which may differ greatly in construction and appearance, yet which must serve the same general purposes.

TELEVISION

LESSON NO. 90

TUNER ALIGNMENT

Training Your Mind
C.T.I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind

C.T.I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON NO. 90

TUNER ALIGNMENT

In other lessons we have aligned the television or video i-f amplifier, the sound system, and the r-f oscillator of the tuner. The i-f amplifier and sound system are aligned before the r-f oscillator because signal voltages employed for oscillator alignment pass through the other two systems, and they should be in correct adjustment to avoid confusing difficulties.

When both the r-f oscillator and circuits associated with the r-f amplifier are to be aligned, it is possible to perform either one of these operations before the other. If, however, voltage from a signal generator is to be introduced at the antenna terminals during oscillator alignment it usually is more satisfactory to first align the couplings for antenna, r-f amplifier, and mixer.

Many tuners are fitted with shielding covers, one example of which is illustrated by Fig. 90-1. If a shield has to be removed while making adjustments on coils or other parts there usually will be major changes in gain, band width, and peaking when the shield is replaced. Some shields have openings through which alignment tools may be inserted. When a service shop handles many of some certain make and model of receiver or tuner it is common practice to make up a dummy shield with small openings for insertion of alignment tools. The dummy is used while adjustments are made, then is exchanged for the regular shield.

If you have to make adjustments with a shield removed, the true effects can be observed only after the shield is back in place. Shielding made of sheet steel adds to the inductance and lowers the frequencies, while aluminum or other non-magnetic metal lessens the effective inductance and raises the frequencies when the shield is replaced. Cover plates on the chassis, or any other metal that comes near the tuned circuits acts like shielding, and when such parts are removed during alignment there are likely to be changes in frequency response when they are replaced.

It seldom is necessary to disable the r-f oscillator when making r-f adjustments. In some cases the oscillator frequency may show up as a small pip or dip somewhere on the response curve, but this seldom is troublesome.

A tuning wand is useful when adjusting any r-f circuit in which the inductor or coil can be reached. If the end of the wand cannot be brought to one end of the coil, or cannot be inserted into the coil form, simply hold the end alongside the tuned inductor. As always is true, if the iron end of the wand causes the desired change, the circuit should be adjusted for more inductance, while if the non-magnetic end makes an improvement the circuit needs less inductance.

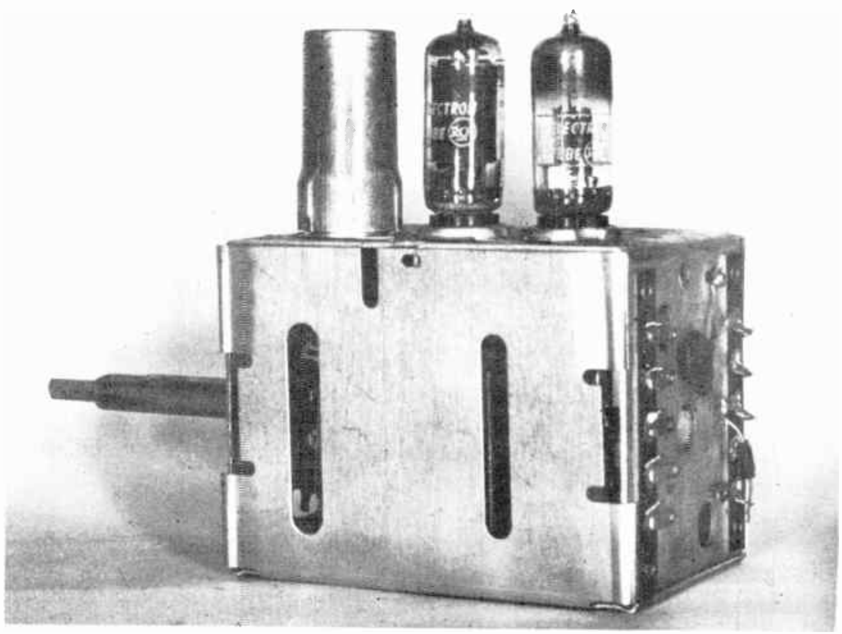


Fig. 90-1. A shielded tuner. The shield slides off the bottom of the frame.

Use only non-metallic alignment tools for tuner adjustments. Even a small metal tip on a plastic or fibre handle will alter the resonance of high-frequency circuits.

INSTRUMENT CONNECTIONS. When using an oscilloscope for r-f alignment the sweep generator is connected to the antenna terminals through an impedance matching pad.

The marker generator high-side cable is connected through a fixed capacitor of almost any convenient value to either of the antenna terminals, and the ground of the cable is connected to chassis ground. The series capacitor is to prevent making a direct conductive connection between the output controls or attenuators of the two generators. These generator connections are shown by Fig. 90-2.

The vertical input of the oscilloscope is most often connected to the grid circuit of the mixer, to which is applied the output of the r-f amplifier. This output is at carrier frequencies, which are much higher than any frequencies to which the oscilloscope will respond. But all mixers act as non-linear detectors or as partial rectifiers for the high frequencies. The mixer often is called the first detector of a superheterodyne system. Consequently there are in the mixer grid and plate circuits rectified voltages which will show up on the oscilloscope screen as a response curve for the r-f amplifier.

Many tuners have a special connection point for the oscilloscope on the mixer grid resistor, as shown in Fig. 90-2. The resistor is in two parts, *Ra* being two to five per cent of the total resistance in *Ra* plus *Rb*. This connection point often is easily accessible on the top of the tube shelf of the tuner in the form of a small wire nib or a small loop of wire. Otherwise it maybe reached from underneath the chassis. When there

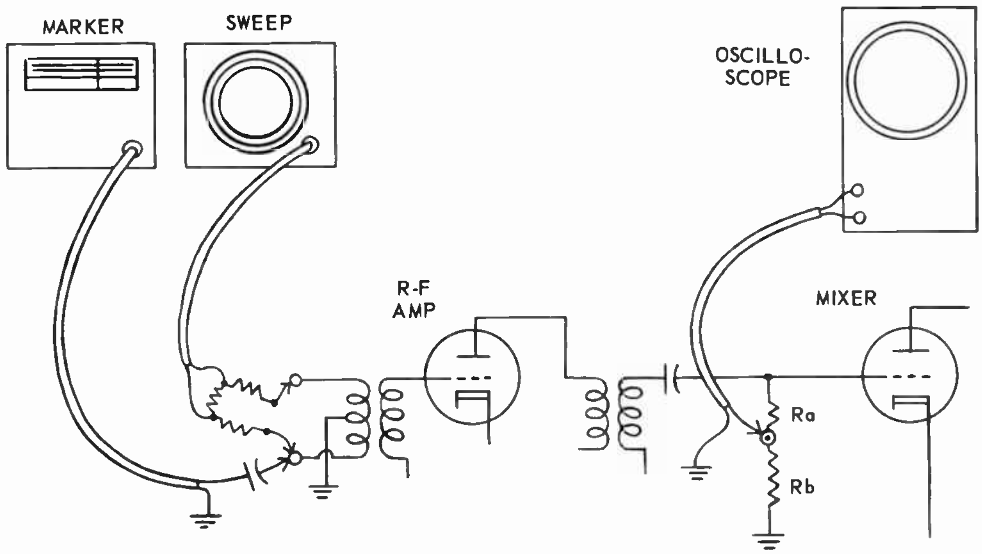


Fig. 90-2. Instrument connections for tuner alignment.

is no divided grid resistor the scope lead may be connected to the grid lug of the mixer socket through a fixed resistor of 10,000 to 50,000 ohms.

If there is no grid leak on the mixer, the grid return for this tube may pass through tuning inductors to a biasing voltage or through a resistor to ground. The oscilloscope then may be connected to some point on this grid return circuit which is lower than the tuning inductors while being above ground potential.

With the generator and scope connections as described there is only the amplification of the r-f amplifier tube. The gain is small because of the broad band tuning. For a readable trace on the scope screen the sweep generator should have an r-f output of at least 0.1 volt, and the scope must have vertical sensitivity on the order of 0.02 volt per inch or better. Even then, some tuners have so little gain on the high-band channels that response curves are difficult or impossible to obtain.

If the contrast control acts on the r-f amplifier this control should be turned to maximum or nearly so to allow all possible gain. When an agc system acts on the r-f amplifier the agc voltage should be overridden with a single dry cell providing 1½ volt bias to maintain fairly high gain. Still greater gain may be had by grounding the low end of the grid return resistor on the r-f amplifier, to allow zero bias. It is not advisable to ground the agc bus because this puts zero bias on all other controlled tubes and may allow their plate currents to become excessive.

A frequency response observed at the mixer grid circuit will be affected by the tuned coupler which is between the mixer plate and the first i-f amplifier. If this coupler is a parallel resonant type, as at the left in Fig. 90-3, it will put a sharp peak on the r-f response, as shown down below. If, as at the right, the coupler is a series resonant type it will put a sharp dip in the r-f response. If a sound takeoff is in the mixer plate circuit there will be similar effects.

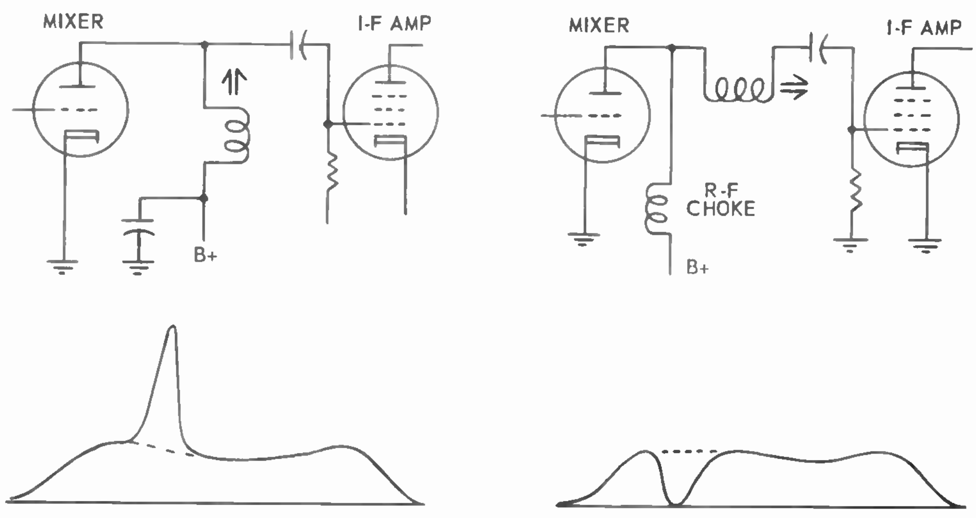


Fig. 90-3. Effects on the r-f response of a tuned coupler between mixer and first i-f amplifier.

To prevent such effects the coupler between mixer and i-f amplifier must be made non-resonant. This can be done by connecting a fixed resistor of something like 100 ohms across the tuned coil. In making such a connection do not run one end of the resistor to ground, because the other end will be at the B+ plate voltage on the mixer. Also, do not get one end of the resistor on the i-f grid side of a coupling and blocking ca-

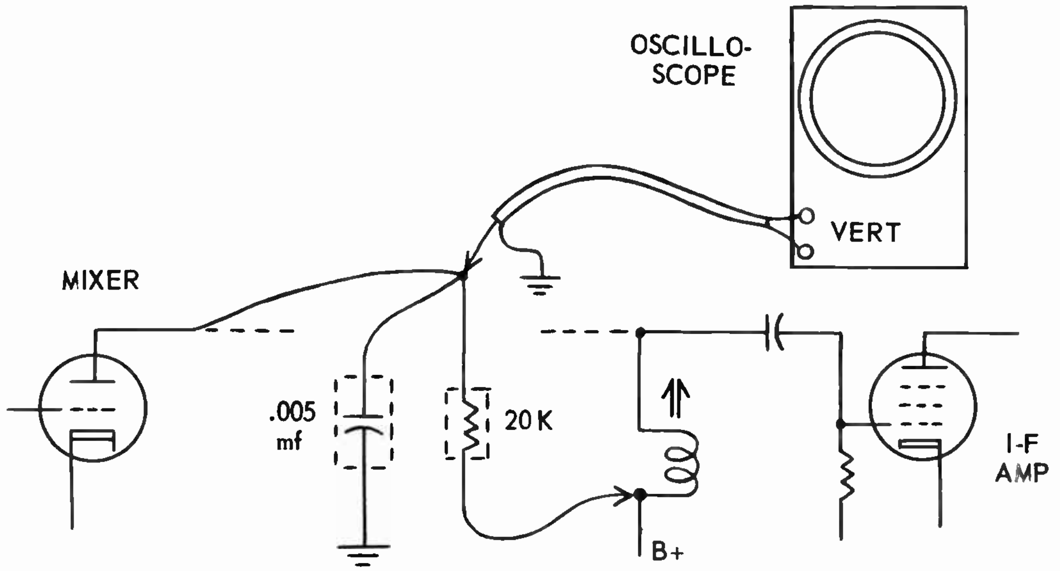


Fig. 90-4. Altering a mixer plate circuit to obtain an r-f response with greater gain.

pacitor, for this would put the high B+ voltage on the i-f grid. In many cases the detuning may be accomplished by connecting from mixer plate to ground a fixed capacitor of 1,000 mmf or more, which should throw the resonant frequency well outside the band pass of the r-f amplifier.

When your oscilloscope is not sufficiently sensitive, or the output of the sweep generator cannot be made great enough to obtain frequency response traces from the mixer grid circuit it is possible to obtain a much stronger response from the plate of the mixer tube. The mixer tube acts not only as a partial rectifier, but also as an effective amplifier for the output of the r-f amplifier.

To obtain a true r-f response from the mixer plate it is essential to remove all affects of whatever coupling is normally used between mixer and first i-f amplifier. There are various methods of doing this, but they vary between receivers and what works in one case may be useless in another. A method which has been found satisfactory with practically any receiver is illustrated by Fig. 90-4 and carried out as follows.

Temporarily disconnect the plate lead of the mixer from everything that follows it. Connect this lead through a carbon resistor of about 20,000 ohms to the point of B-plus voltage normally connected through the coupler to the mixer plate. From the mixer plate lead connect to chassis ground or B-minus a mica or paper capacitor of not less than 0.005 mf and of any value up to 0.01 or 0.02 mf. Connect the vertical input of the oscilloscope to the junction of the mixer plate lead, the fixed resistor, and the bypass capacitor, with the scope ground connections to the chassis or B-minus. The response trace now will be of the same form and band width as at the mixer grid, but much stronger or higher. When you are through making adjustments don't forget to reconnect the mixer plate lead as you found it.

The r-f response may be checked with a constant-frequency signal generator and a vacuum tube voltmeter connected as in Fig. 90-5. The generator output is connected through an impedance matching pad to the two antenna terminals of the receiver. With this connection the housing or case of the generator must be insulated from the test bench and receiver chassis. Otherwise the high-side lead from the generator may be connected to either antenna terminal, and the low side to ground. This latter connection will give a weaker response on most receivers.

The vacuum tube voltmeter is used as a d-c instrument with its high side lead connected to the mixer grid return which is above ground potential, in the same way as explained for the scope connection of Fig. 90-2. With no signal voltage from the generator the VTVM will indicate the negative grid bias on the mixer tube. This bias voltage level will be the zero level so far as response or gain is concerned, and the actual response will consist of variations from this level.

The generator frequency is changed slowly through a range of frequencies from well below to well above the limits of the channel for which the receiver selector is set. The voltage is noted at every change of a few megacycles, also at the minimum for response peaks.

The curve at the bottom of the figure was plotted in the manner described, with the receiver selector set for channel 7. In this example the tuner had relatively little compensation for band width, which accounts for the response peaks being far outside the channel limits. On the low channels this tuner or receiver showed a much narrower band. To draw a response curve the readings of voltage versus frequency may be marked on a sheet of cross section paper, then a smooth curve drawn through the check points. It is not necessary to plot a curve, because your chief interest is in the relations between response peaks and channel limit frequency, and between the peaks and the carrier frequencies for video and sound.

There is still another way of aligning the r-f circuits in the tuner when your oscilloscope is not sensitive enough to produce a readable response trace from voltages in the mixer grid circuit. This other method is

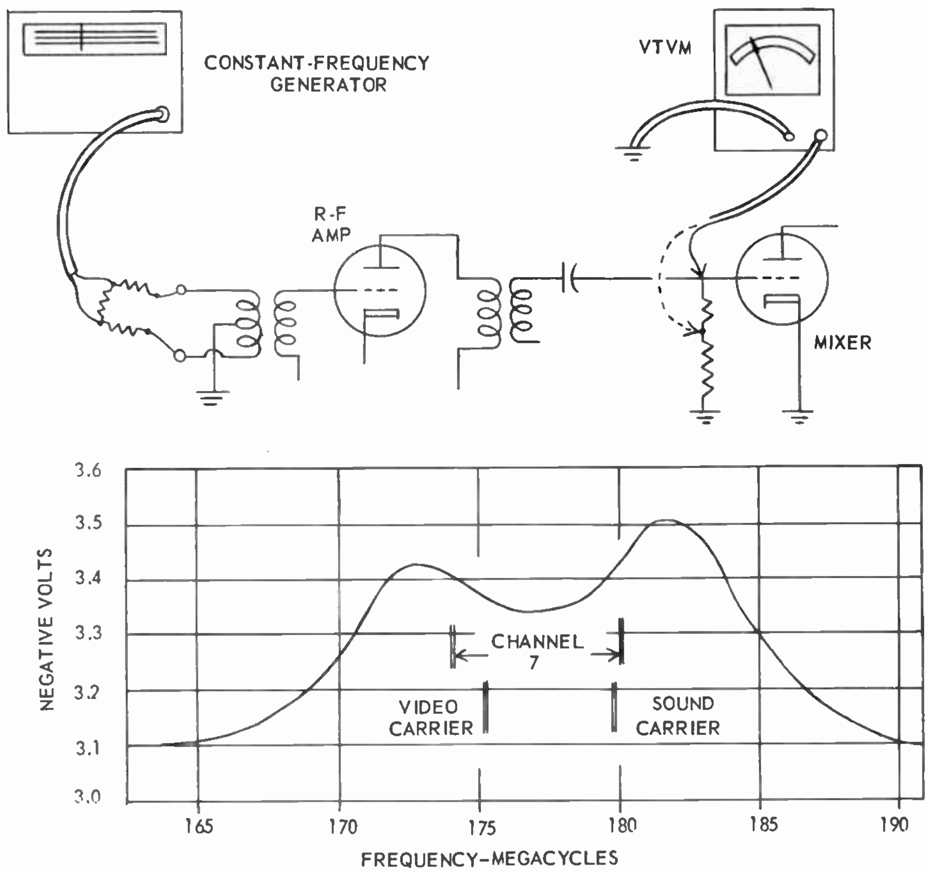


Fig. 90-5. R-f response may be checked or plotted by using the vacuum tube voltmeter.

shown by Fig. 90-6. The oscilloscope is connected across the video detector load resistor, just as when aligning the video amplifier system. At this point we always have a strong response, easily read on any oscilloscope. The sweep generator is connected through an impedance matching pad to the two antenna terminals, and is adjusted to sweep the band of frequencies for each channel being aligned or checked.

The video i-f amplifier system, from mixer to video detector, must be in correct alignment. The response curve which appears on the oscilloscope will not be the r-f response, but will be the video i-f response as affected by alignment of the r-f amplifier circuits in the tuner.

A marker signal may be introduced at either of three places. First, *A*, at one of the antenna terminals and ground, always using a series capacitor on the high-side lead. Second, anywhere on the mixer grid resistor or anywhere on the grid return which is above ground potential, as at *B*. Third, *C*, at the video detector load resistor, which is the same point to which the vertical input of the oscilloscope is connected.

A marker signal introduced at the antenna terminals, *A*, must be in the range of carrier frequencies for

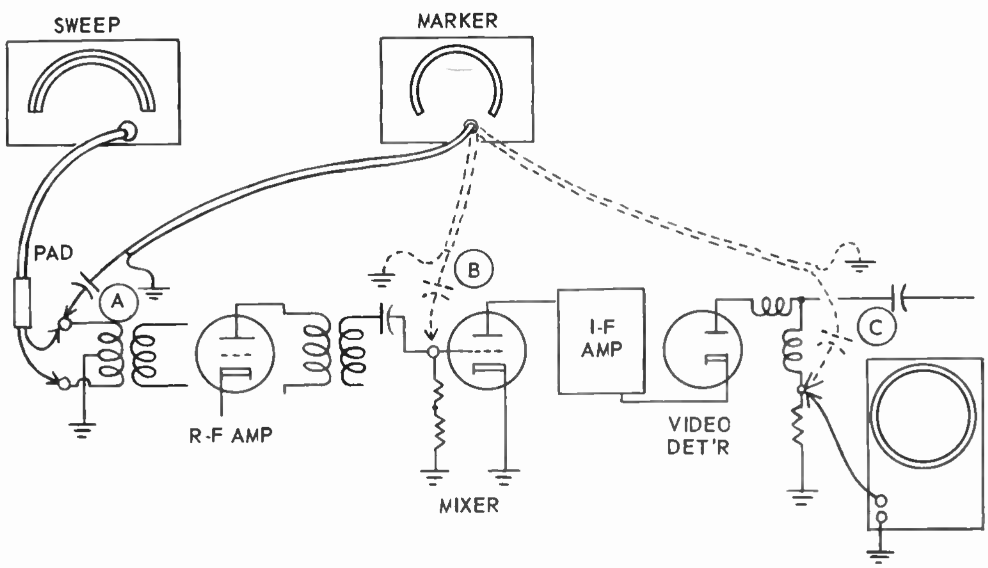


Fig. 90-6. Connections for r-f alignment with the oscilloscope at the video detector load.

the channel being checked. When observing this carrier-frequency marker pip keep in mind that the sound carrier is at a higher frequency than the video carrier, as shown at *1* in Fig. 90-7. Therefore, when you increase the marker frequency this marker pip will move from the video side to the sound side of the trace, in spite of the fact that this is a trace of i-f response being taken from the output of the video detector.

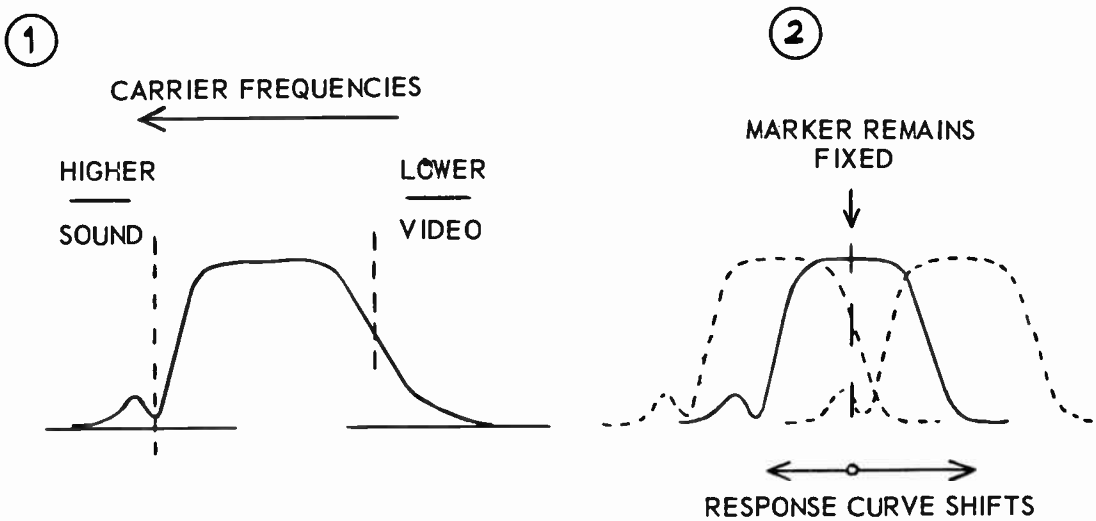


Fig. 90-7. Response frequency relations with marker and sweep generators connected to the antenna terminals.

Now you may tune the marker generator to the video carrier frequency for the tuned channel and note where this frequency falls on the i-f response. It should be at 50 to 60 per cent of maximum gain, just like a video intermediate marker. Then you may tune the marker generator to the sound carrier frequency of the same channel, and note where this frequency falls on the i-f response trace. R-f alignment in the tuner may be altered to bring these two frequencies to their correct points on the i-f response. Such alignment will change the shape of the response trace in obtaining the desired results.

When carrying out this process the frequency of the r-f oscillator must not be altered. This means that the fine tuning control must remain at the approximate mid-position where it was placed during alignment of the r-f oscillator. If you should alter the fine tuning control or in any other way change the frequency of the r-f oscillator, the entire response curve will shift to the left or right on the oscilloscope screen, but the marker pip will remain fixed. That is, you will shift the frequency response with reference to the marker pip.

This shifting of the response by altering the frequency of the r-f oscillator is shown at *2*. You might commence with the marker pip at the center of the top of the full-line response curve. Altering the oscillator frequency by means of the fine tuning control or otherwise may shift the response trace to either of the broken-line positions. But the marker pip will remain in its original position on the screen, and will be at the corresponding positions on the shifted response curves.

All these effects which occur with a carrier-frequency marker signal introduced at the antenna terminals will occur in just the same way with the marker voltage introduced at the video detector load, or at the vertical input of the oscilloscope. This is the connection shown at *C* in Fig. 90-6.

If your marker generator won't produce fundamental frequencies in the carrier range, quite likely it will produce harmonic frequencies in this range. As an example, you might want the sound carrier frequency for channel 7, which is 179.75 mc. You aren't going to get this exact frequency without crystal control or calibration, but you may come close enough to 180 mc to give useful information on the approximate position of the sound carrier on the response trace. Later, during actual reception, the small necessary correction probably could be made by operating the fine tuning control.

To obtain 180 mc as a harmonic frequency, assuming that the marker generator will produce high enough harmonics of good strength, the generator may be tuned to any simple fraction of 180. Sych fractional frequencies would include 90 mc, 60 mc, 45 mc, 36 mc, 30 mc, 22.5 mc, 20 mc, and so on. The accuracy of the carrier-frequency harmonic will be the same as that of your fundamental tuned frequency, no better and no worse. If you miss by something like 2 per cent on the fundamental, the error on the high harmonics will be 2 per cent. Should you happen to have a 4.5 mc crystal of 0.02 per cent accuracy, and a good crystal oscillator, you might calibrate against the 40th harmonic of 4.5 mc, which would be 180 mc, and this check point would come within 0.02 per cent of 180 mc.

A marker voltage introduced at the mixer grid, *B* of Fig. 90-6, should be in the intermediate-frequency range which includes the video and sound intermediates for the receiver being worked upon. Now you must keep in mind that the sound intermediate frequency is lower than the video intermediate frequency. Therefore, as you increase the tuned frequency of the marker generator the pip will move across the response trace from the sound side to the video side, as shown at *1* in Fig. 90-8. Compare this with diagram *1* of Fig. 90-7, which applies to carrier-frequency markers.

When the marker is introduced at the mixer grid, altering frequency of the r-f oscillator or changing the fine tuning control will shift the response trace to the left or right on the screen of the oscilloscope, but the marker pip will move right along with the trace. In other words, change of oscillator frequency leaves the i-f marker pip in its original position on the response. Altering the r-f amplifier alignment in the tuner

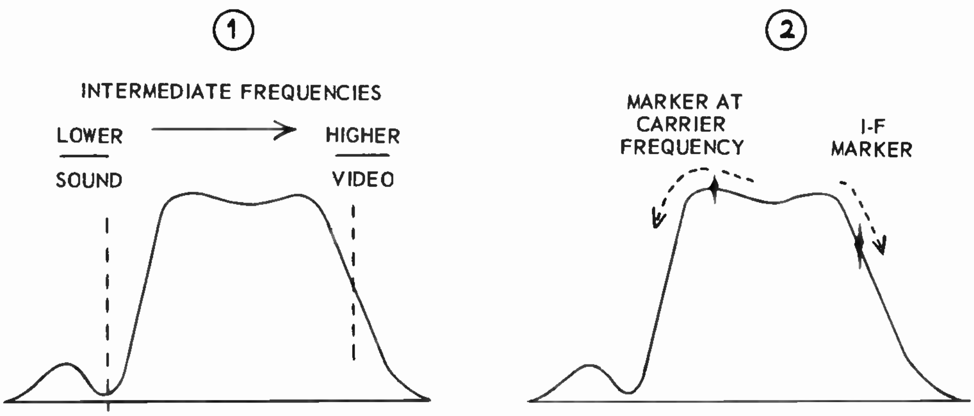


Fig. 90-8. Response frequency relations with the marker generator connected to the mixer grid.

now will alter the form of the i-f response trace. Changing the form or shape of the response will move a pip of any given frequency higher or lower with reference to peaks of the response, or will move the pip up or down on a slope of the response.

If your marker generator produces fairly strong harmonic frequencies which are high enough to get into the carrier frequency range for the tuned channel there may be the rather peculiar performance illustrated at 2 of Fig. 90-8. As a marker generator is tuned to a higher frequency the i-f pip will move toward the video side of the trace. But a carrier-frequency pip, due to a generator harmonic, will at the same time move toward the sound side of the trace. Tuning the generator to a lower frequency reverses the direction of travel for both pips. Until you recall what is happening, it is rather confusing to see two pips chasing back and forth in opposite directions as you tune the marker generator.

R-F ALIGNMENT ADJUSTERS. Now that we know how to observe and measure the results of altering the r-f response of the tuner we may consider some of the ways in which tuning of r-f circuits may be adjusted. We already are quite familiar with movable cores whose positions in their coils are changed with a non-metallic screw driver. Coils for tuning low-band channels may have cores or slugs of powdered iron. When you turn an iron core farther into its coil the inductance is increased and the resonant frequency is lowered. Turning the core farther out has opposite effects.

Coils for high-band channels often have cores of brass or other non-magnetic or weakly magnetic alloy metals. High band coils may be tuned also by brass or other non-magnetic screws mounted outside the coils but arranged to turn closer to or farther from the coils. When a non-magnetic core or screw is turned farther into or closer to its coil the effective inductance is reduced and the resonant frequency is raised, the effects being exactly opposite to those of an iron core or screw. Turning a non-magnetic core farther out of or away from its coil increases the effective inductance and lowers the resonant frequency.

In the tuner of Fig. 90-9 the r-f oscillator coils with individual movable cores are at the left. Back of the oscillator coils and at the top of the picture you will see two more coils having adjustable cores. These are the coils connected between switch taps for channels 7 and 6, or connected between the high- and low-band portions of inductances in the r-f plate circuit and mixer grid circuit. They alter the tuning for all low-band channels, without affecting the high-band channels.

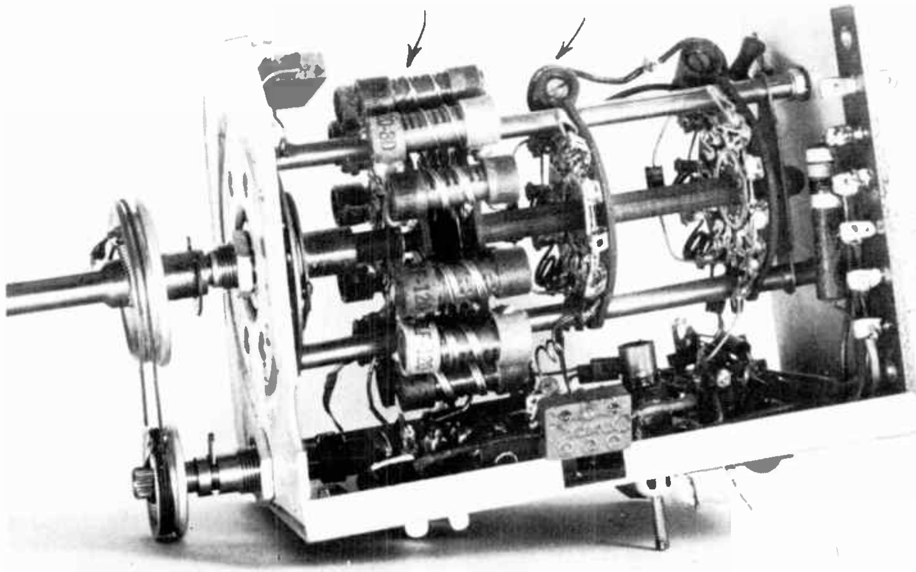


Fig. 90-9. Low-band adjusters having movable cores.

All self-supporting coils, also some of those wound on forms, may be adjusted by spreading or squeezing the turns. Coils that are susceptible to this treatment are pictured in Fig. 90-10. It is plainly evident that some of these coils have had their turns slightly spread apart. Spreading the turns farther apart reduces the inductance and raises the resonant frequency. When the turns are squeezed back closer to one another the inductance is increased and the resonant frequency is lowered.

To spread the coil turns without damaging the insulation use a small non-metallic screw driver or similar tool. This may be done while watching a frequency response or a voltmeter reading. The turns may be squeezed by using two such tools, one for support and the other for pressing. Usually it is easier to turn off the receiver and squeeze the turns with a pair of thin-nosed pliers whose tips have been wrapped with cellophane tape to protect the coil insulation. Then turn on the receiver to note the results. When working on coils supported by a form it is advisable to spread or squeeze only the end turns. The effects on resonance are the same no matter what part of a coil is opened up or pushed closer together.

There are many tuners in which the inductors consist of wire loops or "hairpin coils", of which some forms are shown by Fig. 90-11. Such inductors can be pressed or twisted to change their inductance and the resonant frequency of their circuits. If we consider the frequency to be 100 per cent with a hairpin formed as at *a*, spreading the sides as at *b* will increase the inductance and drop the frequency to about 96 per cent. With a circle, as at *c*, we have maximum possible inductance for the given length of wire, and frequency drops still lower. Squeezing the sides of the hairpin close together, as at *d*, reduces the inductance enough to raise the frequency to about 108 per cent of the original value.

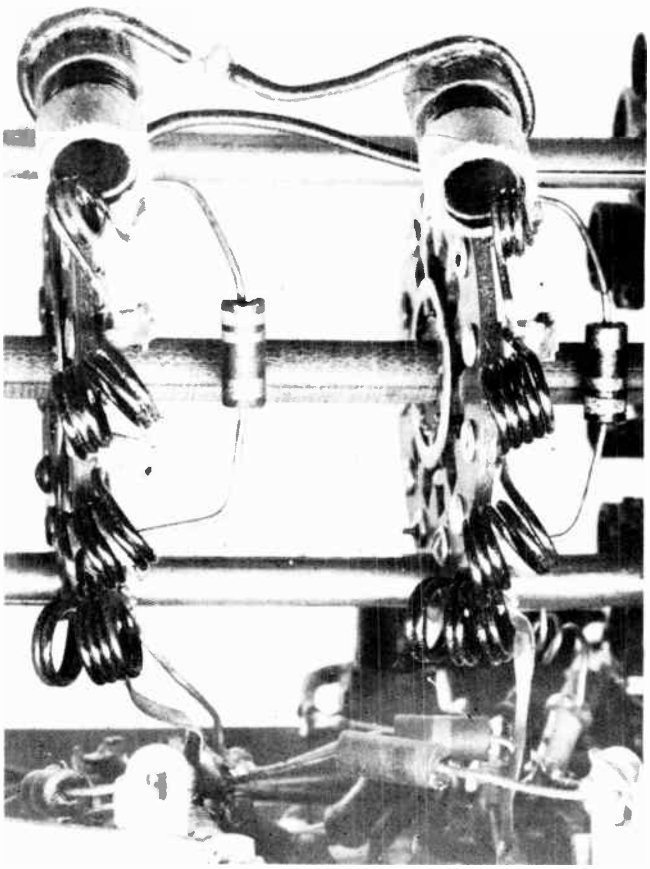


Fig. 90-10. Small self-supporting coils for channel tuning may have their turns spread apart or squeezed together.

The next four sketches *e* to *h* represent a hairpin inductor on which is a metal slider which may be moved to short circuit the portion of the loop between its closed (upper) end and the slider. Moving the slider away from the closed end of the loop reduces the inductance and raises the resonant frequency about as indicated. These sliders usually are held in position, and electrical contact is maintained, by soldering. You can loosen the solder with a hot iron while moving the slider one way or the other with fine nosed pliers or other suitable tool.

In the lower part of the figure is shown a form of inductor often used between two plates or sections of a rotary selector switch. If you find such an inductor shaped as at *i*, and consider the resulting resonant frequency to be 100 per cent, pulling the center loop toward the ends of the inductor reduces the inductance and would raise the frequency to about 105 per cent, as at *j*. Inductance is further reduced, and frequency raised, by squeezing one or both loops closer together as at *k* or *l*. For changes of shape as illustrated the frequency percentages would be about as marked.

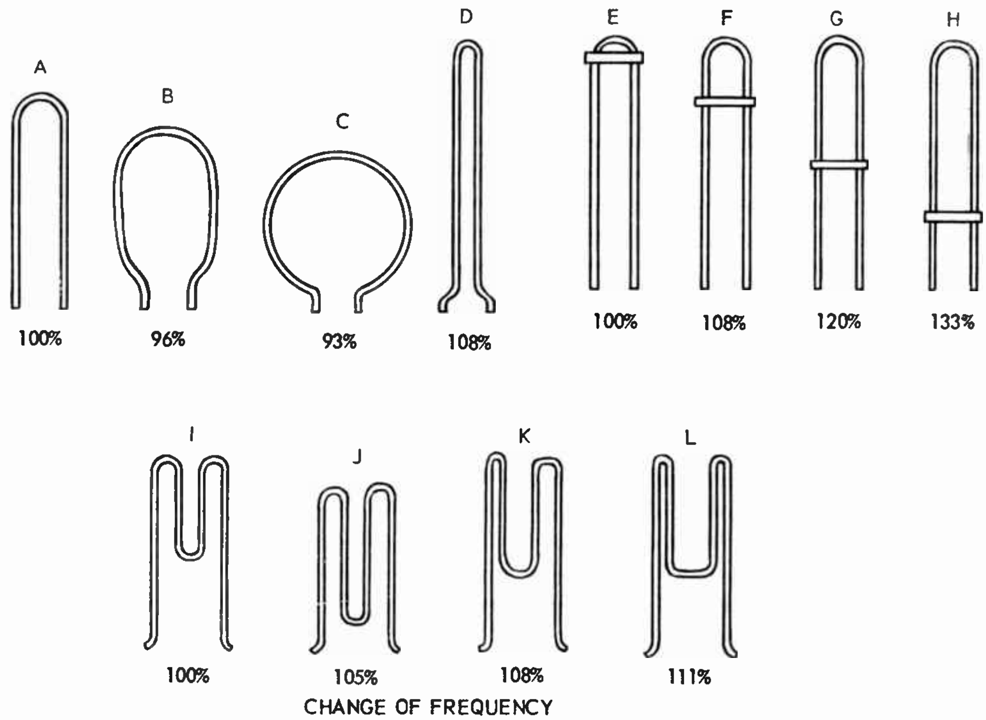


Fig. 90-11. Single-turn loops and hairpin inductors for channel tuning.

Tuning inductors of this general style are used on the turret tuner strips pictured by Fig. 90-12. The general rule for altering the inductance and frequency of any single-turn coil or loop is this. Increasing the area enclosed by the active part of the inductor will increase the inductance and lower the frequency. Lessening the enclosed area drops the inductance and raises the frequency. These results are plainly apparent in Fig. 90-11.

When working with very-high frequency circuits in tuners we must not forget that straight wires possess enough inductance to materially change the resonance point when the length of a wire is changed. A piece of number 22 gage wire one inch long contributes about 0.02 microhenry inductance to the total inductance in a circuit. If a circuit containing 20 mmf of capacitance is resonant at 180 mc, as an example, the inductance must be approximately 0.039 microhenrys. Should you add one inch of wire in this circuit the inductance would become about 0.059 microhenry, and the resonant frequency would go down to 147 mc. Removing an inch of wire would put the frequency up around 258 to 260 mc. The point is, you should not change a very-high frequency circuit by even so little as a fraction of an inch of wire – unless you wish to change the resonant frequency.

The inductance of a straight wire is affected not only by length but also by diameter. Examples are shown at the left in Fig. 90-13. With a diameter of seven thousandths (0.007) inch, each inch of wire length provides approximately 0.030 microhenry of inductance. As diameter is increased, inductance decreases. A

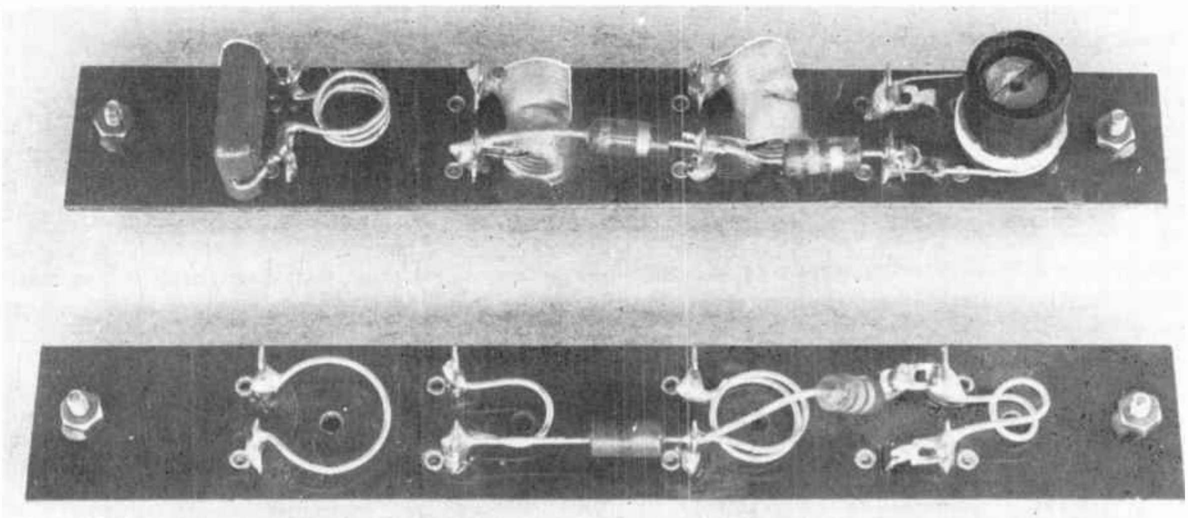
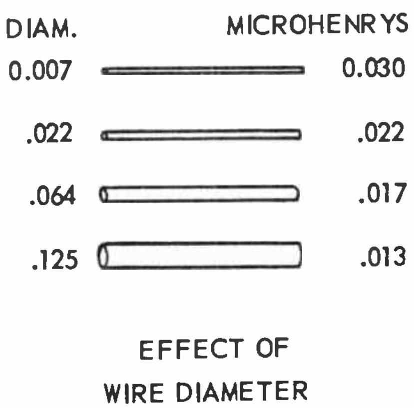


Fig. 90-12. Tuner strips having single-turn coils for some of the tuned circuits.

wire having diameter of one-eighth (0.125) inch has only a little more than 40 per cent as much inductance as the smallest size represented.

Technicians sometimes make good use of this change of inductance with change of wire diameter, in the



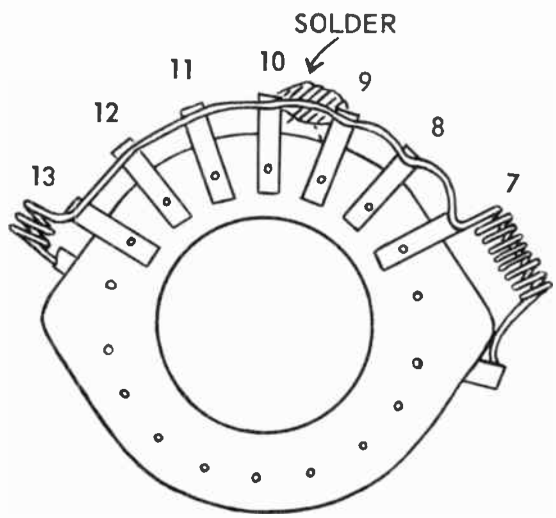


Fig. 90-13. Effect of conductor diameter in inductance (left) and a method of altering diameter and inductance (right).

manner illustrated at the right. Here we have a section or plate of a selector switch to the contacts of which is attached the inductor wire for tuning the high-band channels. The short lengths of wire between contacts may be built up with solder, or spaces between contacts may be partially filled in with solder as shown between the contact points for channels 10 and 9. The increase of effective conductor diameter would lessen the inductance for channels 9, 8, and 7. For all higher channels the lessened inductance would be shorted out of circuit, and would not affect the tuning. The built-up section will affect all channels for which it is in the active circuit, and will affect none of the channels for which it is shorted out.

Wherever the tuning adjustments are held by solder or are made by adding or removing solder it is quite a job to make an alternation. Do not forget that the tuning wand nearly always may be used to determine whether there should be more or less inductance, or more or less capacitance, before making the more permanent changes.

Adjustable cores and screws often are prevented from accidental turning by seals of various cements. A wax seal may be loosened by heating the tip of a screwdriver to a temperature no higher than that of boiling water, then holding the hot tip in the screw slot until the adjustment can be turned back and forth freely. Then the regular non-metallic tool is used for making the alignment.

Various cements require different kinds of solvents to free the adjustments. Solvents are available from radio and television supply stores. Lacquer thinner is effective for many cements, but if applied too freely or for too long it may attack the coil forms. Another solvent for many cements is amyl acetate, which usually is called banana oil.

The types of adjusters and adjustment methods which have been described in connection with r-f circuits in the tuner are used also for the r-f oscillator circuits. The same methods are employed for traps, and anywhere else that small inductances and small changes of inductance are required for tuning.

TUBE REPLACEMENTS. Tubes used in the tuner eventually require replacement because of burnouts or other defects which may develope. For routine replacements use the same type of tube and, if possible, the same make as originally in the receiver. If the new tube alters the performance for the worse in any way, and you have access to a stock of tubes, try different ones in an attempt to find one that gives acceptable performance.

Because tuned antenna circuits and tuned couplers between r-f amplifier and mixer tubes have broad frequency responses it is quite likely that a new r-f amplifier tube will cause no difficulties, but this is not always true. Curve *1* of Fig. 90-14 is the frequency response at the video detector load resistor with sweep voltage introduced at the antenna terminals of a receiver in excellent alignment. Substituting a different r-f amplifier tube changed the response as at *2*, placing the video carrier marker pip too low down and producing a peak on the sound side of the curve.

If there are trimmer adjustments for the grid and plate circuits of the r-f amplifier these adjustments usually will correct a condition such as shown by diagram *2*. If the fault is apparent only on channels of one band, the alignment adjustments for that band may be changed.

A new mixer tube ordinarily makes only slight changes in the overall frequency response. A typical effect is shown by curve *3*. A trimmer adjustment in the mixer grid circuit usually will overcome any difficulty. A change in position of the video carrier marker pip, as at *2* or *3*, may be corrected by altering the setting of a fine tuning control or by realigning the r-f oscillator for channels affected. With a dual sound system such changes of oscillator frequency may have serious effects on the sound response, but with intercarrier sound there usually is plenty of leeway.

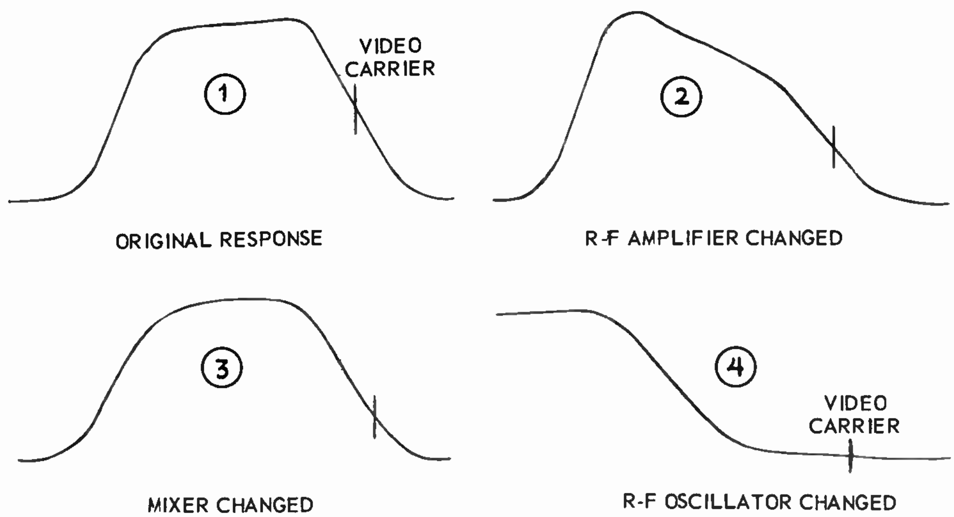


Fig. 90-14. The shape of the r-f response and the positions of marker pips may be altered by changing the tubes.

Replacing the r-f oscillator tube, or a combined oscillator-mixer tube, nearly always will change the oscillator frequency enough to shift the response curve up or down in frequency, and to displace both the video and sound carrier frequencies. In most cases a correction may be made by altering the fine tuning control or, better, by realigning the oscillator tuning when this is not too inconvenient. A new oscillator tube sometimes will shift the response so far that the fine tuning control cannot bring it back. This is illustrated by curve *4*. Then the oscillator circuit must be realigned if the particular tube is the only one available.

PREAMPLIFIERS. A preamplifier or television "booster" is a device containing one or more carrier-frequency amplifying stages arranged for connection between the transmission line and the antenna terminals of a receiver to increase the signal input. Preamplifiers are used in areas where signal field strength is low, or they may be used anywhere with receivers having insufficient gain on some or all channels.

There is as great variety in circuit and construction details of preamplifiers as of television tuners. Many of the common features are illustrated by Fig. 90-15. Amplifier *A* is used for high-band channels and amplifier *B* for low-band channels. Switching is handled by a rotary selector switch having five sections, numbered from *1* to *5*, with three positions or three contact points on each section. The transmission line is connected to the rotors of switch sections *1* and *3*. The rotors of sections *2* and *4* are connected through a short piece of added transmission line to the antenna terminals of the receiver.

In the grid circuit of each amplifier tube is an antenna coupling transformer, and in each plate circuit is an output transformer. The grid and plate transformers are tuned for the various channels in each band by movable cores operated together from a tuning knob or dial. Each of the movable cores is individually adjustable for alignment, and in each grid circuit is a trimmer capacitor used for alignment. Similar trimmer capacitors may be used in the plate circuits.

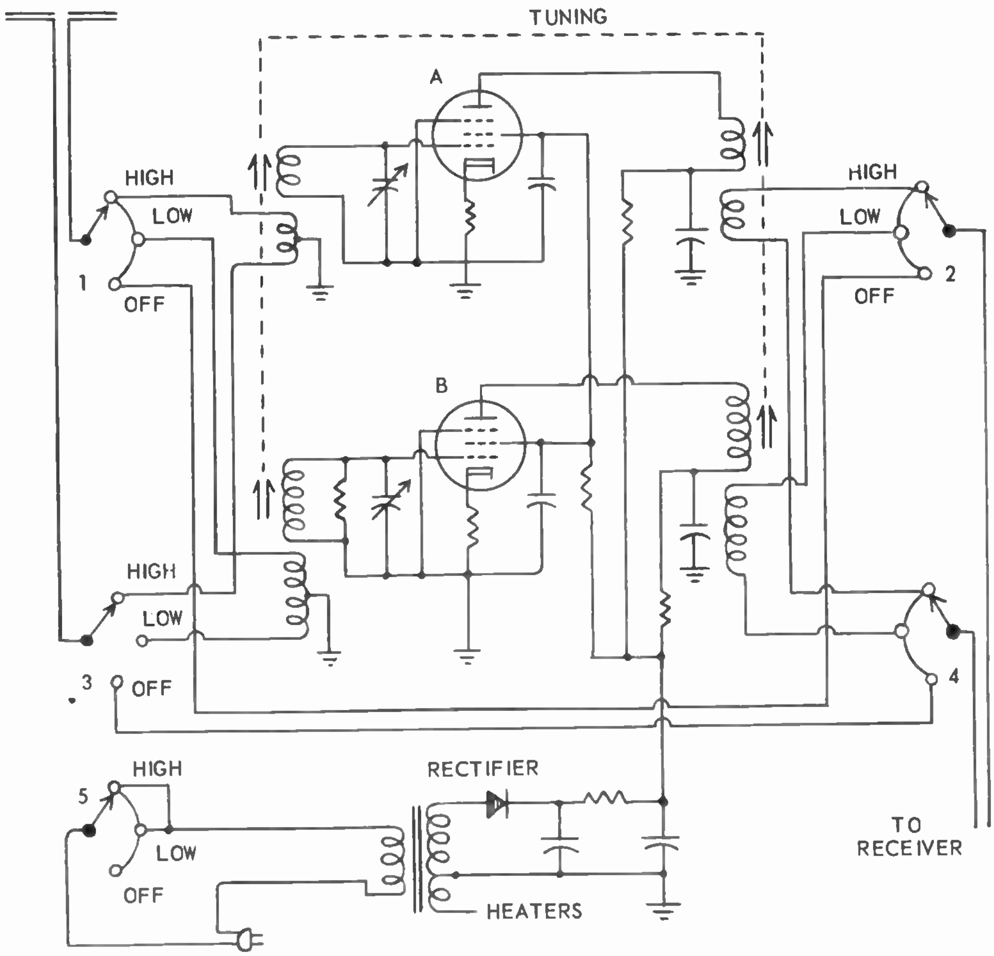


Fig. 90-15. Circuits such as found in preamplifiers or television boosters.

Selector switch sections are shown in their high-band positions, with which the transformer windings for amplifier *A* are connected to the antenna and receiver, while windings for amplifier *B* are disconnected. In the low-band positions the transformers for amplifier *B* are connected, and those for amplifier *A* are disconnected. In the off positions the two conductors from the regular transmission line are directly connected through the preamplifier to the length of transmission line going to the receiver, and the transformers for both amplifiers are disconnected.

The power supply, shown at the bottom of the diagram, contains a half-wave selenium rectifier and a resistance-capacitance filter. The power supply is controlled by switch section 5. In the high-band and low-band positions the primary of the power transformer is connected to the line so that voltages and currents are applied to the plates, screens, and heaters of both tubes. In the off position the transformer primary is disconnected from the line, making both tubes inactive.

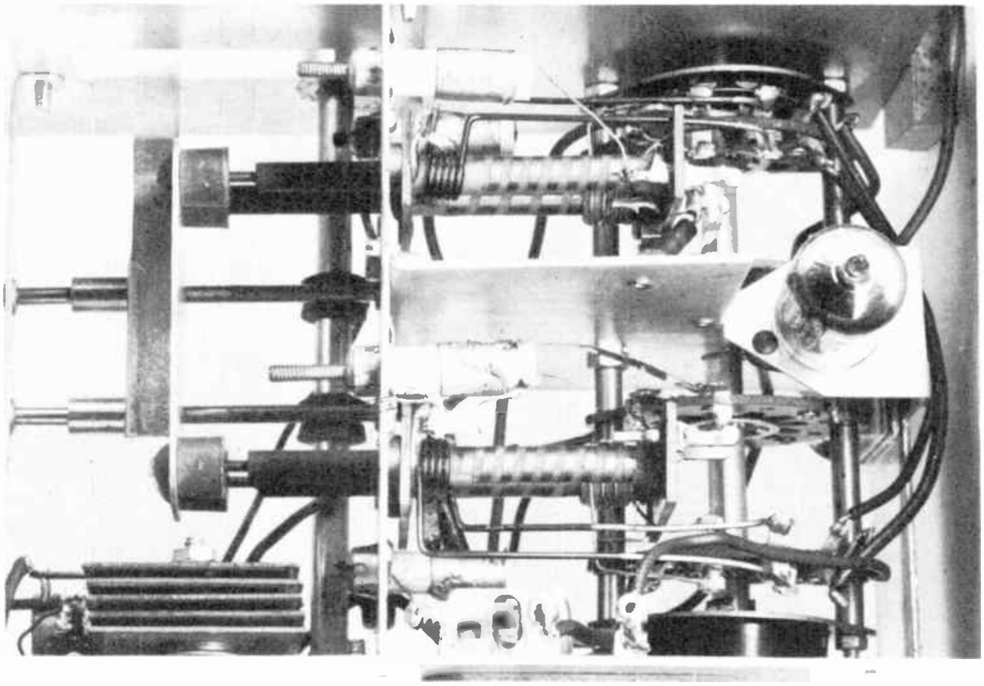


Fig. 90-16. Construction of a single-tube preamplifier having movable cores for channel tuning.

In many preamplifiers a single amplifier tube is used for both bands. This requires, in addition to the switch sections of Fig. 90-15, additional sections for connecting the grid and plate of the single tube to either of the transformers. Fig. 90-16 is a picture of the interior of a preamplifier having only one tube, and employing movable coil cores for tuning all the transformer windings. The amplifier tube or tubes may be either pentodes or triodes. As a general rule both the grid circuit and plate circuit are tuned, although in some designs there is adjustable tuning for only the grid or only the plate.

Instead of tuning the circuits with movable cores for the coils there may be a multisection variable tuning capacitor. Then the core adjustments are used only for alignment. A number of preamplifiers are tuned for each channel by means of tap switches, on the order of the one for which a section is shown in Fig. 90-13. Still others have continuous tuning by means of sliders rotated on coiled or spiral inductors. As mentioned before, there may be two amplifying stages, or a single stage may be used for the low-band channels where signals are relatively strong, and two stages for the high-band channels with their weaker field strengths.

It is important to have the best possible impedance match between the output of a preamplifier and the antenna terminal input to the receiver. The match sometimes may be improved by trying different lengths of connecting line, by transposing the two concuctors either at the preamplifier or the receiver, or by using a matching stub at either end of the connection. A good impedance match is important also between the regular transmission line and the input to the preamplifier. Some of these devices have impedance matching transformers. With others it may help to use a matching stub. Transposing the line conductors at the preamplifier may improve the results.

U.S. Civil Service
Regional Office
433 W. Van Buren 77, ILL.

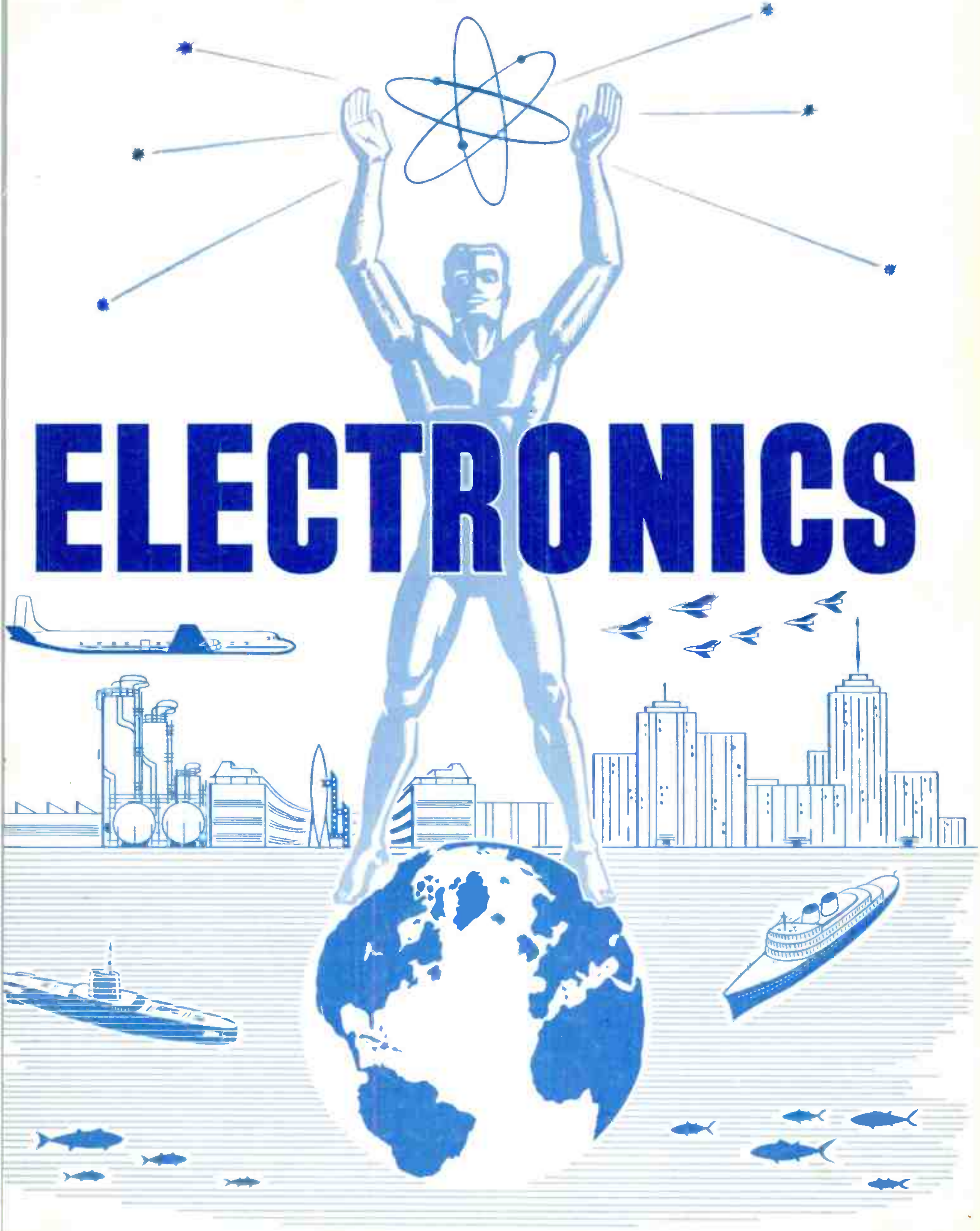
Training Your Mind

C.T.I.

Perfecting Your Skill

91→95

ELECTRONICS



COMMERCIAL TRADES INSTITUTE • CHICAGO 26, ILLINOIS

TELEVISION

LESSON NO. 91

DETECTOR PROBES FOR SERVICE OPERATIONS

"Training Your Mind"

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON NO. 91

DETECTOR PROBES FOR SERVICE OPERATIONS

A number of important service operations cannot be performed satisfactorily with either the vacuum tube voltmeter or the oscilloscope without the help of a detector probe of the kind pictured in Fig. 91-1 or something generally equivalent. When a detector probe is connected to a circuit carrying a signal-modulated high-frequency voltage, the signal or the modulation is obtained from the probe output. A probe is roughly the equivalent of a video detector which may be connected to any circuit, and which will deliver the modulation of a signal existing in that circuit. The modulation from the probe may be applied to a vacuum tube voltmeter, an oscilloscope, or other test instrument.

Within the tubular shielded probe is the detector element, usually a germanium crystal diode, also capacitors and resistors required to complete the demodulation circuit. A high-frequency modulated signal voltage usually is taken into the probe through an insulated metallic point or clip which is applied to the circuit being checked. There is also a ground lead, usually fitted with a spring clip. The output from the probe passes through a shielded cable to a phone plug or other connector which fits the input terminals of the indicating instrument.

Doubtless you have noticed that in none of the service operations previously described has the oscilloscope been connected other than to a point where there is a demodulated or partially rectified signal, such as to the video detector to some point following the video detector, to the sound demodulator, or to a sound limiter. If you connect the oscilloscope directly to the grid or plate of an r-f amplifier or to either a video or sound i-f amplifier there will be no useful trace. This is because no ordinary oscilloscope has appreciable vertical gain at frequencies so high as in these amplifier circuits.

The d-c input of the vacuum tube voltmeter has been similarly connected only where there is a demodulated signal voltage. This is because the a-c input of the usual VTVM is of such low impedance as to virtually short circuit any high-frequency signal voltages, and because the d-c input, while of high resistance, is quite useless for measurement of alternating voltages. A suitable detector probe will demodulate the high-frequency signals and filter the remaining alternating voltage to allow measuring the rectified direct voltage on the d-c ranges of the VTVM.

At 1 in Fig. 91-2 is the circuit for a detector probe adapted for observation of television signal waveforms on the oscilloscope. Note that the crystal rectifier X is shunted across the input and ground sides of the circuit. Blocking capacitor Ca prevents any d-c voltages in the tested circuit from entering the probe, but freely passes the alternating signal voltages. The crystal is paralleled by load resistor Ra. Resistor Rb is in series with the high-side lead that goes to the vertical input of the oscilloscope. The shield of the cable is connected to the probe shield and to the ground lead. The same basic demodulator

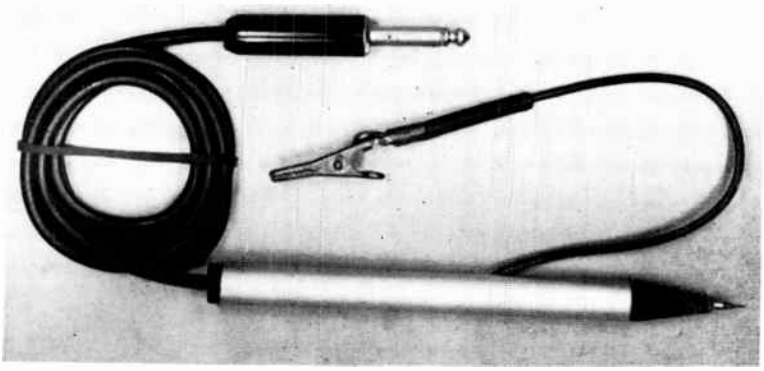


Fig. 91-1. A detector probe with its cable.

circuit may be used to observe frequency responses on the oscilloscope or voltages on the vacuum tube voltmeter, but for such use there usually is an added filter capacitor Cb of diagram 2. Typical values for the various elements are as follows.

	FOR SIGNAL WAVEFORMS	FOR FREQUENCY RESPONSE AND VTVM
Blocking capacitor Ca.	100 to 500 mmf	500 to 5,000 mmf
Load resistor Ra.	1,000 to 15,000 ohms	100K to 1 megohm
Filter resistor Rb.	5,000 to 20,000 ohms	100K to 1 megohm
Filter capacitor Cb.		None to 500 mmf

In Fig. 91-3 the crystal is connected in series with the line that goes from probe input to output. For waveform observation there ordinarily is only a single filter capacitor, Cb in diagram 1. When the probe is to be used chiefly for observing frequency responses or for use with a vacuum tube voltmeter there may be an added filter capacitor, Cc of diagram 2. Values for the elements may be as follows.

	FOR SIGNAL WAVEFORMS	FOR FREQUENCY RESPONSE AND VTVM
Blocking Capacitor Ca.	20 to 500 mmf	500 to 5,000 mmf
Load resistor Ra.	300 to 10,000 ohms	100K to 1 megohm
Filter capacitor Cb	5 to 50 mmf	500 to 2,000 mmf
Filter capacitor Cc.	none	20 to 50 mmf

Most satisfactory values in ranges listed for waveform observation will depend largely on the characteristics of your oscilloscope.

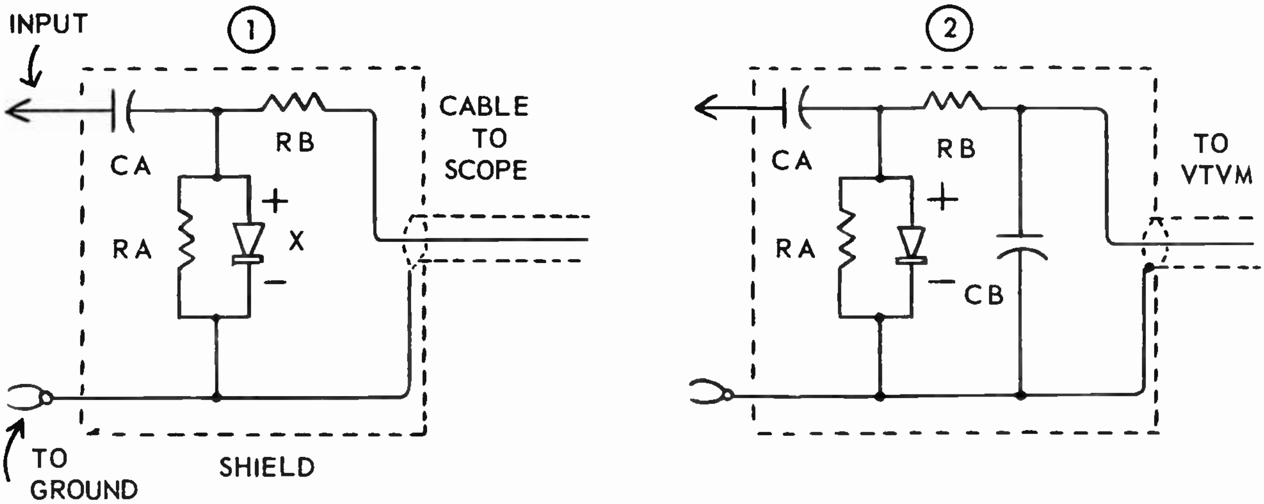


Fig. 91-2. Probe circuits with crystal detectors shunted across the input.

When used with a fairly sensitive oscilloscope the probe circuits of diagrams 1 in both figures will allow observation of television signal waveforms in both the vertical and horizontal scanning intervals when connected to video i-f amplifier circuits anywhere from the mixer output to the video detector input. Unfortunately, these probe circuits which are well suited to waveform observations are seldom satisfactory for observing frequency responses, their tendency being to produce a trace with a single rather sharp peak when the true response may be broad and formed with two peaks. The probe designs of diagrams 2, and the element values listed for them, are satisfactory for observing frequency responses as well as for measurement of voltages with the VTVM. As between the designs with shunted and series crystals there is little to choose when each is suited to its particular application.

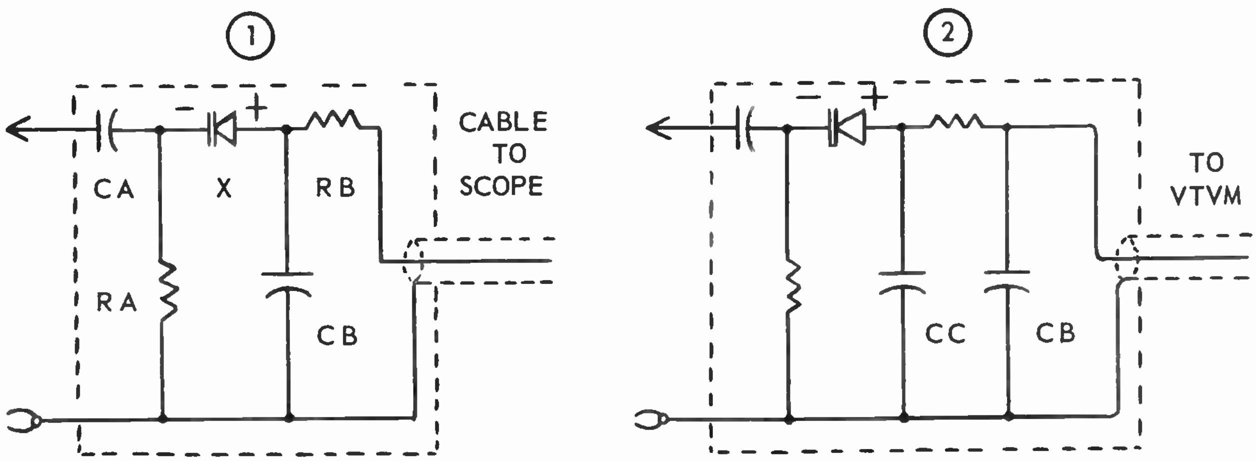


Fig. 91-3. Probe circuits with crystal detectors in series with the line.

When a detector probe is connected to the d-c input of a vacuum tube voltmeter the indicated voltages will not be correct unless the probe is designed or manufactured especially for use with the particular voltmeter. Even then the voltage readings with the probe in use are subject to all the errors discussed in connection with rectifier voltmeters in general, since the probe is a rectifier used with the d-c circuits in the meter.

Probes assembled according to designs and element values of Figs. 91-2 and 91-3 allow comparative voltage readings at frequencies between 1 and about 50 megacycles. At audio and power line frequencies the voltage readings will be much too low, because of the large reactances of input capacitors Ca at these frequencies. These low frequencies may be more accurately measured for voltage by using the regular a-c input connections of the VTVM.

Crystal diodes are small enough to easily fit within probe housings of 3/8 to 1/2 inch diameter. The other elements must be small enough to assemble in the general manner illustrated by Fig. 91-4, so that everything may be slipped into the housing after the wiring connections are complete. All resistors may be of the smallest 1/2-watt size. Capacitors are small cylindrical ceramic types with either radial or axial pigtail leads. Paper capacitors sometimes are used, but the ceramics give better results at high frequencies. The input blocking capacitor, Ca in the circuit diagrams, must have a voltage rating amply high to withstand the B-voltage in any plate circuit to which the probe may be connected. The rating should not be less than 600 volts d-c or peak.

All the elements within the probe must be shielded and the shield must be connected to the ground lead and to the cable shield. Otherwise there will be strong pickup of 60-cycle power fields and other electrical interference, while all indications will be affected by your body capacitance when holding the probe. If wiring joints and element terminals first are covered with insulation the shielding may consist of a wrapping of tin foil or aluminum foil which will be concealed within the outer housing.

The probe housing may be of plastic or fibre. If it is of metal there must be an insulating covering around the outside. Without such a covering the grounded metal might come in contact with high-voltage conductors in a receiver, and you might get a severe shock by accidentally touching such conductors while holding the grounded housing of the probe.

The ground lead with its spring clip terminal must be connected to the circuits inside the probe, as shown by the circuit diagrams. Do not use a separate ground wire or cable extending from the oscilloscope, the VTVM, or other indicating instrument. With the ground lead connected into the probe itself all high-frequency circuits are completed from the input point through this lead back to the apparatus being

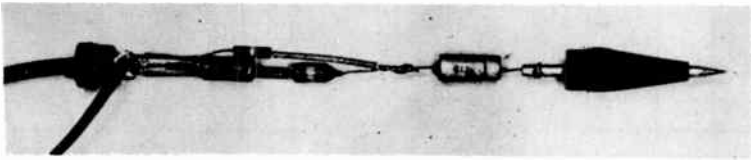


Fig. 91-4. The parts of a probe are small, and are assembled for compactness.

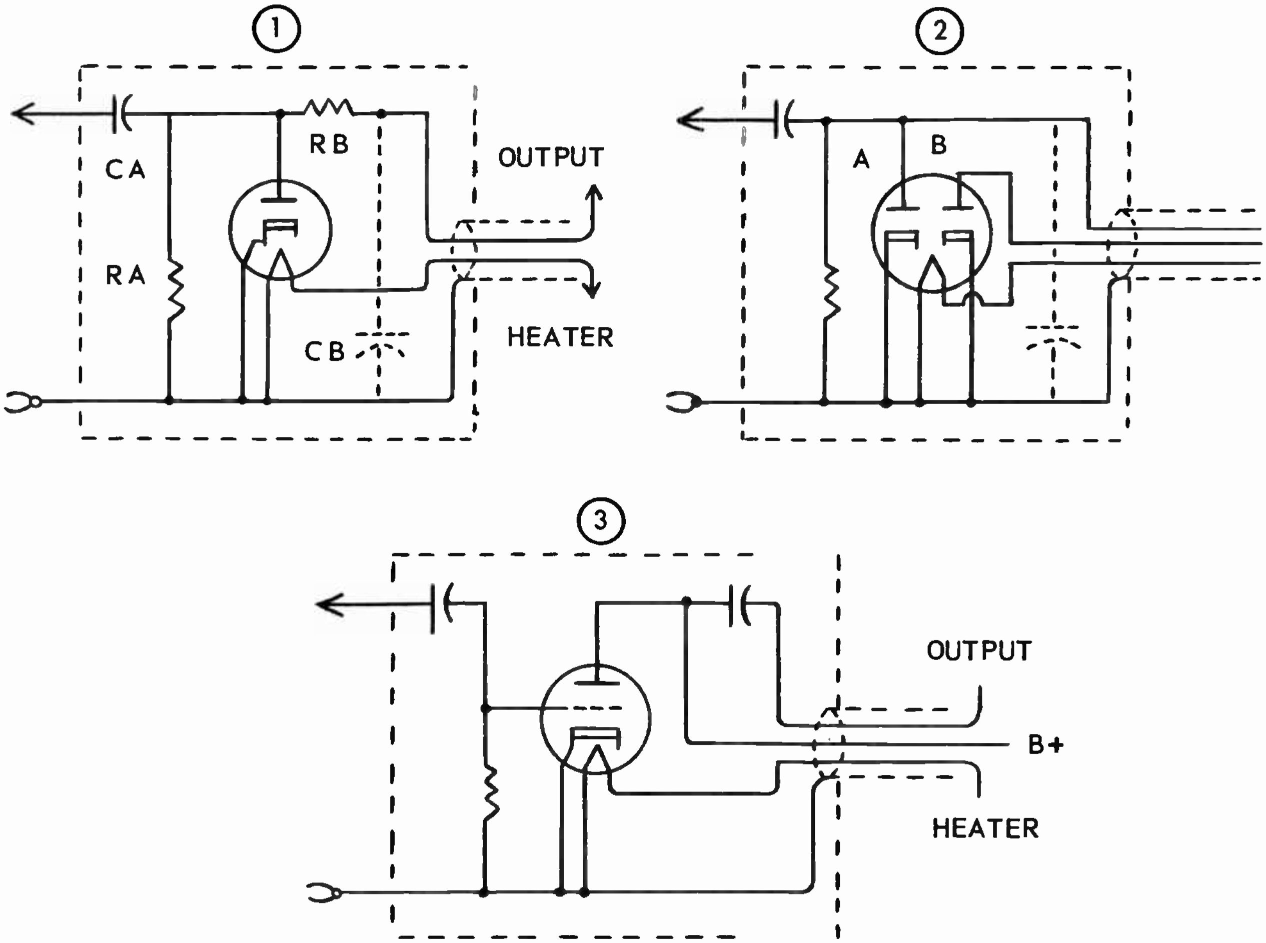


Fig. 91-5. Probe circuits in which the detectors are diode or triode tubes.

tested, and only the low-frequency demodulated currents or the rectified direct currents pass through the output cable of the probe. Otherwise the high-frequency circuits would be extended all the way through an external ground wire and the central conductor of the cable.

Probes may be constructed with miniature tubes instead of crystal diodes as the detector elements. Several detector tube circuits are shown by Fig. 91-5. At 1 there is a single-element diode, such as the type 9006 which is designed for operation at very-high and ultra-high frequencies. The circuit is generally equivalent to those of Fig. 91-2 and the same letter identifications are used for the various parts. The cable must be a shielded two-conductor type, with the shield used for one side of the heater circuit and one of the conductors for the other side. Capacitor Cb may or may not be used, depending on the applications of the probe.

In diagram 2 the tube is a twin diode. Section A is used as the detector, just as in diagram 1. Section

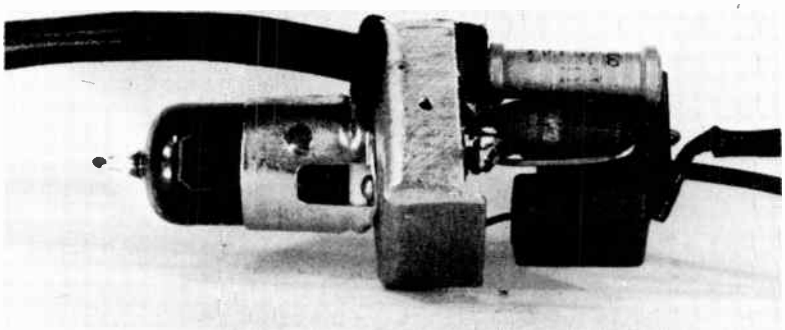


Fig. 91-6. The parts of a probe in which is a high-frequency diode tube.

B is employed for balancing out the contact potential effect of section A. The remainder of the balancing circuit, which would be in the VTVM or other indicating instrument, is like this type of circuit explained in connection with vacuum tube voltmeters. The additional diode for balancing of contact potential sometimes is located within the indicating instrument rather than in the probe.

In diagram 3 the tube is a triode, or it may be a triode-connected pentode. This tube is operated as a grid leak detector which, as you will recall, is a very sensitive detector of weak voltages and has the added advantage of providing a fair amount of amplification between plate and grid. This arrangement, like that in diagram 2, requires a shielded three-conductor cable with the insulated conductors used for the output, the B-plus voltage, and the heater current.

Fig. 91-6 is a picture of a detector probe employing a tube diode. Since the tube and its socket require a probe housing of fairly large size the capacitors and resistors need not be quite so small as with the crystal diode probes. The tube shield and base shield have been removed to allow showing the internal construction in the photograph.

DETUNING AND LOADING EFFECTS. In the wiring, capacitors, resistors, shielding, and the crystal or tube of every probe there are considerable values of capacitance and inductance when considered from the standpoint of high-frequency operation. These properties of the probe will detune any resonant circuit to which the probe is connected for testing. The higher the operating frequency the greater will be the detuning effect, as measured in megacycles off resonance. To have the indications represent actual behavior without the probe connected it is necessary to retune the measured circuit to resonance while the probe is in place. After the probe is removed the tuning must be restored as originally found. Needless to say, this tuning and retuning seldom is done during ordinary service work.

Regardless of how small may be the capacitances of capacitors and of how great may be the resistances of resistors in the probe its input impedance will be quite low at television intermediate frequencies. and very low at television and f-m carrier frequencies. There are dielectric losses in all insulating materials, the ordinary types of resistors drop their impedance to a small fraction of their d-c resistance above about

10 mc. It should be noted that the effective capacitance of a crystal diode is only about 2 mmf and the plate-cathode capacitances of tube diodes are on the order of 1½ to 3 mmf, so these elements are not responsible for much of the impedance drop.

Measurements on three probes constructed according to the circuits of Figs. 91-2 and 91-3 showed impedances at 10 mc of 680, 1,500, and 2,700 ohms, while at 25 mc the impedances were 500, 800, and 1,800 ohms. These low impedances do not cause serious trouble with measurements in the i-f and r-f circuits of television receivers, because the effective load resistances in these circuits are quite low in order to have broad band response. The loading effect of a probe is, however, great enough to require changes of the contrast control and possibly of the fine tuning in order to obtain satisfactory indications.

The capacitances and inductances of the parts and wiring in a probe will be resonant at some fairly high frequency. That is, the probe by itself is a complete resonant circuit. If the self-resonant frequency should fall within the band where you are making tests, as within the frequencies being delivered from a sweep generator, there will appear peaks and dips on the oscilloscope trace which are due to the probe rather than to the circuit being checked. Although carefully constructed probes usually have self-resonances around 100 mc, many of them are resonant in some of the television i-f ranges, and even in the i-f range for f-m receivers.

When using any probe keep the housing and shield on a line extending away from the receiver circuits

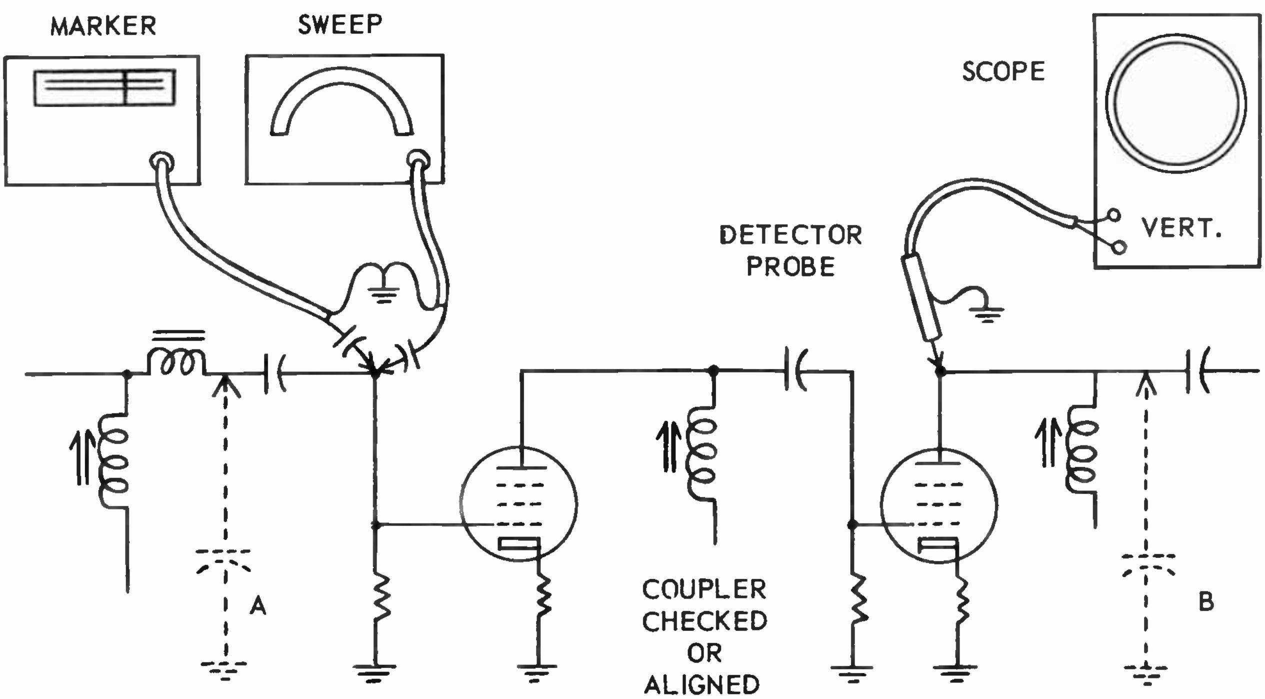


Fig. 91-7. Instrument connections for observing the frequency response of a single amplifying stage.

rather than letting it lie nearly parallel to wiring and parts of tested circuits. This precaution applies also to the probe cable. Connect the ground lead or clip to a chassis ground or a B-minus point as close as possible to the point of high-side connection, thus keeping the path for measured high frequencies well confined.

SINGLE STAGE RESPONSE. The frequency response of any one amplifying stage may be observed with the marker generator, sweep generator, and oscilloscope connected as in Fig. 91-7. The high sides of the generator cables are connected through small capacitors to the grid of the tube which precedes the coupler whose response is to be checked or which is to be aligned in producing some desired response. The vertical input of the oscilloscope is connected through a suitable detector probe to the plate of the tube which follows the coupler.

When checking the first i-f amplifying stag the generators would be connected to the mixer grid, and for all other stages to the grid of an amplifier tube. When checking or aligning the coupler that is between the last amplifier and the video detector the oscilloscope is connected without the detector probe to the high side of the video detector load resistor, but the probe must be used for all other stages.

Do not connect the generators to the plate of the tube ahead of the coupler being checked, and do not connect the probe to the grid of the tube following. Such connections would completely detune the coupler. Do not remove the tube ahead of the coupler nor the tube following it, for the tube capacitances form large parts of the tuning capacitance for the stage.

At the outer ends of the stage responses the gains will be very small, yet great enough that following amplifying stages can bring the total or overall response for these small gains up to whatever is needed at the output of the video detector. In one of the earlier i-f stages the gain at the video intermediate frequency might be about as indicated at the left in Fig. 91-8, yet in the detector output the gain at this frequency will be half or more of the peak gain, as shown at the right.

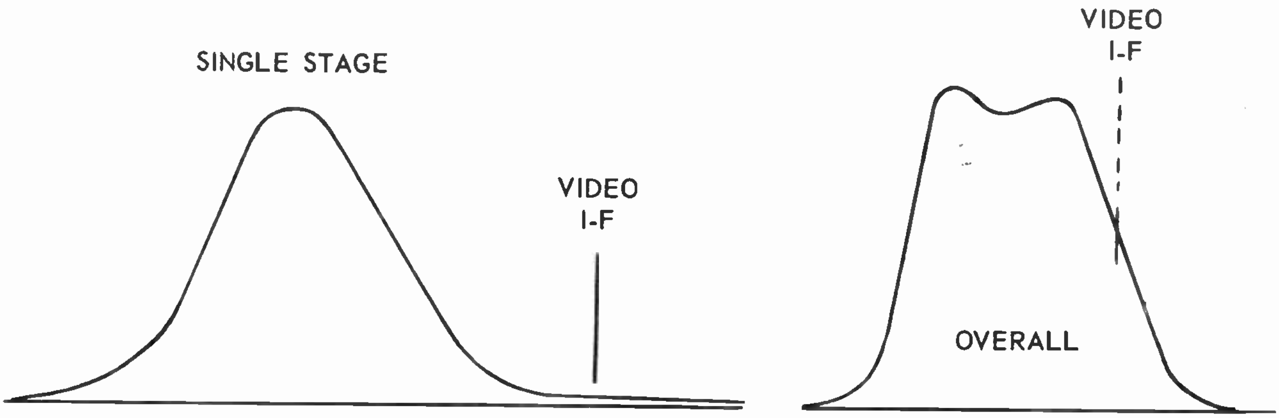


Fig. 91-8. Small gains in early amplifying stages are brought to much higher levels at the output of the amplifier system.

To check the ends of the response traces move the marker pip by tuning the marker generator, and as the pip moves down and out onto the ends of the trace keep increasing the marker output until the pip cannot be seen even with maximum output. Consider this marker frequency to be that for zero or near zero gain in the stage being checked.

Not only any single stage but any combination of stages may be examined in the same general manner. You might connect the generators to the grid of a first i-f amplifier and connect the probe and scope to the plate of the third amplifier. The resulting trace would show the combined response of the two couplers which are between the first and third i-f amplifier tubes.

It is possible also to examine the stage by stage responses by keeping the generators connected to the mixer grid and moving the detector probe successively from the plate of the first i-f amplifier to the plate of the second amplifier, then to the third, and so on. The final check would be with the oscilloscope connected directly, without the detector probe, to the video detector load.

Single stage alignment and stage by stage alignment is by far the most satisfactory way for obtaining correct adjustments of overcoupled transformers used as interstage couplers. By checking each stage, and possibly observing the stage by stage response as more and more stages are included, it is quite easy to align these couplers for approximately equal peak amplitudes and to have the peaks equally distant from the mid-frequency of the response.

Response voltages at various frequencies in any single stage or for any combination of stages may be measured or compared by using a constant-frequency signal generator, such as a marker type, and the vacuum tube voltmeter. Instrument connections are like those of Fig. 91-7 except that the sweep generator is not used and the detector probe is connected to the d-c input of the VTVM instead of to the vertical input of the scope.

