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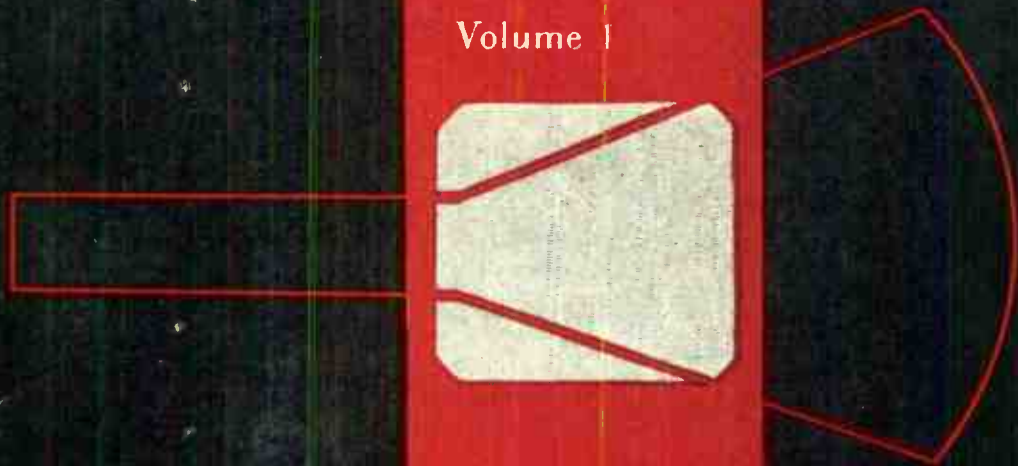
Volume I



TELEVISION for RADIO MEN

EDWARD M. NOLL

TELEVISION FOR RADIOMEN

Volume I

by EDWARD M. NOLL

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Dedicated to my uncle and aunt

Mr. and Mrs. H. Roy Eberly

PREFACE TO REVISED EDITION

When the first edition of this book was published, I wrote of how much I appreciated and recognized the contributions, guidance, and research of the many who had contributed to the development of a practical television system and whose work is necessarily a part of this television text. Today I must extend my gratitude to many, many more organizations and individuals who have since shared in the refinement and colossal growth of television. Despite the impressive expansion of television broadcasting during the few years intervening, its present status represents only a small percentage of its potential.

This edition of *Television for Radiomen* has been revised thoroughly and enlarged. The book should be even more serviceable in revision because the author has included information on new developments in television and eliminated sections which have been little used. New sections have been added and two new chapters included that present comprehensive data on UHF and transistors. The increasingly important subject of color television occupies the entire, new, second part of the book.

It is hoped that the book will encourage the student of television to pursue his studies diligently and that he will not become discouraged over inability to understand all things quickly and completely. But few things in the universe are understood with complete finality. We must learn to work with those phases of television that we do not understand completely. As we work and observe results, our knowledge and understanding increase.

It is essential that the television student develop the habit of having his mind ready to pursue and accept new knowledge, even when he must interpret an old subject differently in the light of new disclosures. He must study continuously to keep abreast of new products designed to meet specified performance levels in various price ranges. In a field of so many variables and cost factors, a "best without reservation" is almost a non-existent product.

Since there are so many differing circuit techniques, as functions of cost and application, each variation cannot be detailed, even in a text of considerable length. Our task has been to present materials necessary to the student in building the solid foundation and gaining the wide, thorough background he will need if he is to progress with the science in the future.

In conclusion I want to thank Stanton Snyderman for preparation of the fine illustrations for the revised edition and Dorothy Meeder for her helpful assistance in stenographic work.

EDWARD M. NOLL

PREFACE TO FIRST EDITION

Television For Radiomen was written to serve as a television text for practical radiomen, and television students in the final semesters of their studies. The text assumes a rather thorough basic knowledge of radio circuits, and, if practicing radiomen have been away from study for some time, it is advisable to review a good radio text. The material is arranged for progressive, orderly study in a sequence which the author has found most effective for classroom or written-lesson presentation of television. Treatment can be mathematical or nonmathematical depending on the scope of the course or the needs of the individual studying the text. Chapter 14 is a lengthy chapter on the mathematical aspects of various television circuits and gives to those who have the necessary background that added understanding necessary to handle effectively television jobs of a more complex nature. At the same time, practicing television technicians who do not require an extensive mathematical background can progress through the book without being impeded by mathematical formulas, interpretations, and derivations. Reference is made at various points in the text to the mathematical presentations in Chap. 14. As a further aid to study, questions and a thorough bibliography follow each chapter.

A technical author must necessarily feel humble when he reflects on the creative labor, organization, and toil of the many persons who have contributed to such a comprehensive field as television and, therefore, have contributed to this book. And when he considers the numerous persons who have contributed to his own education and abilities, he feels that only a little of himself, excluding some tedious work, has gotten into the work. This author feels indebted to many individuals and organizations.

For willing cooperation and permission to use manufacturers' data special thanks is given to the following men and the manufacturers they represent: A. W. Bernsohn, A. Liebscher, E. L. Clark, R. S. Burnap of RCA, E. D. Lucas of Philco, W. L. Parkinson of GE. A. C. Lescarboura representing Dumont, J. F. Bigelow of Farnsworth, F. L. Uhrus of Motorola, A. B. Friedman of J.F.D., R. K. Arbogast of Hickok, L. J. A. van Lieshout of Norelco and S. Horbach of Brach.

I thank Mr. Byron Young for the very fine illustrations. For typing and proofreading, I appreciate the careful work of the Kerner sisters, and Mrs. Evelyn Bondy.

I am grateful for the help and encouragement of officials and members of the faculties of Temple University Technical Institute and Trenton Technical School, and to Mr. Avery Pitt of Old York Road Publishing Company.

It has been my good fortune to have the guidance of the fine group of editors who manage our radio trade journals—Lewis Winner of *Service,* Oliver Read of *Radio News,* Fred Shunaman of *Radio-Electronics,* and Sanford Cowan of *Radio Service Dealer.* I thank them for permission to use again material which the author prepared originally in article form for these journals.

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PART I

Chapter 1

AN INTRODUCTION TO TELEVISION

1. *The Approach*

Television is not an involved, difficult subject—it is a comprehensive subject which touches upon almost every aspect of electronic circuits. It is actually a giant orientation problem in which each individual function must be painstakingly inserted into the complete system. Therefore, text material must consist of the fundamental approach, a discussion of the over-all system plus general functions, and, finally, a detailed discussion of individual circuits and how they fit into the entirety. This method of attack affords an ideal skeleton for assembling the television jigsaw because first you obtain an over-all insight into the system and, then, you discuss and insert each part into the whole.

To derive the most benefit from the text, proceed as follows:

1. Read over an *entire chapter* at one or, at the most, two sittings. Do not worry about details but do read the chapter to obtain the most from it.
2. Start again at the beginning. This time study every sentence and word. Really dig into it. Keep notes on subjects understood with difficulty or not clearly understood for future reference and review.
3. Answer questions carefully and to the best of your ability. If you are weak on the questions, go over the chapter again.
4. Study illustrations. Frequently a descriptive illustration will tell you more than many paragraphs of text.
5. Review often and thoroughly.

2. *Basic System*

Essential units of a basic, modern television system are shown in the block diagram of Fig. 1. At the station location there are two transmitters, picture and sound; at the receiver location, both transmitted signals are picked up by a single receiver. The following paragraphs, numerically keyed to the block diagram, will give you, the beginner in television, an initial, brief insight into the over-all functions of the various components of the system.

The picture or video signal released by the camera or pickup tube (film iconoscope in Fig. 2) is a series of electrical charges which represent the light distribution of the object or scene televised. The light distribution of the object (1) is gathered up by the *lens system* (2) and focused on the *photosensitive plate* (3) (called a *mosaic*) of the *iconoscope* (4). This mosaic is swept or scanned by a thin stream of electrons from the electron gun (5) of the iconoscope, progressively releasing in the form of electrical signals the light distribution of the image. Video signal, therefore, consists of a parade of electrical impulses that are released by a high-velocity stream of electrons which scan

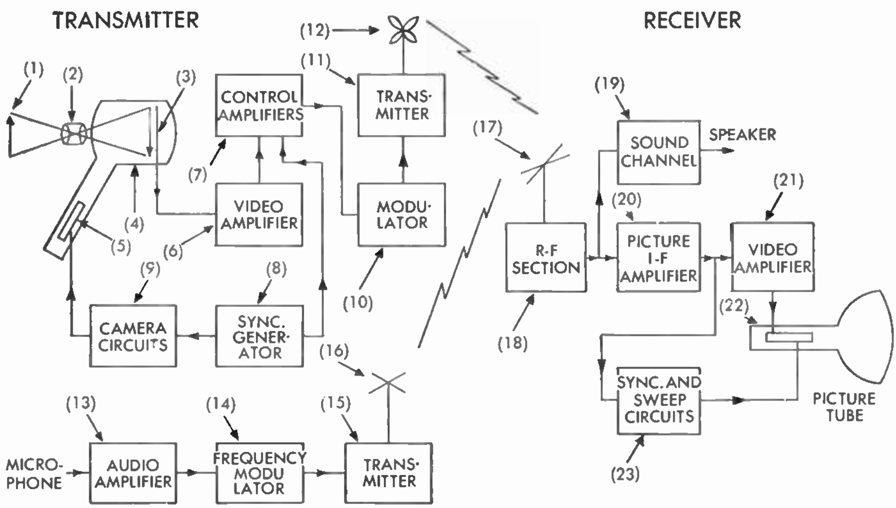


FIG. 1 Functional Diagram of Basic Television System

across and down the image on the mosaic (Fig. 2). Thus, all points comprising a television scene are not transmitted as an entirety but as a series of signals, each representing a very tiny area of the scene. However, they are developed at such high velocity and reassembled so quickly at the receiver that the human eye views the scene as an entirety.

The signal from the camera tube is many times weaker than the output from even the least sensitive microphone and must, consequently, be amplified by a *video amplifier* (6) before it reaches a useful level. Video output from this amplifier is fed into another video amplifier, called a *line-and-control amplifier* (7). Here it is combined with other signals necessary in the transmission of television signals. These signals, called *sync* and *blanking pulses,* originate at the *sync generator* (8) and synchronize, or keep in step, the entire television system. Through the *camera circuits* (9) they control the *electron-gun* scanning (5) of the iconoscope, and, at the receiver, they control the electron-gun scanning of the picture tube, keeping it in step with that of the iconoscope. Thus we have a tightly locked-in system with all timing controlled by the pulses from the sync generator.

The *sound transmitter* is a conventional frequency-modulation system consisting of *audio amplifiers* (13), *frequency modulator* (14), *high-frequency transmitter* (15), and *high-frequency antenna* (16).

Picture and sound signals are intercepted by the *receiver antenna* (17) and conveyed to the *r-f section* of the *receiver* (18). At the output of the r-f section, picture and sound can be separated. Sound i-f signal passes into a conventional *f-m limiter, discriminator,* and *audio system* (19); picture, into the picture *i-f amplifier* (20). At the output of the picture i-f amplifier and detector stages, two signals are present. One is the picture signal and blanking which pass to the control grid of the *picture tube* (22). The second signal is the

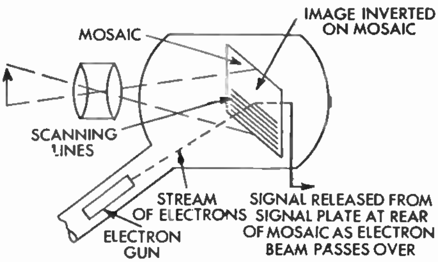


FIG. 2 Scanning with Electrons

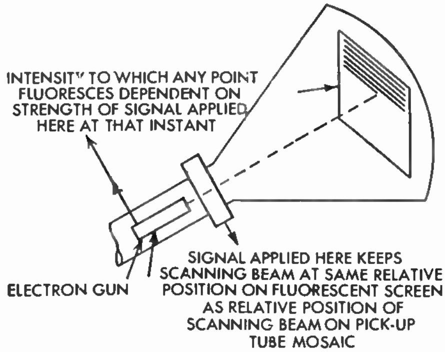


FIG. 3 Scanning at Picture Tube

sync pulses which are fed to the sync and *sweep circuits* (23). Output from this circuit controls the electron gun of the picture tube, causing it to scan the fluorescent screen of the picture tube. See Fig. 3. The picture-tube screen fluoresces or lights up in accordance with the parade of charges or picture signal applied to its control grid. The point at which the sweeping stream of electrons strikes the fluorescent screen must always be at the same relative position as the point at which the stream of electrons at the pickup tube is striking the mosaic image at that particular instant. This is the task of the sync pulses.

In summary, the television system has two primary functions: first, to gather the picture information and deliver it to the receiver; and second, to release this information in proper sequence at the picture tube. The latter of these requirements is fulfilled by transmission of synchronizing signals which cause the picture-tube scanning beam, releasing picture signal, to move in step with the pickup-tube scanning beam, gathering picture signal. Thus the picture-tube scanning beam is directed, at all times, toward the same relative position on its scanning raster as the pickup-tube beam on the photosensitive mosaic. Picture signals, therefore, strike the fluorescent screen in the same order and at the same relative points from which they have been released from the mosaic image.

It can be said that the three major units of the television system (Fig. 4) are pickup tube, picture tube, and transmitting medium. The transmitting medium conveys picture and scanning information between pickup and picture tubes. Picture signal is gathered off the mosaic signal plate and conveyed to the picture tube to modulate the picture-tube beam. Scanning is locked in by synchronizing the deflection systems of pickup and picture tubes with vertical and horizontal synchronizing or "sync" pulses.

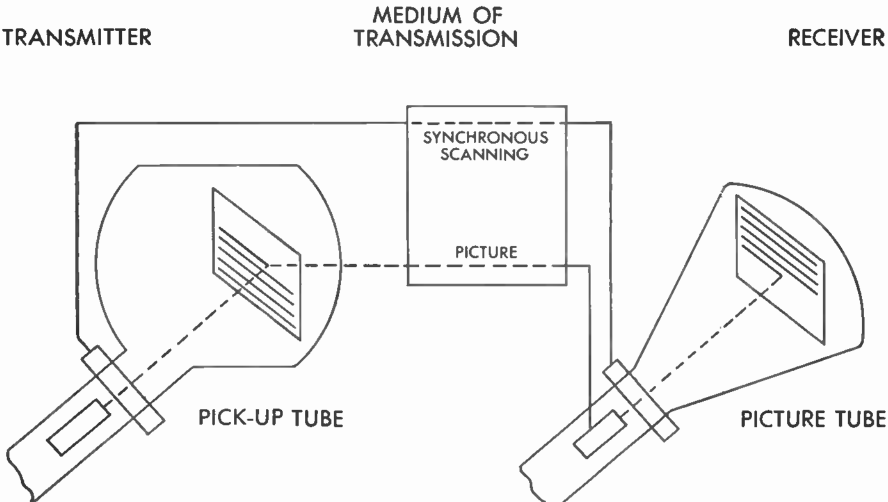


FIG. 4 Needs of the System

3. *Frequency Bands*

A single television channel is 6 megacycles wide; the entire commercial broadcast band, 1 megacycle wide. One broadcast channel is *only* 10 to 20 kilocycles wide. The very wide television channel is necessary to transmit pictures with clarity and sharpness. This wide channel means television information must be transmitted on very high frequencies to retain a satisfactory ratio of carrier frequency to bandwidth.

The commercial television channels are allocated as follows:

Channel Number	Frequency, megacycles
2	54–60
3	60–66
4	66–72
5	76–82
6	82–88
7	174–180
8	180–186
9	186–192
10	192–198
11	198–204
12	204–210
13	210–216

Ultra-high-frequency channels *14–83* occupy a span of six-megacycle channels, from 470 to 890 megacycles.

Television frequencies have also been allocated up to thousands of megacycles for experimental, relay, remote, color, high-definition, and educational telecasting.

4. *Standard Channel*

A standard television channel, shown in Fig. 5, contains both picture and sound carriers and their corresponding sidebands. Picture carrier and sidebands occupy a much greater portion, approximately 5¾ megacycles, of the total 6-megacycle band. The picture carrier is amplitude-modulated; the sound carrier, frequency-modulated with a maximum deviation of plus and minus

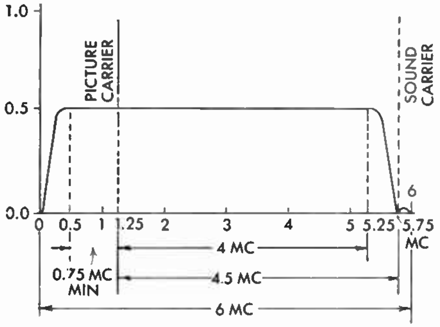


FIG. 5 Signal Distribution, Standard Television Channel

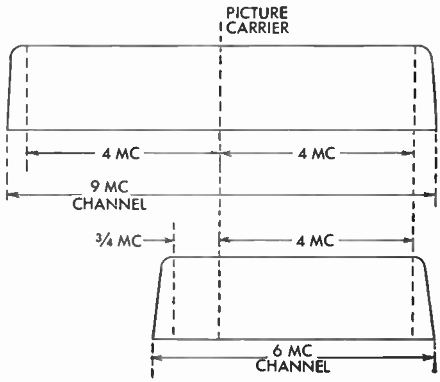


FIG. 6 Reduction of Channel Width by Partial Suppression of One Sideband

25 kilocycles. Figure 5 is scaled in terms of channel width, 0 to 6 megacycles. For example, if we want to consider the distribution of signal for channel 2, the zero point on the horizontal base line represents 54 megacycles, and the 6-megacycle point represents 60 megacycles. Thus, if we want to know the picture-carrier frequency on channel 2, we add 1.25 to 54 $(54 + 1.25 = 55.25)$; for sound-carrier frequency, we must add 5.75 to 54 $(54 + 5.75 = 59.75)$.

One very unusual feature of the amplitude-modulated picture carrier is that the high-frequency sideband is 4.5 megacycles wide and the low-frequency sideband only 1.25 megacycles wide. This unsymmetrical distribution permits transmission of a picture with better definition using only a 6-megacycle channel. It is called *vestigial sideband transmission.*

In the transmission of an amplitude-modulated broadcast carrier there is a maximum total bandwidth of 16 kilocycles, or plus and minus 8 kilocycles about the carrier frequency, which means the highest frequency audio component transmitted is only 8 kilocycles, or 8,000 cycles. Thus, with a 16-kilocycle symmetrical channel, the highest audio component which can be transmitted is 8 kilocycles. Likewise, the highest frequency picture component, leaving spectrum for the sound carrier, which could be transmitted with symmetrical television sidebands in a 6-megacycle channel is a bit more than 2¾ megacycles. Inasmuch as there are many frequency components in the picture-signal megacycles above 2¾, these components would be lost and the picture would likewise lose in definition and sharpness. The modern television station transmits components of signal up to 4 megacycles, or slightly more. For symmetrical sideband transmission, this would require a 9-megacycle channel.

It is still possible, however, to transmit a 4-megacycle component in a 6-megacycle channel by partially suppressing one sideband (Fig. 6). Inasmuch as all the required signal components are present in one sideband, it is not detrimental, if proper compensation is made at the receiver, to suppress one sideband, or, in fact, to remove it altogether. However, inherent difficulties encountered with single-sideband transmission are avoided by only partially suppressing one sideband.

Channel	Low-Frequency End of Channel	Guard Band	Flat Portion of Low-Frequency Sideband	Picture-Carrier Frequency	Flat Portion of High-Frequency Sideband	Guard Band	Sound-Carrier Frequency	High-Frequency End of Channel
2	54	54–54½	54½–55¼	55¼	55¼–59¼	59¼–59¾	59¾	60
3	60	60–60½	60½–61¼	61¼	61¼–65¼	65¼–65¾	65¾	66
4	66	66–66½	66½–67¼	67¼	67¼–71¼	71¼–71¾	71¾	72
5	76	76–76½	76½–77¼	77¼	77¼–81¼	81¼–81¾	81¾	82
6	82	82–82½	82½–83¼	83¼	83¼–87¼	87¼–87¾	87¾	88

FIG. 7 Signal Distribution on Five Lower Frequency Channels

The picture-carrier frequency is located at the 1.25-megacycle point in the channel. The high-frequency sideband is flat to the 5.25-megacycle point and then decays to almost zero in one-half megacycle. The low-frequency picture sideband is flat to the 0.75-megacycle point and then decays to almost zero in the remaining one-half megacycle. The half-megacycle bands in which the amplitude drops off rapidly are known as *guard bands.* Sound-carrier frequency is located at the 5.75-megacycle point and then, approximately, a one-fourth-megacycle guard band exists to the high-frequency end of the channel.