As the signal generator is tuned through the range of frequencies amplified by any stage or stages the pointer of the VTVM will rise and fall in accordance with gains at each frequency. As a general rule the meter will indicate some relatively small voltage no matter to what frequency the signal generator is tuned. This may be noise voltage in the amplifier, it may be amplified voltage from stray fields, and very likely it is amplified signal voltage which enters the receiver even with the antenna disconnected. Make a test by temporarily turning off the signal generator. If the meter voltage remains steady it may be neglected and the stage gains considered as readings in excess of this minimum. If the voltage reading varies more or less continually you should connect shorting capacitors as at a and b of Fig. 91-7 ahead of and following the stage or stages to be checked. These capacitors should be of about 1,000 mmf.

If, when working with the oscilloscope as a frequency response indicator, there appear any peaks or dips which obviously should not be there it is advisable to connect the shorting capacitors before and after the circuits being checked. The peaks or dips are caused by couplers preceding and following the tested circuits.

Fig. 91-9 illustrates a method for checking the frequency response of a single stage without using a detector probe. The oscilloscope or VTVM, whichever is to be used as output indicator, is connected to the video detector load resistor and left there for all tests. The video detector provides the necessary demodulation for the oscilloscope or provides signal rectification to allow using the d-c ranges of the VTVM.

Sweep and marker generators are used when the output indicator is an oscilloscope. For the VTVM as output indicator only a constant-frequency signal generator is needed, not a sweep generator. The generator or generators are connected in the usual way to the grid of the mixer tube. All the interstage couplers except the one whose frequency response is to be checked are temporarily shorted with resistors of 300 to 500 ohms, marked R-R-R on the diagram. Be sure to get the shorting resistors across the coupler inductors or capacitors only, do not get them from the B-plus voltage in a plate circuit to ground or B-minus.

The frequeucy response now becomes practically flat for all couplers having the resistance shorts, leaving the overall response that of the one stage not shorted. This method requires a strong signal voltage from the sweep generator or the constant-frequency generator, since there are no resonant gains from the shorted couplers, although the tubes still provide amplification. Instead of keeping the generators at the mixer grid for all tests they may be moved to the grid of the tube preceding whatever coupler is being checked or aligned, this being the coupler not shorted. If more than one stage is left without a short, the overall frequency response will be that for all the unshorted stages working together. In this way it is possible to make stage by stage checks, first with shorting resistors on all couplers except the one following the mixer, then removing the resistors one by one until all are taken off and there is the overall

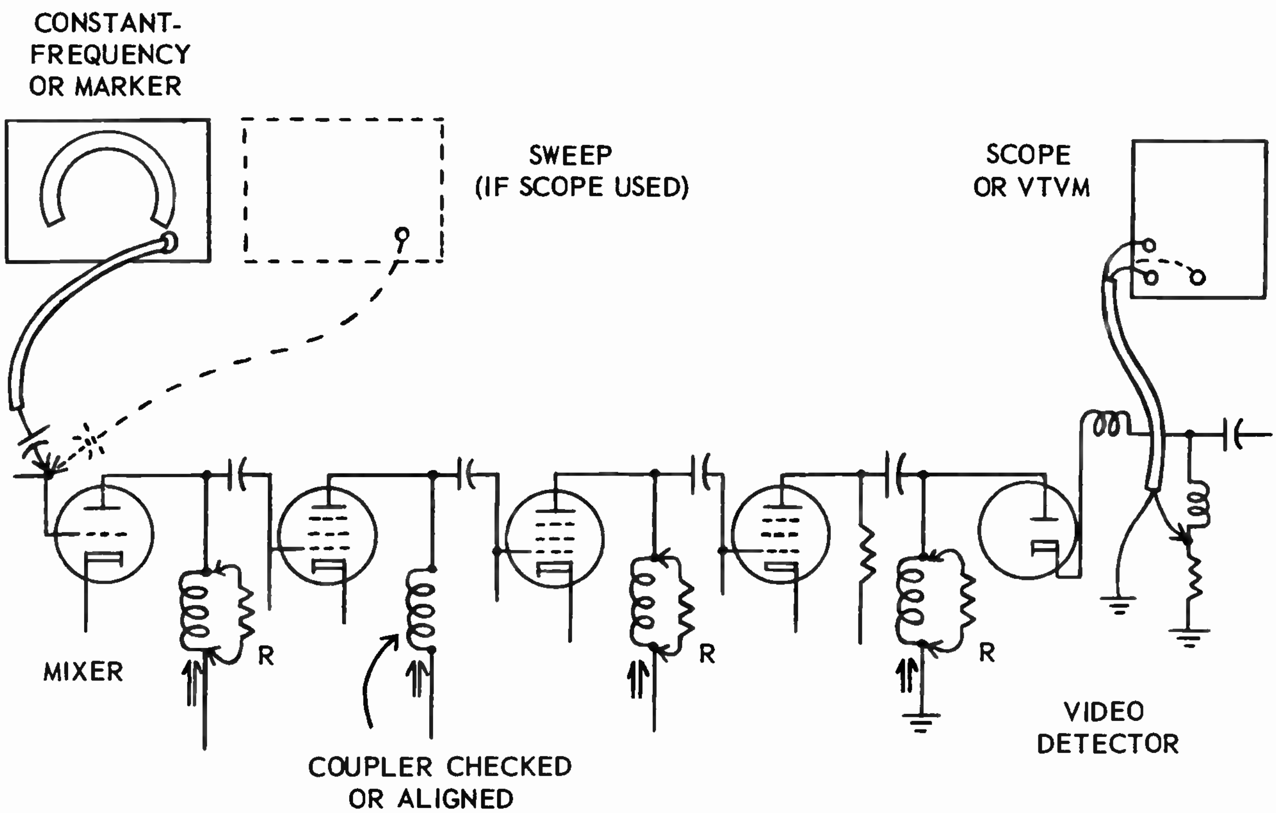


Fig. 91-9. Checking the response of a single stage with all other stages detuned by shorting resistors.

response of the entire amplifier system.

VIDEO AMPLIFIER RESPONSE. The video amplifier system, between the video detector and the grid-cathode circuit of the picture tube, has a great deal to do with picture quality because the response of this amplifier determines the relative gains at high and low video frequencies or picture signal frequencies.

A method of checking the relative gains at higher frequencies is illustrated by Fig. 91-10. To the input of the video amplifier system, which is the same as the output of the video detector, is connected an r-f signal generator operated with audio modulation. This generator should have a calibrated output, or the output should be continually measured with a VTVM and detector probe, in order that the signal fed into the video amplifier may be of constant strength at all frequencies.

The detector probe on the oscilloscope is connected to the output of the video amplifier system, which is the same as the input to the grid-cathode circuit of the picture tube. The scope is used with its internal sweep adjusted to show a trace of the audio modulation from the signal generator. Since the audio modulation at the signal generator should remain of constant percentage, any variation in the height of the trace on the scope will be due to greater or less gain for the r-f signal voltages coming from the signal generator through the video amplifier.

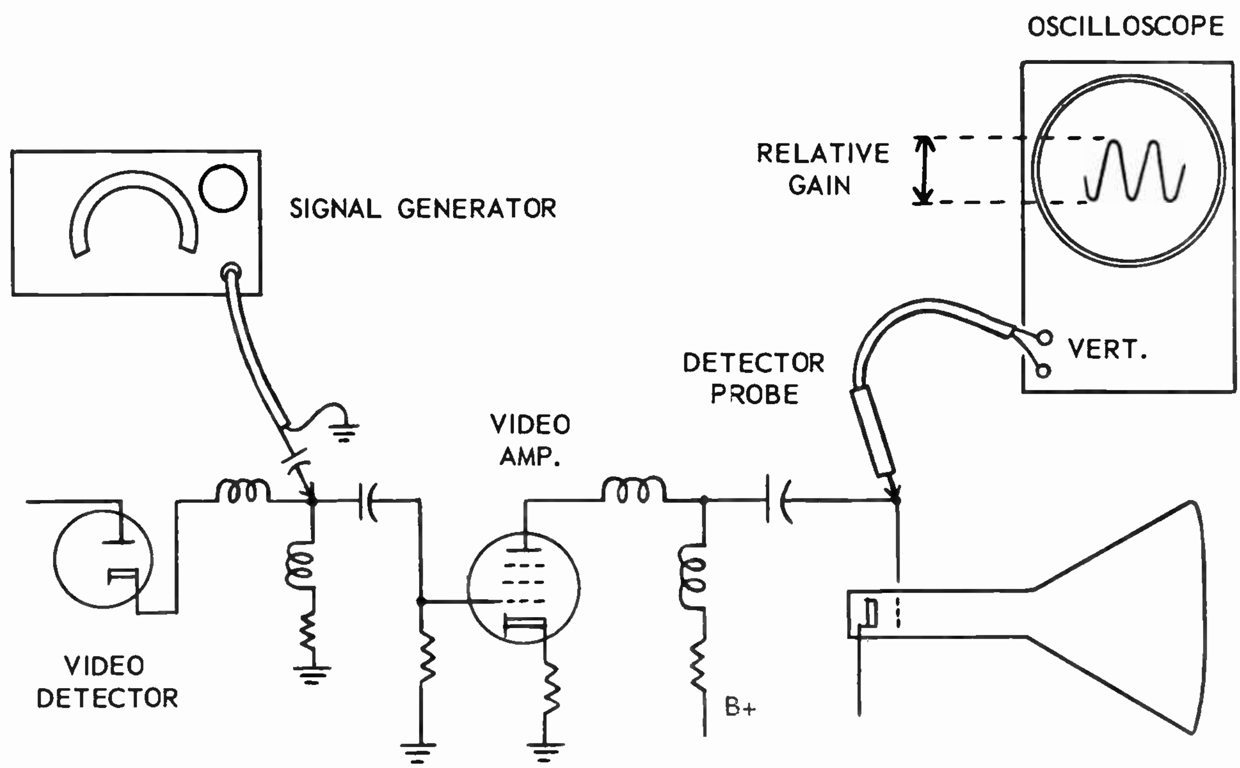


Fig. 91-10. Measuring voltage gain in a video amplifier at various frequencies by using the oscilloscope for a voltmeter.

The signal generator now is tuned from its lowest radio frequency through to something more than 5 mc. The changes of trace height will show, approximately, the relative amplifier gains at the tuned frequencies. The cutoff frequency, between 3.0 and 4.5 mc, will show up very definitely as the trace height drops practically to zero. Since r-f signal generators seldom will deliver frequencies much below 100,000 cycles or 0.1 megacycle, this method is not useful for checking the video amplifier response at the low frequencies.

This check of frequency response may be carried out equally well with the vacuum tube voltmeter instead of the oscilloscope, since the scope is used merely as an indicator of relative voltages. The same detector probe which is used for the scope (the type used for frequency response) is connected to the d-c input of the VTVM. The signal generator need not be operated with audio modulation when the VTVM is used as the output indicator. In fact, it is advisable to turn off the modulation and feed only the radio-frequency signal voltages through the video amplifier.

In making these tests you must realize that the detector probe is not compensated for frequency. Fur-

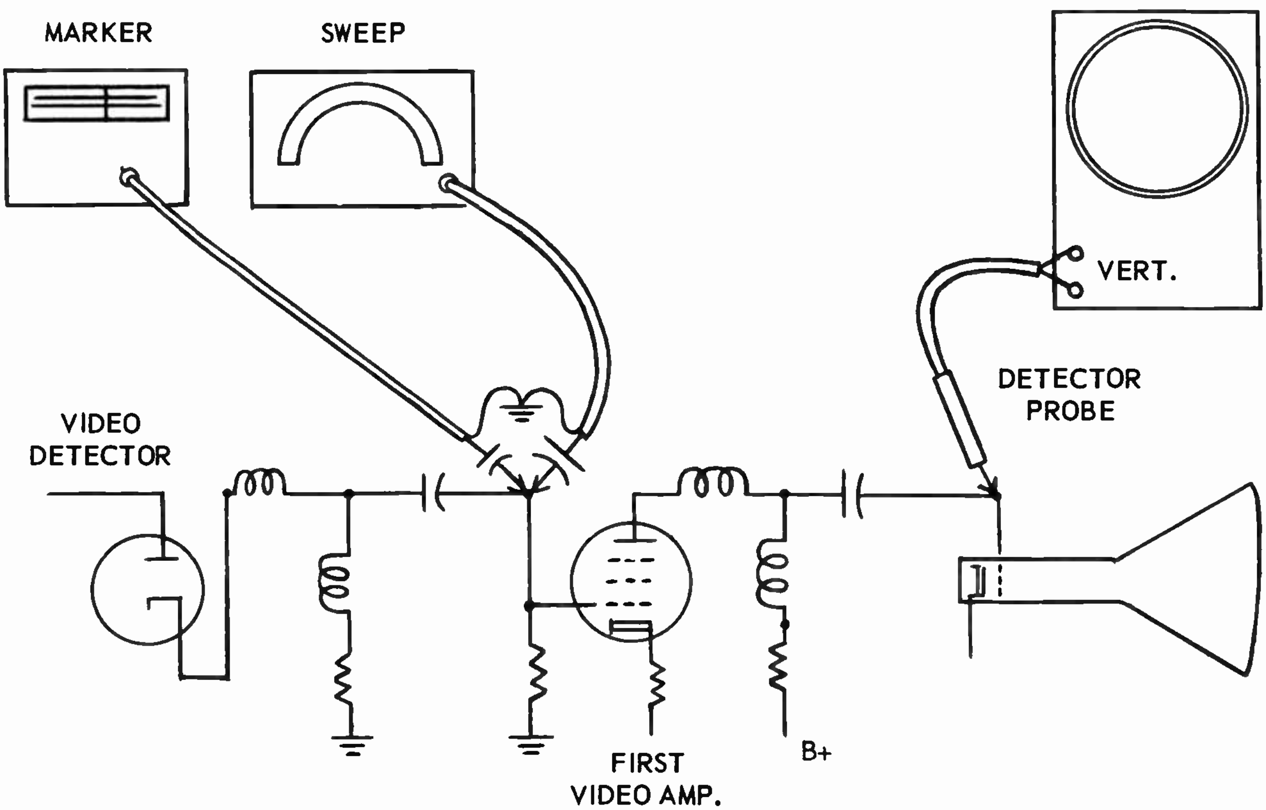


Fig. 91-11. Connections for measuring video amplifier response by using swept frequencies increasing from zero.

thermore, energy losses in the probe and its connections are greater at frequencies of several megacycles than a few hundred cycles. This will cause the indications of a VTVM or the trace height on a scope to decrease to a greater extent than the amplifier gain decreases as the test frequency is increased. Also, some small voltage will be indicated by the meter or there will be some height of the oscilloscope trace even with zero output from the signal generator. These effects are due to varying or alternating voltages present when the receiver is turned on. The response at some frequency such as 7 to 8 mc should be considered as zero amplifier gain.

It is a simple matter to observe a trace of video amplifier frequency response on the oscilloscope if you have a suitable sweep generator. First, the generator must be capable of operating at a center frequency of two to three megacycles so that, with the sweep width set for about five or six megacycles, the total sweep will extend down to zero frequency and up through the high-frequency limits for video amplifiers. Second, the sweep frequency should be produced by beating together of two fundamental frequencies, one fixed and the other varying. Neither of these frequencies should be a harmonic if confusing effects are to be avoided.

Instrument connections are made as in Fig. 91-11. The high-side cable from the sweep generator is connected through a capacitor of about 1,000 mmf to the grid of the first video amplifier tube or to the grid of the single video amplifier when there is only one. The high side of the marker cable is connected to the same point through a capacitor of 20 to 50 mmf. The vertical input of the oscilloscope is connected

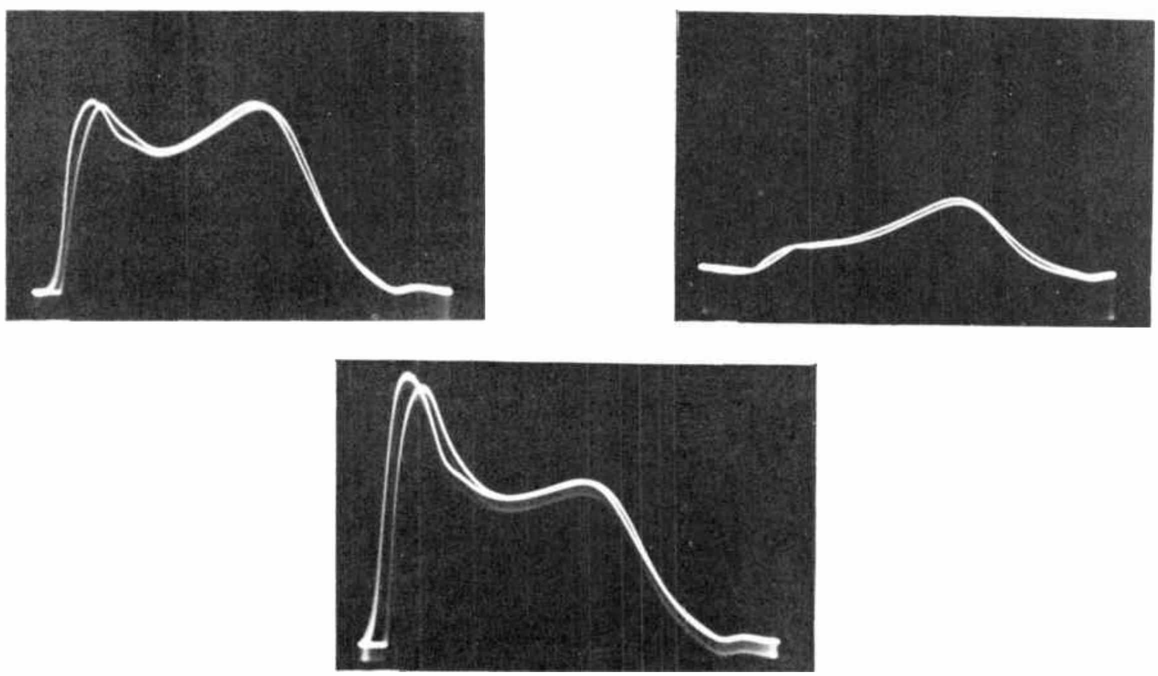


Fig. 91-12. Frequency responses of a video amplifier with the contrast control adjusted for normal reception (1), turned very low (2), and turned too high (3).

through a detector probe to the output of the video amplifier system, which most often is at the control grid of the picture tube.

To begin with, set the sweep center frequency at about 3 mc and the sweep width for about 5 mc. With the contrast control of the receiver set as for normal picture reception the response trace on the oscilloscope should appear somewhat as shown at 1 in Fig. 91-12. The response rises rapidly from the point of zero frequency at the left, remains reasonably constant between one peak at a low frequency and another caused by the peaking inductors between 3 and 4 mc, then drops to zero at 4.5 mc due to the trap effect of a sound takeoff. There is a slight rise beyond 4.5 mc and then a final drop to zero.

If the contrast control is turned almost all the way down the response changes as shown by curve 2. Although the low frequencies now are very weak the peakers still produce a decided hump around 3 to 4 mc. Turning the contrast so high as to cause unsatisfactory picture reproduction raises the low-frequency response very high, as in curve 3, but does not have any great effect on the high-frequency hump. The "fuzz" that shows underneath the trace lines is due to beating of the fixed and varying frequencies in the sweep generator.

Since no marker generator will furnish frequencies all the way from zero to four or more megacycles the frequencies along the response curve can be identified only within the range of whatever marker is available. To check the low-frequency end of the curve it is possible to employ an audio-frequency signal

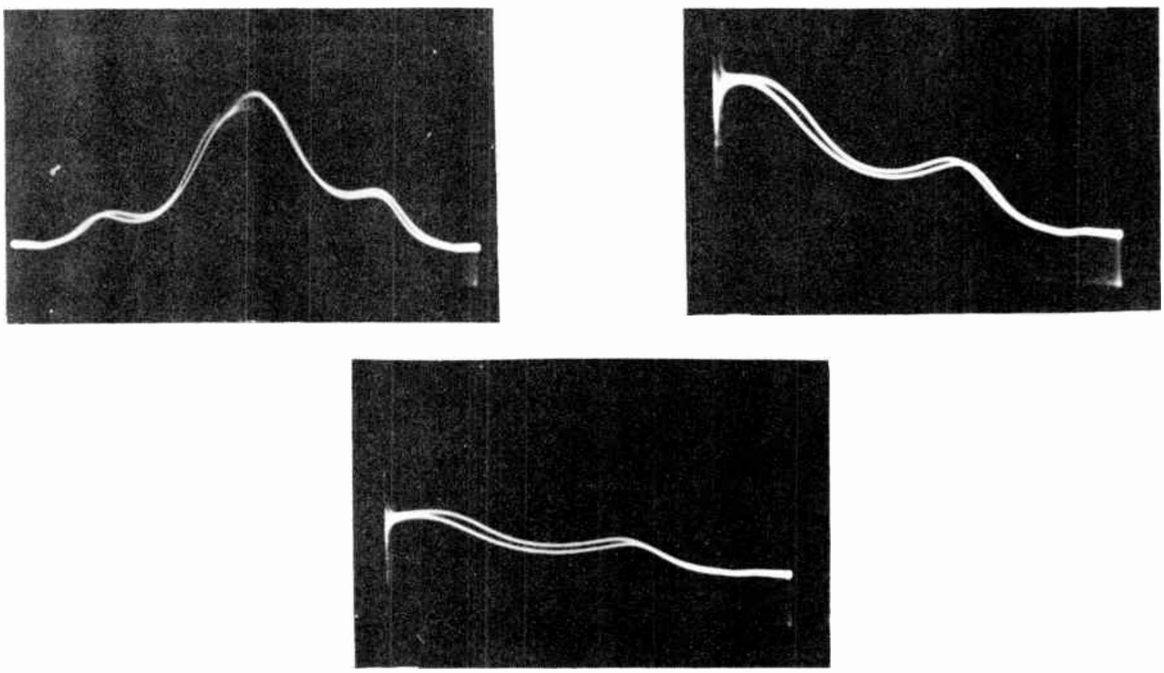


Fig. 91-13. Video amplifier response traces obtained by beating the frequencies of a sweep generator and a constant-frequency generator.

generator with this end of the response spread out as wide as the horizontal gain control of the scope will allow. As a general rule enough information for service work may be obtained by checking frequencies from about 0.1 mc through the high-frequency peak and the cut off point of the curve.

When the sweep generator will not operate at center frequencies so low as three megacycles it still is possible to observe the frequency response of the video amplifier. This is an alternating method but is not recommended, since the desired wave form may be extremely difficult to obtain and interpret with the usual shop equipment. The instrument connections are exactly the same as in Fig. 91-11, but it is necessary to remove the effects of all circuits ahead of the input to the video detector. This may be done by taking out the detector tube in sets whose heaters are in parallel, or the detector input (either cathode or plate) may be connected to ground through a capacitor of about 1,000 mmf, or the circuit between detector and first video amplifier may be temporarily opened at any solder joint. In the latter case be sure not to open the grid return circuit of the video amplifier.

The next step is to adjust the center frequency of the sweep generator to any convenient frequency, such as 10 to 20 mc. Then, with maximum output voltages from both the sweep and the marker generators, vary the frequency of the marker until the oscilloscope displays a trace on the general order of that shown at 1 in Fig. 91-13. Now change the marker frequency very slowly, or else change the sweep width slowly, until you obtain a trace such as shown at 2.

The vertical line at the left of this trace is at the point where the marker frequency is equal to the sweep frequency, and indicates zero frequency so far as input to the video amplifier is concerned. It is zero frequency for the amplifier input because you are feeding into the amplifier the beat frequencies which are combinations of the constant frequency from the marker generator and the varying frequencies from the sweep generator. The sharp vertical line is the frequency of zero beat.

Varying the setting of the receiver contrast control will change the response in the same general way as when using other methods. As an example, turning down the contrast control will change the response from the trace at 2 to the one at 3 in Fig. 91-13. Unless the sweep and marker generators are capable of furnishing high output voltages a readable trace can be secured only with the receiver contrast control turned rather high. Consequently, this method is likely to show greater than normal heights for the low-frequency end of the response. It still allows checking the high-frequency peak and the cutoff. To identify the points at which various frequencies fall on the response trace it is necessary to use still another generator to furnish a variable marker pip.

TRAPS FOR INTERCARRIER BEAT. In receivers having a dual sound system there often is a trap for keeping the intercarrier beat frequency of 4.5 mc out of the picture tube input. The trap may be located anywhere between the output of the video detector and the grid or cathode of the picture tube. To adjust this trap proceed as follows.

Connect a constant-frequency signal generator to the grid of the first video amplifier tube or to the grid of any video amplifier which precedes the trap. The generator must be tuned accurately to 4.5 mc, preferably by a check with crystal calibration or by using a 4.5 mc crystal in the generator. The d-c input of the vacuum tube voltmeter is connected through a detector probe to the grid or cathode of the picture tube, whichever of these elements receives the picture signals. Use the signal generator without audio modulation. If the contrast control of the receiver acts on any video amplifier tube, set this control at maximum.

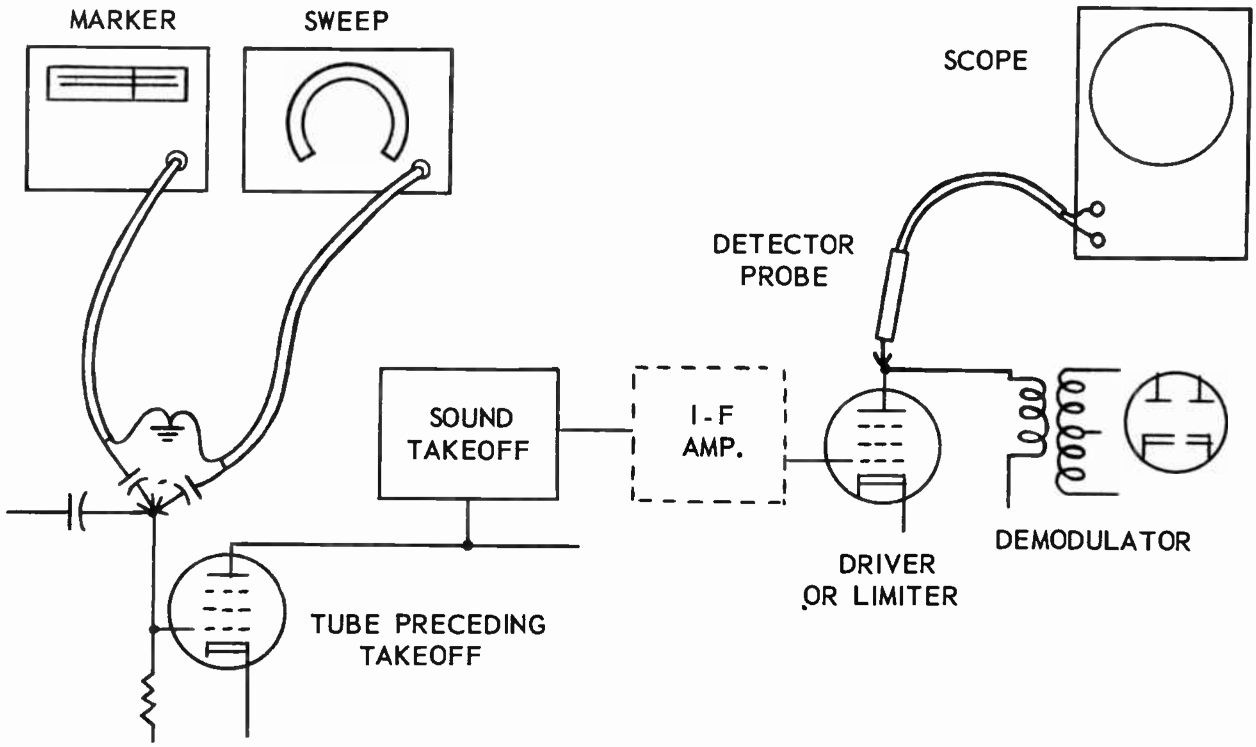


Fig. 91-14. Checking the frequency response of an i-f amplifier in the sound system.

Now adjust the trap for the lowest possible reading on the VTVM. Increase the output of the signal generator as the trap is brought more nearly to its point of final adjustment. The meter reading can be reduced to a minimum, but not to zero, because voltages at frequencies other than 4.5 mc always are present.

Incidentally, this procedure without any changes whatever may be used to check the tuning of the sound takeoff in a receiver having an intercarrier sound system. The takeoff should absorb or remove maximum energy at exactly 4.5 mc, and when it does so the input to the picture tube will be a minimum at this frequency. In other words, the intercarrier sound takeoff behaves just like a trap for removing the intercarrier beat frequency.

SOUND I-F ALIGNMENT. The frequency response of the portion of the sound system from the takeoff through to the primary of the demodulator transformer may be observed with the connections of Fig. 91-14. The sweep and marker generators are connected through the usual capacitors to the grid of any tube which precedes the sound takeoff. This may be any tube from the mixer through to the last video amplifier, it all depends on where the sound signal is taken off. The oscilloscope is connected through the detector probe to the plate of the tube which feeds into the primary of the demodulation transformer. This ordinarily would be the driver with an intercarrier sound system or the limiter with a dual sound system.

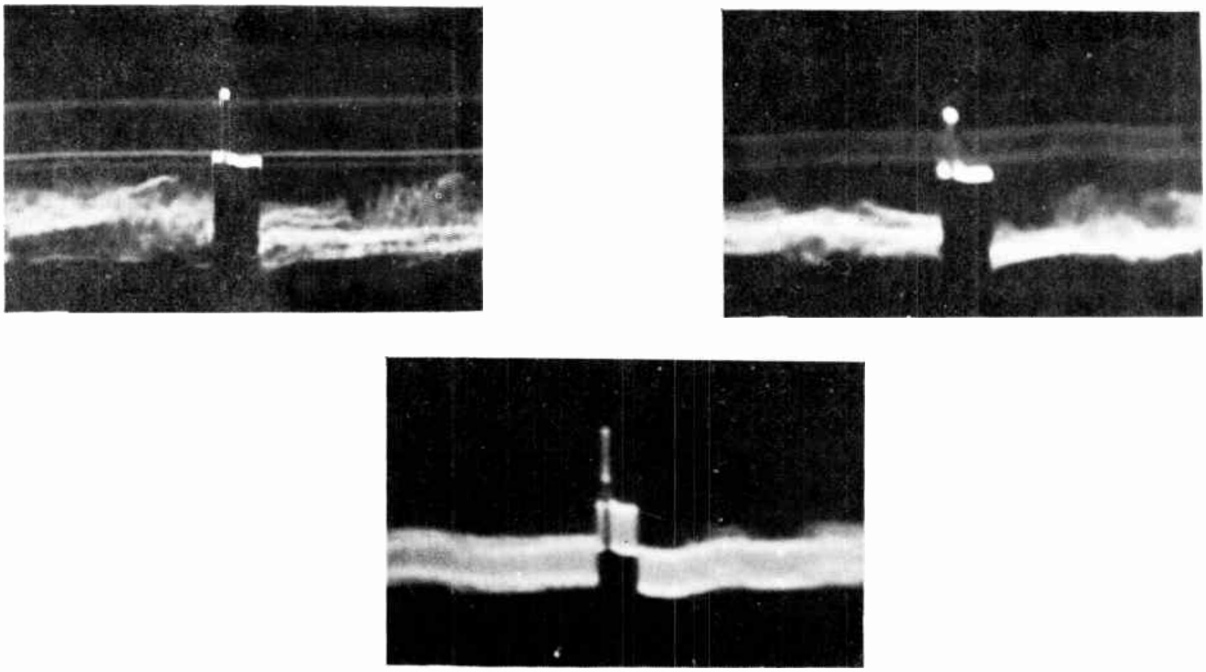


Fig. 91-15. Vertical blanking intervals and sync pulses observed in the video i-f amplifier system.

For an intercarrier sound system the center frequency of the sweep generator is adjusted for about 4.5 mc. For a dual sound system the sweep center is adjusted for the sound intermediate frequency employed in the particular receiver. The sweep width will have to be something like two or three megacycles in order to display the sound system response on the oscilloscope.

The first check would be with the marker turned to precisely 4.5 mc for intercarrier sound or to the sound intermediate frequency for dual sound. The pip for this frequency should be at or very close to the center of the response. Then the marker may be tuned to identify the frequencies or any peaks or valleys on the response, and to check the band width.

Alignment adjustments may be made as required in the takeoff coupler and in any sound i-f couplers which may be used. The general methods of alignment and the results to be secured are the same as explained for methods employing other types of output indicators, the only difference being that here we are using the detector probe. The probe may be connected also to the plate of any sound i-f amplifier tube to check the response through couplers as far as the one in the grid circuit of this i-f amplifier. It is possible to observe the frequency response of any one sound i-f stage by using connections essentially like those of Fig. 91-7, with the generators to the grid of the tube that feeds the coupler in question, and with the scope to the plate of the tube following this coupler.

PREAMPLIFIER ALIGNMENT. The output from a preamplifier or television booster is at carrier frequencies, and requires demodulation before it can be observed or measured on an oscilloscope or any ordinary type of vacuum tube voltmeter. The demodulation is best provided with a detector probe used with an oscilloscope.

The sweep generator should be connected through an impedance matching pad to the input or antenna terminals of the preamplifier. The marker generator high side is connected through a fixed capacitor to either input terminal, and the low side to chassis ground. The preamplifier output terminals, from which a short line normally goes to the receiver, should be connected together with a fixed carbon resistor of about 300 ohms to simulate the 300-ohm input impedance of a receiver. The high side of the detector probe then is connected to one side of the carbon resistor and the ground lead of the probe is connected to the other side of the resistor. The preamplifier may be tuned to any channel and the sweep generator center frequency adjusted to bring a response trace onto the oscilloscope, or the sweep may be adjusted to sweep some channel and the preamplifier tuned to produce the frequency response trace.

Instead of using the detector probe with an oscilloscope it may be used on the d-c input of a vacuum tube voltmeter. Then the preamplifier may be aligned to produce voltage peaks at or near the video and sound carrier frequencies of each channel, or near the channel limits if the band width is great enough, or near the center of the channel frequencies if the response shows only a single peak.

The preamplifier alignment adjustments are made in the same manner as those for the r-f amplifier or r-f-to-mixer coupling in a television tuner. There may be trimmer capacitors, adjustable coil cores, coils

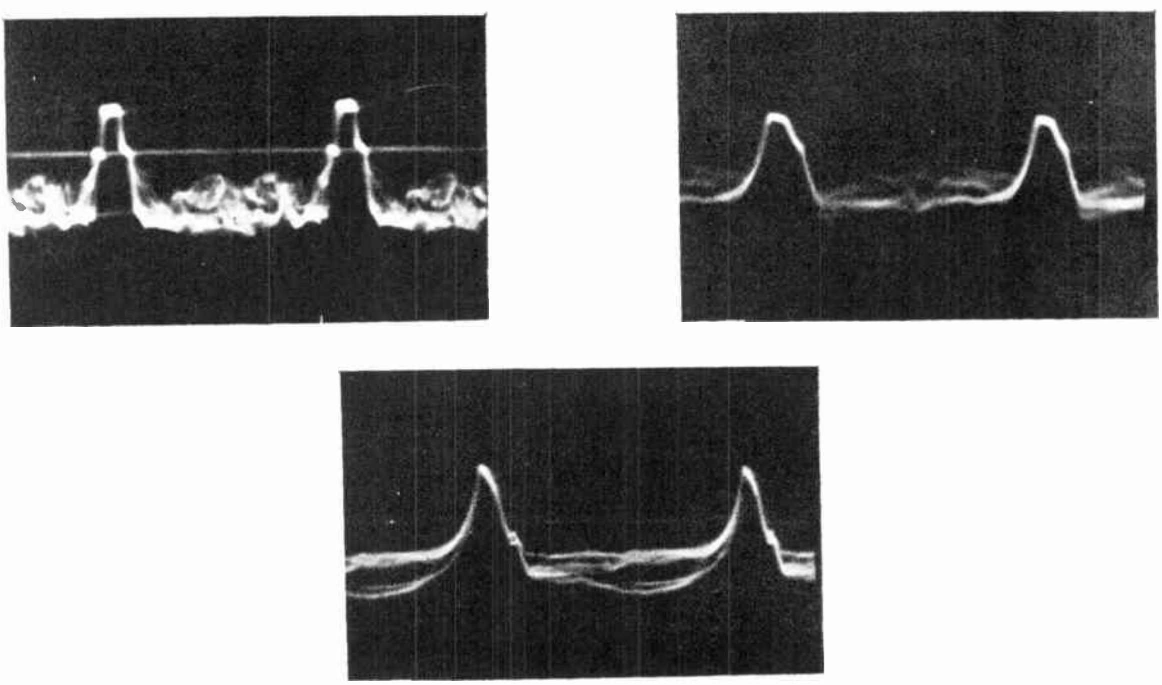


Fig. 91-16. Horizontal blanking intervals, sync pulses, and picture line signals observed in the i-f amplifier system.

suitable for spreading or squeezing, and all the other types of adjusters found in tuners. If good picture quality is important the preamplifier should be aligned for band width not less than 6 mc on every channel. If maximum gain is the chief desire, the alignment may be made to produce a peak at or near the video carrier frequency while maintaining fair gain at the sound carrier frequency.

The peaks, valleys, band widths, and general shapes of the responses for various channels quite often will differ radically. As when aligning the r-f circuits of tuners it usually is necessary to make compromiss adjustments to obtain acceptable results on all channels, or else to make adjustments which favor channels in which there is weakest reception.

SIGNAL WAVEFORMS. In the early part of this lesson we discussed two general types of detector probes, one especially suited for observation of television signals as they exist in circuits between the tuner and video detector, the other better suited for observing frequency responses and for use with the vacuum tube voltmeter. It is the latter type which is most generally employed for service work, and it is the type employed in the various tests and measurements which have been described up to this point.

There are, however, certain kinds of troubles whose causes are fairly easy to locate if we can follow a received signal from the tuner to the video detector input, and for such jobs it is necessary to use a probe that will demodulate the signals and allow their display on the oscilloscope. The receiver is operated just as for normal reception, with its regular antenna and transmission line connected. Pictures may be seen on the picture tube at all times and the audio portion of transmitted programs is heard from the speaker. The detector probe, connected to the vertical input of the oscilloscope, is used at the plates of the video i-f amplifiers and at the mixer plate. The same signal would appear at the grid of any given tube, but there is somewhat less loading and detuning effect with the probe used at the plate. Both plate and grid signals may be observed in checking for a faulty tube or for anything that affects operation of a tube.

For observation of signal waveforms at vertical scanning frequencies the internal sweep of the scope is adjusted for 60, 30, or 20 cycles per second. For checking horizontal line intervals the internal sweep is set for 15,750 cycles, 7,875 cycles or 5,250 cycles, depending on how many line intervals you wish to see at one time.

At 1 in Fig. 91-15 is a trace of a vertical blanking interval and some of the lines on either side with the oscilloscope connected directly to the load resistor of the video detector, without the probe. This trace is shown merely for comparison with the one shown at 2, which is taken with the detector probe connected to the plate of the second i-f amplifier. This trace does not differ a great deal from the one taken at the detector load resistor. A trace from the plate of a third i-f amplifier would be still more like the one from the video detector.

As we move back toward the mixer the traces lose some of the features of the actual signal, and take on others. At 3 is shown a trace of a vertical blanking interval and some of the horizontal lines with the detector probe at the plate of the mixer tube. We still have enough of the signal characteristics for easy identification of faulty performance.

At 1 in Fig. 91-16 is shown a trace of two horizontal blanking intervals with the horizontal sync pulses and the intervening picture signal. This trace was taken without the detector probe, with the oscilloscope connected directly to the video detector load resistor. At 2 is a trace of the same horizontal interval taken with the detector probe connected to the plate of the second i-f amplifier. There has been quite a loss of detail, but the sync pulses and even part of the form of the "back porch" still are clearly recognizable and would allow identification of faulty performance.

A trace of horizontal intervals taken from the third i-f amplifier would be much more like the one at the detector load, but in going the other direction there is greater deformation of the waveform, as illustrated at 3, which is a trace with the detector probe at the plate of the first i-f amplifier. At the plate of the mixer there usually is so much distortion that you would not recognize the vertical blanking intervals and sync pulses other than knowing from experience what you are looking at.

It always is much easier to observe the vertical than the horizontal intervals at any point along the amplifying system. This is no great handicap, because in the great majority of cases the presence of a good signal as observed at the vertical frequency means that the horizontal pulses are satisfactory. This is not always true, but observation of the vertical signals will allow either the elimination of trouble possibilities or else the identification of faulty performance. Of course, if you have the composite television signal in one stage, and it becomes lost or badly deformed in a following stage, the hunt for trouble is narrowed to a very small section of the receiver.

TELEVISION

LESSON NO. 92

SYSTEMATIC TROUBLE SHOOTING

Training Your Mind...

C.T.I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

COMMERCIAL TRADES INSTITUTE

Training Your Mind — C.T.I. — Perfecting Your Skill

Chicago, Ill.

LESSON NO. 92

SYSTEMATIC TROUBLE SHOOTING

There are all kinds of television service men, good, bad, and indifferent. The ones who are most successful in the profession, and make the most money, are those who can shoot trouble fast. They are the technicians who spend a relatively short time on a faulty receiver and determine what is wrong. Anyone possessing reasonable skill in handling tools then can replace or repair the defective part or parts – while the expert trouble shooter goes on to analyze other sets which still other helpers will repair, once they have been told what to do.

Hit and miss methods of trouble shooting often get by when handling sound radios which contain only a few dozen parts. Within an hour at most it is possible to check everything from antenna to speaker. But this won't work when you are faced with the four or five hundred circuit elements and parts of a television receiver. You must follow a system based on picture analysis, signal tracing, waveform analysis, and sectionalizing of possible faults. Then, in a matter of minutes, you can eliminate whole sections and divisions of the receiver as being incapable of producing the trouble which exists, and quickly narrow your search to a single circuit or a small group of parts in which the fault must lie.

There are three preliminary steps to be taken when commencing to look for any kind of trouble.

1. Check the performance on more than one channel. The transmitted signal itself may be at fault. If the trouble disappears on another channel, suspect the first station. If the trouble remains, suspect the receiver or its installation.

2. Observe the effect of all controls which are accessible to the operator. Turn each control slowly to note any erratic behavior. Make a final adjustment for the best possible performance. The owner or operator may need instruction for correct use of the controls.

3. Turn the selector to a channel in which there is no transmission. Increase the brightness control to illuminate the picture tube screen and produce a raster, if possible. Should the original trouble remain, the fault probably is in parts of the receiver from sweep oscillators to picture tube. If the trouble disappears, the fault probably is between antenna and sweep oscillators, or may be in the received signals.

When checking service adjustments which are generally accessible from the rear of the chassis, place a fairly large mirror in front of the picture tube and facing the rear so that you may observe the effects of

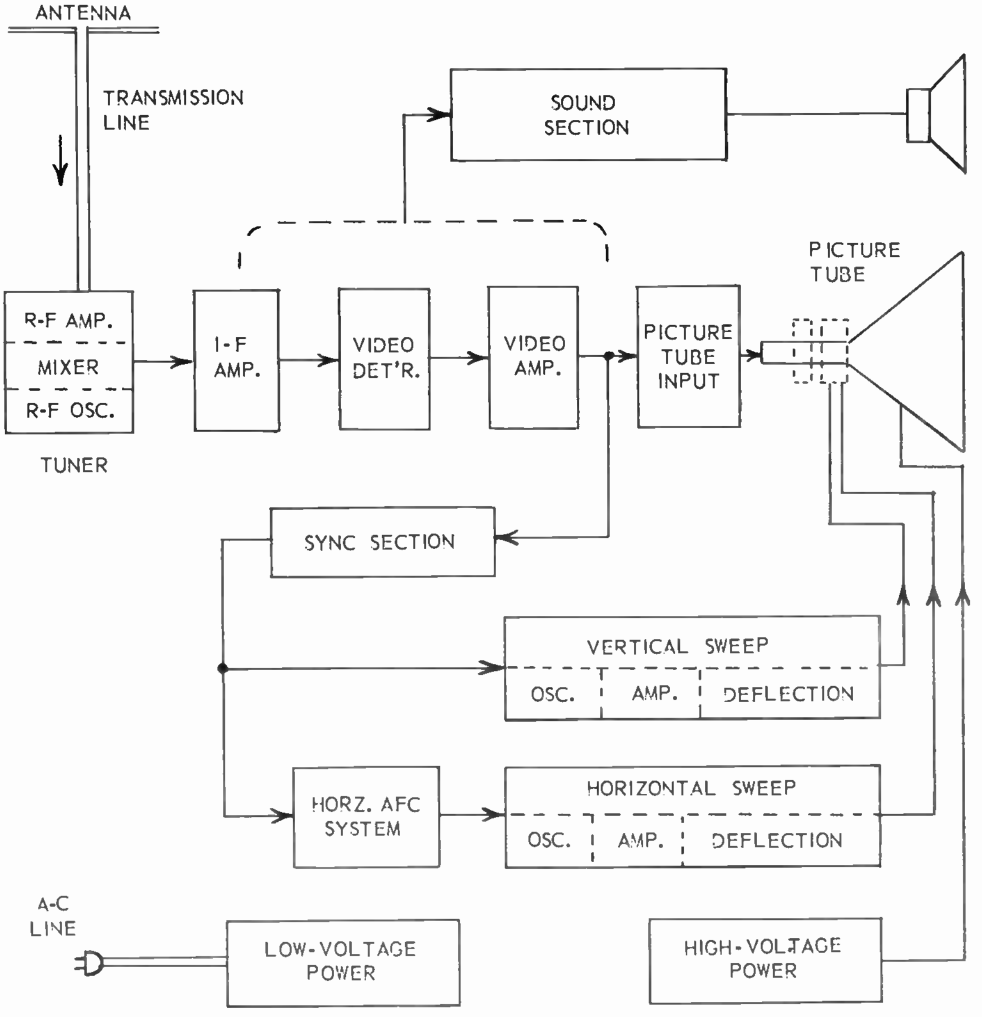


Fig. 92-1. The receiver is to be considered as consisting of various separated sections, in each of which certain faults may occur.

settings. An unbreakable light-weight mirror suitable for use on service calls consists of a 10 by 14 inch chrome ferrotype plate, obtainable from any photographic supply store.

SECTIONALIZING. The basic principle in fast trouble shooting is to divide and subdivide the possible sources of trouble. For example, supposing there are 100 possible faults. Your first test or observation may show that 60 of them cannot cause the existing difficulty, which leaves 40 to be suspected. A second check may eliminate 24 additional causes, leaving 16 possibilities. A third check may eliminate 10 more, leaving 6 parts or circuit elements for intensive examination. This, of course, is an ideal example. Real jobs may not be quite so easy. But the principle of dividing and subdividing must be followed if you are to attain speed.

Think of any television receiver as consisting of the divisions and subdivisions named in Fig. 92-1. Visualize the signal paths from antenna to picture tube and speaker. This would work out in a typical case as follows. Assume there is a raster, but neither picture nor sound. You should draw these conclusions.

1. To produce a raster, both the vertical and the horizontal sweep sections, also the high-voltage power supply, must be working. They need not be examined.

2. All the sections mentioned draw energy from the low-voltage power supply, so it may be eliminated as the cause for trouble.

3. There is neither picture nor sound. This could result from failure of the low-voltage power supply, but we have determined that this section is working. It is highly unlikely that anything serious would develope in the picture section and sound section at the same time. If this is a dual I.F. receiver, we may assume for the time being that the sound section is OK, and that everything from the i-f amplifier through the picture tube input is working.

4. All that remains is the tuner, the transmission line, and the antenna as the most likely locations for trouble.

After noting any trouble symptoms and combinations of symptoms, decide which sections of the receiver, if at fault, could cause all these symptoms at once. The fault is in one of these sections. Note also whether the picture, the raster, or the sound is satisfactory, and decide which sections must be working correctly to produce these results. The fault won't be in these sections.

Once you determine what sections, sub-sections, or circuits are probably at fault, it becomes a matter of substituting tubes, of measuring voltages, and of measuring resistances in these sections or circuits. Thus the trouble finally is traced to a tube, a capacitor, a resistor, an inductor, an adjustment, or a faulty circuit connection.

PICTURE ANALYSIS

A television picture may be perfect, or it may be faulty in more than fifty different ways. Each of all these faults may result from any one of a number of troubles. Recognizing the picture defect and knowing its most probable causes is the process of picture analysis.

Just what causes a particular picture defect depends largely on the type of receiver. Lists of troubles which are to follow apply in a general way to all kinds of receivers. A given defect may result from a certain trouble in one receiver, and from something entirely different in another set. This is especially true of troubles in the sync, sweep, and horizontal afc systems.

Any trouble serious enough to produce a major picture defect is almost certain to produce minor accompanying defects at the same time. For example, poor resolution or lack of detail accompanies a great many other faults, and so does non-linearity of one kind or another. Many troubles are intermittent, they come and go at irregular intervals. The causes are the same as for similar continual troubles, usually aggravated by poor wiring or terminal connections or by resistors that heat and cool as they work and stop working.

Picture analysis is far from being a certain means for spotting particular troubles. It gives you the most probable causes. There are many unusual combinations of troubles which might cause the same defect as some simple single trouble. The troubles which are illustrated in following pages are given certain names as a matter of convenience and as a means for classification. Different technicians use different names for the same things. What we call a wavy picture may be called a rippling picture. Our bending often is called pulling. We speak of resolution, which is just as correctly called definition. We speak of smears, which someone else may call trailers.

TEST PATTERNS. Faults in reproduction are more easily recognized by observing test patterns than actual program pictures. A test pattern is a design on the general order of the one shown in Fig. 92-2. All the major television broadcasting stations transmit their individual test patterns for short periods of time before the regular daily program schedules, and sometimes between programs. The chief difficulty in using test patterns for trouble shooting is that the patterns are available for only a brief period of time.

A test pattern is designed to emphasize faulty reproduction, and to clearly bring out faults which seldom would be noticed in a picture where there is perspective and more or less continual movement of people and objects. As an example, horizontal resolution or rendering of details along horizontal trace lines may be observed and measured by the vertical wedges. Lines in these wedges become continually narrower and closer together toward the center. The farther toward the center these lines remain distinctly separated, the greater is the number of picture elements reproduced along each horizontal trace line, and the better is the video frequency response toward the higher frequencies. Numbers along the wedges show how many elements are resolved at each point where the lines remain distinct. Vertical resolution may be similarly checked by means of the converging lines in the horizontal wedges.

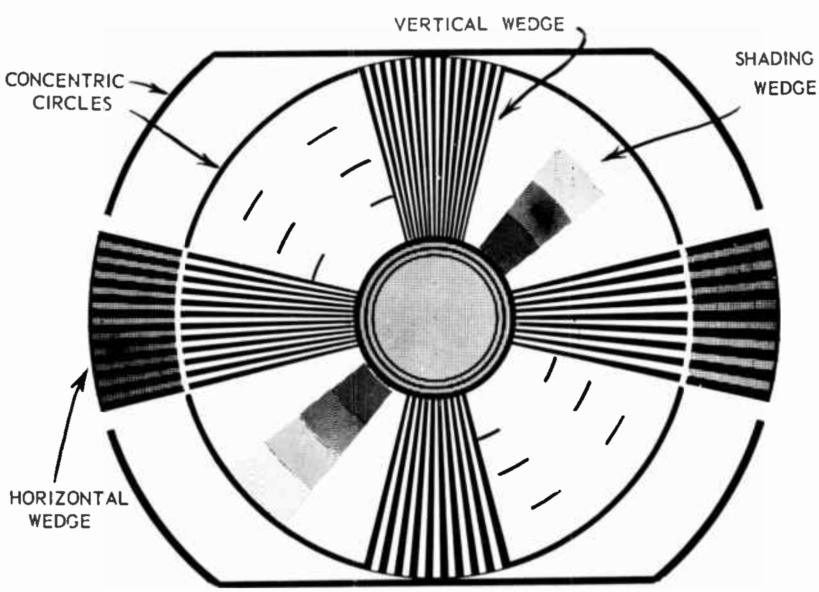


Fig. 92-2. Some features of a television test pattern.

Concentric circles and arcs, as well as straight lines, allow observing both vertical and horizontal non-linearity, which causes the circles to become egg-shaped or otherwise distorted. Some test patterns are ruled in squares for checking linearity at all points.

Relations between contrast and brightness are checked by means of shading wedges or, in some patterns, by a number of concentric circles having graduated shadings. There should be a gradual change from darkest to lightest tone along a wedge, or from circle to circle. Diagonal lines in wedges, or along arcs and parts of circles, appear jagged, or have a "moire" effect like some kinds of satin, when interlacing is faulty. Smears or trailers show up clearly at the ends of horizontal lines or blocks in a pattern.

A test pattern may show various faults, yet picture reproduction may be entirely acceptable to all except persons who, like yourself, are trained to look for defects. Usually it is poor policy to explain the features of a test pattern to customers. Later they may watch the patterns, and call you back for free service in case there is the slightest imperfection – which wouldn't affect their enjoyment of program pictures in the least.

There are three general classes of faulty video reproduction.

I. No picture. The screen may be dark, or there may be only a raster. There may or may not be sound.

II. Picture is defective, but nothing other than the picture appears on the screen.

III. Bars or lines appear on the screen, in addition to the picture.

Each general classification is subject to further subdivision. For class I, with no picture, the subdivisions are as follows.

I. No picture.
 A. No raster. No sound.
 B. No raster. Sound OK.
 C. No raster. Bright horizontal or vertical line.
 D. No raster. Bright spot on screen.
 E. Raster OK. No sound.
 F. Raster OK. Sound OK.

Now we shall proceed to list the most probable causes for each class and each subdivision of faulty reproduction. When the faulty appearance can be illustrated it will be shown by a photograph or drawing. Troubles are listed in either of two ways; in the order of most likely occurance, or in the order of ease in locating.

I. A. NO PICTURE. NO RASTER. NO SOUND.

1. All glass tubes are dark, all metal tubes cold. No heater voltage.

a) Receiver switch not turned on. Switch might be defective.

b) Power cord not plugged in, contacts loose or dirty at plug, cord broken inside its insulation.

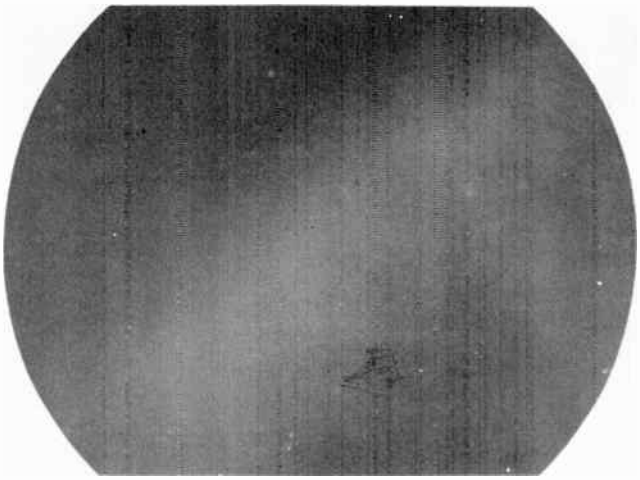


Fig. 92-3. The screen is dark, there is neither picture nor raster.

c) Building power line dead. Check house fuses.

d) Interlock connection at receiver end of power cord not plugged in, contacts dirty, loose, or broken.

e) Fuse blown in primary circuit of power transformer.

f) Power transformer primary winding open, disconnected, burned out. Check for a-c voltage on secondary side, using highest range of voltmeter.

2. Glass tubes lighted, metal tubes slightly warm. There is heater voltage but no B-voltage.

a) Low-voltage power supply rectifier(s) weak or burned out.

b) Interconnecting cable in receiver, as for power supply, not plugged in.

c) Speaker plug not making contact, when power filter choke is in speaker.

d) Tube removed, during tests, and not replaced.

e) Focus coil open, disconnected, burned out. B-voltage cannot come through this coil.

f) Low-voltage power filter choke or resistor open, disconnected, burned out.

g) Time delay relay not operating, remains open.

I. B. NO PICTURE. NO RASTER. SOUND OK.

1. High-voltage power supply inoperative, no anode voltage to picture tube.

Note.—A safe and simple way of checking for presence or absence of anode voltage is as follows: Obtain any small neon glow lamp such as type NE-48 or NE-45. Fasten the lamp in any convenient way to one end of a fibre rod or clean, dry wooden rod at least one foot long. The lamp terminals may be either open or shorted. With the receiver operating, hold the lamp near the rim of a metal cone picture tube or near the anode terminal of a glass tube. If the picture tube is glass, before turning the receiver on, fold back the rubber or plastic cup which is around the anode terminal. The lamp will glow if high voltage is present, will remain dark if this voltage is absent or very low. You may make a similar test at the cap or base of the high-voltage rectifier tube, and close to the cap of a horizontal output amplifier tube in a flyback power system. If there is no anode voltage, proceed as follows.

a) Any type of high-voltage power supply.

(1) Rectifier tube(s) weak or burned out.

(2) Rectifier filament winding open due to poor solder joint and etc.

(3) High-voltage filter capacitor shorted, filter resistor open.

b) Flyback type high-voltage power supply.

(1) Fuse blown in line to horizontal output amplifier. Do not replace this fuse until you have examined the following parts, one of which probably caused the fuse to blow.

(a) Tube weak or burned out; damper, horizontal output amplifier, high-voltage rectifier.
(b) No sweep voltage at grid of horizontal output amplifier. Check from here back to the horizontal sweep oscillator.
(c) Faulty inductor, capacitor, or resistor in voltage boost circuit of linearity control on the output transformer.
(d) Horizontal output transformer windings shorted or grounded.
(e) Deflection yoke grounded, shorts between vertical and horizontal coils.

(2) Tube weak or burned out, with fuse not blown. Check damper, horizontal output amplifier, horizontal sweep oscillator.

(3) Coupling capacitor open or leaky, or coupling resistor defective, between horizontal sweep oscillator and output amplifier.

(4) Width control inductor (on transformer) shorted.

(5) Linearity control (on transformer) voltage boost circuit. Leaky capacitor. Open connections, etc.

(6) Horizontal output transformer disconnected, open circuited, shorted.

c). R-f type high-voltage power supply.

(1) Oscillator tube(s) weak, operating with low plate or screen voltage.

(2) Coupling and decoupling chokes, capacitors, resistors. Open or shorted.

2. High-voltage power supply operating, according to previous test.

a) Brightness control. Turned too low, open circuit, bypass capacitor shorted.

b) Selector of combination receiver not at TV position.

c) Drive control misadjusted, too little signal to horizontal amplifier.

d) Ion trap magnet. Incorrect adjustment. Positioned for wrong polarity, reversed front and back, top and bottom. Wrong type of magnet for the picture tube.

e) Picture tube.
(1) Anode lead or connector disconnected or defective.
(2) Socket not secure on tube base, broken lugs, etc.
(3) Heater circuit open, shorted out.
(4) Cathode circuit open, no grid return path.
(5) Control grid bias is too negative.
(6) Voltage too low or zero on second grid of picture tube.
(7) Picture tube defective, must be replaced.

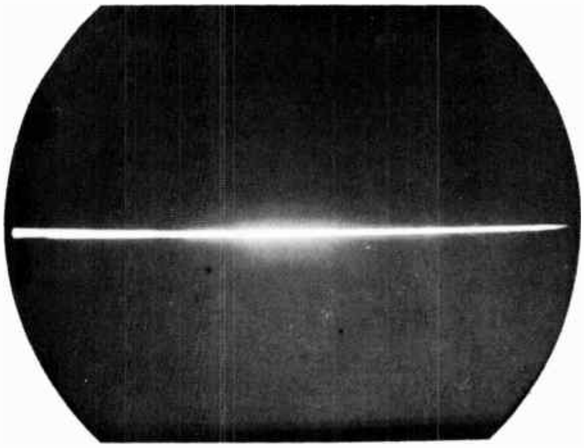


Fig. 92-4. This bright horizontal line indicates failure of the vertical sweep. A similar vertical line indicates failure of the horizontal sweep. Either line will damage the screen if brightness is high enough to make the line more than faintly visible.

I. C. NO PICTURE. NO RASTER. BRIGHT LINE ON SCREEN.

1. Horizontal line (as in illustration) indicates that the horizontal sweep is operating, but that there is no vertical sweep.

   a) Vertical oscillator circuit. Tube weak or burned out. Defective capacitor or resistor. Poor connections. Check the coupling capacitor(s) between oscillator and vertical amplifier for opens, leakage, wrong value.

   b) Vertical amplifier circuit. Tube weak or burned out. Check capacitors and resistors. Wrong voltage on a tube element.

   c) Vertical output transformer and connections to deflection coils. Open winding, poor connections.

   d) Vertical deflecting coils in yoke. Open, shorted, grounded.

2. Vertical line indicates that the vertical sweep is operating, but that there is no horizontal sweep.

   a) Flyback high-voltage power supply.
   Check leads and connections of the horizontal deflecting coil circuit. Other parts of the horizontal sweep and deflecting systems probably are OK because there is anode voltage causing the picture tube beam to strike the screen and produce the vertical line.

   b) R-f high-voltage power supply.
   Check all tubes and component parts in the horizontal sweep and deflecting systems, from the horizontal sweep oscillator through to the horizontal deflecting coils in the yoke. Tubes may be weak, operating with element voltages too low, burned out.

I. D. NO PICTURE. NO RASTER. BRIGHT SPOT ON SCREEN.

A stationary spot indicates that there is neither horizontal nor vertical sweep, but that there is anode voltage causing the beam to strike the screen of the picture tube.

Check all B-voltage lines and dropping resistors, also decoupling capacitors, which furnish plate and screen voltages to both the horizontal and vertical sweep sections, to both oscillators, to both amplifiers, etc.

Deflecting yoke cable disconnected, faulty contacts.

Note.—A bright stationary spot will quickly destroy sensitivity of the picture tube screen at that point. The brightness control must be immediately turned down until the spot is barely visible in a dimly lighted room while making tests. This is true also of either a horizontal or a vertical line across the screen.

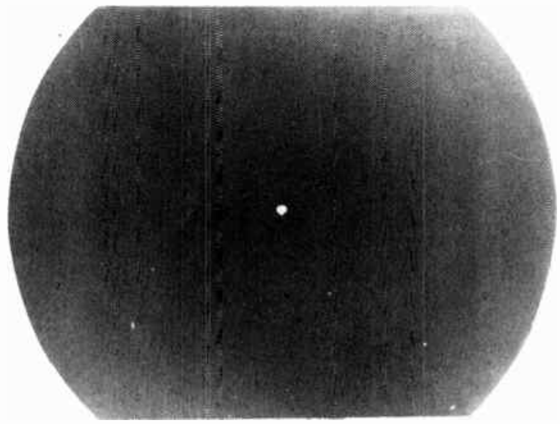


Fig. 92-5. A spot, without picture or raster, indicates failure of both horizontal and vertical sweep.

I. E. NO PICTURE. RASTER OK. NO SOUND.

It would be unlikely that serious trouble would develope at the same time in both the video and sound sections without either trouble having been noticed before the other one occured. Therefore, absence of both picture and sound indicates that trouble probably is in parts or circuits that carry both kinds of signals.

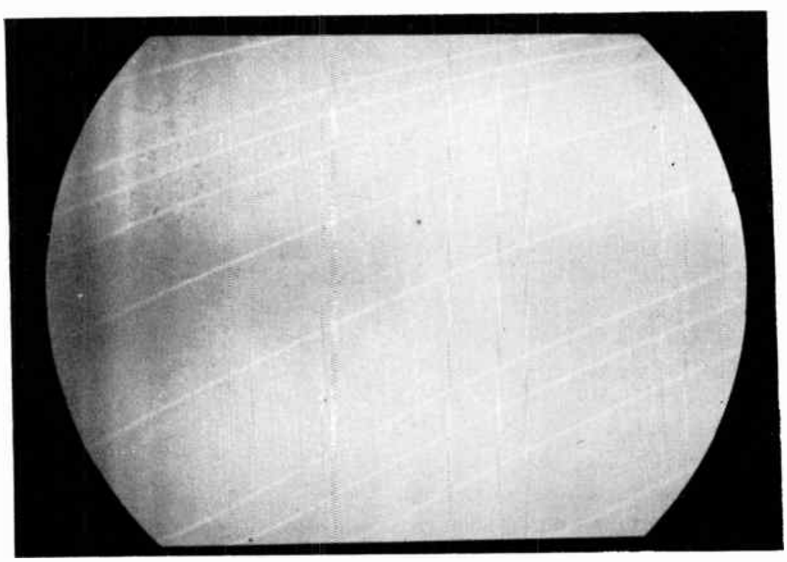


Fig. 92-6. A raster without a picture.

Trouble often is due to low gain or a weak signal combined with frequency response so narrow as to greatly reduce both video and sound signals. Adjusting the fine tuning or making realignments in the tuner or i-f amplifier may shift the peak of response to bring in either the picture or the sound, but not both together.

Because there is a raster we need not consider the sweep sections nor the high-voltage power supply. Also, there is nothing to indicate trouble in the sync section.

1. Either intercarrier or dual sound system.

   a) Fine tuning adjustment. Note preceding explanation.

   b) R-f oscillator tube weak, burned out, operated with wrong voltage.

   c) Tuner circuits aligned for too narrow response. Aligning for wider response may drop the gain very low for both video and sound.

   d) Transmission line weakening the signal, wrong placement, poor connections.

   e) Antenna disconnected, dirty terminals, oriented wrong.

   f) Too far distant from stations with which trouble occurs.

   g) I-f amplifier (carrying both video and sound) improperly aligned, tubes weak or operating with wrong voltages.

   h) Tuner. Check the r-f amplifier and mixer tubes for emission, wrong voltages. Dirty, loose, or broken contacts on channel switch.

   i) Automatic gain control circuit faults. Bias too negative on r-f and/or i-f amplifiers.

2. Only with intercarrier sound system.

   a) Check entire i-f amplifier system, as above. Check video detector and video amplifier for low output or low gain.

   b) Contrast control open circuited, faulty capacitor or resistor in circuit.

I. F. NO PICTURE. RASTER OK. SOUND OK.

The trouble almost certainly is in parts of circuits carrying only the video or picture signals. Since the sound is OK we need not consider the sound section itself, nor anything from the antenna through to the point of sound takeoff.

This assumes that the fine tuning control is correctly adjusted, and that the r-f oscillator is correctly aligned. Were either of these incorrect it could still be possible to bring in the sound without the picture, for the sound carrier or sound intermediate frequency could be at a satisfactory point on the frequency response while the video carrier or video intermediate could be down too low.

Trouble of this nature is most easily located by the process of signal tracing, using either an oscilloscope, a VTVM with high-frequency detector probe, or a regular signal tracing instrument with a similar probe.

In the absence of signal tracing it is advisable to check all the tubes, preferably by substituting others known to be operative. Also check plate voltages, screen voltages, and grid biases on all tubes. Trouble may be in any coupling capacitors or resistors, or in any decoupling units. Check through the following sections.

1. I-f amplifier in stages that carry only the video (and sync) signals without sound signals.

2. Video detector circuit.

3. Video amplifier circuit.

4. Automatic gain control and contrast control circuits which are connected to the above mentioned video i-f stages.

5. Picture tube grid-cathode input circuit, including connections to the d-c restoration system if such a system is used.

G. NO SOUND. PICTURE AND RASTER OK.

Although this condition is not an example of faulty video reproduction, it is closely allied to the two preceding subdivisions. The most probable causes are as follows.

1. Fine tuning control misadjusted, especially with dual sound systems.

2. Signals very weak. With some receivers a weak signal from a distant transmitter will produce pictures, but no accompanying sound. Check the antenna type, orientation, use of reflectors and directors, etc. Check condition of transmission line as to placement with reference to metallic objects, excessive length, impedance matching, etc.

3. I-f amplifier tubes weak. Alignment brings sound intermediate frequency too low on response curve.

4. Tuner frequency response and gain.
   a) Aligned with sound carrier down too low.
   b) R-f oscillator overall alignment incorrect.
   c) R-f amplifier tube weak, operating with wrong voltages.
   d) Mixer tube weak, operating with wrong bias or plate voltages.

Fig. 92-7. There may be a satisfactory picture, also a raster, but no reproduction of sound.

5. Sound section alignment.
   a) Check the sound takeoff, the i-f amplifier stage(s), and the demodulator.
   b) Audio amplifiers or speaker inoperative.

II. Picture is defective, but nothing other than the picture appears on the screen. There are no added bars or lines.

In this general classification there are four major subdivisions, with a number of faults in each. The subdivisions and the troubles or faults in each one are as follows.

II. A. Tone defects. Incorrect relative shadings.

1. Brightness excessive.

2 Brightness lacking.
3. Brightness varies.
4. Contrast excessive.
5. Contrast lacking.
6. Negative picture. Light and dark tones nearly reversed.
7. Shadows
8. Ion burn.

II. B. Detail lacking. Blurring of lines.

1. Focus incorrect.
2. Interlacing faulty.
3. Resolution or definition is poor.
4. Smears or trailers.
5. Snow in picture, from a weak signal.

II. C. Distortion. Portions of picture out of place.

1. Bending or pulling.
2. Centering incorrect, horizontal or vertical or both.
3. Linearity faulty, horizontal or vertical.
4. Reversed picture. Upside down or left and right reversal.
5. Size incorrect both horizontally and vertically.
6. Size incorrect horizontally only or vertically only.
7. Size varies intermittently.
8. Tilting or skewing within the mask.
9. Wavy or rippled lines and objects.
10. Wedge shaped, or non-rectangular outline.

II. D. Movement. Of the picture as a whole.

1. Flickering.
2. Jumping or jittery.
3. Sync failure, horizontal and vertical at once.
4. Sync failure, horizontal only.
5. Sync failure, vertical only.
6. Sync failure, tear out at top of picture.

II. A. 1. BRIGHTNESS EXCESSIVE.

It is assumed that brightness always remains excessive, that the brightness control is ineffective in darkening the picture. Slanting retrace lines are difficult to remove from the picture without using excessive contrast. The general cause for this condition is picture tube grid bias insufficiently negative.

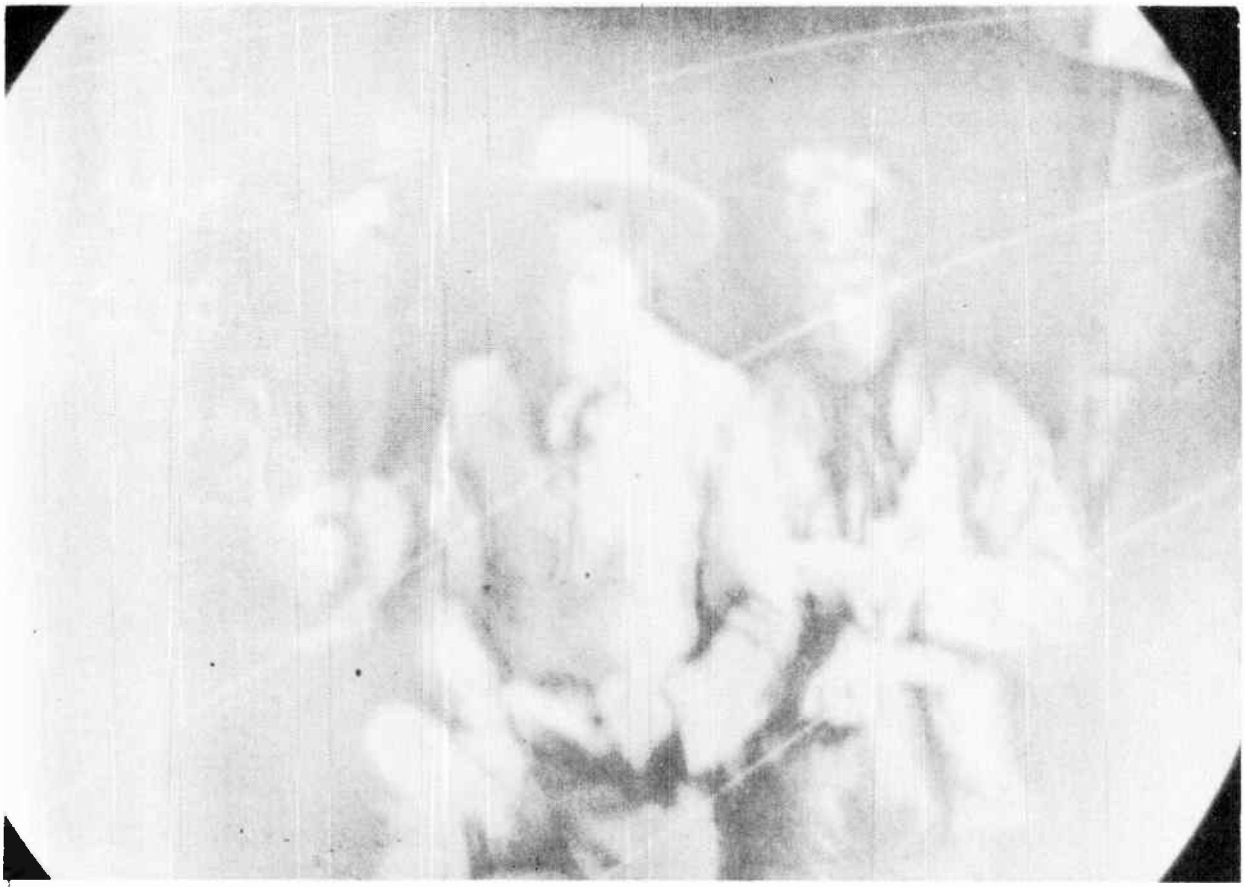


Fig. 92-8. Brightness too great. Note that vertical retrace (slanting) lines appear in the picture.

a) Brightness control.
   (1) Potentiometer shorted, open, or otherwise defective.
   (2) Leaky bypass capacitor.
   (3) Resistor open or of wrong value.
   (4) B-voltage applied to the brightness control circuit is either too high or too low, depending on how the circuit is arranged.

b) Video amplifier weak or operating with wrong voltages.

c) Coupling capacitor from video amplifier plate circuit is leaky.

d) Picture tube input (grid-cathode) circuit.
   (1) Cathode line open or shorted.
   (2) High-resistance short from grid to cathode conductors.

e) Picture tube, cathode-heater leakage.

Note.—If the screen changes to a brilliant all-over white, it indicates self oscillation in the i-f amplifier section. This usually is due to wrong alignment, with two or more stages operating at the same or nearly the same frequency. The receiver must be immediately turned off to prevent permanent damage to the screen. Oscillation cannot be stopped by operating any of the controls or service adjustments.

II. A. 2. BRIGHTNESS LACKING.

Here we are considering troubles which prevent sufficient screen illumination to bring out dif-

Fig. 92-9. Brightness lacking. The effect is somewhat similar to excessive contrast, in that there are not enough different gray tones to bring out details in the picture.

ferences between shading over various parts of objects. It is assumed that the received signal is strong enough to prevent appearance of "snow" in the picture.

a) B-voltage too low.
   (1) Check the voltage applied to the brightness control circuit. It may be too high or too low, depending on the type of circuit.
   (2) Low-voltage power supply furnishing insufficient voltage.
   (3) Power line voltage low.

b) Ion trap magnet adjusted wrong, or magnet weak.

c) Picture tube and circuits.
   (1) Control grid bias remains too negative at all settings of brightness control.
   (2) Low voltage to second grid.
   (3) Anode voltage too low. Check the high-voltage power supply.
   (4) Picture tube faulty, requires replacement.

II. A. 3. BRIGHTNESS VARIES IRREGULARLY.

a) D-c restoration circuit. Check for weak tube, capacitors leaky or open, resistors open, shorted, wrong value, faulty connections, etc.

b) High-voltage power supply filter capacitor open or disconnected.

c) Measure B-voltage output of low-voltage power supply. If erratic, look for trouble in rectifier tube(s), filter resistors, filter capacitors, and connections.

II. A. 4. CONTRAST EXCESSIVE.

Excessive contrast and lack of brightness cause generally similar effects. In either case the pictures are of the "soot and whitewash" variety. All objects appear nearly black or nearly white, with details lacking in parts of normally light tone, and with shadow areas going black.

If contrast is only moderately excessive, an increase of brightness will restore the tone balance but will make the entire picture brighter than for best reproduction. When brightness is lacking, reducing the contrast will often restore the tone balance but will leave the picture darker than it should be.

Troubles considered here are those which prevent the contrast control from producing correct tone balance without making the picture too bright.

a) Contrast control circuit.
   Check capacitors, resistors, and connections for opens, shorts, and leakage.

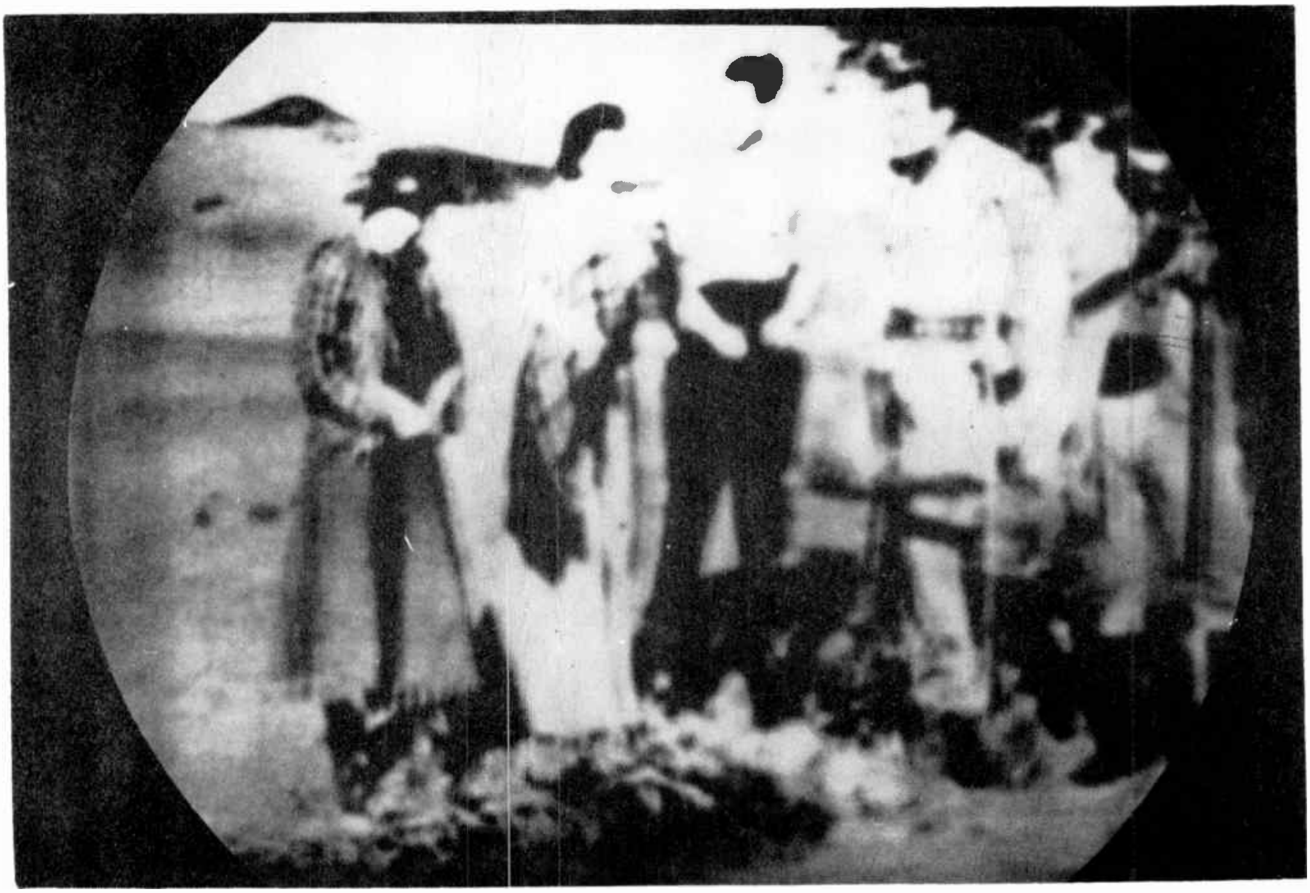


Fig. 92-10. Contrast excessive. Parts of the picture are nearly all white or all black, without smoothly graduated intermediate gray tones.

b) Automatic gain control circuit.
Look for leaky filter or bypass capacitors along the agc line, and for high-resistance shorts from this line to ground or B-minus.

c) I-f amplifier section.
Leaky coupling capacitor, making the bias insufficiently negative and increasing the gain

d) Picture tube circuits.
(1) Open circuit in lead to second grid, no voltage to this grid.
(2) Measure bias voltage between control grid and cathode while operating the brightness control. Bias may be too negative at all times, requiring that the contrast or gain control be set too high.

Fig. 92-11. Contrast lacking. There is not a full range of shading from nearly black to nearly white.

II. A. 5. CONTRAST LACKING.

Lack of contrast gives somewhat the same appearance as excessive brightness, in that there is insufficient range of shades or tones. If lightest portions are white, darkest portions are light gray rather than nearly black, while with darkest portions nearly black the lighter portions are dark gray rather than nearly white.

With excessive brightness the sloping vertical retrace lines usually are clearly visible. With lack of contrast these lines do not appear or show up only occasionally.

Lack of contrast ordinarily indicates lack of gain somewhere between antenna and picture tube grid-cathode input circuit. This should not be confused with the effect of a weak received signal or defective antenna installation, which generally causes snow in the picture.

a) Fine tuning control misadjusted.

b) R-f oscillator misaligned, check overall alignment adjustment if trouble appears on all channels.

c) Tuner.
   (1) Weak tube, r-f amplifier, mixer, or r-f oscillator.
   (2) Alignment is incorrect.

d) Transmission line mismatch at antenna, receiver, or both.

e) Weak interference due to oscillator radiation from other television receivers.

f) I-f amplifier tubes weak, operating with wrong voltages. Misalignment may cause video intermediate frequency too low on response curve.

g) Video detector weak.

h) Automatic gain control system maintaining grid biases too negative.

i) Video amplifier.
   (1) Tube(s) weak or operating with too low voltage.
   (2) Peaking inductor open circuited.

j) D-c restoration circuit, defective tube, capacitor, resistor, or connections.

k) Picture tube.
   (1) High resistance short between grid and cathode lines.
   (2) Cathode-heater leakage in tube.

In addition to the points listed it is possible for trouble to occur in coupling or decoupling capacitors or resistors anywhere in the tuner, the i-f amplifier, the video detector, or the video amplifier circuits.

II. A. 6. NEGATIVE PICTURE.

A negative picture is one which appears somewhat like a photographic negative, as compared with the positive print produced from the negative. Light and dark tones are partially interchanged, and details appear flattened out or in uniform grays.

a) Interference.
   Any rather strong radio-frequency or high-frequency interference. Strong radiation from the r-f oscillator of a nearby television receiver.

b) R-f oscillator. Extreme misalignment for the channel or channels on which the trouble appears, or in the overall alignment when the fault exists on all channels.

Fig. 92-12. A negative picture has generally "flat" tones, with actual or apparent interchanging of light and dark tones.

c) I-f amplifier. Overloading of one or more tubes. Grid bias may be insufficiently negative, or plate and/or screen voltage may be too high for the existing bias.

d) Automatic gain control. Allowing amplifier grid bias to become zero or even slightly positive.

e) Defective picture tube.

II. A. 7. SHADOWS.

A dark shadow at any corner or at either side of a television picture is caused by the electron beam in the picture tube striking the neck of the tube or the edge of the front opening in the

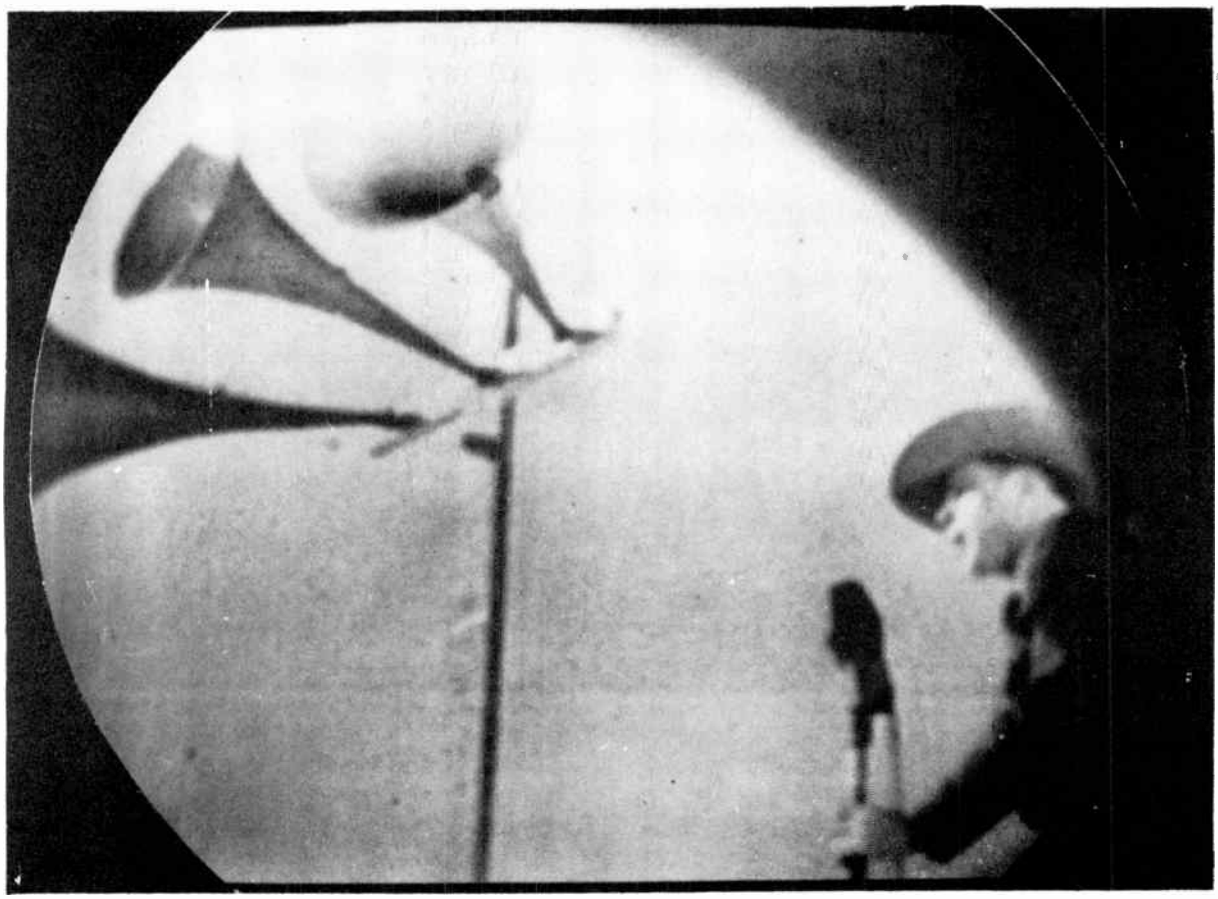


Fig. 92-13. Shadowing at the upper right-hand corner. A shadow may appear at any corner, at either side, or at top or bottom of the picture.

electron gun. Then the beam cannot reach the shadowed area of the screen. The shadow will appear on the raster as well as on all pictures.

a) Ion trap magnet.
   This magnet should be adjusted only in such manner as produces maximum brightness on the screen. It should not be adjusted to remove a shadow if such adjustment drops the brightness below maximum.

   The trap magnet may be of the wrong type for the picture tube.

b) Focus coil or focus magnet.
   (1) Placed too far back of the deflecting yoke.
   (2) Not centered around the neck of the picture tube. The coil or magnet should be so

mounted that, when tilted for picture centering, its center still is centered on the picture tube axis.

(3) Inner ring or other movable part of a permanent magnet device may have slipped out of its correct position.

(4) Leads reversed to a focusing coil.

c) Deflecting yoke.
Too far back from flare of picture tube. The yoke must be close against the cushion, and the cushion close against the tube flare. The yoke should be carefully centered with reference to the tube neck.

To maintain centering, avoid tilting of the picture, and to allow removal of shadows, it may be necessary to rotate the tube around its neck axis.