The distribution of signals on the lower five commercial television channels is shown in Fig. 7. It is evident from the study of the television channel that the antenna used to radiate the picture signal must not be sharply tuned to one frequency but must be flat and have a uniform radiation over a wide band of frequencies. Likewise, the receiver antenna must have a uniform sensitivity over a still wider range of frequencies, for it must also be sensitive to the sound carrier and sidebands.

5. *What's So Different about Television!*

TRANSMITTER FREQUENCY

The television picture and sound transmitters are operated on frequencies approximately fifty to two hundred times higher than the commercial broadcast frequencies. Transmission on these frequencies is necessary to radiate the very high-frequency modulation components of a picture signal. These frequencies have approximate line-of-sight range; consequently, the area over which the television signal can be picked up is limited. Thus, antenna height and location become as much of a problem as transmitter power.

FREQUENCY RESPONSES AND BANDWIDTH

The highest frequency audio component of the signal radiated by a commercial broadcasting station is approximately 8000 cycles; the highest frequency picture component of the signal radiated by a commercial television station is in excess of 4 megacycles. Consequently, television transmitter and receiver r-f, i-f, and video amplifiers must have a uniform response up to this limit. These requirements make the entire system broad, and circuits have low gain and wide bandwidths instead of the conventional high gain, high Q circuits of broadcast practice. Television circuits are designed for maximum gain and required bandwidth. It should be pointed out that the response of the system may some day be in excess of 4 megacycles as the state of the art advances and progress is made toward still higher carrier frequencies and color television. For only by increasing the upper frequency response do we obtain pictures of improved clarity and definition.

The bandwidth of a commercial broadcast station is 16 kilocycles; of a commercial telecasting station, 6 megacycles. Antennas must be very broad to radiate or intercept uniformly the extended sidebands. As the state of the art advances and definition improves, the bandwidths will become correspondingly broader. In fact, the broadness of the entire system (high video response and broadness of i-f, r-f, and antenna circuits) hinges on this one requirement—to obtain a better picture, you need higher frequency components.

VESTIGIAL SIDEBAND TRANSMISSION

The sidebands of a commercial broadcast station are symmetrical; the sidebands of the picture transmitter extend farther on the high-frequency side of the carrier than on the low-frequency side. This system permits transmission of 4-megacycle definition with only a 6-megacycle channel width.

TWO TRANSMITTERS

A television station requires both picture and sound transmitters. The picture transmitter is amplitude-modulated and occupies a 5¾-megacycle bandwidth. The sound transmitter is frequency-modulated and occupies an

80-kilocycle band. Its carrier frequency is 4½ megacycles higher than that of the picture carrier. The same antenna and receiver pick up both picture and sound, segregating one from the other in the converter, i-f amplifier, or video amplifier.

SIGNAL CONSTRUCTION

The useful modulation on the carrier of a broadcast transmitter is audio (speech, music, sound effects, etc.—all audio information). The modulation applied to the picture carrier is not always picture signal (signal which represents light distribution on the image) but also rectangular sync and blanking pulses which stabilize the entire system. Thus, the picture information is frequently interrupted to insert sync and blanking pulses. While radiation from the aural broadcast station contains sinusoidal variations, the television signal contains some nonsinusoidal waveforms as well.

6. *What Is a Sine Wave?*

The sine wave is the basic, alternating waveform. All other waveforms, square, rectangular, sawtooth, etc., can be considered to consist of combinations of sine waves with various amplitudes and frequencies. A sine wave is generated when a single conductor is rotated at a constant velocity around a uniform magnetic field. The generation of a one-cycle-per-minute sine wave is demonstrated in Fig. 8. When the conductor is moving parallel to the magnetic field, no voltage is induced; such is the case at the 0-, 30-, and 60-second points. At the 15- and 45-second points, the motion of the conductor is perpendicular to the field and maximum voltage is induced. Note that the polarity of the voltage at the 15- and 45-second points would be opposite because the conductor travels through the lines of force in the opposite direction.

Actually, therefore, a sine wave is a circle stretched out in terms of time just as we have done in Fig. 8. The circumference and horizontal axis have been divided into equal time segments. However, it is to be observed that equal time segments do not set off equal voltage segments, which would result in a triangular wave. Instead, as shown, the voltage increments become less and less as the point of maximum induced voltage is approached, generating the characteristic sine wave. In the generation of higher frequencies, the conductor revolves around the magnetic field more often and more sine waves per minute are generated.

A similar wave is generated in high-frequency tuned circuits by the charge and discharge of the tuned-circuit capacitor and the expanding and contracting field of the tuned-circuit inductor.

An important fact to remember is that it is impossible to distort a sine wave with a resistor, capacitor, or inductor; or to distort a sine wave with any combination of resistors, capacitors, and inductors normally used in electronic circuits. The only characteristic changed is the amplitude. If it is desired to change a sine wave to some other type of waveform, a vacuum tube is generally used. It is only a *complex wave* made up of combinations of sine waves which can be distorted or changed in shape by combinations of resistors, capacitors, and inductors. This change in wave shape occurs because of the changing reactance of capacitors or inductors with variations in frequency and

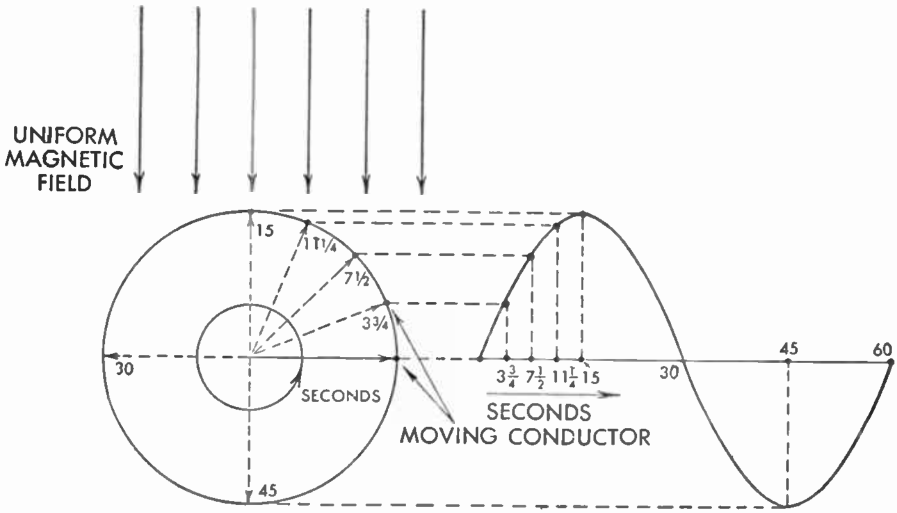


FIG. 8 Generation of a Sine Wave

because of the phase shift of some frequencies with respect to others. Such changes in wave shape are due to the fact that complex waves are combinations of many sine waves of different frequencies, all multiples of the basic frequency; emphasizing or eliminating any of these frequencies changes the shape of the wave.

7. *What Is a Pulse?*

A rectangular pulse (Fig. 9) is spoken of as a *composite wave* because it does not consist of one sine-wave frequency but rather is considered to be made up of many sinusoidal waves harmonically related up to at least the tenth harmonic. Actually, in the case of a square or nearly square wave (drawing *A*, Fig. 9), the fundamental frequency of the pulse is considered to be the repetition rate of the pulse (number of pulses which occur per second), and to have a properly squared appearance it is necessary that the pulse contain harmonic frequencies up to the tenth. The more harmonic frequencies present, the sharper the wave rises and falls.

For rectangular sync pulses used in television, the interval between pulses (Fig. 9, *B*) is long in comparison to the duration of the pulse. For example, the duration of one of the television pulses is only 5 microseconds (horizontal sync pulse) and the spacing between pulses some 58 microseconds. When the pulse is short in duration in comparison to the separation between pulses the important frequency components of the pulse and of the system are dependent on the duration of the pulse and less dependent on the repetition rate. It is true that the fundamental frequency is the repetition rate of the pulses and that the pulses consist of a series of harmonics of this frequency extending up to almost 2 megacycles. However, the small energy content of the fundamental and low-

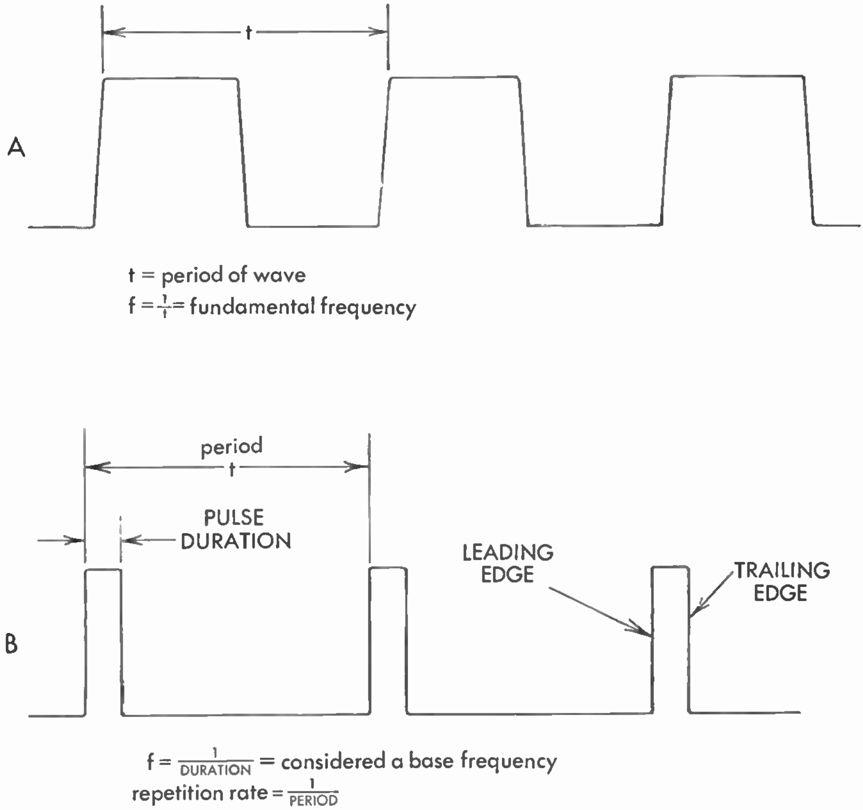


Fig. 9 Characteristics of Rectangular Pulses

order harmonics compared to the total energy vested in the pulse means they can be attenuated without affecting the fidelity of the pulse. In fact, a base frequency from which performance of pulse circuits can be judged is a frequency the period of which equals the duration of the pulse. To form or pass a pulse with a good squared appearance, the pulse must contain frequency components up to the tenth harmonic of this so-called "base frequency." For example, the base frequency of a horizontal sync pulse is one over its duration, or 1/0.000005, or 200,000 cycles per second. The tenth harmonic of this frequency is 2 megacycles.

More exactly, the highest order harmonic is set by just how fast the pulse must rise or fall between minimum and maximum. In the case of horizontal sync pulses, this sets the highest order harmonic almost to 2 megacycles. The extreme of the low-order harmonics is set by just how flat the top of the pulse must remain; the flatter the level, the lower the response requirements must be. Again, the low-frequency limit is related to the duration of the pulse, and the longer the pulse, the lower is this frequency.

Inasmuch as the video amplifiers of the television receiver and transmitter equipment have a very much extended frequency range to pass the picture information, no difficulty is involved in transferring the sync and blanking pulses.

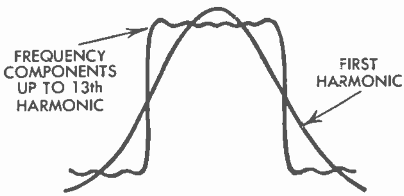


FIG. 10 Formation of a Pulse

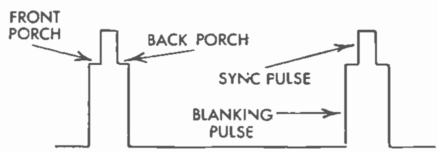


FIG. 11 Short-Duration Pulse on a Pedestal

The cumulative effect of the fundamental and harmonics in the formation of a squared pulse can be observed in Fig. 10. It is important to note that a pulse is a complex waveform constructed of widely separated frequency components. In an amplifier which is to pass this pulse with fidelity, it is necessary that the amplitudes and the phase relations between fundamental and harmonics not be disturbed. Thus, the amplifier must have a linear phase and frequency response up to at least the tenth harmonic of the fundamental frequency of the pulse. It is possible to distort the shape of a pulse for particular applications with a simple resistor-capacitor or resistor-inductor combination because of the varying reactance and phase with frequency of inductors or capacitors.

The sharp rise of voltage from the base line to peak value, the pulse duration, and the sharp drop from peak amplitude back to the base line are characteristics of pulses which are used to advantage in television circuits. It is the sharp rise in voltage, called a *leading edge,* or occasionally the sharp fall, called a *trailing edge,* of the rectangular pulses which accurately times the television system and synchronizes the motions of the two scanning beams. An opposite relation exists for a negative pulse.

It is also possible to construct a waveform consisting of one rectangular pulse with another shorter-duration pulse mounted on top of it. It is often spoken of as a *pulse mounted on a pedestal* (Fig. 11). The flat portion of the lower pulse just ahead of the leading edge of the top pulse is called a *front porch.* In television vernacular the lower pulse is a blanking pulse and the top a sync pulse.

8. *What Is a Sawtooth?*

A sawtooth waveform (Fig. 12) consists of its fundamental frequency (number of sawteeth per second) plus, for a satisfactory sawtooth, harmonics of the fundamental up to the tenth or fifteenth (Fig. 13). This means, of course, that the frequency response of the circuits through which the sawtooth voltage passes must be reasonably flat up to the frequency of the final harmonic. As you will learn later in this chapter, the frequency of the horizontal sweep of the television receiver is 15,750 cycles per second, and therefore the horizontal circuits must be capable of passing frequencies to at least ten times this value, or 157,500 cycles.

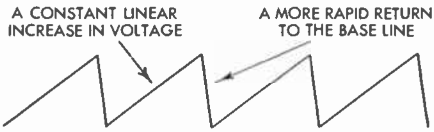


Fig. 12 Sawtooth Waves

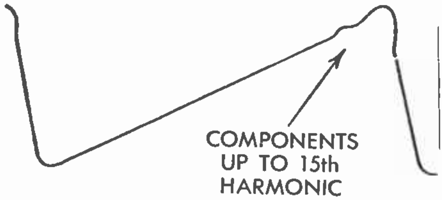


Fig. 13 Components of a Sawtooth

The sawtooth voltage can be used to advantage in television circuits because its waveform (Fig. 12) represents a slow, fixed rate of voltage increase and a fast decrease back to the base line. This characteristic, you will learn, is used in television scanning because it is necessary to move the scanning beam at a constant velocity across picture-tube screen and camera-tube mosaic and then quickly return it to begin a new line. If a sine wave were used to deflect the scanning beam, the velocity of beam as it moves across the screen would decrease as it approaches the right or left side, causing the reproduced images to appear out of proportion.

When the scanning beam of the pickup or picture tube is deflected by electrostatic deflection plates, the sawtooth voltage is used; when electromagnetic coils are used to deflect the beam, a sawtooth current is needed.

9. *What Is a Signal with a D-C Component?*

It is a well-known fact that in the plate circuit of an audio amplifier, for example, there is an a-c component of plate voltage and a d-c component. In the case of a resistance-coupled amplifier, the d-c component of plate voltage might be 100 volts and the a-c component might have a peak-to-peak amplitude of 100 volts with the instantaneous plate potential varying between 150 and 50 volts, as in Fig. 14. In transferring the signal to the grid circuit of the next stage, the d-c component is blocked by the coupling capacitor. Likewise, in television circuits we have signals with d-c components; however, the d-c components of a television signal contain useful information and must be transmitted. Thus, circuits must be used which pass d-c and a-c signal components.

A typical portion of the television signal (Fig. 15), therefore, contains continuously varying picture signal plus the d-c blanking and sync-pulse components plus a d-c component of average brightness, which varies with average illumination of the televised scene. Notice that the television signal, consequently, has a single polarity; that is, it varies from some base voltage to some fixed positive voltage or, with a 180-degree polarity shift, from zero to some fixed negative voltage. The picture signal consists of amplitude variations between blanking pulses and confined within the voltage range between base and the so-called "blanking level." When blanking and sync are plus with respect to base, the signal is referred to as a *positive-going composite signal; negative-going* when sync and blanking are negative.

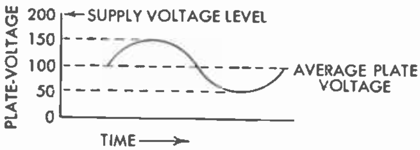


FIG. 14 A-C and D-C Plate Current Components

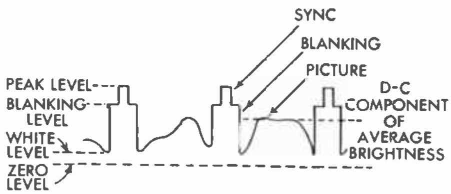


FIG. 15 Portion of Television Signal Showing Uni-Polarity Voltage

In the last few paragraphs we have discussed the various waveforms encountered in the television circuits. In subsequent paragraphs we will discuss the application of these waveforms.

10. *Television Scanning*

Television scanning can be understood by comparing the television scanning process with the scanning process of the conventional cathode-ray oscilloscope. In the conventional oscilloscope a horizontal sawtooth is applied to the horizontal deflection plates or deflection coils. This sawtooth voltage causes the

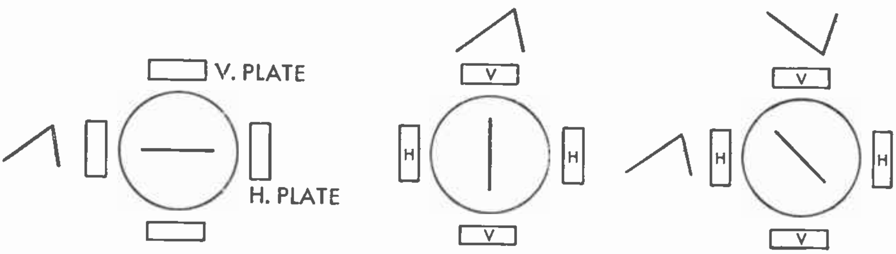


FIG. 16 Scanning Process

beam to travel from left to right across the screen, and then the retrace causes the beam to snap back quickly to the left side again. Thus, with no signal applied to the vertical amplifier of the oscilloscope, we find that a straight line is traced on the fluorescent screen. (Refer to Fig. 16.) Likewise, it is possible to apply a sawtooth to the vertical deflection system of the oscilloscope. If no signal were applied to the horizontal plates under this condition, a vertical straight line would be traced on the fluorescent screen. If the same frequency sawtooth were applied to the horizontal deflection plates as applied to the vertical deflection plates (Fig. 16), a diagonal straight line would appear on this screen.