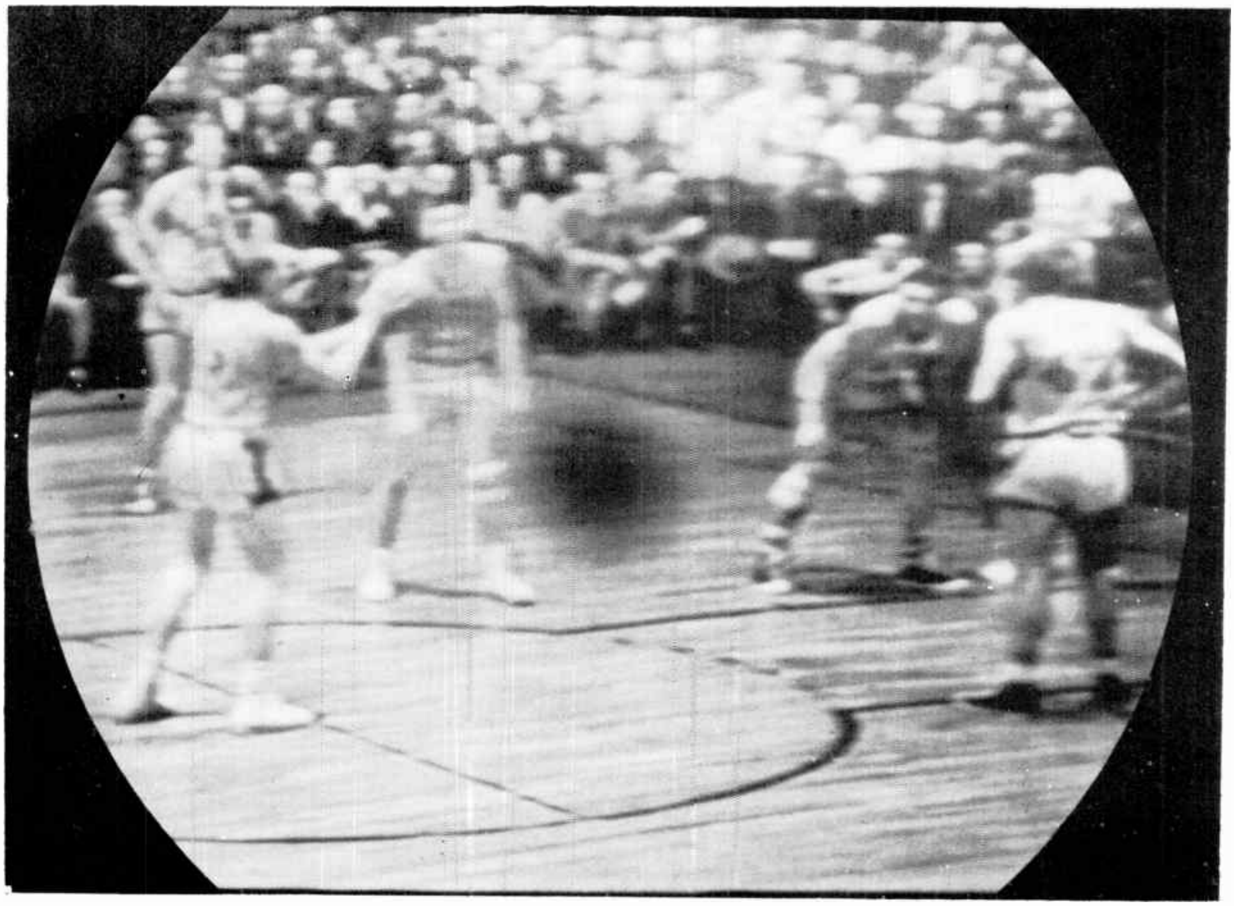


Fig. 92-11. An ion burn is apparent in darkening of the screen at the center of the picture area.

d) Electrostatic picture tube.
Interchange the connections for front and rear deflecting plates, then rotate the tube through 180° on its neck axis.

II. A. 8. ION BURN.

An ion burn makes itself evident as a darkened area of generally circular form or sometimes as a cross with irregular outline near the center of the picture tube screen. The darkening indicates loss of sensitivity at this part of the screen, which is due to failure or misadjustment of the ion trap magnet. This allows heavy ions to bombard the screen. With the receiver turned off the affected area may appear yellow or light brown, and the yellowish tinge may persist when viewing pictures.

If the burn is objectionable to the set owner, the only remedy is replacement of the picture tube. The ion trap magnet and its adjustment should be carefully checked.

TELEVISION

LESSON NO. 93

SYSTEMATIC TROUBLE SHOOTING—PART TWO

"Training Your Mind"

C.T.I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON NO. 93

SYSTEMATIC TROUBLE SHOOTING—PART TWO

Now we shall look at additional examples of faulty picture reproduction in which there are no added bars or lines, but only some defect in the image. Previously we examined various defects of tone or relative shadings. The next sub-division includes lack of detail, or blurring of lines which should be sharp. Here we find troubles related to focus, interlace, resolution, smears, and snow.

For many of the faults listed throughout all the tables of troubles and remedies there are service adjustments or controls accessible to the operator. It is assumed that these adjustments and controls have been set for best possible reproduction, or to minimize whatever faults exist. Our listings give items to be checked when the regular controls and service adjustments will not remove the trouble. For example, poor focus most often is due to incorrect adjustment of the focusing control. This is not listed as a trouble, because you are supposed to have made all necessary service adjustments when first checking receiver performance.

II. 2. a. FOCUS INCORRECT.

It is necessary to distinguish between faulty focusing and lack of resolution or definition. The difference becomes apparent when looking at the raster, without a picture. If the focusing control of the receiver can be adjusted to produce fairly distinct and separate horizontal trace lines, there is nothing wrong with the focusing and you should check the causes for poor resolution. If horizontal trace lines cannot be seen, or if they can be made to appear at only one side or the other of the screen, proceed as follows.

Focusing coil or permanent magnet, or combination.

Too far back from yoke, or sometimes too close to yoke. Not centered around the neck of the picture tube.
Coil connections have bad joints, loose terminals.
Coil open circuited, short circuited.
Coil current too great or too little.
Permanent magnet is weak.
Note: The adjustment range of any type of focusing control should extend somewhat beyond the position of best focus, and should be capable of shifting the area of sharpest focus from side to side.

Fig. 93-1. Focus incorrect. The appearance here is not due to poor resolution, but to an electron beam spot which is too large to form distinct horizontal trace lines.

Ion trap magnet misadjusted, or weak.

Deflecting yoke too far back from flare of picture tube, is not against the cushion.

Picture tube, any type.
Anode voltage low, allows electron beam to produce a large and poorly defined spot.
Anode voltage too great for the current flowing in the focusing coil, the magnetic field of the coil is too weak to bring the beam to a sharp point at the screen.
Cathode-heater leakage in the picture tube.
Tube has become gassy, possibly due to previous misadjustment of the ion trap magnet.

Magnetic fields affecting the picture tube beam. A PM speaker, a low-voltage power filter choke, or other part may be too close to the neck or flare of the picture tube.

Picture tube, electrostatic type.
Check the entire circuit leading to the focusing anode. Incorrect voltage, leaky bypass capacitor, defective resistor or control pot.

II. 2. b. INTERLACING FAULTY

Dark spaces are clearly apparent between horizontal trace lines when the picture is seen from a normal viewing distance.

Lines and edges having a slight downward slope in the picture appear jagged, zig-zag, stepped, or have beads. This effect appears on the horizontal wedges of a test pattern, which may have a moire effect, like some kinds of satin, toward the center of the pattern.

When faulty interlacing causes alternate horizontal trace lines to overlap or coincide the effect may be called pairing. Pairing may be produced by carefully and slowly adjusting the vertical hold control to the point where the picture is just ready to slip downward. The effect is more easily observed if you temporarily increase the picture height, by adjusting the vertical size control.

Faulty interlacing is apparent when observing the bright, sloping, vertical retrace lines with the brightness control turned high. Alternate retrace lines are in the alternate fields of each frame. With good interlacing the alternate lines are evenly spaced, while with poor interlacing they tend to come close together or to pair.

Received signal may be temporarily at fault. Try other channels.

Fig. 93-2. Interlacing faulty. Note the fine horizontal lines which show up most clearly along the top of the picture.

Strong r-f interference is affecting the vertical sync action.

Vertical hold control adjusted too close to where the picture will roll downward. Defective capacitors or resistors in the hold control circuit may be keeping the free running vertical frequency too close to this point.

Sync section. Vertical and horizontal sync pulses not well separated by the integrating and differentiating filters. Check capacitors and resistors in these filters.
If parts of the sync section serve only the vertical sweep oscillator, and other parts serve only the horizontal sweep oscillator, check the tubes, tube voltages, capacitors, and resistors in these separate sections.

Vertical sweep oscillator. The horizontal sweep frequency of 15,750 cycles per second may be getting into the vertical oscillator circuit. Check the dressing of conductors in the oscillator circuits.
Resistors and/or capacitors in the grid circuit of the vertical oscillator may be leaky, of wrong value, or otherwise defective.

II. 2. c. RESOLUTION OR DEFINITION POOR

Note the difference between poor resolution and incorrect focus as explained with reference to focus, (Trouble Section II. 2. a.) Observe a raster, with no picture. A magnifying glass will help. If trace lines are distinct on the raster, but pictures are blurred, the trouble is not in focusing, and may be in lack of resolution.

Note also that excessive brightness almost always is accompanied by poor rendering of details. Check the brightness control setting.

Signal reflections which cause ghost images very close to the regular picture will give the appearance of poor resolution. Refer to the Trouble Section dealing with Ghosts.

The most common cause for poor resolution is lack of response or gain at the higher video frequencies. These are the frequencies which produce fine lines and small details in the picture.

Television pictures reproduced from motion picture films originally prepared for theatre use, rather than especially for television reproduction, practically always have poor definition or resolution. This cannot be improved at the receiver.

With some television transmissions carried wholly or in part by wire lines, the maximum video frequency is somewhat less than 3 mc. This limits the resolution, but not to an extent noticed by the viewer or owner unless he has been instructed in this matter.

A weak signal, whatever may be the reason, is likely to result in poor rendition of picture details because the set operator will employ excessive contrast, excessive brightness, or both, in attempting to improve the reception. Some of the reasons for lack of gain are discussed in the Trouble Section dealing with Snow.

Fig. 93-3. Resolution or definition is poor. This is not due to incorrect focusing, but rather to lack of gain at the high video frequencies and inability to reproduce sharp changes of illumination along horizontal trace lines.

When poor resolution is due to lack of high video frequencies the lights and shadows along horizontal trace lines cannot change from one to the other quickly enough to produce sharp vertical lines and edges in the picture, especially when such lines are closely spaced. This shows up when observing a test pattern, wherein the converging lines of the vertical wedges will blur and run together toward the center of the pattern when the resolution is poor.

Video amplifier circuit.
- Tube(s) weak, or operating with wrong voltages.
- Coupling capacitor leaky.
- Decoupling capacitor in plate or screen circuit is open.
- Peaking inductors shorted. Try placing a temporary short circuit around each of the inductors. If this causes no change in the picture rendition, it is probable that the peaker already is shorted and should be replaced with one of the correct inductance.
- Peaking inductor may be open circuited. This would cause a weak picture in addition to poor resolution. One of the inductors may be of wrong inductance for its position, in case repairs and replacements have been made.
- Overloading of a video amplifier tube due to excessive plate or screen voltage, or grid bias not

sufficiently negative. Overloading results also from applying a grid signal too strong for the bias. This could result from defects in the contrast control circuit or in the automatic gain control system.

Video detector circuit.
Weak tube or crystal.
Load resistor of wrong value.
Coupling capacitor leaky.
Check the peaking inductors as explained for the video amplifier circuit.

Fine tuning control misadjusted.

Tuner.
R-f oscillator incorrectly aligned for channel(s) in which trouble appears, or in the overall alignment adjustments.
Band width too narrow. R-f amplifier grid and plate couplings not aligned correctly.

I-f amplifier section.
Alignment wrong. Response does not extend far enough toward the sound side of the curve, where the higher video frequencies are carried as modulation.
Tube(s) weak.
Coupling capacitor open.
Grid resistors of wrong value.

Transmission line poorly matched to antenna or receiver. Signal reflections in the line may cause loss of high frequencies.

II. 2. d. SMEARS OR TRAILERS.

Light-toned vertical or sloping lines or edges are followed on the right by dark areas, while dark-toned lines or edges are followed by light areas. In some cases the reversals of shading appear on the left rather than on the right.

When smearing is severe it is most noticeable near the larger objects in the picture. Dark objects may be closely followed by a narrow light band which is followed by a dark trailer extending to almost any distance across the picture. Light objects may be followed by a narrow dark band which is followed by a light trailer.

Horizontal lines or bands which are normal parts of the picture may appear to extend right on through objects toward the right, so that these objects are seen rather indistinctly through the line or band. This may be called the X-ray effect.

Ghosts might be mistaken for trailers or smears. The difference is that usually there is reversal of tone, light and dark, with smearing, while a ghost image repeats the same tones as the picture, but at a greater or less separation from lines in the picture.

Fig. 93-4. Smears or trailers. This picture shows an extremely bad condition. Even slight smearing causes unsatisfactory reproduction.

Very slight smearing causes the appearance of poor resolution, but the causes for the two troubles are not the same.

The usual cause for smearing or trailers is poor response or gain at the low video frequencies, with low-frequency phase shift.

Contrast control turned too high, overloads the video amplifier.

Received signal may be temporarily at fault. Try other channels.

Video amplifier circuit.
- Tube(s) weak.
- Plate or screen voltages too low.
- Grid bias not sufficiently negative, allows flow of grid current on peaks of strong signals.
- Coupling capacitor leaky, open, of too small value. This applies also to the coupling capacitor between amplifier and picture tube.
- Decoupling capacitor in plate or screen circuit open or of too small capacitance. Check also the cathode resistor bypass capacitor. Decoupling resistors may be of too small value. These faults may allow degeneration and loss of gain at low frequencies.

Peaking inductors open. Shunt resistor open. Wrong inductance, allowing excessive peaking or response within the video frequency band.

Video detector circuit.
Load resistor of too great value.
Check coupling capacitors and peaking inductors as noted above for video amplifier circuit.

I-f amplifier section.
Incorrect alignment, with video intermediate frequency misplaced on response curve.

II. 2. e. SNOW IN PICTURE

The appearance of snow in a picture indicates a low ratio of signal to noise. That is, the picture signal is too weak to ride well above the noise impulses which are inherent in tubes and amplifier circuit elements. The flecked or snowy appearance of the picture is due to the noise impulses which originate ordinarily in the tuner circuits, and are amplified by following i-f and video amplifier stages.

Snow may be prevented in two ways. First, by bringing a stronger signal to the receiver. This

Fig. 93-5. Snow in the picture. This indicates a received signal which is weak in proportion to tube and circuit noise impulses in the first amplifying stages of the receiver.

requires attention to the antenna and transmission line. Second, by increasing the gain in the r-f stage or stages, in the mixer, and possibly in the first i-f stage. Gain added in stages farther along in the receiver will increase the noise as much as the signal, and will not reduce the snow.

Snow may appear when the receiver is tuned to a channel in which the signal field strength is low, and may disappear when tuned to another channel in which signal field strength is relatively high at the locality of the receiver. Snow increases as the contrast control is turned higher. It is least noticeable in light-toned areas, and most apparent in dark gray and black areas of the picture.

Antenna.
Check the orientation and location.
Connections to transmission line corroded, open, short circuited.
The antenna may be of a type having too little gain for signal field strengths in its locality.

Transmission line.
Poor, dirty or loose, connections. Open circuit or short circuit.
Mismatched at the antenna, the receiver, or both.
Of excessive length.
Running too close to metal objects, or where interference is picked up.

Tuner.
Tube(s) weak, operating with wrong voltages.
Switch contacts dirty, loose, making insufficient contact.
Incorrect alignment in antenna and r-f couplings.

Automatic gain control maintaining an r-f amplifier grid bias too negative when receiving weak signals. It may be desirable to operate the r-f amplifier with a fixed bias of low value.

I-f amplifier section.
Tube(s) weak, operating with wrong voltages.
Misaligned, with video intermediate frequency too low on the response.
Insufficient gain may require settings of the contrast control so high as to emphasize the snow effect. This is true also of the video detector and video amplifier.

Video detector. Weak tube or crystal.

Video amplifier. Tube(s) weak, operating with wrong voltages.

Corona or brush discharge in the high-voltage power circuits may cause an effect somewhat similar to that of a weak signal by producing a snow effect.

<u>Increasing the gain.</u>
The following methods sometimes are used for increasing the gain in localities where signal field strength is weak.

R-f amplifier.

Align for narrower passband and higher peak of response, with the video carrier at or near the peak.

Substitute a tube of greater transconductance. This may or may not require changing the socket wiring to accomodate the elements of the new tube. Many of the present high-gain tubes were not available when older receivers were designed.

If the r-f amplifier is on the agc system, disconnect the grid resistor from the agc bus and connect its low end to chassis ground or to B-minus.

Substitute a grid resistor and/or a plate load resistor of greater resistance. An increase of more than 20 to 30 per cent of the original value may result in poor frequency response. A high grid resistor may affect the input impedance of the receiver and cause mismatching.

When video carrier or video i-f response is raised, there is likely to be weakening of the sound. This must not be carried so far that satisfactory sound response cannot be had by increasing the volume control setting.

I-f amplifier section.

Substitute a tube or tubes of greater transconductance, as mentioned in connection with the r-f amplifier.

Change the alignment to bring the video intermediate frequency up around 80 per cent of maximum gain, instead of the usual 50per cent. Raise the response peak as much as possible without losing the sound. If carried too far, this may cause regeneration and possibly oscillation in the i-f amplifier due to similar frequencies.

Operate the tubes with higher plate and screen voltages. Where the original voltage is 135 or 150, it may be raised to something like 180 or 200 volts.

Our next subdivision deals with troubles which cause various kinds of distortion, with portions of the picture moved out of their normal positions. Here we find a large number of defects in picture reproduction, as listed in the original outline for systematic trouble shooting.

II. 3. a. BENDING OR PULLING.

Bending refers to curvature of vertical lines toward either the right or left. This fault usually occurs only at the top of the picture, but may extend downward to any distance.

Except for curvature of the vertical outlines, objects remain of correct proportions.

When there is slight bending at the top of the picture, vertical lines sometimes may be made to slope toward either the right or the left, or to remain straight, by adjustment of the horizontal hold control.

Bending occurs when horizontal trace lines of the picture do not hold precise synchronization, some of them commencing and ending later than others. The cause most often is weak or erratic horizontal sync pulses.

Received signals very weak. There may be bending on some channels and not on others.

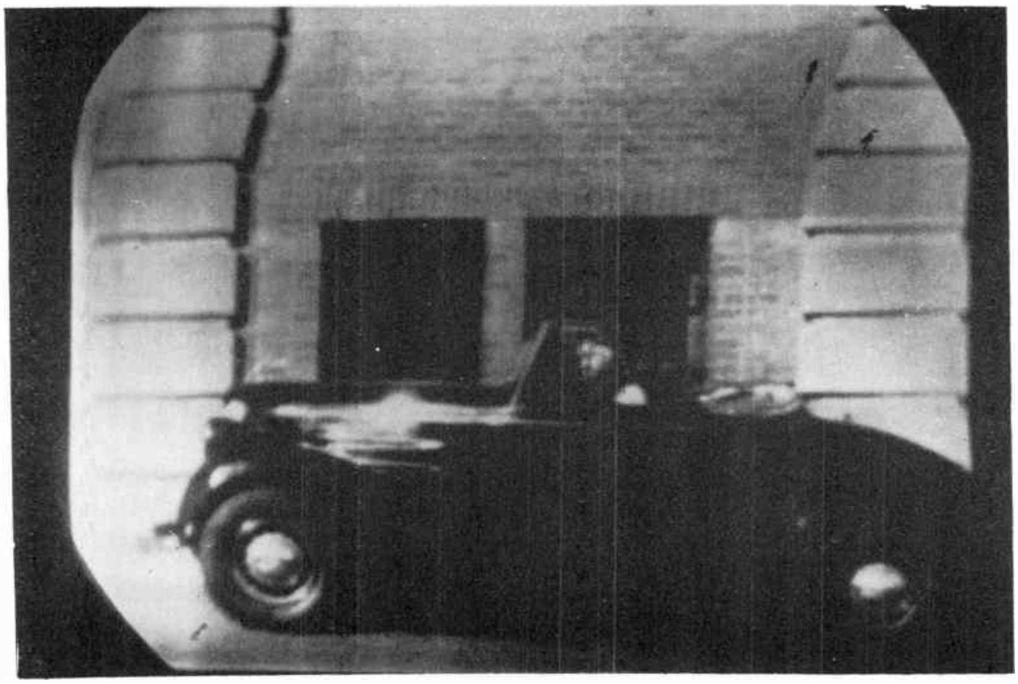


Fig. 93-6. There is moderate bending or pulling toward the right, at the top of the picture. There might be similar bending toward the left.

Interference, usually of the spark type, may be interfering with sync.

Heater-cathode leakage may exist in any tube of the r-f amplifier, the mixer, the i-f amplifier, the video amplifier, the sync section, or the horizontal afc system. Such leakage may allow the 60-cycle heater frequency to enter the other circuits and vary the horizontal sync timing.

R-f alignment. Video carrier too low on the response.

I-f alignment. Video intermediate too low on the response.

Horizontal afc system.

The frequency, free running, of the horizontal sweep oscillator may be too far off for the control pulses to synchronize it. This is a matter of adjustment according to the type of afc system.

Control tube microphonic. Try lightly tapping this tube to note whether the bending is affected.

Voltages at other than the horizontal sweep frequency may be getting into the afc circuits, as from circuits for an audio amplifier, a video amplifier, a vertical sweep oscillator, etc.

With a phase detector or discriminator as the afc tube, the d-c voltages reaching the two sides of the twin diode may be unequal or may vary. Check all capacitors and resistors in circuits leading to elements of this tube, especially where equal resistances are connected to each of the diodes.

Fig. 93-7. Severe bending or pulling extending all the way from top to bottom of the picture. This picture has been moved to the left in order to show bending of the right-hand edge, which would show up also on a raster.

Sync section.

Tube(s) weak, operating with wrong voltages.

Check all coupling and decoupling capacitors and resistors for opens, leakage, poor connections, wrong values.

The signal reaching the input to the sync section may be too weak, or possibly too strong. Weak input signals weaken the sync pulses at the output, while excessively strong signals prevent necessary clipping and limiting.

D-c restorer.

When the input for the sync section or the afc system is taken from the restorer circuit, check for a weak tube and for incorrect operating voltages.

Video amplifier.

Weak tube(s), operating with wrong plate or screen voltages.

Grid bias voltage too negative. Check cathode bias resistor.

Coupling capacitor open, or of too little capacitance.

Plate load resistor of too low resistance.

Excessively strong signal to the grid may cause plate current saturation if the tube(s) are normally operated with fairly low plate voltages. This will result in cutting of the sync pulse peaks.

Check the action of the agc system.

II. 3. b. CENTERING INCORRECT.

Centering of the picture in the mask usually is accomplished by tilting the focusing coil or magnet with reference to the neck of the picture tube. In some receivers there are electrical centering controls which vary the amount of direct current in the deflecting coils. When these electrical controls will not center the picture, they should be set at the center of their adjustment ranges while the focusing coil is tilted for correct centering. If the picture cannot thus be centered, or if excessive tilting of the coil is required, check all capacitors and resistors, including the pots, in the electrical control circuits.

Focusing coil or magnet.

Not centered with reference to the axis through the picture tube neck.

Too far back from the deflecting yoke.

Permanent magnet weak.

Fig. 93-8. This picture is off center toward the right and toward the bottom. Centering may be faulty in any direction, upward, downward, or toward either side.

Ion trap magnet in wrong position, very weak.

Deflecting yoke too far from flare of picture tube, not against cushion.

PM speaker located too close to neck or flare of picture tube.

Horizontal hold control cannot be adjusted far enough one way or the other. Check capacitors and resistors in the hold control circuit.

Horizontal afc misadjusted to an extent that the operator's hold control adjustment lacks enough range to move the picture to either the left or right.

Electrostatic picture tube.
Coupling capacitor from sweep amplifier to deflecting plates is leaky.
A magnetic shield around the picture tube has become permanently magnetized.

II. 3. c. LINEARITY FAULTY, HORIZONTAL.

Transmitted pictures are not always perfectly linear. The variations are, however, much too

Fig. 93-9. Horizontal non-linearity here causes crowding at the left and stretching at the right.

Fig. 93-10. Horizontal non-linearity here is causing compression at the center of the picture area.

is small to be noticed in any picture program and can be detected only by observation of a test pattern. Receivers very seldom are adjusted for perfect linearity, and some cannot be so adjusted. Slight defects in linearity are far more likely to result from receiver faults than from the received signal.

A quick check for linearity consists of watching objects of some regular or symmetrical form as they move from place to place in the picture. If the form or shape of such objects appears to change as they move, the reproduction doubtless is decidedly non-linear.

As shown by Fig. 93-11, horizontal non-linearity may show up as (a) crowding at the left with stretching at the right, (b) crowding at the center with some stretching on both sides, or (c) crowding at the right with stretching at the left. Which fault exists depends on circuit design in the receiver and on the kind of trouble present.

Linearity control.
  Capacitor or resistor shorted or open.
  Inductor on output transformer shorted.

Drive or peaking control.
  Incorrect adjustment.

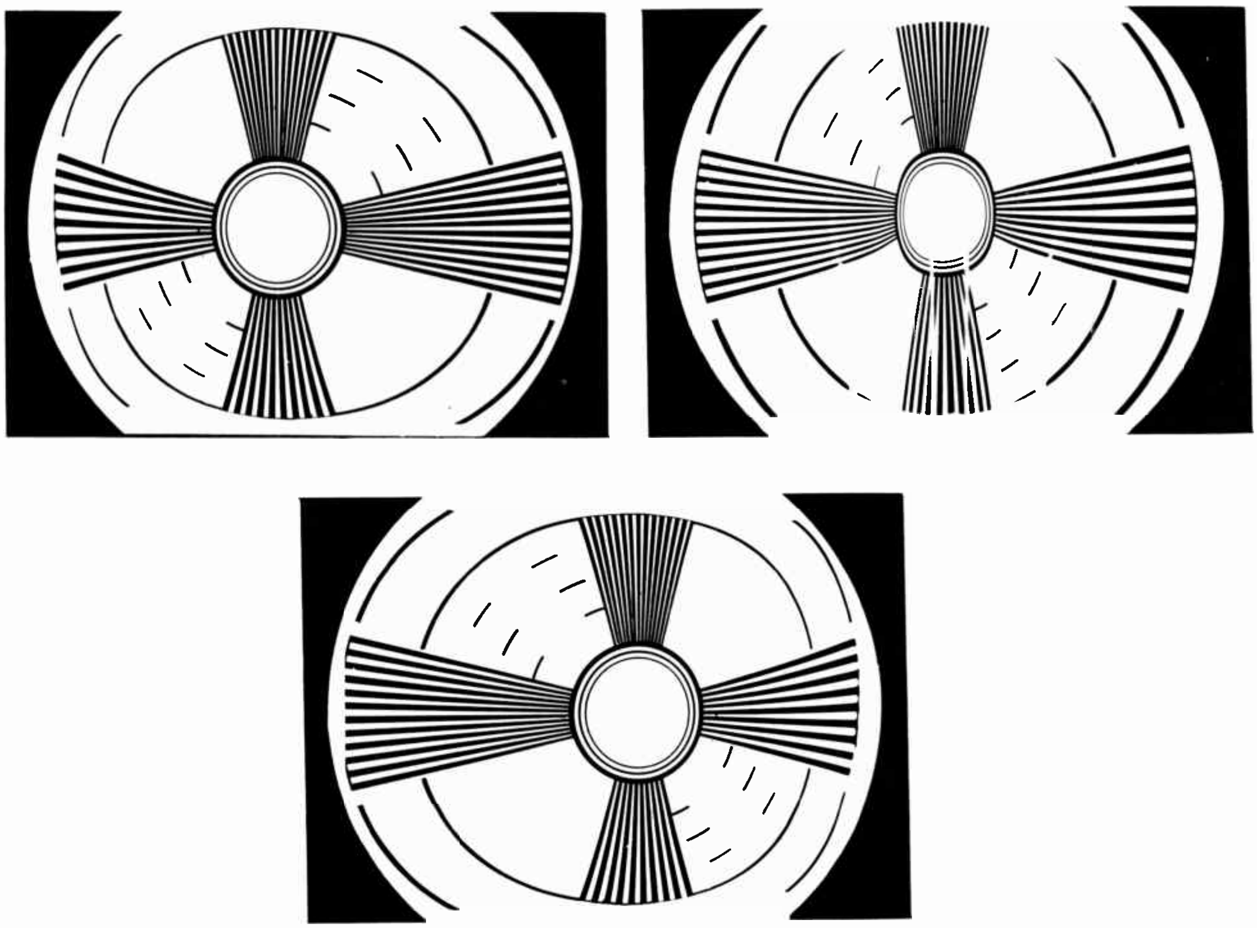


Fig. 93-11. Horizontal non-linearity as shown by test patterns. A, the left side is crowded. B, the center is crowded. C, the right side is crowded.

Open capacitor or resistor in peaking circuit.

Horizontal size control.
Adjustment of any type of size control usually affects linearity. Size, drive, and linearity adjustments have to be readjusted to secure a correct balance.

Focus coil or magnet in wrong position along neck of picture tube.

PM speaker too close to picture tube flare or neck.

Damper circuit.
Tube weak, plate voltage too high.

Shunt or series resistor of wrong value, or open.
Connection to wrong tap on resistor shunting the damper.

Horizontal sweep amplifier circuit.
Weak tube, operating with wrong voltage on screen or control grid.
Capacitor on grid, cathode, or screen is open or leaky.
Grid resistor of too small resistance, overloads the sweep oscillator.

Horizontal sweep oscillator circuit.
Tube weak, operating with wrong voltages.
Coupling capacitor leaky.
Coupling resistor open, or of wrong value.
Sawtooth capacitor of wrong value, leaky.

Horizontal output transformer defective.

Deflecting yoke circuit.
Excessive series resistance, high-resistance short or ground.
Capacitor from this circuit to B-plus line open or leaky.
Yoke not suited to deflection circuit, replacement unit is of wrong type or wrong Q-factor.

Fig. 93-12. Vertical non-linearity most often causes stretching at the top of pictures. Horizontal trace lines then are farther apart at the top than lower down.

II. 3. d. LINEARITY FAULTY, VERTICAL

Vertical non-linearity most often is apparent in stretching at the top of pictures, but sometimes there is crowding at the top combined with inability to make the picture high enough to fill the mask.

Linearity control.
Resistor or control "pot" open circuited.
Resistor of wrong value.
Capacitors, usually in bypass positions, leaky or wrong values.

Vertical size control.
The vertical size control and the vertical linearity control must be adjusted during the same operation to produce an undistorted picture which fills the mask from top to bottom.

PM speaker too close to picture tube flare or neck.

Vertical sweep amplifier circuit.
Tube weak, operating with low voltage on plate and/or screen, incorrect grid bias.

Fig. 93-13. Vertical non-linearity may cause crowding at the top. The more closely spaced trace lines then cause excessive brightness at the top.

Coupling or decoupling capacitor open, small capacitance, leaky.
Coupling or decoupling resistor of wrong value.

Vertical sweep oscillator circuit.
Tube weak, operating with wrong voltages.
Coupling or bypass capacitor open, wrong value, leaky.
Sawtooth capacitor of wrong value, leaky.

Vertical output transformer shorted, grounded, defective.

Deflecting yoke circuit.
Vertical coil shorted or grounded.
This circuit may be picking up a 60-cycle voltage from a heater circuit, with the result that there is faster vertical deflection at the top or bottom of the picture. However, this condition will usually be moving up or down.

Note: Vertical non-linearity may be detected by close observation of horizontal trace lines on a raster, with no picture. With satisfactory linearity the lines will be uniformly spaced from top to bottom. With non-linearity the spacing will be greater at either the top or the bottom than at the other end of the raster.

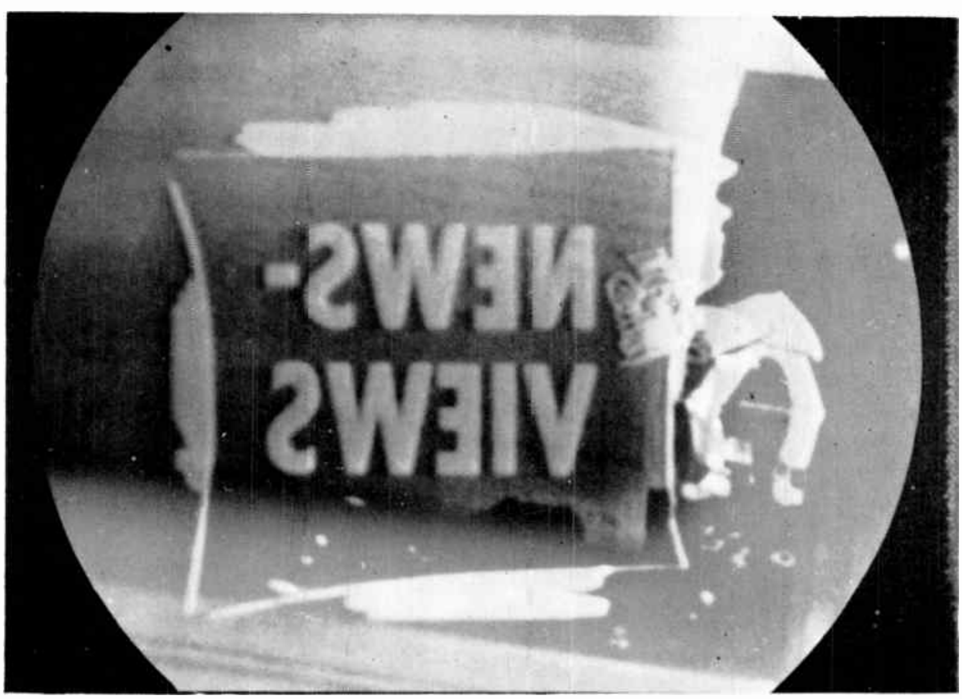


Fig. 93-14. Reversed pictures may appear backward, from right to left, or they may be upside down, or they may be both backward and upside down.

II. 3. e- REVERSED PICTURE. (When caused by previous repair or "tampering" by the owner.)

If the reversal is left to right, as in the picture, the leads to the horizontal deflecting coils in the yoke have been interchanged.

If the picture is upside down, with reversal top to bottom, the leads to the vertical deflecting coils have been interchanged.

Reversal in both directions at the same time indicates that leads have been interchanged to both sets of coils in the yoke.

Ordinarily there will be no other distortions in the pictures, although centering and size may be rather difficult to adjust correctly.

II. 3. f. SIZE INCORRECT, HORIZONTAL

Too narrow, will not fill mask sideways.

Note: The controls for horizontal size, horizontal linearity, and drive or peaking must be ad-

Fig. 93-15. Horizontal size here is too narrow. It might also be too wide.

justed together or during the same service operation in order to produce a picture of correct width without distortion in the horizontal direction.

Horizontal size control.
Control potentiometer shorted, open circuited.
Bypass capacitor shorted or very leaky.

Horizontal sweep amplifier.
Tube weak, operating with plate and/or screen voltage too low, or with grid bias excessively negative.
Weak sawtooth voltage to grid, check the horizontal oscillator.
Grid resistor of too little resistance, partial short.
Coupling capacitor or bypass capacitor of wrong value.

Damper circuit.
Tube weak, operating with plate voltage too low.
Shunt or series resistor of wrong value, connection to an incorrect tap which reduces boost voltage.

Horizontal sweep oscillator circuit.
Tube weak, operating with voltage too low.
Coupling capacitor of too small capacitance.
Sawtooth capacitor of too great capacitance.
Blocking transformer defective.

Horizontal output transformer.
Open circuited (high resistance), short circuited, grounded.

Horizontal deflecting coil circuit.
Too little sawtooth current. High resistance or impedance in series.
Capacitor open, disconnected, too little capacitance.

Too wide, sides of picture cannot be brought within mask.

Check the adjustment of the drive or peaking control.

Horizontal size control on flyback transformer.
Inductor disconnected or open circuited.
Capacitor across taps on secondary winding of wrong value or connected between wrong taps.

Horizontal sweep amplifier.
Check as for, "size too narrow", but for opposite conditions; tube voltages too high, excessive sawtooth voltage, etc.

Horizontal sweep oscillator.
Check as for "size too narrow", but for opposite conditions; voltages too high, sawtooth capacitor too little capacitance, etc.

Note: Folding in the horizontal direction may produce the appearance of too little width, since part of the picture is folded over the remainder. Refer to Trouble Section on Folding.

II. 3. g. SIZE INCORRECT, VERTICAL.

Note: The controls for vertical size and vertical linearity must be adjusted at the same time to have the picture fill the mask vertically without distortion in the vertical direction.

Size insufficient, will not fill the mask vertically.

Vertical size control.
Control potentiometer defective.
Bypass capacitor defective.

Vertical hold control.
Control potentiometer defective.
Coupling or bypass capacitor leaky or shorted.

Vertical sweep amplifier.
Tube weak, operating with low plate and/or screen voltage.

Fig. 93-16. Vertical size here is too little. It might also be too great.

Grid bias too negative.
Weak sawtooth voltage to grid, check the vertical oscillator.
Coupling and/or bypass capacitors leaky, open, of wrong value.
Coupling resistor, change in value.

Vertical sweep oscillator circuit.
Tube weak, operating with voltage too low.
Grid resistor, change in value.
Plate load resistor (if used) of too little resistance.
Blocking transformer defective.
Coupling capacitor to amplifier, too small capacitance.
Sawtooth capacitor of too great capacitance.

Integrating filter, sync section to vertical oscillator.
Check for leaky or shorted capacitor, for resistor of wrong value, partial short, open connections.

Vertical output transformer.
Partial short circuit grounded, series resistance.

Note: Folding in the vertical direction may produce the appearance of too little height. Refer to Trouble Section on Folding.

<u>Picture too high, top and bottom of picture are outside the mask.</u>

Vertical linearity control misadjusted. Keep in mind that the vertical size and linearity must be adjusted at the same time.

Vertical size control.
Control potentiometer misadjusted or defective. Bypass capacitor defective.

Vertical sweep amplifier.
Excessively great sawtooth voltage to the grid. Check the vertical oscillator circuit.

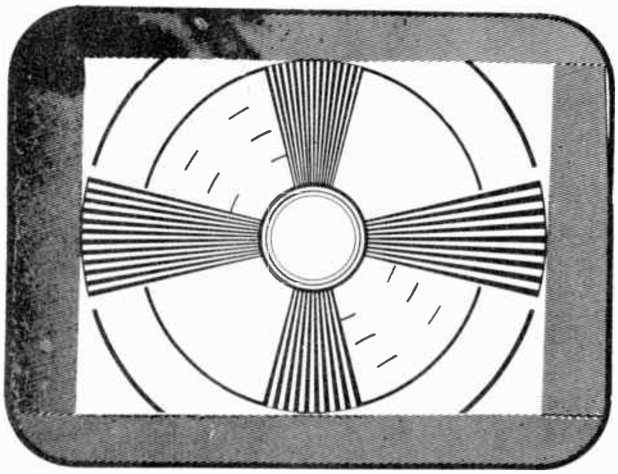
Vertical oscillator circuit.
Excessive plate voltage on tube.
Sawtooth capacitor of too small capacitance.

II. 3. h. SIZE INCORRECT, BOTH VERTICAL AND HORIZONTAL.

<u>Too small, picture will not fill mask in either direction.</u>

Power line voltage low.

B-voltage low from low-voltage power supply.
Check rectifier(s), preferably by substituting new ones.



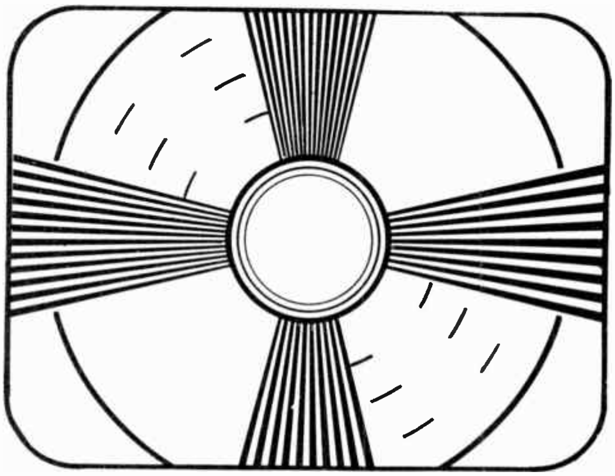


Fig. 93-17. The picture is too narrow, and lacks sufficient height, to fill the mask. Size too small or too large shows clearly on test patterns.

Measure a-c voltage from secondary of power transformer. Low voltage usually indicates an overload (short or ground) somewhere in the B-voltage lines or circuits of the receiver, but might indicate a defective power transformer.

Low B-voltage on line supplying both the vertical and horizontal sweep amplifiers and/or sweep oscillators. Check series resistors, bypass and decoupling capacitors for shorts or leakage after comparing voltage readings with chart.

Picture tube anode voltage too high, prevents sufficient deflection of the electron beam.

Damper tube weak, or faults in the damper circuit, if boosted voltage goes to both the horizontal and the vertical sweep circuits.

Picture tube defective. If gassy will cause poor resolution as well as reduced picture size.

Too large, pictures extend beyond mask in both directions.

Picture tube anode voltage too low, allows excessive deflection of the beam. This will be accompanied by lack of brightness. Check the high-voltage power supply rectifier, filter resistor and capacitor, connections to the anode cap or rim.

Expansion or "blooming" of the picture when the brightness control is turned up. Too much filter resistance or other high resistance in series with picture tube anode lead. Defective picture tube.

Brightness control.
B-voltage to the control circuit too high or too low. (Excessive brightness).
Bypass capacitor leaky or shorted causing excessive brightness.

II. 3. i. SIZE VARIES IRREGULARLY.

Magnetic field from power transformer reaching picture tube. If the power line frequency and vertical deflection frequency are not exactly alike there may be slow expansion and contraction of the picture. This may be due to lack of a closed copper band, single shorted turn, around the core of the power transformer.

Varies when brightness control operated. Some variation is normal in many receivers. Otherwise there may be excessively high resistance from high-voltage rectifier to picture tube anode.

Varies with adjustment of centering control of the electric type, not a focus coil or magnet. Defective bypass capacitor across the centering control potentiometer.

II. 3. j. TILTING OR SKEWING.

Magnetic deflection picture tube.
Loosen and rotate the deflecting yoke in its support until the picture is square within the mask or until horizontal trace lines of a raster are parallel with the top or bottom of the mask.
Focus coil or magnet may be too close to the deflecting yoke. With a focusing coil the picture may tilt one way or the other when the focusing control is adjusted.

Electrostatic deflection picture tube.
Loosen the tube in its supports and rotate the tube around its lengthwise axis to make the

Fig. 93-18. Tilting of the picture within the mask.

picture square within the mask, or until trace lines of the raster are parallel to the top or bottom of the mask.

II. 3. k. WAVY OR RIPPLED.

Usually the greatest waviness is at the left in the picture, decreasing gradually toward the right, although the entire picture area may have waves running vertically. The left-hand edge of the picture may be ragged or stepped. In some cases the entire picture may appear to shift rapidly back and forth from left to right.

When the trouble is due to power line or low audio frequencies there may be alternate light and dark horizontal bands. The number of pairs of bands or the number of complete wave cycles is equal to the troublesome frequency divided by 60. If the troublesome frequency is rather high, say 500 to 1,000 cycles, everything in the picture appears to wriggle horizontally in many small waves.

Contrast control turned too high.

Drive control incorrectly adjusted.

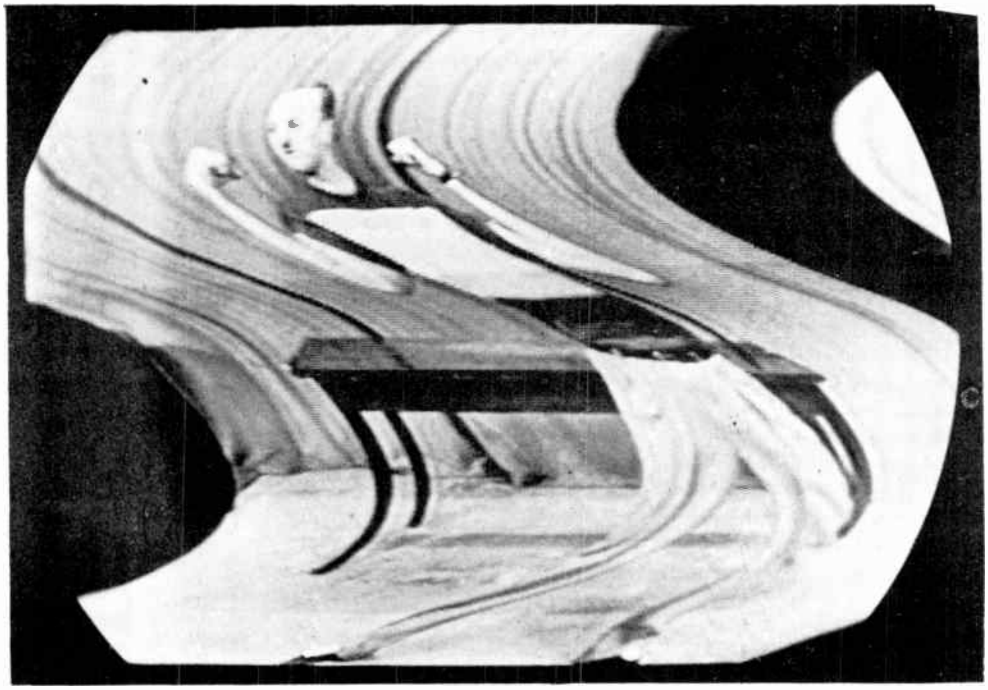


Fig. 93-19. Wavy picture, a severe case. Note that black horizontal blanking bars cut into both sides. Here there is only one complete wave. There may be any number, appearing as vertical ripples.

Strong interference, of either smooth waveform or of spark types.

Magnetic field too near the flare or neck of picture tube, as from a power transformer, filter choke, or speaker field coil.

Low-voltage power supply, excessive ripple voltage.
Filter capacitor open, too little capacitance, poor connections.
Filter resistor or choke shorted, of too small value.
Defective capacitor anywhere in a transformerless power supply.

Sync section.
Poor filtering of B-voltage to tubes.
Decoupling capacitor open in a plate or screen circuit.

Horizontal sweep oscillator.
Poor filtering of B-voltage to tube(s).
Cathode-heater leakage.
Coupling or decoupling capacitor open or disconnected.
Grid bypass or decoupling capacitor open.
Oscillation due to unwanted self-resonance, try another tube.

Horizontal afc circuit.
Correction voltage is varying too rapidly, or with too great amplitnde. There is hunting action, with horizontal sync frequency going alternately too high and too low.
Capacitor or resistor of wrong value, or otherwise defective. Check units in a noise filter, if used.

Horizontal sweep amplifier.
Tube defective, cathode-heater leakage.
Poor filtering of B-voltage to tube.
Oscillation due to internal capacitance of tube, try another tube.

Picture tube input circuit.
Grid or cathode lead too close to conductors in the horizontal sweep oscillator circuit, or power line circuits.

Damper circuit.
Tube weak or otherwise defective.
Plate voltage too low.
Wrong connection to a tapped resistor, or a tapped output transformer.

Horizontal output transformer defective, high resistance short or ground.

Deflecting yoke circuits.
Coupling or leakage between vertical and horizontal coils.
Hum voltage at 60 or 120 cycles getting into the horizontal deflecting coil.
Shunt capacitor on horizontal coil leaky, open, or too great capacitance.
Shunt resistors on vertical coils wrong value or otherwise defective.
Deflecting coils have been forced out of shape or displaced with reference to one another.

II. 3. l. WEDGE SHAPED PICTURE.

The picture may taper toward the bottom, as in the illustration, or toward the top, or toward either the left or right-hand side.

The usual cause is that one horizontal deflecting coil or one vertical deflecting coil is delivering a stronger magnetic field than the other coil of the same pair. Vertical tapering, as in the illustration, is caused by faults in the horizontal coils, while horizontal tapering is caused by faults in the vertical coils.

Deflecting yoke.
One coil of a pair shorted or open circuited, or has shorted turns.
Capacitor or resistor across one coil shorted, open, wrong value.
Yoke tilted with reference to axis of picture tube.
Coil has shifted position, replace the yoke.

Focusing coil or magnet.
Wrong position on tube neck.
Non-uniform magnetization in permanent magnet type.

Fig. 93-20. Wedge shaped picture. Note that everything tapers inward toward the bottom. The wedge form might taper upward, or toward either side.

Ion trap magnet.
 Non-uniform magnets. Might possibly be due to wrong adjustment.

Magnetic fields.
 PM speaker or filter choke carrying nearly pure direct current is too close to neck or flare of picture tube.
 Metal cone of picture tube may be magnetized. The distortion will rotate when the tube is rotated, regardless of the angle of the deflecting yoke.

Electrostatic picture tube.
 High series resistance value.

The final subdivision in this class of faulty pictures, where no bars or lines are added to the image, includes troubles which cause movement of the picture as a whole. The movement may be great enough to make the picture unrecognizable, but when the movement is stopped the picture will appear in correct proportions.

II. 4. a. FLICKERING.

Flickering, as here considered, consists of a rapid variation in brightness, which causes apparent movement of the picture without actual change of image position.

Interference affecting strength of received signal. One example is so-called airplane flutter caused by signal reflections from low-flying planes.

Temporary trouble at the transmitter, in the signal.

Automatic gain control circuit.
Defective agc tube.
Time constant too short, caused by faulty capacitor or resistor.
Loose connections.

D-c restorer circuit.
Defective tube.
Wrong time constant, due to faulty capacitor or resistor.

I-f amplifier.
Regeneration and slight intermittent oscillation, due to incorrect alignment, wrong dressing of leads, faulty shields or shield grounding, etc.

II. 4. b. JUMPING, JITTERY

Contrast control set too high. May overload the video amplifier and reduce the amplitude of sync pulses.

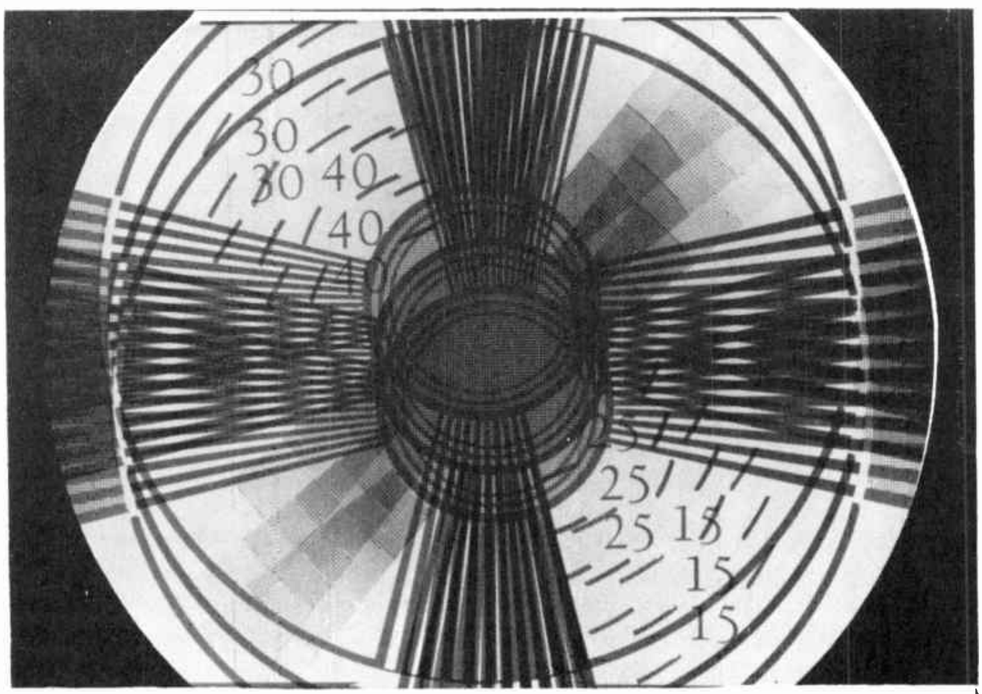


Fig. 93-21. Jumping or jittery picture. The jumpiness may be in either the vertical or the horizontal direction.

Received signal temporarily at fault. Try other channels.

Interference, of spark type, affecting the sync pulses.

Microphonic tube in the tuner, i-f amplifier, video detector, or video amplifier. If lightly tapping a tube increases the jumping, that tube may be microphonic provided there are no loose connections at the socket.

Tuner.
Poor connections or joints. Examine especially the switch contacts, and connections in the fine tuning circuit.

Sweep oscillator circuits.
Tube defective, possibly contains shorted or loose elements.
Capacitor leaky, faulty resistor.
Loose connections.
Grid resistor or capacitor of wrong value.

Vertical hold control circuit.
Examine the capacitors, resistors, and connections. Check operation of the control potentiometer for smoothness.

Sweep amplifier circuits.
Tube defective, especially in the horizontal amplifier position.

Picture tube input circuit.
Leads to the grid or cathode may be too close to the cable for the deflecting yoke.

Drive control.
Incorrect adjustment, defective capacitor (either fixed or adjustable), or defective resistor sometimes will cause jumping.

II. 4. c. SYNC NOT HOLDING HORIZONTALLY.

When the frequency of the horizontal sweep oscillator differs from horizontal pulse frequency of a received signal, the reproduction contains a greater or less number of horizontal or sloping bars. Between each pair of dark bars is a complete picture or pattern, greatly compressed. The greater the number of bars, the further the sweep oscillator is from synchronization. As oscillator frequency is brought closer to that of the received signal, the number of bars decreases and they become more nearly vertical. Each dark bar results from a period of vertical blanking. Adjustment of the horizontal hold control should reduce the number of bars and cause the picture to lock into horizontal sync.

Horizontal hold control circuit.
Capacitor leaky, or value has changed.
Resistor open, or of wrong value. Defective control pot.
Insulation leakage in circuit wires, examine the dressing.

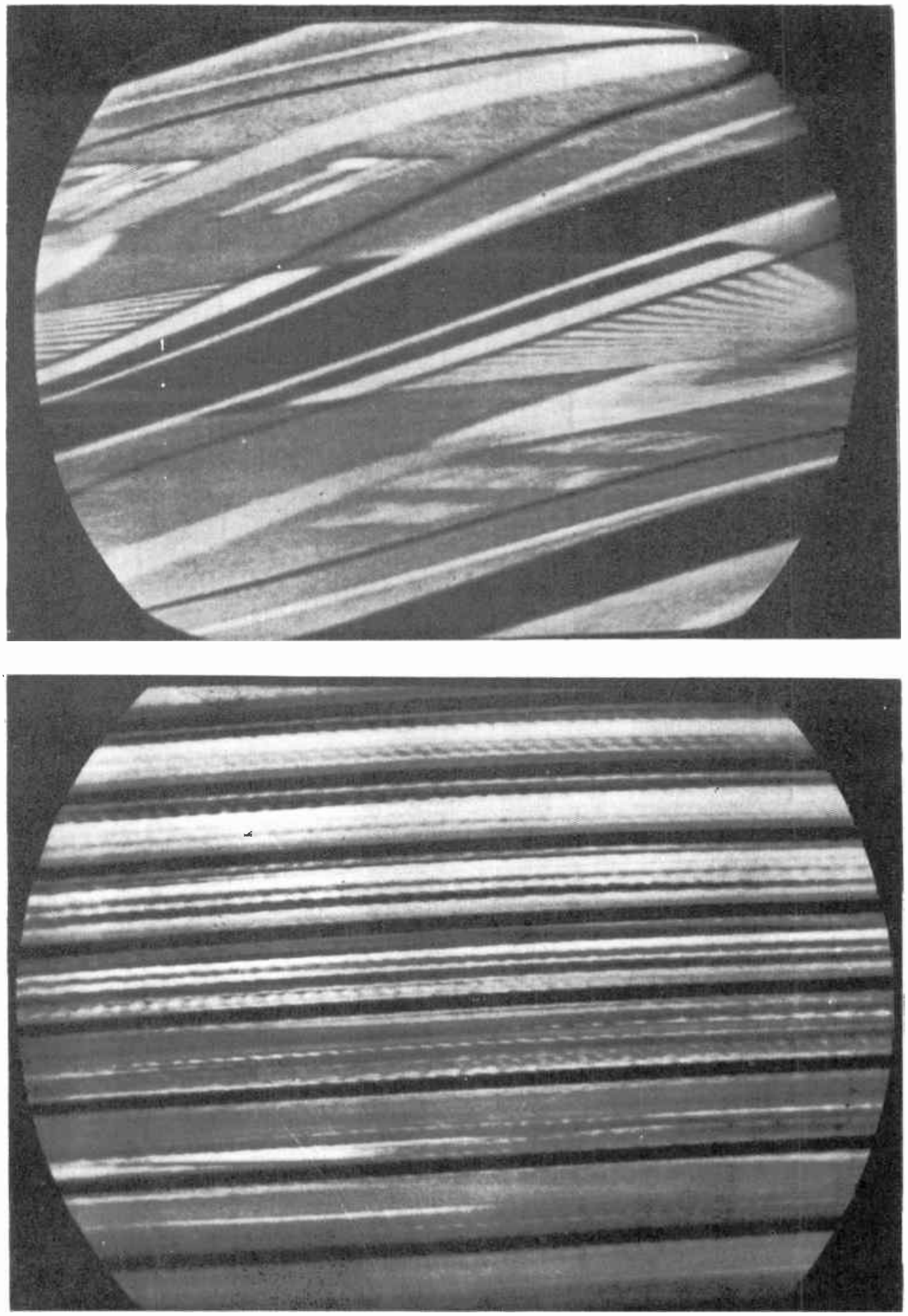


Fig. 93-22. Horizontal sync not holding. The number of bars will decrease, and they will become more nearly vertical, as the picture approaches synchronization or lock-in.

Horizontal afc system.
Tube weak. Incorrect voltage on plate, screen, grid, or cathode.
Capacitor leaky, of wrong value. Input pulses too weak.
Resistor open, of wrong value.

Horizontal oscillator.
Tube weak, operating with wrong voltages.
Coupling capacitor open, leaky, wrong value. Blocking transformer faulty.
Resistor open, of wrong value, partially shorted.

Differentiating filter between oscillator and sync section.
Check for opens, shorts, leakage in capacitor and resistor.

II. 4. d. SYNC NOT HOLDING VERTICALLY.
When the frequency of the vertical sweep oscillator differs from vertical sync frequency of the received signal, the picture will move upward or downward on the screen. With a great difference between the frequencies the motion will be rapid, and as the difference becomes less the picture will jerk upward or downward, holding only briefly.

Vertical hold control circuit.
Capacitor leaky, or value has changed during use.
Control pot or fixed resistor open, wrong value, otherwise defective.

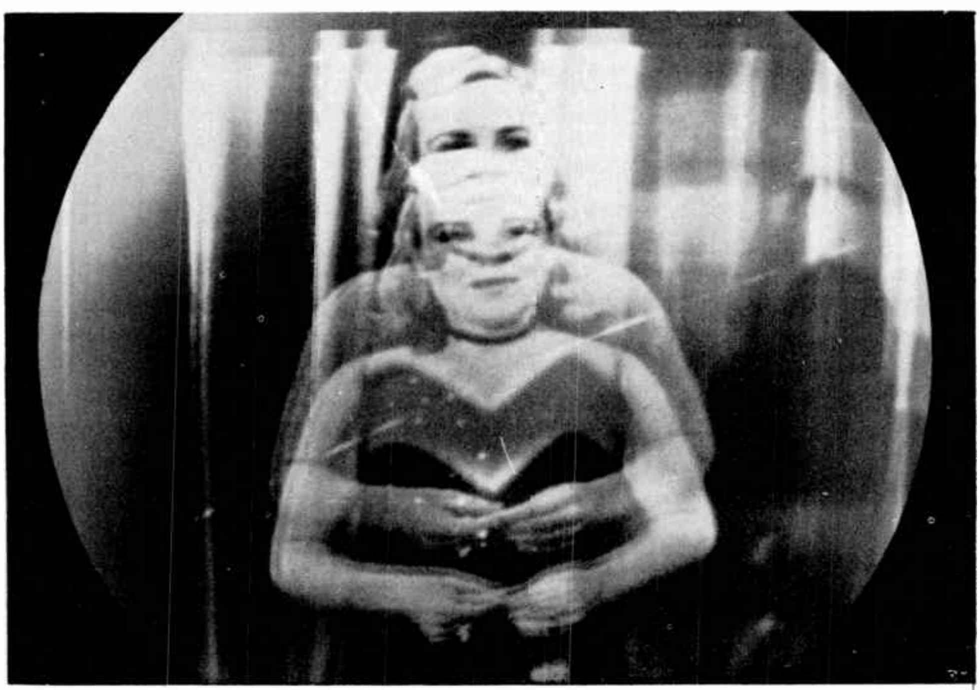


Fig. 93-23. Vertical sync not holding. The picture moves either upward or downward, steadily or in successive jumps.

Vertical oscillator circuit.
Tube weak, operating with low plate or screen voltage, wrong bias.
Coupling capacitor, leaky, open, shorted, wrong value.
Blocking transformer defective.
Input pulses from integrating filter too weak.

Integrating tilter between oscillator and sync section.
Capacitor open, leaky, shorted. Resistor open, of wrong value. Makes hold control adjustment very critical.
Check parts from the oscillator grid circuit back to a point preceding the integrating and differentiating filters.

Vertical afc system.
Some receivers have automatic control for frequency of the vertical sweep oscillator. Check adjustments and circuit elements in the same way as for a horizontal afc system.

II. 4. e. SYNC NOT HOLDING, BOTH HORIZONTALLY AND VERTICALLY.

When the picture does not hold synchronization in either direction it is unlikely that the fault will be in either the horizontal or vertical sweep sections, since both these sections seldom would develope serious trouble at the same time. The fault may be in the sync section, which feeds both the horizontal and the vertical sweep sections, or it may be anywhere ahead of the sync section.

A common cause for sync failure is lack of response or gain at the lower video frequencies, where the signal sync pulses are carried.

Sync section.
Tube weak. Operating with wrong voltage, usually too low.
Coupling capacitors open, leaky.
Resistor open, of wrong value.

D-c restorer circuit, when sync pulses taken from this circuit.
Weak tube, wrong voltage, defective capacitor or resistor.

Contrast control.
Adjusted too high or too low.
The setting may be critical for holding the sync on weak signals, indicating lack of gain in the receiver.

Weak signal and lack of gain. Refer to Trouble Section on Snow.

Interference causing temporary loss of sync. Strong ghost (reflection) signals sometimes do this.

Automatic gain control holding amplifier biases too negative on weak signals.

Video amplifier.
Tube weak, plate or screen voltage low, grid bias too negative.

I-f amplifier.

Incorrect alignment, with video intermediate frequency too low on the response curve. This decreases gain at low video frequencies.

Tuner.

Incorrect alignment, or passband too narrow. Video carrier frequency too low on the response curve.

II. 4. f. TEAR OUT.

There is failure of horizontal synchronization at the top of the picture, and sometimes also at the bottom. This trouble is more common with receivers having no automatic control for horizontal sweep oscillator frequency than with receivers having such a control.

Contrast control set too high.

Excessive gain. Check the action of the agc system on various channels to note whether amplification is suitably reduced on strong signals.

Fig. 93-24. Tear out refers to lack of horizontal synchronization in part of the picture but not over the entire area.

Interference, usually spark type, is upsetting horizontal sync.

Microphonic tube anywhere between antenna input and sweep oscillators. Try tapping the tubes lightly to note whether this causes or affects the tear out. A tube causing a marked difference should be replaced.

Horizontal hold control.
Check capacitors, resistors, and control pot for faults.

Low-frequency response is poor.
I-f amplifier aligned to bring video intermediate frequency too low on the response curve.
Video amplifier tube may be weak, or operating with low voltage on plate or screen, or grid bias too negative.

Horizontal afc control, if used.
Check as for sync not holding horizontally, Trouble Section II. 4. c.

This method of trouble shooting by picture analysis will be continued, and concluded, in the following lesson. There we shall consider the various kinds of bars and lines which should not appear in the picture.

TELEVISION

LESSON NO. 94

SYSTEMATIC TROUBLE SHOOTING—PART THREE

Training Your Mind...
C. T. I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind
C.T.I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON NO. 94

SYSTEMATIC TROUBLE SHOOTING—PART THREE

Our final major division of faulty pictures, Class III, includes all those in which something that is not transmitted is added to the reproduced picture. Most of these additions have the form of bars or lines. The bars or lines may be bright or dark, and they may run horizontally, vertically, or at a slant across the picture area. The distinction between a bar and a line is not always definite. A thin bar might be called a heavy line, and a wide line might be called a narrow bar. Our grouping into bars and lines is merely for convenience of identification and to save time when looking for troubles which may exist. The subdivisions in this Class III are as follows:

III. A. Bars or lines.

1. Bright, horizontal.
2. Bright, vertical.
3. Bright, slanting.
4. Dark, vertical.
5. Alternate light and dark, horizontal.
6. Picture split horizontally, left and right.
7. Picture split vertically, top and bottom.
8. Picture folded horizontally.
9. Picture folded vertically.

III. B. Interference effects.

1. R-f signal effects.
2. Streaking, horizontal.
3. Flashes, horizontal.

III. C. Added picture or image lines.

1. Ghosts
2. Multiple images.

Now we shall proceed to list the troubles which most often produce the added bars, lines, or image effects.

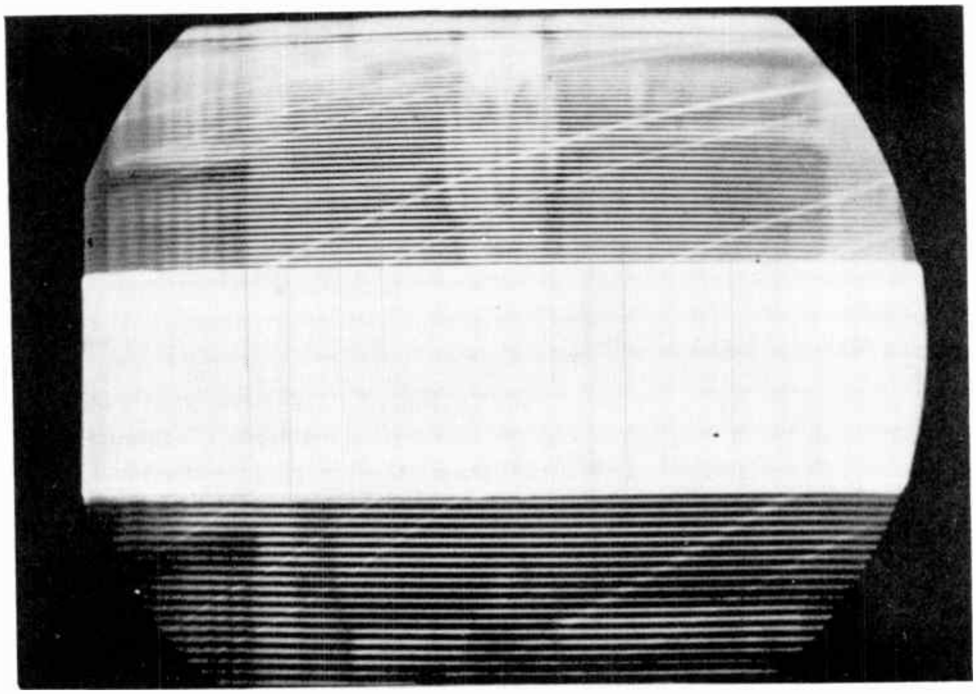


Fig. 94-1. A bright horizontal bar combined with vertical non-linearity.

III. A. 1. BRIGHT HORIZONTAL BAR.

Bars which show up horizontally indicate that something has gone wrong in the vertical control or sweep circuits, since the picture is not correctly distributed in a vertical direction.

Check for weak tubes, for incorrect tube voltages, for changed values of resistors or capacitors, for shorted or open resistors, for shorted, leaky, or open capacitors, and for defective connections in the following circuits.

a) Vertical hold control.
b) Vertical linearity control.
c) Vertical size control.
d) Vertical sweep oscillator and vertical sweep amplifier.

III. A. 2. BRIGHT VERTICAL LINES.

Lines, or bars, which show up vertically generally indicate trouble in a horizontal control or sweep circuit. Most often the trouble is one which affects every horizontal trace line at the

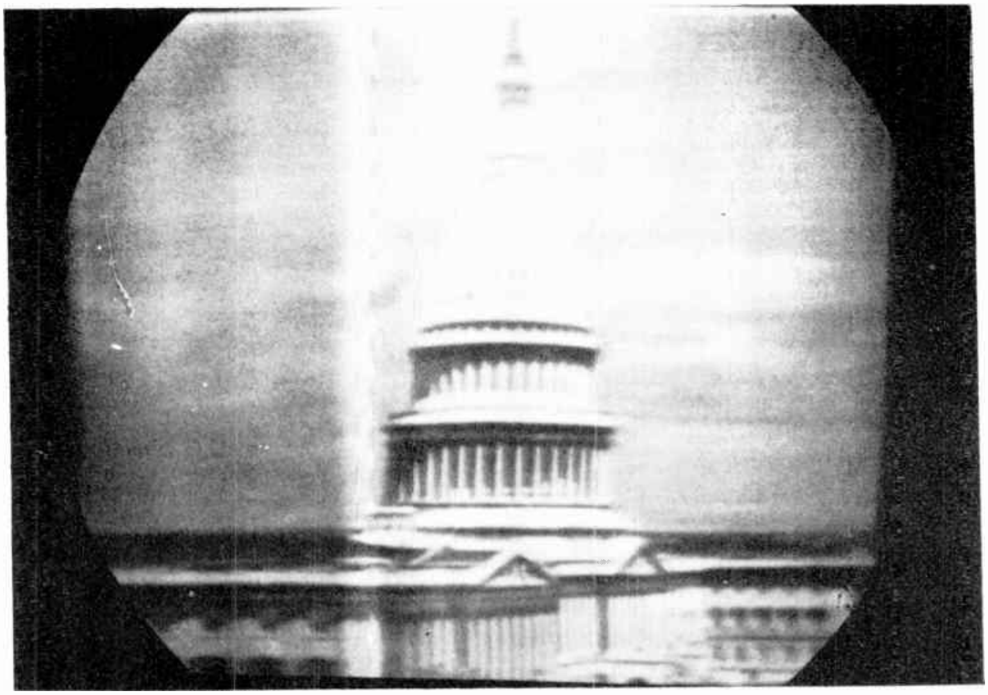


Fig. 94-2. A bright vertical line appears to the left of the dome.

Fig. 94-3. Bright vertical lines of varying intensity are at the left-hand side of the picture.

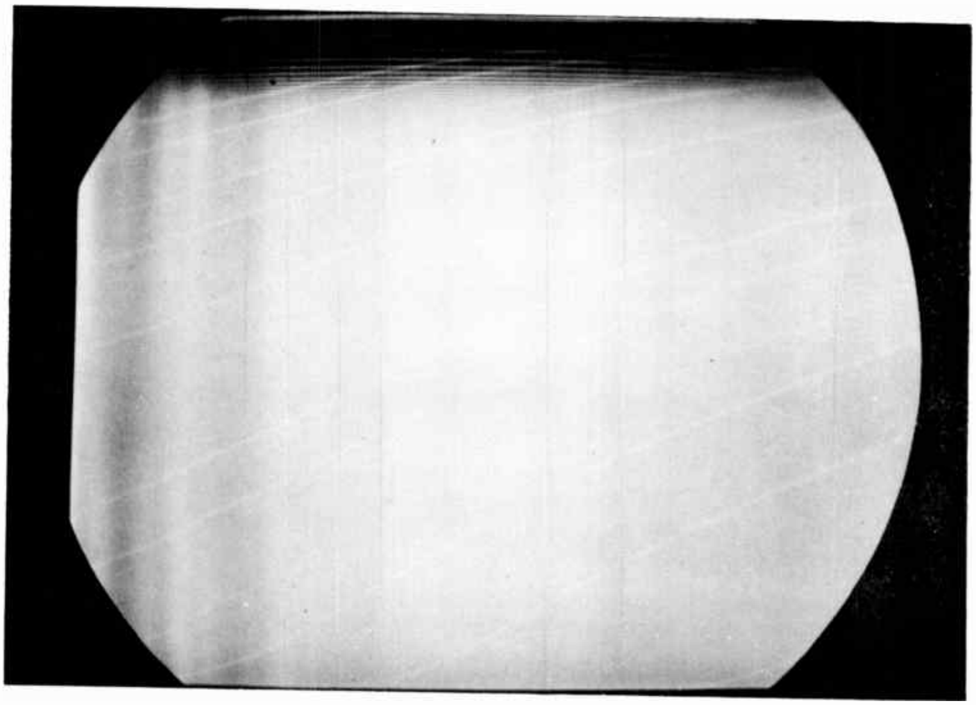


Fig. 94-4. Troubles producing bright vertical lines cause darker vertical bands to appear in the raster.

same point or points during travel of the electron beam from left to right. The horizontal sweep rate is not uniform from left to right, and the sawtooth voltage or current is not linear. When the same effect occurs on every horizontal trace line, the overall result appears as vertical bars or lines when successive horizontal trace lines are formed from top to bottom of the picture.

Faults of this general nature ordinarily will appear on the raster as well as on pictures. A raster may be examined by turning down the brightness control to emphasize differences in shading across the picture area.

When the trouble consists of high-frequency damped oscillations anywhere in the horizontal sweep and deflection systems the effect may be called ringing.

a) Drive control or horizontal peaking control incorrectly adjusted.
b) Horizontal sweep oscillator circuit.
   (1) Tube defective.
   (2) Plate voltage or grid bias incorrect. Check voltage dropping resistors, grid resistor, and grid capacitor.
   (3) Coupling capacitor from oscillator to sweep amplifier leaky.

c) Damper circuit.
  (1) Plate voltage too low, not enough damping action.
  (2) Resistor shunting the damper has too little resistance, too much series resistance, wrong tap connection being used.
  (3) Connection to wrong tap on secondary of output transformer.

d) Deflection yoke circuit.
  (1) Check the capacitor which is across a horizontal coil.
  (2) Yoke (replacement type) not suited for the circuit.

Fig. 94-5. Vertical retrace (slanting) lines on a picture. Note, incidentally, that the lines are uniformly spaced, indicating good interlace.

III. A. 3. BRIGHT SLANTING LINES.

These are vertical retrace lines, usually caused by setting the brightness control too high and/or the contrast control too low in relation to each other. Blanking voltage applied to the grid-cathode circuit of the picture tube does not cut off the beam during retrace.

a) Picture tube control grid bias insufficiently negative.
b) Coupling capacitor from video amplifier to picture tube leaky.
c) Vertical hold may be far enough out of adjustment to keep picture slightly high or slightly low.

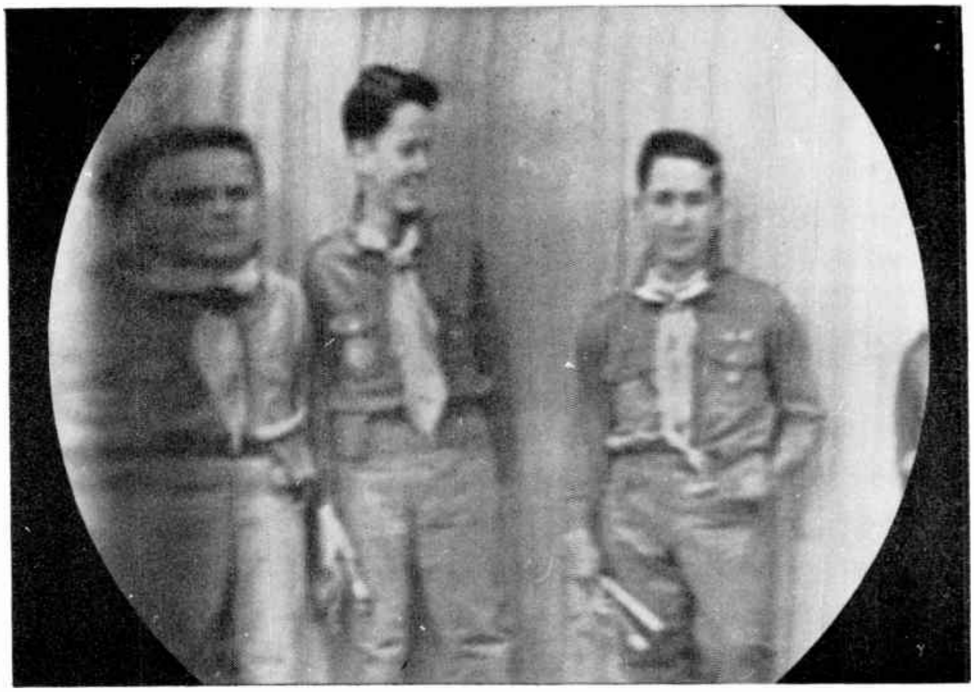


Fig. 94-6. There is a dark vertical band at the left. Note that this side of the picture is stretched, indicating too rapid travel of the electron beam where illumination is low.

III. A. 4. DARK VERTICAL BAND.

The same troubles which cause bright vertical lines may, in more severe cases, cause darkening of a vertical bar or band on the left side of the picture. Examination of the raster will show that the darkened band is followed by alternate light and dark vertical bars which become successively narrower.

Check the troubles listed for subdivision III. A. 2.

III. A. 5. DARK VERTICAL LINE.

This fault usually appears as a narrow line, dark on its left-hand edge and light on its right-hand edge. The width may vary more or less continually, and occasionally the line may break up into a series of flecks with a total width of about a quarter inch. Instead of a single line there may be two or more lines or parts of lines.

The cause is Barkhausen oscillation. This is a very high frequency parasitic oscillation occuring in the screen and plate circuit of the horizontal sweep amplifier. During such oscilla-

Fig. 94-7. A dark, or dark and light, vertical line at the left indicates Barkhausen oscillation.

tion the electrons flowing to the plate change their velocity or momentarily reverse their direction as the plate becomes instantaneously negative with reference to the screen, causing a sudden variation of plate current. Frequency of oscillation depends on spacing of elements within the amplifier tube, on inductances of connected leads, and on stray capacitances in the circuits.

The fault most often appears with weak received signals and high setting of the contrast or gain control, although sometimes it appears with reduction of gain or contrast control setting. Reception in high-band channels is affected more than in the low band. The position of the line usually varies with the channel received.

a) Drive or peaking control misadjusted, for too much drive (not enough capacitance) or for too much negative peaking.
b) Horizontal linearity control, on output transformer, misadjusted.
c) Horizontal size control, on output transformer, adjusted for an excessively wide picture.
d) Horizontal sweep amplifier.
   (1) Defective tube.
   (2) Plate and/or screen voltage incorrect, usually too low.
   (3) Screen bypass capacitor too small or open.

III. A. 6. ALTERNATE LIGHT AND DARK BANDS, HORIZONTAL.

a) Sound bars. More than two dark and two light bands between top and bottom of picture.

Sound bars result from a sound or audio frequency getting into the circuits for the video detector, video amplifier, or picture tube signal input.

If the troublesome voltage is that of voice or music modulation from any source, the bars will move or weave vertically with changes of modulation frequency, and the bars will disappear when the voice or music stops. If the frequency is constant, and an exact multiple of 60 cycles, the bars will remain stationary. If the constant frequency is not a multiple of 60, the bars will move up or down on the screen at a steady rate.

The frequency of the troublesome voltage is equal to 60 times the number of pairs of dark and light bars.

(1) Volume control too high.
(2) Fine tuning control misadjusted.
(3) Strong interference from amplitude-modulated transmission of short-wave or broadcast stations. Amateur, police, etc.

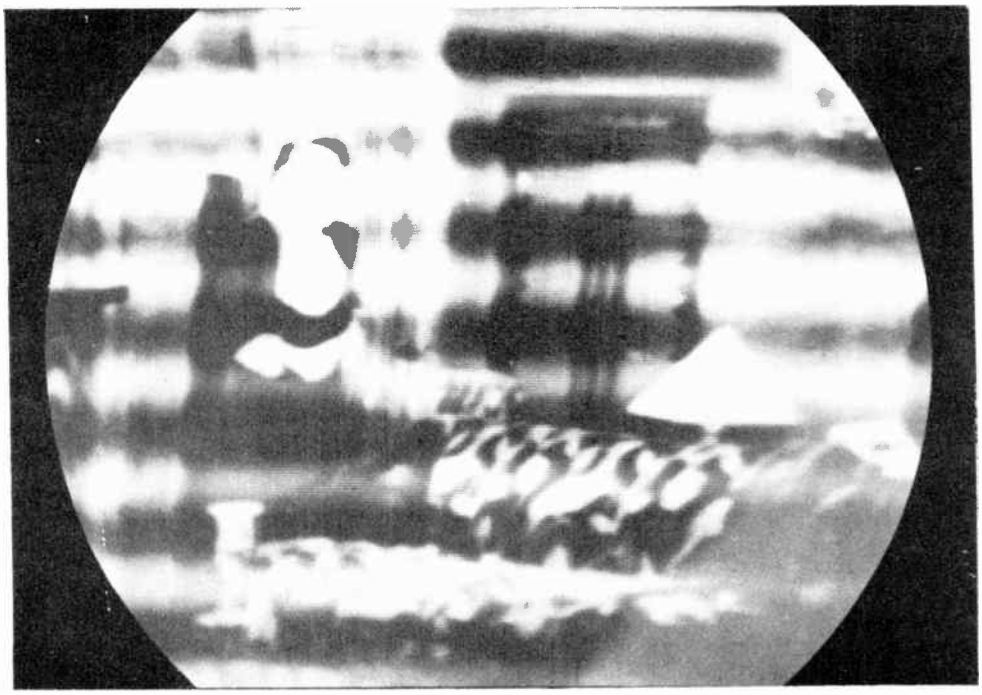


Fig. 94-8. Sound bars. There may be any total number of alternate light and dark bars.

(4) Microphonic tube.
  (a) Vibrates with volume turned up, usually stops when volume turned down. Try tapping tubes with volume low.
  (b) Usually the r-f oscillator or the mixer, but may be in i-f amplifier, sync section, or sweep oscillator positions.

(5) R-f oscillator. Aligned to bring sound intermediate frequency too high on the response, or outside the trap frequency when sound traps are used. With dual sound systems this would prevent reproduction of sound from the speaker.

(6) I-f amplifier.
  (a) Aligned to bring sound intermediate too high on the response. The sound bars usually persist with the volume turned down.
  (b) Oscillation, usually due to alignment at incorrect frequencies in stagger tuned amplifier.

(7) Traps misadjusted.
  (a) Bars usually persist even with volume turned down.
  (b) Check the following traps, if used.
    i) Accompanying sound.
    ii) Adjacent sound.
    iii) Intercarrier beat.
    iv) Trap that shapes response or prevents excessive peaks.

(8) Sound section.
  (a) Sound takeoff incorrectly aligned.
  (b) Demodulator secondary incorrectly aligned.
  (c) Tube defective. Sound i-f amplifier, demodulator, or audio output.

(9) B-supply system with sound section cathodes feeding plates in other sections. Filtering ineffective between sections. Check the decoupling resistor and capacitor(s).

b) Hum bars. Only one or two dark bands and light bands between top and bottom of picture.

These voltages are due to power line frequency (60 cycles) or at the output frequency of a full-wave rectifier (120 cycles) entering circuits of the video detector, video amplifier, or picture tube input.

Hum bars ordinarily remain stationary, since their voltages are at the 60-cycle vertical deflection frequency or at twice this frequency. Strong hum voltages which are at or close to 60 cycles or a multiple of 60 cycles will synchronize the vertical deflection, and the bars will remain stationary.

Hum bars usually are accompanied by greater or less vertical waving of the picture, as may be seen in the illustration of this fault.

(1) Excessive ripple voltage from low-voltage power supply.
  (a). Filter capacitor open, of too small capacitance.
  (b) Filter resistor or choke shorted, of too small value.

(2) Power transformer or power filter choke too close to picture tube neck or socket.

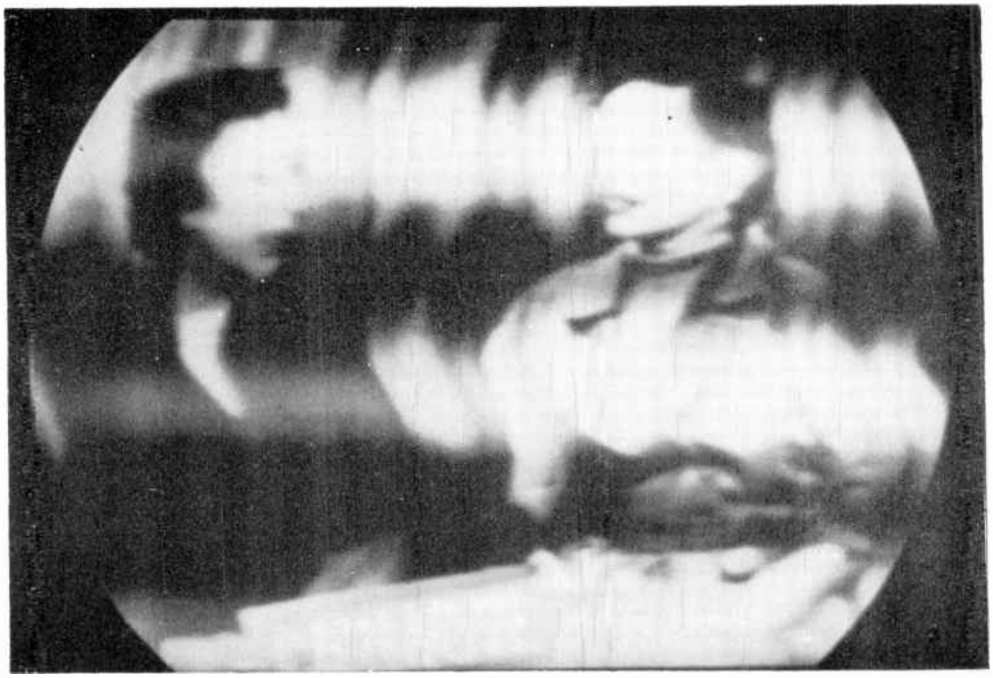


Fig. 94-9. Hum bars, caused by voltages at 60 or 120 cycles per second. Note that the picture is wavy from top to bottom.

(3) Decoupling capacitor open, or decoupling resistor shorted. Check video amplifier, sync section, horizontal sweep oscillator and amplifier circuits.
(4) Heater-cathode leakage. Check tubes in video amplifier and in i-f amplifier sections, especially the latter because its output is amplified before reaching the picture tube.

III. A. 7. SPLIT PICTURE, HORIZONTALLY.

The picture is divided into two parts by a horizontal dark bar, which is a vertical blanking bar occuring between successive fields and frames. The bottom of the reproduced picture appears above the lower part. The blanking bar may remain stationary, but more often it moves up or down rather slowly.

This fault is due to incorrect vertical sweep frequency, to a frequency that does not exactly synchronize with vertical pulses of the received signal. If there is but one complete picture, split into two parts, the frequency is off to only a slight degree. If there are two complete pictures, separated by horizontal bars, the vertical sweep is occuring at 30 cycles instead of at 60 cycles per second. If there are three complete pictures the vertical sweep is at 20 cycles, or at 1/3 of the correct rate.

Fig. 94-10. Split picture, with horizontal bar. Note that the legs of the two boxers appear at the top, while their heads appear below.

Vertical sweep oscillator circuit.

a) Capacitor of wrong value or leaky, or resistor of wrong value, in vertical hold control circuit.
b) Capacitor or resistor of wrong value or defective in the vertical timing circuit, which includes the sawtooth capacitor and all resistors and other capacitors associated with it.
c) If part of the sync section serves only the vertical oscillator, check tubes, capacitors, and resistors in this part.

III. A. 8. SPLIT PICTURE, BY VERTICAL BAR.

The appearance of a dark vertical bar between parts of the picture indicates trouble in horizontal synchronization. The dark area is a horizontal blanking bar, or it represents blanking between successive horizontal trace lines all the way from top to bottom of the picture.

a) Horizontal afc system, between sync section and oscillator.
   (1) Service controls misadjusted, as lock control, phasing control, etc.
   (2) Capacitor open, leaky, wrong value.
   (3) Resistor of wrong value.

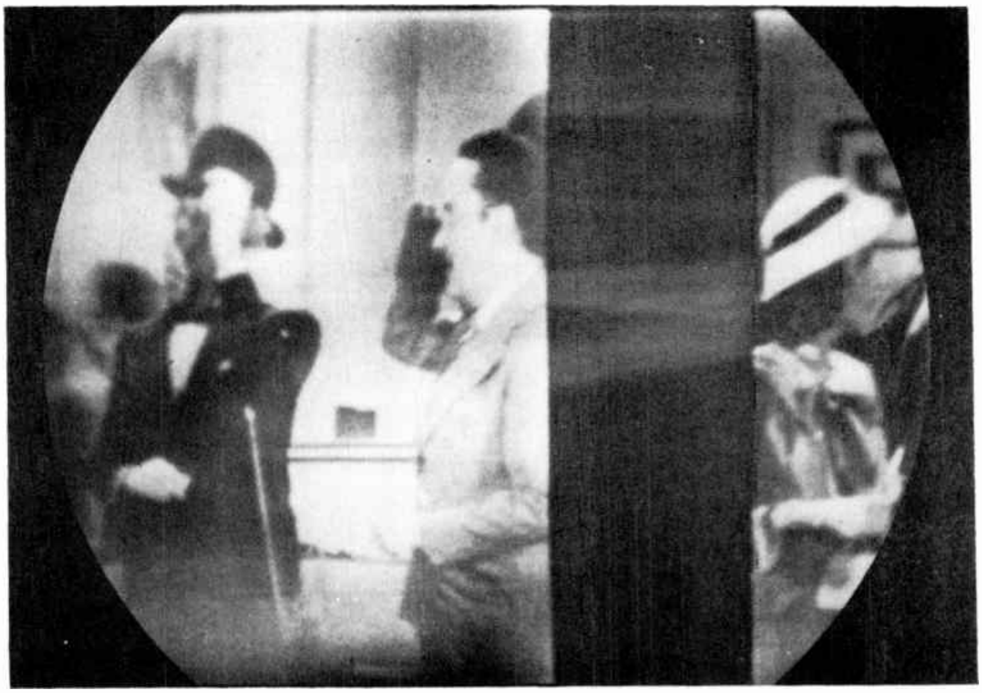


Fig. 94-11. Split picture, with vertical bar. The man at the left-hand side of the picture is tipping his hat to the woman who appears at the right.

(4) Control tube defective, for this position.
(5) Leads reversed to a phase detector.

b) Horizontal hold control circuit.
Defective capacitor or resistor.

c) Sync section.
(1) Coupling capacitor leaky, too little capacitance, open.
(2) Wrong connections in portion serving horizontal control.

d) Horizontal output transformer.
Wrong connections on secondary; for yoke, damper, controls, or shunting capacitor.

e) Horizontal deflecting coils, in yoke.
Shunting capacitor defective.

Note.–Strong interference from a channel other than the one tuned may cause this fault.