When a sawtooth voltage of one frequency is applied to the vertical plates and a sawtooth voltage of a higher frequency is applied to the horizontal plates, the beam will make a number of trips left and right as it moves from the top to the bottom of the screen. This series of tilted lines is generated in approximately the same manner as the television scanning raster is formed on the front of the picture tube. In the case of the television system, the frequency of the vertical sawtooth is 60 per second and the frequency of the horizontal saw-

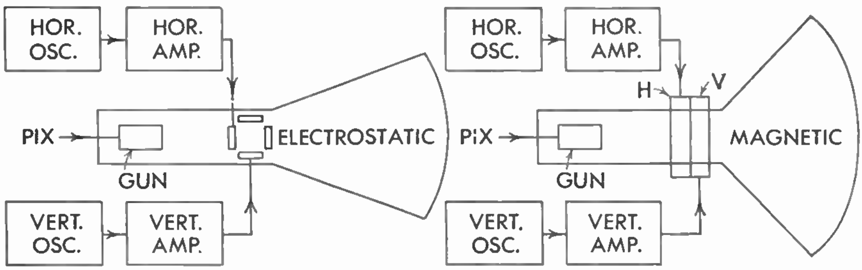


Fig. 17 Television Scanning

tooth is 15,750 per second. It is evident how much faster the horizontal sawtooth is than the vertical sawtooth. In fact, so many lines are generated horizontally in comparison to the time required for the beam to move from top to bottom that each of the lines generated appears almost as a straight, horizontal line. The tilt of these lines, although it is present, is so small that it is not discernible, and the lines themselves are so near to each other that they are discernible only upon close observation of the picture-tube screen. Actually, approximately 500 lines are traced and are present from top to bottom of the television screen.

The basic units associated with the picture-tube deflection system are shown in Fig. 17. The essential differences, as compared to a conventional test oscilloscope, are:

1. A sawtooth voltage is also applied to the vertical deflection plates of the picture tube which is much lower in frequency than the sawtooth applied to the horizontal plates.

2. The entire front of the television screen is illuminated because many, many lines are generated from top to bottom of the screen.

3. In a test oscilloscope, the waveform to be observed is always applied to the vertical deflection plates. In television, the picture signal is applied to the

control grid of the picture tube and the electron beam is modulated. This modulated beam causes the intensity of the fluorescent screen to vary as the beam travels line by line from top to bottom of the screen. In a test oscilloscope the control grid of the oscilloscope tube is usually unmodulated.

In summarizing the scanning action with reference to Fig. 18, assume that the beam is at the top left-hand corner of the screen. A sawtooth voltage is applied to the horizontal plates, and a slower sawtooth is applied to the vertical plates to trace a simple scanning pattern on the screen. The first horizontal

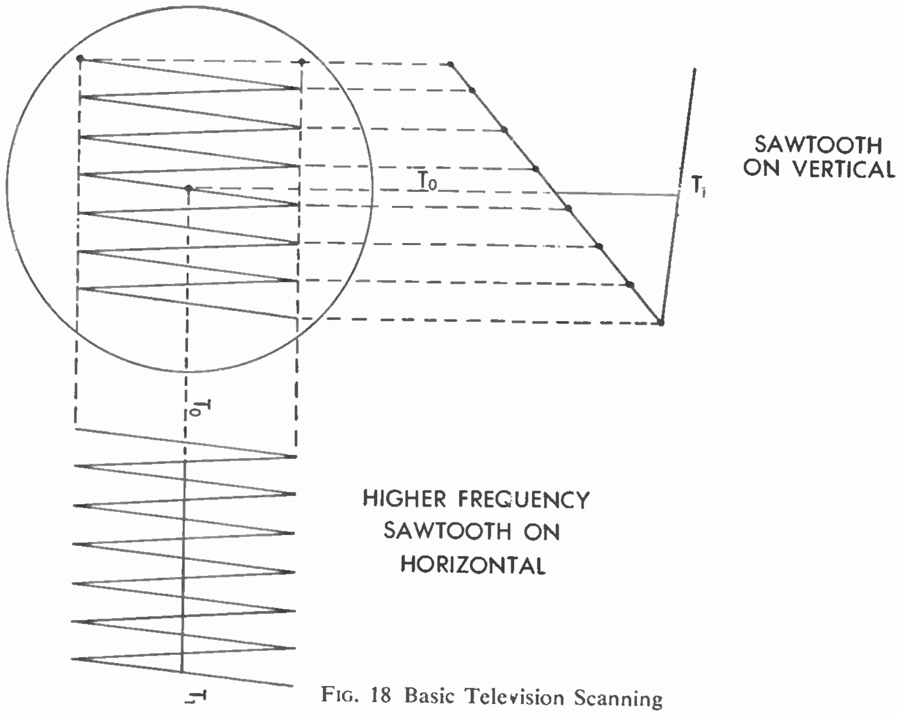


FIG. 18 Basic Television Scanning

sawtooth causes the beam to sweep across the screen and then quickly return to the left side again. The beam always returns much faster than it first sweeps across the screen from left to right because of the much faster change in voltage per unit time associated with the retrace period of a sawtooth wave. Usually, in test oscilloscope application the return of the beam is so fast that the fluorescent screen is not illuminated because of the speed with which the beam passes over the screen. At any rate it is very dark in comparison with the trace and does not interfere with the observation of the desired waveform. In television we make doubly sure that the retrace is not visible because, in addition to retracing the beam very quickly with the sawtooth retrace, a blanking pulse is also applied to the grid of the picture tube during this interval, which completely shuts off the picture beam.

After the beam has made one trip across and back, it is ready to start another line. This time, however, it starts a bit below the previous sweep line because the slow, vertical sawtooth voltage has decreased a small amount. In this manner the beam is swept back and forth across the screen starting at the upper left and ending at the lower right. Notice that the scanning line slopes more from left to right than it does during the snapback or horizontal retrace from right to left. This is understandable when we consider that the beam moves much slower in going from left to right than it does in returning from right to left. Likewise, the beam moves slowly down the screen, tracing line after line, and then quickly snaps back to the top left to begin a new series of scanning lines. This pattern is similar in many respects to television scanning. At any rate, it gives you that first, simple, but necessary, glimpse of what is meant by scanning patterns.

11. *Scanning Raster and Aspect Ratio*

The scanning raster is that part of the fluorescent screen illuminated by the sweeping electron beam. When the receiver deflection circuits are properly adjusted, this illuminated scanning raster has width-to-height ratio of 4:3. That is, if the scanning raster is 4 inches wide, to have properly proportioned images on the screen it must also be 3 inches high. This is termed the *aspect ratio* of the television system.

The scanning raster is present whether a signal is present or not. For example, when we turn on our home television receiver, the illuminated scanning raster appears whether signal is received or not. In fact, if the scanning raster is observed closely with no signal being received, the actual scanning lines can be seen, and still clearer are the vertical retrace lines, which are returning the beam from the bottom to top of the scanning raster. Of course, these return lines are visible because, with no signal received, there are no blanking pulses present on the grid of the picture tube.

12. *Standard Interlace Scanning*

In modern television systems, scanning consists of 525 horizontal lines, which are covered 30 times (called *frames*) per second, making a total of 15,750 (525×30) lines scanned per second. Thus, the electron beam at the pickup tube, under control of the electron gun, moves across the photosensitive mosaic, covering the mosaic image left to right and top to bottom, progressively releasing and gathering the light distribution of the image. Also under control of an electron gun, the picture-tube beam scans the fluorescent screen in the same manner. The electron beam at the picture tube, however, is modulated and reproduces the original variation in light and dark as it travels line by line. To reproduce the picture exactly as it is picked up at the camera tube, it is necessary that the electron guns of both tubes direct their beam at the same relative position on both fluorescent screen and scanned camera image.

In the television system the rise of voltage of the horizontal sawtooth moves the electron beam from left to right across the screen in approximately 53 microseconds. When the beam reaches the right side, the faster change in voltage (abrupt decline of the sawtooth) rapidly returns the beam from right to left in less than 10 microseconds. This motion is repeated 15,750 times per second. However, in the television picture tube we not only apply a horizontal sawtooth but also a much slower vertical sawtooth of only 60 cycles per second. This linear fall of the vertical sawtooth moves the beam from top to bottom of the scanning raster in approximately 15,800 microseconds (almost 300 times longer than it takes the beam to move from left to right across the screen). When a beam reaches the bottom, the faster change in voltage of the vertical sawtooth returns the beam to the top in approximately 850 microseconds.

With application of the sawtooth waveforms to both sets of plates or coils, the beam is moved rapidly across the screen (horizontal sweep), and at the same time it is moved slowly down the screen by the slower vertical sweep. However, in television scanning the downward tilt of the scanning line is not discernible because the horizontal sweep is so much faster than the vertical sweep. The tilt is just sufficient to cause each horizontal line to begin slightly below the previous one. When the beam reaches the bottom of the scanning raster, it is returned to the top by the abrupt increase of the vertical sawtooth. The return to the top of the raster is not made directly, for even during the relatively fast vertical retrace a number of horizontal sweeps occur and the returning trace weaves back and forth up to the top of the raster. Thus we find that, although the modern television system is a 525-line system, not all of these lines are active lines (by active lines are meant lines upon which actual picture information appears); some of them are not used, occurring during the vertical retrace intervals.

Another unusual feature, but a very logical one so far as performance is concerned, is the use of interlaced scanning. Scanning of the horizontal lines is not performed in sequence but instead the odd-numbered lines are scanned first (1, 3, 5, 7, etc.), and then the beam returns and scans the even-numbered lines (2, 4, 6, 8, etc.). This means the entire raster is scanned twice—first time, odd-numbered lines; second time, even-numbered—before the complete scene is scanned. We have learned that 30 complete frames or scenes are transmitted each second. Because of the inertia of the human eye and fluorescent screen, motion can be simulated by transmitting individual scenes at this rate. However, during the vertical retrace the scene is completely blacked out (to make vertical retrace invisible), and if the blacking-out occurs at only a 30-cycle rate, there is some noticeable flicker. It is apparent with interlaced scanning that the beam is retraced twice each frame, or 60 times per second, which is fast enough to eliminate annoying flicker. In the same manner in motion picture practice each motion picture frame, of which there are 24, is projected twice to produce an equivalent repetition rate of 48 projections per second. Thus, in television scanning there are 30 frames (complete pictures) per second and 60 fields (two fields per frame).