III. A. 9. FOLDED PICTURE, HORIZONTALLY.

Part of the picture may appear reversed from left to right, and as though folded under or over the other part, but with the images in both parts remaining visible. One part is greatly ex-

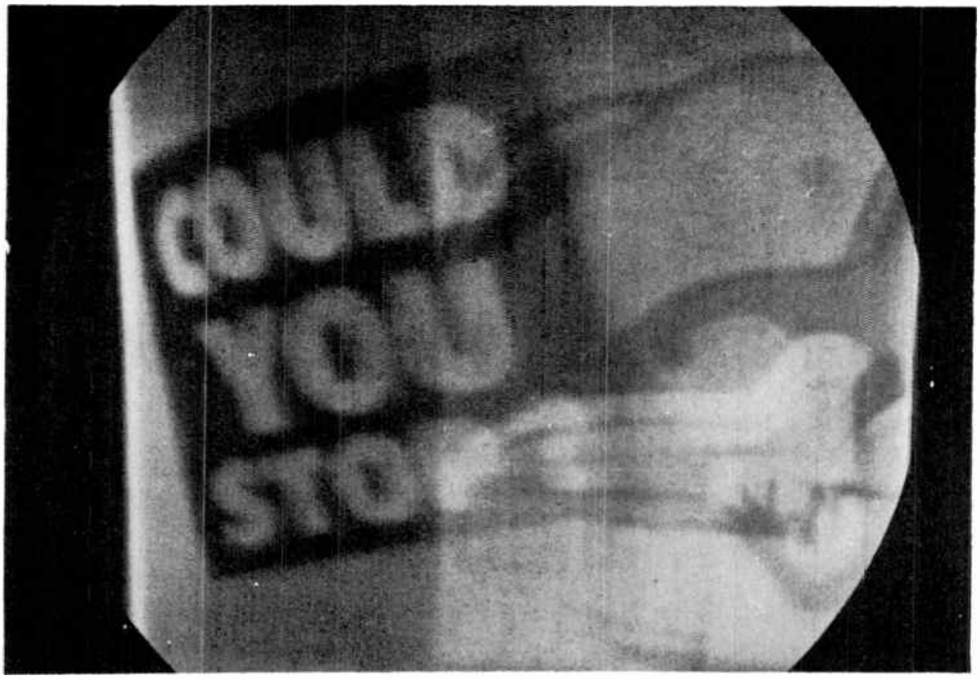


Fig. 94-12. Folded picture, horizontally. The part of the image which is stretched horizontally, is at the left in this particular case. It might be at the right.

Fig. 94-13. Horizontal folding combined with lack of horizontal size. The stretched and folded part of the picture is here at the right.

tended in a horizontal direction, so that objects in it appear much too wide. The two parts of the picture may nearly fill the width of the mask, or both together may take up much less than the distance across the mask.

The fault is due to an incorrect timing relation of horizontal trace lines (the picture elements) and horizontal blanking. Part of each trace line is formed during the period when the beam should be blanked. Instead, the trace is reversed in direction while picture elements still are being formed, with the result that these elements are reversed.

a) Drive or peaking control misadjusted.
b) Horizontal afc control adjustments are incorrect.
c) Damper circuit.
   (1) Tube weak or otherwise defective.
   (2) Capacitor from cathode line to B-plus is open.
   (3) Plate voltage too low. Shunt resistor adjusted wrong, of wrong value, connection made to wrong tap. Connection to wrong tap on secondary of horizontal output transformer.
d) Horizontal sweep amplifier grid resistor of too little resistance, or otherwise defective.
e) Horizontal linearity control, on transformer secondary.
   (1) Open circuited.
   (2) Inductor defective.

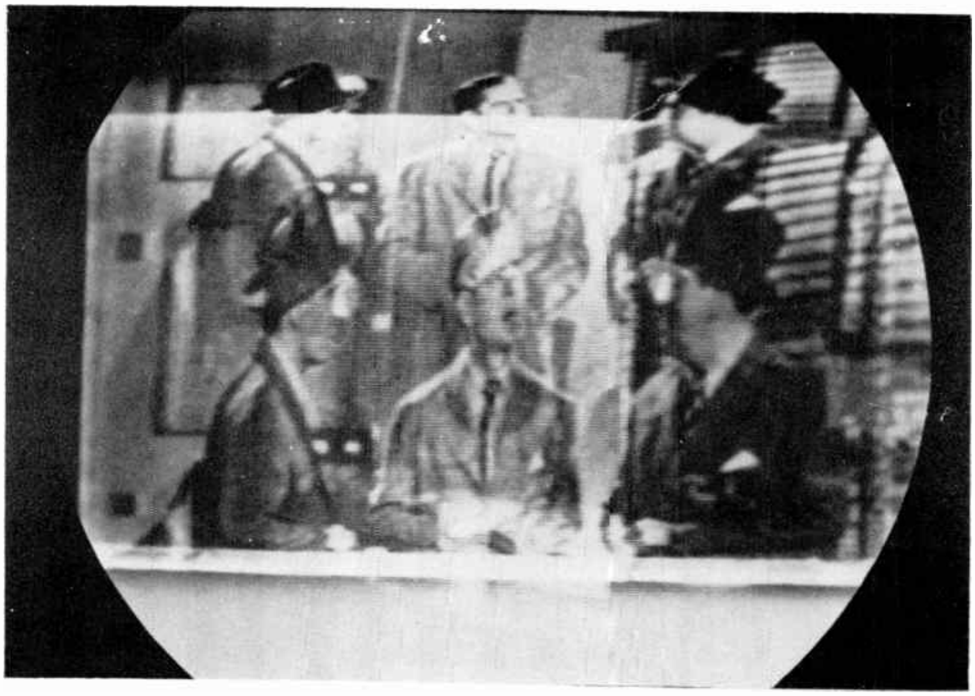


Fig. 94-14. Folded picture, vertically. In the central portion of the picture there is excessive height while there is compression of the duplicated images higher up.

III. A. 10. FOLDED PICTURE, VERTICALLY.

Part of the picture is compressed in a vertical direction, with the same part or the remainder appearing as though folded over or under. Both parts are visible at the same time. The entire picture is of less height than the vertical size of the mask opening. The picture, as folded, may remain stationary or nearly so with the vertical hold control in one position, and lose vertical sync with the control in other positions.

a) Vertical hold control circuit. Check capacitors and resistors for wrong values, opens, leaky capacitors, partially shorted resistors, etc.
b) Vertical sweep oscillator.
   (1) Grid capacitor leaky.
   (2) Grid resistor of too little resistance, shorted, wrong value.
c) Vertical sweep amplifier circuit.
   (1) Grid resistor of too small value, otherwise defective.
   (2) Coupling capacitor from oscillator to amplifier shorted, leaky.

III. B. 1. SLOPING OR VERTICAL LINES. R-f Interference.

High-frequency interference causes fine lines to slope to the right or left, or they may be ver-

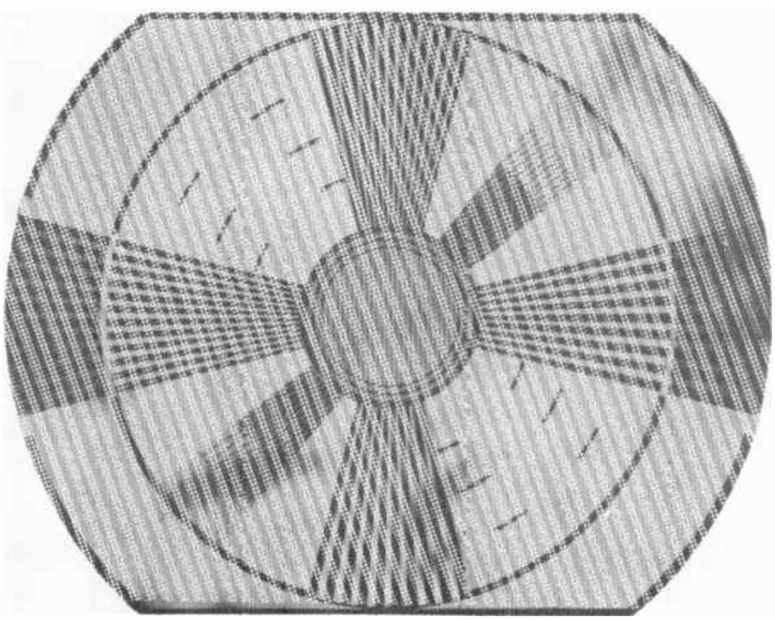


Fig. 94-15. R-f interference, with or without modulation, may cause narrow lines sloping either direction, or shifting back and forth.

tical. The lines shift and weave while the interfering signal is modulated, and remain nearly stationary when there is no modulation. The more common sources of such interference include:

a) Other television channels, f-m broadcasts, and radiation from oscillators or television receivers. One or more of the following procedures may help.
   (1) Careful adjustment of the fine tuning control.
   (2) Change of antenna orientation or position.
   (3) Change of transmission line position.
   (4) Careful adjustment of trap(s) for adjacent sound, or of traps on the antenna and r-f amplifier circuits.

b) Intercarrier beat, at 4.5 mc, causes similar effects.
   (1) Adjustment of accompanying sound trap(s).
   (2) Adjustment of trap for intercarrier beat frequency.
   (3) I-f amplifier may be aligned to bring sound intermediate frequency too high on the response, or out of the frequency range of accompanying sound traps.

c) Feedback of video frequencies, in the higher range.
   (1) Check the dressing of all coupling capacitors from the video detector through to the picture tube grid-cathode input. Capacitors may be brought closer to chassis metal.
   (2) Check the dressing of picture tube signal input leads, either grid or cathode, which may run too close to other circuit elements and wires, or too close to chassis metal.

d) High-voltage power supply, r-f type.
   (1) Shield enclosure may be grounded to chassis when it should be isolated or connected only through a capacitor and resistor.
   (2) Shield may not be grounded in designs which require its grounding.
   (3) Defective filter chokes or filter resistors on line from B-plus to oscillator circuits.
   (4) Filter capacitor (high-voltage) open.
   (5) Oscillator plate and grid leads not dressed close to chassis.
   (6) Loose or dirty connections anywhere in this power supply.

   Note.—This kind of interference sometimes may be identified by counting or estimating the number of lines, which would be equal to the oscillator frequency divided by 15,750 cycles. The number of interference lines and their degree of slope will vary if the power supply oscillator frequency can be altered by the control for output voltage.

III. B. 2. CROSSHATCH LINES. R-f Interference.

This type of interference effect usually is due to amplitude modulated signals from any class of transmitter. If interference is in the very-high frequency band it may appear on some channels and not on others. If in the television intermediate-frequency range the trouble will appear on all channels.

Remedies are, in general, the same as for sloping or vertical lines, subdivision III. B. 1.

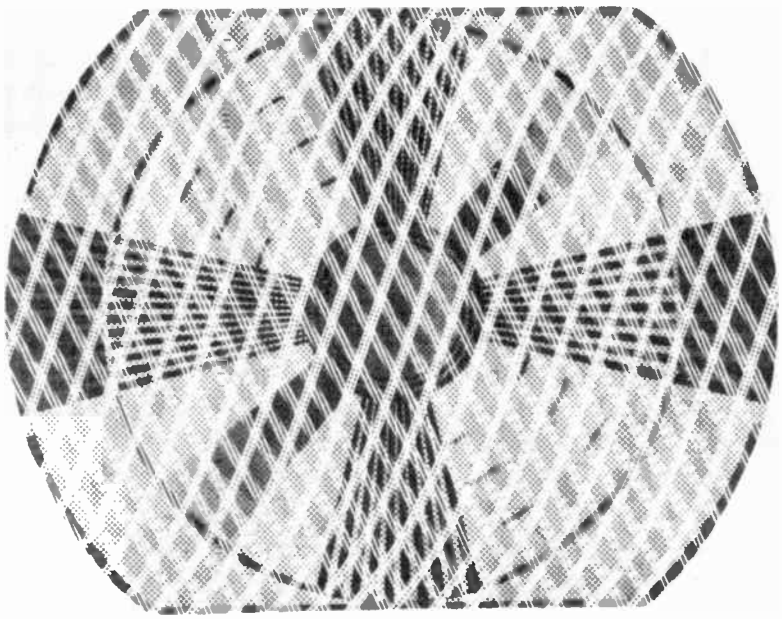


Fig. 94-16. R-f interference, with modulation, may cause a cross-hatched pattern.

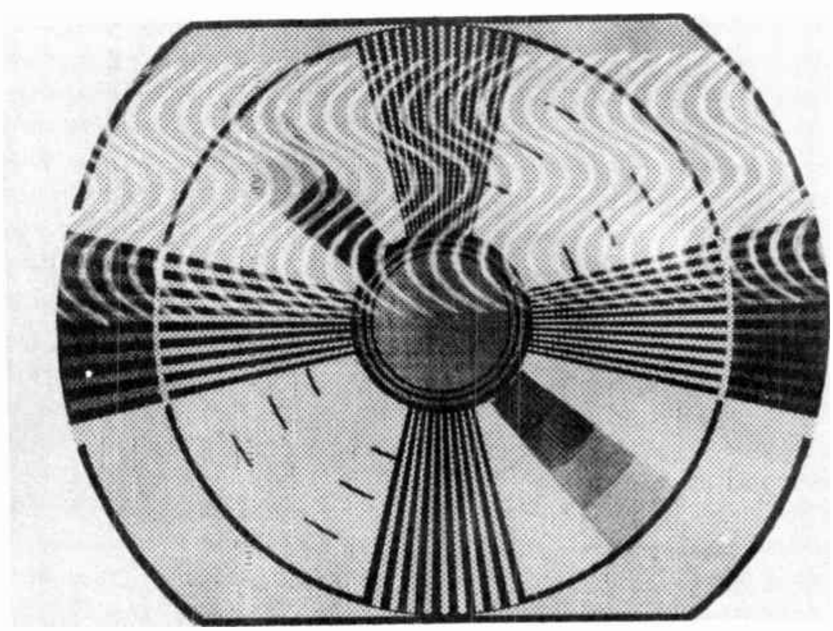


Fig. 94-17. R-f interference from medical, commercial, or industrial apparatus operating at moderately high frequencies may cause a herringbone pattern anywhere on the picture. Other interferences may cause a similar pattern.

III. B. 3. HERRINGBONE LINES. R-f Interference.

a) Possible sources of interference causing this effect are as follows.
   (1) High-frequency medical apparatus of any kind, such as X-ray machines, diathermy apparatus, violet-ray or infra-red apparatus.
   (2) F-m broadcast signals.
   (3) A-m broadcast or short-wave signals, while modulated.
   (4) Oscillator radiation from other television receivers.
b) Remedies are the same as for sloping or vertical interference lines, subdivision III. B. 1.
c) Sources within the offending receiver may be,
   (1) Oscillation in the i-f amplifier section, due usually to alignment of two or more stages at the same or nearly the same frequency, or to incorrect dressing of parts and leads, or to shields which are loose or which have been removed and not replaced.
   (2) Feedback in the sound i-f amplifier, usually due to defective decoupling capacitors.

III. B. 4. STREAKING OR SPECKLING, HORIZONTAL.

This fault, usually due to some kind of spark interference, consists of very brief darkened

Fig. 94-18. Horizontal streaking or speckling which was due to an electric motor on the same power line as the receiver.

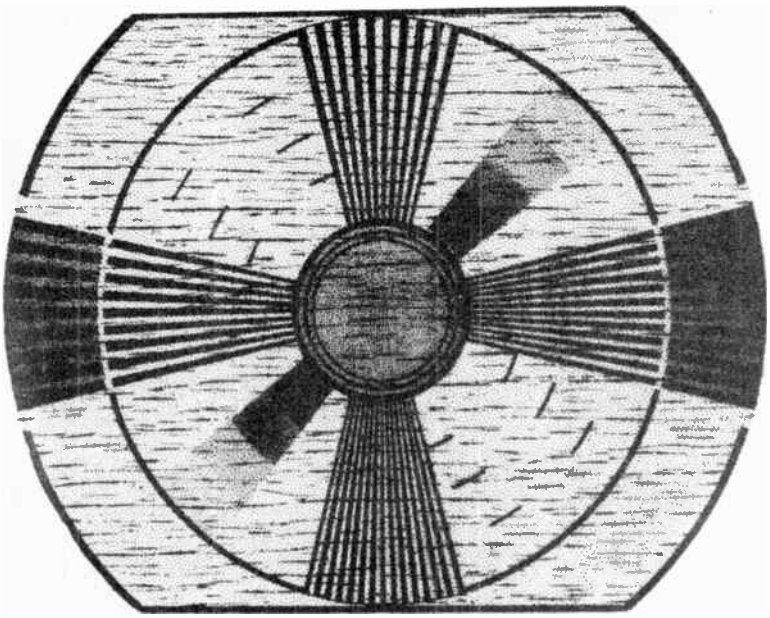


Fig. 94-19. Horizontal streaking appears as irregular dark lines occuring intermittently across the picture or pattern.

intervals along the horizontal lines. The duration of the streaks is so brief that the camera hardly can catch their effect. To the eye, the appearance is more as illustrated by the drawing of a test pattern in Fig. 94-19.

Spark interferences, their causes and possible remedies, are discussed in detail by the lessons dealing with interference in general. Other causes for streaking are as follows.

a) High-voltage power supply corona, arcing, or flashover Causes and remedies are discussed in lessons on high-voltage power supplies. These troubles may be identified with room lights turned off while the receiver is in operation Corona appears as a blue glow on high-voltage parts and conductors.
   (1) Loose or corroded connections in any high-voltage lines.
   (2) Dirt or moisture on the high-voltage rectifier socket.
   (3) Splices in high-voltage leads. They should be continuous wires.
   (4) High-voltage leads or parts too close together. Check carefully at the rectifier socket. Rectifier plate and filament leads should be at least one inch from other metallic objects.
   (5) Poor connections between rectifier base pins and socket contacts.
   (6) Poorly soldered joints. There must be no sharp points or loose wire strands.
   (7) A corona ring may be mounted on or connected otherwise to one of the rectifier filament pins, at the socket. See that this ring is secure and evenly spaced around socket terminals.

Note.—Additional insulation on high-voltage conductors and parts may be provided by applying high-voltage cements sold for this purpose, or by polystyrene or polyethylene cement, wherever there are electrical discharges.

b) Oscillation in the i-f amplifier section, due to incorrect alignment, wrong dressing of leads, loose or omitted shields, etc.
c) Microphonic tube anywhere between the antenna and picture tube input. Try tapping each tube while watching for any change in the streaking effect. If a tube causes any increase, check its socket connections, a cushion mounting if used, or replace the tube.
d) Picture tube second grid intermittently open circuited. This may cause a coarse variety of streaking.

III. B. 5. FLASHES ON PICTURE.

a) Very strong spark interference occuring intermittently, as when electrical devices are switched on and off.
b) Coupling capacitors in high-voltage circuits have intermittent or momentary shorts, or there may be flashing over their insulation because of dirt or moisture.

Fig. 94-20. Flashes at irregular intervals across the picture

c) Tube in which elements momentarily short circuit. Try lightly tapping the tubes between antenna and picture tube input to note whether any of them cause flashing.
d) Flashover or arcing in high-voltage power supply or picture tube anode circuit. Check as explained for Streaking or Speckling in subdivision III. B. 4.

In our next classification we consider troubles which cause added picture lines. The normal image is present, but, in addition, there appear to be one or more additional sets of image lines. One group of troubles includes signal reflections which reach the antenna at an instant of time slightly different from the regular signal, causing what usually is called a ghost image. Another group of troubles occurs within the receiver itself, sometimes causing only one additional set of image lines, but usually causing two or more sets of graduated intensity. These are called multiple images.

III. C. 1. GHOST IMAGE.

The causes for ghost images, and the usual remedies, are discussed in lessons dealing with interference in general. Ordinarily there is only a single signal reflection, which causes a single added image displaced from the normal image by any distance horizontally on the screen. The ghost image usually is toward the right from the normal image, indicating that the

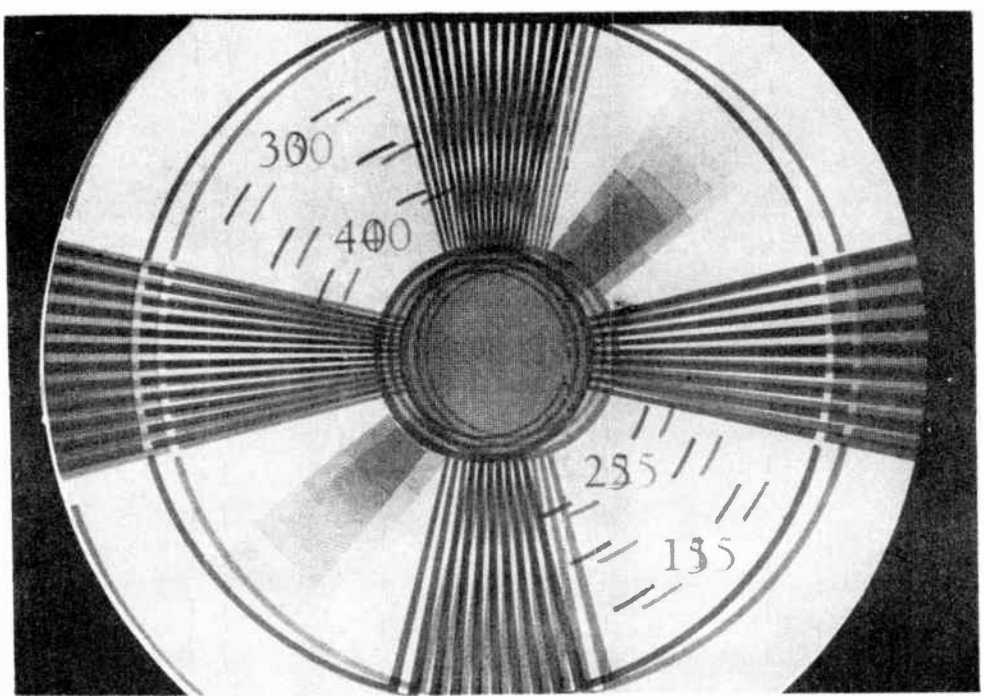


Fig. 94-21. A ghost image is a complete duplicate of the regular image, although usually of lessened intensity or strength.

reflection has traveled farther than the direct signal, and has taken longer to arrive at the antenna. In rare cases the ghost may appear at the left from the normal signal, indicating a relatively weak signal pickup directly in the receiver circuits slightly earlier than through an antenna and a long transmission line. Remedies most often are applied at the antenna, as follows.

a) Change the orientation.
b) Change the location.
c) Raise the antenna higher or, in rare cases, drop it lower.
d) Use a reflector element when signal reflections are from the rear.
e) Use a director, in addition to a reflector element, when signal reflections arrive at the front of the antenna system.
f) Use a more directional type of antenna.

III. C. 2. MULTIPLE IMAGES.

Multiple images almost always indicate the presence of damped oscillation in an amplifier circuit. There are oscillatory waves which quickly die away. Portions of the picture which are produced by video frequencies at and near the oscillation frequency then are caused to repeat by successive peaks of oscillation.

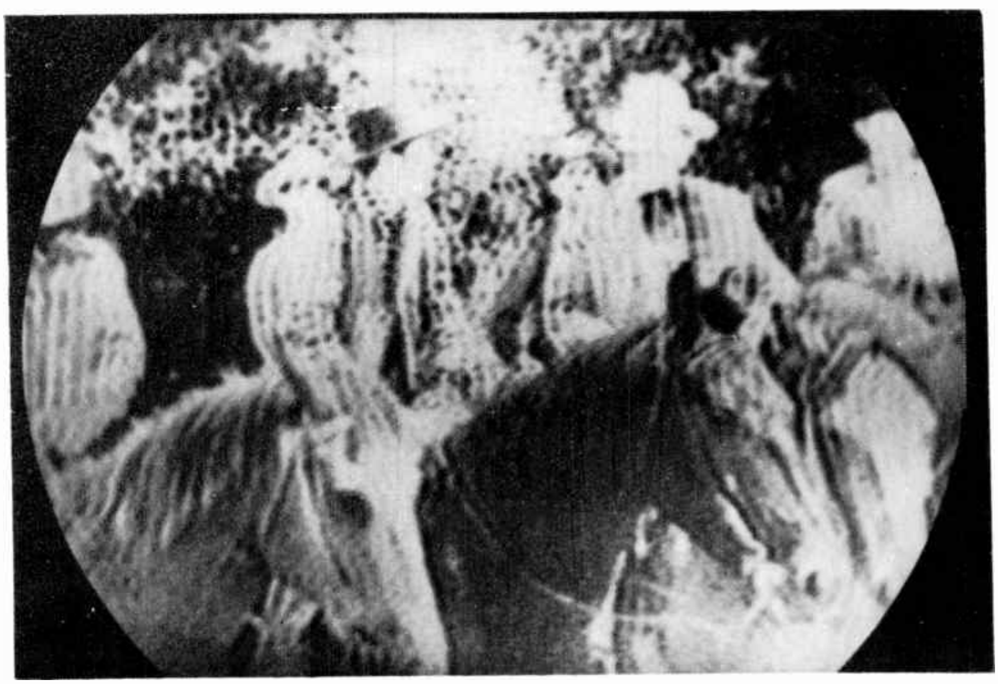


Fig. 94-22. Multiple image lines follow closely at the right from the normal image lines. The effects may be disagreeable, even when the fault is much less pronounced than in the extreme case illustrated.

Multiple images are closely and equally spaced from each other, and to the right from normal image lines, at distances dependent on the frequency of damped oscillation. Successive images decrease in strength, as the damped waves decrease in amplitude.

The images occur most commonly where there are sudden variations of video frequency, as at abrupt changes from bright to dark and vice versa. These sudden changes tend to excite the offending circuit into oscillation. Ghost images, as distinguished from multiple images, follow all shadings and variations of shadings, whether abrupt or gradual.

a) Contrast or gain control too high.
b) Drive or peaking control misadjusted.
c) Video amplifier circuits.
  (1) Peakers of incorrect value, or defective, causing excessive peaking at high or medium video frequencies.
  (2) Coupling capacitor of too small capacitance, open.
  (3) Decoupling capacitor open.
  (4) Check the dressing of leads, capacitors, and resistors.
d) Video detector circuit. Load resistor of wrong value, shorted.
e) I-f amplifier circuits.
  (1) Incorrect alignment, with excessive gain toward the sound side of the response, where the higher video frequencies are carried.
  (2) Coupling capacitor of wrong value, open.
  (3) Decoupling capacitor open.
  (4) Check the dressing of leads and capacitors.
  (5) Shields loose, removed and not replaced.
f) Transmission line mismatch, standing waves. This would cause very close spacing of images, and usually would result only in blurring rather than in separated lines.

PATTERN GENERATORS

When neither television pictures nor test patterns are available it is possible to check the performance of many receiver circuits and to make certain service adjustments with the help of a pattern generator. This is an instrument for producing on the picture tube screen a pattern consisting of a series of bars or lines. Fig. 94-23 shows horizontal bars thus produced on the picture tube of a receiver which is in satisfactory adjustment. Fig. 94-24 shows how the bars look when there is severe vertical non-linearity, which here is causing excessive separation and widening of bars toward the top, with crowding and narrowing at the bottom.

Most commercial pattern generators will produce either horizontal bars or vertical bars, while some will produce both kinds of bars at the same time. Fig. 94-25 shows how vertical bars may appear when there is severe horizontal non-linearity, causing crowding and narrowing at both left and right, with stretching and widening just at the right of center.

Pattern generators find their greatest usefulness in making adjustments of linearity. They are used also during adjustment of size controls, for adjusting drive or peaking controls as these affect linearity and width, and for correcting skew. It is possible also to check the behavior of hold controls, of contrast

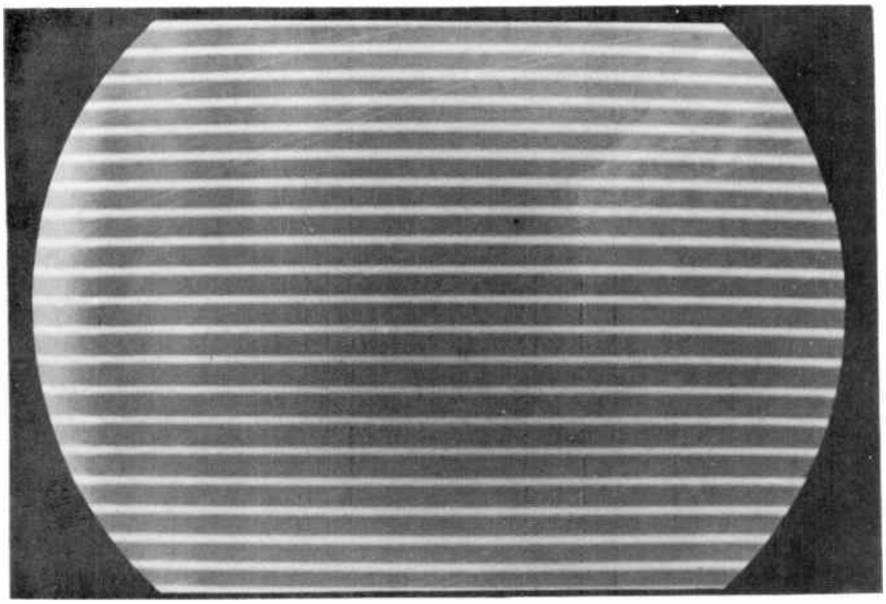


Fig. 94-23. A pattern of horizontal bars showing satisfactory receiver operation.

Fig. 94-24. Vertical non-linearity as shown by a bar pattern.

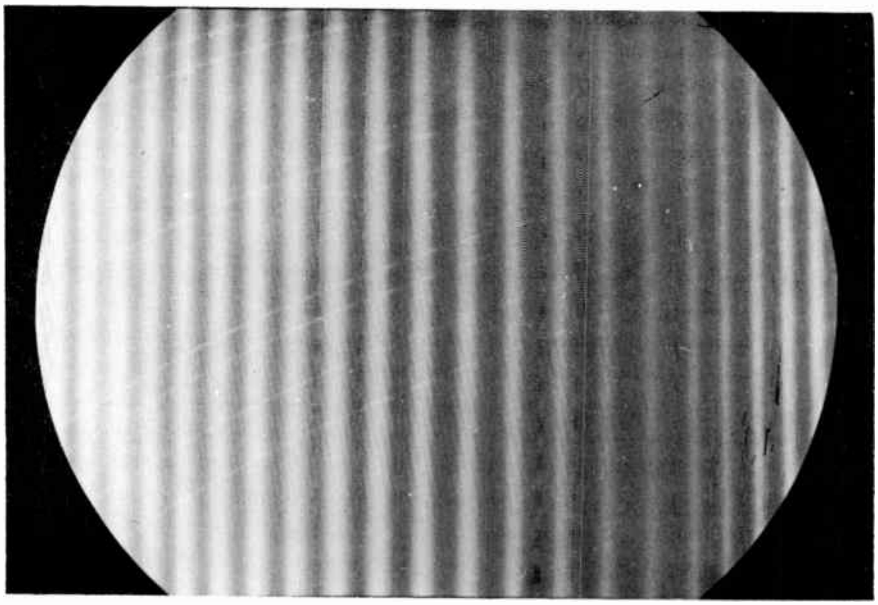


Fig. 94-25. Horizontal non-linearity as shown by a bar pattern.

or gain controls, and of brightness controls. The pattern generator feeds to the antenna terminals of a receiver a carrier-frequency signal modulated with the bar pattern. This signal passes through all receiver sections and to the picture tube. Consequently, the appearance of satisfactory bar patterns indicates that there is selection, transfer, and amplification of signals in the tuner, in the i-f amplifier, the video detector, video amplifier, sync section, and sweep sections.

A pattern generator employs one oscillator for producing the carrier frequency and a separate oscillator for producing either horizontal or vertical bars. If bars are to be formed in both directions at the same time there will be three oscillators; one for the carrier frequency, a second for horizontal bars, and a third for vertical bars.

The output of the carrier-frequency oscillator, which goes to the antenna terminals of the receiver, is modulated by frequencies from the other oscillators. The oscillator for horizontal lines must produce a frequency which is some multiple of the vertical sync frequency, 60 cycles per second. For each 60 cycles in the oscillator output there will be one horizontal dark bar and a following light bar during each 1/60 second of time. A frequency of 600 cycles per second thus should produce 10 dark bars and 10 intervening light bars. Since vertical blanking takes up about 7 per cent of each field or frame period, no more than 93 per cent of the theoretical number of pairs of bars can actually appear on the screen. In our present example only 9 pairs of bars would be seen with the oscillator operating at 600 cycles per second.

The oscillator producing vertical bars must operate at a multiple of 15,750 cycles per second, the horizontal sweep frequency. For 14 pairs of dark and light bars, as an example, the oscillator frequency would be 14 times 15,750 cycles, or 220,500 cycles (220.5 kc) per second. Horizontal blanking takes up about 16 per cent of each horizontal line period, so no more than 84 per cent of this number of pairs of bars can appear. In the present example there would be 12 pairs of bars on the screen.

The outputs of the oscillator or oscillators producing bar-forming frequencies may be applied to the output of the carrier-frequency oscillator in any of the ways commonly used for modulating service type r-f signal generators with audio frequencies. Any marker or sweep generator capable of producing carrier frequencies, and with provision for external modulation, may have any unmodulated r-f signal generator connected to its external modulation jack for production of vertical bars. Any audio-frequency generator may be similarly connected to the external modulation jack for production of horizontal bars.

Another method of modulation employs a mixer, which may be a diode tube or a crystal diode. The principle is illustrated at 1 in Fig. 94-26, where there is one additional oscillator for producing either horizontal or vertical bars, depending on the operating frequency of this oscillator. At 2 are shown, in principle, the connections for using two additional oscillators for forming both horizontal and vertical bars at the same time. In actual practice there must be provision for making the two sets of bars of equal strength, and for preventing undesired interaction between the several oscillators.

To use a pattern generator, first remove the regular antenna or transmission line connections from the receiver, attach the generator output leads to the antenna terminals, turn on both the receiver and generator, and allow them to warm up thoroughly. Place the receiver selector switch at any channel for which

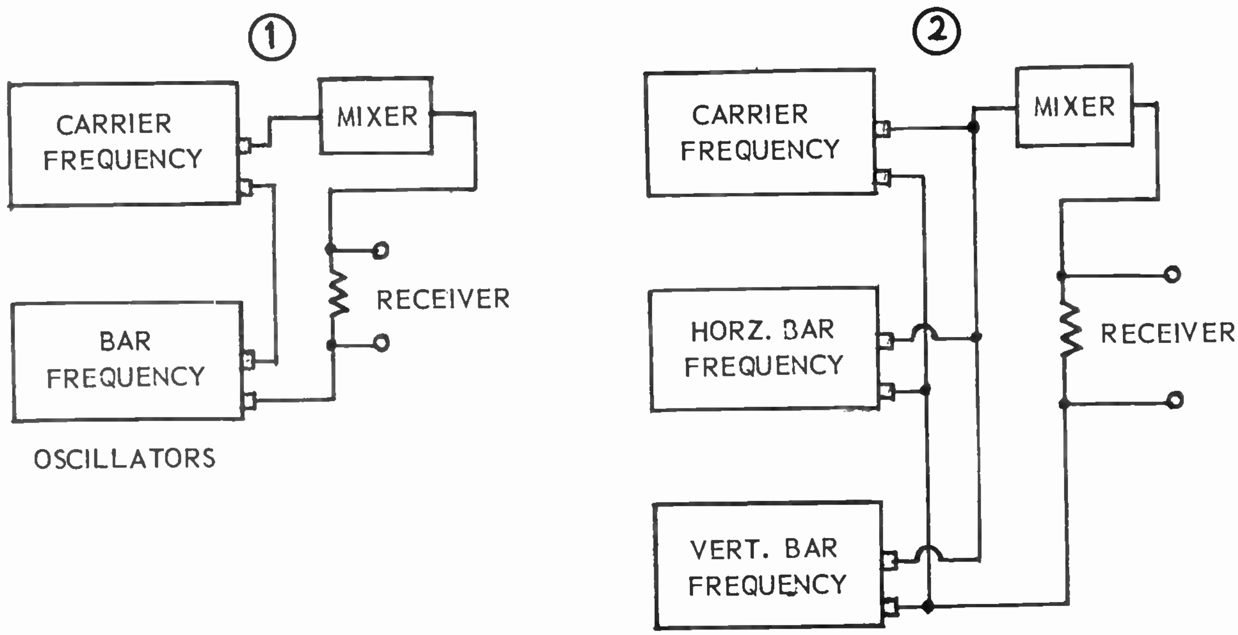


Fig. 94-26. Modulation of the carrier-frequency oscillator by other oscillators which produce the bar-forming frequencies.

the pattern generator may be tuned. Tune the generator for the same channel, using the video carrier frequency or one somewhat higher, but still within the channel limits. Tests and observations may be made on any channel. If the generator is marked or graduated only for low-band channels, it nearly always will cover the high-band channels on third harmonics of the lower frequencies.

Place the receiver contrast and brightness controls at about their mid-positions to begin with. Adjust the receiver horizontal hold control to obtain stationary vertical bars, and adjust the vertical hold control for steady horizontal bars. It may be necessary to slightly readjust the generator frequency. To obtain good contrast, and equal brightness of horizontal and vertical bars, it will be necessary to adjust the receiver contrast and brightness controls, also the output controls or attenuators of the pattern generator.

When generator modulation frequencies are multiples of vertical and horizontal sweep frequencies, as explained earlier, the receiver hold controls will act in the same way as with transmitted pictures. The horizontal hold will tilt the vertical bars one way and the other, or will hold them in a true vertical position. The vertical hold will cause the horizontal bars to roll either upward or downward, or will keep them stationary.

TELEVISION

LESSON NO. 95
SIGNAL TRACING

Training Your Mind

C. T. I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind

C.T.I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON NO. 95
SIGNAL TRACING

Signals which arrive at the antenna of a television receiver contain all the picture and sound information and all the synchronizing pulses, presumably in correct relative amplitudes, or correct waveforms, and with correct phase relations. If these correct signals or their intended effects reach the picture tube and speaker there will be satisfactory performance. But if there is trouble, all or part of the signals will get lost, become distorted, or become excessively weakened somewhere along their route. All this is equally true of sound-modulated signals which reach the antennas of broadcast receivers.

Disappearance of a correct or satisfactory signal at the point where there is a fault is the basis of a method of trouble shooting called signal tracing. Signal tracing consists of observing the signal and its effects at various points in the receiver circuits. To observe modulated signals which are at carrier frequencies, intermediate frequencies, or other radio frequencies we employ some form of detector for demodulating the signals wherever they may be picked off the receiver circuits. The modulation then is passed through a suitable low-frequency amplifier and reproduced by a test speaker or other suitable indicator. To observe signals which are at audio frequencies or at sync pulse frequencies in the receiver, it is possible to omit the detector and apply these signals directly to the low-frequency amplifier of the tracing instrument.

A signal tracer for use on broadcast sound receivers consists of a detector, which most often is a crystal diode type, mounted within a probe and connected through a shielded cable to a sensitive amplifier whose output operates either a speaker or headphones. The interior of one such instrument is pictured by Fig. 95-1. The speaker maybe seen in the center of the panel. The two tubes in the foreground are an a-f voltage amplifier and a power amplifier. The rectifier tube for the self-contained power supply is at the far left. On the panel is a control for sensitivity or volume, also a jack or other suitable connection for the cable of the detector probe. A tracer of this style is also useful for working on f-m broadcast receivers and on many of the circuits in television receivers, provided the detector probe is suitable for pickup of the high frequencies found in such receivers.

TRACER PROBES. Fig. 95-2 shows the design of a crystal diode detector probe which is satisfactory for tracing in standard broadcast sound receivers and for much of the work in f-m and television receivers. By using tubular ceramic capacitors and 1/2-watt composition resistors this probe maybe built into a small tube with the arrangement of parts and connections illustrated below the circuit diagram.

Blocking capacitor *C1* may be of 1,000 to 1,500 mmf, rated at no less than 500 d-c volts. Capacitor *C2*, of 10 mmf, prevents harmful effects of self-resonance in the probe when used at television and f-m frequencies. Capacitor *C3*, of about 2.2 mmf, bypasses very-high-frequencies found in television circuits. Capaci-

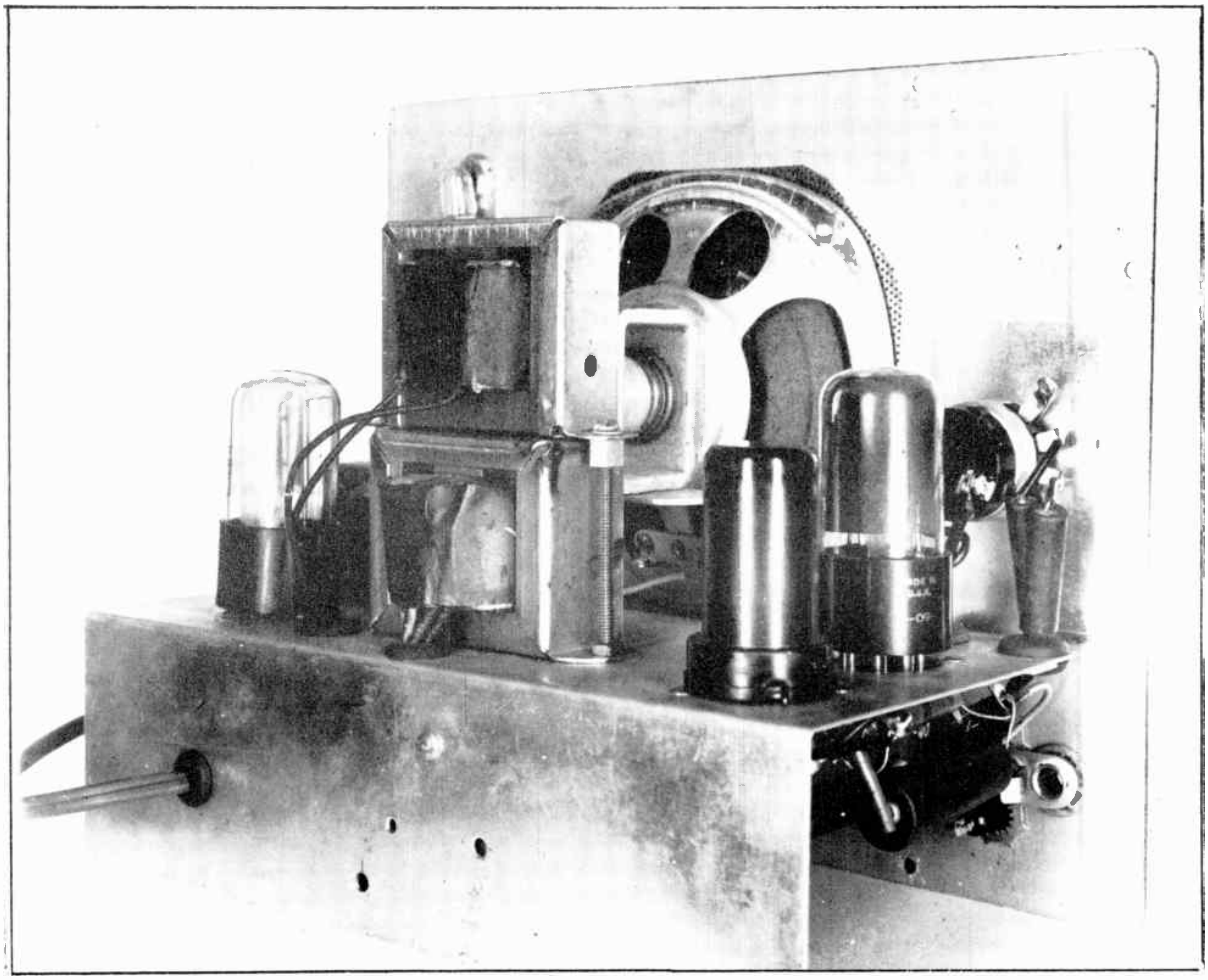


Fig. 95-1. A signal tracing instrument consisting of a sensitive audio amplifier and a speaker.

tor *C4*, of 500 mmf, bypasses the lower radio frequencies of all types of receivers. Resistor *R1*, of 100K ohms, completes a d-c return circuit for the signal. Filter capacitor *R2* should be of about 250K ohms.

This probe places a very considerable capacitance across the points tested, and tends to detune any circuit which is normally tuned to resonance. This disadvantage is balanced to a large extent by the fact that this one probe has good pickup for weak r-f signals, yet gives good response when applied to audio circuits. If the detuning causes severe loss of a signal it will be necessary to temporarily change the tuning adjustments of the tested circuit, the operating frequency of the test generator, or to use a probe of less capacitance such as will be described presently.

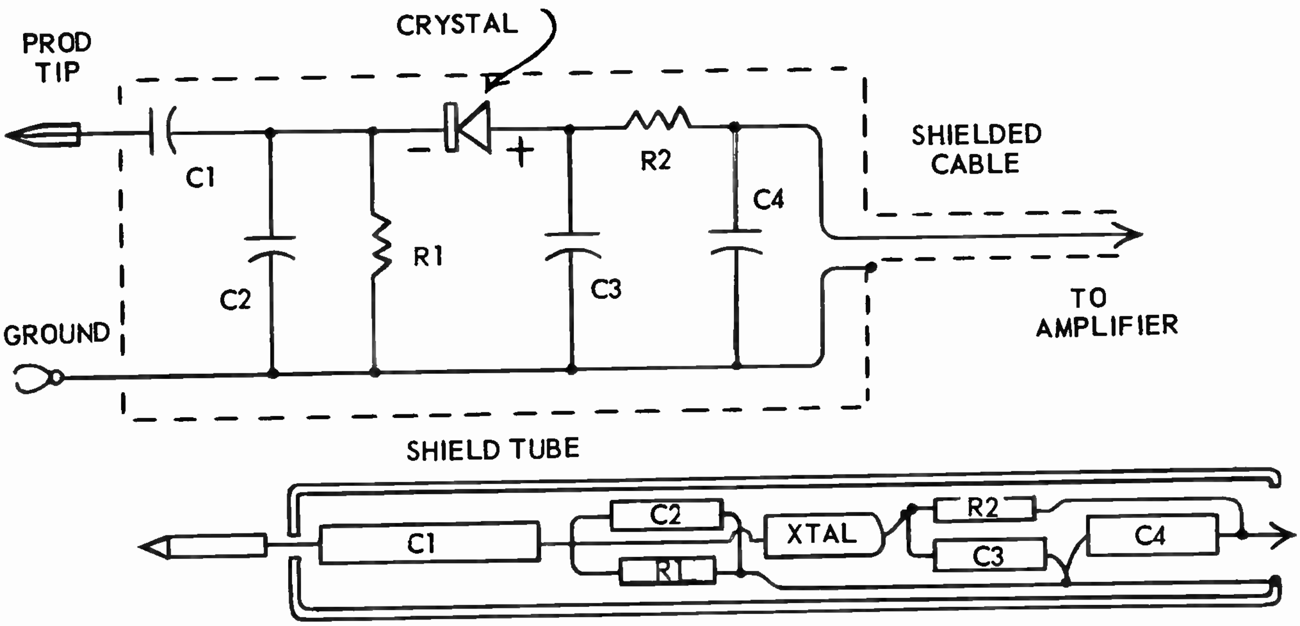


Fig. 95-2. Circuit and arrangement of parts in a detector probe suitable for nearly all methods of signal tracing.

For use only with standard broadcast receivers the detector probe may be as simple as illustrated by Fig. 95-3. Resistor *R* may be of any value between 50K and 500K ohms, with capacitor *C1* of 50 to 100 mmf. Since there is no blocking capacitor in this probe, one must be used in the amplifier as shown at *C2* or with some equivalent arrangement.

It is absolutely essential to use shielded cable between probe and amplifier, with the cable shield se-

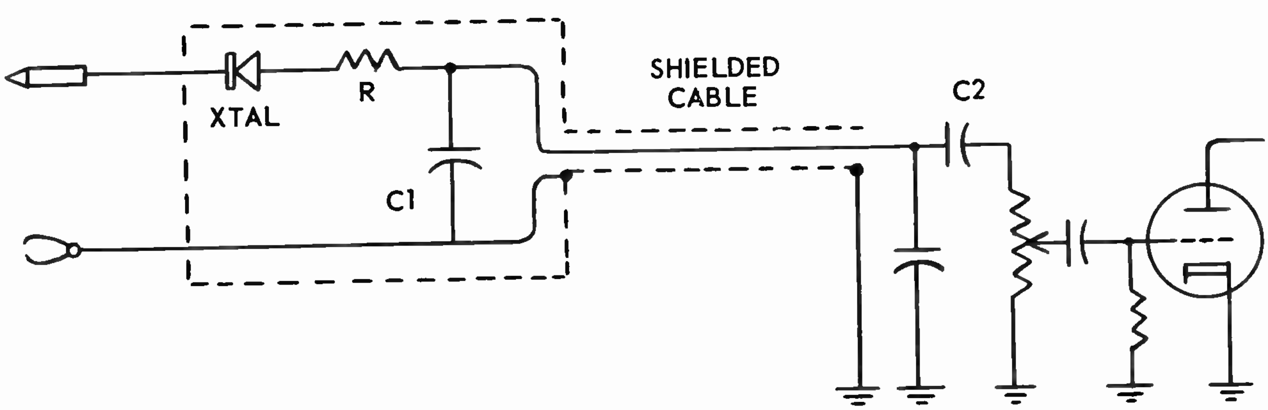


Fig. 95-3. A detector probe suitable for signal tracing in broadcast sound receivers.

curely connected to the probe shield and to chassis ground or B-minus in the amplifier. The lead for the ground clip on the probe is connected to the probe shield and to the cable shield.

TUNED TRACERS. Detector probes for signal tracing usually are untuned, they respond to any and all frequencies within the demodulation range of the detector element. The probes in Figs. 95-2 and 95-3 are untuned types. Amplifiers in the smaller and less costly signal tracing instruments are likewise untuned, they are simply audio amplifiers which operate on any low frequency or audio frequency which is applied to them from a detector probe.

In order to provide greater pickup sensitivity and also make it possible to identify frequencies existing at various points in receiver circuits, either the probe or the amplifier may be tunable. A tuned probe is somewhat similar to the untuned types so far as the detector and filter are concerned. The tuned probe has a small coil which may be tuned to resonance at various frequencies by means of a miniature ceramic adjustable capacitor connected across the coil. The coil, or a single turn in series with it, is brought near any point at which it is expected to pick up a signal frequency to which the probe is tuned. There is enough selectivity in the tuned probe to allow identifying a frequency, and enough resonant gain to provide some increase of signal voltage going to the detector portion of the probe. A tuned probe may contain either a crystal diode or a grid-leak tube detector.

Many of the larger signal tracers contain tuned radio-frequency amplifiers. An elementary circuit diagram for such a type is shown by Fig. 95-4. In this diagram are two tuned r-f amplifier stages followed by a combined diode detector and a-f voltage amplifier tube. The probe is not a detector type, but contains only a

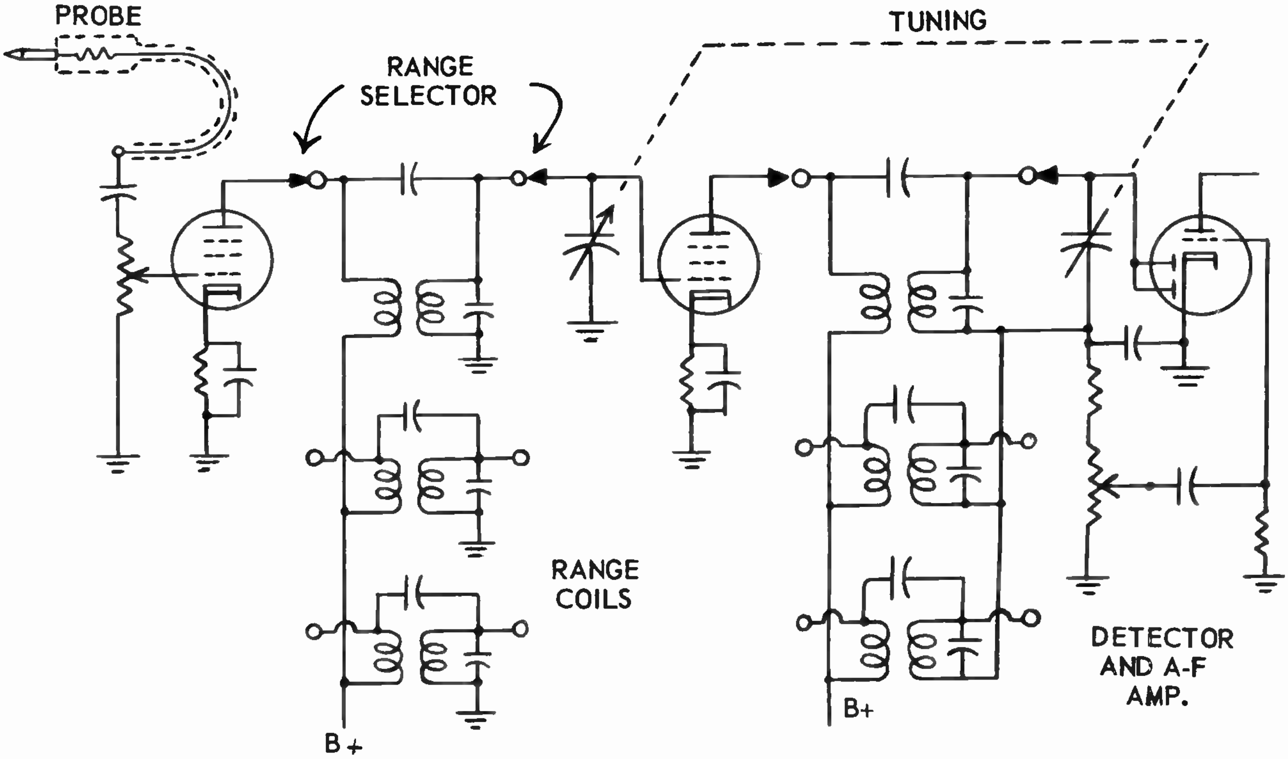


Fig. 95-4. Amplifier circuits and range selector connections for a tuned signal tracer.

resistor of 0.5 to 1.0 megohm for reducing the effect of cable capacitance. This resistor is connected through a shielded cable to an input attenuator and thence to the grid of the first r-f amplifier.

The tuned amplifier stages are designed to operate in any one of a number of radio-frequency bands. For each band there is a suitable coupling transformer having its individual trimmer. Any one of these transformers may be connected to the amplifier tube circuits by means of a range selector switch. Tuning throughout any selected band or range of frequencies is accomplished by a ganged variable tuning capacitor having one section for each amplifier stage. It is common practice to use more than the two amplifier stages shown in the diagram, also to have more than the three frequency bands represented here. Following the a-f voltage amplifier will be an output power amplifier and a speaker or other indicator. Electron-ray tubes often are used to insure correct tuning of the amplifier to whatever frequency is being picked up.

An amplifier of this general type may be tuned to the frequency at which a tested circuit should be operating when checking for such matters as gain. It is possible also to tune to any frequency at which there appears to be high response, thus identifying any frequencies which should not be present in the tested circuit. This is a means for determining the kind of interference which may be present.

AUDIO AMPLIFIERS. The audio or low-frequency amplifier of a signal tracer should operate at a low hum level or with small ripple voltage. This requires good filtering of the power supply, effective decoupling, and careful dressing of leads and parts. Tone fidelity seldom is of great importance, since our chief concern is with the presence or absence of any recognizable signal. If the test signal is to be provided by a signal generator, rather than from pickup of broadcast programs, the tracer amplifier should have good sensitivity at the frequency of tone modulation in the generator, which usually is 400 cycles.

The output of the signal tracer audio amplifier may be applied to various indicators other than a built-in speaker. From the plate circuit of an amplifier tube a connection may be made through a blocking capacitor of 1,000 mmf or more to a vacuum tube voltmeter which is set for a-c voltage indications. The secondary of the speaker coupling transformer may be connected to an a-c output meter such as used for alignment of sound receivers. Some tracers have a built-in meter, with a selector switch for using either the speaker, a meter, or any external indicator.

The generator, for providing a test signal may be any service type r-f generator, which will tune to carrier and intermediate frequencies, and which can be operated with audio or tone modulation. If the generator does not tune as high as the required carrier frequencies on fundamentals, its harmonic frequencies usually will be satisfactory. With the output of the signal generator coupled to the antenna, to the antenna terminals, or to the mixer grid of the receiver, the demodulated signal at the tone frequency will be heard from a speaker or will actuate any other output indicator when the detector probe is applied to any point in r-f, i-f, or a-f circuits reached by the signal. Methods of coupling the signal generator to the receiver are the same as for making alignment adjustments.

It may be assumed that receiver circuits are in working order as far as all points where the test signal can be detected and reproduced. When the signal no longer can be detected, or when it becomes noticeably faulty in some respect, some trouble may be assumed to exist between this latter point and the previous one yielding a good signal.

METHOD OF TRACING. The principal points at which a test signal would be checked in a conventional standard broadcast receiver are numbered in order on Fig. 95-5. With the signal generator coupled to the receiver, the tuning dials of both are set for the same frequency. It is preferable to use a test frequency near the low end of the reception band in order that capacitance of the probe may have minimum detuning effect. As mentioned before, the signal generator should be tone modulated. We shall assume that neither the probe nor the tracer amplifier is of a tunable type.

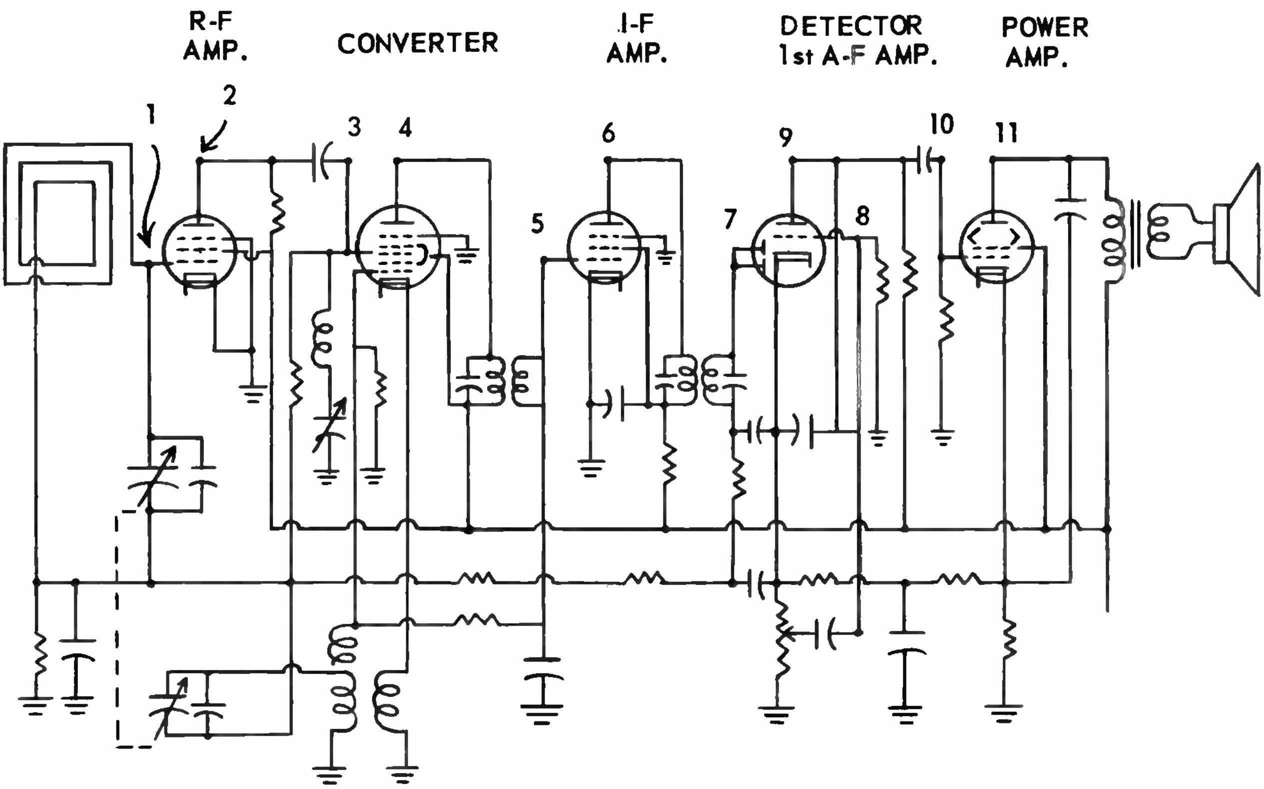


Fig. 95-5. The order of signal tracing in a typical standard broadcast sound receiver.

For a preliminary check the detector probe should be applied directly across the output of the signal generator while the attenuator of the generator is adjusted for a barely audible sound in the speaker or a low reading on any other indicator. This should be done with the volume control of the tracer amplifier at or near maximum. Then, with the probe applied to the grid of the r-f amplifier, point *1* of the diagram, either the receiver or the signal generator should be carefully retuned for maximum response, since dial markings for a given frequency may not be alike on the two pieces of apparatus.

The probe is connected next to the plate of the r-f amplifier, point *2*. Then response or output indication should be noticeably stronger than at the grid. The third check point is at the r-f signal grid of the converter tube, point *3*. Here the sound or other indication should be about the same as at the plate of the r-f amplifier. This completes the checks of the r-f stage.

With the probe still on the r-f signal grid of the converter, and the tracer volume control still at maximum, the attenuator of the signal generator should be turned down until the sound response or other indication is as weak as can be clearly recognized. Then the probe is shifted to the plate of the converter tube, point *4*. Here the signal should be much stronger than at the r-f grid of this tube. Without changing the adjustments of the generator attenuator or the tracer volume control, the probe is applied next to the grid of the i-f amplifier, point *5*. Signal strength should be about the same as at point *4*. This completes checking the converter stage.

Once more the signal generator ontput is reduced to bring about the weakest tracer response which is clearly recognizable at point *5*. Then the probe is moved to the i-f plate, point *6*, where a very decided increase of signal strength should be apparent. It is necessary to reduce the signal generator output after each stage is tested to avoid possibility of overloading an amplifier and thus causing very misleading indications.

The tests are continued in the numbered order until arriving at the plate of the power tube or output amplifier, point *11*. After checking at point *7*, the diode plates of the detector, the receiver volume control should be turned rather high, the signal generator output reduced to minimum, and the probe applied to the a-f grid, point *8*, while signal generator output is increased only enough to cause a recognizable response. In case the generator output cannot be attenuated enough to provide a weak indication with the probe at the grid, it will be necessary to turn down the receiver volume control.

The probe is applied next to the plate of the a-f voltage amplifier, point *9*, where signal indication should be very much stronger than at the a-f grid. This same signal indication, or nearly the same, should be found at the power tube grid, point *10*. Now, with the probe still at point *10*, reduce the receiver volume control to make the response just clearly recognizable. Shift the probe to the power tube plate, point *11*, where the response should be much stronger than at the grid.

Unless the receiver volume control is adjusted as described, the a-f voltage amplifier and the output amplifier are almost sure to be badly overloaded. It always is essential to keep signal generator output and receiver volume control setting below the points at which any of the amplifiers may be overloaded. An overloaded amplifier might appear to have no gain at all, or the signal at its plate might actually be weaker than one checked at the grid.

Should there be serious trouble in the r-f stage it might be impossible to obtain signal indications beyond the r-f plate, and possibly not even there. In this event the signal generator may be coupled through a blocking capacitor to the r-f signal grid of the converter tube, as shown by Fig. 95-6. The blocking capacitor may be of 1,000 mmf or larger.

The signal generator, still with tone modulation, is now tuned to the intermediate frequency used in the receiver. Then the detector probe is applied to the plate of the converter while the signal generator is retuned for maximum response from the tracer amplifier. It is, of course, not necessary to set the receiver tuning dial at any particular frequency when feeding the intermediate frequency directly to the converter grid. Remaining tests are carried out as previously described. It is essential, as always, to adjust the attenuator of the signal generator and the volume control of the receiver to avoid overloading any of the amplifiers.

Were a response to be absent even beyond the converter, the signal generator may be coupled through a blocking capacitor to the grid of the i-f amplifier, point *5* in the diagrams. The generator must be tuned to the intermediate frequency of the receiver. The probe then is applied at all points from the i-f plate, point *6*, through to the plate of the output amplifier. When the signal generator is connected to any point beyond the antenna input of a receiver in good condition, there will be less amplification between the generator and the point of application of the detector probe than when the generator is coupled to the antenna input. This will make it necessary to use somewhat greater output from the signal generator to have camparable signal indications.

F-M RECEIVERS. At first thought it might seem necessary to use a frequency-modulated test signal for tracing in f-m sound receivers, but this is not the case. Except that they are designed to operate at higher frequencies, the tuner and i-f amplifier circuits of f-m receivers are similar to those in an a-m set. It is not until we come to the f-m demodulator that there is a material difference. Consequently, if an amplitude-mod-

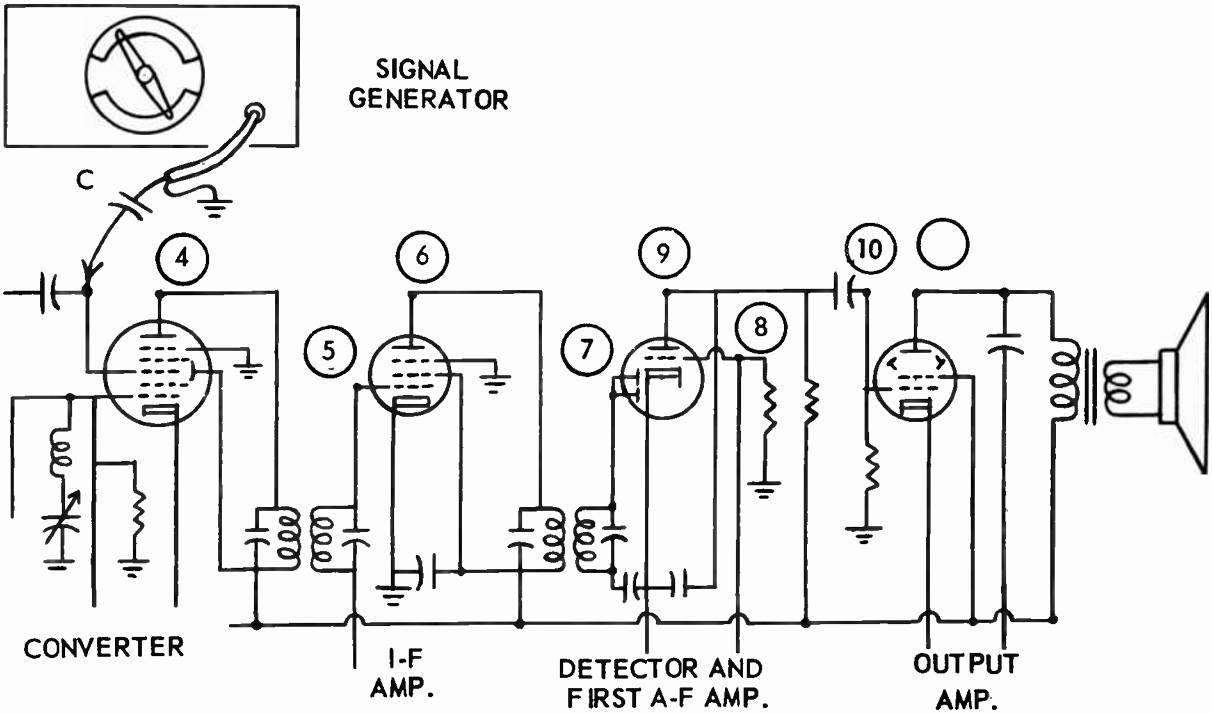


Fig. 95-6. Test points used for tracing when the signal generator is coupled to the r-f grid of the converter.

ulated signal of suitable frequency is applied at the antenna of an f-m receiver or a combination fm-am receiver this signal may be traced just as in a straight a-m set. The tracing can be carried even through the f-m demodulator when it is a discriminator type.

The similarity between high-frequency circuits in f-m and a-m sound receivers becomes apparent upon examining the partial diagram for a combination set as shown by Fig. 95-7. To simplify this diagram the output of the converter is coupled to the f-m intermediate-frequency amplifier. Actually there would be an additional combination fm-am stage with its dual coupling transformer like the transformer shown between converter and i-f amplifier in this diagram.

Even though we were to use a frequency-modulated test signal it could not be demodulated by a diode detector such as used in probes, since the diode is useful only for demodulation of a-m signals. This is true also of the triode grid-leak detector used in some probes. It is possible to construct an f-m discriminator with two crystal diodes, but probes employing this type of demodulator are not in general use.

Because of the high frequencies involved when tracing in f-m receivers, the signal generator must not have excessive leakage, and any portions of its output leads extending beyond the cable shield must be very short. Otherwise the signal fields may be picked up throughout the receiver circuits and carried into the amplifier of the tracer regardless of receiver tuning.

To avoid excessive detuning of f-m receiver circuits at their high operating frequencies, the detector probe must have small capacitance and high resistance. A probe constructed as in Fig. 95-8 has shunting

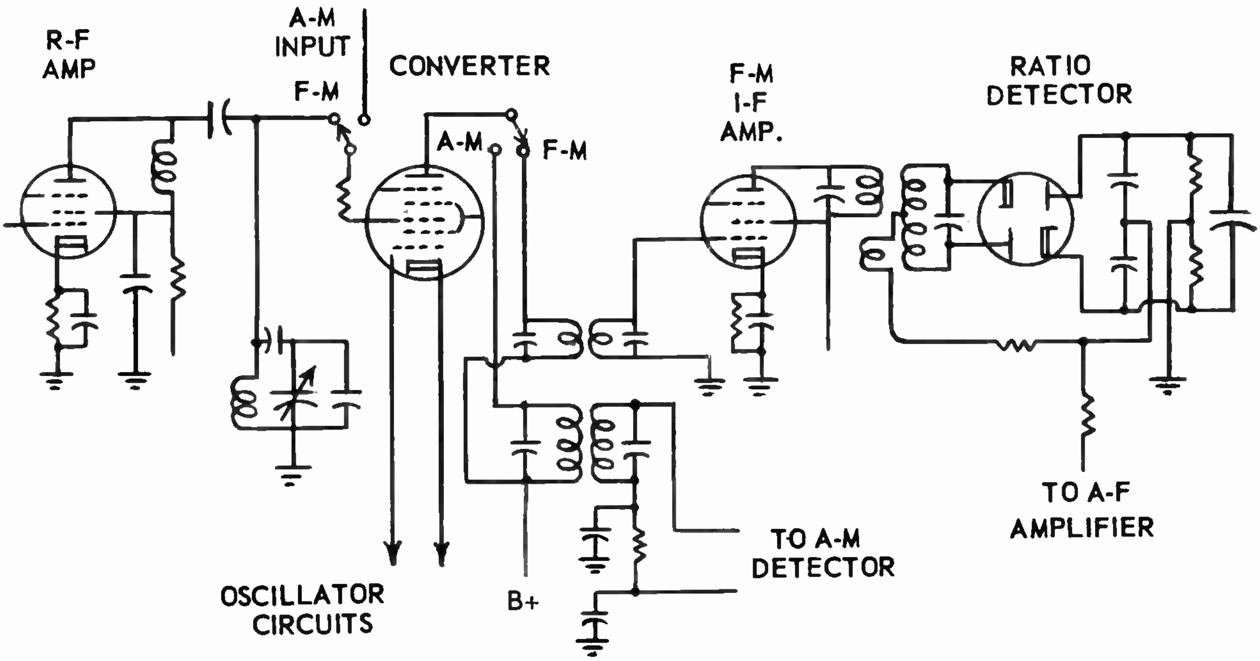


Fig. 95-7. Some of the circuits in a combination fm-am receiver.

capacitance of only about 6 to 8 mmf. Capacitor *C* is of 3 mmf. Resistor *R1* is 5 megohms, and *R2* is 100K to 500K ohms. This low-capacitance probe or something similar may be useful for checking narrow-band f-m amplifier circuits, if detuning of other probes prevents signal pickup. Because a low-capacitance probe greatly lessens the signal pickup it will be necessary to use fairly strong signals from the test generator.

Signal response on the input side of the f-m demodulator may be checked when using any of the probes

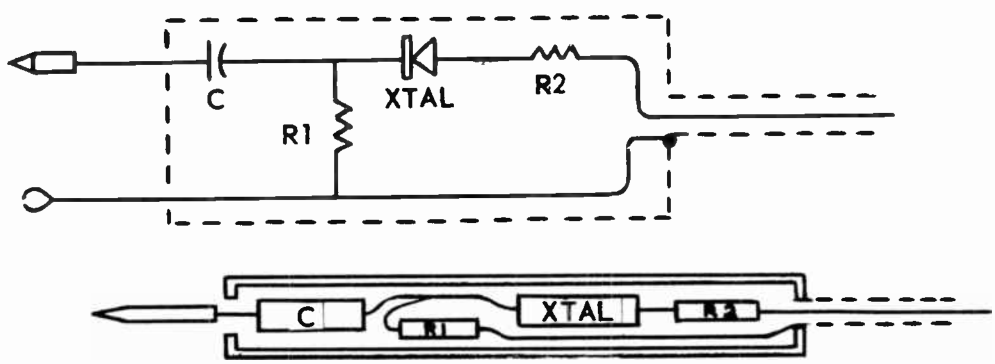


Fig. 95-8. A low-capacitance detector probe which causes only slight detuning of high-frequency circuits.

which have been shown. The tracing can be continued on the output side of a discriminator provided the generator output is kept so low as to prevent limiter action. There will, of course, be no response from an element which is grounded on the output or load side. The low-capacitance probe will not be satisfactory for tracing in the audio amplifier of an f-m receiver or any other receiver, since its pickup as audio frequencies is very weak.

VTVM FOR SIGNAL TRACING. A vacuum tube voltmeter equipped with a high-frequency detector probe may be used for signal tracing in circuits where signal voltage is strong enough to actuate the meter. As a general rule the voltmeter and detector probe, without additional amplification, will not be sensitive enough for r-f stage measurements unless the applied signal is strong enough to overload the amplifier tube. The VTVM will indicate the sum of all alternating voltages reaching it. This will include not only a desired signal voltage, but also hum voltage and noise voltage from tubes and circuit elements. The meter reading with no signal from a generator is due to these random voltages. The difference between this minimum and a high voltage with the generator in use is the voltage due to the desired signal.

Measurements which are more satisfactory may be made with the setup of Fig. 95-9. The output of the detector probe is fed to a sensitive low-frequency amplifier, such as found in untuned signal tracers. The output of the amplifier is applied to the a-c input of the VTVM. Noise voltage must be allowed for when interpreting the meter readings, as when using the VTVM alone, but now there will be added noise from the amplifier. As a general rule there will be less noise voltage when the receiver and signal generator are tuned to the lower carrier frequencies.

The receiver should be tuned to a carrier frequency at which no broadcast signals are picked up, if this is possible. Station signals are indicated by wavering of the VTVM pointer, in step with changes of amplitude in transmitted speech or music. If the gain control of the low-frequency amplifier is set too high the meter is likely to indicate pickup of transmitted signals practically everywhere on the receiver tuning dial, since there are 100 different carrier allocations between frequencies of 550 and 1550 kilocycles.

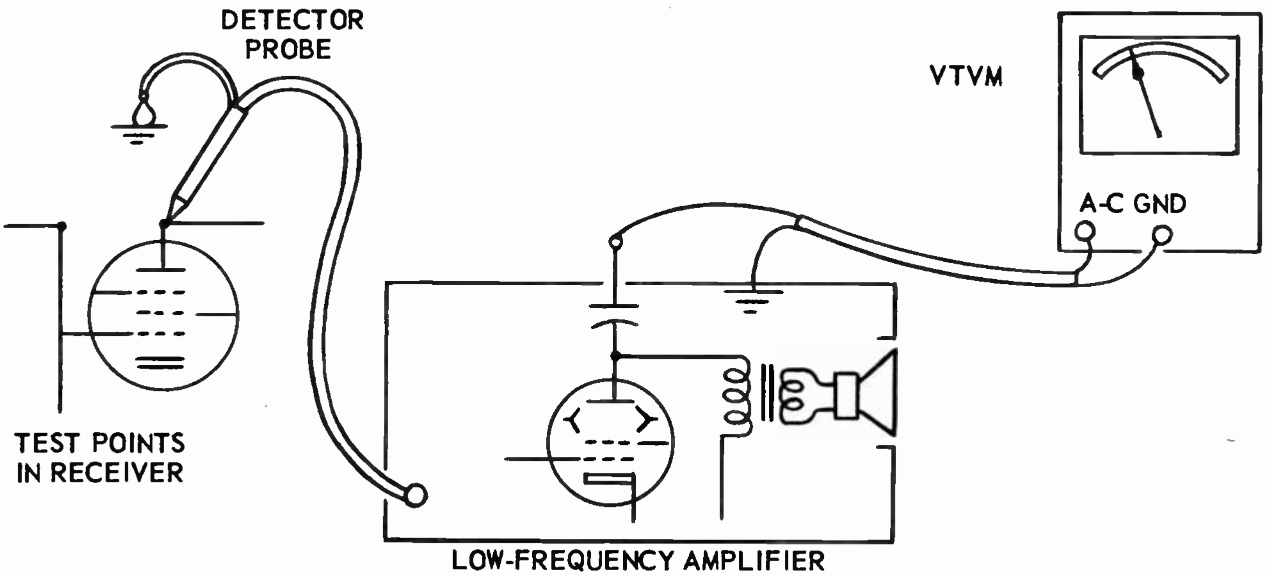


Fig. 95-9. How the output of a detector probe is strengthened by a tracer amplifier for application to an a-c vacuum tube voltmeter.

To avoid confusion between generator signal voltages and those due to transmitted signals, a speaker normally used with the tracer amplifier may be kept in operation while reading the VTVM. If the modulation tone of the generator rises and falls in intensity with rise and fall of the meter pointer, the pointer is following the changes of generator signal response and it will be known that the receiver and generator are tuned together.

When using the VTVM with a low-frequency amplifier it is important to adjust the generator output to avoid overloading of tubes in the tuner and i-f amplifier, and to adjust the receiver volume control to prevent overloading of the audio amplifier tubes in the receiver. When these precautions are observed, the VTVM allows fairly precise measurements of responses and of the differences between signal strengths at the input and output of any amplifier stage. This makes it possible to compute stage gains upon dividing the output volts by the input volts.

R-F OSCILLATORS. No signal or tone modulation will be obtained from any of the elements of an r-f oscillator in the tuner section of receivers, unless there is some incidental pickup from the mixer or converter grid. The r-f oscillator is checked by observing its d-c grid voltage or operating bias by means of the vacuum tube voltmeter.

As a general rule an r-f oscillator which is operating correctly will have negative d-c grid voltage of at least three or four volts, and often as high as 20 volts. If the grid voltage is zero or nearly zero it may be assumed that the oscillator is not operating. Only a d-c VTVM is satisfactory for this check.

TELEVISION SIGNAL TRACING. The processes of signal tracing in television receivers are best explained by considering the receivers as consisting of four principal sections. Certain general methods are used through the tuner, the video or combination i-f amplifier, video detector, and video amplifier. These parts make up our first principal section. The other three sections are the sound and audio portions of the receiver, the vertical sync and sweep sections, and the horizontal sync and sweep sections. In these latter three sections some special methods are of advantage.

The first section and the points through which a signal is traced may be represented as in Fig. 95-10. The generator is a constant-frequency type operated with tone modulation. The instrument may be a television marker generator, if it allows tone modulation, or it may be any service type r-f signal generator that will tune to television carrier frequencies. Harmonic frequencies usually are satisfactory when the generator won't reach the channel frequencies on fundamentals.

The signal generator output is normally connected to the antenna input terminals of the receiver, either with or without a matching pad. When it is necessary, in order to temporarily bypass a faulty stage, the generator may be connected to the mixer grid, to the grid of any i-f amplifier, or to the grid of the video amplifier. Grid connections always must be made through a blocking capacitor, as for aligning.

It is best to work in one of the lower channels of the low television band. As a general rule the responses will be satisfactory with the signal generator tuned to any channel frequency between the video and sound carriers, although the tuning may be varied within this range to cause best responses after the tracing is begun. The generator output always must be kept as low as will allow easily recognizable responses at the modulation tone frequency.

It is not necessary, nor is it desirable, to use a low-capacitance detector probe. General purpose probes or those suitable for waveform observation give good results. They do not cause excessive detuning, because it is possible to operate almost anywhere in the normal frequency response range of three to four megacycles. The probe usually will pick up enough of the modulation tone when brought close to but not in contact with the test points of the various circuits.

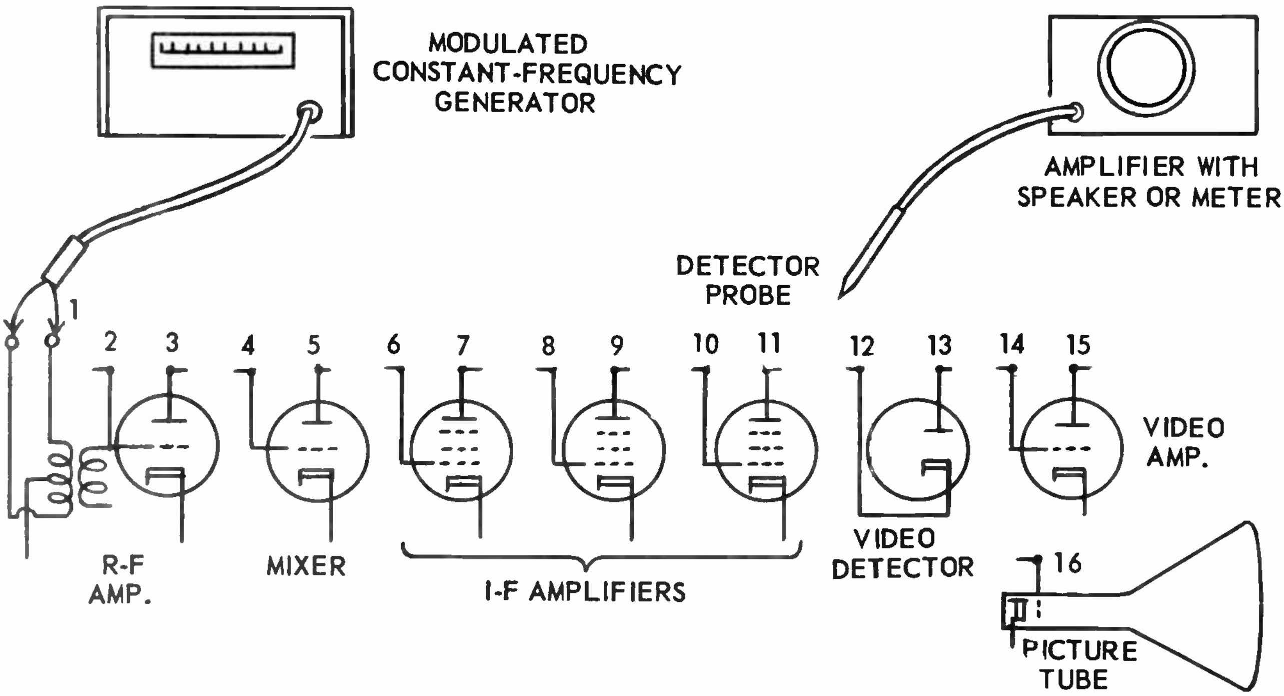


Fig. 95-10. Points at which a signal may be traced through the tuner and video sections of a television receiver.

In following explanations it is assumed that the detector probe is connected to the input of an audio-frequency amplifier which drives a speaker, this being the usual arrangement of signal tracer instruments. The amplifier may be connected also to an a-c VTVM, as described for work with broadcast radio receivers. The first check, after the signal generator is connected and operating, is to touch the probe tip to the antenna terminals while adjusting the attenuator of the generator and the volume control of the tracer amplifier to provide a weak response.

The next tests are made in order at the grid and then the plate of the r-f amplifier, and at the grid and then the plate of the mixer. During these tests the signal generator may be retuned to allow the clearest sound The responses should become progressively louder or stronger, and signal generator output should be reduced after completing the checks on each stage – just as in any signal tracing process.

The tests are carried on through the i-f amplifiers, to the input and output of the video detector, through the video amplifier or amplifiers, and to the signal input element of the picture tube, all as numbered in Fig. 95-10. Failure of the response at any point indicates trouble between that point and the preceeding one at which the modulation tone is clearly heard.

As mentioned earlier in this lesson, the action of the r-f oscillator in the tuner is checked by measuring its grid voltage with a d-c VTVM. The grid will be highly negative if the oscillator is working correctly.

TRACING IN SOUND SECTION. When the television receiver has a dual sound system rather than an intercarrier system the modulated constant-frequency generator may be coupled to the antenna terminals, the

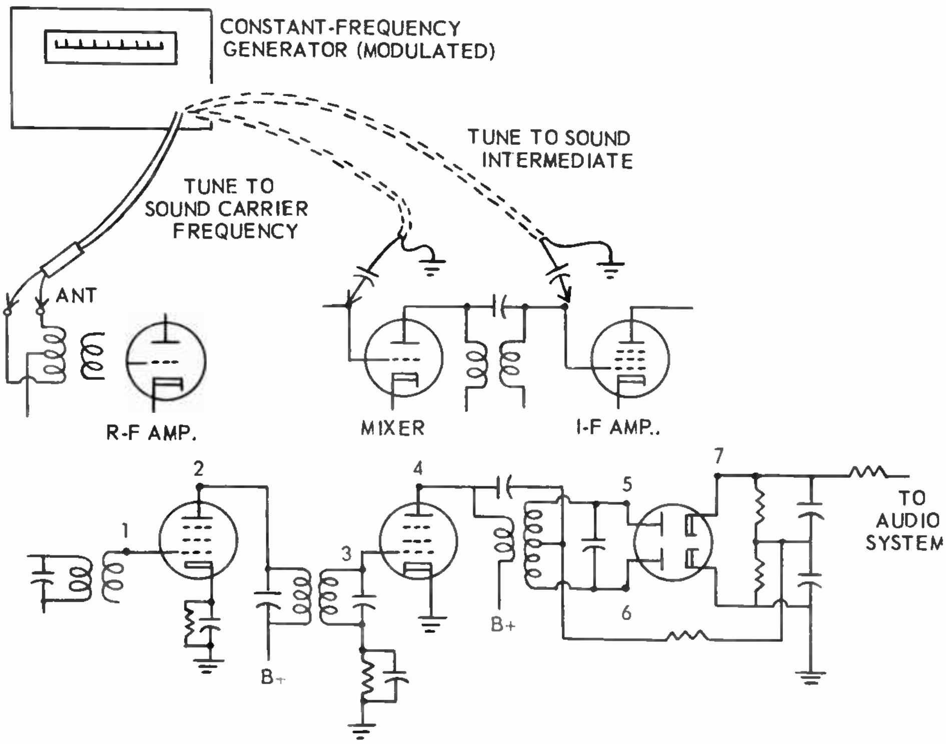


Fig. 95-11. Signal tracing through the sound section of a television receiver in which there is dual sound and a discriminator.

mixer grid, or to the grid of an i-f amplifier which precedes the sound takeoff. With coupling to the antenna terminals, the generator is tuned to the sound carrier frequency of the channel being used for tests. When coupled to the mixer grid or an i-f grid, the generator should be tuned to the sound intermediate frequency of the receiver. Signal generator connections are shown at the top of Fig. 95-11.

As mentioned in connection with signal tracing for f-m sound receivers, the f-m sound system of a television receiver will amplify or carry amplitude-modulated signals as far as the frequency demodulator. Although the limiter is intended to get rid of amplitude modulation, it does this only when the applied signal exceeds some minimum strength. If the output of the signal generator is kept low enough, the amplitude-modulation tone will come through a limiter. Then, if the demodulator is a discriminator type, the tone signal will appear at both the input and output of the discriminator.

Check points for a dual sound system are numbered in order on the diagram at the bottom of Fig. 95-11. Commencing at the sound takeoff or grid of the first i-f amplifier, tests are made at the plate of this amplifier, at grid and plate of any other i-f amplifiers, at grid and plate of the limiter, and at the input and output of the discriminator. One side of the discriminator output usually is grounded, and no response would be obtained there. If the amplitude-modulation tone passes through the discriminator it will continue through the audio amplifier system, and checks may be made at grids and plates of audio voltage and power amplifiers with direct connections from test points to amplifier input, omitting the detector probe.

For checking intercarrier sound systems the modulated constant-frequency signal generator may be con-

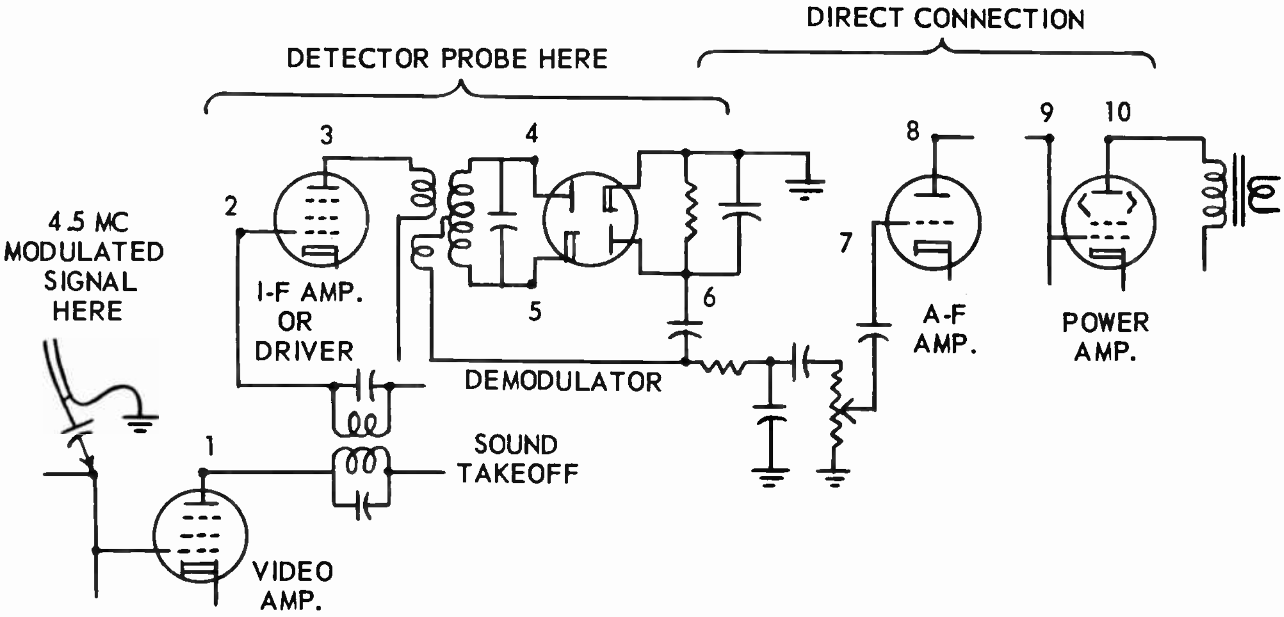


Fig. 95-12. Points at which the signal is traced through an intercarrier sound system.

nected to the receiver antenna terminals and tuned at or near the sound carrier frequency of the test channel. Otherwise this generator may be connected to the grid of the tube preceeding the sound takeoff, which usually is a video amplifier. This connection is shown by Fig. 95-12. The generator now is tuned to 4.5 megacycles, which is the sound intermediate for intercarrier systems.