The interlaced method of scanning is demonstrated in Figs. 19 and 20. Scanning begins at line 1 at the top left of the scanning raster and proceeds to the right; at this point the beam retraces to the left and begins scanning line 3. The odd-numbered scanning cycle continues until the beam reaches the lower

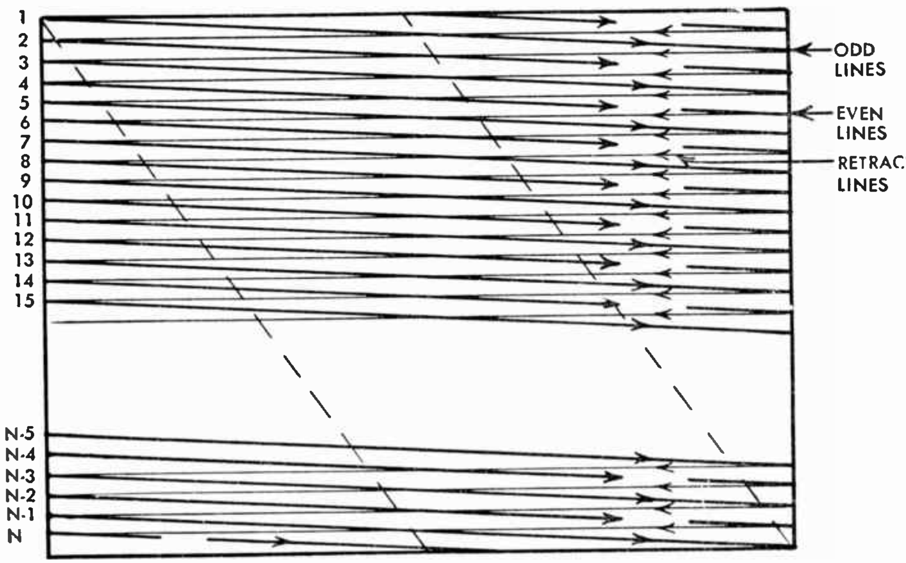


FIG. 19 Scanning Raster Showing Interlaced Pattern

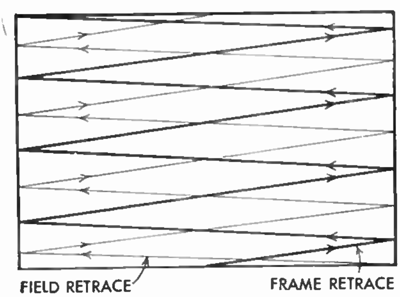


FIG. 20 Vertical Retrace Paths of Scanning Beam

right-hand corner of the raster. Here the vertical retrace returns the beam to the top center of the scanning raster and the scanning of the even-numbered lines begins. When the last active line is scanned, ending at the bottom center of the raster, the beam is returned to the top left and initiates the start of a new frame. Each frame consists of 525 lines, and 30 frames or complete pictures are transmitted each second. However, because of the interlaced system, the vertical sweep rate is not 30 but 60 cycles per second because each frame is scanned twice from top to bottom.

Although each frame consists of 525 horizontal lines, only approximately

500 of these lines actually carry picture information; the remainder of the horizontal lines are unused during the vertical retrace intervals. There are two vertical retrace intervals for each frame—one between fields, the other between frames. The vertical-retrace interval lasts for approximately 850 to 1,300 microseconds, twice each frame, and consumes approximately 25 to 40 horizontal lines. Thus, it is apparent that the beam does not return directly from bottom to top, as shown in Fig. 19, but returns over a path similar to that shown in Fig. 20, there being quite a number of horizontal sweeps during the vertical retrace.

It is evident that a considerable amount of time is consumed in retracing the beam, and for these intervals no picture information is transmitted. Furthermore, it is necessary to blank out the electron beam (causes no excitation of fluorescent screen) during these intervals to prevent visible retrace lines from streaking across the picture on the screen. In fact, the only time the electron gun emits a strong beam, and the only time useful picture information is transmitted, is during traces or scans from left to right of the 485 to 500 active lines. During the retrace intervals, both horizontal and vertical, the electron gun cuts off the beam and no picture information is conveyed from transmitter to receiver.

13. *Signal on Control Grid of Picture Tube*

Two primary functions of the signal on the control grid are to impress picture information on the fluorescent screen during active horizontal traces and to cut off the beam to prevent screen excitation during all retrace intervals. Thus, for a certain length of time (approximately 53.3 microseconds for each active horizontal line), picture information is impressed on the grid. At the conclusion of the active scan the voltage goes sharply negative to the blanking or black level (grid-voltage level at which the beam is cut off) and holds at this level or becomes more negative for approximately 10.16 microseconds. The horizontal sync pulse is actually more negative and is used in another section of the receiver to initiate the horizontal retrace. Its presence on the control grid of the picture tube has no detrimental effect because it extends into the so-called "blacker than black" region, there being no indication of it on the fluorescent screen.

The blanking or black level of the signal (Fig. 21) is always held constant regardless of picture signal or its average voltage level. The maintenance of this fixed-level d-c component is necessary because it is the instantaneous grid-voltage level at which the beam is cut off. We want the beam to be cut off during all retrace intervals regardless of what has been transmitted between retrace intervals. The voltage level at which the control grid causes the beam to cut off is comparable to the cutoff grid bias of an ordinary vacuum tube, it being fixed for a given set of operating voltages. In order to reach this level, a signal of so many volts must be applied. This condition is clearly shown in

Fig. 22, in which the grid-voltage level of $-Ec_2$ must be reached by the signal before the screen blacks out.

Observation of the signal also shows that the greater the signal amplitude, the less the screen is illuminated. Thus, the higher amplitude level of signal at *A* (Fig. 21) causes brighter instantaneous illumination (less negative grid voltage) than the lower level of signal at *B* (more negative grid voltage), which has driven the grid almost to the screen cutoff level (Fig. 22).

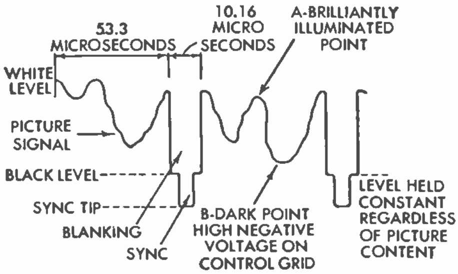


FIG. 21 Signal on Control Grid of Picture Tube

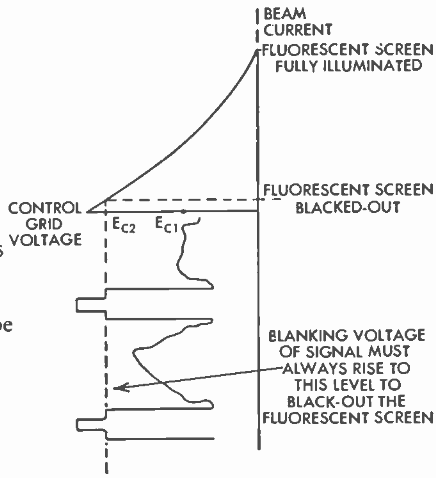


FIG. 22 Positioning of Signal on Picture-Tube Characteristic

The signal applied to the control grid also rises to the blanking level during the vertical retrace intervals. In fact, during these intervals the beam is off not merely for 10 microseconds but for more than 850 microseconds to give the beam time to retrace from bottom to top of the scanning raster. When the beam is off for such a long interval, it must occur often enough to prevent flicker. This condition can be compared to our home lighting power, which alternates 60 times per second but still the light appears continuous.

14. *Television Image*

We have studied the scanning process and mentioned the manner in which the signal is removed from the mosaic progressively, and then reassembled at the receiver in the same sequence.

The mosaic upon which the object is focused at the camera tube is constructed of many thousands of photocells insulated from each other. Each cell or group of cells emits electrons in accordance with the light intensity of the very tiny portion of the scene focused on it. An electrical representation of this illumination is released when the beam passes over this particular group of cells. If each cell contributed a signal singly, the number of cells would represent the number of elements which make up the television image. Actually, each signal component is contributed by a number of cells, and the number of elements making up the picture is limited by other factors.

The television image, similar to a photoengraving or motion picture frame, is made up of a number of elements, each element representing the light gradation at one tiny spot of the scene. Unlike the motion picture, which flashes all elements of a scene on the screen simultaneously, the television system transmits the elements in sequence, one at a time. The disassembly and assembly of all elements occur at such a rapid rate that all elements of the scene appear to be present simultaneously as far as the eye can tell.

The idea of picture elements can be understood by observing the poorer grade newspaper halftones. Observation shows that the picture is constructed of black dots or elements. On the darker parts of the scene the black dots are larger while light portions contain very small dots. A greater number of elements of a smaller size produce a more detailed and clearer photograph. The elements are so numerous and small in a first-grade reproduction that they are not discernible.

Although the scanned surface of the camera tube contains millions of elements, the actual picture element size is considerably smaller because of the finite diameter of the scanning beam of the camera tube. The camera-tube spot size is large enough to encompass many, many elements on the scanned surface at any one instant. Although the camera-tube beam imposes this restriction on picture elements, a more severe restriction is placed on effective picture elements by the frequency response of the system and the size of the picture-tube spot. Thus, in the television system, element size is more or less uniform (except for limited change in picture-tube spot with intensity of beam), and picture gradations are set by the intensity of the spot instead of its size.

There are approximately 400 to 600 elements along a single line of picture. This number is largely a function of the camera, frequency response of the system and picture-tube spot size. Inasmuch as the camera-tube spot is in most cases smaller than the picture-tube spot, the number of elements is limited by the picture-tube spot. The number of elements along each scanning line determines the picture detail horizontally and is spoken of as the *horizontal resolution.*

The number of elements vertically, which is spoken of as *vertical resolution,* is a function of the number of active scanning lines—485 or more. The number of vertical elements is less than the number of active lines because of the inability of the scanning beam always to reproduce exactly a change in illumination vertically. For example, if a thin, black, horizontal line (one line wide) is to be conveyed, it must be completely covered by the beam to reproduce accurately. If the scanning beam is half on and half off this line (it could just as well be a point of change from white to black or black to white), the line at that point will reproduce grey. Thus, the change from one degree of illumination to another is gradual instead of sharp because of the relation between picture information and position of scanning beam (Kell factor), and vertical resolution is thereby limited. There is also a small, finite spacing between lines and a slight shifting of lines. Thus, the number of vertical elements is somewhat less than the number of active lines—approximately 350 or more elements vertically. The total number of picture elements (product of elements spaced vertically and elements along each horizontal line) is approximately 200,000, substantially fewer than the number associated with motion pictures. As the state of the art progresses, the television image will contain more elements and improved detail.

The actual picture signal representing the element-to-element illumination is applied to the control grid of the picture tube. Picture signal does not vary the velocity of the beam nor in any way affect the deflection mechanisms However, it does modulate or vary the number of electrons in the beam and consequently varies the degree of illumination of the points it strikes. Picture information is present on the control grid of the picture tube when the beam is sweeping left to right across the screen. It is not present when the beam is being retraced by the deflection system. Instead, during these intervals a so-called blanking signal appears on the control grid to make the retrace invisible.

The point to remember is to keep picture and scanning separated. Picture signal varies the strength of the beam and, consequently, the instantaneous illumination of the screen; scanning is the process by means of which the beam follows a continuous and repeating path across and down the fluorescent screen.

15. *Video Frequencies*

Just as the sinusoidal variations associated with sound are called *audio frequencies,* those associated with picture information are spoken of as *video frequencies.* These frequencies are generated when the electron beam of the pickup tube moves from left to right across the mosaic, releasing electrical charges in accordance with the intensity of the light focused on the various elements scanned. One complete frame or picture is transmitted in 1/30 second, and one complete line of this picture is transmitted in 53 microseconds. Now, let us assume that as our beam is moving left to right across the screen it hits one element which is illuminated and then passes on to an adjacent element which is brighter and on to a third element which is illuminated to the same degree as the very first one. Thus, in this short time, if we consider the signal released to be sinusoidal, we have gone from a maximum to a minimum and back to a maximum, or *one complete cycle.* The frequency of this cycle can be approximated if we consider the time necessary to scan these three elements. If there are 500 elements along a horizontal line and it requires 53 microseconds to scan a full line, it requires approximately 0.2 microsecond to travel between the three elements. This represents a base frequency of 5,000,000 cycles per second.

It is apparent from the above figures how necessary it is to use wide-band amplifiers and broad channels to convey a picture with clarity and sharpness. The video or picture signal has frequency components from 30 cycles to 5 megacycles.

16. *Picture Resolution*

Resolution of a television image is a measure of picture definition and clarity. A picture with good resolution has sharply defined objects and no blurring or running together of closely spaced image lines or points. Resolution of the television image is measured both horizontally and vertically. Close-spaced vertical lines of a televised image or chart are used to measure horizontal resolution; close-spaced horizontal lines, vertical resolution. This is apparent when we consider that as the beam scans the elements horizontally it is crossing, line after line, any series of vertical lines. As the beam moves downward, it will be crossing any horizontal lines that might be present on the televised picture. Actually, the station *test charts* transmitted by the television stations contain converging horizontal and vertical lines to check resolution of the system.

Vertical resolution is limited by Kell factor and the number of lines—the more active horizontal lines, the better the vertical definition. Horizontal resolution is limited only by the capabilities (frequency and phase response in particular) of the system and the size of the pickup and picture-tube scanning spots. The iconoscope pickup scanning spot and density of the silver globule deposits are at present capable of much better performance than the remainder of the television system is able to handle. Chief limitations are the frequency response of the various picture video, i-f, and r-f amplifiers, plus the size of the picture-tube scanning spot. It must be pointed out that video, i-f, and r-f amplifiers can be made broader with proper design.

17. *Persistence of Vision and Flicker*

It is a characteristic of the human eye to retain for a finite time an impression of light after the source has been removed. It is this very characteristic which permits transmission of a picture, element by element, and permits the creation of the impression of motion by transmitting individual pictures at a fast repetition rate.

At any one instant in the picture-tube scanning cycle, only one element is being illuminated by the impact of the scanning beam. The remainder of the elements remain illuminated, or appear to remain illuminated, for two reasons: one, the actual element remains illuminated or dims slowly after the passage of the beam; and two, the human eye retains an impression of light after the beam has passed on. Thus, to the eye, all elements appear illuminated, each in accordance with its relative light value in the televised scene. In the television system, if each element (approximately 200,000 of them) is scanned once every 1/30 second, the human eye retains an impression of continuous illumination. That is, light impression from the first element in the first row still persists at the time that the last element in the last row is being scanned.

To create an impression of motion, a certain number of closely associated yet individual scenes must be transmitted each second. To give motion to the characters in a motion picture, 24 individual scenes are transmitted each second. Now, during the interval that one scene is being replaced by the next, there is no light projected on the screen. However, the screen remains illuminated to the human eye because of its persistence of vision. Likewise, the impression of motion is created in television by transmitting 30 individual scenes each second, and the human eye retains impression between frames. To avoid even the smallest trace of flicker, however, each frame is subdivided into two fields (interlaced scanning) to jump the repetition rate to 60 per second.

Flicker is caused when light impulses are presented to the eye at a slow rate, allowing illumination to decay appreciably before being reinforced by a new impulse. Inasmuch as the fluorescent screen of the picture tube is blanked or blacked out a considerable portion of the frame-scanning cycle, proper precautions must be taken to avoid flicker and consequent viewing fatigue.

After each line is scanned, the beam is cut off until the beam retraces. The beam is again blanked for a longer duration after a field is scanned and the beam is retracing from bottom to top of the raster. Furthermore, each picture element is scanned individually and at any one instant only one element is receiving light impulse. To avoid flicker from the decay of individual elements, therefore, each element is scanned once every 1/30 second. Actually, to create an impression of motion, only 15 individual scenes per second need be transmitted; however, at this low repetition rate—in fact, even at 30 blanking periods per second—flicker is apparent. The higher the brightness of the system, the more apparent flicker becomes for a given blanking period and repetition rate.

18. *Contrast, Brightness, and Gamma*

There are three classifications related to the image illumination of a televised scene. *Brightness* refers to the background illumination of the scene; *contrast,* to the light range between darkest black and brightest white; and *gamma,* to the log relation between change of image brightness to change of object brightness.

A conception of contrast and average brightness can be obtained by reference to ordinary snapshots. For example, an object snapped in front of a bright, sunny landscape may have the same contrast range as the same object taken in front of a shady scene. However, the average brightness of the sunny scene is much higher. Usually, the relative-brightness range or contrast is also higher. However, so far as television image brightness (scene on fluorescent screen) is concerned, the peak brightness and black-level limitations often confine the contrast to the same range for dark and bright scenes, and only the background brightness changes.

A scene televised under banks of concentrated light in the studio will have a higher average brightness than an event televised outdoors on a cloudy day, and yet it is possible that the darkest and brightest points of both scenes will have the same intensity. Contrast of a television system, therefore, represents the range of variations in light and dark as the picture-tube beam reproduces the original scene. It is possible to control relative and average brightness over a limited range at the receiver. Average or background brightness is controlled by varying the d-c bias between control grid and cathode of the picture tube (brightness control). Thus, the average strength of the beam is regulated and the average illumination of the screen can be set at a comfortable level. The contrast, or relative brightness, can be adjusted by varying the peak-to-peak amplitude of the picture signal applied to the picture-tube control grid (contrast control). It must be pointed out that the electronic methods of varying brightness, while necessary, are not a cure-all for weak signals and poorly designed receivers. For, with electronic control we also amplify the noises the same amount as the signal. Thus, if the original signal is weak and noisy, the image on the fluorescent screen will appear washed out and noisy.

Inasmuch as it is not possible to reproduce the range of brightness of the original scene, a method of increasing apparent contrast is achieved by having a nonlinear light-transfer characteristic. That is, the image brightness is made to vary logarithmically. Using this method, the brightest bright and darkest dark still have the same absolute values, but the brightness increments at the brilliant end of the range are greater per given change in object brightness than the increments at the dark end of the light scale. This apparent improvement in contrast is illusionary because it is a characteristic of the human eye that an increment in light intensity is more noticeable to the eye in dark parts than in bright parts (logarithmic eye sensation). Thus, in transmission the high lights are overemphasized and the shadows are underemphasized, improving apparent contrast.

Gamma is a measure of contrast and is the ratio of *image log brightness* to *object log brightness*. Since this is a log ratio, it means it is actually a ratio of eye sensation of the image to eye sensation of the object. For example, if the gamma is one, it does not mean that the object brightness and image brightness are directly proportional (as just mentioned, it is logarithmic), but it does mean that a logarithmic rise in object brightness causes a similar logarithmic rise in image brightness. If the gamma were two, the light increments of the screen presentation are twice as great as object light increments over the linear portions of the light transfer between brightest bright and darkest dark. Generally, the response of the pickup equipment has a gamma less than unity, and unity or higher over-all gamma is obtained by the fluorescent screen, which has a gamma higher than unity. Thus, the brightness changes are overemphasized a limited amount to make image change more realistic. At the same time the emphasis on high lights and compression of lows improves apparent effect of noise variations because the ratio of signal to noise is decreased in the highs where changes are less apparent. A scene with good contrast means one which contains light distribution from very dark to an extremely intense white. Unfortunately, the modern television system is not capable of conveying the wide range of light gradations or contrast of a sunlit day when the bright-to-dark ratio may be as high as 10,000:1.

Peak performance of the modern television system is approximately 100:1, or less, with 1000:1 contrast ratio as an ultimate goal.

In terms of photography, a snapshot with a wide range between whites and blacks has good contrast, but an overexposed or underexposed snapshot, in which the distribution is concentrated at one end or the other of the light register, has poor contrast.

QUESTIONS

1. Describe in your own words the basic television system.
2. What is meant when we say the television transmitter does not transmit picture information continuously?
3. What is picture signal and where is it utilized at the receiver?
4. What is sync and where is it utilized at the receiver?
5. Are all elements of a televised scene transmitted simultaneously? Explain.
6. Why is such an extended video frequency range necessary in the transmission of a picture?
7. Why is it necessary to have a wide television channel?
8. Describe the standard television channel.
9. What is the megacycle separation between picture and sound carriers?
10. Does a sine wave have a linear change in voltage?
11. What characteristics of a pulse do we utilize in television?
12. What characteristics of a sawtooth do we utilize in television?
13. What is meant by a fixed-rate or linear change in voltage?
14. Why is it possible to distort a pulse or a sawtooth with a simple resistor-capacitor combination?
15. What is a single-polarity signal?
16. Describe signal on control grid of picture tube.
17. Describe the television image and the limitations of its horizontal and vertical resolution.
18. How is flicker eliminated, and why is it the eye sees a continuous presentation although the picture is frequently interrupted?
19. Differentiate between picture contrast and background brightness.
20. What causes each horizontal line to begin below the previous one?
21. What is the scanning raster?
22. How are retrace lines made invisible in the television system?
23. What are active and inactive lines?
24. Describe the interlaced scanning system.
25. Why are the sync pulses not seen on the fluorescent screen?