Check points on a typical intercarrier sound system are numbered in order on the diagram of Fig. 95-12. The detector probe must be used at points *1* through *6*. Response should be about the same at *1* and *2*, much stronger at *3*, while at *4* and *5* the strength as indicated by sound should be much the same as at *3*. At *6* the response will be weak when taken through the detector probe, because now we are on the audio output side of the demodulator, and the capacitor in the probe has high reactance at audio frequencies such as those of the modulation tone.

Tests through the audio amplifying system are made without the detector probe, by using a probe having only a series resistor connected to the input of the low-frequency amplifier in the tracer instrument. These checks commence at point *6* and continue through to the plate of the power amplifier. This method of making direct connections, without the detector probe, may be used for all audio amplifying systems in all types of receivers.

The sound section may also be checked by using a television sweep generator or any service type f-m signal generator as shown by Fig. 95-13. When coupled to the receiver antenna terminals the generator is tuned to the sound carrier of the channel being used. At the mixer grid, tuning is to the sound intermediate. If the generator is coupled to the grid of a video amplifier preceeding the sound takeoff the generator is tuned to 4.5 megacycles for an intercarrier sound system or to the sound intermediate frequency for a dual sound system.

The audible response during tracing will be a 60-cycle hum or buzz when the sweep rate of the signal generator is 60 cycles, as nearly always is the case. Audio amplitude and loudness in any f-m reproducer varies with the amount of frequency deviation. Deviation in a sweep generator usually is called sweep width, so the loudness of sound response when tracing can be increased by increasing the sweep width, or reduced by decreasing the sweep rate. A wide sweep gives a sharper note which is more easily distinguished from the effects of 60-cycle hum voltages and fields such as radiated by heater circuits. This sharper note is, however, more easily confused with the response picked up from vertical sync and sweep circuits of the receiver, where sync pulses occur at a 60-cycle rate. The tracer response may be varied also by the output attenuator of the signal generator.

When using a frequency-modulated signal, test points are the same as shown by Fig. 95-12 for an intercarrier sound system, or as shown by Fig. 95-11 for a dual sound system. In either sound system the detector probe is used as far as it will pick up the high-frequency signals and demodulate them, with a direct connection to the tracer amplifier used beyond this point.

The 60-cycle note due to frequency deviation may be picked up all the way along an amplifier system without the detector probe, using a direct connection to the low-frequency amplifier. This method is not too satisfactory because, when the amplifier volume control is turned high enough for satisfactory response there is likely to be strong pickup of all 60-cycle and 120-cycle fields which exist in and around a receiver. This random pickup will occur even when the test leads are of shielded cable with the shield well grounded.

SYNC AND SWEEP SECTIONS. Signal tracing in the sync and sweep sections of television receivers, using a speaker or VTVM as response indicator, is not so satisfactory as similar work in the video and sound sections. Difficulties are due chiefly to the facts that sync circuits do not treat a signal as do ordinary amplifier circuits, and there is such wide variety in sync circuits that indications from one receiver

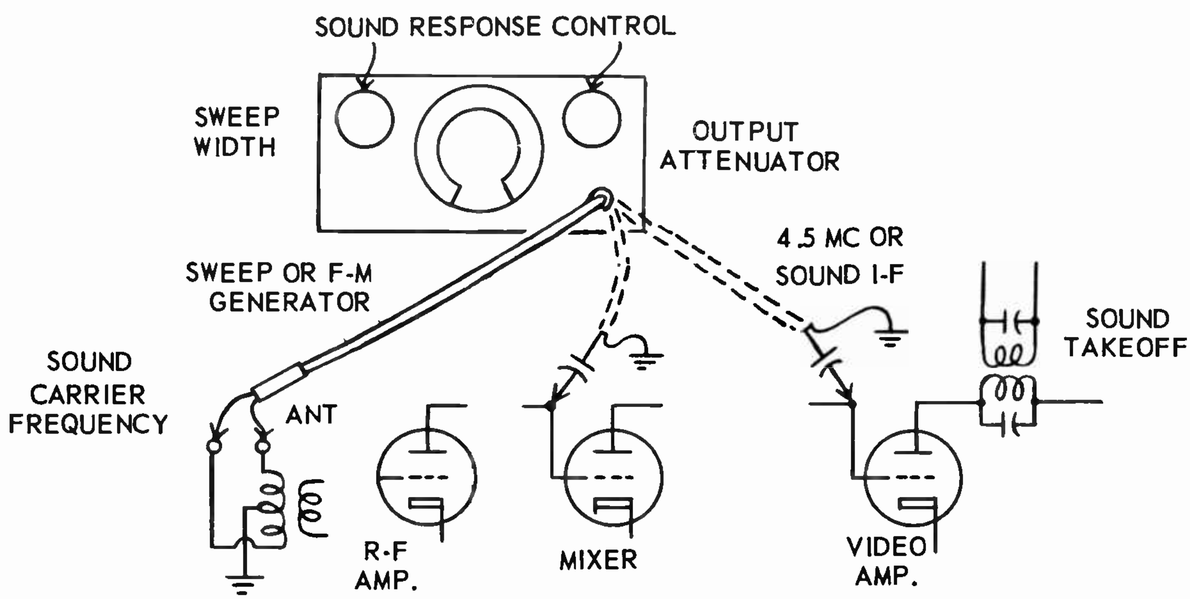


Fig. 95-13. Generator connections when using a sweep frequency or an f-m signal for tracing through television sound sections.

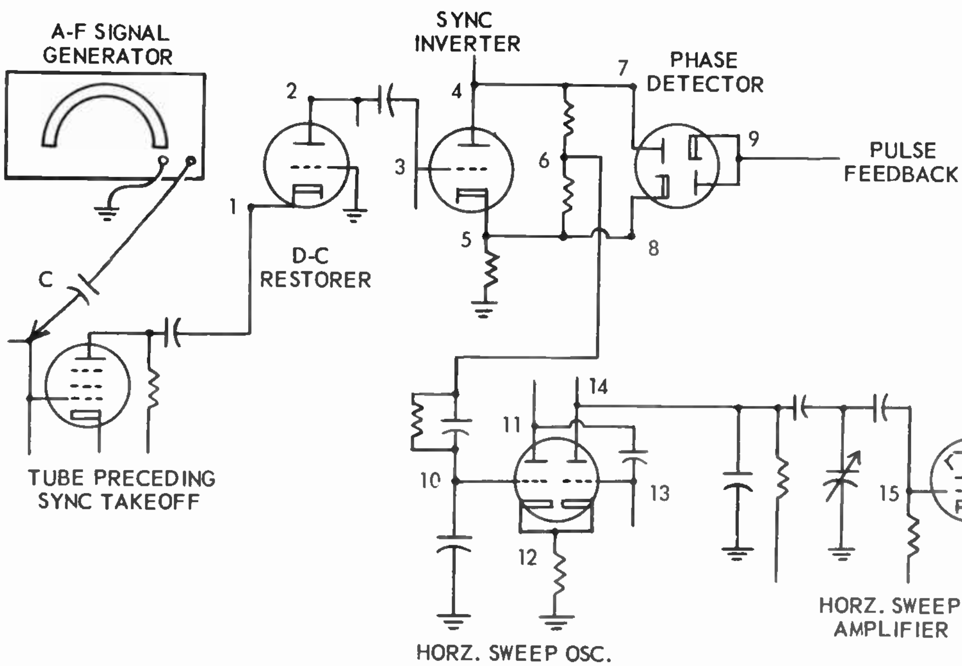


Fig. 95-14. Connection of an audio-frequency generator, and the check points followed, when tracing through a horizontal sync and sweep section.

may be quite unlike those from a different set. If you do a great deal of service work on one type of receiver it is a good idea to trace through circuits known to be operating correctly, and make notes of the responses. Failure to obtain generally similar responses from other sets of the same kind will then indicate the locations of faults.

One method of tracing through a sync section and a horizontal sweep section is illustrated by Fig. 94-14. The circuits shown here are only one of the many combinations which may be found. Since tracer responses will truly indicate normal operating conditions only when the test frequency is that for which the circuits are designed, it is necessary to have an andio-frequency signal generator. The horizontal line frequency is 15,750 cycles per second, a frequency which is not audible to one person in a hundred. To have audibility, the generator frequency should be made a few hundred cycles either higher or lower than 15,750 cycles. If tested circuits are operating, their horizontal line frequency will beat with the generator frequency, and you can easily hear the difference or beat frequency when using the detector probe as a mixer for the two frequencies.

The audio generator is to be coupled through a capacitor of 0.1 mf or larger to the grid of the tube preceeding the sync takeoff, which is usually a video amplifier. With the particular circuits of Fig. 95-14 the detector probe should pick up the audible beat note at all the points which are numbered in the order of checking. If you are not certain that the beat note is heard at any one of the test points, keep the probe at this point while slightly varying the operating frequency of the audio generator. If the audible tone pitch rises and falls, you are obtaining a correct indication. Should the audio generator be tuned to the same frequency as the sweep oscillator of the receiver there will be zero beat, and no tone can be heard.

If the sweep section is operating normally, sound bars will appear on the picture tube. The number of bars will correspond to the audible beat frequency. These bars will disappear, and leave only the raster, when there is zero beat between the sweep oscillator and the audio generator frequencies.

For tracing in the vertical sync and sweep sections no signal generator is needed, since 60-cycle buzz due to vertical sync pulses will be present and clearly audible from circuits which are operating normally. A detector probe may be used, but equivalent responses will be obtained with a direct connection to the tracer amplifier. The response should be a buzz due to the square waveform of the pulses and to sharp voltage pips. The response should not be a soft hum.

Check points in a fairly typical vertical sync and sweep section are numbered in order on the diagram of Fig. 95-15. The response should be followed through the integrating filter, points *5* to *8*. Usually there will be some 60-cycle hum from the tracer amplifier. This hum will beat with the sync and sweep response to cause wavering variations of response volume. The rate at which the response thus varies may be changed by adjustment of the vertical hold control of the receiver.

SOUND BARS FOR SIGNAL TRACING. Signal tracing between the television picture tube and the antenna input may be carried out by utilizing the appearance or lack of appearance of sound bars on the picture tube as the response indication. The principles of this method are illustrated by Fig. 95-16 Since the response indicator (the picture tube) cannot shift around like a detector probe it is necessary to shift the connection of the signal source, which usually is some type of signal generator. Test points then will commence as near as possible to the picture tube and will end with the signal source at the antenna input as shown by the numbering on the diagram.

Sound bars will appear so long as tubes and circuit components between the picture tube and the point of signal application are in working order. The bars will disappear or undergo decided changes of appearance as soon as a fault has been passed, because this fault will then be in the path of the signal between the source and the picture tube.

For checking the video amplifier and video detector stages the signal source is preferably an audio-frequency signal generator operating at 60 cycles or any higher audio frequency. The audio signal is applied to the grids of video amplifiers and then to the input side of the video detector, points *1*, *2* and *3* on the

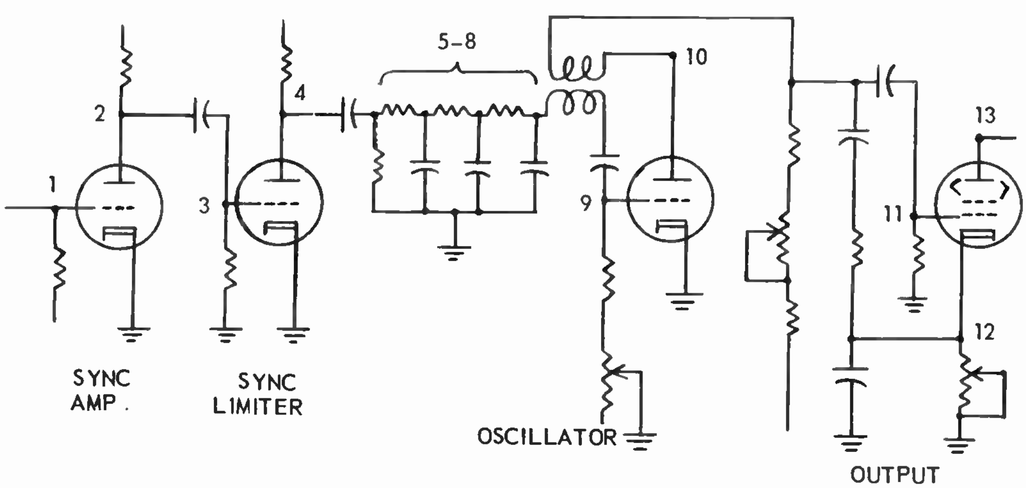


Fig. 95-15. Check points which are followed when tracing through a vertical sync and sweep section.

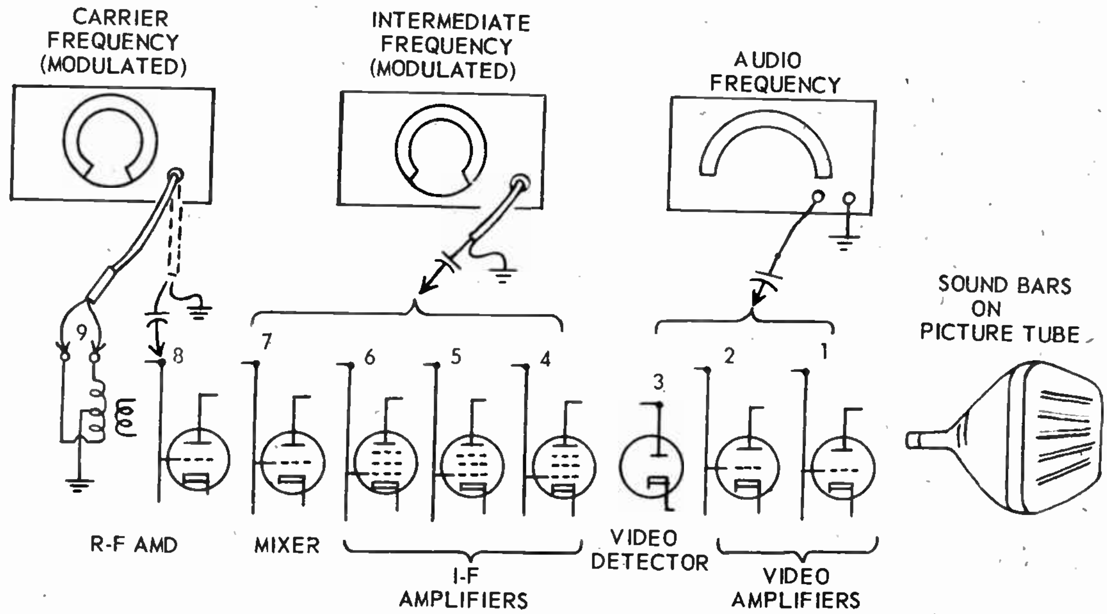


Fig. 95-16. Frequencies applied at various points, and their order, when signal tracing with picture tube sound bars as the indication.

diagram, with a blocking capacitor in series. The lower the test frequency the greater must be this capacitance. About 0.25 mf is satisfactory at 60 cycles, dropping to around 0.01 mf for 600 to 1,000 cycles.

For checking the i-f amplifier stages and the mixer output circuits, points *4* to *7*, the signal source should be a constant-frequency generator tuned within the frequency response range of the i-f amplifiers of the particular receiver being tested. That is, with a video intermediate frequency of 25.75 mc and sound intermediate of 21.25 mc, the generator might be tuned to whatever frequency between 22 and 25 mc allows the most distinct bars on the picture tube. The generator must be operated with tone modulation. Coupling to the tube grids is through a capacitor of about 1,000 mmf.

For checking the tuner circuits, points *8* and *9*, the tone-modulated constant-frequency signal generator should be tuned to any frequency in the higher portions of the carrier response or the channel frequencies of the channel being used. The receiver must, of course, be tuned for the same channel as the generator. The generator signal output may be quite low for tests at points *9* back to *7*, but will have to be progressively stronger for points *6* back to *4*.

The number of sound bars on the picture tube will correspond to the operating frequency of the audio generator or to the modulation frequency of the constant-frequency r-f generator. The bars may be held stationary, if desired, by altering the modulation frequency or the audio frequency, or by adjusting the vertical hold control of the receiver.

Tests may be made at points *1*, *2*, and possibly at *3* with a 60-cycle frequency taken from any tube heater circuit through a capacitor of about 0.25 mf. A test frequency at 120 cycles may be obtained from the output side of a full-wave power rectifier of from the input to the power filter. In this case it is necessary to use a series capacitor rated for at least 600 volts, and to use also a series resistor to drop the high voltage which appears at the rectifier or filter output. A series combination of about 0.05 mf and 50K to 100K ohms usually will provide enough signal, with voltage low enough at the test points to be safe.

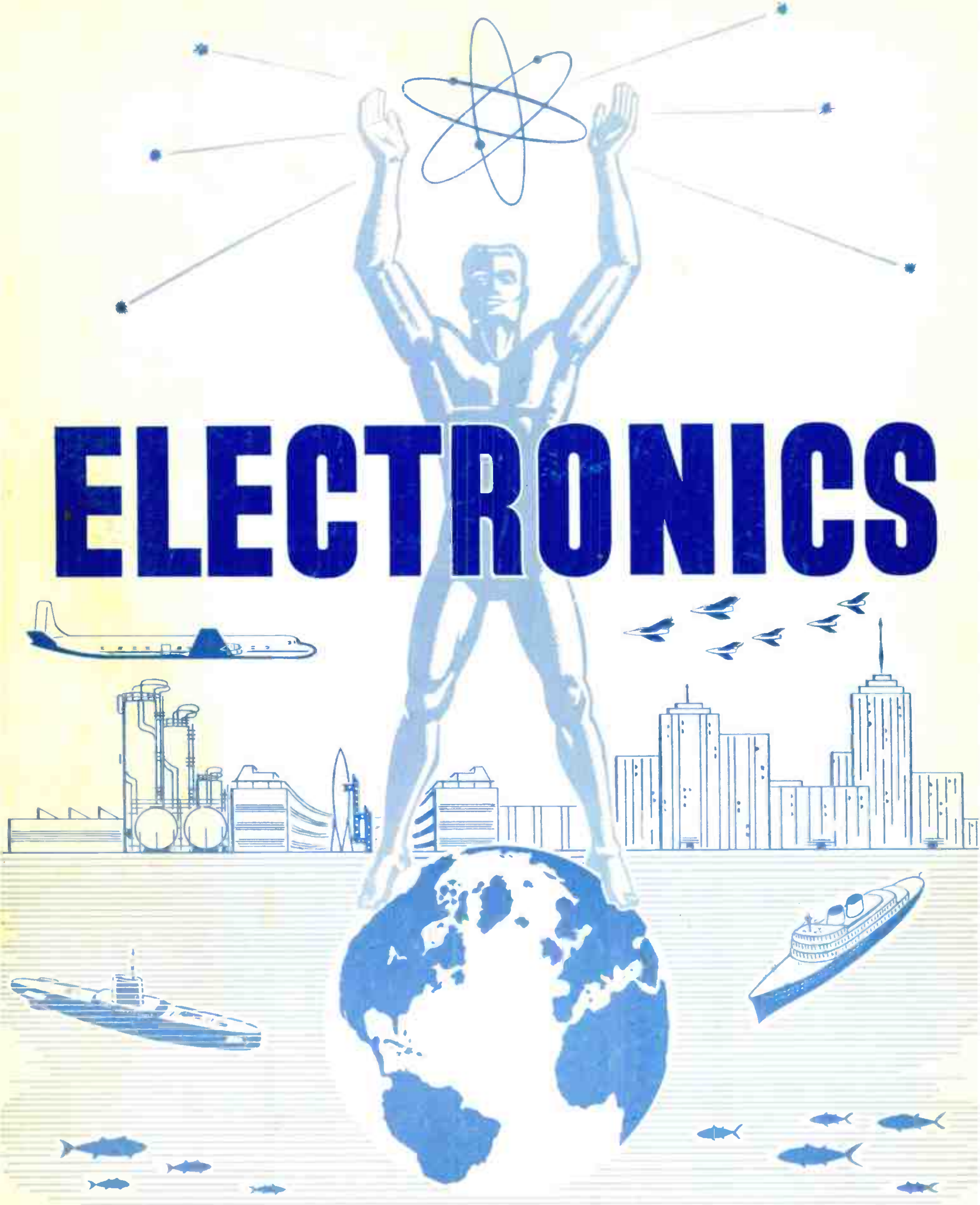
Training Your Mind

C.T.I.

Perfecting Your Skill

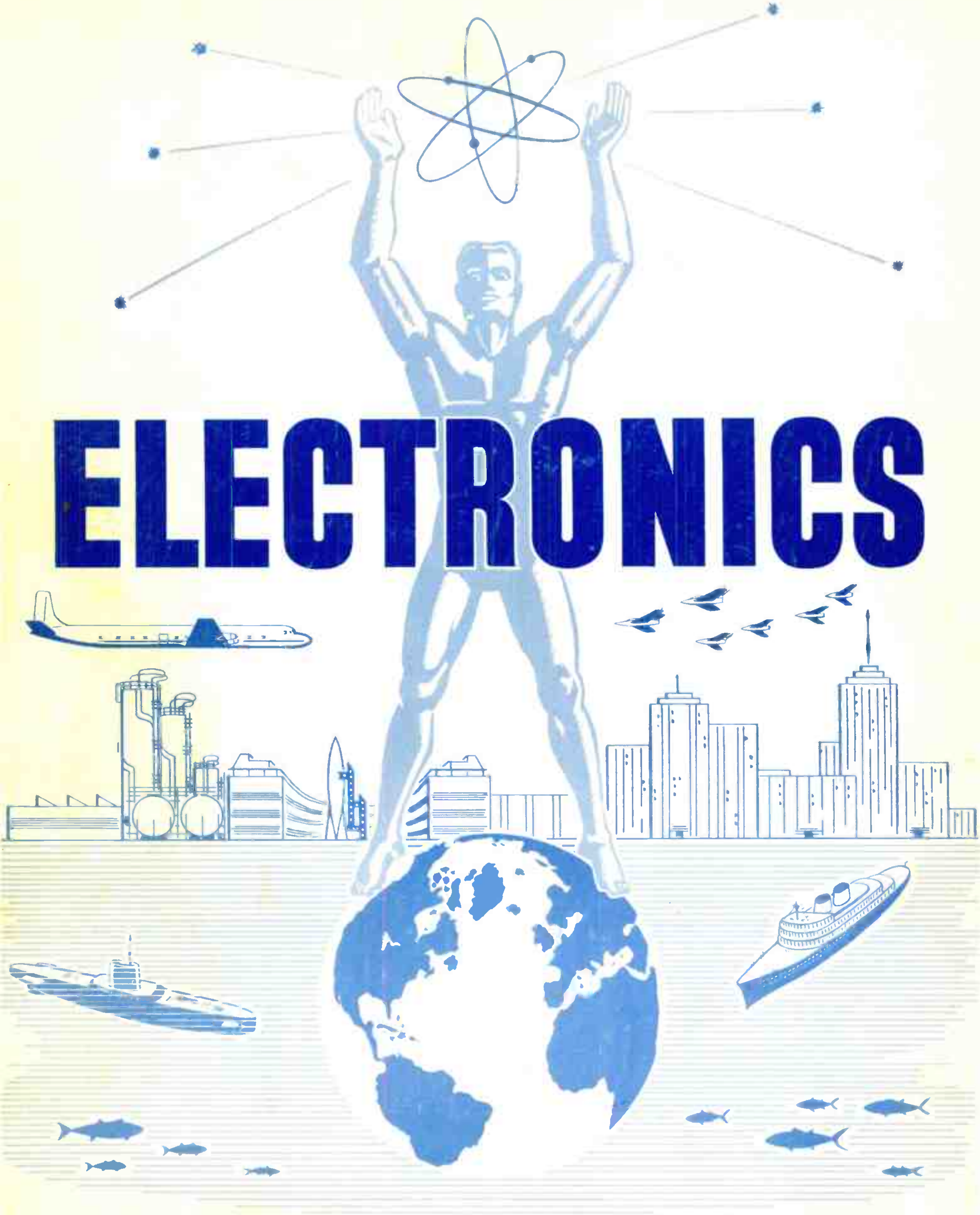
96→100

ELECTRONICS



COMMERCIAL TRADES INSTITUTE • CHICAGO 26, ILLINOIS

96→100



ELECTRONICS

COMMERCIAL TRADES INSTITUTE • CHICAGO 26, ILLINOIS

TELEVISION

LESSON NO. 96

TROUBLE SHOOTING WITH THE OSCILLOSCOPE

Training Your Mind
C.T.I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind

C.T.I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON NO. 96

TROUBLE SHOOTING WITH THE OSCILLOSCOPE

An oscilloscope makes a very satisfactory indicator in various processes of signal tracing. One of the methods commonly employed is illustrated by Fig. 96-1. The signal source is a tone-modulated r-f generator. The output of the generator may be connected to the antenna terminals of the receiver, and tuned anywhere in the range of strong carrier-frequency response for the channel in which tests are to be made.

If the tuner circuits are not to be traced, the signal generator may be coupled through a small capacitor to the grid of the mixer, and tuned in the range of i-f response of the particular receiver being checked. It is possible also to couple the signal generator output through a capacitor to the grid of any i-f amplifier tube, with tuning still in the stronger portion of the i-f response. This latter connection is used when serious trouble exists between antenna input and any i-f stage, which would prevent a signal introduced at the antenna or mixer from coming through all of the i-f amplifiers.

The vertical input of the oscilloscope may be connected to the signal input of the picture tube, to the plate of any video amplifier, or to the load of the video detector. No detector probe is needed, because the oscilloscope always is connected after the regular video detector in the receiver.

The internal sweep of the scope is set for the modulation frequency of the signal generator, or to some simple fraction of the modulation frequency. Then the scope will display one or more sine waves or approximate sine waves of the modulation when all circuits between the points of generator and scope connections are operating.

Usually it is convenient to commence tracing with the scope connected to the video detector load, and the signal generator to the grid of the last i-f amplifier. Then the generator connection is moved back, stage by stage, until it is at the antenna. Should the sine-wave trace on the scope disappear or become noticeably defective when the generator connection is shifted to some one point, trouble is indicated between that point and the last one at which the trace was satisfactory.

With the signal generator connected at the antenna terminals, and no trouble yet apparent, the scope connection may be shifted, stage by stage, through the video amplifier until reaching the picture tube input. This will check from the video detector to the picture tube, with trouble indicated between any point of unsatisfactory trace and the last point at which there is a good trace.

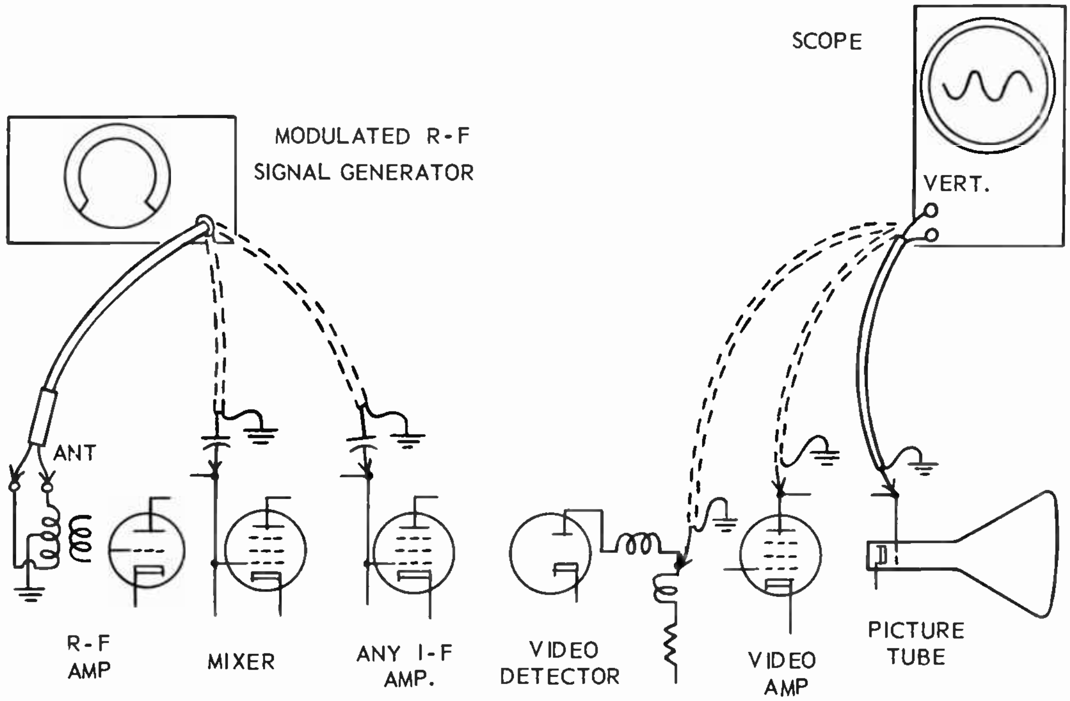


Fig. 96-1. Tracing a tone-modulation signal with the oscilloscope connected beyond the video detector.

A different method of signal tracing with the oscilloscope is illustrated by Fig. 96-2. The modulated r-f signal generator is coupled to the antenna terminals and tuned to a channel frequency, or is coupled to the mixer grid and tuned to the intermediate-frequency range of the receiver being tested – just as with the method first described.

To the vertical input of the oscilloscope is attached almost any type of detector probe except one having small capacitance and low sensitivity. The internal sweep of the scope is adjusted for display of one or more waves at the modulation frequency being used in the signal generator. The detector probe then is applied successively at the input for the video detector, to the plate of each preceding i-f amplifier, and finally to the plate of the mixer. Trouble is indicated between the first point at which there is a defective trace or no trace, and the last point showing a satisfactory trace. The probe may be applied to the grids of the i-f amplifiers, but the detuning effect probably will be too great to allow satisfactory indications

When tracing a signal with the oscilloscope and a generator it is advisable to keep the generator output just as low as yields a satisfactory or readable trace with the vertical gain of the scope turned high. It is just as necessary to avoid overloading any of the tubes with this method of tracing as when using any other kind of response indicator.

It is possible for amplifier stages to successfully handle a high-frequency generator signal with sine

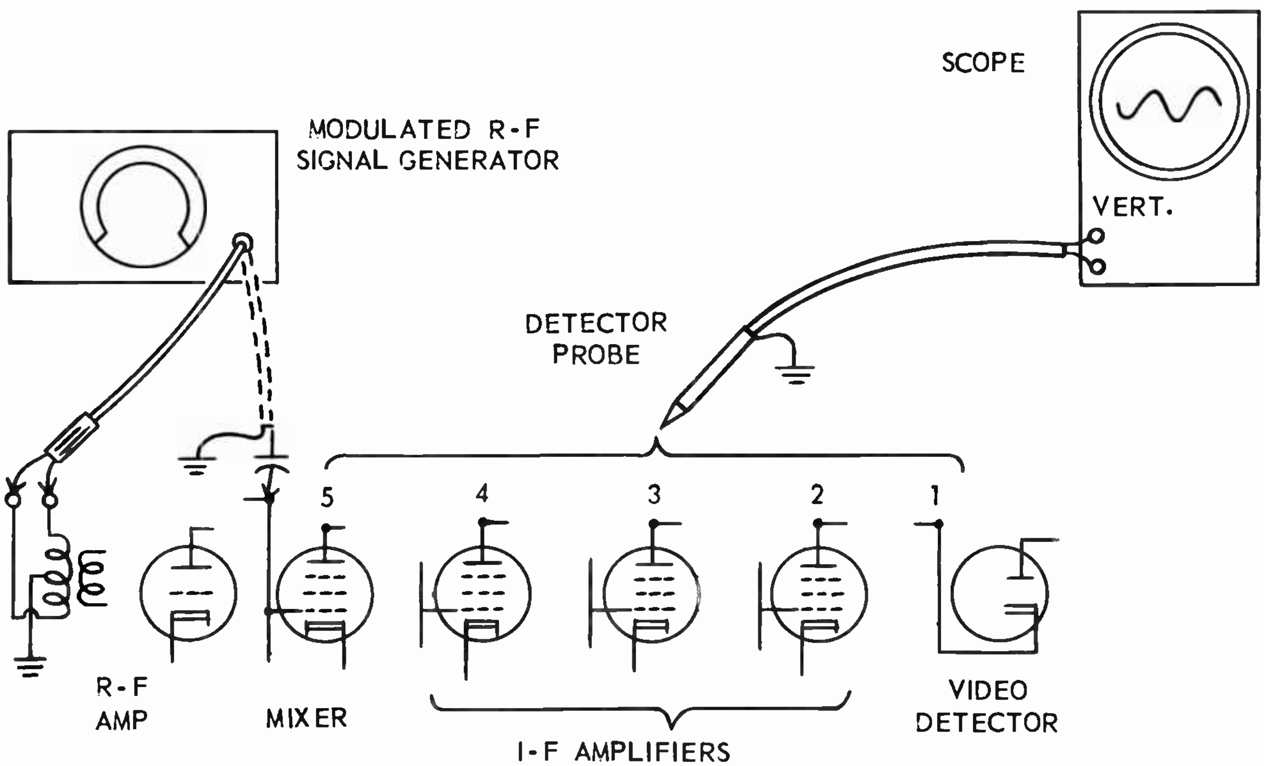


Fig. 96-2. Tracing of tone modulation with a detector probe on the oscilloscope.

(2) wave modulation at an audio frequency, yet fail in some manner to handle the highly complex television signals wherein the modulation frequencies extend from 60 cycles to three or four megacycles. This is especially true in the video detector and video amplifier stages. It is also true that tracing a sine wave through sync and sweep sections is by no means a proof that these portions of the receiver are operating correctly. Here it is the waveform of the signal, and to some extent the peak-to-peak voltages, that are all-important. The square waves and voltage pips of sync pulses may become badly distorted even though some other kind of signal will pass through the circuits.

(4) Tracing in the sync and sweep sections, with a vacuum tube voltmeter, indicates only average gain or lack of it. The meter, unless of some special type, will not read peak-to-peak voltages, and its indications have no particular relations to waveforms.

WAVEFORM TRACING. By far the most positive method of signal tracing in television receivers is one utilizing an oscilloscope to display signal waveforms picked off at various points between antenna input and picture tube.

Some principles of waveform tracing are shown by Fig. 96-3. The signal input is any regularly trans-

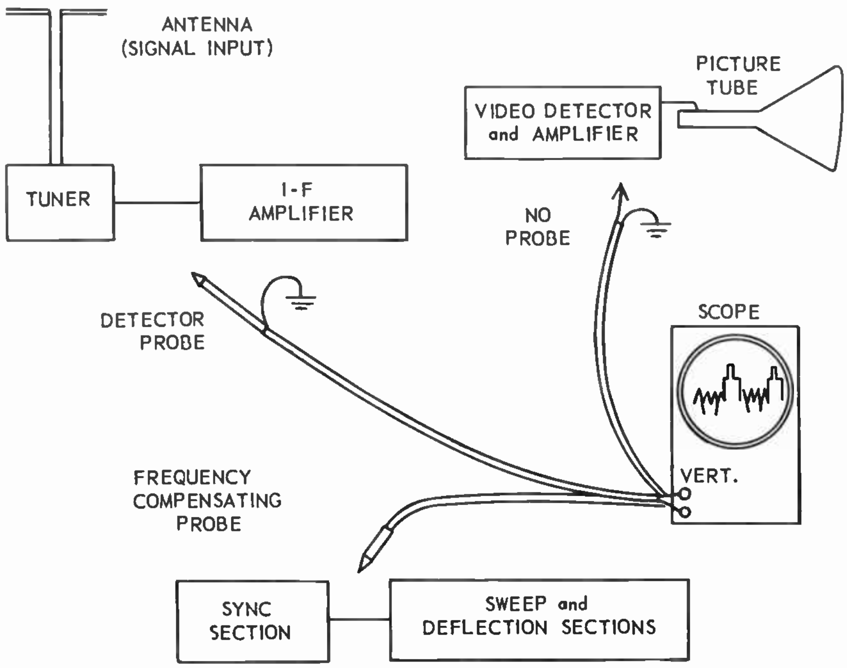


Fig. 96-3. Tracing a television signal with the oscilloscope as indicator.

mitted television signal which is received through the antenna and carried to the tuner. Signals in the lower channels usually give strongest responses, although this depends on the locality and its reception conditions.

For tracing through the tuner and i-f amplifier as far as the input to the video detector the vertical input of the scope must be fitted with a detector probe. This probe is preferably of a type well suited to waveform demodulation. Such types are discussed in lessons which deal specifically with detector probes.

5 For tracing between the output or load of the video detector and the picture signal input to the picture tube no probe is needed, since the video detector takes care of demodulating the signals. Usually it is necessary to use some kind of capacitor bypass across the scope input to prevent fuzzy traces should the scope have high gain at high frequencies. Such filtering is explained in lessons dealing with oscilloscopes.

If the oscilloscope does not have a frequency-compensating input attenuator, the tracing through sync and sweep sections may be carried out with no probe on the vertical input of the scope. Waveforms will be somewhat distorted, particularly in rounding of the corners on square-wave pulses and in making these pulses appear as tapering peaks rather than of square-wave form. The sync and sweep waveforms still should be recognizable to an extent which allows identifying the more serious kinds of troubles.

If the oscilloscope contains a frequency-compensating input attenuator it is well worth while to use a frequency-compensating probe on the vertical input when tracing through sync and sweep sections. Circuits used in such probes are shown by Fig. 96-4. The essential parts consist of resistor R1, usually of a value between 1 and 10 to 20 megohms, which is shunted by capacitor C2 of some value between 2 and 8 mmf. The required value of C2 depends on the oscilloscope with which the probe is used, so this unit may be a miniature adjustable ceramic type.

Capacitor C1 seldom is used, but when included its value should be less than 0.1 mf. This blocking capacitor has no effect on performance of the probe. Filter resistor R2 is not often included. If used, it may be of 50K to 200K ohms. It is important that the shielded cable have low capacitance and low energy loss. Coaxial cables such as used for transmission lines are satisfactory. The cable should be no longer than necessary, in order to reduce its capacitance.

A compensating probe does not greatly alter the input resistance of the probe and scope, but it greatly reduces the input capacitance. This results in a large reduction of input impedance and great improvement of waveform reproduction. The vertical sensitivity of the scope and probe usually is between 1/10 and 1/15 that of the scope without the compensating probe. This makes it necessary that the scope have high vertical sensitivity if waveform amplitudes are to be high enough for satisfactory observation.

The easiest way to adjust the capacitance at C2 is to apply the probe, on the vertical input of the scope, to the video detector load resistance while receiving any regularly transmitted television signal. Assuming that the receiver circuits are in good order and correctly aligned, capacitor C2 then may be adjusted to obtain traces in which sync pulses have the most nearly vertical leading and trailing edges, and the flattest tops. Another way is to use the compensating probe on a square-wave generator of known waveform while adjusting C2 for best possible waveform in the scope trace. When correct capacitance has been determined, the adjustable capacitor may be replaced with a fixed type varying by no more than 5 per cent. Such small capacitances can be measured by substitution in a tuned resonant circuit.

The oscilloscope may be used to best advantage for trouble shooting by making the first checks at the outputs and inputs of large sections of the receiver, rather than going through every separate stage and circuit. As an example, the first check might be at the signal input to the picture tube and the next one at

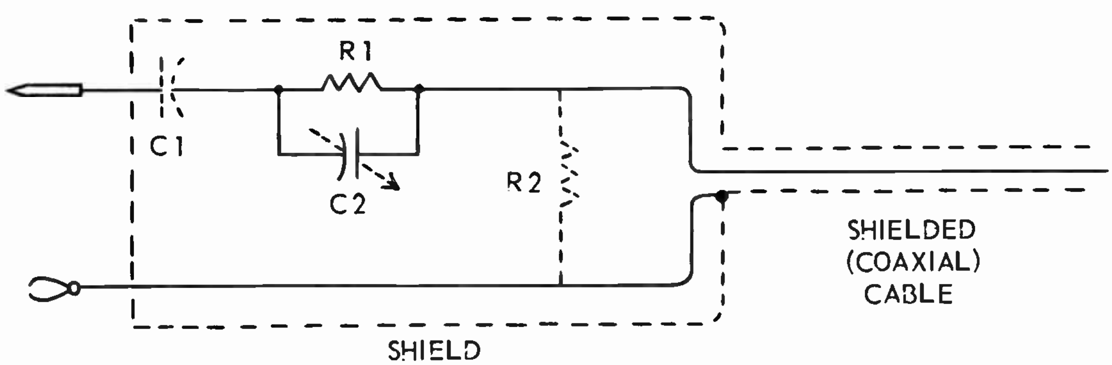


Fig. 96-4. A frequency-compensating probe for waveform tracing in sync and sweep sections.

the video detector load. If there is a signal at the detector and none at the picture tube, it is in order to check through the video amplifier in detail, but if there is no signal at the detector load, the third test may be at the plate of the first i-f amplifier. If there is no signal present, the trouble would be in either the first i-f amplifier or the tuner. If a signal is present, the probe must then be touched to each i-f amplifier stage going toward the detector.

When carrying out a series of tests while employing a transmitted signal, do not change channels if it can be avoided. Do not alter the receiver contrast control, but regulate trace heights by the vertical gain control of the scope. Do not alter the setting of the fine tuning control during the tests. If a picture appears, and can be synchronized by adjusting the receiver hold controls, the traces will be clearer and sharper if synchronization can be maintained by occasional readjustment of the receiver hold controls.

A transmitted signal is needed for all tests between tuner and picture signal input to the picture tube, also for tests through the sync section and as far as the inputs to vertical and horizontal sweep oscillators. This will include any automatic frequency control circuit for the horizontal sweep. A transmitted signal is not necessary when checking between the grids of the sweep oscillators and the sweep amplifiers, since the oscillators generate their own free-running frequencies without a signal. When working without a received signal in these latter sections, adding a signal (as by tuning the channel selector) will change the oscillator frequencies and require readjustment of the internal sweep frequency in the scope. The oscillator frequencies should be somewhat higher or faster with the signal than without it.

WAVEFORM ANALYSIS. It is especially desirable to observe signal waveforms in the sync and sweep sections of the television receiver. Here it is that the original composite signal, as it should appear at the output of the video amplifiers, undergoes separation of the sync pulses from picture signals, also clipping and limiting as well as changes into voltage pips and stepped pulses before reaching the sweep oscillators. Then we find sawtooth voltage and current waves and modifications of these waves between the oscillators and the deflection elements at the picture tube.

Receiver manufacturers often show typical waveforms in their instruction or service manuals, thus enabling the technician to make comparisons with waveforms actually obtained from receivers subject to various troubles in deflection. We shall examine waveforms existing in the sync and sweep circuits shown by Fig. 96-5.

Photographs were made from an oscilloscope having flat frequency response to 500 kilocycles. The vertical input is frequency compensated, and a frequency compensated probe was used. This particular probe reduced the vertical sensitivity from about 0.03 volt per inch to about 0.18 volt per inch, maximum. As explained in lessons dealing with oscilloscopes, photographs showing two vertical pulses are made with a sweep frequency of 30 cycles per second, which is half the field frequency, and those showing one vertical pulse are made with a 60-cycle sweep, which is the field frequency. Pictures showing two horizontal pulses are made with a sweep of 7,875 cycles, which is half the horizontal line frequency. To show a single horizontal pulse the sweep rate would be at line frequency, which is 15,750 cycles per second.

Referring to Fig. 96-5, note that the composite signal from the plate circuit of a video amplifier is applied to the cathode of a sync separator tube. The separated sync pulses are taken from the plate of the sync separator tube; then the signal goes through an inverter, from whose cathode it goes through an integrating filter to the first grid of a multivibrator type vertical oscillator. The signal goes from this oscillator to the vertical output amplifier, and through a transformer to the vertical coils in the deflecting yoke.

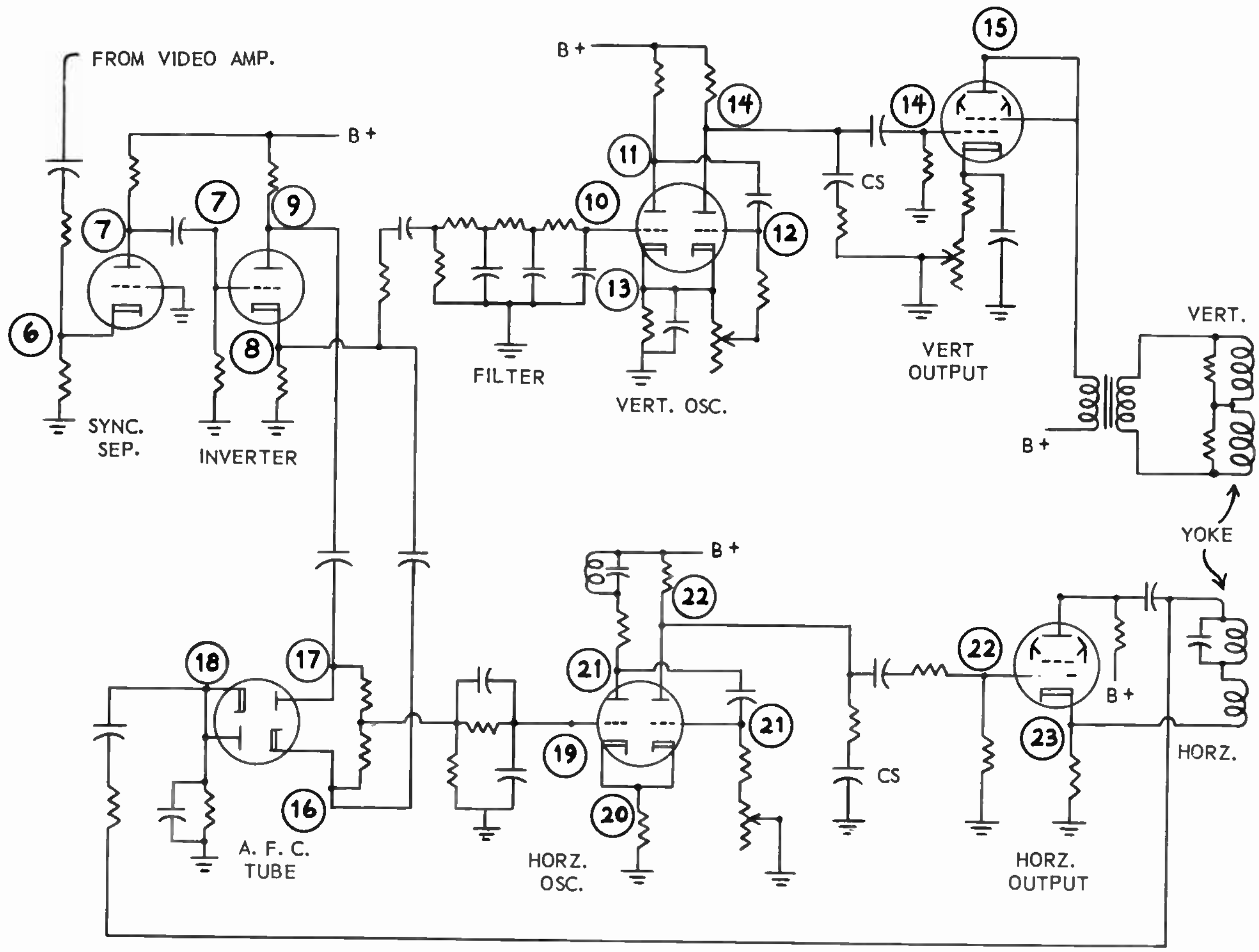
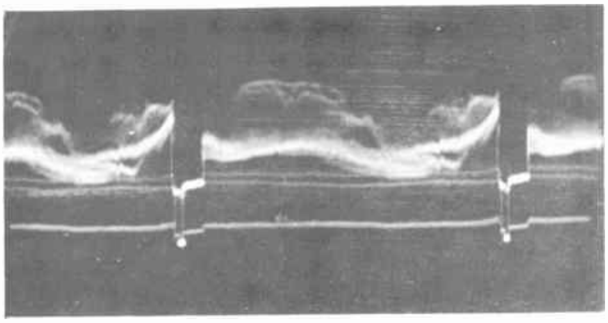


Fig. 96-5. Circuit diagram of the sync and sweep sections of a receiver for which waveforms are shown by following photographs.

Going back to the output of the sync inverter tube, you will see that signals of opposite polarity are taken from the plate and cathode, and are applied to a plate and cathode of the horizontal afc tube. To the other plate and cathode of this control tube come pulses at the actual horizontal line frequency, as taken from the horizontal deflecting yoke and the output of the horizontal sweep amplifier. Frequency control voltages go to the first grid of the horizontal multivibrator sweep oscillator, and from the output of this oscillator a sawtooth is fed to the grid of the horizontal output amplifier.

At the left in Fig. 6 is shown the composite television signal which comes from the video amplifier into the restorer cathode, as this signal appears with the scope sweep at 30 cycles per second to display two fields and two vertical sync pulses. At the right is shown the same signal with the scope sweep at 7,875 cycles, to display two horizontal sync pulses. Note that picture signals are positive, and sync pulses are



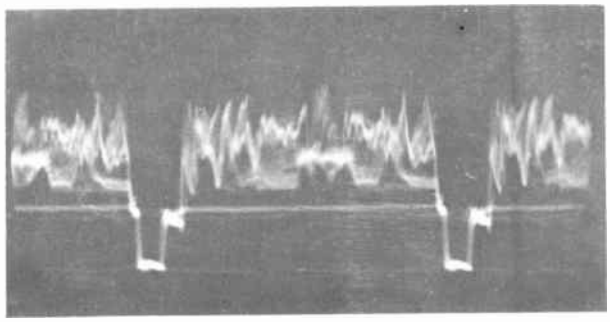
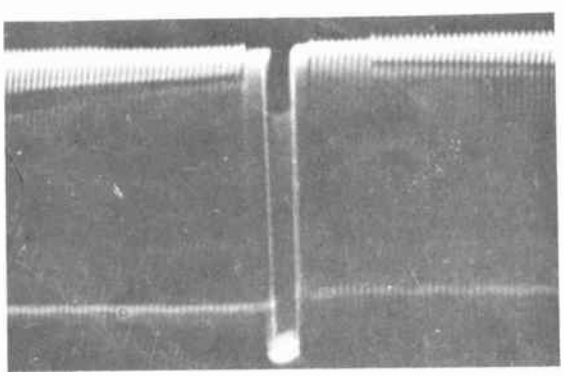


Fig. 96-6. The composite television signal from the video amplifier, as it appears at the restorer cathode. Vertical at left, horizontal at right.

negative. It is customary to show vertical and horizontal signals or pulses together when both are taken from the same point with only a change of scope sweep frequency, and with the probe remaining connected to the same place for both waveforms. For convenience in following these explanations the test points on the circuit diagram of Fig. 95-5 are numbered to correspond with the figures which show the trace or traces obtained at each point.

We go next to the plate of the restorer tube, point 7 on the diagram, whose signal waveform with vertical timing is shown at the left in Fig. 96-7, and with horizontal timing at the right. Note that the signal has been almost completely stripped of the picture elements to leave practically nothing more than the vertical nd horizontal sync pulses. The pulse polarity still is negative, because there is no inversion between hode and plate of the restorer tube.

The next check point is at the grid of the sync inverter. The vertical and horizontal waveforms obtained here are shown respectively at the left and right in Fig. 96-7. This one figure shows the waveforms at



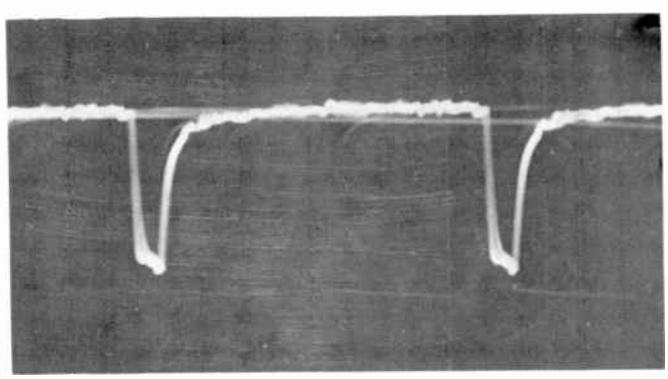
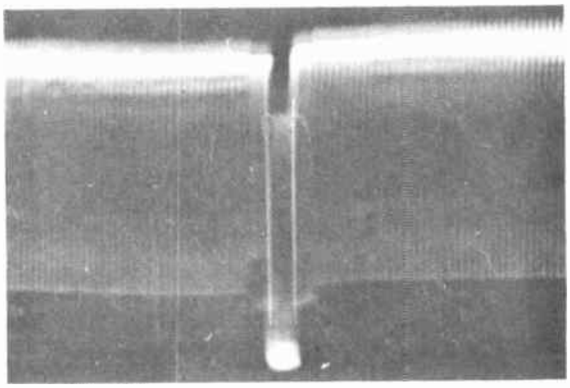


Fig. 96-7. Waveforms at the plate of the restorer tube. Vertical at left, horizontal at right. The same waveforms appear at the inverter grid.



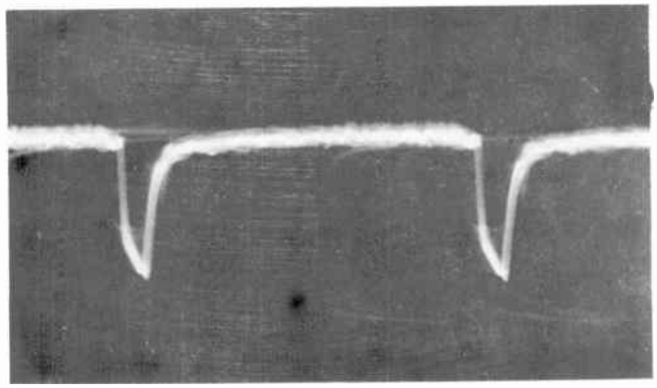
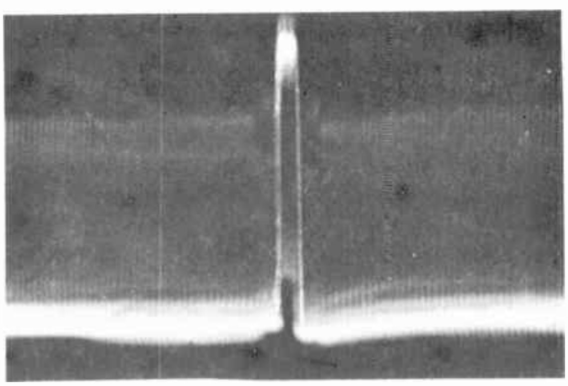


Fig. 96-8. Voltage waveforms at the inverter cathode. Vertical at left, horizontal at right.

both the plate of the restorer and at the grid of the inverter, because there is nothing between these two places except a coupling and blocking capacitor. Such a capacitor should cause no change of a waveform.

Now we move the probe to point 8, the cathode of the inverter, and pick up the vertical waveform pictured at the left in Fig. 96-8, and the horizontal waveform shown at the right. Pulse polarity still is negative, for polarity at the cathode of a triode is the same as at the grid. Upon comparing Figs. 7 and 8 it becomes apparent that there is no material change of waveform between the plate of the restorer and the cathode of the inverter, which is as it should be.

At the left in Fig. 96-9 is the vertical waveform picked up at the plate of the inverter, point 9 on the diagram, and at the right is the horizontal waveform from this same point. Still there is no material waveform change, but the polarity is inverted with respect to polarities at the grid and cathode of the same tube.



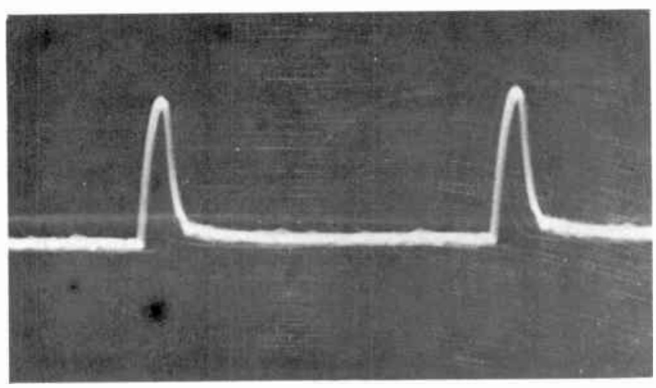


Fig. 96-9. Positive pulses at the inverter plate. Vertical at left, horizontal at right.

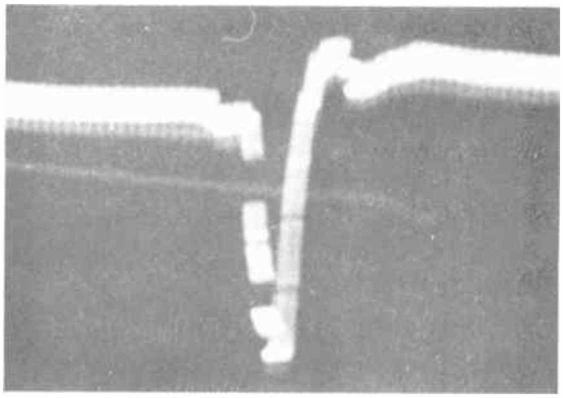


Fig. 96-10. The voltage wave for triggering at the grid of the vertical sweep oscillator.

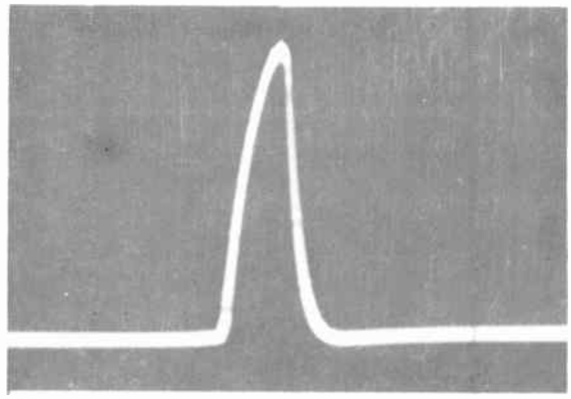


Fig. 96-11. This is the pulse at the plate of the first section of the vertical oscillator.

While proceeding through the sync section of this receiver from test point 6 to points 8 and 9 we have been interested in both vertical and horizontal waveforms, which pass together through this portion of the receiver circuits. Now we come to where the vertical and horizontal pulses are separated. Vertical pulses go through the integrating filter to the vertical oscillator, while horizontal pulses go to the automatic frequency control system for the horizontal oscillator.

We shall follow first through the vertical sweep section, and later come back to the horizontal sweep circuits. Fig. 96-10 shows the waveform picked up at the triggering grid, point 10, of the vertical oscillator after a group of serrated vertical sync pulses have come through the integrating filter and have been combined into a single pulse for triggering the oscillator. The effect of the original separate serrated pulses is evident in the broken appearance of the triggering pulse on the leading side. On the trailing side

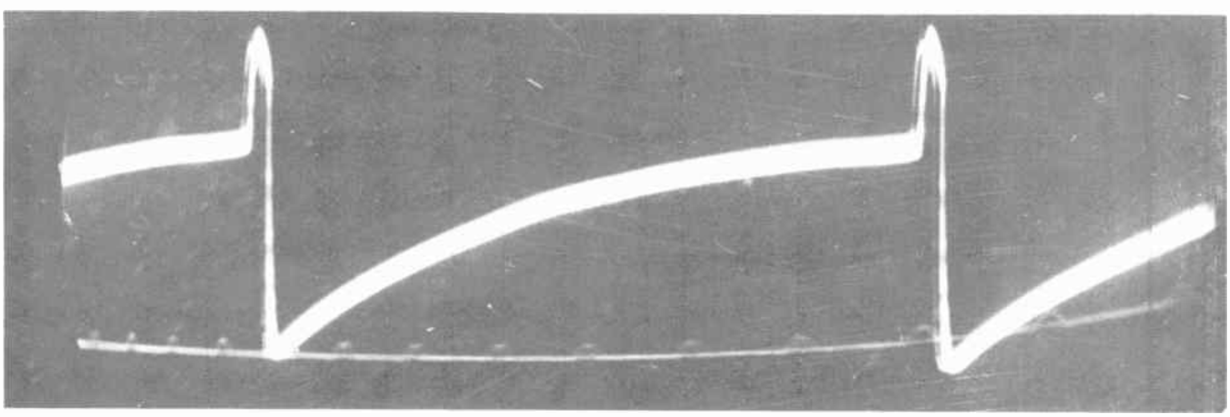


Fig. 96-12. Positive pulses from the first plate of the vertical oscillator cause this waveform at the grid of the second section.

9

is the sharp drop of voltage during the equalizing pulses which follow the serrated pulses in the television signal. Peak-to-peak voltage of the triggering waveform in Fig. 96-10 was measured as only about 0.3 volt, which is sufficient to control the multivibrator type of oscillator.

We proceed next to the plate of the first section of the vertical oscillator, for which the waveform is shown by Fig. 11. This is a very strong pulse; it measured about 60 peak-to-peak volts. The great difference between peak-to-peak voltages at grid and plate, points 10 and 11, does not mean that this section of the tube is amplifying the grid voltage something like 200 times, it means only that the strong plate voltage developed by the oscillator section is controlled, in time, by the weak voltage pulse at the grid. Note that the pulse at point 11 is of positive polarity. It will be used on the grid of the second section of the oscillator to make that second section conductive, and thus to allow discharge of the sawtooth capacitor.

What happens at the grid of the second section of the vertical oscillator is shown by Fig. 96-12. There is a gradual rise of voltage which ends on the sharp positive peak which is due to the positive pulse coming to the grid from the plate of the first oscillator section. That is, the positive pulse of Fig. 96-11 appears in Fig. 96-12 as the narrow peak at the termination of the gradual rise of grid voltage. The pulse comes through the coupling and blocking capacitor connected between points 11 and 12, yet the waveforms on opposite sides of this capacitor are very different. This is a case where we have different waveforms on the two sides of a coupling capacitor, but the difference is due to voltages added on the grid side of the oscillator circuit.

Fig. 96-13 shows the voltage pulse at the two cathodes of the vertical oscillator, point 13 in the circuit diagram. The cathode circuit, which is common to both sections of the oscillator, allows coupling between these sections. It is the voltage pulse at the cathode of the first section of the oscillator that accounts for the strong voltage peak at the plate, which was shown by Fig. 96-11. We really have what amounts to cathode input of a signal pulse for the first section of the oscillator, with the grid of this section acting only to control the timing.

The final output of the vertical oscillator, from its second plate at point 14, is shown by Fig. 96-14.

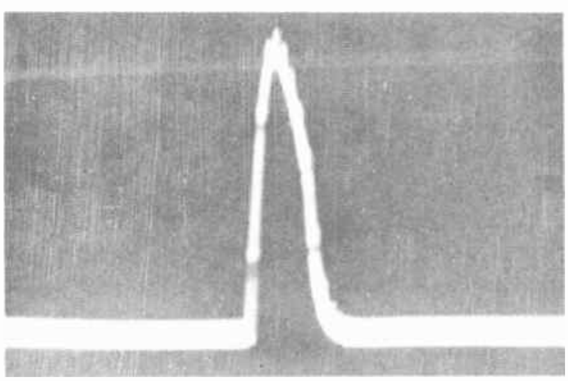


Fig. 96-13. This voltage wave, at the cathodes of the vertical oscillator, shows the feedback pulse between the two sections.

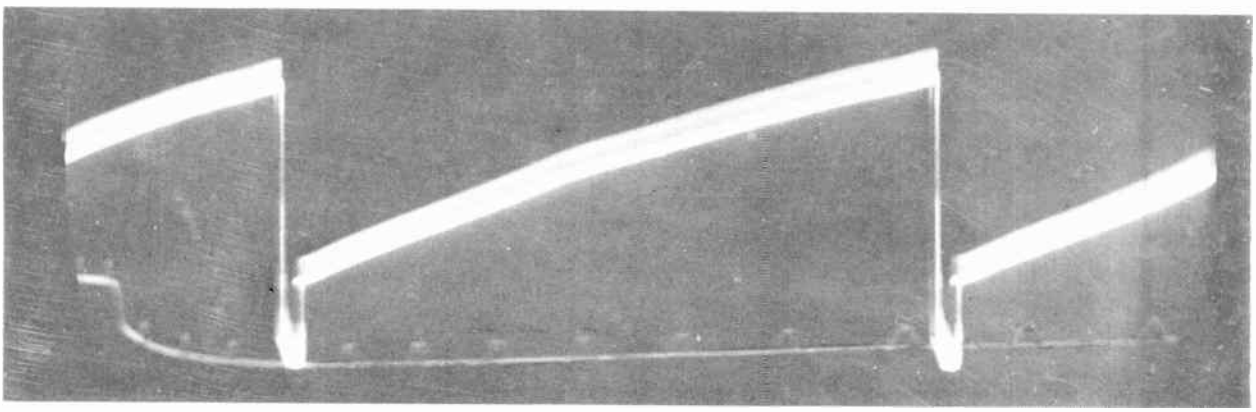


Fig. 96-14. The voltage output of the vertical oscillator, and voltage input at the grid of the vertical output amplifier.

Compare this waveform with the one of Fig. 96-12, which exists at the second grid. There has been inversion of polarity, as would be expected. There has also been a decided straightening of the line showing a gradual rise of voltage. This means that we have come much closer to a linear or well-formed sawtooth voltage at the output plate of the oscillator. The voltage wave of Fig. 96-14 is the voltage on sawtooth capacitor Cs of Fig. 96-5 as well as the voltage on the output plate of the vertical oscillator. The negative peaks added to the sawtooth wave are the result of time constant effects in the capacitor circuit. All this should remind you of many things which were studied in earlier lessons.

The voltage wave at the sawtooth capacitor and at the output plate of the oscillator is applied through a coupling and blocking capacitor to the grid of the vertical output amplifier, which is marked 14 on the diagram. The voltage at this grid is of the same form as at the output plate of the vertical oscillator, and may be represented by Fig. 96-14.

Our final check point in the vertical sweep section is at the plate of the vertical output amplifier, point 15 on the diagram of Fig. 96-5. Here the waveform is as pictured in Fig. 96-15. Compared with the waveform at the grid there has been inversion of polarity, but very little change otherwise. There appears to be somewhat more peaking. This is because we are now into a circuit containing a great deal of inductance and inductive reactance in the transformer which couples the output amplifier to the vertical deflecting coils in the yoke of the picture tube. You will recall that in an inductive circuit the voltage wave tends to lose much of its sawtooth slope and to become more nearly a series of brief peaks. In the particnlar case being examined here we have inductance from the output transformer and plate resistance in the amplifier tube, consequently obtain a waveform in which a sawtooth is combined with voltage peaks.

When making tests, the results of which are being shown by waveforms, peak-to-peak voltage at the grid of the vertical output amplifier, point 14, measured about 43 volts. Measurement at the amplifier plate, point 15, showed about 370 peak-to-peak volts. Methods of measuring peak-to-peak voltages are explained in the lesson which describes tests with the oscilloscope. If the vertical gain controls of the scope are operated to make all traces of about the same height, the changes of the gain control which are required

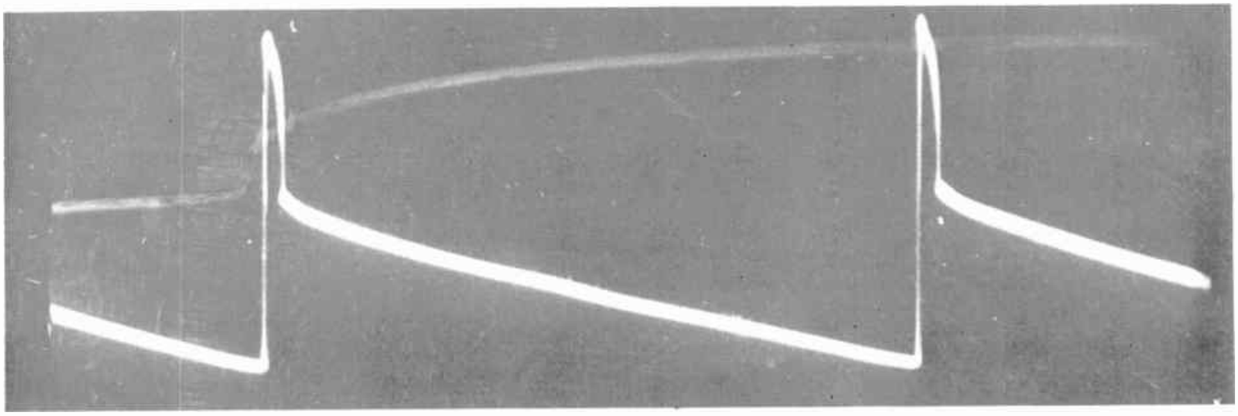


Fig. 96-15. The waveform at the plate of the vertical output amplifier.

when going from one test point to another will give you a fair idea of relative voltages. When you know that there should be an increase or decrease of peak-to-peak voltage between two points, and the required change of gain controls indicates otherwise, there usually is trouble between those two points.

TRACING IN THE HORIZONTAL SWEEP SECTION. To trace the progress of sync and sawtooth waveforms through the horizontal sweep section we must go back to the inverter tube of Fig. 96-5. From the cathode of this tube, point 8, sync pulses are taken to one of the cathodes in the control tube for the horizontal sweep, point 16, and there appear as shown by Fig. 96-16. Compare this waveform with the one at the left in Fig. 96-8, which is taken at the inverter cathode. There is quite a difference between the two, because at the afc tube the voltage wave is being affected by other factors.

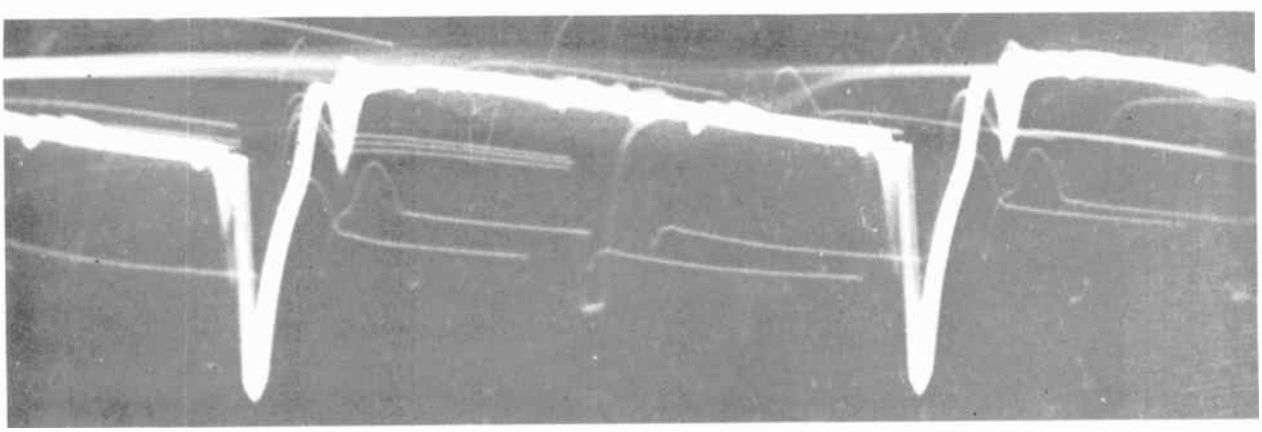


Fig. 96-16. This waveform appears at the signal cathode of the horizontal afc tube.

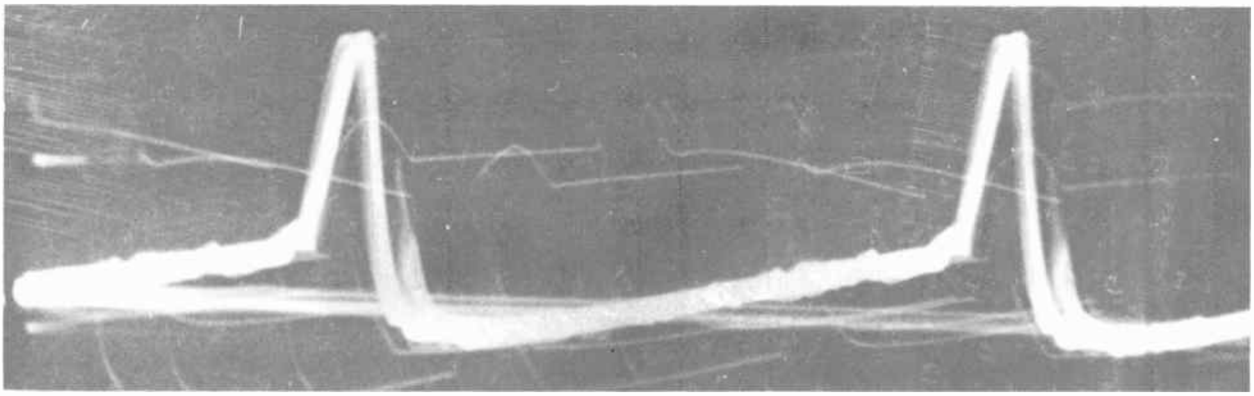


Fig. 96-17. This is the waveform at the signal plate of the horizontal afc tube.

From the plate of the inverter, at 9, a sync pulse is taken to one of the plates of the afc tube, at 17. The waveform at point 17 is shown by Fig. 96-17. You should compare this afc waveform with the one at the plate of the inverter, which is shown at the right in Fig. 96-9. Again there is considerable difference between the waveforms, because additional factors are present at the control tube.

The remaining plate and cathode of the afc tube are tied together at point 18. This point connects through a series capacitor and resistor to one of the horizontal deflecting coils in the yoke, and through one more capacitor back to the plate of the horizontal sweep amplifier or output amplifier. The voltage at the output amplifier plate and at the yoke coil consists of a series of narrow, sharp peaks. This is the kind of voltage wave which accompanies a sawtooth current wave in the deflecting coils. The capacitor

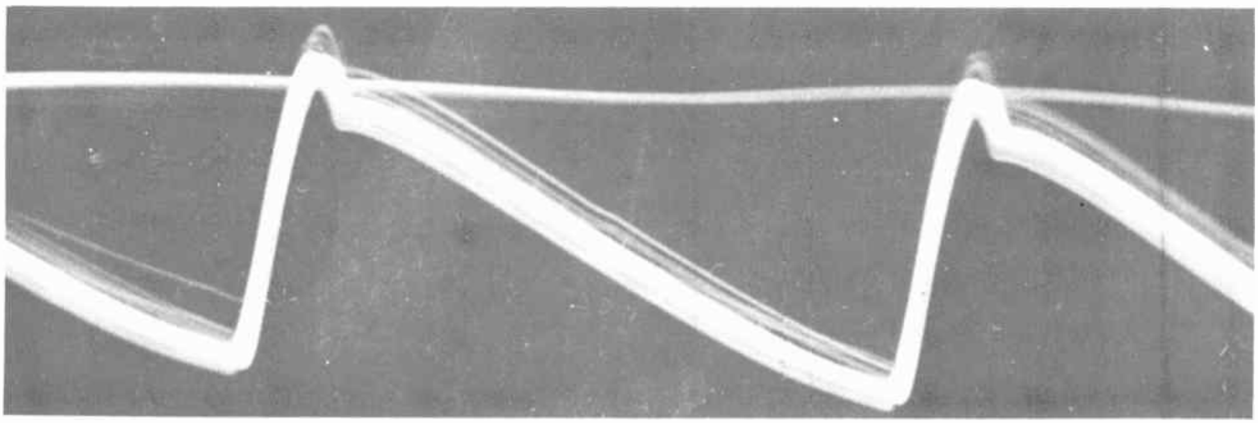


Fig. 96-18. The voltage wave which is obtained from the output of the horizontal sweep section and applied to the feedback plate and cathode of the horizontal afc tube.

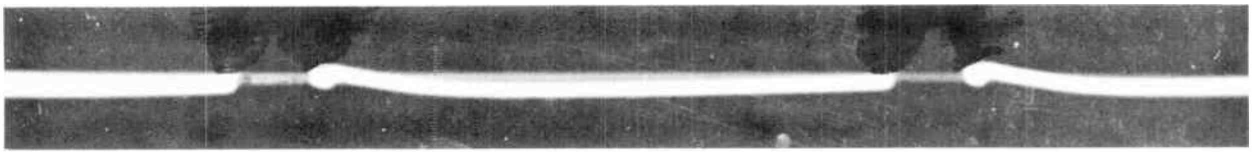


Fig. 96-19. This is the waveform at the grid of the first section of the horizontal multivibrator oscillator.

and resistor between the yoke coil connection and point 18 change the voltage wave to a sawtooth form. This sawtooth voltage which is applied to the afc tube is pictured by Fig. 96-18.

Sync pulses coming from the inverter combine in the afc tube with the sawtooth voltage obtained from the horizontal output amplifier. Timing of the pulses depends on the received television signal, while timing of the sawtooth voltage depends on the frequency at which the sweep system of the receiver actually is operating. The result is a frequency-correcting DC voltage, delivered to the first grid or triggering grid of the horizontal oscillator at point 19. This voltage, as observed for correction conditions at one certain time, is shown by Fig. 96-19. This, remember, is the voltage which controls the timing or the operating frequency of the horizontal oscillator, and which keeps pulling the oscillator frequency into synchronization with the received horizontal sync pulses.

The cathodes of the two sections of the horizontal oscillator are tied together and connected through a resistor to ground. This resistor provides coupling between the two sections. The voltage wave observed at the oscillator cathodes is shown by Fig. 96-20. It is of about 14 peak-to-peak volts. At the plate of the first oscillator section, point 21, appears a very similar waveform which is shown by Fig. 96-21. The voltage at this plate is about 42, peak-to-peak.

Voltage from the plate of the first oscillator section goes to the grid of the second section through a

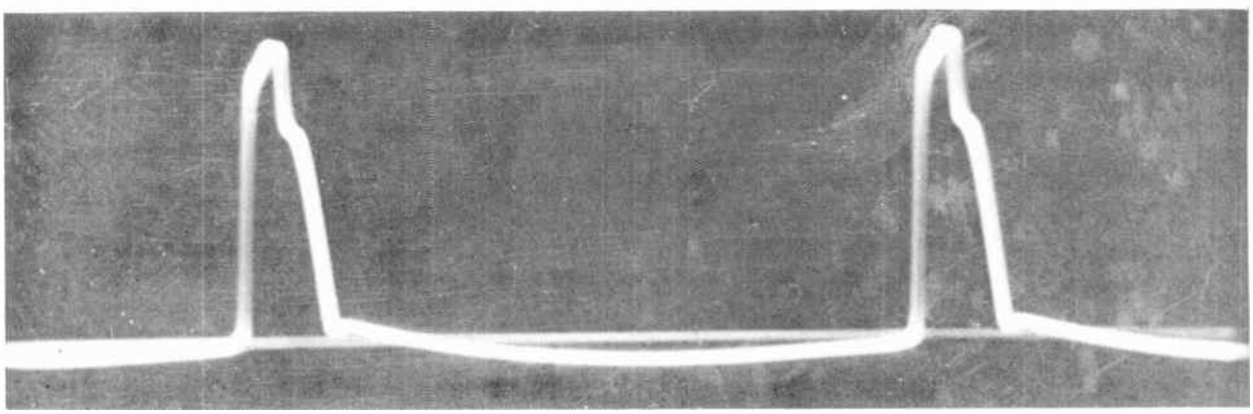


Fig. 96-20. This voltage wave, at the cathodes of the horizontal multivibrator, is the feedback between sections.

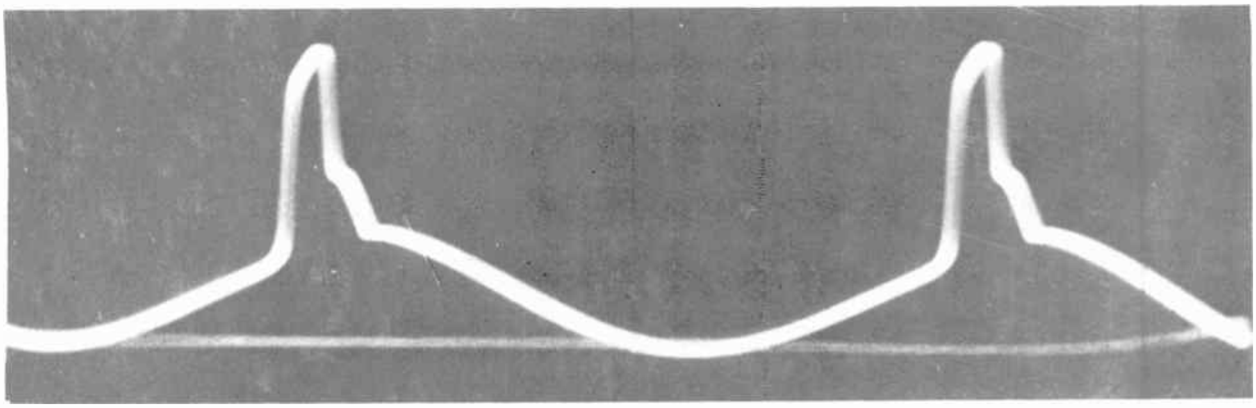


Fig. 96-21. Waveform at the plate of the first section of the horizontal multivibrator and at the grid of the second section.

coupling and blocking capacitor, and appears at the grid, also numbered 21 on the diagram, in the same form as at the plate and of practically the same peak-to-peak voltage. Probably you have noticed that action in the horizontal multivibrator oscillator is not quite like that in the vertical multivibrator oscillator. This is because the horizontal oscillator is a straight cathode-coupled type, with coupling through only a resistor between cathodes and ground, while there is a modified form of cathode coupling on the vertical oscillator.

At the output plate of the horizontal oscillator, point 22, we find the voltage waveform shown by Fig. 96-22. There has been inversion of polarity between the grid and plate of the output section. Peak-to-

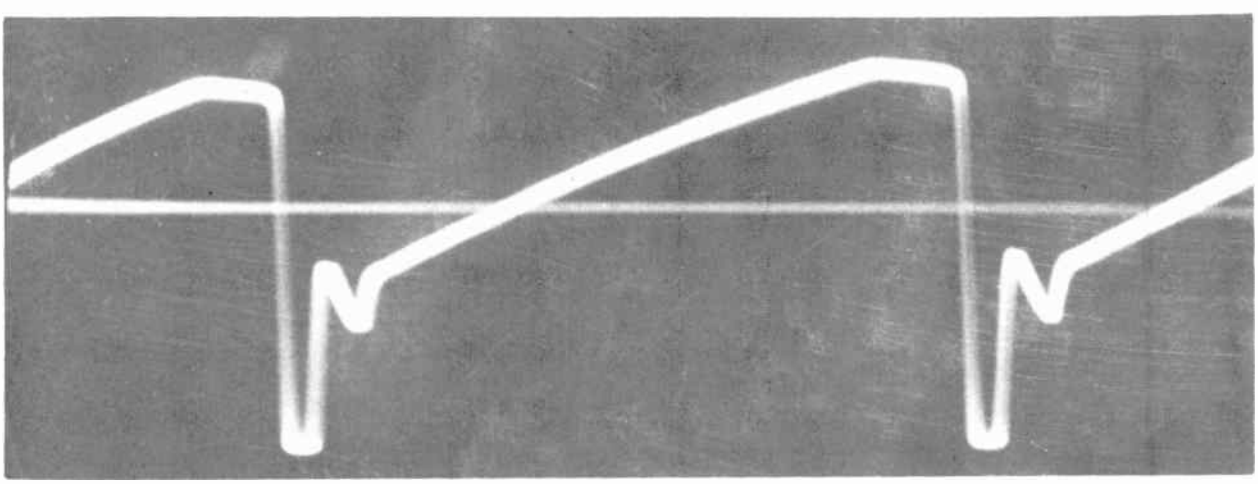


Fig. 96-22. This waveform appears at the output plate of the horizontal multivibrator, also at the grid of the horizontal output amplifier.

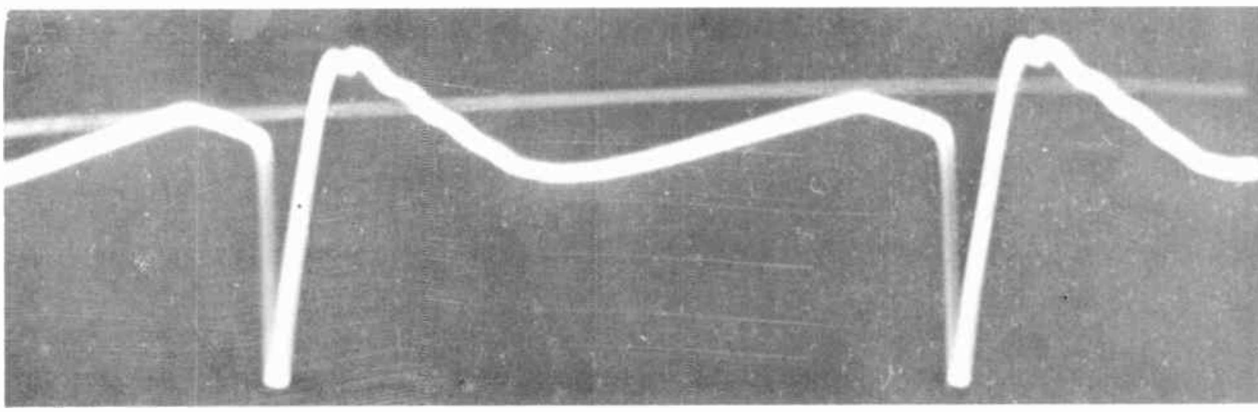


Fig. 96-23. The voltage waveform at the cathode of the horizontal output amplifier.

peak voltage at the output plate is about 95, as compared with about 42 peak-to-peak volts at the grid of this section. This same voltage waveform appears on the connection to the sawtooth capacitor, Cs, and at the grid of the horizontal output amplifier. This grid is numbered 22, and the waveform is shown by Fig. 96-22.

The waveform observed at the cathode of the horizontal output amplifier, point 23, is shown by Fig. 96-23. This is also the voltage wave which would be picked up at one end of the horizontal deflecting coils in the yoke, to which there is a direct connection from the amplifier cathode. The receiver whose sweep circuits have been shown operates with an r-f type high-voltage power supply rather than the more common flyback power supply. This accounts for the direct connection of the horizontal output amplifier to the deflecting coils, instead of having a horizontal output transformer.

6 We are not showing the voltage waveform at the plate of the horizontal output amplifier, because you should not attempt such an observation without a high-voltage capacitor-type divider on the oscilloscope. Peak-to-peak voltage at the horizontal output amplifier plate of the receiver examined measured about 1,940 volts, which is far too strong a pulse to put into the oscilloscope. The waveform consisted of narrow, sharp positive peaks, with the voltage almost flat between successive peaks. At the plate of the horizontal oscillator operating with a flyback high-voltage power supply the peak-to-peak voltage often is on the order of 5,000 to 6,000 volts.

In the receiver with which we have been working it is safe to observe the waveform at the plate of the vertical output amplifier, because at this point (15) we have only about 370 peak-to-peak volts. In other receivers, especially when the picture tube is of the short-neck wide-angle variety, the peak-to-peak voltage at the vertical output plate is likely to be between 1,500 and 2,000 volts. This could be ruinous when applied to the vertical input of any ordinary oscilloscope.

It is unwise to attempt measuring voltages at the plates of sweep output amplifiers with a VTVM or with any other type of meter not fitted with some high-voltage type of divider. Voltage pulses may be so narrow

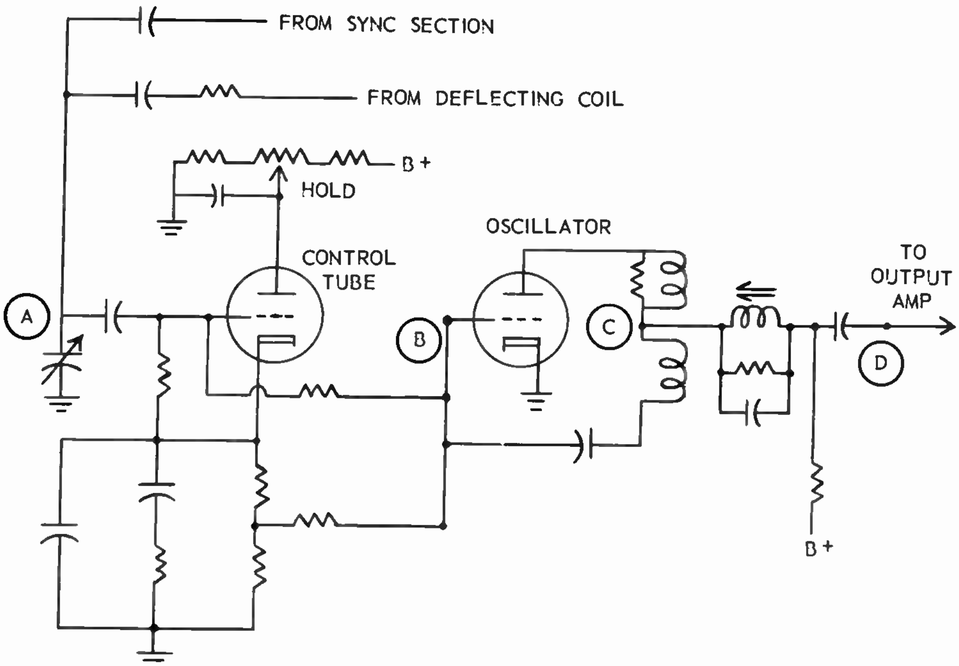


Fig. 96-24. An automatic frequency control system in which a triode varies the grid voltage of a blocking oscillator.

or brief as to have very little average or effective value, and they may hardly move the pointer of the meter. Yet the peak voltages easily may burn out the instrument by puncturing the insulation.

OTHER WAVEFORMS. We have traced through the sync and sweep circuits of one particular receiver, and have observed the waveforms at every point of importance. In a different receiver, employing different kinds of circuits and controls, the waveforms at many points might be quite unlike the ones illustrated in this lesson. It is a great convenience to have available a set of normal waveforms for the receiver with which you are working. Such waveforms, at least for the sync and sweep sections, are photographically reproduced in many service manuals put out by set manufacturers, or are drawn on the service diagrams near the points from which the traces are obtained.

It is, however, not at all essential to have reference to normal waveforms in order to do effective trouble shooting with the oscilloscope. With your present knowledge of television circuits and their behavior it is quite easy to recognize any waveforms which are decidedly wrong in any type of sync system. This is because the input should show a practically complete composite television signal, and the output, or outputs to sweep sections, should consist of pulses suitable for triggering a vertical oscillator and for application to afc systems for horizontal oscillators.

Most receivers have either multivibrator or blocking type sweep oscillators. Waveforms for typical multivibrator oscillators have been shown in this lesson, also in lessons devoted especially to this type

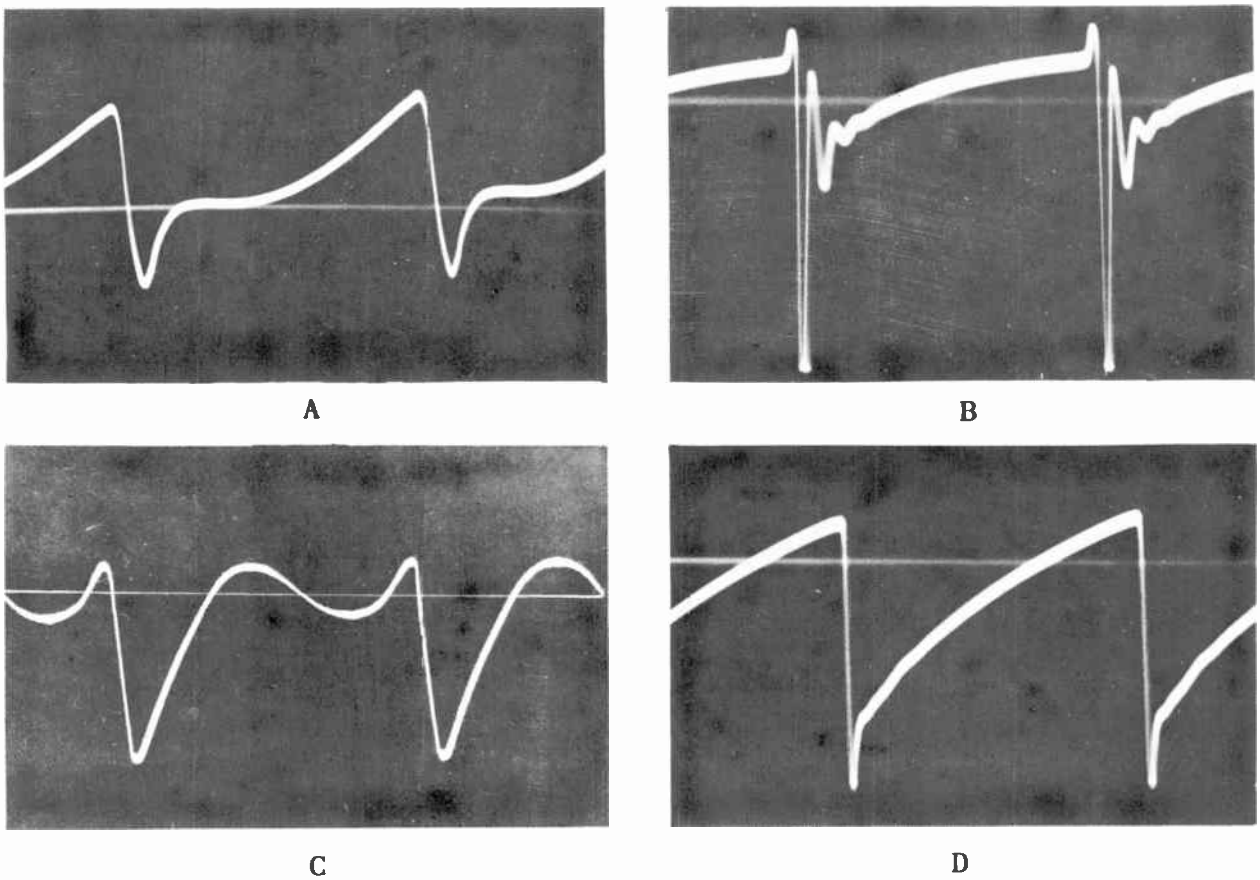


Fig. 96-25. Typical waveforms for the afc system of Fig. 96-24. Letters on the waveforms appear on the circuit diagram at the points of observation.

of oscillator. Waveforms which should appear at blocking oscillators are shown and discussed in lessons dealing with blocking oscillators.

Between the sweep oscillators and the output amplifiers we must find sawtooth waves. Although these waves may carry various added peaks and other characteristics, the essential sawtooth form is easily recognized, and its absence is easily noted.

It is in the automatic frequency control systems for horizontal sweep oscillators, and sometimes for vertical sweep oscillators, in which we are likely to find the greatest diversity of waveforms. But even here you need not be at a loss provided you remember that all these afc systems act to combine received sync pulses with some voltage at the actual operating frequency of the sweep system, and from the combination they produce a correction voltage which is applied to the oscillator. Waveforms that should be found in most of the commonly employed afc systems are shown in lessons on automatic control of sweep frequency.

Each afc system has characteristic waveforms which you will soon learn to look for and to recognize, even when there are minor differences due to variations of parts values and layouts. As an example, many afc systems employ a triode control tube in whose cathode circuit a filter controls the voltage on the oscillator grid, and thus controls the oscillator frequency. One way of using this general type of control is shown by Fig. 96-24. Characteristic waveforms are shown by Fig. 96-25. The waveform at A, or something very similar, should appear at the grid of the control tube. The waveform at B is taken from the grid of the oscillator, and is typical of waveforms at the grids of all blocking oscillators. At C is the waveform at the oscillator output, taken from between the plate coil and the grid feedback coil. The waveform at the grid of the following sweep amplifier is shown at D. The shape of this latter wave depends to some extent on whether or not there is a filter between the oscillator transformer and the amplifier grid, but there is not so much variation that you cannot recognize any serious fault.

FOR GRADING.

TELEVISION

LESSON NO. 97

TUBE TESTING AND SUBSTITUTION

Training Your Mind...
C. T. I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON NO. 97

TUBE TESTING AND SUBSTITUTION

Tubes are one of the few parts in television and radio receivers which have only a limited life even when not subjected to abuse. As a consequence we find that tube failures of one kind or another are the cause for more service calls than failures in all other components combined. A burned out tube is easily located, because it remains relatively cool after a set has been turned on for five or ten minutes, while tubes with live heaters become too hot to touch with comfort. If a burned out tube has a glass envelope it may be possible to see that the heater is not glowing, and thus locate the burned out unit.

It is the tubes which are not burned out, but are otherwise faulty, which cause the real difficulties in servicing. When tests described in preceding lessons have indicated that trouble probably is in some certain section of the receiver, and when there remains some doubt as to exactly where the fault may lie, it is good practice to commence by temporarily replacing the tubes in that section before making other checks.

Always there is the possibility that more than one tube may be defective. For this reason it is desirable to replace all the tubes in a section at the same time, rather than one by one. Then the new tubes can be taken out one at a time, and the original ones put back, until this brings back the trouble or until all the original tubes are back in place with the trouble uncorrected. Obviously, if putting back some certain original tube brings back the trouble, that tube is defective.

A tube which causes trouble in one receiver often will operate satisfactorily in another receiver of different type. It is true also that a tube which gives trouble in one position may be satisfactory at some other position in the same receiver. For instance, a tube causing trouble in the i-f amplifier may work well in the sync section. Such switching of positions usually works out only when the change is from a high-frequency to a low-frequency circuit, for the tube defect is quite likely to be one involving interelectrode capacitances or lack of transconductance.

When trying new or substitute tubes in the high-frequency circuits of tuners, or in video amplifiers or sound i-f amplifiers, you must remember that tube capacitance helps tune the circuits to resonance or helps in peaking the response. Any new tube probably will have internal capacitances slightly different from those in the original tube, even when both are of the same type. Then the substitution will alter the circuit tuning and performance may be worse than before. After retuning or aligning the grid circuit and possibly the plate circuit too, the new tube may bring about a great improvement. When putting a different tube in a tuned circuit you may align that circuit to obtain the best possible performance.

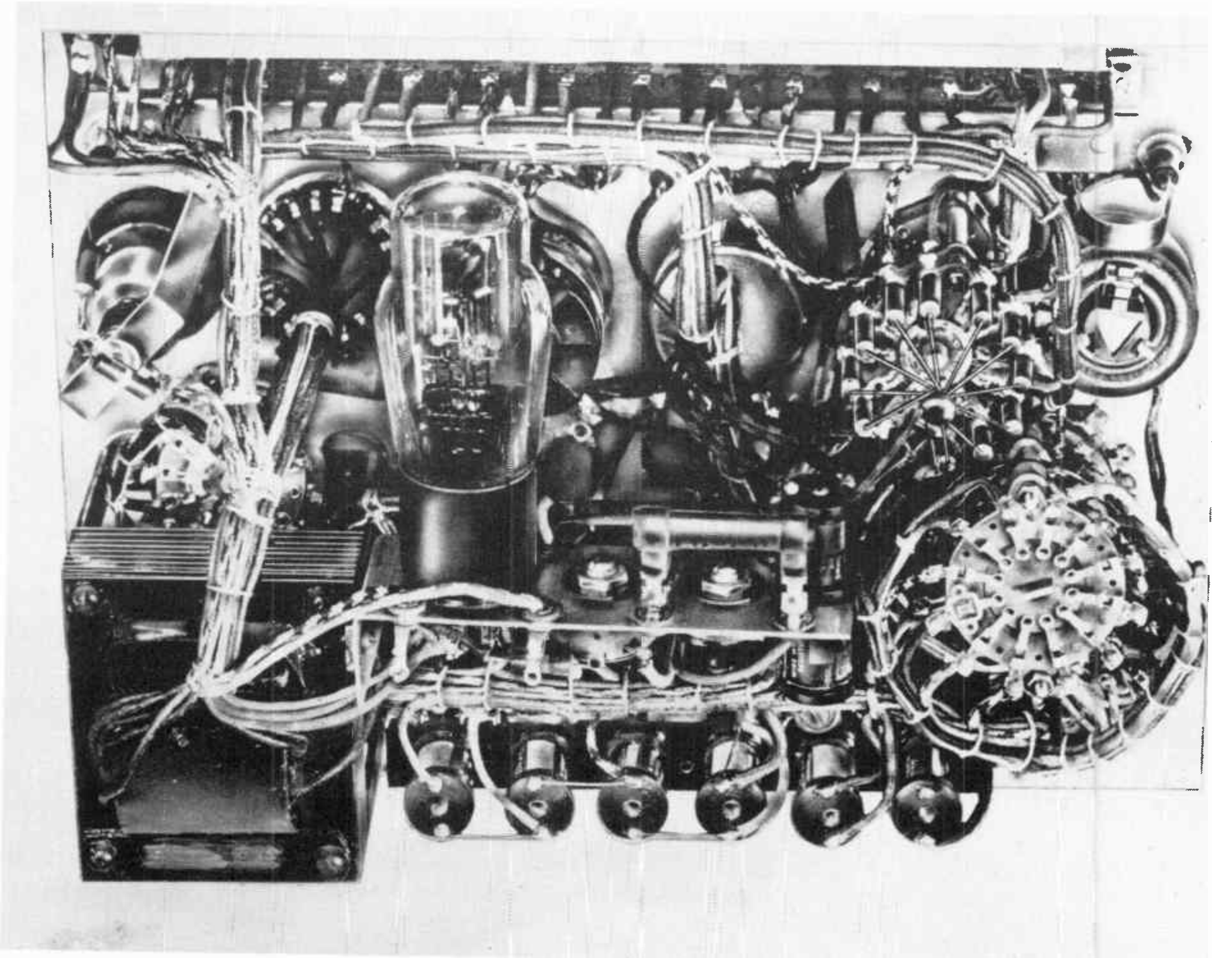


Fig. 97-1. The interior of a tube tester contains many control switches and much intricate wiring.

It is unfortunate that service types of tube testers do not always show a tube as faulty even though the tube fails to perform satisfactorily in a television receiver. The tester may read good or satisfactory on some tubes which give trouble in critical circuits. Therefore, even when all tubes from a section or from the entire receiver give a good reading, one or more of them may be the cause for trouble. It is true, however, that when a tester shows a tube as bad you may be almost certain that the tube should be replaced.

No tube tests which are possible with the usual facilities of a service shop are so dependable as substitution of new tubes in a receiver which is in trouble. Instead of using brand new tubes it is just as well to use those which are known to perform satisfactorily in a generally similar circuit of some other receiver. A tube which has operated well in a similar circuit may give more dependable clues to trouble than a new one right out of the carton. Some new tubes may fail to give satisfaction in certain circuits, and oftentimes it is necessary to try several before finding one that is entirely satisfactory.

Although tube testers are not infallible for television servicing, they are of great help in locating tubes which positively are defective. This leaves only relatively few to be checked by substitution. In following pages we shall examine the principles of some of the more common test methods employed in tube testers.

EMISSION TESTING. In the majority of tube testers the indications depend on the ability of the cathode to emit electrons at a certain minimum rate when test voltages are applied. The elementary principle of emission testing is shown by the circuit arrangement of Fig. 97-2.

When the tube contains more than a cathode and a plate as its elements, all except the cathode are connected together and to one side of the test circuit, with the cathode to the other side. The tube is in series with a d-c current meter, several tapped or otherwise adjustable resistors, and the secondary of a transformer whose primary is on the a-c line. The transformer supplies something like 30 rms volts to the tube and the test circuit. The one-way conductivity of the tube rectifies the alternating voltage and allows pulsating direct current to flow in the meter.

The meter is provided with an adjustable shunt resistor. This shunt is adjusted according to the type of tube being tested, so that normal emission current or a greater current causes the meter to indicate a satisfactory tube, while any current less than a predetermined minimum causes an unsatisfactory reading.

The selected series resistance limits the rectified current in accordance with the class of tube being tested. Minimum current is allowed for diodes such as employed for detectors and demodulators. Somewhat greater current is allowed for battery-type and other tubes having a small normal emission current. Test currents may differ also for voltage amplifiers, power amplifiers, and rectifiers.

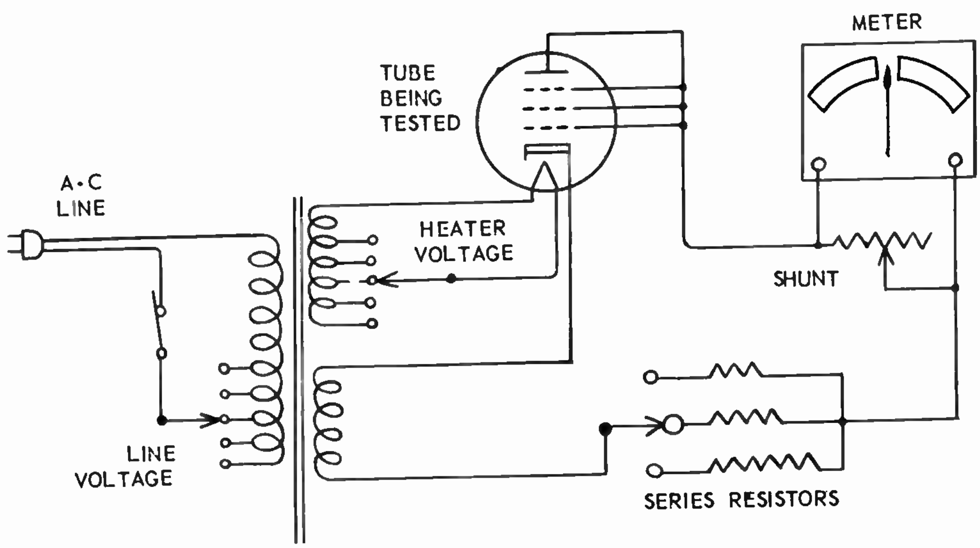


Fig. 97-2. The essential circuits for an emission type of tube tester.

The heater or filament winding of the transformer has taps, allowing selection of the correct voltage. The transformer primary has taps which compensate for line voltage variation and also for transformer voltage regulation, as different test currents are taken from the secondaries.

Any emission current greater than the predetermined minimum is equally satisfactory. Emission in tubes of the same type, but of different makes may vary greatly without affecting their performance in receiver circuits. Unusually great emission does not indicate a superior tube, in fact it may indicate a gassy one. The low limit of emission usually is that causing mutual conductance of amplifiers to drop to 70 per cent of the average in good tubes.

Abnormally small emission current usually means that an amplifier will have low mutual conductance. This is because mutual conductance is equal to the ratio of amplification factor to plate resistance. Low emission increases the plate resistance, and this lowers the mutual conductance. An emission test is based on the assumption that cathode material is equally active everywhere on its surface, and is everywhere equally controlled by the grid. In a tube which has been abused, the cathode material may have excessive emission from one or more spots, while the remainder of the surface is deficient. Total emission may be normal or greater, but the tube will not be satisfactory in a circuit employing grid control.

MUTUAL CONDUCTANCE TESTING. Mutual conductance or grid-plate transconductance may be defined as the ratio of amplification factor to plate resistance, or to the quotient of dividing the amplification

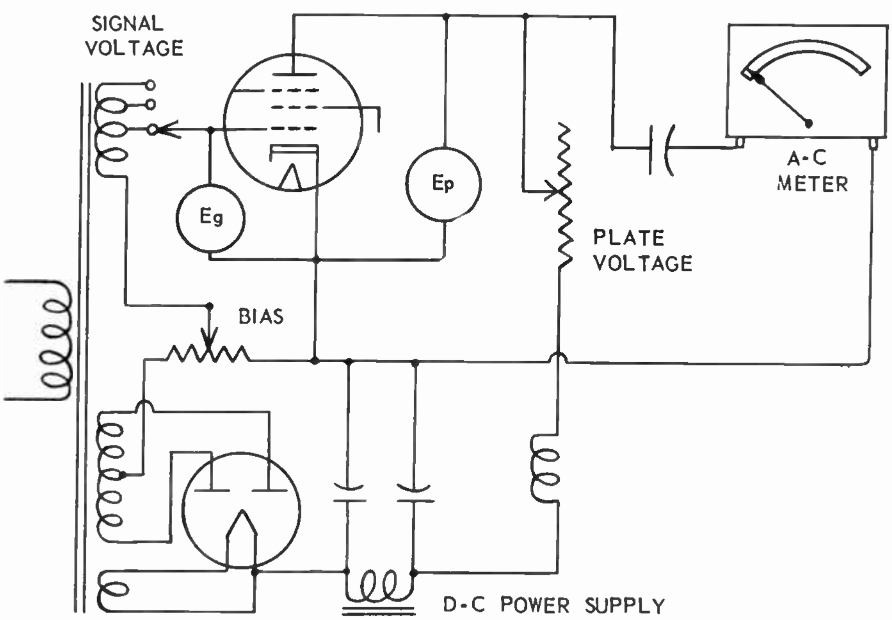


Fig. 97-3. Simplified circuits for measuring mutual conductance.

factor by the number of ohms plate resistance. This gives mutual conductance in mhos. The usual unit is the micromho. Mutual conductance in micromhos is 1,000,000 times its value in mhos.

A more useful definition says that mutual conductance in micromhos is equal to the number of microamperes by which plate current is changed by a change of one volt in grid potential. This is the same as saying that the number of micromhos is 1,000 times the number of milliamperes change of plate current per volt change of grid potential. Test potential applied to the grid is alternating, and only the alternating component of the plate current is measured. The measurement may be with a dynamometer type of meter or with a rectifier meter in series with a large capacitor to block the direct plate current and voltage.

An elementary circuit for measuring mutual conductance is shown by Fig. 97-3. Adjustable d-c voltages are applied to the plate, screen, and other elements as required. Only the adjustment for plate voltage is shown. Additional meters may be provided for measuring and correctly adjusting each of the element voltages. The grid usually is operated from a line-frequency alternating voltage, with the grid return to an ad-

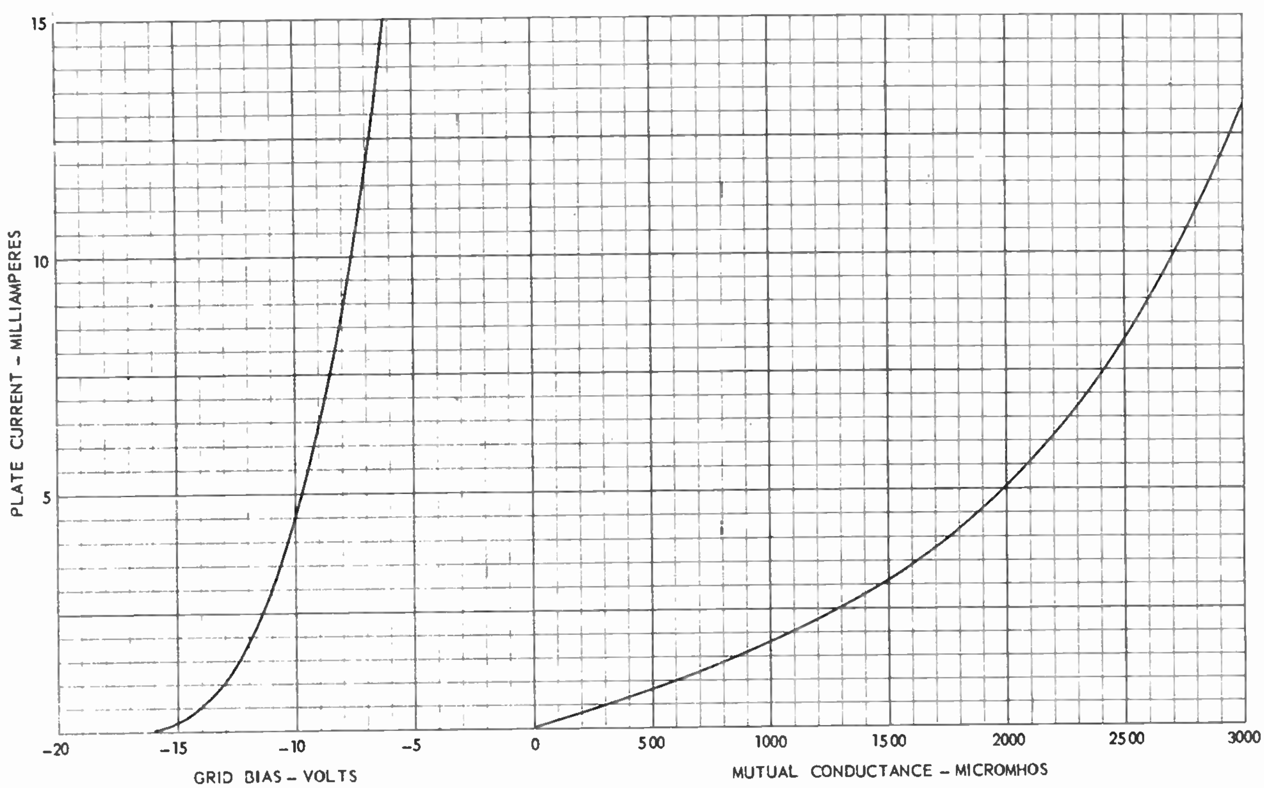


Fig. 97-4. Typical relations between grid bias and plate current (left) and between plate current and mutual conductance (right).

justable negative bias. The alternating component of plate current goes to the meter which is graduated in values of mutual conductance. Usually there is a separate mutual conductance scale for each selected grid signal voltage.

There is no such thing as a standard mutual conductance for any given tube. Instead there is a wide range of possible values which depend on plate voltage, screen voltage, and grid bias. Fig. 97-4 shows how mutual conductance changes with changes of plate current which result from variations of grid bias when plate voltage is held at a constant value for a typical voltage amplifier tube.

The curve at the left in Fig. 97-4 shows the effect of grid bias voltage on plate current. The right-hand curve shows how mutual conductance will vary as the plate current is changed. Were a tube tester to provide element voltages, allowing plate current as shown by the left-hand curve, the correct average mutual conductance would be that corresponding to this plate current.

A mutual conductance tester is calibrated for values of average mutual conductance, corresponding to operating voltages provided by the tester. These mutual conductances may or may not be the same as listed in tables of typical operating conditions for the tube being tested, but they are correct for the particular testing instrument.

Satisfactory mutual conductance usually is considered to be anything greater than 70 per cent of the average value for the type of tube tested. Mutual conductance is affected very little by differences in emission, provided the emission remains above a certain minimum. Were the plate current during a test to be much smaller than for typical receiver operation, the mutual conductance might be satisfactory for the test, but deficient when greater plate current flows in the tube.

Test voltage values for mutual conductance measurement are critical. As an example, Fig. 97-4 shows mutual conductance of about 2,570 micromhos with negative grid bias of 8 volts. Were the test bias to be made 9 volts negative, the mutual conductance would drop to about 2,250 micromhos. Variations of plate voltage for triodes and of screen voltage for pentodes can cause large variations of mutual conductance.

SHORTED ELEMENTS. A tube which has been carelessly handled or otherwise abused may develope short circuits of high or low resistance between its elements. The element supports may become bent, or possibly warped from excessive temperatures. Insulating materials or the coatings of heater and cathode may become loosened. Because of structural or manufacturing faults some tubes become internally shorted during normal use.

One method of checking for shorts between elements is illustrated by diagram A of Fig. 97-5. A rotary switch makes contact with any one element while keeping all other elements connected together. The rotor tongue which contacts the one element is connected to the positive side of the d-c source, while the rotor ring which connects with all other elements is on the negative side of the source and in series with a current meter.

The tube cathode remains connected to the positive side of the source, thus keeping the cathode at the same positive potential as other elements, or else the other elements are negative with reference to the cathode. This prevents emission current from the cathode to other elements. Such current might be confused with short-circuit current.

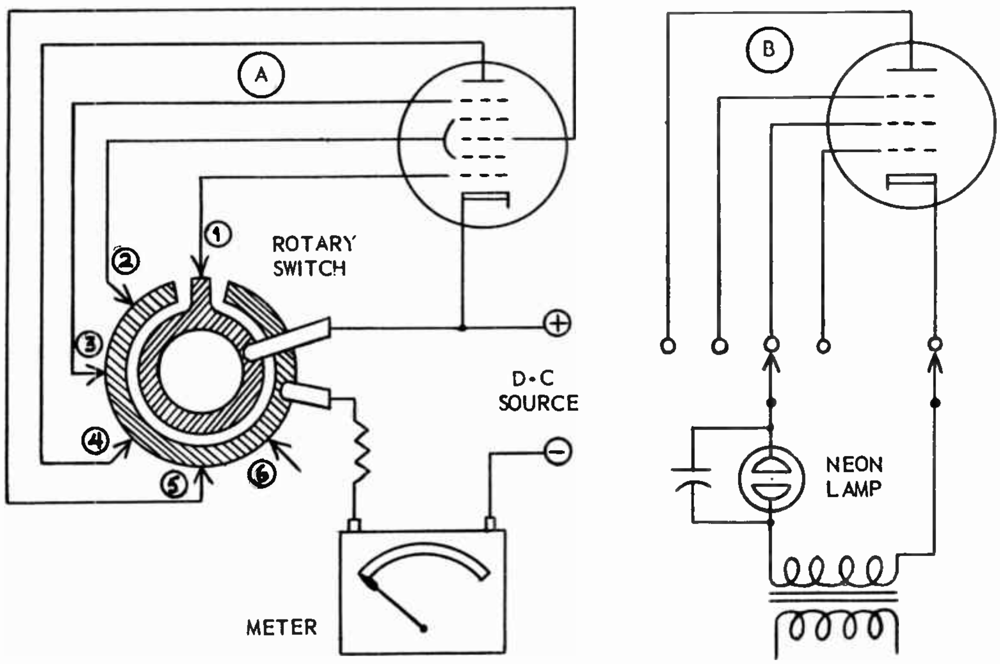


Fig. 97-5. Two methods of testing for short circuits between elements of a tube.

The diagram shows a pentagrid converter tube being tested. With the switch in position 1 the oscillator grid is made positive and all other elements except the cathode are negative. With a short between the oscillator grid and any other element except the cathode, direct current would flow in the meter. Remaining positions of the switch, connect other elements to the positive side of the source, one at a time. In position 6 all elements except the cathode are on the negative side of the test circuit, with the cathode remaining positive in order to check shorts between the cathode and any other element.

Short tests may be made also with a neon lamp indicator as in diagram B of Fig. 97-5. The switching arrangement, not shown in detail, might be similar to that of diagram A or anything equivalent. Alternating voltage in series with the neon lamp is applied to various combinations of elements. Any short allows alternating current in the lamp, and both of its electrodes will glow.

Normal conduction current in an unshorted tube is rectified by the tube, and only one electrode of the neon lamp will glow. When the cathode is not one of the tested elements there is no emission current, and neither lamp electrode will glow provided there are no shorted elements. Consequently, shorts are indicated when both lamp electrodes glow, and freedom from shorts is indicated when only one or neither electrode glows. The neon lamp is bypassed with a small capacitor to prevent the electrodes from glowing due to small alternating currents passed by capacitances of wiring and parts in the tester.

CATHODE-HEATER LEAKAGE. Damaged or defective insulation on the heater conductor or defective insulation anywhere between heater and cathode may allow more or less alternating heater current and voltage to enter the cathode and its circuits. This leakage may cause hum and other troubles unless both the cathode and one side of the heater are connected to the same chassis ground or unless the cathode has a very large bypass capacitor to ground.

Cathode-heater leakage may be troublesome in tubes whose grid circuits contain high resistance or impedance, as is the case with many voltage amplifiers and control tubes. Trouble is less likely where the grid circuit is of relatively small resistance or impedance, as with most power amplifiers. Cathode-heater leakage may be troublesome when it exists in audio and sound section tubes whose cathode currents go to plates and screens of tubes in i-f or r-f sections, and where the audio or sound cathodes are at potentials far higher than ground.

Tests for shorts and leakage should be made after the cathode is hot, since a short which exists at high temperature may disappear when the tube cools. During any test for inter-element shorts the tube should be tapped lightly in order to expose shorts due to loose elements or supports. Inter-element shorts may severely overload grid circuits which should carry no current, and they may reduce the normally negative grid bias or even make the bias positive.

GASSY TUBES. When a tube not highly evacuated is operated with a high plate voltage or large plate

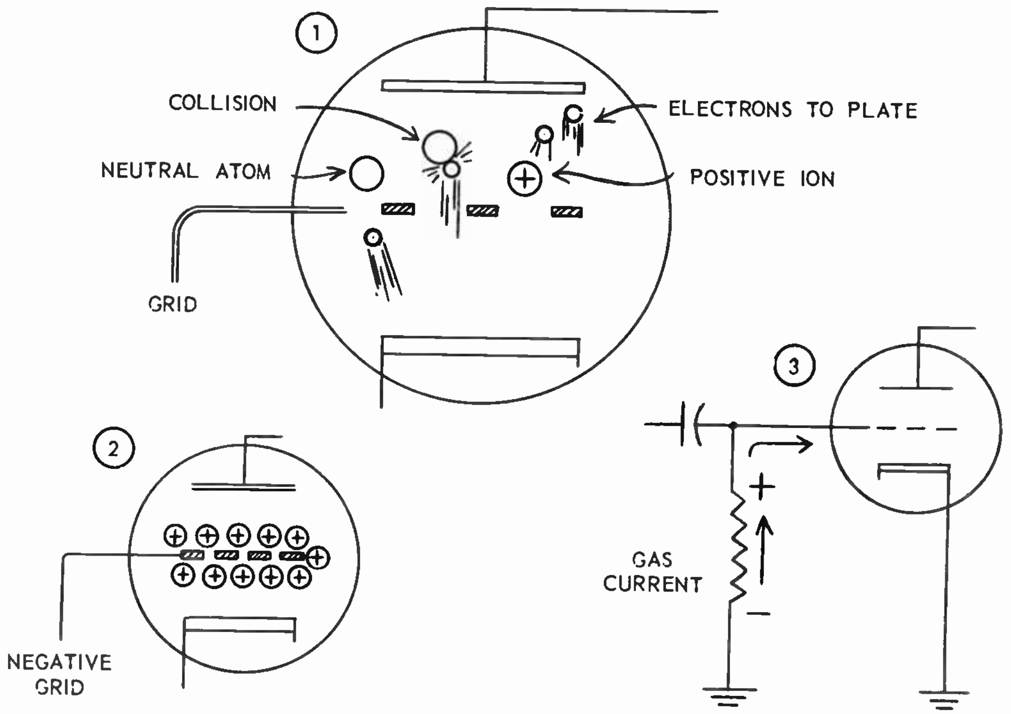


Fig. 97-6. The process of ionization in a gas, and how it may affect the grid bias.

current there is likely to be ionization of gas atoms within the tube. Ionization, as shown at 1 in Fig. 97-6, is the process in which electrons traveling from cathode to plate or screen collide with gas atoms and knock one or more negative electrons off these atoms. This leaves the atoms positive, with a deficiency of negative electrons. Such atoms are called positive ions. The normally negative grid attracts the positive ions and they form a positive sheath around the grid structure as at 2.

Now, as at 3, negative electrons tend to flow toward the grid through the external grid circuit in an effort to neutralize the sheath of positive ions which have gathered around the grid. This grid current or "gas current" flows through the circuit resistance in a direction making the grid end positive with reference to the cathode. This is the reverse of the direction for current which biases the grid with the grid-leak method. The tendency to form an opposite current in the grid circuit may partially or wholly overcome the normally negative grid bias.

A tube may become gassy due to slow leakage of air around the seals for base leads. Overheating of the elements may allow gradual release of gas which has been absorbed or "occluded" in certain of the internal parts and supports.

Symptoms of a gassy tube are as follows. Grid bias is less negative than normal, or may become positive. Less than the normal biasing current flows in a grid-leak biasing resistor, or there may be reversed current. The incorrect grid bias may allow excessive plate current, and this current may overheat the tube. Oscillation may occur in voltage amplifiers, and severe distortion in power amplifiers. A tube which has lost practically all its vacuum, possibly due to a cracked envelope, may become very hot without the heater showing any glow.

When a tube contains considerable gas and is operated with large plate current, ionization may cause a faint blue glow to appear between plate and cathode. This may happen in some large power tubes and in rectifiers without causing any particular trouble. A blue-green glow visible on the inside of a glass envelope on large tubes is caused by fluorescence at the glass surface, not by gas.

A tube must have a minimum of gas to perform well where the grid circuit is of high resistance, as with many voltage amplifiers. Power tube grid circuits usually are of comparatively low resistance, and some gassiness in the tube may not cause trouble. In no case should grid circuit resistances exceed the maximum value recommended by the tube manufacturer for that particular tube.

Most tests for gas depend on the effects of grid current. One method is shown by Fig. 97-7. A switch in the grid circuit is shifted from a to b, cutting off any a-c grid signal and connecting the grid through about one megohm resistance at R to the bias voltage. A test button momentarily short circuits resistance R. When grid current flows in R, due to gas in the tube, the grid becomes less negative than the bias voltage, and the resulting large plate current is indicated by the meter. Operating the test button shorts out the gas-induced grid voltage and allows the grid to become more negative, thus reducing the plate current.

The greater the change of plate current when operating the test button the more gas the tube contains, since change of plate current results from removal of the positive biasing effect which is due to gas current. Change of plate current varies not only with the amount of gas, but also with mutual conductance of the tube. A tube of high mutual conductance may undergo more change of plate current than one of lower mutual conductance, yet may not be so gassy.

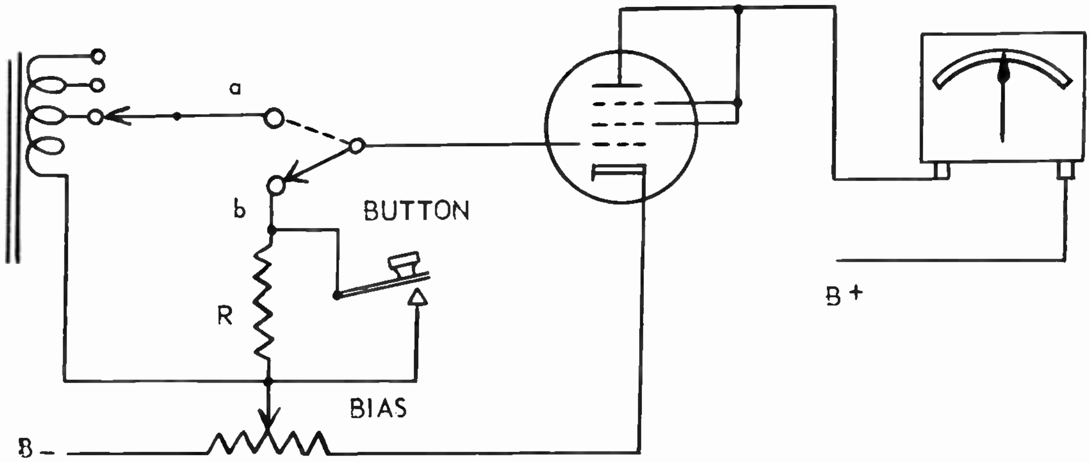


Fig. 97-7. A method commonly employed in testing for excessive gas in amplifier tubes.

OTHER TUBE CHARACTERISTICS. Amplification factor, plate resistance, and power output may be measured with suitable apparatus, but seldom if ever are they measured during service work. Fig. 97-8 illustrates principles sometimes employed for measuring amplification factor and plate resistance.

A low-frequency alternating voltage, at possibly 1,000 cycles, is taken from opposite sides of a source and is therefore of simultaneously opposite polarity at the grid and the plate of the tube being measured.

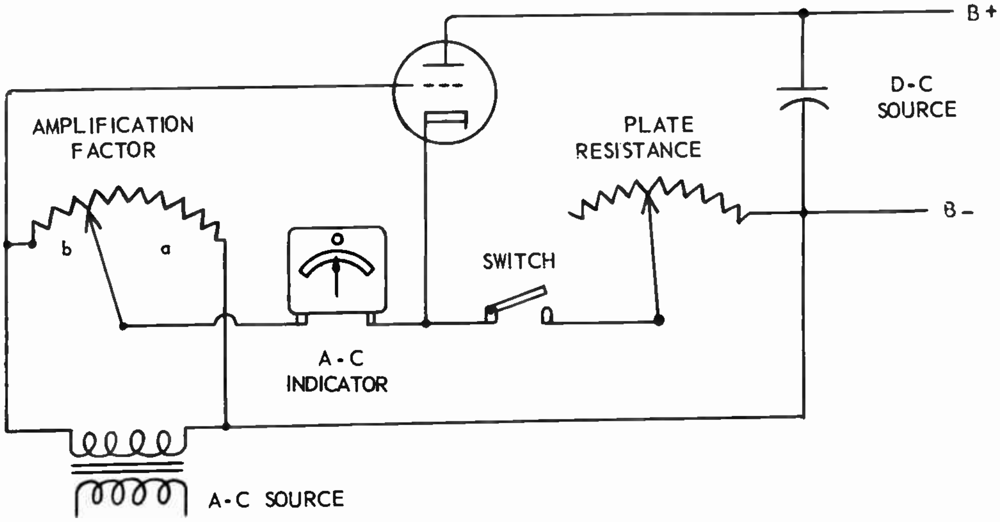


Fig. 97-8. Principal parts of a test circuit which allows measurement or computation of amplification factor and plate resistance.

This alternating voltage is applied also across the amplification factor potentiometer, whose slider is connected through a sensitive current indicator to the tube cathode. When the slider is adjusted so that there is no current in the indicator, the alternating voltage at the plate must exceed the opposing alternating voltage at the grid by a number of times which is the amplification factor. Then this factor is equal to the ratio of resistance at a to resistance at b. The potentiometer may have a scale graduated in values of amplification factor.

Now a note is made of the measured amplification factor, and that potentiometer is adjusted to have equal resistances at a and b. Next, the switch going to the plate resistance rheostat is closed, and that rheostat is adjusted until no current is shown by the indicator. Plate resistance of the tube is the product of rheostat resistance and a number which is equal to the previously measured amplification factor minus one.

There are a number of ways for measuring power output. One of the simplest employs a circuit similar to that for measuring mutual conductance, but, as in Fig. 97-9, substitutes for the current meter an adjustable load resistance and an a-c voltmeter connected to measure alternating potential difference across the load. Power output in watts is equal to the square of the a-c voltage divided by ohms of load resistance. The meter dial may be graduated for several ranges of power, with load resistance and grid input adjusted to suit each range.

TUBES OTHER THAN AMPLIFIERS. Power rectifiers usually are considered to have reached the end of useful life when rectified direct current drops to 80 per cent of the normal average when a-c plate voltage and filter loads are such as commonly used for the tube being tested. This general rule applies also to small diodes employed for detectors and control tubes.

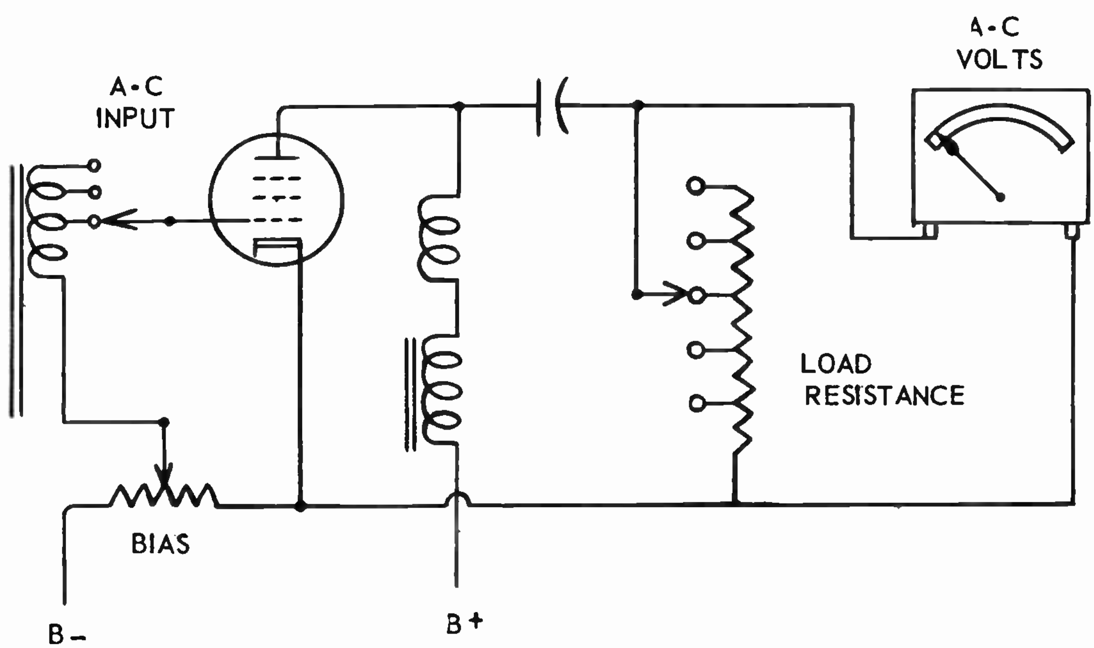


Fig. 97-9. A modification of a mutual conductance tube tester which allows measurement of power output.

COMMERCIAL TRADES INSTITUTE

TELEVISION

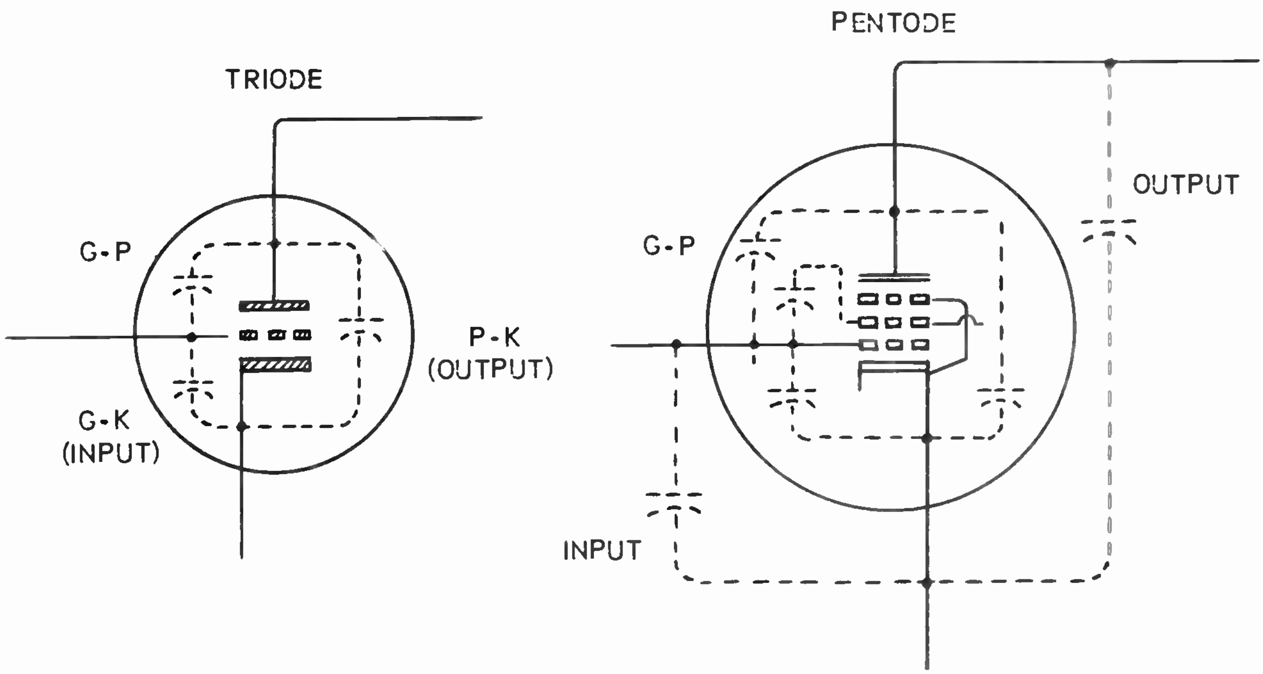


Fig. 97-10. The principal interelectrode capacitances in a triode and in a pentode.

Oscillator tubes are difficult to test satisfactorily other than by operation in their regular circuits. Whether or not a given tube will oscillate depends very largely on circuit impedances and on the intended oscillating frequency. A tube which does not oscillate in one circuit often is satisfactory in another circuit.

The usual test for oscillation is measurement of negative grid voltage with a vacuum tube voltmeter while the tube is in its regular position. Practically all oscillators operate with grid-leak bias. Unless the grid is at least three or four volts negative with reference to the cathode, the tube probably is not oscillating. Oscillator grid voltage often is between 10 and 20 volts negative.

The mixer section of a converter tube may be tested as a voltage amplifier with suitable test voltages on the mixer elements. The oscillator section is tested like any other oscillator, using the elements which function in this capacity. The effectiveness of the tube as a converter depends on conversion transconductance.

Conversion transconductance is comparable to mutual conductance. When measuring mutual conductance the alternating voltages on grid and plate are at the same frequency. But conversion conductance in micromhos is equal to the number of microamperes of mixer plate current at the intermediate frequency, per volt of radii-frequency signal on the r-f signal grid of the converter tube. The different frequencies at grid and plate make it difficult to measure conversion transconductance.

INTERELECTRODE CAPACITANCES. There are capacitances between the internal elements of a tube because the elements are conductors and the vacuum between them acts as a dielectric. These capacitances are important in all high-frequency operation, since they form a large part of the circuit capacitance which tunes r-f and i-f amplifiers and which enters into peaking and other performance factors of many circuits.

The three internal capacitances of a triode are shown by the triode symbol of Fig. 97-10. Grid-plate (G-P) capacitance is the one which affects tendency of the tube to oscillate due to energy feedback from plate to grid through the tube. Grid-cathode (G-K) capacitance is that which the tube places across the external grid circuit. It may be called input capacitance. Plate-cathode (P-K) capacitance is placed across the external plate circuit. It may be called output capacitance. The accompanying table gives typical values of interelectrode capacitances for some single and twin triodes often found in television circuits. All are miniature tubes except the 6SN7, which is an octal type.

INTERELECTRODE CAPACITANCES OF TRIODES, IN MMF.

Tube	Grid-Cathode	Plate-Cathode	Grid-Plate
6AB4	2.2	0.5	1.5
6BQ7, 1st section	2.7	0.15	1.15
2nd section	4.9*	.15	1.15
6C4	1.8	1.3	1.6
6J6, 1st section	2.2	0.4	1.6
2nd section	2.2	.4	1.6
6SN7, 1st section	2.8	0.8	3.8
2nd section	3.0	1.2	4.0
12AT7, 1st section	2.4	0.5	1.5
2nd section	2.4	.4	1.5

* With grounded grid connection.

The more important interelectrode capacitances in a pentode are shown by the pentode symbol of Fig. 97-10. The combined effect of capacitances between the control grid and all other elements operating at the same signal potential as the cathode is called the input capacitance. The combined effect of capacitances between the plate and all elements operating at the signal potential on the cathode is called the output capacitance.

An accompanying table gives average input, output, and grid-plate capacitances of a number of miniature voltage amplifier pentodes found in television receivers. All the capacitances are measured with no external shield on the tube. Addition of a close fitting shield, connected to the cathode, causes a considerable increase of output capacitance, possibly a slight increase of input capacitance, and there may be a decrease of grid-plate capacitance.

INTERELECTRODE CAPACITANCES OF MINIATURE PENTODES, IN MMF.			
Type	Input	Output	Grid-Plate
6AG5	6.5	1.8	0.025
6AH6	10.0	2.0	.03
6AU6	5.5	5.0	.0035
6BA6	5.5	5.5	.0035
6BC5	6.5	1.8	.03
6BH6	5.4	4.4	.0035
6CB6	6.3	1.9	.02

Grid-plate capacitances of pentodes are only about 0.2% to 2.0% of the grid-plate capacitances in any of the previously listed triodes. On the other hand, the input capacitance of the triodes always is much smaller than in any pentode, while triode output capacitances also run smaller than those of pentodes. Other than the possibility of feedback from plate to grid, it is easier to tune to high frequencies with triodes than with pentodes of the same general structure, since the unavoidable and uncontrollable input and output capacitances are smaller in the triodes.

TUBE SUBSTITUTIONS. A tube of one type may be substituted for one of a different type because of two principal reasons. First, certain tubes occasionally become temporarily unavailable, and to keep a receiver in operation it is necessary to substitute another tube which is available or which you have on hand.

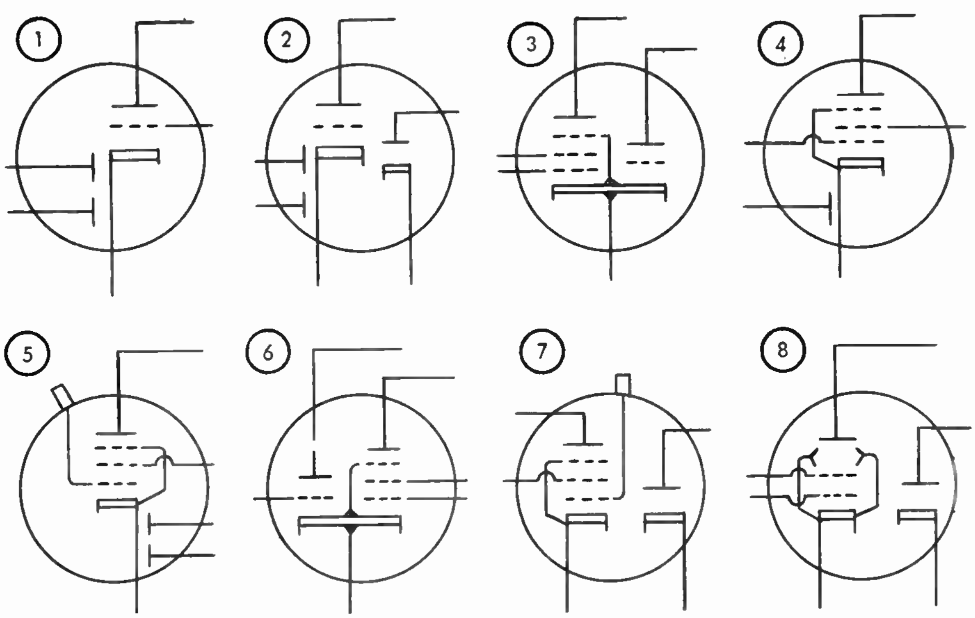


Fig. 97-11. Combination tubes in which the sections are of different types.

COMMERCIAL TRADES INSTITUTE

TELEVISION

Second, sensitivity and output of some older receivers and of a few recent ones may be improved by substituting tubes which provide greater transconductance. A mere listing of possible substitutions would be of little value, since there are hundreds of possible changes with present types and new types are continually added by the makers. It is better to become familiar with principles which govern the possibilities.

Any substitution is bound to change the performance, unless operating voltages can be altered to compensate for change of tube characteristics. It is desirable to improve performance when possible, but if results are somewhat less satisfactory than with the original tube, but still passable, this may be tolerated when no other replacement tube is available.

Most substitutions are in r-f or i-f voltage amplifier stages and in video amplifiers, where difficulties usually are not too great. More trouble may be encountered in some sync circuits, in limiter and demodulator circuits, in sweep oscillators and automatic sweep frequency controls, and in automatic gain controls.

Tubes having only a single set of elements present the least difficulty. Some twin types might be replaced by two tubes, each similar to one of the twin sections, although it would be necessary to find chassis space and sometimes to make an extra socket hole for the added tube.

Combination tubes having sections of different types are difficult to replace other than with a generally similar type. The elements in some such tubes are as shown by symbols in Fig. 97-11. At 1 is the familiar voltage amplifier triode with two detector diodes. A fairly common combination of a triode with three diodes is shown at 2. There are and have been voltage amplifier pentodes combined with a triode (3), with one diode (4), or with two diodes (5). Power pentodes have been combined with a triode (6) or with a rectifier (7). Beam power amplifiers may be combined with a rectifier as at 8. The types at 3, 5, and 7 are obsolete, but may be found in older receivers.

The arrangement of base pins and the type of socket which will take the pins may determine the feasibility of a substitution. In general use are 7-pin and 9-pin miniature sockets, also octal and lock-in sockets. A tube which fits any one of these won't fit in any other type socket. The 7-pin miniature socket requires a chassis hole 5/8 inch in diameter, while the 9-pin type takes a 3/4-inch hole.

Most octal and lock-in sockets will fit into an opening 1 1/8 inch in diameter, which allows interchanging these two types and shifting the wiring connections. For many lock-in tubes there are octals which serve for emergency substitution, or the change may be from octal to lock-in type. Some of the equivalents are as follows.

Lock-in	Octal	Lock-in	Octal	Lock-in	Octal
7A4	6J5	7B5	6K6	7C7	6J7
7A6	6H6	7B7	6SK7	7H7	6AB7
7A7	6SK7	7B8	6A8	7N7	6SN7
7AD7	6AG7	7C5	6V6	7Q7	6SA7
7B4	6SF5	7C6	75	7Y4	6X5

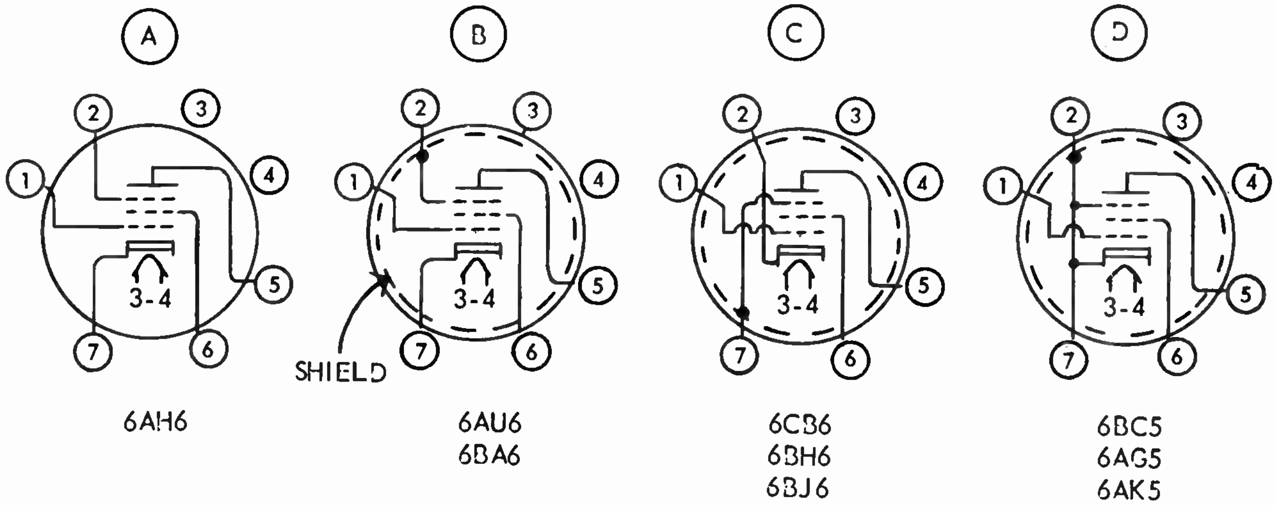


Fig. 97-12. Connections of elements to base pins in some voltage amplifier miniature pentodes.

Even though the same socket will take both the original and the substitute tubes and when base pins and internal element connections are almost alike, the wiring on the socket always must be examined closely. Some reasons are illustrated by Figs. 97-12 and 97-13 which relate to 7-pin miniature voltage amplifier pentodes used for television r-f and i-f amplifiers and for some video amplifiers.

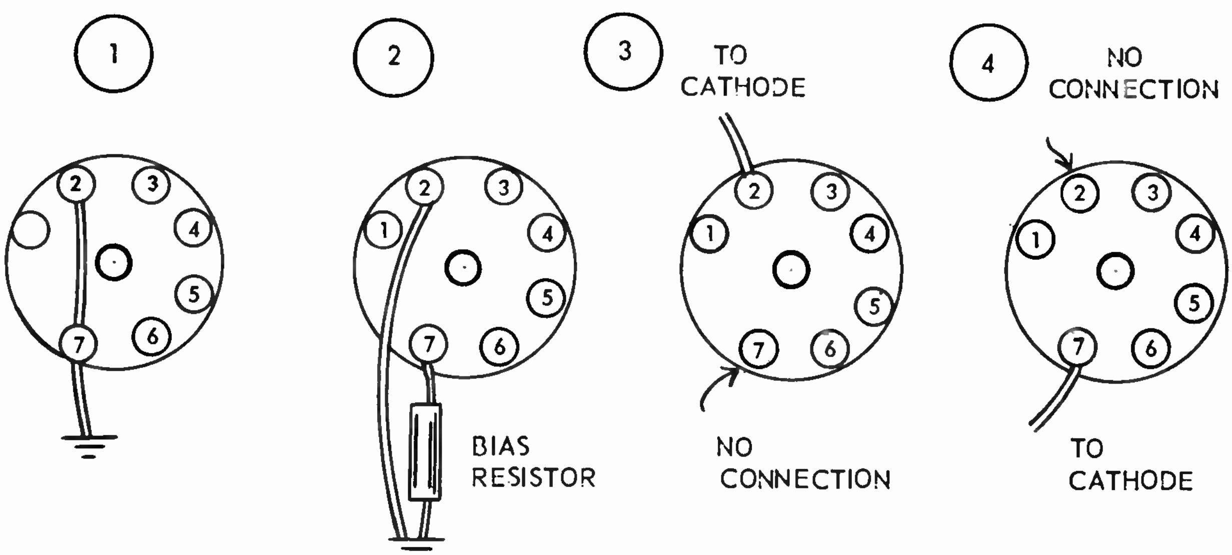


Fig. 97-13. Some of the socket connections which may cause difficulties when making tube substitutions.

Fig. 97-12 shows internal connections for nine such tubes. The heater always is connected to pins 3 and 4, the control grid is connected to pin 1, the plate to pin 5, and the screen to pin 6. Differences are in connections of the cathode, the suppressor, and the internal shield to pins 2 and 7.

At A the cathode is connected to pin 7 and the suppressor to pin 2. There is no internal shield. Diagram B shows the same cathode and suppressor connections, but there is an internal shield connected to the same pin as the suppressor, pin 2. At C the cathode and suppressor connections are reversed on pins 2 and 7, with the internal shield still connected to the suppressor pin. At D the cathode, suppressor, and internal shield are connected together, and all are connected both to pin 2 and pin 7. This use of two pins for the cathode allows arranging the external circuit to lessen the effective inductance of the leads, which helps at high frequencies.

In many receivers you will find socket lugs 2 and 7 connected together and to ground, as at 1 in Fig. 97-13. This may be done when grid bias is by the grid-leak method or by connection of the grid return to some point which is negative with reference to ground and the cathode. Then, so far as internal connections are concerned, any of the nine tubes under discussion could be used without rewiring the socket, for the external connection between lugs 2 and 7 makes all the element connections equivalent to those at D in Fig. 97-12.

Supposing you were to find socket connections as at 2 in Fig. 97-13, with a cathode bias resistor on lug 7 and with lug 2 grounded. This would be suitable for all tubes whose internal connections are shown at A and B of Fig. 97-12. But were one of the tubes at C to be put into the socket wired as at 2 there would be no cathode bias and the suppressor would be connected through a resistor to ground. Inserting a tube whose pin connections are shown at D of Fig. 97-12 would short circuit the biasing resistor to ground, and there would be no bias voltage.

When the original tube is one of those at D of Fig. 97-12, the cathode end of the external circuit may be connected either to socket lug 2 or 7, with the remaining one of these lugs left unconnected, as at 3 and 4 of Fig. 97-13. This would not do for tubes whose internal connections are as shown by any of the other diagrams. The examples of socket wiring which have been illustrated cover only a few of the possible variations, but they show how necessary it is to examine the actual socket wiring before making a tube substitution.

Heater voltages and currents may determine whether a substitution is practicable. As an example, all of the voltage amplifier pentodes used to illustrate base pin and socket connections require 6.3 volts for their heaters. But the 6BH6 and 6BJ6 take only 0.15 ampere of heater current, the 6AK5 takes 0.175 ampere, the 6AH6 takes 0.45 ampere, and all the others require 0.30 ampere.

When all heaters are in parallel on a 6.3-volt transformer winding, as in the upper diagram of Fig. 97-14, each tube will take its required current at this voltage and differences between rated currents will not affect the possibility of substitution. But on a series heater circuit, as in the lower diagram, tubes of different heater current ratings would not be interchangeable.

The hot resistances, with rated heater voltage and current, are shown for each of the tubes. Using a substitute tube of lower than the original current rating would reduce the current and emission in other series tubes, and soon the first one would burn out due to excessive current unless it had a higher voltage rating. With a substitute tube of higher heater current rating its cathode would not be heated enough for normal emission.

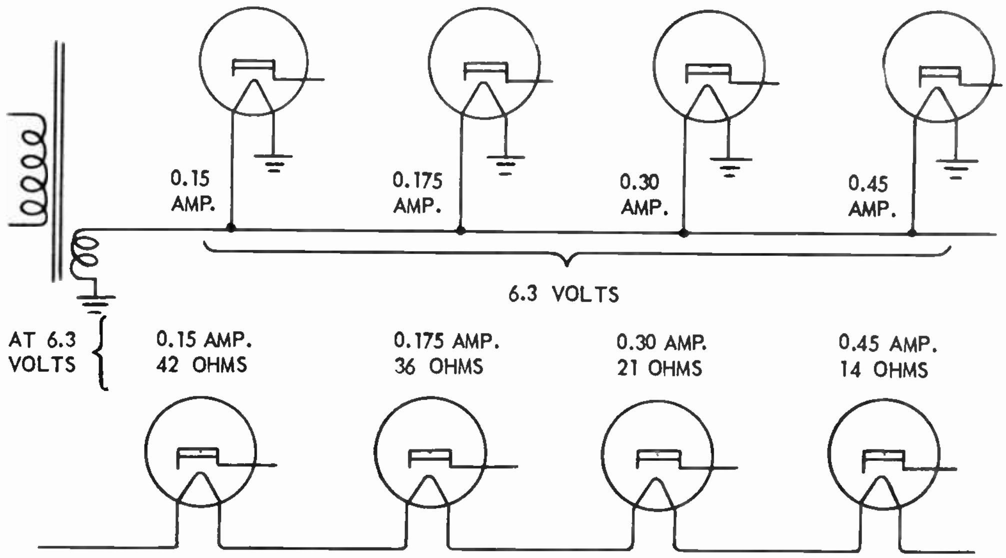


Fig. 97-14. Tubes rated for different heater currents cannot be on a series heater circuit without some form of current equalization.

A tube of lower heater current rating may be used in series with higher current tubes by paralleling the low-current heater with a resistor, as at the top of Fig. 97-15. To determine the required shunting resistance, first subtract from the current rating of other tubes, or the series circuit current, the current rating for the heater in the substitute tube. Then divide the rated heater volts of the substitute by the amperes of current difference. The quotient equals the required resistance in ohms. The power rating of the shunt resistor should be at least double the actual dissipation as computed from resistor current and voltage drop.

To use a tube of greater heater current rating than that of other tubes in a series string would require bringing extra heater current to this substitute tube through resistors shunting the other tubes, with some arrangement such as at the bottom of Fig. 97-15. Nearly always this would overload ballast resistors and other elements in the series circuit or its current source, and would not be practicable without making careful calculations and circuit alterations appropriate for each individual case.

If you merely remove one tube and substitute a different type, with no adjustments of operating voltages, the performance may be better or worse, and the new tube may be overheated by excessive plate or screen currents and voltages. To obtain best performance without overheating, it is usually necessary to alter the grid bias and possibly also the plate and screen voltages.

Mutual conductance or transconductance of either a triode or pentode increases with plate current. Fig. 97-16 shows relations between plate current and mutual conductance for two voltage amplifier triodes. Plate currents are varied by changes of grid bias, with plate and screen voltages held constant. Curves for other tubes would be of similar shape but would not have the same values.

Plate current in a triode depends on plate voltage and grid bias, while in a pentode the plate current depends almost wholly on screen voltage and grid bias, very little on plate voltage. With either type of tube it is usually possible to obtain a suitable and safe plate current by altering the grid bias, as by changing a cathode resistor.

The accompanying list shows how changes of grid bias vary the plate currents and transconductances of some of the common voltage amplifier pentodes when there are no changes in plate or screen voltages.

TUBE	NEGATIVE BIAS, VOLTS	PLATE CURRENT, MA	TRANSCONDUCTANCE, MICROMHOS
6AU6	1.75	7.0	4300
	1.0	10.6	5200
6BC5	2.0	6.2	5100
	1.7	7.5	5800
6BH6	1.1	7.0	4500
	1.0	7.4	4600
6CB6	2.7	7.0	4900
	2.2	9.5	6200
6BJ6	1.6	7.0	2800
	1.0	9.2	3800
6AH6	2.6	7.0	6600
	2.0	10.0	9000

A pentode may be replaced only with another pentode. If the original pentode operates with remote cutoff for gain control, the substitute tube must be another remote control pentode. Otherwise the remote and sharp cutoff types may be interchanged in most circuits.

A triode sometimes can be replaced with a pentode connected to operate as a triode. If the suppressor is not internally connected to the cathode (A, B, and C of Fig. 97-12) this is done by wiring together the socket lugs for plate, suppressor, and screen. If the suppressor is internally connected to the cathode (D of Fig. 97-12) only the socket lugs for plate and screen are connected together for triode operation.

When operating a pentode as a triode it usually is necessary to employ less plate voltage and a more negative grid bias than for pentode operation, although this depends on the particular tube. As a rule it is satisfactory to limit the plate current as a triode to the same value as plate current recommended for typical operation as a pentode. Grid-plate capacitance will be increased by the triode connection, and it may be necessary to add an external shield to prevent excessive peaking or oscillation. Changes of input and output capacitance ordinarily are no greater than can be compensated for by realignment of the circuits.

The maximum power which a tube is capable of dissipating without overheating determines the allowable combinations for voltages and currents of plates and screens. Every tube is rated for some certain maximum plate voltage and maximum screen voltage, which are voltages respectively between the cathode and the plate or screen. There are maximum ratings also for watts dissipated as heat by the plate and the

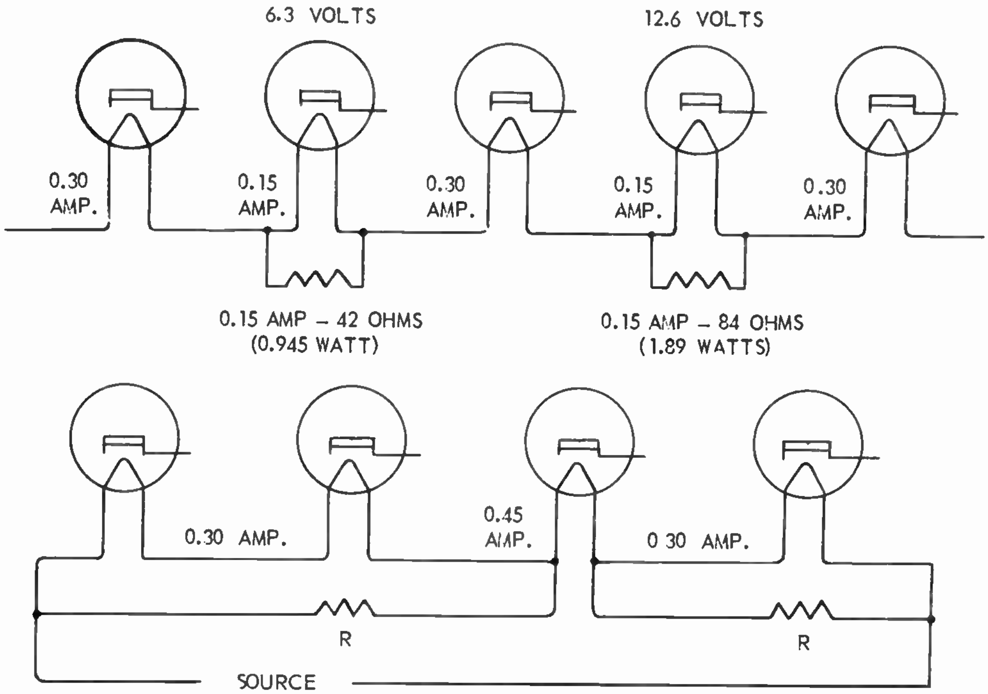


Fig. 97-15. Paralleled or shunting resistors carrying part of the circuit current allow using series heaters of different ratings.

screen. Actual dissipation in watts is equal to the product of element volts and milliamperes, divided by 1,000. Tubes seldom are operated to dissipate much more than 90 per cent of the maximum allowable watts.

To illustrate relations between power dissipation, voltage, and current, we may take the 6BC5, whose plate ratings are 2 watts and 300 volts and whose screen ratings are 0.5 watt and 150 volts. Operating with 250 volts and 7 milliamperes for the plate, the dissipation is 250 times 7, divided by 1,000, which comes to 1.75 watts. This is about 88 per cent of maximum permissible power.

The curve for the 6BC5 in Fig. 97-16 is drawn for 250 volts between plate and cathode. The solid line portion of the curve covers operation within the plate dissipation limit of 2 watts which, with 250 plate volts, allows no more than 8 ma of plate current. The curve for the 6AH6 is drawn for 300 plate-cathode volts. Maximum plate dissipation for this tube is 3 watts, so maximum plate current with 300 plate volts is 10 ma. Operating on the broken line portions of the curves would overheat the tubes.

With screen voltage no higher than the rated maximum, and a grid bias which prevents plate current from exceeding the maximum plate power limit, there seldom is danger of too much dissipation from the screen.

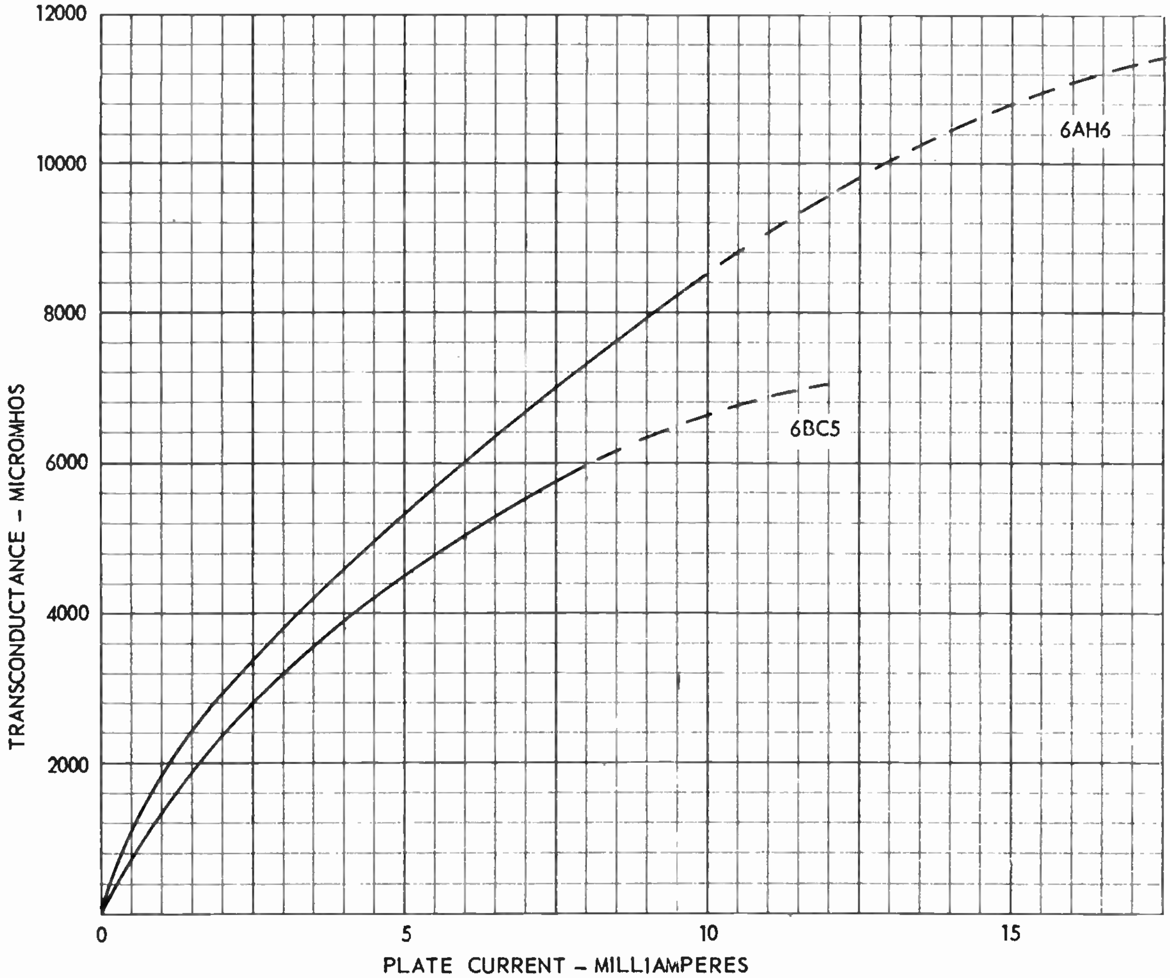


Fig. 97-16. Relations between plate current and transconductance in two voltage amplifier pentodes.

For any given values of plate and screen voltage, which are applied to a tube, the grid bias must be sufficiently negative to prevent a plate current which would cause excessive plate dissipation in watts. Unless you have reliable information relating to the tube or tubes involved, it is not safe to make a substitution without measuring the final plate voltage and current, computing the watts of dissipation, and checking this against the limits set by the tube manufacturer.

TELEVISION

LESSON NO. 98

TELEVISION AT ULTRA-HIGH FREQUENCIES

Training Your Mind...

C. T. I.

Perfecting Your Skill

DO NOT TEAR — CUT ALONG THIS LINE AND SEND IN FOR GRADING.

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind
C.T.I.
Perfecting Your Skill

COMMERCIAL
TRADES
INSTITUTE

Chicago, Ill.

LESSON NO. 98

TELEVISION AT ULTRA-HIGH FREQUENCIES

Throughout the whole history of radio and television the trend has been toward the use of higher and higher carrier frequencies. Every increase of frequency has brought new problems in the manufacture of receivers which must be designed to sell at prices people are willing to pay, which will perform well in the hands of inexperienced operators, and which can be serviced without too great difficulty when trouble appears. The extension of television broadcasting into the ultra-high frequency channels 14 through 83 has been no exception to this rule.

In this first lesson on ultra-high frequency practices we shall examine most of the difficulties. After reading the first few pages you will wonder how it is possible to receive and reproduce television signals and programs in this range of frequencies—everything seems to work against it. Then we shall commence studying practical solutions for the many problems, and learn how ultra-high frequency reception may be made entirely satisfactory.

The entire extent of frequencies used for radio transmission and reception is divided into several classifications, as follows.

NAME	ABBREVIATION	RANGE
Medium frequencies		0.3 mc to 3.0 mc
High frequencies	hf or h.f.	3.0 mc to 30.0 mc
Very High frequencies	vhf or v.h.f.	30.0 mc to 300.0 mc
Ultra High frequencies	uhf or u.h.f.	300.0 mc to 3,000.0 mc
Super High frequencies	shf or s.h.f.	3,000.0 mc to 30,000.0 mc

Relations between the several television bands are shown by Fig. 98-2. The ultra-high frequency channels extend from 470 through 890 mc. This is an average frequency nearly four times that of the very-high frequency channels 7 to 13, and almost ten times as high as in channels 2 to 6.

One wavelength in the middle of vhf channel 4 measures more than 11 feet. In the middle of vhf channel 10 a wave is about 5 feet long. In the middle of the uhf band a wavelength is only about 17½ inches long, and a half-wave dipole antenna can be spanned with the fingers of one hand.

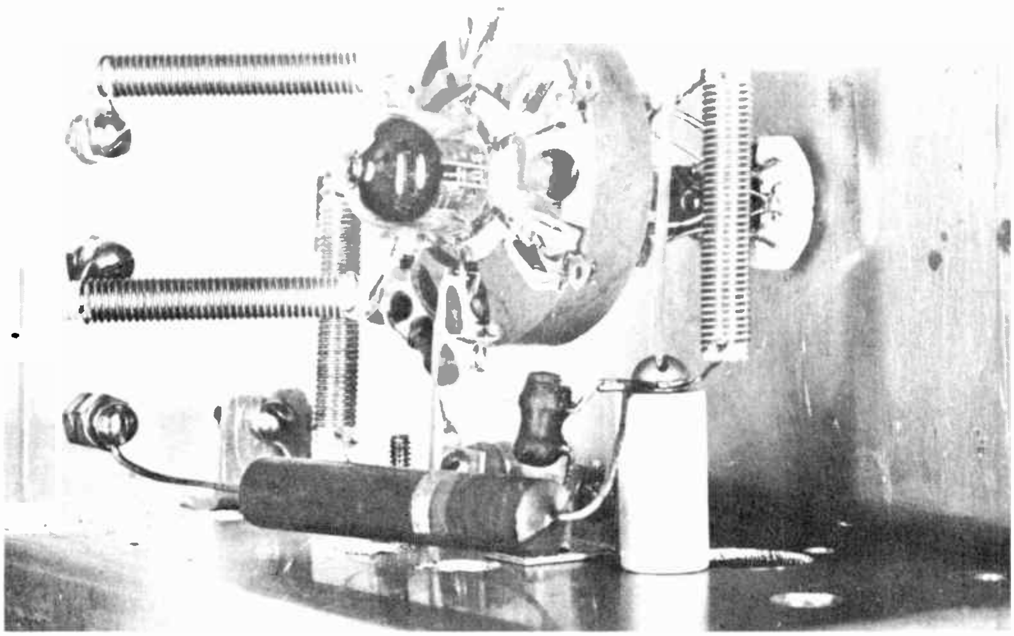


Fig. 98-1. An experimental ultra-high frequency oscillator which will operate at frequencies up to 1,000 megacycles. The tube is a 6F4 Acorn type.

The differences between receiving equipment for the vhf channels 2 through 13 and for the uhf channels 14 through 83 are entirely in the antenna, the transmission line, and the tuner. Once the modulated carrier frequencies for any of the bands have been converted to modulated intermediate frequencies by the mixer, the following sections of all receivers may be identical in design and operation.

The uhf carrier waves follow a line-of-sight direction even more closely than the vhf carriers. There

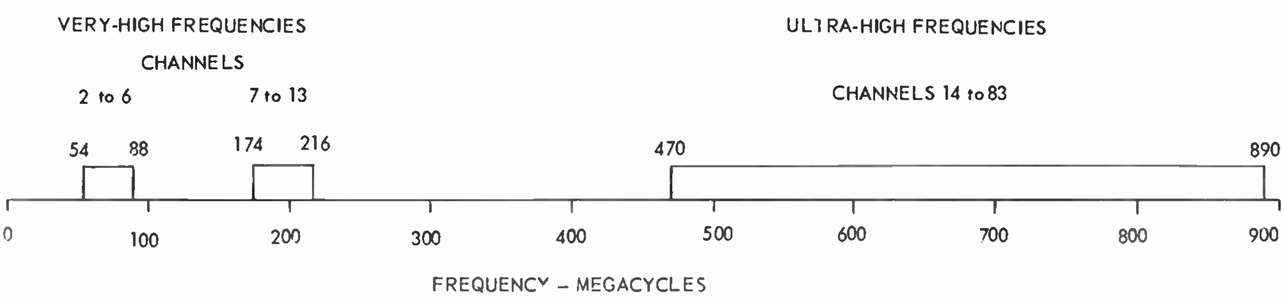


Fig. 98-2. Frequency relations between the very-high and ultra-high television channels.

is even greater energy absorption and greater "shadowing" effect due to hills, buildings, trees, and other objects along the line from transmitting to receiving antennas. In localities where uhf signals come close to the ground before reaching the receiving antenna, such things as moving trains and large trucks can cut off reception as they pass by. Maximum reception distance for uhf signals usually is somewhat shorter than for vhf signals.

Uhf signals are highly directional. Reflections such as cause ghost images are more troublesome than at the lower television frequencies. This is because the size of an object which can cause total wave reflection is roughly proportional to the signal wavelength. The very short waves in the uhf band can be wholly reflected by proportionately small surfaces.

Partially compensating for the many shortcomings in uhf signal propagation is the relative freedom from certain kinds of interference. This is true of most spark type interferences, also of interference from medical and industrial apparatus working at very-high frequencies. Television was extended into the ultra-high frequencies not because of operating advantages, but because in no other range were there enough frequencies not already assigned to other services.

Each uhf channel covers 6 mc, just as does each vhf channel. Uhf channel 14 extends from 470 to 476 mc. Channel 83 extends from 884 to 890 mc. For both uhf and vhf work there is the same general type of carrier signal. There is vestigial sideband transmission, 4.5 mc separation between video and sound carriers, amplitude modulation for video signals, and frequency modulation for sound signals. The video carrier in any uhf channel is 1.25 mc higher than the low limit of that channel, and the sound carrier is 0.25 mc below the high limit, just as in the vhf channels.

With all of the uhf channels in use there will be about 1,500 television stations working in this band. Combined with the possible total of about 550 vhf stations, there will be hardly a populated area of the United States where television programs cannot be received. Effective radiated carrier powers may be anything between one and 1,000 kilowatts, with the greatest powers used in the larger cities and the least powers in the smallest cities.

UHF TUBES. None of the tubes commonly used in the carrier tuning sections of standard broadcast receivers are suitable for use in uhf tuners. Furthermore, many tubes which are satisfactory in tuners for the vhf television bands become almost useless at frequencies in excess of 300 to 400 mc. A few will perform quite well up to 600 mc or thereabouts, but cannot reach the high end of the uhf band.

One reason why ordinary tubes fail at ultra-high frequencies is that their internal capacitances are too great. The capacitive reactance of the tube drops lower and lower as frequency goes up. The small reactance of the tube, connected across the grid circuit, makes it impossible to obtain large voltage gains. We learned this when studying video amplifiers. As an example, the input capacitive reactance of most of the miniature pentodes used in vhf receivers would be less than 40 ohms at a frequency in the middle of the uhf television band. No worth-while signal voltage can be built up across such a small reactance.

In addition to lack of gain, the large input capacitance of the tube may add greatly to the total capacitance in a resonant grid circuit, and make it difficult or impossible to tune the circuit at ultra-high frequencies. It is largely because input and output capacitances of triodes are smaller than in pentodes that the majority of uhf tubes are triodes.

Ordinarily we think of the input capacitance of a triode as that between the grid and cathode, and of the output capacitance as that between the plate and cathode. But there is also a considerable capacitance between the cathode and heater, possibly 3 mmf or more, and unless the cathode is directly connected to one side of the heater this capacitance may add still further to the total.

Input capacitances of triodes especially designed for uhf work may be as small as 1 to 2 mmf, while output capacitances of many of these types are much less than 1 mmf. This is accomplished by suitable shaping and positioning of the elements.

At the left in Fig. 98-3 are the elements of a typical r-f pentode. At the center are the elements of a 6J6 twin-triode, a type which can give satisfactory performance up to about 600 mc. At the right you can see the still smaller element structure of a 6AF4 tube, a special uhf type which operates well to about 1,000 mc or even somewhat higher.

A second reason why ordinary tubes fail to perform well at ultra-high frequencies is that the inductances of internal leads between elements and base pins are too great. The addition of these internal inductances to other inductances in the external circuit may make a total so great as to prevent tuning in the ultra-high frequencies.

The internal inductances also may combine with the internal capacitances of the tube to form a complete resonant circuit at some frequency within the uhf band. Such self-resonance of the tube limits the

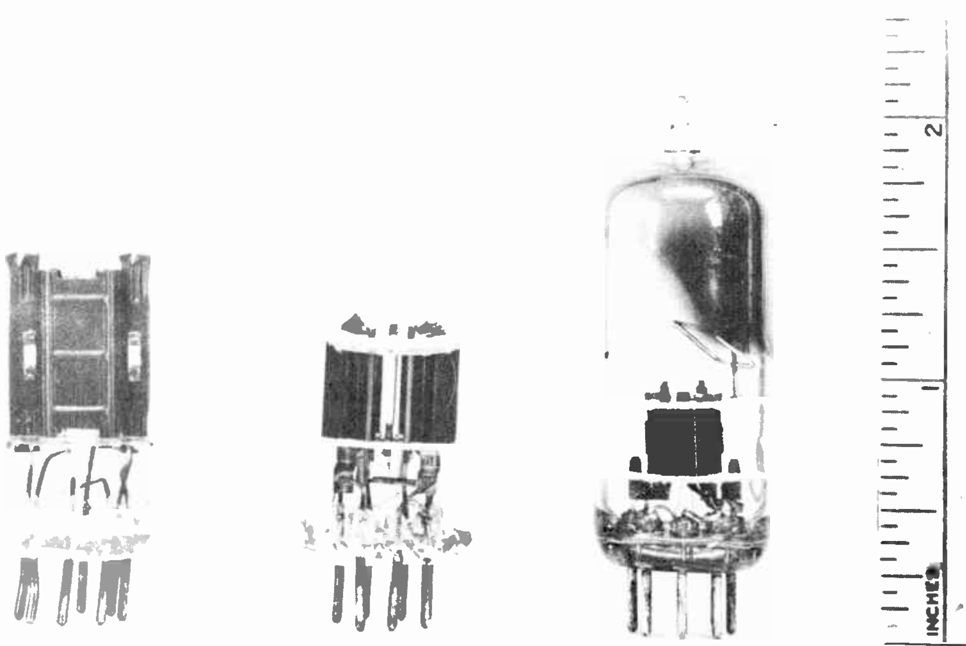
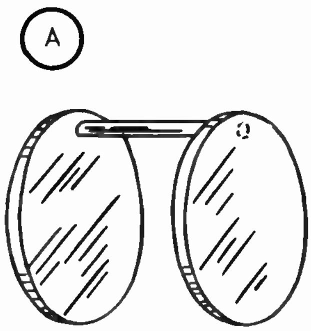


Fig. 98-3. As tubes are designed for higher and higher frequencies, the sizes and spacings of the elements becomes progressively less.



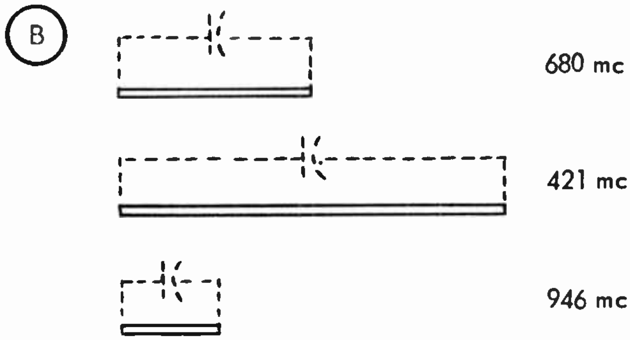


Fig. 98-4. Inductances of short, straight wires become highly important at the ultra-high frequencies.

maximum operating frequency because it prevents oscillation in an externally connected circuit at the ultra-high frequencies, and also makes amplification impossible at these frequencies.

After having worked with the relatively low frequencies of the vhf television channels for so long it is difficult to realize the importance of inductances in leads and wiring connections at the ultra-high frequencies. At 680 mc, the middle of the uhf band, the inductance of one inch of number 14 copper wire (about 1/16 inch in diameter) is approximately 0.0142 microhenry. If this one inch of straight wire were connected across a capacitance of only about 3.85 mmf we would have a circuit resonant at 680 mc.

It is quite evident that there would be no way of connecting one inch of wire across a capacitor without bending the wire. To loop the wire around the capacitor would increase the inductance of the wire. What we can do is connect the straight wire to two discs, between whose extended surfaces there is capacitance. This is illustrated at A in Fig. 98-4. It is one possible form of resonant circuit for ultra-high frequencies.

Doubling the length of a straight wire more than doubles its inductance, and cutting the length to half drops the inductance to less than half the original value. Resonant frequency does not, however, vary directly with changes of inductance. The variation involves "the inverse of the square root" of the inductance. Omitting the mathematics, the changes in resonant frequency when using the same capacitance in all cases would be as shown at B in Fig. 98-4. Commencing with one inch of wire, and resonance at 680 mc, doubling the length would drop the resonant frequency to about 421 mc, and halving the length would raise the frequency to about 946 mc.

The foregoing example is intended only to bring out the exceeding importance of having the shortest possible leads and connections in uhf circuits. Reducing the length of a connection by a half-inch may make the difference between a circuit which will tune to some desired frequency and a circuit which cannot be tuned.

In addition to reducing the length there are two other ways of lessening the inductance of wires or leads. One way is shown at the left in Fig. 98-5. We may use a conductor of larger diameter. If you com-

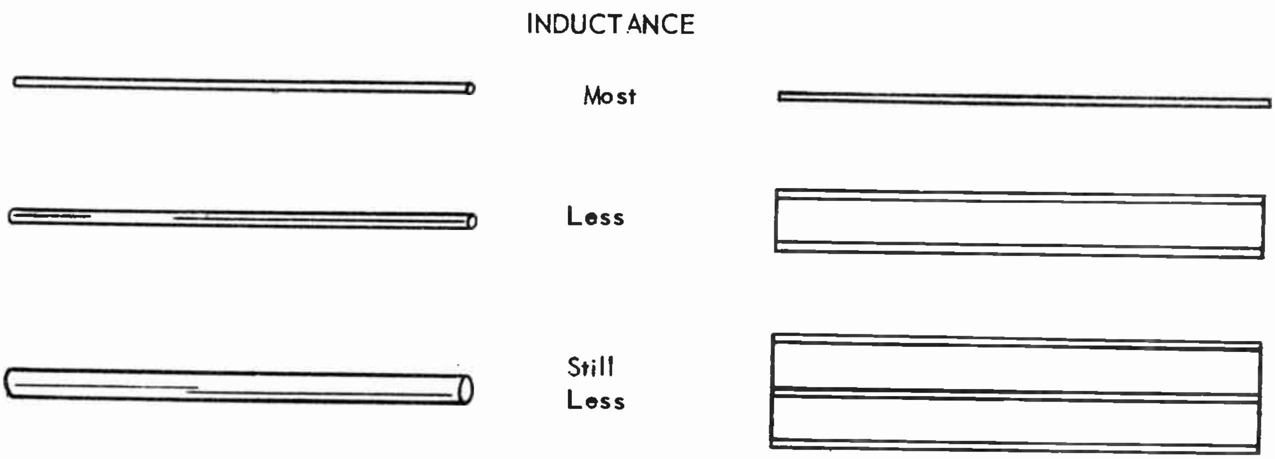


Fig. 98-5. Inductances of straight wires may be lessened by increasing their diameters or by using two or more wires connected in parallel.

mence with a very thin wire, the diameter may have to be increased by three or four times to drop the inductance a worth-while amount But if the diameter is originally about 1/16 inch, and it is doubled, the inductance is dropped almost 30 per cent.

The other way of reducing lead inductance is shown at the right. We use two or more conductors in parallel with one another. Two paralleled wires of equal diameters and lengths *may* have only half as much inductance as one of the wires alone. Three similar wires in parallel *may* have only one-third as much inductance as one of the wires by itself.

We say that paralleled conductors *may* have much less inductance than a single conductor, because the actual reduction depends largely on how far the paralleled leads are separated from one another. If all the wires are close together, almost touching, the effect is about the same as when increasing the diameter of a single conductor. As the paralleled wires are spaced farther and farther apart, we gain increasing advantage of the paralleling. When the conductors are spaced apart by something like 20 times their diameter they very nearly obey the regular rule for inductances in parallel. That is, the combined inductance then becomes nearly equal to the inductance of one divided by their number.

Some uhf tubes have paralleled internal leads from two separate contact pins to the same elements. Connections for a few such types are shown by Fig. 98-6. The 7E5 lock-in tube has two base pins and paralleled internal leads for plate, grid, and cathode. The 9002 miniature tube has paralleled leads for plate and cathode, with a single grid lead. In the 6AF4 miniature and in the 6F4 and 6L4 Acorn tubes there is a single cathode lead with paralleled leads for plate and grid.

The paralleled internal leads reduce the inductances within the tube itself. When external circuit connections are completed with well separated paralleled wires between the tube and tuning units there is considerable reduction also of external inductance.

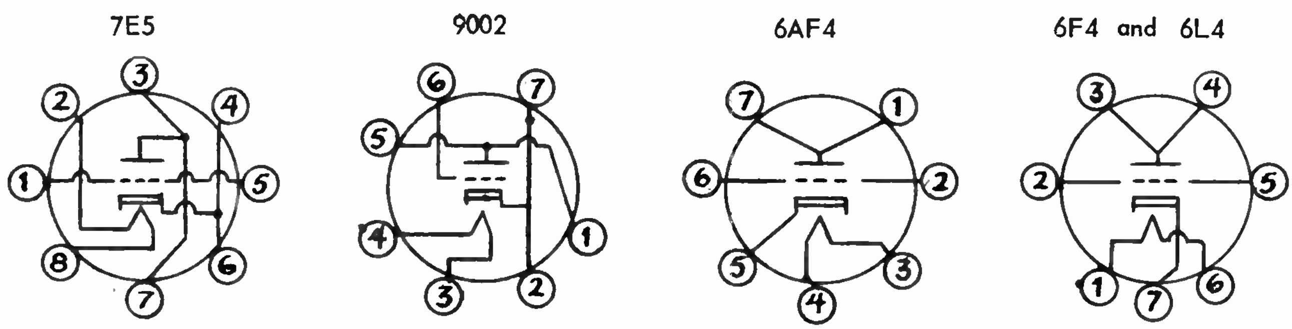


Fig. 98-6. Some of the uhf tubes which have paralleled internal leads for some of their elements.

If a circuit is tuned by a variable capacitor, reducing the inductance by half will raise the maximum tunable frequency by somewhat more than 40 per cent. For instance, were the maximum frequency to be 500 mc with a certain inductance, cutting the inductance to half the original value would raise the maximum frequency by a little above 700 mc.

Reducing the inductance in a capacitor-tuned circuit will not increase the ratio of highest to lowest tunable frequencies. That is, were the original range to be 250 to 500 mc (a ratio of two-to-one), and were the maximum raised to 700 mc, the ratio still would be two-to-one and the minimum would be about 350 mc.

The frequency ratio of a capacitor-tuned circuit can be increased only by getting rid of some of the stray, distributed, and tube capacitances. All these are fixed capacitances, and when they are very large the capacitance variation in an adjustable unit will have relatively little effect on total capacitance. When a circuit is tuned by variable inductance, a reduction of inductance will not only raise the maximum tunable frequency but also will increase the tuning range or ratio.

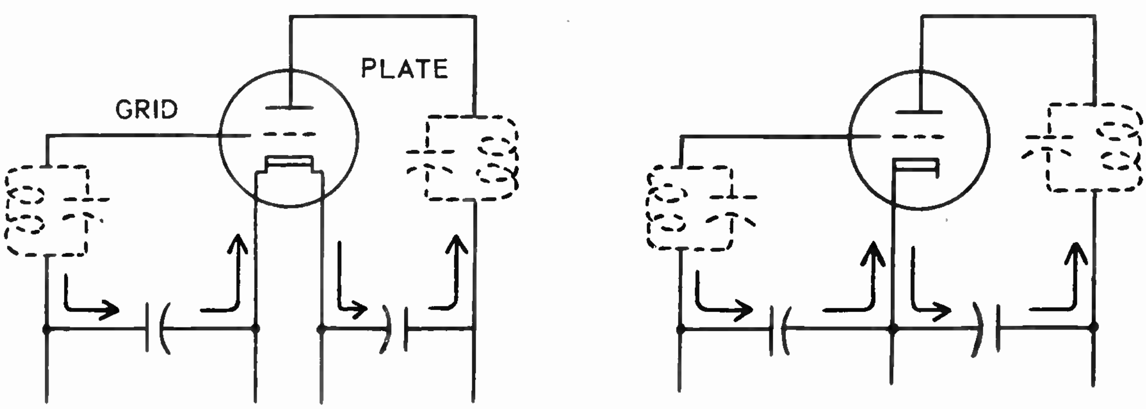


Fig. 98-7. How inductance common to grid and plate circuits is avoided by using two cathode leads.

Some uhf tubes and many of those used at very-high frequencies have paralleled internal leads and two base pins for the cathode. The purpose of two cathode leads is not primarily a lessening of lead inductance, it is to permit completing the r-f grid circuit to one cathode lead and completing the r-f plate circuit to the other cathode lead, as at the left in Fig. 98-7.

Separating the two signal carrying circuits prevents the inductance and inductive reactance of a single cathode lead from appearing in both the plate and grid circuits, as would be the case at the right in Fig. 98-7. Any cathode lead reactance which carries signal voltages in both the plate and the grid circuits acts much like a cathode resistor in causing degeneration and loss of gain. Of course, the two separate cathode connections may also reduce the total circuit inductance, but, as mentioned, this is not the principal purpose of such construction.

TRANSIT TIME. Transit time is the length of time required for electrons to pass from one element to another inside a tube, usually from the cathode to the plate. The greater the plate voltage, with reference to the cathode, the faster the electrons are caused to travel and the shorter becomes the transit time. Transit time is shortened also by designing the tube with less separation between the elements.

Transit time has no important effects at moderate frequencies, such as those used in i-f amplifiers. This fact is illustrated by the diagram at the left in Fig. 98-8. The distance from cathode to plate is assumed to be 8/100 inch. Plate potential is 150 volts at the instant in a signal cycle at which the action is supposed to take place. Transit time from cathode to plate is about 0.0056 microsecond.

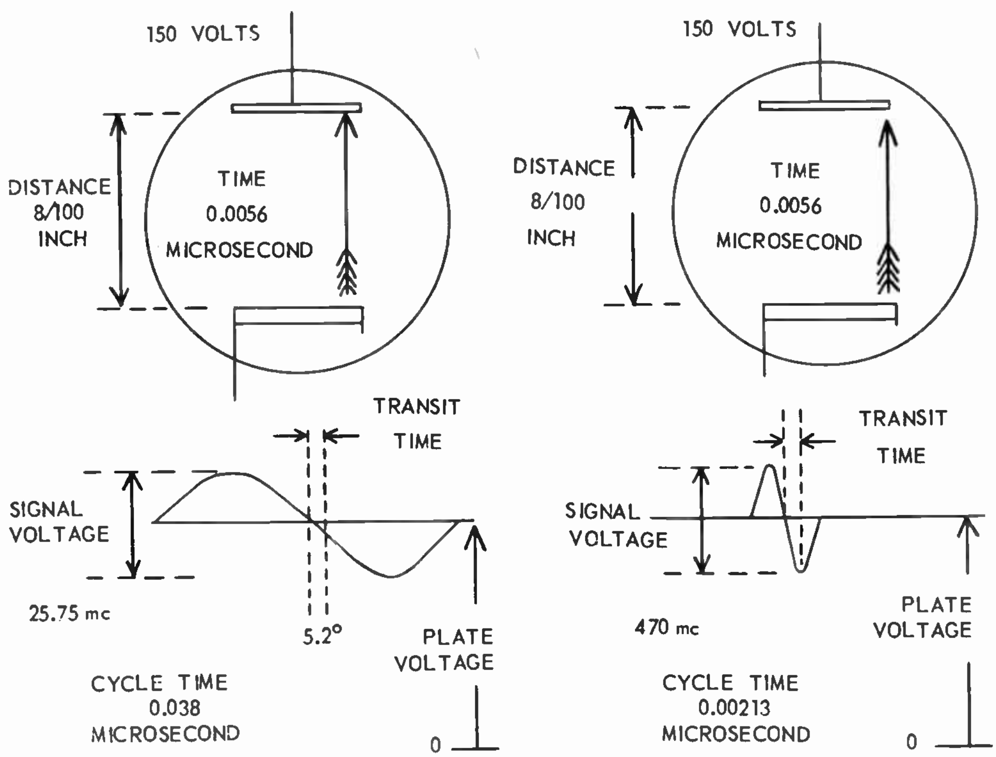


Fig. 98-8. At high-frequencies the transit time takes up very little of a signal cycle, but at ultra high frequencies it may take a large portion of a cycle.

Down below is represented one signal cycle when the frequency is 23.75 mc. The time for one cycle is about 0.038 microsecond. If you divide the transit time by the time for one cycle, and change the fraction to electrical degrees, it will appear that the electrons go from cathode to plate during only about 5.2 degrees out of the total 360 degrees in the signal cycle. During these few degrees the alternating signal voltage on the plate can change very little. This means that electrons will travel at practically uniform speed all the way from cathode to plate.

Now look at the right-hand diagram. Distance from cathode to plate is the same as before, as are also the instantaneous plate potential and the transit time. But signal frequency has been raised to 470 mc. As shown below the tube symbol, one signal cycle now takes only about 0.00213 microsecond, and the transit time is equal to more than 90 degrees of the cycle.

At the ultra-high frequency of 470 mc a bunch of electrons would start from the cathode toward the plate while plate potential is 150 volts. Before these electrons would have time to reach the plate, the signal cycle would have gone through more than 90 degrees and plate potential would have changed by approximately the full signal amplitude.

Because now there are large changes of plate potential during electron transit time, the electrons between cathode and plate are caused to travel at varying speeds during every signal cycle. At the higher speeds the electrons are spread farther apart, and at lower speeds they crowd closer together. It is like the spreading and crowding of contestants on a race track when some of them run faster while others run slower.

During any one signal cycle at the frequency of 21.75 mc all the electrons in the space between cathode and plate are uniformly spread out, as at the left in Fig. 98-9. This is because plate potential remains almost unchanged during each entire signal cycle. But at the ultra-high frequency, the electrons are not uniformly distributed. As shown at the right, there is greater concentration where the electrons have been slowed by drops of signal voltage, and there is less concentration where they have been speeded up by a rise of signal voltage.

What happens is this. Electrons arrive at the plate in bunches rather than at a rate which is uniform during each signal cycle. These electrons form the plate current. Because they do not arrive at the plate in phase with changes of signal voltage on the plate, the plate current gets out of phase with plate voltage. The greater this phase difference the less becomes the signal power output from the tube. When frequency becomes so high that transit time is equal to a half-cycle of the signal, no signal power can be obtained from the plate circuit.

INPUT IMPEDANCES OF TUBES. While studying broad-band high-frequency amplifiers we learned that many difficulties arise because of the small reactances or impedances in grid circuits. We found that reactance due to input capacitance of the tube drops rapidly with increase of frequency, and seriously limits the amplification even when using tubes of high transconductance. At ultra-high frequencies this loss of gain is far greater than at very-high frequencies.

Impedance at the grid-cathode input of the tube will be lowered still further if the cathode lead is part of both the grid circuit and the plate circuit, as at the right in Fig. 98-7. This comes about because inductive reactance of the common lead has much the same effect as a resistance connected across the grid circuit. It is true also that degeneration becomes greater as frequency goes up.

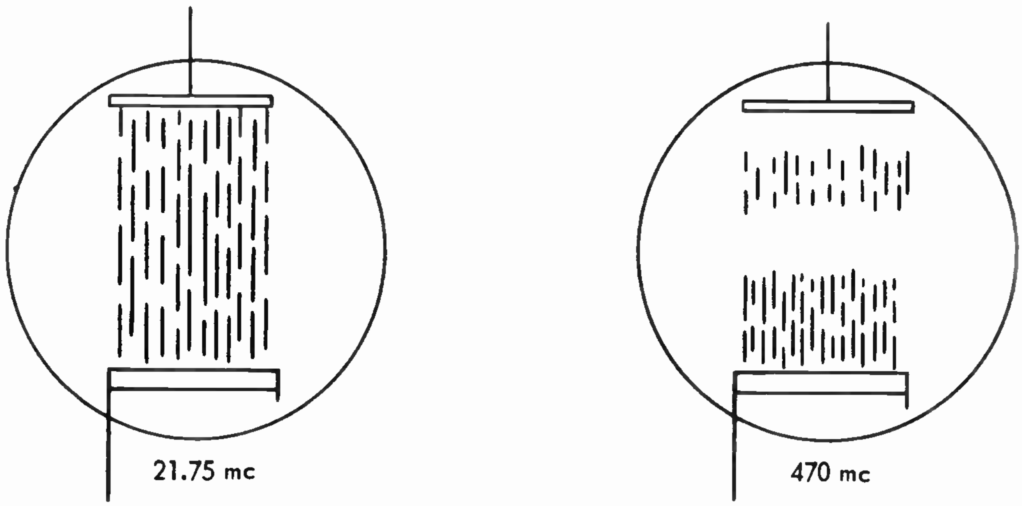


Fig. 98-9. Electrons which are uniformly distributed in the cathode-plate space at high frequencies tend to travel in bunches at ultra-high frequencies.

Next we come to a trouble called electron loading of the grid, which results from transit time effects. If you look back at the diagrams illustrating transit times you will notice that no grid was shown in the tube symbols. Now we shall add the grid. In discussing what happens at the grid you must keep in mind that, at frequencies where transit time becomes a considerable portion of a signal cycle time, the electrons are not uniformly distributed in the space between cathode and plate. This was illustrated at the right in Fig. 98-9.

The diagram at the left in Fig. 98-10 represents conditions during the signal half-cycle in which the grid is made less negative than the bias voltage. A heavy concentration of electrons is arriving at the grid from the cathode, while electrons between the grid and plate are spread more thinly. It is apparent that more electrons are arriving at the grid than are leaving the grid.

The difference between arriving and leaving quantities of electrons must flow through the external grid circuit in the direction shown by arrows. This direction of external flow through any impedance or resistance in the grid circuit is such as to make the grid *more* negative. It is counteracting, to a greater or less extent, the signal voltage which is making the grid *less* negative.

The right-hand diagram illustrates conditions during the opposite half-cycle of signal voltage, during which the signal is making the grid *more* negative than the bias voltage. Now there is a heavy concentration of electrons between the grid and plate. More electrons are leaving the grid than are arriving at the grid. The difference must come through the external grid circuit in the direction indicated by arrows. This direction is such as to make the grid *less* negative, and again we find the external electron flow acting to oppose the signal voltage.

These alternating electron flows in the external grid circuit are maintained only at the expense of signal energy. Part of the signal energy is being used to maintain grid currents instead of being amplified. This "electron loading of the grid" increases as the square of the operating frequency. It would be three to four times as bad at 890 mc as at 470 mc.

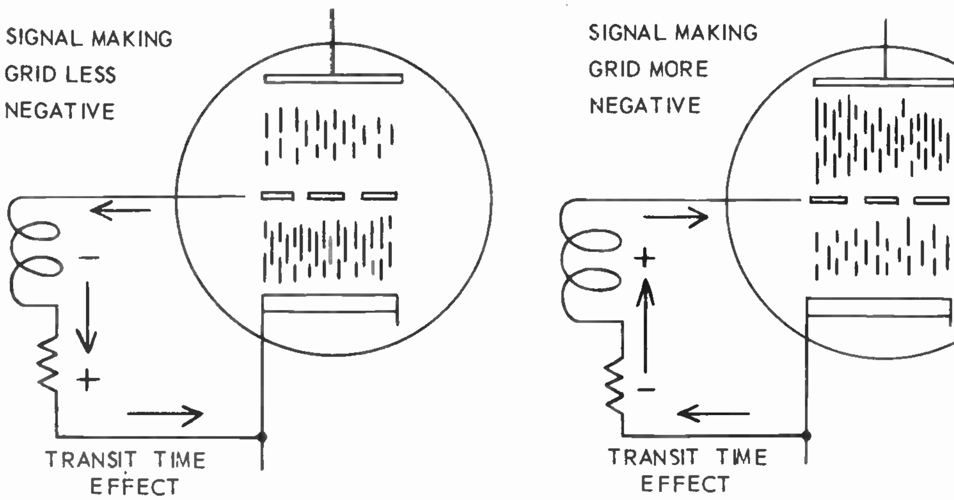


Fig. 98-10. How transit time effect may cause alternating currents to flow in the grid circuit at ultra-high frequencies.

Now for one final difficulty in the tube. In order to have enough voltage gain to overcome all the energy losses it is necessary to use tubes having high transconductance. But in any tube of high transconductance the grid must be made with many turns per inch, and must be located close to the cathode. More grid wires, with very little vacuum space between them and the cathode, will increase the capacitance between grid and cathode. This lessens the input reactance or impedance. It is unfortunate, but always true, that input impedance goes down as transconductance goes up.

Fig. 98-11 pictures an Acorn tube. It is one of the earliest designs developed especially for satisfactory operation at ultra-high frequencies, and is still one of the types best suited for such operation. The element leads in this particular Acorn type do not come out through either end of the envelope, they pass radially outward through a ring-like extension of the glass and are thus so widely separated as to minimize the capacitance between leads. The elements are closely spaced, to reduce transit time. These tubes require a special kind of socket, called a radial socket.

TUNING AT ULTRA-HIGH FREQUENCIES. We learned long ago that any combination of capacitance and inductance is resonant at some certain frequency. If we increase the capacitance while keeping the frequency unchanged it is necessary to reduce the inductance. It is necessary to reduce the capacitance if more inductance is to be used with no change of resonant frequency. It so happens that, for any given resonant frequency, there is only one possible product of capacitance and inductance. This product of capacitance multiplied by inductance is called the oscillation constant for the given frequency.

When measuring capacitance in micro-microfarads and inductance in microhenrys the oscillation constant for 1 mc (1,000kc) is 25,330. A circuit which is to be resonant at this frequency may be made with any value of capacitance and any value of inductance, so long as their product is 25,330. If the product is anything else the circuit cannot be resonant at 1 mc. For resonance at this frequency we might use 2533 mmf of capacitance and 10 microhenrys of inductance, giving a product of 25,330. We might use 10 mmf of capacitance and 2533 microhenrys of inductance, and the circuit would be resonant at 1 mc. We could use

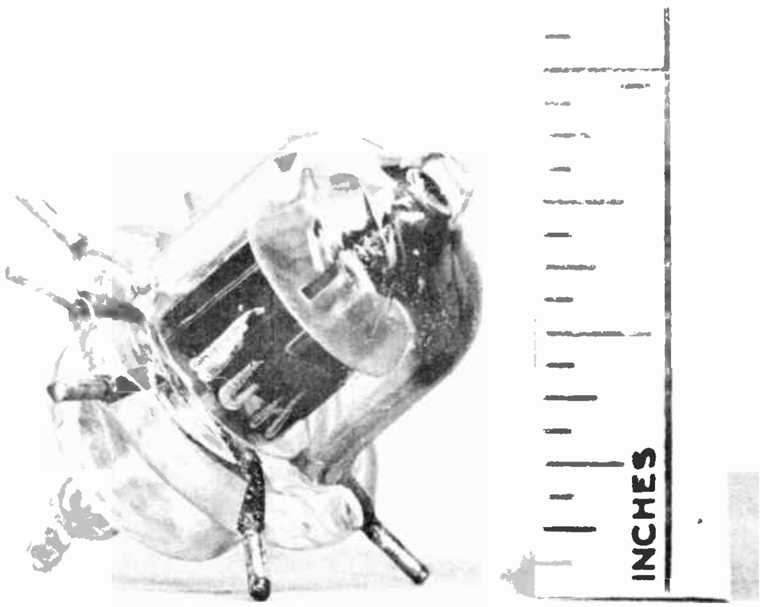


Fig. 98-11. This is a type of Acorn tube which will operate at frequencies as high as 1,200 megacycles.

any other imaginable combination of capacitance and inductance which, multiplied together, give 25,330, and the circuit would be resonant at 1 mc.

The oscillation constant for the ultra-high frequency of 470 mc is 0.1147, and for 890 mc it is 0.0312. It is possible to use any combinations of capacitance and inductance whose products are these oscillation constants, and the circuits will tune to 470 mc and to 890 mc. Both constants are such small fractions that it is plain we shall need very small inductance and very small capacitance in this tuning range.

As we learned earlier in this lesson it is not difficult to obtain small inductances if we employ short connections, large conductors, and maybe use paralleled connections. Supposing, for an example, we were to build a circuit in which the total inductance is 0.0100 microhenry. What capacitances will tune this inductance to 470 mc and to 890 mc? We need only divide the respective oscillation constants by the assumed value of inductance to find the answers, thus.

For 470 mc. $\dfrac{0.1147 \text{ (the constant)}}{0.0100 \text{ (microhenrys)}} = 11.47$ mmf capacitance.

For 890 mc. $\dfrac{0.0312 \text{ (the constant)}}{0.0100 \text{ (microhenrys)}} = 3.12$ mmf capacitance.

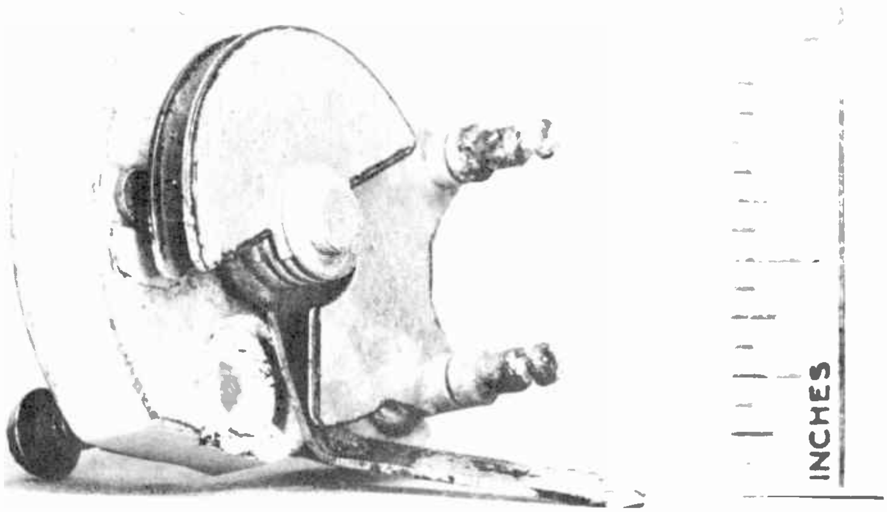


Fig. 98-12. The three-plate capacitor with minimum capacitance of 3 mmf.

The variable capacitor of Fig. 98-12 has minimum capacitance of about 3 mmf and maximum capacitance of about 15 mmf. According to our figures, this variable capacitor, with an inductance of 0.0100 microhenry, should more than cover the tuning range from 470 to 890 mc.

But we have forgotten a few things. To make either an amplifier or an oscillator we need more than capacitance and inductance. We need a tube, and we need connections between the tube and the resonant circuit. The tube will have input capacitance, and there will be stray capacitance in the wiring connections and between all metal parts of the circuit which are separated by air and by insulating supports.

Among the tubes which will operate at frequencies up to 1,000 mc there are some having input capacitances as small as 1.8 to 2.0 mmf. By very careful design and layout it might be possible to keep the stray circuit capacitances as low as 2.5 mmf. The tube capacitance and the stray capacitance will be fixed, they cannot be varied during tuning. Their sum will be at least 4.3 mmf.

Now look at diagram A of Fig. 98-13. With the plates of our variable capacitor all the way out of mesh its capacitance is 3 mmf. To this we must add the fixed capacitance of 4.3 mmf to make a total minimum capacitance of 7.3 mmf. With the plates of the variable capacitor fully in mesh, as at B, its capacitance is 15 mmf. Adding the fixed capacitance of 4.3 mmf gives a maximum tuning capacitance of 19.3 mmf.

When these actual minimum and maximum capacitances are used with the inductance of 0.0100 micro-

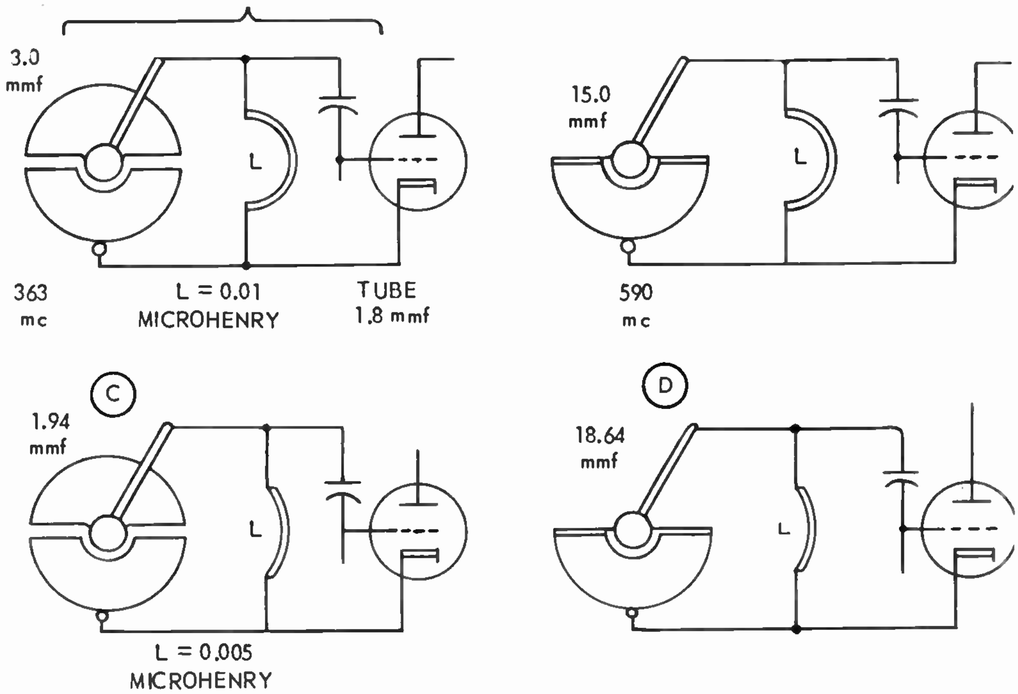


Fig. 98-13. Tuning throughout the uhf television band with any ordinary variable capacitor is a practical impossibility.

henry the circuit won't tune from 470 to 890 mc, it will tune from about 363 mc to 590 mc. This is a tuning range much too low for the uhf television band.

The inductance of 0.0100 microhenry cannot possibly be tuned to 890 mc in a circuit having the assumed tube and stray capacitances. This is evident from examination of our original figures, for the total tube and stray capacitance is 4.3 mmf and we require minimum capacitance of 3.12 mmf. Even with no variable capacitor at all, we still couldn't tune as high as 890 mc with our assumed circuit.

Supposing we could bring the inductance down to half its former value, or to 0.0050 microhenry. What now will happen is illustrated at C and D of Fig. 98-13. Dividing this new inductance into the oscillation constants shows that minimum and maximum tuning capacitances must be 6.24 mmf and 22.94 mmf. From these totals we must subtract the total fixed capacitance of 4.3 mmf in order to find the change needed in a variable tuning capacitor, like this.

Required total capacitance, mmf.	6.24 minimum	22.94 maximum
Fixed tube and stray capacitance.	4.30	4.30
Change in variable capacitor.	1.94 minimum	18.64 maximum

Now we run into another difficulty. No commercial design of variable capacitor having minimum capacitance of less than 2 mmf has a maximum as great as 18 mmf, and none with a maximum of 18 mmf can have a minimum so small as 2 mmf. If there are enough plates to reach the maximum, the minimum remaining between the rotor and stator with the plates all the way out of mesh is too great. If the plates are small enough and few enough for the desired minimum capacitance they won't have enough total surface area to reach the required maximum capacitance.

We might start all over again with a variable inductor for tuning, and use fixed capacitance. The end result would be generally the same. We couldn't cover the entire uhf television band with any variable tuning inductor which could be sold for a reasonable price. If the inductor is small enough to get down to the required minimum it won't be big enough to reach the maximum. If there is enough conductor to reach the required maximum inductance there will be too much to get down to the necessary minimum.

Tuned circuits employing common forms of variable capacitors or variable inductors can be used successfully at the very-high frequencies, up to 300 mc, and also at all lower radio frequencies. At these frequencies most of the circuit capacitance and inductance can be concentrated in a capacitor or an inductor. Capacitors are said to have "lumped" capacitance, because it is concentrated within small space. Inductors or coils are said to have "lumped" inductance, for a similar reason.

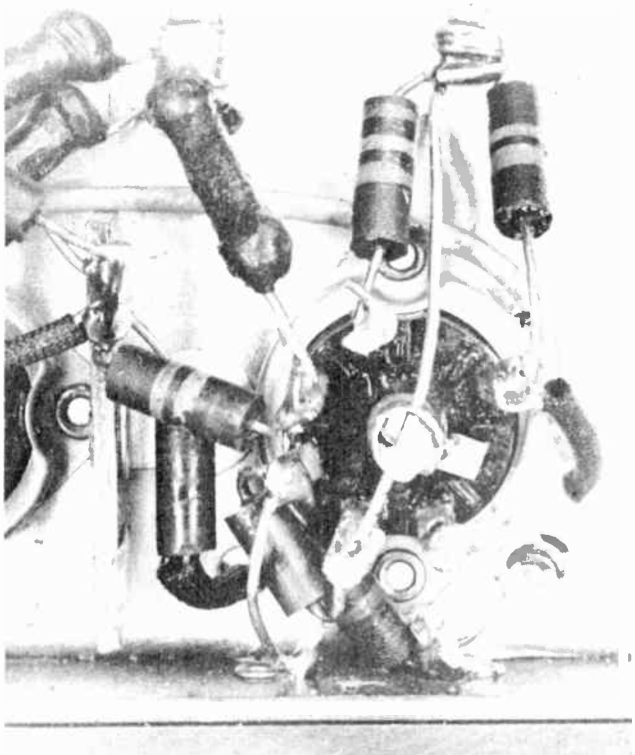


Fig. 98-11. Fixed capacitors, resistors, and r-f chokes are mounted on tube socket lugs or as close as possible to the lugs.

Ordinary lumped capacitances and inductances are difficult or impossible to use for continuously variable tuning throughout the uhf television band. Instead we must resort to circuits in which the capacitances and inductances are distributed instead of lumped, or must use special modifications of lumped elements.

The tuning troubles which have been outlined are those encountered when we attempt to build circuits which can be tuned continuously through the whole range of the uhf television band. On the other hand, it is not hard to build a circuit which will be resonant at any one ultra-high frequency, for then we are not confronted with the difficulties of varying the capacitance or the inductance over wide limits.

As you know, most tuners for the vhf television channels 2 through 13 are made with separate resonant circuits or parts of circuits individually tuned for each channel. This avoids the difficulties of wide-band tuning where there are fixed tube and circuit capacitances and inductances to be considered. This practice is followed also in uhf tuners which are to work in only one channel, or, at the most in a limited number of channels. To extend the practice to a tuner with separate resonant circuits or sections of circuits for all of the uhf channels would be commercially out of the question, in both size and cost. In order to have continuous tuning through all these channels we shall have to play various interesting electrical tricks with resonant circuits.

CIRCUIT PARTS AND WIRING. Certain methods and practices must be followed in all uhf apparatus, no matter what the type of resonant circuit employed. As we have learned, all leads and connections must be very short in order to reduce their inductances. Capacitors, resistors, and r-f chokes are mounted as close as possible to the lugs of tube sockets. These circuit elements usually are supported at one end on the socket lugs, about as shown by Fig. 98-14. When one side of a capacitor, resistor, or r-f choke connects to a tube element, and the other side to ground or to a supply voltage, any extra length of lead should be on this latter side of the unit.

Capacitance, other than used intentionally in resonant circuits, must be kept at a minimum. We must remember that there is capacitance between every metal part and every other metal part from which it is spaced or supported by any kind of insulation, including air. The less the separation the greater will be the undesired capacitance, and the greater the transfer or loss of energy. All parts containing metal should be of small physical size, because the metal acts like the plates of a capacitor and the insulation acts as the dielectric.

All insulating supports and spacers must be considered as dielectric. The only permissible kinds are special low-loss phenolic compounds, ceramics, and plastics such as polystyrene. Regardless of the kinds of material used, there should be the least practicable quantity of dielectric substances, other than air.

Uhf currents and voltages must be confined to the resonant or tuned circuits so far as is feasible. In diagram 1 of Fig. 98-15 the resonant circuits represented by symbols for inductance and capacitance actually may be any of the types to be described later.

High-frequency currents in circuit A return to the cathode through a direct conductor. High-frequency currents in circuit B complete their path to the cathode through bypass capacitor Cb. Passage of these currents into the B-supply portions of the apparatus is opposed by voltage dropping resistor Rb. In diagram 2 the grid return is not made directly to ground and the cathode. This requires that the grid circuit be bypassed to the common ground and the cathode in the same manner as the plate circuit. Separated grounds to different points on the chasses metal are likely to cause unwanted couplings and either regeneration or degeneration.

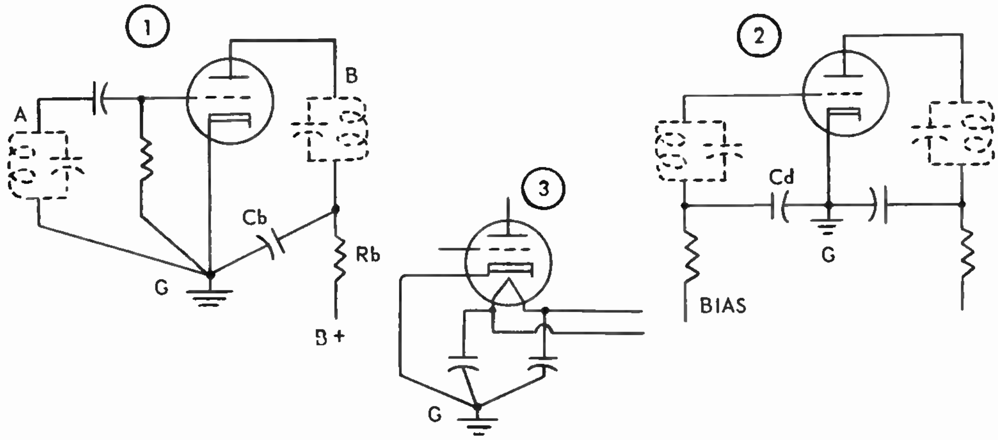


Fig. 98-15. Confining uhf currents to resonant grid and plate circuits with the help of bypass capacitors.

With tubes having heater cathodes it is desirable to keep the cathode and heater at the same r-f potential in order to prevent energy transfer and loss from occuring in the high resistance between these two parts. This may be accomplished as in diagram 3, by connecting bypass capacitors from each heater lead to the cathode or the common ground.

Bypass capacitors for uhf circuits ordinarily need be of values no greater than 100 mmf. Reactance of this capacitance at the low end of the uhf television band is about 3.4 ohms, and at the high end is only 1.8 ohms. The conductors within a capacitor possess inductance, and when a capacitor of large physical size is used at ultra-high frequencies this inductance may act with the capacitance to cause resonance at some frequency within the band. This is one of the reasons for using small tubular ceramic capacitors in most uhf circuits.

Wire wound resistors must not be used in uhf apparatus, other than in power supply circuits, because of the large inductance in such units. All resistors, no matter how constructed, possess some small inductance, and all have some capacitance between their leads or terminals. Instead of pure resistance at ultra-high frequencies there is likely to be capacitive reactance combined with the resistance to provide impedance much lower than the rated resistance value. Trouble is unlikely from resonance effects of inductance and capacitance because the relatively high resistance prevents appreciable peaking.

Voltage-dropping resistors in plate circuits and isolating resistors in grid circuits help to confine uhf currents to the resonant circuits, as illustrated at 1 and 2 of Fig. 98-15. When such resistors are not present the resonant circuits may be isolated with the help of r-f chokes. Often it is an advantage to use chokes in addition to resistors. A number of circuit-isolating r-f chokes are visible on the oscillator pictured by Fig. 98-1.

Fig. 98-16 illustrates several methods of using r-f chokes, all of which are marked RFC. In diagram 1 there is grid-leak biasing of the tube. RFC-1 is in series with grid resistor Rg to oppose flow of uhf currents through this path. RFC-2 is in series with the B-supply lead to oppose passage of uhf currents into the power supply, while an easy path is provided through bypass capacitor Cb to the cathode. RFC-3 and RFC-4 are in series with the heater leads. They oppose escape of uhf currents which otherwise might pass from the cathode through the cathode-heater capacitance to the heater circuit.

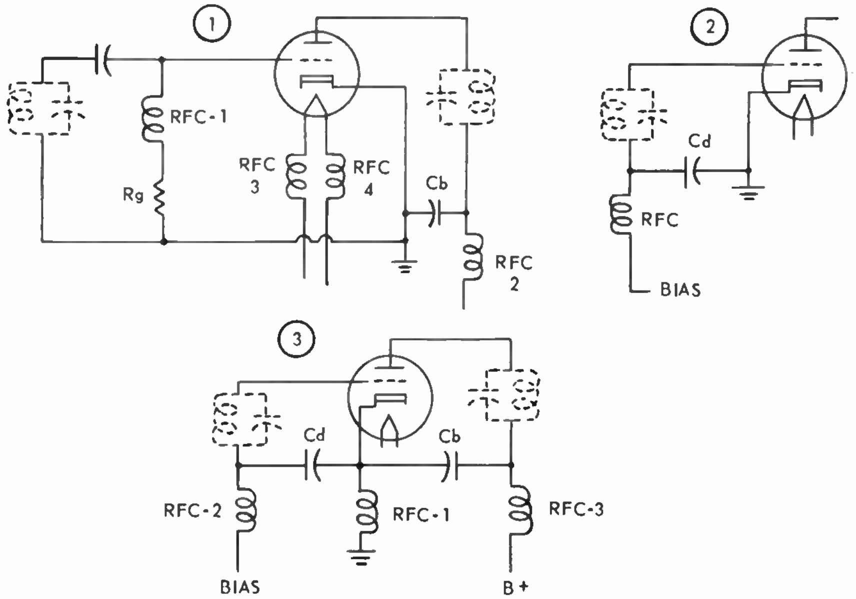


Fig. 98-16. **Isolating the resonant circuits and their uhf currents by means of r-f chokes and bypass capacitors.**

Diagram 2 shows a grid circuit from which there is a connection to a fixed bias voltage. An r-f choke is in series with the lead to the bias voltage, opposing escape of uhf currents to the bias supply and forcing them to take the path through bypass capacitor Cd to the cathode. The remainder of the tube circuit might be as shown by diagram 1.

Diagram 3 of Fig. 98-16 illustrates a method of isolating the uhf grid and plate circuits from chassis ground, while allowing a d-c grid return path and also completion of the B-minus side of the plate power supply through chassis ground. The essential feature is RFC-1 connected between the cathode and ground. The uhf grid circuit is completed to the cathode through bypass capacitor Cd, while being isolated from the bias supply and ground by RFC-2. The uhf plate circuit is completed to the cathode through bypass Cb, while being isolated from the plate supply and ground by RFC-3. Note that all parts of the resonant circuits, including the cathode connections, are isolated from ground and grounded circuits by the r-f chokes.

R-f chokes for uhf circuits usually are wound on tubular or cylindrical forms of low-loss dielectric material. The chokes must have small distributed capacitance, which requires using single-layer, honeycomb, or spaced-turn windings on forms having a low dielectric constant. Inductances usually are between ¼ and 1½ microhenry. The chokes visible in Fig. 98-1 are space wound with bare copper wire on polystyrene rods to have measured inductance of about one microhenry. The reactance of one microhenry at 470 mc is about 2950 ohms, and at 890 mc is about 5600 ohms.

If chokes of identical inductance and construction are used in both the plate circuit and the grid circuit, the two chokes may be self-resonant due to their inductance and distributed capacitance. The chokes and the tube then may act as a tuned-grid tuned-plate oscillator at some frequency in or outside the operating band, and may cause considerable loss of energy.

When two or more r-f chokes are connected to the same tube or in the same amplifying system the several chokes should be positioned with reference to one another so that there is the least possible coupling. Wide separations and positions with axes mutually at right angles will help. Otherwise the chokes should be separated by shielding.

DO NOT TEAR — CUT ALONG THIS LINE AND SEND IN FOR GRADING.

TELEVISION

LESSON NO. 99

TUNERS FOR ULTRA-HIGH FREQUENCIES

Training Your Mind

C.T.I.

Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind
C.T.I.
Perfecting Your Skill

COMMERCIAL
TRADES
INSTITUTE

Chicago, Ill.

LESSON NO. 99

TUNERS FOR ULTRA-HIGH FREQUENCIES

In the design of r-f amplifiers for operation at ultra-high frequencies it is necessary to overcome two major difficulties. First, in order to use a triode we must prevent the oscillation which tends to occur because of feedback through the rather large grid-plate capacitance in this type of tube. This feedback problem may be overcome by using a grounded grid circuit. Second, because uhf signals brought to the r-f amplifier will be very weak, the matter of noise voltages becomes serious. You will recall that signal voltage must be well in excess of all noise voltages if reception is to be satisfactory. Since the uhf signal is bound to be weak, the only solution is reduction of tube and circuit noise voltages. This can be done by using what is called a cascode r-f amplifier.

GROUNDED GRID AMPLIFIER. A circuit arrangement commonly used for grounded grid r-f amplifiers is shown by Fig. 99-1. Antenna signal input is through capacitor Cc to the cathode of the amplifier. Inductor L provides a reactance or impedance across which the signal voltage is developed. This inductor may be varied to suit the frequency band or to help in matching the antenna impedance. Resistor Rk is for cathode bias, and capacitor Ck is the bypass for the bias system.

The grounded grid acts as a shield between the cathode input circuit and the plate outpnt circuit. This shielding so reduces back coupling between output and input circuits as to prevent oscillation. An incidental effect is reduction of radiation from r-f oscillator potentials which otherwise could pass through the r-f amplifier to the antenna. The bias resistor and its bypass may be anywhere in the d-c path between cathode and ground or B-minus. Two alternative positions are shown by the small sketches at the bottom of Fig. 99-1.

Grounded grid r-f amplifiers are used in some vhf television tuners. The tubes sometimes are triodes and again are pentodes connected as triodes. Special grounded grid amplifier tubes, such as the 6J4 triode, are more suitable for uhf circuits. As shown by Fig. 99-2, this tube has three paralleled internal leads and three base pins for the grid. With all three pins connected to the common ground there is minimum inductance and inductive reactance in the leads and connections. Consequently, there can be only negligible reactive voltage developed between the grid and ground, and the shielding effect of the grid is much improved. Maximum capacitance between the plate and the combined cathode-heater is only about ¼ mmf. The heater should be kept at the same r-f potential as the cathode, this being accomplished by placing r-f chokes in both heater leads.

Two triodes or a twin triode may be used in a push-pull grounded grid amplifier circuits as shown by

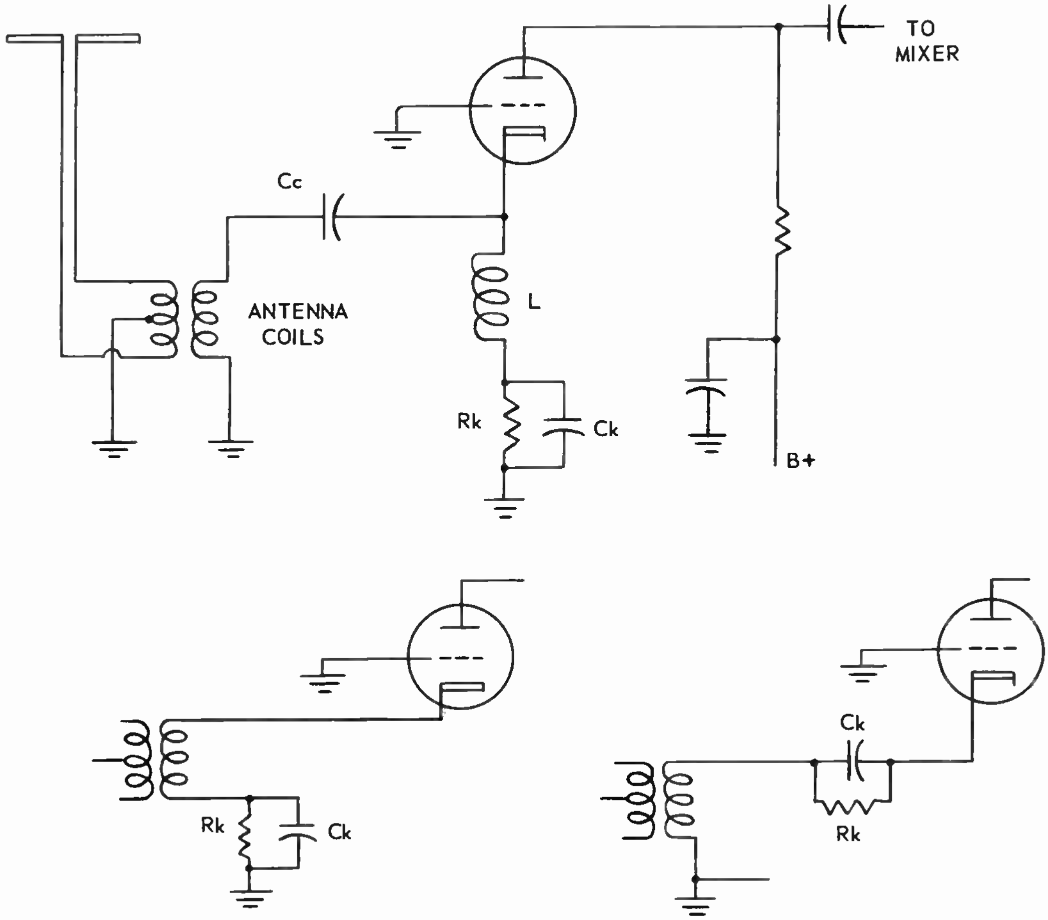


Fig. 99-1. The grounded grid r-f amplifier which may be used at ultra-high frequencies.

Fig. 99-3. The bias connection is made at the center of the tapped winding which is between the two cathodes, while plate voltage is fed to the center tap of the plate coil.

CASCODE AMPLIFIERS. The elementary principle of the cascode amplifier system is illustrated by Fig. 99-4. There are two triodes which, in actual practice, usually are the two sections of a twin triode tube. Signal input to the first section is between its grid and cathode. The plate of this section is directly connected to the cathode of the second section. The second section is a grounded grid amplifier, with signal input to its cathode, with the grounded grid acting as a shield, and with signal output from the plate.

The plate load for the first section is the impedance between the cathode and the grounded grid of the second section. This small load impedance and the resulting small gain in the first section prevent plate-to-grid feedback and oscillation in this section. There will be no oscillation in the second section

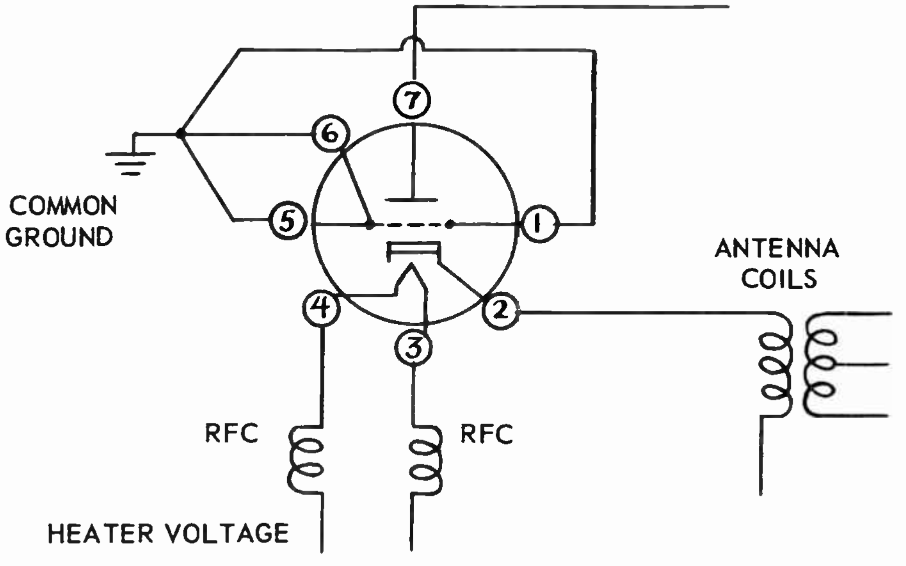


Fig. 99-2. Connections to a grounded grid amplifier tube in which there are three paralleled leads and three base pins for the grid.

because of the grounded grid. The grounded grid section has high mutual conductance. The overall result is an amplifier having the low noise level which is possible with triodes, combined with the high gain and freedom from oscillation such as usually found only with pentodes.

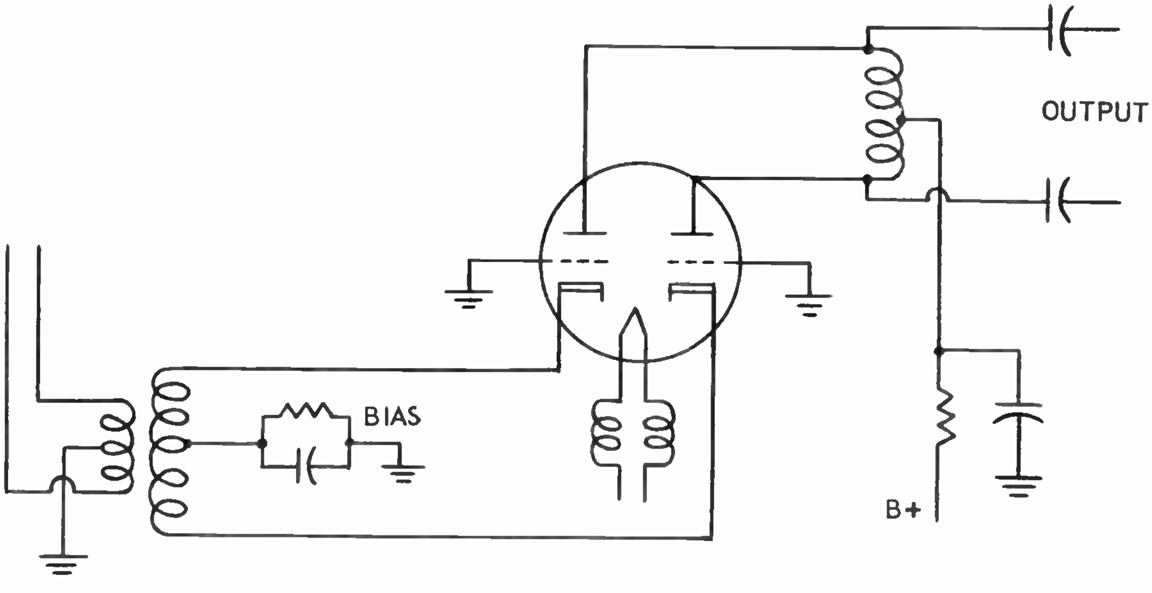


Fig. 99-3. A push-pull grounded-grid r-f amplifier circuit.

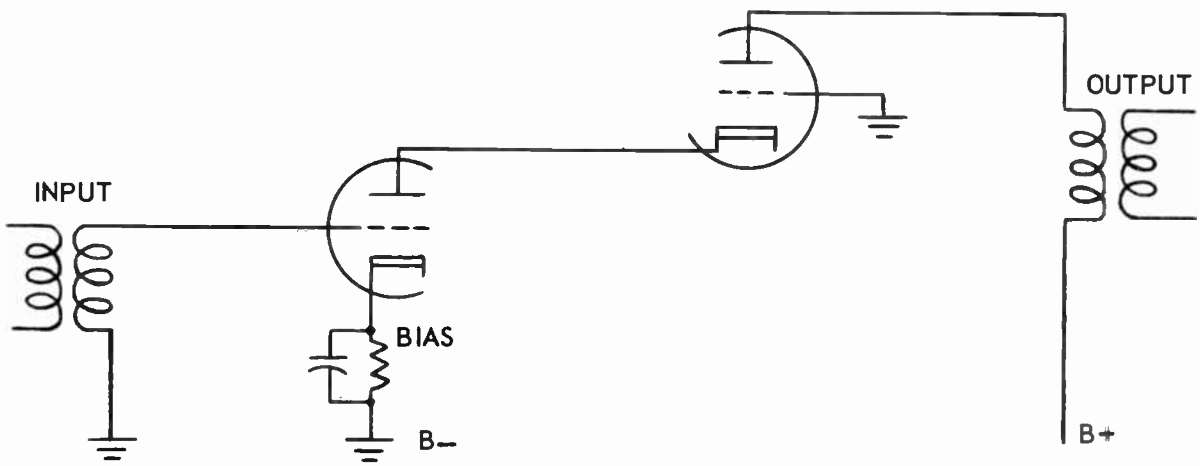


Fig. 99-4. In the cascode amplifier system the plate of the first tube is directly connected to the cathode of the second grounded-grid amplifier.

The two triodes are in series so far as electron flow is concerned. This flow from B-minus of the power supply is through the bias resistor to the cathode of the first section, thence from this cathode to the first plate, and from this plate to the cathode of the second section. The electron flow goes from cathode to plate in the second section, and from the second plate to B-plus of the power supply. The overall B-voltage divides between the two sections. Plate voltages must be measured between plates and cathodes of the respective sections.

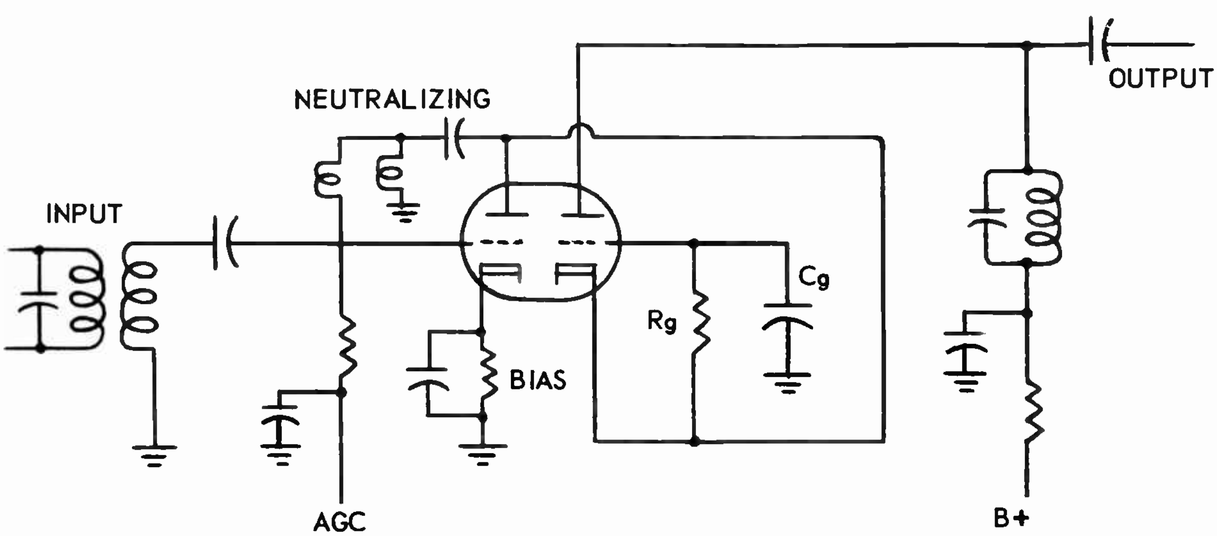


Fig. 99-5. Connections for a complete cascode amplifier system.

Cathode bias is applied only to the first section. Any automatic gain-control voltage is applied only to the grid of the first section. Every change of bias or of grid voltage on this first section changes the plate current in both sections, since they are connected in series on the B-supply. Thus the effect of biasing and of grid voltage on the first section is applied also to the second section. Whatever causes increase of plate current in the first section causes an equal change of plate current in the second section, and every decrease of plate current occurs equally in both sections.

A more complete cascode amplifier system is shown by Fig. 99-5, where the tube is a twin triode. Between the plate and grid of the first section is a neutralizing network which reduces the effect of grid-plate capacitance in this section and the effect of internal coupling between elements of the two sections. Instead of this opposite-phase feedback from plate to grid of the first section, a feedback arranged to be equivalent in phase may be taken from the plate circuit of the second section. This is one of the minor variations found in different applications of the cascode amplifier.

The grid of the second section in Fig. 99-5 is not directly grounded, but is effectively grounded for all r-f currents through Capacitor Cg. This unit usually has capacitance of 1,000 to 1,500 mmf. The d-c grid

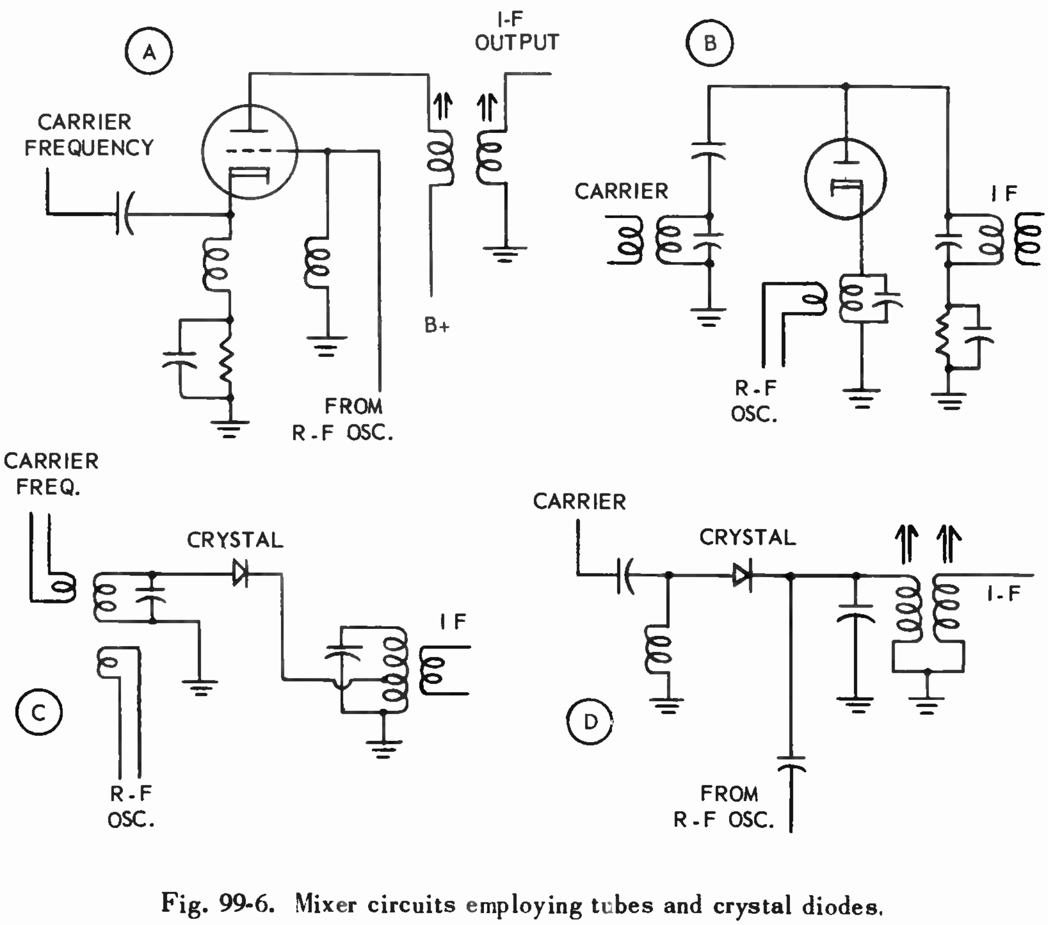


Fig. 99-6. Mixer circuits employing tubes and crystal diodes.

return circuit is completed to the second cathode through resistor Rg, of about one-half megohm in most cases.

Several twin triodes have been designed especially for use in cascode amplifiers, among them the 9-pin miniatures of types 6BQ7 and 6BK7. One of the more noticeable features is a large internal shield between the two sections for the purpose of reducing intersection coupling to a minimum. Although the two plate structures have areas of about 0.14 square inch and are separated by less than 0.2 inch, the plate-to-plate capacitance has a maximum value of only about 0.01 mmf.

The twin triodes just mentioned are suitable for operation at frequencies no higher than about 300 mc, consequently cannot be used as r-f amplifiers at ultra-high frequencies. They are, however, widely employed for amplification of the mixer output in uhf tuners. Thus operated at a fixed frequency these tubes allow high gain and an exceptionally low noise level.

1 *UHF MIXERS.* The mixer in a uhf tuner may be a triode tube or it may be either a diode tube or a crystal diode. There are many possible circuit arrangements for any of these elements. The essential feature is that carrier signal voltage and r-f oscillator voltage be brought together in a tube or crystal which acts as a rectifier. Then, as with any mixer circuit, the output will contain sum and difference frequencies, of which one will be the desired intermediate frequency.

One type of triode mixer circuit is shown at A in Fig. 99-6. Carrier frequency voltage is applied to the cathode, in whose lead to ground and B-minus there is an r-f choke in series with the bias resistor and its bypass capacitor. Voltage from the r-f oscillator is applied to the grid. At B is a simple mixer circuit employing a diode tube. Carrier-frequency voltage is applied to the diode plate while voltage from the r-f oscillator is applied to the cathode.

Any mixer tube must be capable of operating at ultra-high frequencies. Triodes of types suitable for uhf amplifiers or oscillators make satisfactory mixers. There are a few diode tubes which will be satisfactory at ultra-high frequencies. With others their capacitances and inductances in elements and leads become self-resonant in the uhf television band. For example, the familiar 6AL5 twin diode will be self-resonant somewhere around 700 mc.

2 There are crystal diodes of both germanium and silicon types which have been especially developed for operation as uhf mixers. In this class are the 1N72 germanium diode and the 1N82 silicon diode. In all the crystals there is a signal energy loss of 60 to 70 per cent during the conversion process, which makes it necessary to follow these mixers with an r-f amplifier stage. This amplifier usually is a cascode type. Crystal mixers have the advantages of freedom from transit time effects and of operation at low noise levels. Typical crystal mixer circuits are shown at C and D of Fig. 99-6.

Fig. 99-7 shows at the left a uhf mixer crystal of the 1N72 type. At the right is a type 9006 diode tube which will operate satisfactorily at frequencies up to about 1,400 mc. This tube is in an envelope even smaller than that used for most of the miniature tubes.

UHF OSCILLATORS. There are many oscillator tubes which will operate at frequencies all the way up to 30,000 mc, although most of the types designed for the upper end of the ultra-high frequencies and for the super-high frequencies would be too costly for use in commercially priced tuners and receivers.

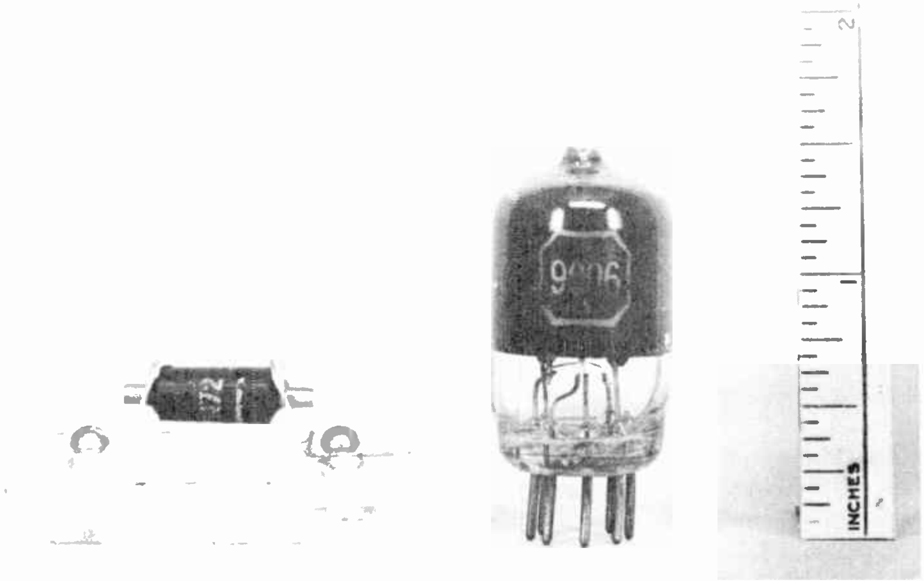


Fig. 99-7. A crystal diode designed for uhf mixing, and a diode tube which operates satisfactorily at ultra-high frequencies.

Fig. 99-8 is a cut-away view of a magnetron in which electrons are deflected by a strong magnetic field in their travel from cathode to plate. The complete resonant circuit is built into the tube structure. These units seldom are made for work below 2,000 mc, and go from there to the top of the super-high frequency range. Any one tube has only a limited range of frequency adjustment.

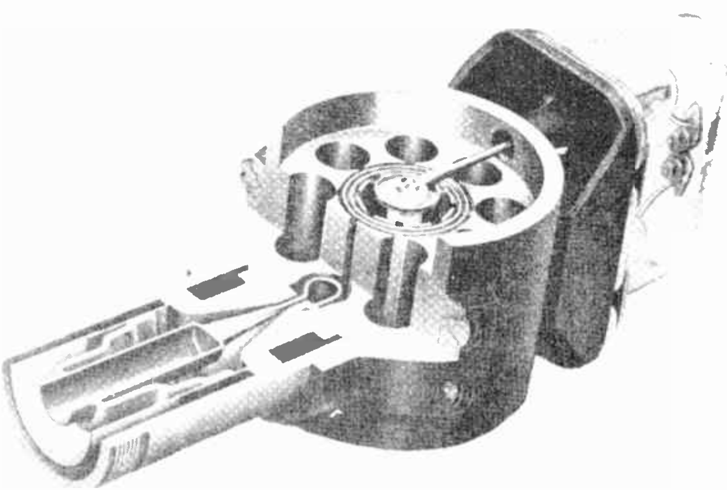


Fig. 99-8. Construction of a magnetron oscillator used at super-high frequencies.

Another tube used as either an amplifier or oscillator in bands from around 1,200 mc. all up through the super-high frequencies is the reflex klystron. This type operates by virtue of the electron bunching which results from transit times that are long in relation to signal cycles. Of more interest in our present work are the miniatures and the Acorn types which are designed for the uhf television band.

One of the principal difficulties with uhf oscillators is frequency drift. Unless such drift is reduced by effective methods it may be necessary to retune a receiver several times during the first 10 to 20 minutes of operation. In the middle of the uhf television band, at 680 mc, a change through the frequencies of one entire channel means a drift of only about 0.9 per cent in frequency, and a drift half way through the range of video frequencies (about 2 mc) would require a drift of only about 0.3 per cent. At the top of the uhf band, 890 mc, all the frequencies in a channel extend over only about two-thirds of one per cent of the operating frequency, and between the video carrier of one channel and the sound carrier of an adjacent channel the difference is only about one-sixth of one per cent.

Oscillator frequency is altered by any variation of tube voltages. The result is reduction of oscillator power output, a change in relation of harmonic frequencies to the fundamental, and a change of internal reactances due to alteration of the internal capacitances and inductances of the tube. The most effective remedy is to maintain nearly constant plate and grid voltages with the help of voltage regulating tubes or voltage regulating power transformers, although either of these methods adds quite materially to the cost of a receiver

Changes of temperature causes circuit parts to expand or contract. In inductors of all types, including resonant lines, a rise of temperature nearly always causes the operating frequency to decrease. The effect is increased by increase of skin effect in all the conductors. The obvious remedy is to employ temperature compensating capacitors in the oscillator resonant circuit.

Sudden variations of oscillator frequency may occur due to vibration when the mechanical design and construction is not rigid and free from parts which are unduly elastic. There may be gradual changes of frequency due to the use of dielectric materials which shrink or warp with age. High grade ceramics suffer the least from such effects. Ultra-high frequencies may be varied quite materially even by changes of atmospheric humidity in warm weather.

It is essential that all parts of the oscillator system be well shielded in order to prevent direct radiation not only within the tuner or receiver but to other nearby receivers.

When the r-f oscillator is capable of delivering much more output energy than actually fed to the mixer the loading of the oscillator has little effect on operating frequency. If, however, the mixer takes a considerable part of the possible oscillator output, any changes in the other circuits connected to the mixer will alter the oscillator frequency. Under such conditions the frequency will be more stable when the oscillator resonant circuit contains large capacitance in relation to inductance. This makes it practicable to track the frequency of the r-f oscillator with the frequency picked up in the channel selector by using very small adjustable capacitors for both ends of the band. The low end of the band may be tracked with a capacitor in parallel with the resonant circuit, and the high end with another capacitor in series with the grid or between the grid and the resonant circuit.

TUNED CIRCUITS. There are numerous ways in which small inductances, capacitances, or both may

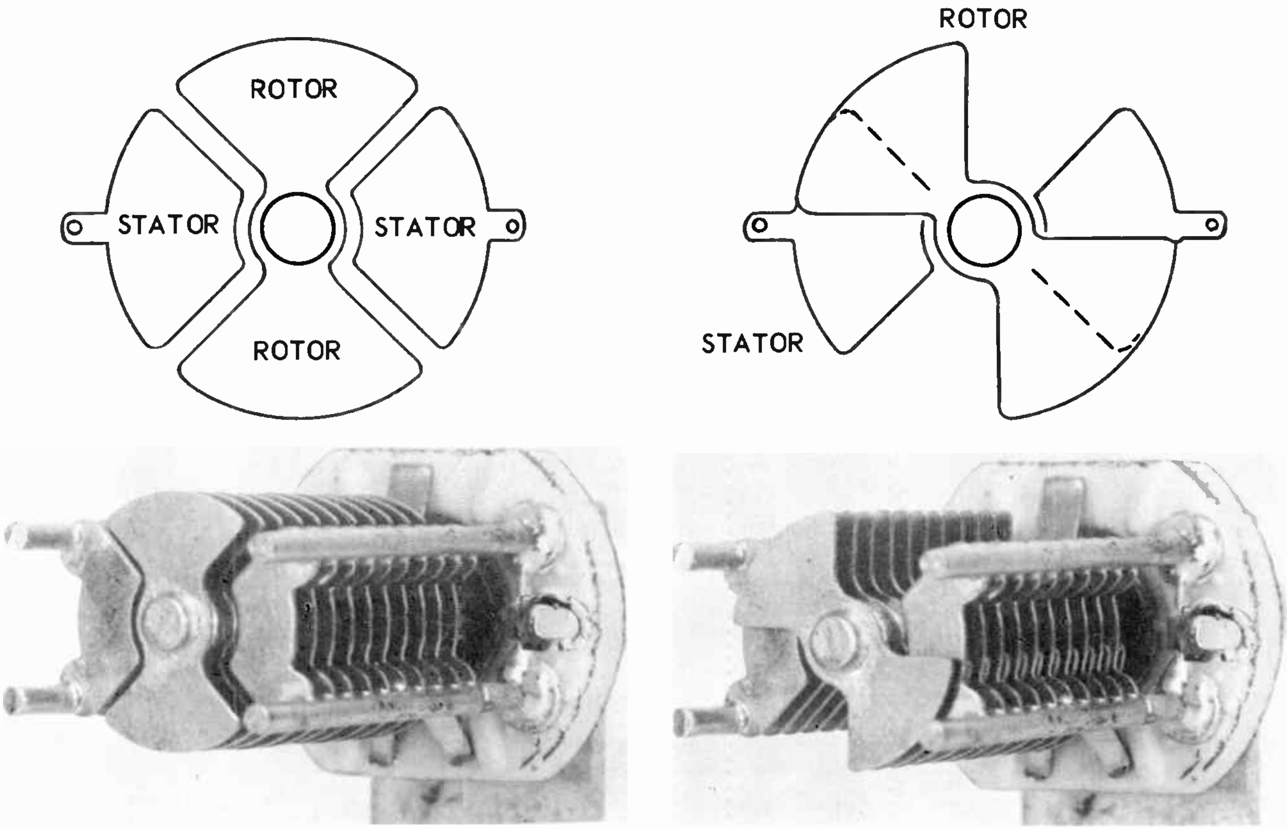


Fig. 99-9. A butterfly capacitor adjusted for minimum capacitance and for about half the maximum capacitance.

be varied in order to tune a circuit to resonance through a range of frequencies. For example, the total inductance of paralleled conductors may be altered by connecting a greater or less number of such conductors into the resonant circuit. A single conductor will provide maximum inductance, two similar conductors in parallel will drop the inductance to approximately one-half, three conductors will drop it to about one-third, and so on.

Another way of reducing the inductance of paralleled conductors is to move strips or vanes of metal into the spaces between the conductors. The greater the portions of the spaces which are taken up by the movable vanes, the less becomes the inductance. Inductance is reduced because less space remains for magnetic lines of force around the active inductors.

When an inductor is in the form of a small loop or part of a circle, moving a copper plate across the face of the loop, but not making contact, will reduce the inductance. Although these and generally equivalent methods of altering the tuning inductance sometimes are employed in uhf apparatus, none of them used alone give enough change for tuning through the entire uhf television range without dividing the channels into several bands and employing band switching.

Tuning capacitors of ordinary construction may be designed to have small minimum and maximum capacitances, but they have the objectionable feature that uhf currents must flow in the sliding contacts of the rotor shaft bearings. This objection is overcome in a design called a butterfly capacitor. The principle of the butterfly capacitor is illustrated by the drawings at the top of Fig. 99-9, while actual construction of a typical unit is shown i, the pictures down below.

The butterfly stator consists of two opposite positioned and separately insulated groups of plates, each plate having the approximate form of a quarter circle. There is no conductive connection between the two sections of the stator. The rotor consists of plates having the form of two quarter-circles joined at the center and mounted on the tuning shaft. It is the shape of these rotor plates that gives the butterfly capacitor its name.

The ends of a circuit with which the butterfly capacitor is to be in series are connected to the two opposite stator sections. Alternating currents pass into one stator section and out of the other stator section, and from each stator section there is capacitance to the common rotor. No alternating currents have to pass into and out of the rotor. There is minimum capacitance with the rotor plates all the way out of mesh with the stators, as at the left, and capacitance is increased as the rotor plates are moved into mesh with the stators, as at the right.

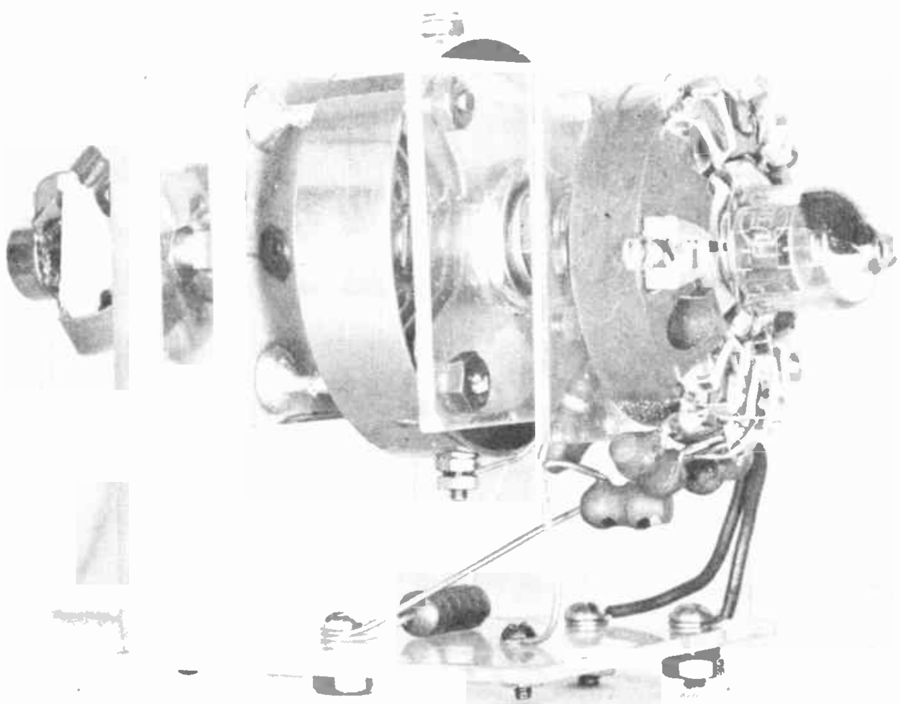


Fig. 99-10. A butterfly oscillator which is designed to operate at very-high frequencies and in the lower range of ultra-high frequencies.

A

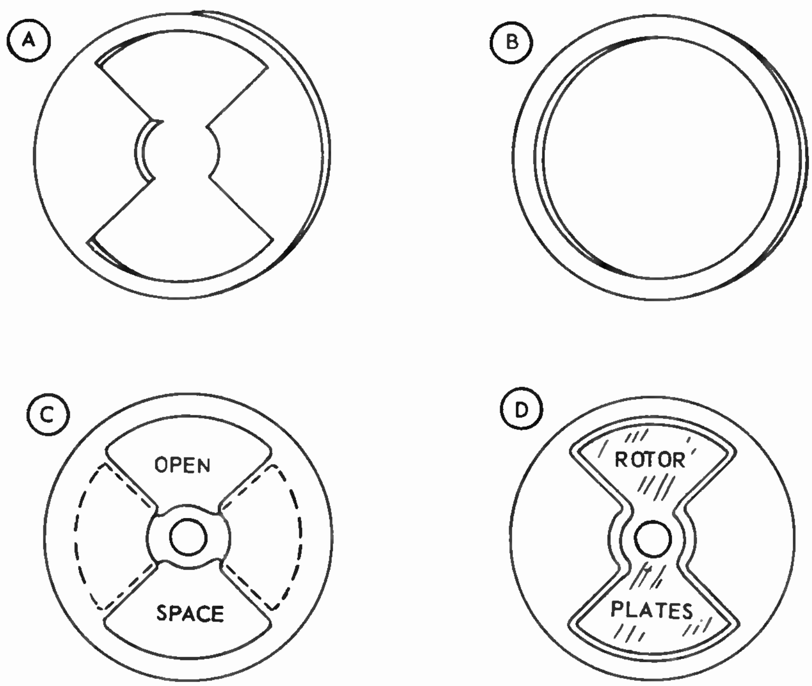


Fig. 99-11. The plates and spacers for the stator portion of a butterfly tuner.

The rotor of the butterfly capacitor usually is connected to ground, which tends to lessen the ill effects of body capacitance during tuning. When this type of capacitor is used in the manner described, with the two stator sections in series, the effective capacitances are half those at which the unit is rated. For instance, a butterfly capacitor having ratings of 4 mmf minimum and 16 mmf maximum capacitance will have series values of 2 mmf minimum and 8 mmf maximum.

BUTTERFLY TUNER. The butterfly capacitor and inductance variation by change of magnetic field space are combined in a most ingenious manner to form the butterfly tuning circuit. Fig. 99-10 is a picture of an experimental oscillator of this general type, constructed with a commercial type of butterfly capacitor and with the necessary parts added. The tuning range is approximately from 200 to 460 mc. Oscillators employing the butterfly tuning circuit, but made with parts designed for the particular purpose, have tuning ranges as great as from 250 to 920 mc, without band switching.

The stator assembly of a butterfly tuner is made with a number of platesshaped as shown at A in Fig. 99-11. Adjacent plates are spaced apart with metal rings of the form shown at B. The plates of a butterfly rotor of ordinary form rotate in the spaces between these stator plates, without, of course, touching the stators. The outside of the completed stator assembly consists in effect of a metal ring with inward extensions, but with a large air space inside the ring. It is this ring that provides inductance for the tuned circuit. Variable tuning capacitance is provided by the stator and rotor plates, with air between them for the dielectric.

When the rotor plates are turned all the way into mesh with the stators, as represented at C, the capacitance of the butterfly capacitor is maximum. It is to be noted especially that, with the rotor in this position, there remains maximum clear air space inside the inductor ring. Consequently, there is room for maximum magnetic field and there is maximum inductance. Now we have maximum inductance combined with maximum capacitance, and resonance will occur at the lowest frequency.

At D the rotor plates are shown all the way out of mesh with the stators. In this position of the butterfly there is minimum capacitance. The metal of the rotor plates now almost fills the space within the outer ring of the stator assembly. Little clear space remains for magnetic lines of force, and inductance is reduced to its minimum. Now we have the combination of minimum inductance and minimum capacitance, and resonance will occur at the highest frequency.

In most tuning circuits either the capacitance or the inductance is reduced to raise the resonant frequency, and either of these factors is increased to lower the resonant frequency. The frequency range is limited by the possible change of either capacitance or inductance. In the butterfly tuning circuit there is simultaneous reduction of both capacitance and inductance, and simultaneous increase of both factors. This accounts for the wide frequency range of the butterfly tuning circuit.

Examination of Fig. 99-11 shows that the full tuning range is covered by moving the rotor one-quarter turn, as from the position at C to the position at D. This is true also of the simple butterfly capacitor illustrated by Fig. 99-9. Inductance of the butterfly tuner is decreased by using a thicker outer ring, also by using a wider ring together with the added stator plates that go with extra width. The inductance varies also approximately as the square of diameter of the outer ring. Inductance of the ring is much less than that in a similar turn of a coil, because the ring is a shorted turn and also because so much of the space within the ring always is taken up by the stator and rotor plates even when these are all the way in mesh.

The manner of making circuit connections to the butterfly tuner is illustrated at A in Fig. 99-12, where a simple ultraudion oscillator is used as an example. Two points on opposite sides of the outer ring, in line with the centers of the stators, are equivalent to the opposite ends of an ordinary parallel resonant circuit consisting of a coil and capacitor. The plate of the oscillator tube is connected to one of these points, either one might be used, and the grid is connected to the other point.

Midway between the two opposite points which correspond to the ends of a parallel resonant circuit is a point which is equivalent to a center tap on a coil, and to this point is connected the B-plus lead of the oscillator circuit. There is another neutral point or equivalent center tap on the opposite side of the ring, at b, and the B-supply lead might be connected there instead of at the bottom.

Energy at the oscillation frequency may be taken from the butterfly tuner with a small loop for inductive coupling placed near the ring as at B. This coupling is made preferably between the plate connection and either of the neutral points. The entire oscillator circuit is electrically equivalent to the more familiar form shown at C. The butterfly tuner circuit may, of course, be used for applications other than oscillators. The capacitor and stator ring assembly, also an inductive coupling, may be used in any manner that the parallel resonant circuit of diagram C might be used.

RESONANT LINES FOR TUNING. When we studied transmission lines and resonant lines or stubs in

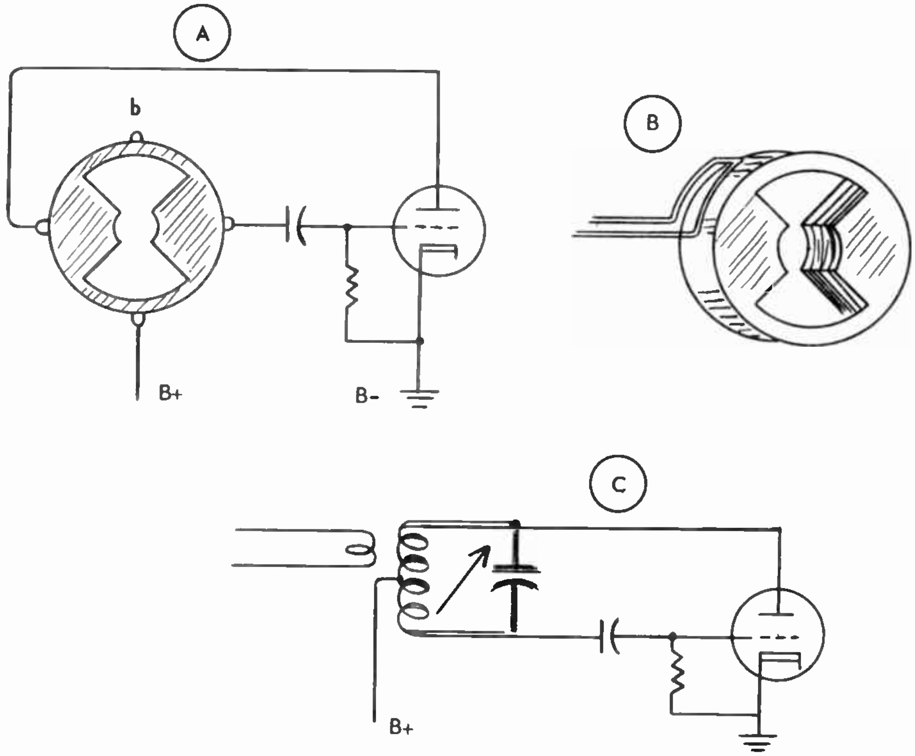


Fig. 99-12. Connections for a tube and an output coupling to a butterfly tuner, and the equivalent circuit.

earlier lessons it was stated that resonant lines are important in ultra-high frequency circuits. These lines, which are tuned by varying their length, are used in uhf tuners for antenna circuits, mixer circuits, and oscillator circuits. Fig. 99-13 is a picture of a uhf tuning assembly in which adjustable shorted lines are used in all three circuits.

Doubtless you will recall that resonant lines, either open or shorted at their far ends, are electrically equivalent to either series resonant circuits or to parallel resonant circuits when the lines are of certain fractions of a wavelength at the desired tuned frequency. Since tuning practically always is accomplished with parallel resonant circuits, we must use resonant lines which have the properties of parallel resonance.

The properties of parallel resonance are possessed by a half-wave open line and also by a quarter-wave shorted line. Were we to use a half-wave open-ended line it would be difficult to vary the effective length during tuning by any simple mechanical means. It is easy to vary the effective length of a shorted line by sliding a shorting conductor or bar from point to point, thus giving a continuous variation. As a consequence, we always use quarter-wave shorted lines for tuning. There is the further advantage that a quarter-wave line for any minimum frequency is only half the physical length of a half-wave line for the same frequency.

In the structure of Fig. 99-13 the shorting brushes, clearly visible, are carried on rotor arms which are

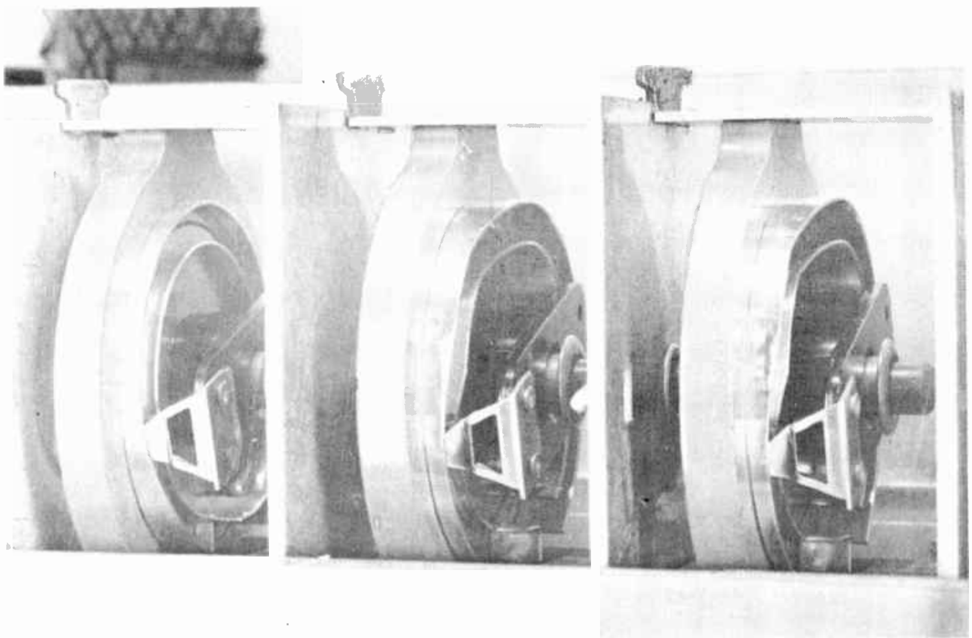


Fig. 99-13. A uhf tuner employing resonant lines which are shorted at various lengths by adjustable contact fingers.

turned by the tuning shaft. Each of the three resonant lines, for antenna, mixer, and oscillator, consists of two parallel conductors formed into nearly complete concentric circles to make the tuner more compact than were the lines to be straight. At one end of each pair of conductors are the terminals for connected circuits. The other ends of the lines, so far as tuning performance is concerned, are formed by the movable shorting brushes.

Quarter-wave shorted lines may be used at ultra-high frequencies in any of the ways that resonant circuits with lumped elements are used at very-high frequencies and at all lower radio frequencies. Some examples are illustrated by Fig. 99-14. At A the grid circuit of a triode is tuned by means of a shorted line. The effective length of the line is varied by moving the shorting bar. At the grid-cathode end of the shorted line there is maximum impedance at a frequency for which the effective or unshorted length of the line is equal to a quarter wavelength. This is just what happens when an ordinary parallel resonant circuit is tuned to some certain frequency.

The quarter-wave tuning line at A performs in the same manner as the lumped-element tuning circuit at B. The two sides of the resonant line and the shorting bar take the place of the familiar tuning coil and variable tuning capacitor. All other parts of the two circuits are similar.

A plate circuit might be tuned with a quarter-wave shorted line as shown at C in Fig. 99-14. This arrangement is equivalent in performance to the coil-capacitor tuning circuit shown at D. Both the grid circuit and the plate circuit could be tuned to the same or different frequencies by combining the resonant line arrangements of diagrams A and C with a single tube.

The portion of a shorted quarter-wave line which is on the far side of the shorting bar has no effect on the frequency of resonance, and so far as electrical performance is concerned this portion of the line could be removed. As the shorting bar is moved toward the tube, all of the line on the far side of the short becomes electrically dead. The far ends often are connected together, as at A in Fig. 99-14, and used for any d-c connections to the tube elements. In the circuit shown at C the B-plus connection might be made to the far ends of the line conductors instead of to the plate of the tube.

When a shorted line is connected to the elements of a tube there is maximum voltage and maximum impedance across which the voltage is developed at the tube end of the line. At the point of shorting there is maximum current and minimum voltage. All this is as described in our earlier study of resonant lines. In a parallel resonant circuit consisting of a coil and capacitor there is maximum impedance at opposite ends of these two elements at resonance, but at the same time there is maximum current circulating back and forth between the coil and capacitor. This we learned when first becoming acquainted with the behavior of resonant circuits.

Resonant line tuning as commonly used for oscillators operating at ultra-high frequencies is illustrated by Fig. 99-15, with the electrical equivalents in the form of lumped capacitances and inductances. With

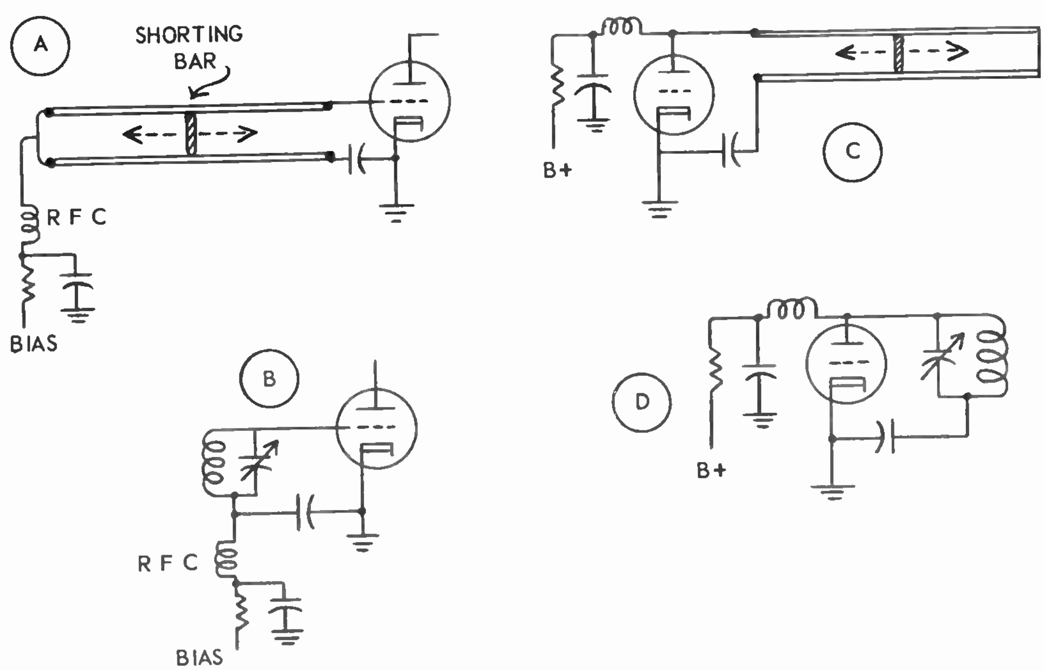


Fig. 99-14. How quarter-wave shorted lines may be used for tuning grid circuits and plate circuits.

the circuit shown at A, and its equivalent at B, there is biasing by the grid-leak method and there is a connection from the plate of the tube to the B-voltage supply. Were the tuning capacitor omitted from diagram B the circuit would be recognized as the common form of Colpitts oscillator.

In diagram C we again have grid-leak bias, as used for nearly all oscillators, but here the plate voltage and current are fed to the outer ends of the quarter-wave shorted line. When the equivalent circuit is made up as at D it often is referred to as an ultraudion oscillator.

At 470 mc, the bottom of the uhf television band, a quarter-wavelength in air or free space is about 6.28 inches long. At 890 mc, the top of the uhf band, a quarter wave is about 3.32 inches long. To tune from the lowest to the highest frequency in this band the shorting bar on a quarter-wave line would have to be moved the difference between these distances, which is about 2.96 inches.

As we learned when studying transmission lines, the actual length of lines of ordinary construction is somewhat less than the length in free space. This is because transmission lines ordinarily have insulation, which brings in the velocity factor. Quarter-wave tuning lines are made with bare conductors, but in order to have mechanical stability these conductors must be mounted on or in some kind of insulating material. This makes the physical length of the line less than it would be were there no supports. Therefore, the shorting bar does not have to be moved quite as far as 2.96 inches to cover the uhf television range.

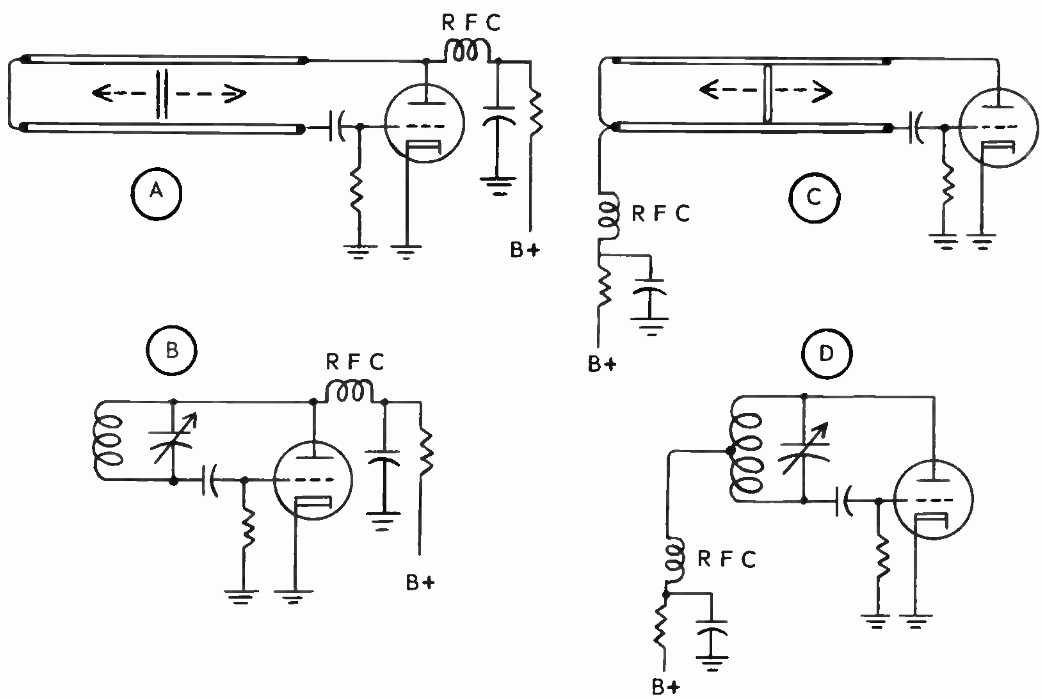


Fig. 99-15. How an oscillator may be tuned by means of a quarter-wave shorted line.

The actual line length is still further shortened by the effects of capacitances and inductances in the tube and connections. Were there no capacitances and inductances, the distance from the tube terminals or element terminals would be a quarter-wavelength at the tuned frequency. That is, when tuned to 890 mc the shorting bar would be about 3.32 inches from the tube elements. Actually the bar must be brought much closer to the tube elements The distance from the socket lugs to the tube end of the resonant line conductors usually is no more than an inch, or it may be less. The end of the resonant line which is toward the tube or between the shorting bar and the tube, includes not only any wiring connections, but also the socket lugs, the tube base pins, the internal leads of the tube, and any coupling or blocking capacitors which are found along this path.

The effective length of a shorted line is varied by altering the capacitance of any capacitors which are in series with the line as shown by diagrams in Figs. 99-14 and 99-15. This applies to capacitors which are on either the grid side or the plate side of the circuit. These series capacitors often are of adjustable types, to allow alignment of the tuned circuits during manufacture and servicing. The effective length of a resonant line may be altered also by connecting an adjustable capacitor between the two line conductors. When such a capacitor is connected near the tube end of the line it will have a large effect on effective line length and tuning. Such paralleled capacitors are used for alignment in many uhf circuits.

In preceding diagrams which have shown the principles of resonant line tuning, all the circuits have been made as simple as possible. Actually these circuits must include most or all of the features required by ultra-high frequency practice in general. This is true regardless of the method of tuning, whether with resonant lines, butterfly circuits, or anything else.

At A in Fig. 16 are shown connections such as used in a typical oscillator circuit. There are r-f chokes in both heater leads. One side of the heater circuit is here completed through chassis ground, as is common practice. There is an r-f choke between the cathode and ground, and another between the plate and the B-supply. The grid returns and plate returns for uhf currents are made to a common ground.

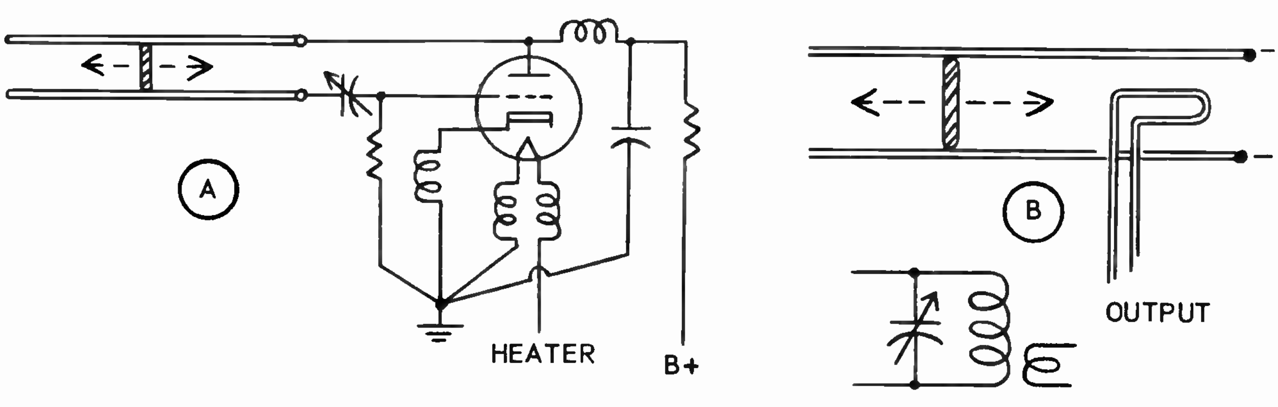


Fig. 99-16. Shorted line circuits must include the usual means for circuit isolation, as at the left. An inductive output coupling is shown at the right.

One method of taking output energy from a resonant line oscillator is shown at B. A small loop is mounted in such position that it always is between the shorting bar and the tube terminals of the line, or so that it always is near an active part of the line. This is equivalent to the inductive coupling to an oscillator coil shown in the lower sketch.

It is more common practice to take the oscillator output through a very small capacitor, as is usually done in apparatus working at lower frequencies. This oscillator coupling capacitor sometimes is connected to the plate side of the resonant line. It seldom if ever is connected to the grid side. To avoid unbalancing the line or adding unduly to its effective length by loading of the output connection, the oscillator energy sometimes is taken to the mixer through a small capacitor connected to one of the heater leads. Oscillator energy then passes through the cathode-heater capacitance of the tube.

In order to reduce radiation, the two conductors of a resonant line are placed no farther apart than about 2 per cent of the shortest wavelength to be received, and often they are but little more than 1 per cent of a wavelength from each other. For a frequency of 890 mc this means a separation of about a quarter-inch, or even less. On the other hand, if the two conductors are too close together there may be excessive energy loss due to eddy currents.

At ultra-high frequencies every part of the oscillator system must be enclosed within grounded shields. This applies to the oscillator tube, to whatever type of tuning circuit may be employed, and to all connections between the tube socket and the tuning device. If the shielding is too close to the tuning device it adds unduly to capacitance, although it may reduce inductances, and in any case makes for erratic performance and non-linear tuning. If the shielding enclosure is too large it very possibly may act as a cavity resonator to generate many frequencies other than the one desired. Everything connected with ultra-high frequency is "critical" in dimensions, positioning, materials, and almost every structural detail in the apparatus.

It was mentioned in connection with the butterfly tuning circuit that one of the chief advantages is freedom from flow of uhf currents through sliding contacts. One of the chief difficulties in the design of satisfactory tuners using shorted lines is in making good contact of the shorting bars on the line conductors. Why this is so difficult is explained by the electrical characteristics of a shorted line. At the point of shorting there is maximum current. This means that there is minimum voltage - theoretically there is zero voltage. In any case there is not enough voltage to force the current through even the slightest insulating film on the conductors. Unless the contacts are of practically zero resistance there will be no conduction between the two conductors, there will be no effective shorting, and there will be no tuning of the circuit. This is both a manufacturing problem and a service problem.

CAVITY RESONATORS. At and above the highest frequencies handled by lumped inductances and capacitances in resonant circuits we may use resonant lines. At the highest frequencies for which resonant lines may be used, also at far higher frequencies, the tuned circuits may be in the form of cavity resonators.

Fig. 99-17 illustrates the basic principle of the cavity resonator by showing its development. At A we have a lumped-element resonant circuit consisting of a coil and a capacitor. For very-high and ultra-high frequencies the capacitance may be reduced by using two well separated discs, as at B, joined by a conductor which provides the inductance. To further reduce the inductance, and thus raise the resonant

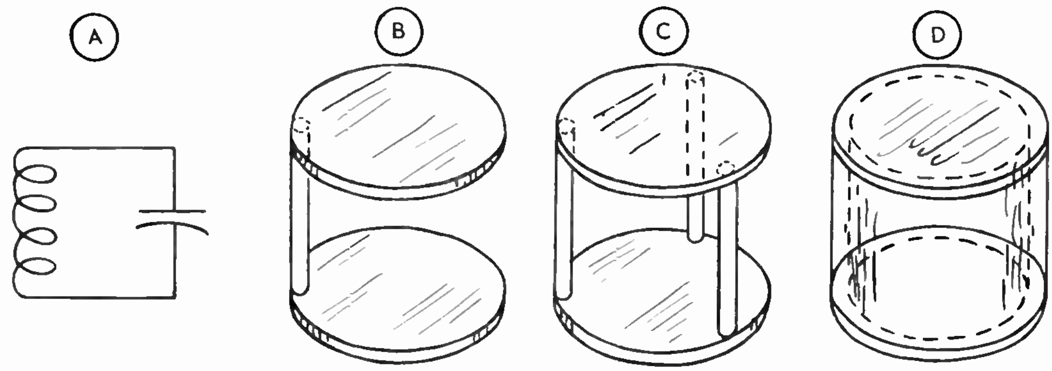


Fig. 99-17. The relation of a cavity resonator to a resonant circuit consisting of lumped inductance and capacitance.

frequency still higher, we may use several conductors in parallel, as at C. The least possible inductance and highest frequency for the existing capacitance will result with the structure at D, where we have added so many paralleled conductors that they become a continuous ring extending between the end plates. This is one form of cavity resonator.

The magnetic field lines which formerly whirled around the separate conductors now are wholly confined within the cylinder. They follow circular paths which are concentric with the circular wall of the cylinder. There is no external magnetic field. The lines of the electric field still extend from one end plate to the other, just as they extend between the plates of any capacitor. There are no electric lines outside the cylinder, because its walls form a shield.

Inside the cylinder we now may have magnetic and electric lines which everywhere are at right angles

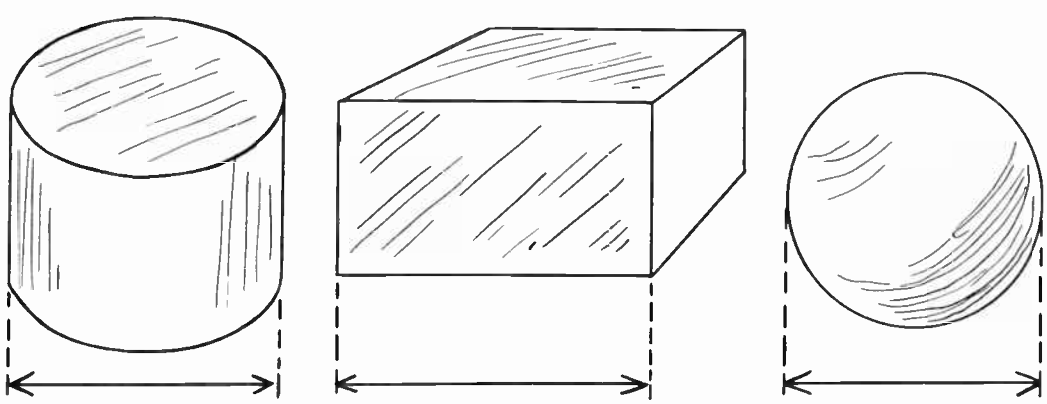


Fig. 99-18. Cavity resonators of these relative sizes would all be resonant at the same frequency.

to each other. This, as you will remember, is the definition of an electromagnetic wave. When energy is put into the cavity there will be internal electromagnetic waves at a frequency for which the cavity is resonant.

Cavity resonators may be of various shapes. With any shape the resonant frequency depends on the dimensions, becoming higher as the cavity is made smaller, and lower as the cavity is enlarged. Fig. 99-18 shows a cylinder, a square box, and a sphere or ball, all drawn to such relative sizes as to have the same resonant frequency. For the cylinder and the sphere this frequency is determined by their diameters, and for the box it is determined by the length of one of the sides. For the resonant frequency to depend on these dimensions, the heights of the cylinder and the box must be less than a half-wavelength, or they must be proportionately no greater than shown by the drawings.

The cylindrical cavity is the form which lends itself best to ordinary mechanical constructions and to adjustment of the resonant frequency. If the diameter of the cylinder is made small in relation to the internal height, the resonant frequency becomes dependent on this height. The resonant frequency is

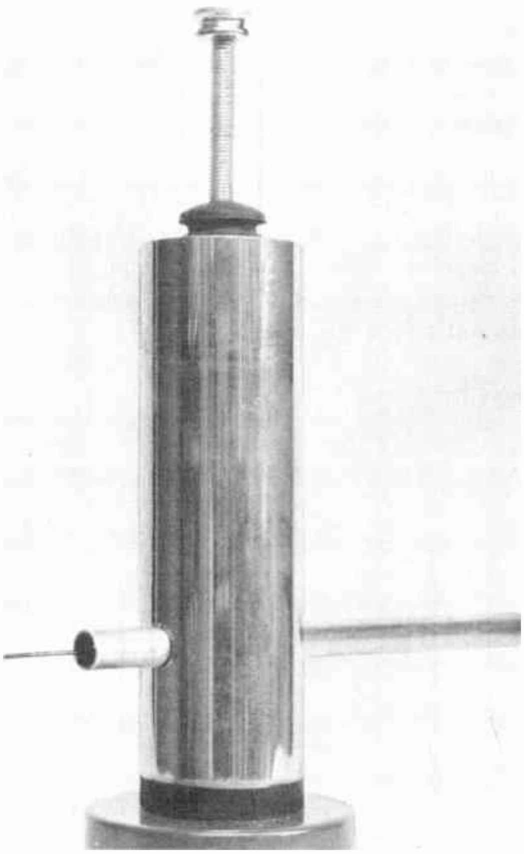


Fig. 99-19. An experimental cavity resonator with an adjustable disc for effective length and with coaxial lines for terminals.

that for which the internal height equals a half-wavelength. and is equal in megacycles to the number 5,905 divided by the inches of height. Stating this the other way around, the height in inches required for any resonant frequency is equal to 5,905 divided by the number of megacycles.

The internal height of a cylinderical cavity resonator for 890 mc, the top of the uhf television band, would have to be about 6.64 inches. For the bottom of the band, 470 mc, the height would have to be about 12.57 inches. To make the height adjustable, the cylinder may be closed at one end with a flat circular plate, and in the other end may be fitted a movable plate or plug. The experimental resonator of Fig. 99-19 is constructed in this manner. This particular cylinder is only long enough, on the inside, to resonate at a minimum frequency around 1,200 mc. Resonant heights do not correspond exactly to the formula, because there is some effect of velocity factor. Whatever mechanical device is used to move the adjustable plate may be calibrated and graduated in megacycles of frequency.

The circuit terminals of the cavity resonator usually consist of the central conductors of air-insulated coaxial lines, as may be seen in the photograph. The central conductors extend a short distance into the internal cavity. No form of external coupling can be used, because the resonant electromagnetic fields are wholly within the cavity.

The physical size of a cavity resonator for the uhf television band may be rather large for convenient mounting in a tuner mechanism. The most troublesome electrical problem is in making an effective contact between the sliding plate and the wall of the cylinder. At the resonant frequency there is, theoretically, zero voltage at the point of contact. The least trace of insulating film will prevent completion of the conductive path and will cause unreliable tuning or complete failure to tune to a desired frequency.

TELEVISION

LESSON NO. 100

UHF TUNERS AND CONVERTERS

Training Your Mind...
C.T.I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

C.T.I.

Chicago, Illinois

Training Your Mind
C. T. I.
Perfecting Your Skill

COMMERCIAL TRADES INSTITUTE

Chicago, Ill.

LESSON NO. 100

UHF TUNERS AND CONVERTERS

Television has been established in the very-high frequency channels for so much longer than in the ultra-high frequency channels that all receivers, old and new, are capable of receiving any of the vhf channels. Many of the more recent models have built-in tuners of types which will handle both uhf and vhf signals. Any of the sets designed to handle only vhf signals may be used with an external converter which makes possible the reception of uhf signals. Regardless of the manner in which uhf reception is made possible, it always is combined with vhf reception.

In the tuners of very-high frequency television receivers we often find a circuit arrangement which is essentially like that shown at the top of Fig. 100-2. Details may differ widely, but always we find three tubes or tube sections which work as r-f amplifier, r-f oscillator, and mixer. Usually there are at least three tuned circuits; one for the plate of the r-f amplifier, another for the grid of the mixer, and another for the grid-plate circuit of the oscillator. Often there is tuning also between the antenna and the r-f amplifier.

At the bottom of Fig. 100-2 is a fairly typical circuit arrangement for ultra-high frequency tuning. Here again are three tuned circuits. They are shown as shorted quarter-wave lines, but might be of any other kind suitable for ultra-high frequencies. The big difference between the two tuners is in the absence of the r-f amplifier tube and the mixer tube in the uhf arrangement. The oscillator tube remains. The mixer tube is replaced with a crystal diode.

The chief reason for omitting tubes is to avoid adding to the received signal the noise voltages which originate in any tube. With no r-f amplifier, no tube noise is added between antenna and mixer. The crystal mixer adds very little noise, although this type of mixer cuts down the signal strength instead of giving amplification, as does a mixer tube.

Without the amplification which would be provided by r-f amplifier and mixer tubes, the signal output from the crystal mixer is very weak, but normally is well above the reduced noise level. This mixer output, with its favorable ratio of signal to noise, is applied to an amplifier which we shall call the beat-frequency amplifier. Because this amplifier works at only the moderately high frequency from the mixer, often around 100 mc, its gain can be made great enough to compensate for what is lacking due to absence of r-f amplifier and mixer tubes.

The crystal mixer, like any other mixer, delivers sum and difference frequencies or beat frequencies which are produced by combination of uhf carrier frequencies and oscillator frequencies. These frequencies

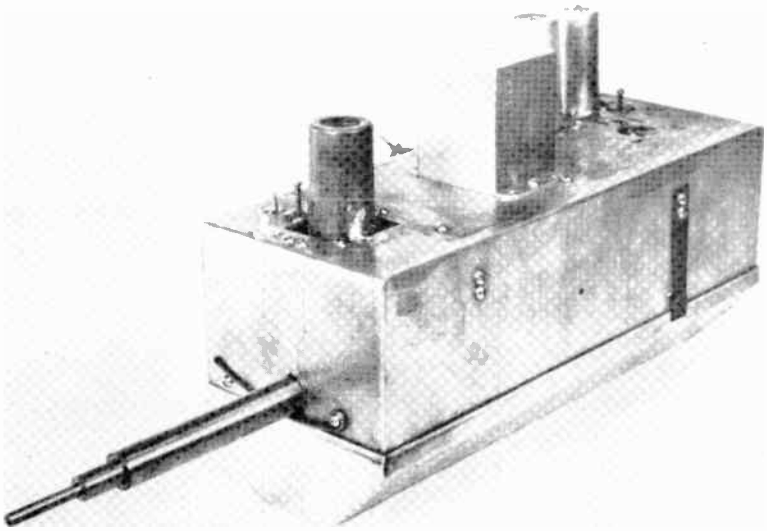


Fig. 100-1. A Standard Coil all-channel television tuner for 70 ultra-high frequency and 12 very-high frequency channels.

in the mixer output will, of course, be the same for all channels. The beat-frequency amplifier circuits are tuned to whatever mixer output frequency it is desired to use, just as the i-f amplifier in any vhf receiver is tuned to some constant intermediate frequency.

The constant-frequency output of the beat-frequency amplifier may be used in various ways. In Fig. 100-3 the output of this amplifier is fed to the antenna terminals of the vhf section of a receiver. The beat-frequency amplifier is permanently tuned for carrier frequencies in some vhf channel in which there is no local reception. The channel selector of the vhf receiver is tuned to this channel, and allowed to remain there for all uhf reception. It is necessary to provide a switch for connecting the vhf tuner to the regular vhf antenna for vhf reception, and to the beat-frequency amplifier for uhf reception.

Uhf channel selection is performed with the uhf channel selector, which simultaneously changes the resonant frequencies of the uhf antenna tuning circuits and of the uhf oscillator, just as any vhf channel selector simultaneously changes the frequencies of the r-f amplifier and r-f oscillator in the vhf tuner. Contrast, brightness, and sound volume are controlled in the vhf sections for reception in either band. This general method of tuning and control is employed in nearly all systems of combined uhf and vhf reception.

If the vhf tuner is a continuous tuning type, rather than one which tunes in steps to each of the vhf channels, it is possible to employ the system illustrated by Fig. 100-3 in another way. The output of the beat-frequency amplifier may be tuned to some frequency below vhf channel 2 or else for some frequency between vhf channels 6 and 7, but not in the f-m broadcast band or in any other band used for other radio services.

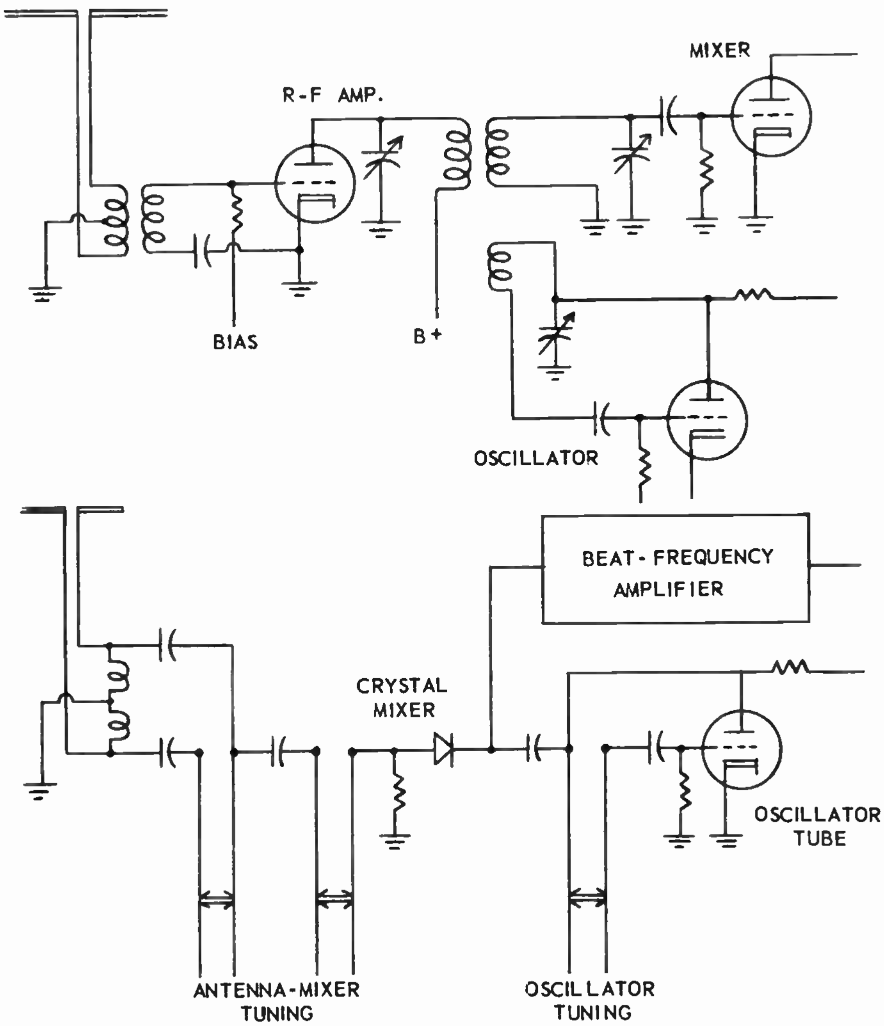


Fig. 100-2. Similarities and differences in tuners for very-high and ultra-high television frequencies.

If the vhf tuner is not a continuous type, but is of the step-by-step variety, this same general method may be employed by providing on the vhf tuner one extra position at which the frequency is that of the uhf beat-frequency amplifier. In other words, the vhf tuner will have an extra position at which it is resonant somewhere below channel 2 or else somewhere between channels 6 and 7 where there won't be interfering signals.

Another system of combined uhf-vhf reception is illustrated by Fig. 100-4. The vhf tuner is constructed with a cascode r-f amplifier. There is no beat-frequency amplifier in the uhf section. For uhf reception the oscillator of the vhf tuner is disabled, and the output of the uhf crystal mixer is connected to

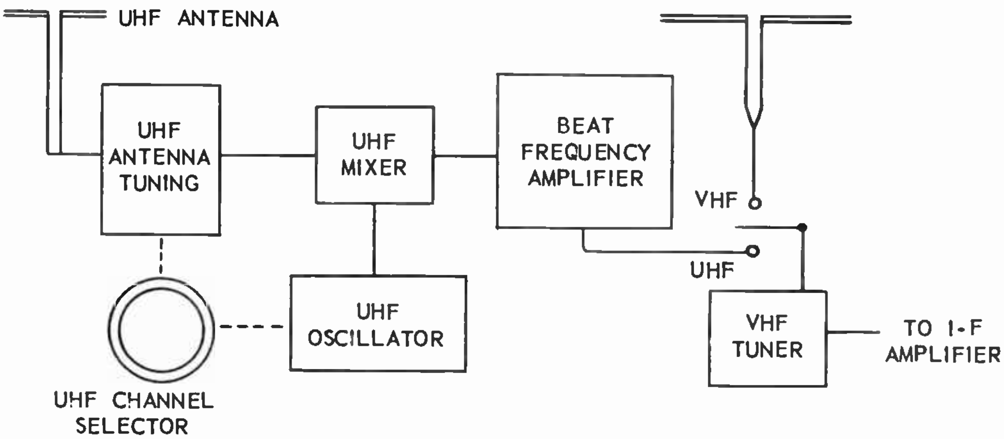


Fig. 100-3. Principal parts of a typical uhf television tuner.

the input of the cascode amplifier in the vhf tuner. There will be a position of the vhf channel selector that tunes this cascode r-f amplifier to the constant beat-frequency which is the output of the uhf crystal mixer.

Fig. 100-5 shows the principle of still another method for receiving both uhf and vhf signals. The uhf tuner is like those in Figs. 100-2 and 100-3, including the beat-frequency amplifier which follows the mixer. The uhf oscillator is tuned to produce, with the uhf carriers, a beat frequency which is the intermediate

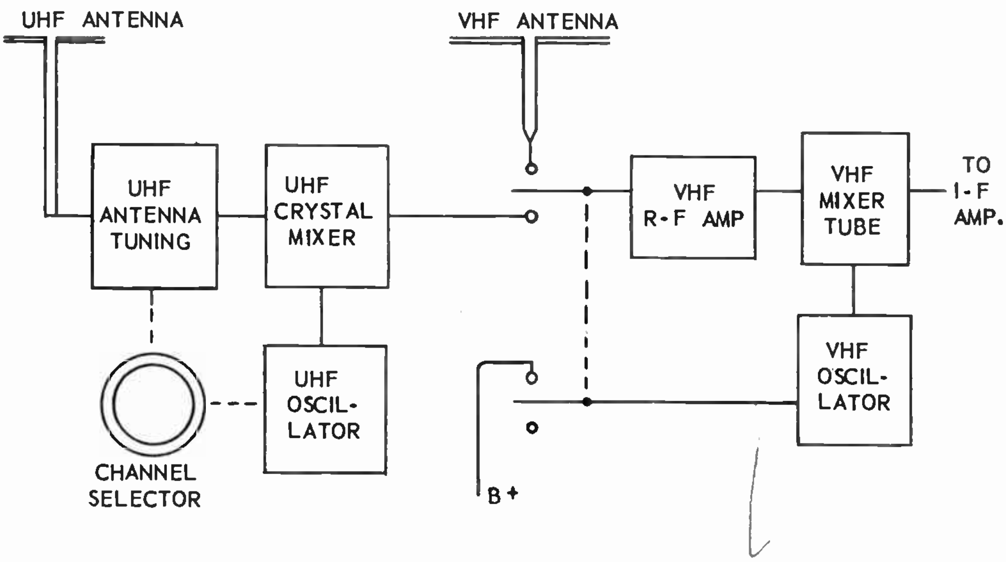


Fig. 100-4. How the r-f amplifier of the vhf section may be used to strengthen signals from the uhf mixer.

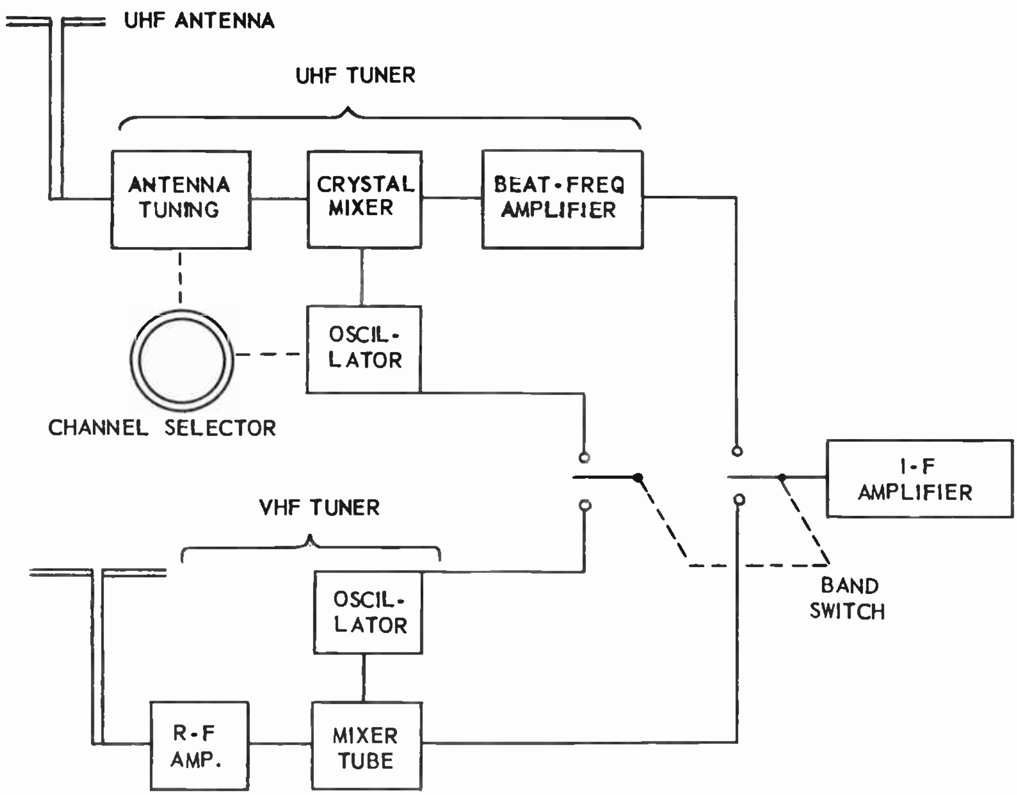


Fig. 100-5. A uhf tuner and a vhf tuner, either of which feeds directly to the i-f amplifier of the receiver.

frequency of the regular i-f amplifier in the vhf section of the set. The beat-frequency amplifier is tuned to this intermediate frequency.

For uhf reception the band switch connects the beat-frequency amplifier to the i-f amplifier of the receiver. In this method the intermediate frequencies usually will be 45.75 mc for video and 41.25 mc for sound. For vhf reception the band switch connects the output of the vhf mixer to the i-f amplifier. One section of the band switch cuts off the uhf oscillator during vhf reception, and cuts off the vhf oscillator during uhf reception.

Fig. 100-6 illustrates the general principles of one method for receiving all the uhf channels and all the vhf channels, not by continuous tuning through all the bands but with a step-by-step method. That is, there are definitely fixed positions for each of the 70 uhf channels and for each of the 12 vhf channels, just as in many vhf tuners we go from one fixed position to another for each of the vhf channels. The tuner utilizing this principle is pictured by Fig. 100-1.

Consider first the uhf section of Fig. 100-6. The uhf channel selector consists of eight tuned circuits

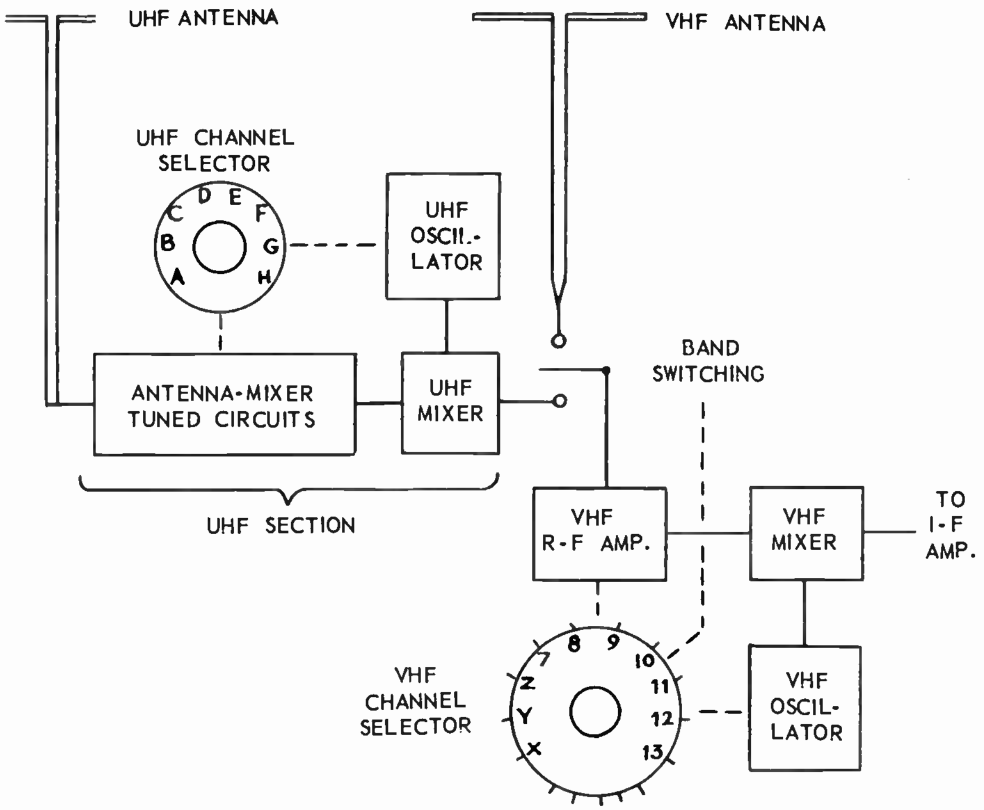


Fig. 100-6. The principle of the Standard Coil method of decade or decimal tuning for each of 82 television channels.

connected between antenna and mixer. Each circuit is broadly responsive over a range of frequencies about 60 mc wide. The eight 60-mc ranges thus cover the entire uhf band from 470 to 890 mc with something to spare, since 8 times 60 mc equals 480 mc, while the difference between 890 and 470 mc is only 420 mc. The switch that adjusts the uhf channel selector to any of its eight positions also tunes the uhf oscillator to any of eight fixed frequencies.

Now let's see what can be done while the uhf channel selector remains in one position which, merely for identification in this example, we shall call position B. We shall assume that, in this position, the antenna-mixer circuits respond to all carrier frequencies between 506 and 566 mc. These are the frequencies included in uhf channels 20 through 29. We shall assume further that the uhf oscillator, in our position B, is tuned to 350 mc. The resulting beat frequencies will be the difference between the carrier frequencies (506 to 566 mc) and the oscillator frequency (350 mc), and will extend from 156 mc to 216 mc, according to whatever uhf carrier frequencies may be reaching the antenna. These beat frequencies are applied to the regular antenna input of the vhf tuner.

The channel selector switch for the vhf section is shown in Fig. 100-6 as having ten positions. Positions numbered 7 through 13 are the regular positions for receiving vhf channels 7 through 13. There are three extra positions, here identified as X, Y, and Z, at which tuning is in three additional ranges, each 6 mc wide. This gives a total of ten 6-mc ranges from which selection may be made. Actually there are still other positions for vhf channels 2 through 6, but these positions are not used for uhf reception.

Now, keeping in mind that the uhf section, with its selector in position B, is delivering beat frequencies all the way from a minimum of 156 mc to a maximum of 216 mc, we shall see what happens when this range of frequencies is applied to the vhf tuner.

UHF SECTION, IN "POSITION B"			VHF SECTION		
Antenna-Mixer Tuning	Oscillator Frequency	Beat Frequency	Selector Position	Response, mc	UHF Channel
Minimum	Constant				
506 mc –	350 mc =	156 mc	X	156 - 162	20
			Y	162 - 168	21
			Z	168 - 174	22
			7	174 - 180	23
(This entire range remains in tune so long as the uhf channel selector is in our position B.)			8	180 - 186	24
			9	186 - 192	25
			10	192 - 198	26
			11	198 - 204	27
			12	204 - 210	28
566 mc (Maximum) –	350 mc (Constant) =	216 mc	13	210 - 216	29

By keeping the uhf channel selector at position B, to deliver any beat frequencies between 156 and 216 mc, and by turning the vhf channel selector to some one of its ten positions, we may pick any 6-mc range out of the beat frequencies. Every 6-mc range must cover one channel, whether in the uhf or vhf band. In the uhf band we thus are enabled to select any one of the channels 20 through 29 merely by shifting the vhf selector.

With the uhf selector in any other position the antenna-mixer tuning will be changed to some other 60-mc range of uhf carrier frequencies, and at the same time the uhf oscillator frequency will be changed so that the beat frequencies from the uhf mixer again extend from 156 to 216 mc. With the uhf selector in any one of these other positions, the vhf selector will pick out any 6-mc range from the beat frequencies. Each 6-mc range will cover one uhf channel. Thus we find that each of the uhf selector positions will allow reception in some certain group of ten uhf channels. The first position actually is used only for channels 14 through 19, and the eighth position only for channels 80 through 83. The intervening six positions are for all other uhf channels in groups of ten.

In the tuner as actually constructed there are many refinements not discussed in this explanation of principles. For instance, the two selectors are combined in one dial having knobs for selecting the "tens" and the "units" for the number of any uhf channel to be received. The numbers of all channels, both uhf and vhf, appear in windows of the dial. None of the positions are lettered, as has been done merely to simplify our explanation of principles.

DOUBLE CONVERSION. In all uhf tuning systems we have, in addition to the uhf section, the tuner which is regularly used for vhf reception. With the systems of Figs. 100-2 and 100-3 the uhf section may be constructed as a separate unit and connected to the antenna terminals of any vhf receiver, so that programs may be received in the uhf band as well as in the vhf bands. Such a separate unit is called a uhf converter. It may be designed to handle only the few uhf channels in which there is local reception, or it may be of a type which tunes continuously through the entire uhf band. The converter does not prevent using the receiver for vhf reception.

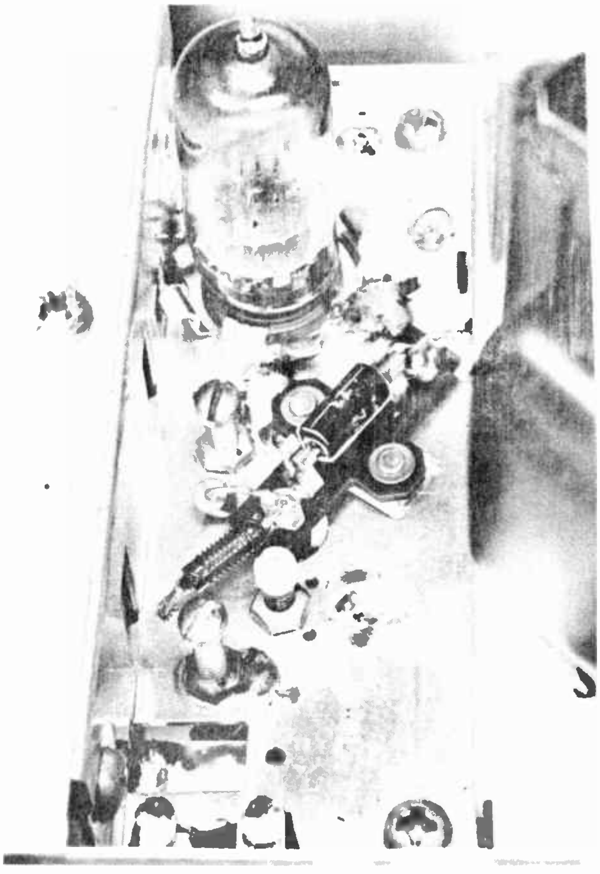


Fig. 100-7. The oscillator tube and the crystal mixer of an uhf tuner.

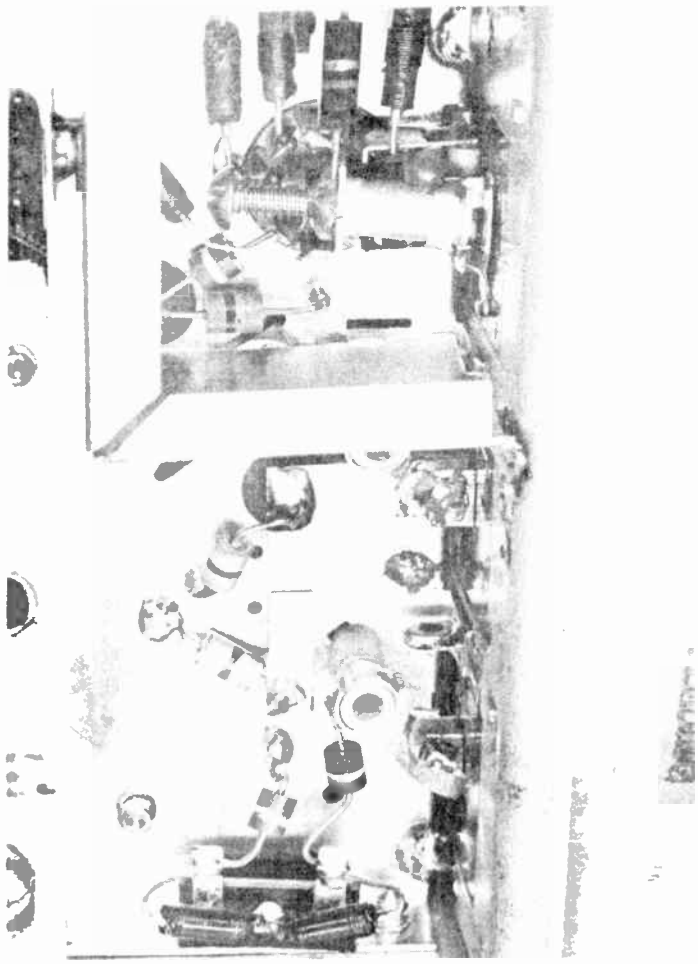
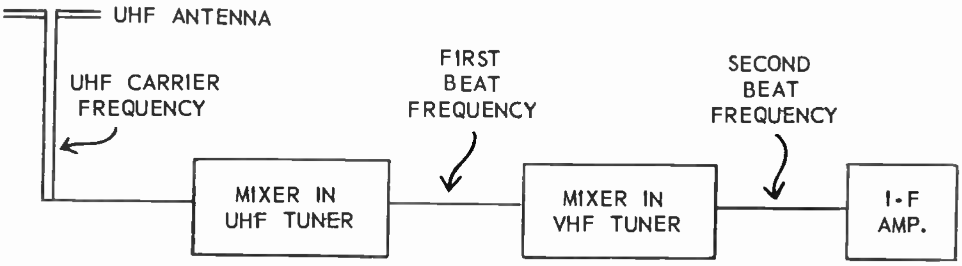


Fig. 100-8. The underside of the uhf tuner chassis, showing the oscillator socket compartment and some of the antenna-mixer connections.

Fig. 100-7 shows portions of the uhf section of a converter as seen from above the chassis. The oscillator tube is at the upper left. At the center is the crystal mixer. Fig. 100-8 shows parts which are visible from underneath the same chassis with some of the shielding removed. At the top is the oscillator socket compartment in which are some of the r-f chokes, resistors, and capacitors for the uhf circuits. Antenna and mixer connections are at the bottom of this picture.

With converters of this type, also in most uhf tuner systems, the received signals pass through two mixers, one after the other, as illustrated at the top of Fig. 100-9. Uhf carrier frequencies are converted to lower beat frequencies by the uhf mixer. The signal output from this mixer then is converted to still lower beat frequencies by the vhf mixer. This process is called double conversion. The first beat frequency,



CHANNEL 19	VIDEO	SOUND	VIDEO	SOUND	
UHF CARRIERS	501.25	505.75			
UHF OSC'R.	418.00	418.00			
			129.00	129.00	VHF OSC'R.
1st BEAT	83.25	87.75 →	83.25	87.75	1st BEAT
			45.75	41.25	2nd BEAT

Fig. 100-9. A typical combination of frequencies for uhf reception by the double conversion method.

which goes to the antenna terminals of the vhf tuner, must be the one for which this tuner is set. The second beat frequency, which goes to the i-f amplifier, is the regular intermediate frequency for the receiver.

For uhf reception the vhf tuner is set for the constant beat frequencies coming from the uhf mixer, and is left there. Oftentimes the output of the uhf mixer will be very broad, and centered between two vhf channels. Then the vhf tuner may be set for either of these channels in which there is no local vhf reception. The output of the uhf mixer might be centered between channels 2 and 3, or between 5 and 6, or between 12 and 13. Of any of these pairs of channels only one will be active in any locality and the other may be used for uhf reception. Vhf tuner circuits for the channel employed in uhf reception must be accurately aligned.

An example of double conversion frequencies is given at the bottom of Fig. 100-9. Uhf channel 19 is used for illustration, although frequencies for other channels would work out in a similar manner but with different figures. Video and sound carriers are respectively at 501.25 and 505.75 mc in channel 19. For reception in this channel we assume that the uhf oscillator is tuned to 418.00 mc. Subtracting this oscillator frequency from the video and sound carrier frequencies gives the first beat frequencies as 83.25 mc for video and 87.75 mc for sound. Note that these beat frequencies are the same as the vhf carrier frequencies for channel 6. They are applied to the vhf tuner.

For all uhf reception, with our assumed frequencies, the vhf tuner will be set for channel 6. For this channel the vhf oscillator frequency will be 129.00 mc. From this vhf oscillator frequency we subtract the beat frequencies that come from the uhf mixer. The differences are second beat frequencies of 45.75 mc for video and 41.25 mc for sound. These are among the standard intermediate frequencies for vhf receivers. They are applied to the regular i-f amplifier.

In the vhf tuner of this example, as in all such tuners, the oscillator frequency is higher than the received frequency. This higher oscillator frequency produces a video intermediate which is higher than the sound intermediate. It is this relation of video and sound intermediates for which all i-f amplifiers are designed and aligned.

In order not to reverse the relation between video and sound intermediate frequencies, the video and sound "first beat" signals coming to the vhf tuner must have the same relation as video and sound frequencies in any vhf carrier. That is, the sound signal frequency must be higher than the video signal frequency. This can be accomplished only by operating the uhf oscillator at a frequency lower than the uhf carrier frequencies, as in our example. In all double conversion systems the uhf oscillator frequency must be lower than the received uhf carrier frequencies. Unless this frequency relation is observed, we would go to the i-f amplifier with the sound intermediate higher than the video intermediate.

In television sets designed for reception in only the vhf bands there is only single conversion, because there is only one mixer. In Fig. 100-4 is an example of single conversion used for uhf reception.

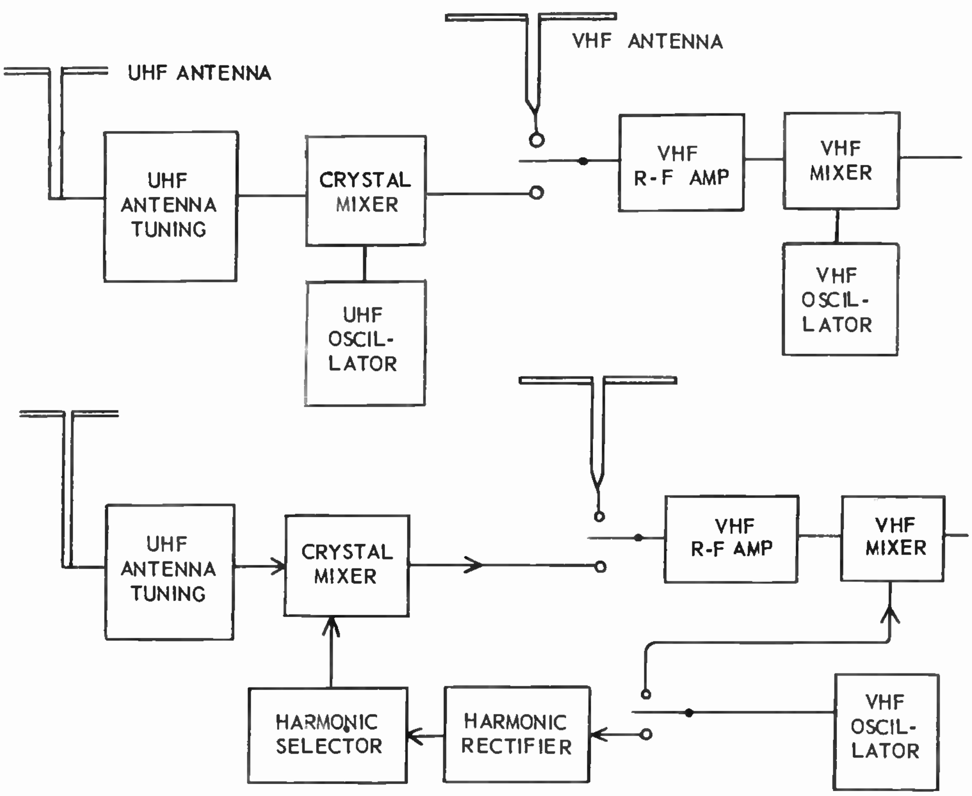


Fig. 100-10. The uhf oscillator of the upper diagram is replaced by a harmonic rectifier and selector in the lower diagram.

Only the uhf mixer precedes the i-f amplifier during uhf reception, and only the vhf mixer precedes the i-f amplifier during vhf reception. In this case the frequency of the uhf oscillator must be higher than the uhf carrier frequencies, just as the vhf oscillator frequency is higher than vhf carrier frequencies in all vhf receivers.

USING OSCILLATOR HARMONICS. At the top of Fig. 100-10 is a block diagram of a uhf tuning system that is like the one of Fig. 100-1. In the uhf section there is tuning of the antenna-mixer circuits, also a uhf mixer of the crystal type, and a uhf oscillator tube. The output of the uhf mixer goes to the antenna terminals of the vhf tuner.

At the bottom of Fig. 100-10 is a block diagram which differs from the one above in that there is no uhf oscillator. Instead there is a harmonic rectifier and a harmonic selector, forming what may be called a harmonic generator. This harmonic generator will produce a second or third harmonic frequency from the output of the vhf oscillator. This harmonic is applied to the uhf mixer instead of the equivalent frequency formerly obtained from the separate uhf oscillator tube in the uhf section.

The principle of a harmonic generator such as often used for uhf tuning is illustrated by Fig. 100-11. The frequency at which the vhf oscillator is operating for any selected vhf channel is applied through blocking capacitor Cb to the crystal rectifier. Capacitor Cb and resistor Rb are of such values that a biasing voltage maintained across Rb causes the crystal to operate at the sharpest bend of its current-voltage characteristic curve. Then, as is the case with any rectifier or detector operating in this manner, the crystal circuit contains harmonics of the vhf oscillator frequency. One of these is the third harmonic frequency, which we shall assume is the one desired for use in the uhf mixer.

In series with the crystal rectifier is a parallel resonant circuit, the harmonic selector, which is tuned

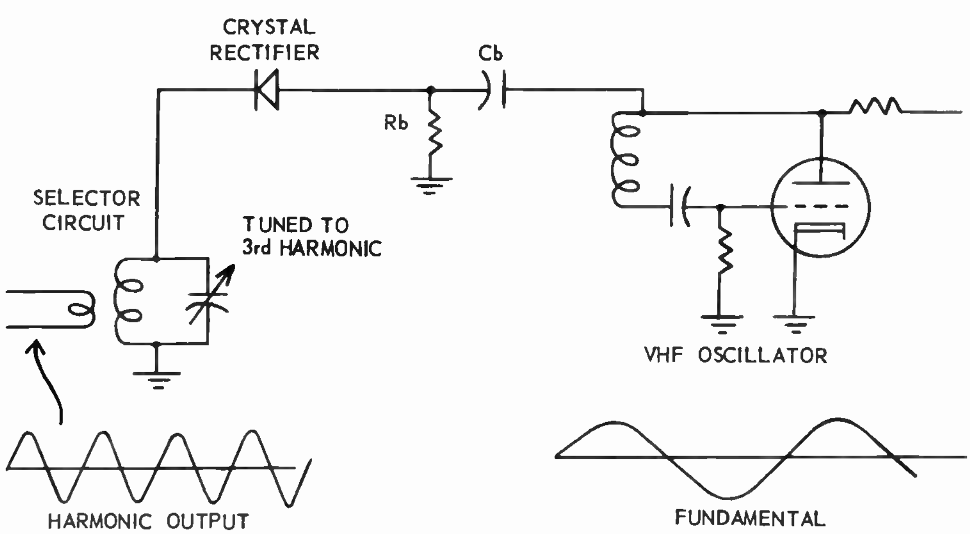


Fig. 100-11. Elementary circuits for one type of harmonic generator.

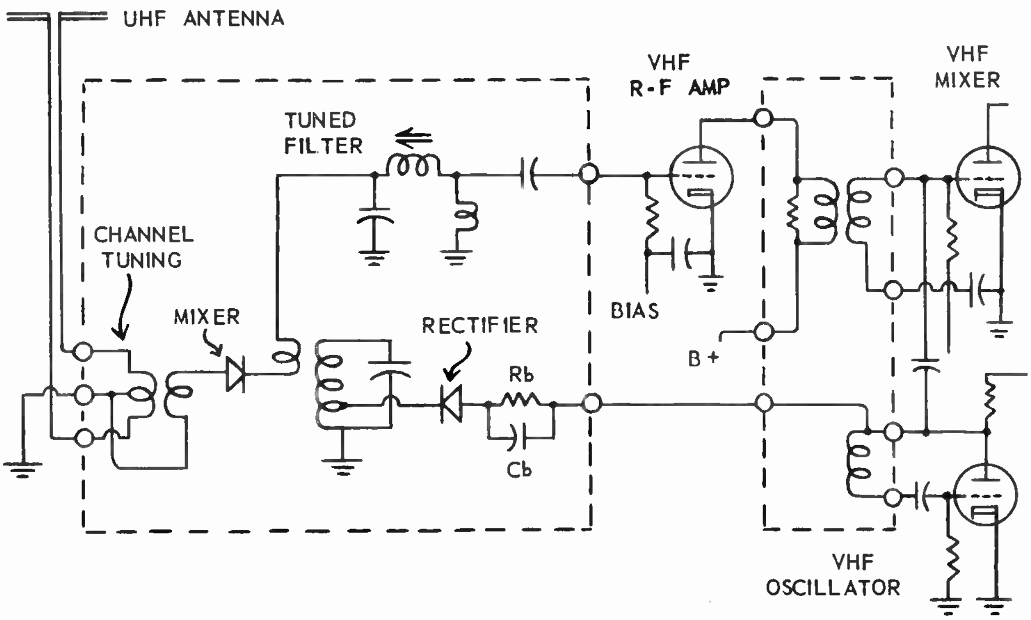


Fig. 100-12. A method of using the harmonic generator system in a uhf channel strip for a turret tuner.

to the third harmonic frequency. In this circuit the third harmonic voltages have the benefit of resonant gain, which we studied long ago, and at the same time the other harmonics and the fundamental of the vhf oscillator are greatly attenuated. By means of any suitable coupling method, the third harmonic voltage may be taken from the harmonic selector circuit and applied to the uhf mixer, just as a uhf oscillator voltage would be applied to this mixer.

Fig. 100-12 shows one way in which the harmonic generator system may be employed in connection with the tuner of a vhf receiver for reception of uhf signals. The harmonic rectifier and selector circuits are practically the same as in Fig. 100-11, except that now the harmonic rectifier crystal is biased by a small voltage taken from the vhf oscillator circuit through a high resistance at Rb. The channel tuning method will vary with the make and model of tuner or receiver, and may be of any type suitable for uhf reception.

The beat-frequency output of the uhf crystal mixer goes to the grid of the vhf r-f amplifier through a tuned filter which opposes and bypasses remaining ultra-high frequencies, while passing the desired beat frequency. This is equivalent to the tuned coupler used between the mixer and following i-f amplifier in most vhf sets. The vhf r-f amplifier now acts to strengthen what we have called the first beat frequency, coming from the uhf mixer. Oftentimes this r-f amplifier is a cascode type.

Inasmuch as the harmonic generator method does away with the separate uhf oscillator tube, and may be employed with a crystal mixer, it lends itself well to use in turret tuners which have removable and interchangeable channel strips. One of the vhf strips, for a channel not active in the locality, may be removed and replaced with a strip for any uhf channel. This uhf strip need carry only the parts which are enclosed within the broken lines of Fig. 100-12, including only a few inductors and capacitors, and two crystal diodes. The tubes of the vhf tuner then are employed for all uhf and vhf television bands, as is also the regular i-f amplifier.

The intermediate frequencies going from the vhf mixer to the i-f amplifier must remain the same for all channels. To allow this, it is necessary to change the vhf oscillator frequency for each channel to be received in the uhf band. This necessity arises because the harmonic frequency fed to the uhf mixer always will be a multiple of the vhf oscillator frequency, and to secure the correct harmonic frequency it is necessary to change the fundamental vhf oscillator frequency. This means that a vhf oscillator coil for each uhf channel must be on the uhf strip for that channel. Also, to suitably tune the vhf r-f amplifier and the mixer, their tuning coils must be on the uhf channel strip.

The various frequencies for one particular uhf channel might work out as follows. We assume that the receiver employs a video intermediate of 45.75 mc and a sound intermediate of 41.25 mc, and that we wish to tune to uhf channel 31.

	Video	Sound
Vhf oscillator frequency (the fundamental)	154.75 mc	154.75 mc
Subtracting the intermediate frequencies	45.75	41.25
Beat frequencies required from uhf mixer	109.00	113.50

The fact that the uhf mixer will furnish these required beat frequencies may be shown as follows.

	Video	Sound
Carrier frequencies for uhf channel 31	573.25 mc	577.75 mc
The third harmonic of the vhf oscillator frequency will be 3 times 154.75 mc, which is 464.25 mc. Subtract this from the carriers	464.25	464.25
Beat frequencies furnished by the uhf mixer	109.00	113.50

These combinations of frequencies are marked, in megacycles, on the block diagram Fig. of 100-13.

Were the receiver to employ a video intermediate of 25.75 mc and a sound intermediate of 21.25 mc, and were we still tuning to uhf channel 31, the frequencies would be as follows.

Vhf oscillator frequency, 149.75 mc. Third harmonic frequency, 449.25 mc. Beat frequency from uhf mixer, 124.00 mc video, 128.50 mc sound. Vhf oscillator frequencies for any other intermediates and any other uhf channels may be computed.

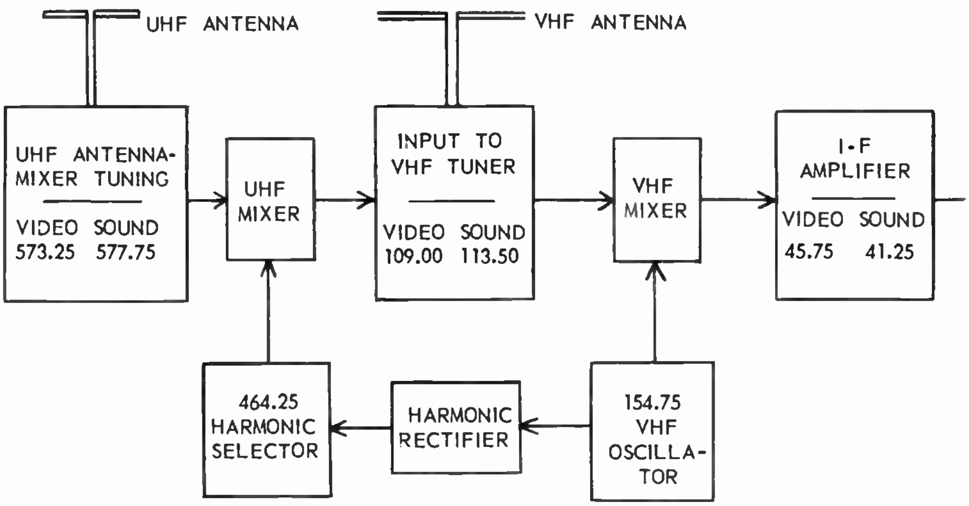


Fig. 100-13. Frequencies in the various sections of a uhf-vhf tuning system employing a harmonic generator.

UHF ANTENNAS. When studying television antennas in general we learned that the performance of any receiver is greatly improved by a good outdoor antenna unless the receiver is within a few miles of transmitting stations. This is true to an even greater extent for uhf reception. Except in primary service areas, the uhf antenna usually determines the success or failure of the installation.

Because signal reflections and ghosts are more troublesome at ultra-high frequencies, the antenna should be highly directional in its reception pattern. Because signal shadowing is worse at ultra-high frequencies, the location, height, and orientation of the antenna become of great importance. Moving the antenna as little as a foot to one side or the other may bring in signals otherwise unobtainable. Added height usually is well worth the effort of attaining it, since this gets the antenna above the shadows cast by objects in the line of signal travel. A high antenna must be well guyed to prevent swaying and twisting effects of wind, snow, and icing.

Many types of antennas used for vhf reception are satisfactory for uhf signals when well constructed and of suitable dimensions. Folded dipoles with reflectors may be used. With this type it is usual practice to stack four bays or even eight bays. Methods of correctly phasing the vertical or sloping connectors between bays are the same as for vhf reception, as are also the spacings between active elements and reflectors, and the effects on gain and directional pattern of varying this spacing.

V-antennas with reflectors usually give good results when two or four bays are stacked and correctly phased. The forward angles between conductors or rods in the active elements are made sharper than for vhf reception, in order to sharpen the directional pattern. That is, the angle formed by the extended conductors of the dipole may be 90 degrees or even less. The best angle is found by trial.

Built-in antennas may not prove satisfactory, even close to transmitting stations. The signal may be weakened or cut off by any metal parts of the building and its fittings, as by metal beams, metal window frames, metal blinds, metal awning frames, and all other metal objects. The signal may be affected also when people come near the receiver. Indoor antennas, as contrasted with built-in types, usually are more satisfactory provided they can be placed high enough to avoid shadowing effects of all metal structures.

To preserve as much of the uhf signal as possible, the transmission line must be of some low-loss variety. Where signal field strength is weak the best line is an open wire type with the conductors supported and spaced apart with insulators having a low dielectric constant. Twin-conductor flat line, such as used for most vhf installations, is satisfactory under ordinary conditions. In localities where interference may be severe it becomes necessary to use coaxial cable of large diameter. Correct matching of impedances at the antenna or receiver, or possibly at both places, is of prime importance.

Dimensions of antenna elements, and their spacings, usually are specified as fractions or multiples of wavelengths at the operating frequencies. One wavelength, in inches, is found from dividing the number 11,811 by the frequency in megacycles. Any fractional wavelength is, of course, that same fraction of one full wavelength. For example, 0.6 wavelength at 680 mc would be found thus.

$$\frac{11{,}811}{680} = 17.37 \text{ inches, approximately.}$$

$$17.37 \times 0.6 = 10.42 \text{ inches, approximately.}$$

Half-wavelengths and quarter-wavelengths, the two values most often needed, are found thus.

$$\text{Half-wave, inches} = \frac{5905.5}{\text{mc}}$$

$$\text{Quarter-wave, inches} = \frac{2952.7}{\text{mc}}$$

CORNER REFLECTOR ANTENNA. A type of antenna not previously examined, but which gives good reception in the uhf television band, is called the corner reflector. Fig. 100-14 is a picture of such antenna whose elements are cut for peak response around 680 mc. The structure is essentially a half-wave dipole with a large number of reflector elements arranged in two planes which, in the design illustrated, come together at an angle of 90 degrees. The dipole element is on a line or plane that passes through the center of the angle. Overall dimensions of this particular antenna are: height 26 inches, width 9 inches, depth from front to back 14 inches.

The corner reflector antenna constructed with a simple straight dipole has a gain as great as 10 to 12 db, compared with the same dipole in free space with no parasitic elements. The reception pattern is highly directional, with front-to-back ratio up to 30 or more db, with little pickup from either side, and with additional lobes which are almost negligible so far as signal pickup is concerned.

The frequency response of this antenna may be extended over a fairly wide range by providing relatively long reflector planes and by spacing the reflector rods close together. There is, of course, some

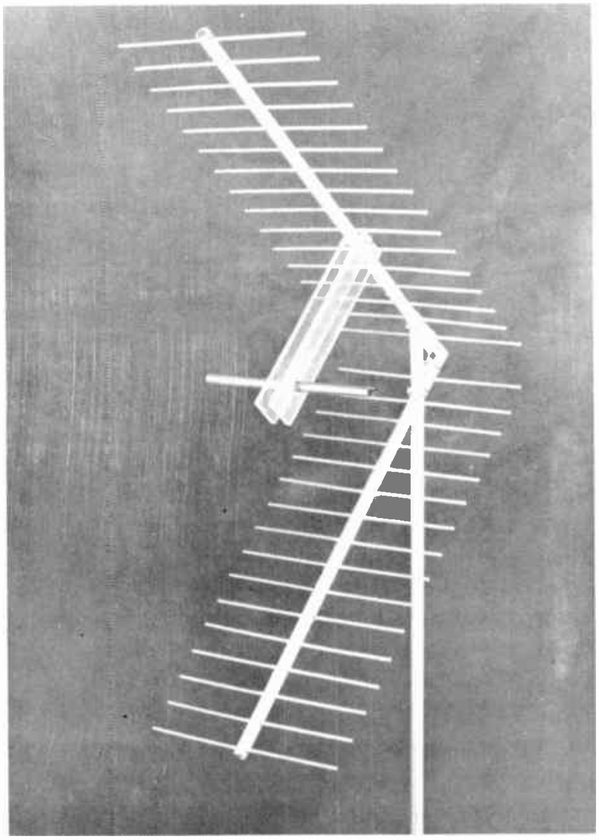


Fig. 100-14. A corner reflector antenna for uhf reception.

drop in maximum gain when the frequency range is widened. As with other antennas, this type may be stacked for added gain. Although the corner angle most often is 90 degrees, it may be as small as 45 degrees. There is some increase of gain with smaller angles provided the reflector planes are lengthened.

Impedance depends on the type of antenna element, on its distance from the corner of the angle (where the reflector planes meet), and on the number of degrees in the angle between reflector planes. Instead of a straight dipole, as pictured, the active element may be a folded dipole or a parallel conductor dipole to increase the impedance.

In the design pictured, the corner angle is 90 degrees and the dipole is slightly more than a half-wavelength from the corner. The impedance is approximately 150 ohms. With this same dipole spacing from the corner, but with an angle of 60 degrees, the impedance would drop to about 75 ohms. Impedance decreases rapidly with still smaller angles.

To increase the impedance when using any corner angle, the active antenna element may be moved farther from the corner. This may be necessary in avoiding an impedance so small as to make it difficult to match the antenna and line when using corner angles of less than 90 degrees. Regardless of the spacing, the antenna element is kept on a line or plane which passes through the center of the reflector angle.

Instead of making the reflector of metal rods it may be formed from a wire mesh screen, preferably of some good conductor such as copper, bronze, brass, or aluminum. This gives somewhat greater gain because of a more nearly complete reflector surface. Were it not for difficulties with wind resistance the reflector could be made from a solid sheet of any good conductor.

It is desirable that the transmission line or a matching transformer (section of resonant line) extend from the antenna element through the back corner of the reflector before being sloped down and away to the receiver connection.

Following are dimensions, in wavelengths and fractions at the center frequency, for corner reflector antennas as usually constructed. These dimensions are marked on Fig. 100-15.

A. Length of each side of reflector, or length of each reflector plane, 1.0 wavelength or more. Greater length adds somewhat to gain.

B. Center of dipole element to reflector corner, about 0.5 wavelength.

C. Center-to-center spacing of reflector rods from one another, 0.05 to 0.10 wavelength. The smaller spacing adds to the gain.

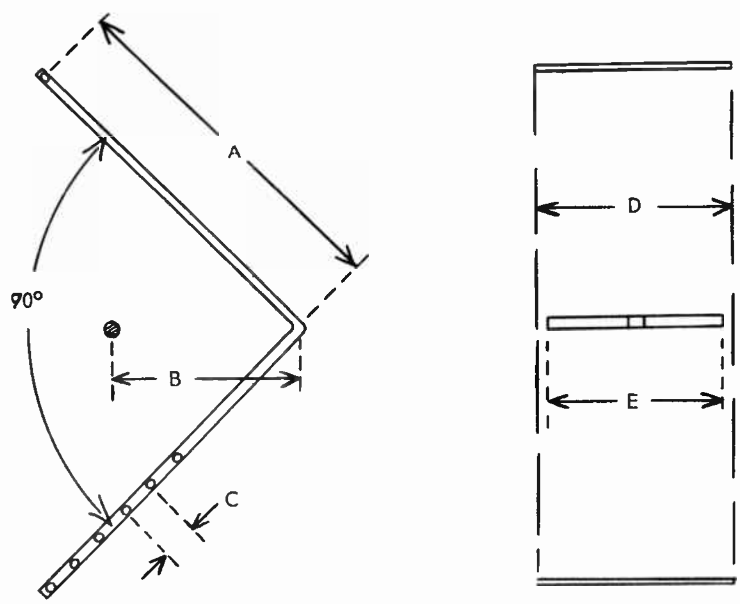


Fig. 100-15. Principal dimensions of the corner reflector antenna.

D. Length of reflector rods or width of reflector screen in a horizontal direction, about 0.55 to 0.60 wavelength.

E. Overall length of dipole element, 0.5 wavelength or slightly less.

SERVICING AT ULTRA-HIGH FREQUENCIES. Because the differences between receivers for ultra-high and very-high frequencies are entirely in the tuners, it is only in this section that any special methods are required when working at ultra-high frequencies. Even in the tuners, alignment and adjustment procedures follow the same general methods with which we are familiar. For instance, there may be adjustments for the high- and low-frequency ends of the uhf band, or for various portions of the band when there is step tuning rather than continuous tuning. Adjustments most often are made by small variable capacitors or trimmers, but sometimes by varying the inductances.

Constant-frequency signal generators or marker generators may furnish fundamental frequencies in the uhf band. For much servicing it is practicable to work with harmonic frequencies from a signal generator whose fundamentals do not extend above the vhf television band. Anything other than a crystal controlled frequency is unlikely to have sufficient accuracy for making adjustments with any degree of precision. Earlier it was shown that small fractions of one per cent of the total uhf band may throw a signal completely out of a given channel.

Sweep generators for ultra-high frequency work may operate on fundamentals or on harmonics. Accuracy of frequency calibration is of minor importance because exact frequencies must be determined by

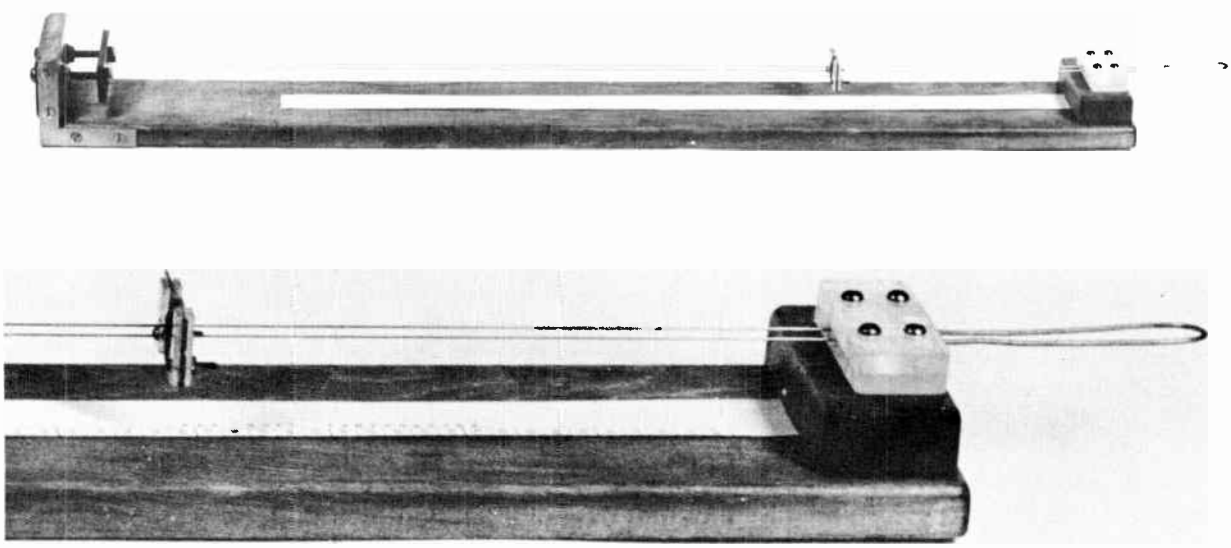


Fig. 100-16. Lecher wires whose length and spacing are suitable for frequency measurements in the uhf television band.

means of a marker. In many uhf tuned circuits the frequency response extends over several channels, with final selection of any one channel made in following circuits. This makes it desirable that a sweep generator have a greater sweep width than for most vhf work.

Vacuum tube voltmeters and other instruments commonly employed for servicing on vhf tuners are just as useful for uhf servicing, and are employed in the same ways. It is necessary to take great care in keeping instrument prods and leads as far as possible out of uhf magnetic and electric fields.

A grid dip meter or grid dip oscillator operating at fundamental ultra-high frequencies is of wide usefulness. This instrument will furnish frequencies for various tests and also will check the resonant frequency of any tuned circuit whether or not that circuit is alive at the time. If the grid dip oscillator is modulated at any frequency up to about 20,000 cycles its signals may be picked up on an oscilloscope, or as sound bars on the picture tube of the receiver. These and other uses of the grid dip meter were explained in the lesson dealing with this instrument.

In the uhf band it is not difficult to check the operating frequency of an oscillator by means of Lecher wires. As shown by Fig. 100-16, these are two parallel bare wires arranged to form an adjustable half-wave shorted line. At the right-hand end, shown by the enlarged picture, the wires are formed into a small hairpin or loop for signal pickup, and are held by an insulating clamp. The wire which extends into the pickup loop must be insulated to avoid short circuiting of tuner conductors.

At the left-hand end of the larger picture the wires are attached to a tension adjusting bracket which holds them straight and fairly taut. Whether this end of the wires is left open or is short circuited by the support makes no difference. A small bar of metal with a clean, smooth, and rather narrow edge is arranged to slide along the exposed length of the parallel wires to form the adjustable short circuit. This bar is moved back and forth by grasping it with your fingers.

With the looped end of the Lecher wires close to a signal source, waves are induced in the wires and are reflected from any point at which the shorting bar is held. The signal source may be almost any conductor in the plate or grid circuits of the oscillator. When the effective length of the line from the source to the shorting bar is any number of half-wavelengths at the oscillator frequency, standing waves are formed on the wires. The current maximum for these standing waves is at the shorting bar, as shown by Fig. 100-17. When standing waves are formed, the line absorbs maximum energy from the signal source.

You can identify the condition of maximum energy absorption by measuring the voltage across the grid resistor that biases the oscillator. This is done by connecting a vacuum tube voltmeter across the grid resistor. Then, commencing with the shorting bar close to the pickup end of the Lecher wires, and at right angles to the wires, the bar is moved very slowly toward the other end. The bar must remain in good contact with both wires at all points.

At some position of the bar the oscillator grid voltage or bias will decrease quite sharply. By making the coupling continually looser while carefully changing the position of the shorting bar it is possible to locate the position for maximum voltage dip with considerable accuracy. This first point is identified by an inch scale placed under the wires.

Now the bar is moved slowly farther away from the pickup end until locating a position at which there is another dip of oscillator grid voltage. The difference between bar positions for the two dips is found by

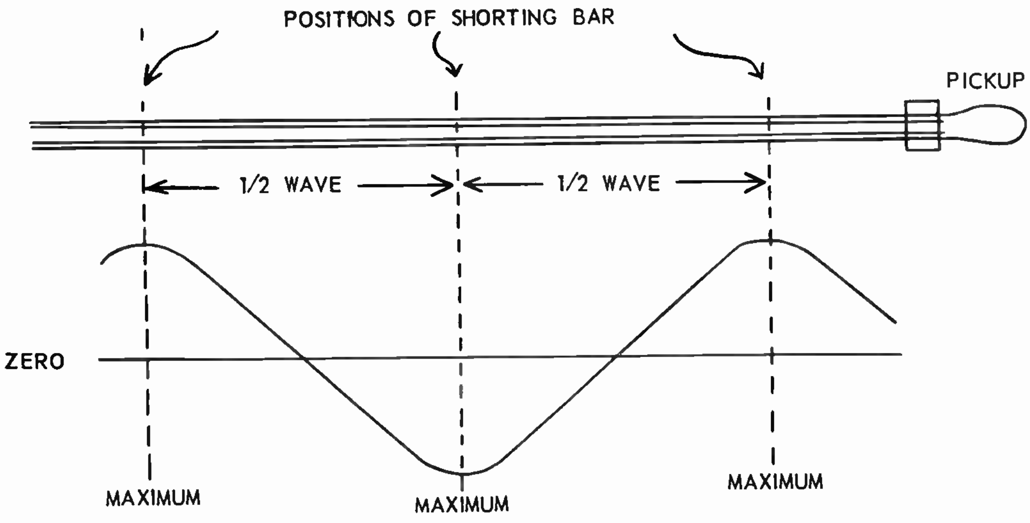


Fig. 100-17. How standing waves are formed when the Lecher wires are shorted at certain points along their length.

subtracting the first scale reading from the second one. This difference is a half-wavelength at the oscillator frequency. By dividing the number of inches of difference into 5,905 you will determine the oscillator frequency in megacycles.

The length of the exposed bare wires should be at least one wavelength, and preferably somewhat more, at the lowest frequency to be measured. This insures getting a complete half-wave onto the line, so that two dips may be located. For work in the uhf television band the parallel wires should be separated by 1/4 to 3/8 inch. The gage size or diameter of the wires is not important, almost anything available will be satisfactory. Accuracy of frequency measurement depends on using the loosest coupling which allows a recognizable dip, on moving the shorting bar back and forth until positively locating the position for maximum dip, and on carefully measuring the two positions and determining their difference.

If it is inconvenient to bring the looped end of the wires to the signal source, the coupling may be through a link. The link is made from a foot or so of twin-conductor transmission line. Both conductors at both ends of this line are bared and formed into closed loops about 3/4 inch in diameter. Both loops are to be insulated. One of them is placed close to and parallel with the pickup end of the Lecher wires while the other loop is brought near the conductor with which coupling is desired. This coupling link does not alter the effective distances between dip positions of the shorting bar.

UHF TELEVISION CHANNELS, FREQUENCIES IN MEGACYCLES

Channel Number	Limits	Video Carrier	Sound Carrier
14	470-476	471.25	475.75
15	476-482	477.25	481.75
16	482-488	483.25	487.75
17	488-494	489.25	493.75
18	494-500	495.25	499.75
19	500-506	501.25	505.75
20	506-512	507.25	511.75
21	512-518	513.25	517.75
22	518-524	519.25	523.75
23	524-530	525.25	529.75
24	530-536	531.25	535.75
25	536-542	537.25	541.75
26	542-548	543.25	547.75
27	548-554	549.25	553.75
28	554-560	555.25	559.75
29	560-566	561.25	565.75
30	566-572	567.25	571.75
31	572-578	573.25	577.75
32	578-584	579.25	583.75
33	584-590	585.25	589.75
34	590-596	591.25	595.75
35	596-602	597.25	601.75
36	602-608	603.25	607.75
37	608-614	609.25	613.75
38	614-620	615.25	619.75
39	620-626	621.25	625.75
40	626-632	627.25	631.75
41	632-638	633.25	637.75
42	638-644	639.25	643.75
43	644-650	645.25	649.75
44	650-656	651.25	655.75
45	656-662	657.25	661.75
46	662-668	663.25	667.75
47	668-674	669.25	673.75
48	674-680	675.25	679.75
49	680-686	681.25	685.75

Channel Number	Limits	Video Carrier	Sound Carrier
50	686-692	687.25	691.75
51	692-698	693.25	697.75
52	698-704	699.25	703.75
53	704-710	705.25	709.75
54	710-716	711.25	715.75
55	716-722	717.25	721.75
56	722-728	723.25	727.75
57	728-734	729.25	733.75
58	734-740	735.25	739.75
59	740-746	741.25	745.75
60	746-752	747.25	751.75
61	752-758	753.25	757.75
62	758-764	759.25	763.75
63	764-770	765.25	769.75
64	770-776	771.25	775.75
65	776-782	777.25	781.75
66	782-788	783.25	787.75
67	788-794	789.25	793.75
68	794-800	795.25	799.75
69	800-806	801.25	805.75
70	806-812	807.25	811.75
71	812-818	813.25	817.75
72	818-824	819.25	823.75
73	824-830	825.25	829.75
74	830-836	831.25	835.75
75	836-842	837.25	841.75
76	842-848	843.25	847.75
77	848-854	849.25	853.75
78	854-860	855.25	859.75
79	860-866	861.25	865.75
80	866-872	867.25	871.75
81	872-878	873.25	877.75
82	878-884	879.25	883.75
83	884-890	885.25	889.75

Training Your Mind

C.T.I.

Perfecting Your Skill