Chapter 2

COMPOSITE TELEVISION SIGNAL

19. *Components of the Television Signal*

Three types of information are transmitted on the picture carrier: (1) *picture* or *video signal,* a progressive series of impulses which convey the light distribution of the televised scene; (2) *sync* (horizontal and vertical), a series of rectangular pulses which keep the pickup and picture-tube beams locked in synchronism and prevent displacement of the pattern on the picture-tube screen; and (3) *blanking* (horizontal and vertical), a series of rectangular pulses of longer duration than the sync pulses which blank or black out the fluorescent screen during all retrace intervals, making the retrace lines invisible. A fourth group of signals called *equalizing pulses,* of shorter duration than sync pulses, are transmitted to insure uniform spacing of the interlaced scanning lines and to prevent loss of synchronism of the horizontal circuits during the vertical retrace intervals between fields and frames.

To be certain that all models of television home receivers function properly on all commercial telecast stations, the Federal Communications Commission has set down definite standards on the content, construction, and sequence of the television signals. Much useful information can be obtained by studying the composite signal shown in Fig. 23. This composite waveform chart is drawn to scale vertically but not horizontally. Actually, picture is transmitted over a much longer period of time for each line in comparison to the horizontal blanking interval. The chart is also numerically keyed to portions of the text material of this chapter.

20. *Picture Signal*

The picture signal is transmitted during the active horizontal scans (Fig. 23, points 1–6–7) and is located between horizontal blanking periods of the composite signal. The picture signal is transmitted line after line and is interrupted for each horizontal blanking or retrace interval and again for each vertical blanking interval. The picture signal varies in amplitude between the blanking

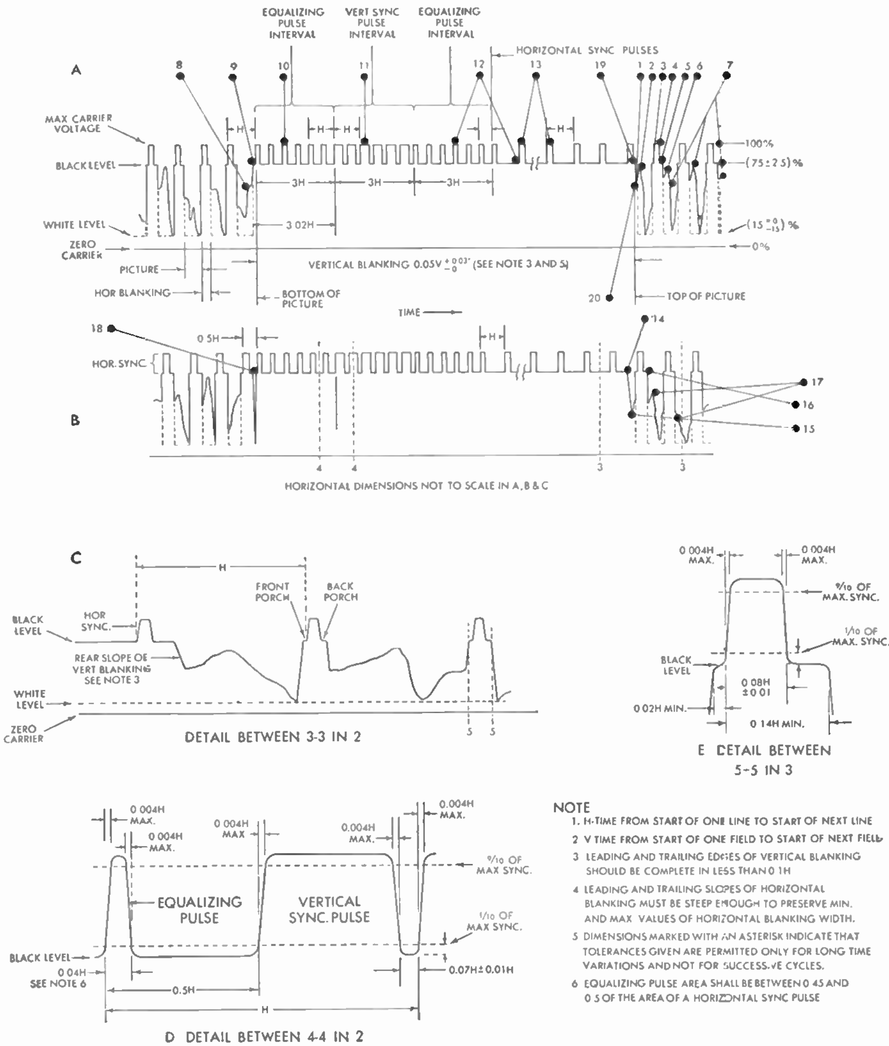


FIG. 23 Composite Television Signal

level, which represents black, and the 15-per cent amplitude level, which represents peak white.

Signal variations for two lines of picture are shown in Fig. 24, which is also drawn to approximate scale horizontally. It is apparent that picture is transmitted for a much longer time than is required for the beam retrace, during which time picture is removed while blanking and sync are transmitted. The time required for the beam to cover one horizontal scan (trace and retrace) is spoken of as the *horizontal-line interval.*

Inasmuch as 525 lines are transmitted per picture or frame (we must remember that only approximately 450 to 500 of these are active; the remainder occur during the vertical retrace and contain no picture information) and there are 30 complete pictures per second, the horizontal or line rate is 15,750 per second and a single horizontal-line interval occurs in 1/15,750 second or 63.5 microseconds.

In the two lines represented in Fig. 24, notice also that although the second line has the same relative variations in brightness, the average brightness of this line is somewhat darker because the picture variations as a whole are nearer the blanking or black level of the composite signal.

21. *Horizontal Blanking*

The horizontal blanking pulse occurs at the end of the scanned line (point 3, Fig. 23) and cuts off the picture-tube beam prior to the retrace of the beam to the left side. The blanking pulse keeps the beam blanked out all during the retrace time until the beginning of another line of picture.

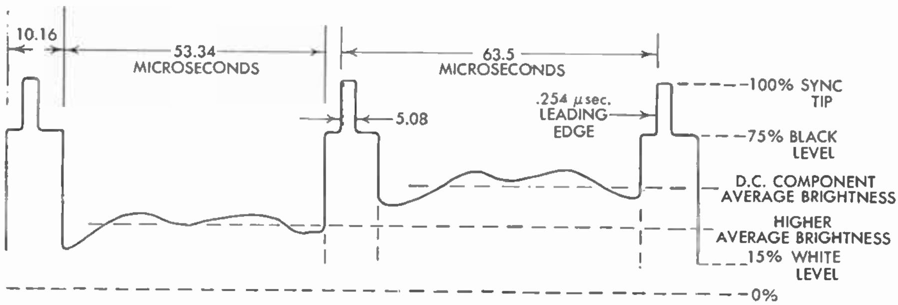


FIG. 24 Signal Distribution for Two-Line Interval, Showing Blanking and Sync Pulses

The duration of the horizontal blanking pulse (Fig. 24), according to the Federal Communications Commission, must be set at some duration between 10.16 and 11.43 microseconds, which represent the minimum and maximum tolerances. Thus, each horizontal line consists of some 50 microseconds of picture and some 10 microseconds of blanking, or an approximate trace-to-retrace ratio of 6:1. Inasmuch as the actual sawtooth retrace does not begin until after the blanking front porch (initiated by leading edge of sync pulse), and every precaution must be taken to make certain the beam does get back to the left side in the allotted time, the trace-to-retrace ratio of the sawtooth voltage or current which deflects beam is approximately 7:1 or higher.

22. *Horizontal Sync Pulse*

The horizontal sync pulses, of which there are nearly 500 each frame, keep the horizontal motion of the receiving-tube scanning beam in sync with the same motion of the pickup tube. This is accomplished by directly controlling frequency and phase of the sawtooth sweep voltage which deflects the picture-tube beam. Therefore, at any one instant, the two beams are striking their respective scanning rasters at the same relative positions horizontally. Thus, these horizontal sync pulses hold the pattern stationary horizontally (*i.e.*, prevent tearing out). Another group of pulses, called *vertical sync pulses,* hold the pattern stationary vertically. A typical horizontal sync pulse (Fig. 24) has a duration of approximately 5.08 to 5.6 microseconds and initiates the horizontal sawtooth retrace cycle by control of the receiver horizontal sawtooth generator. For a short interval before the transmission of the sync pulse, a blanking pulse is transmitted which immediately blacks out the screen to make the retrace invisible. Actually, during this front porch of the blanking pulse, the sawtooth voltage continues its trace, and it is not until the leading edge of the sync pulse arrives that the sawtooth begins to retrace. This horizontal blanking continues for an interval after the sync pulse, permitting sufficient time for the return to the left-hand side of the raster before the screen is unblanked.

It is important to consider at this point that the beam itself is not retraced because it is, in truth, cut off by the blanking pulse. Actually, what does occur is that the sawtooth retrace reshapes the electrostatic or magnetic field (which normally would retrace the beam if beam were present) to have it ready to move the beam left to right at the start of the next active scan. Thus at instant beam is again turned on it is at the left side.

The composite signal, as it appears on the control grid of the picture tube, is negative in polarity—that is, the higher the signal amplitude, the more negative the grid is driven and the darker the fluorescent screen becomes. After the completion of a horizontal line, the arrival of the blanking pulse drives the grid to the black-out or blanking level of the fluorescent screen. A short time later the still greater amplitude sync pulse arrives to initiate, by way of the receiver deflection system, the horizontal retrace of the sweep. At the conclusion of the sync pulse, the blanking continues to hold the screen blanked until the sweep has had sufficient time to retrace and is ready to begin a new scanning line.

23. *Vertical Blanking*

The vertical blanking pulse (Fig. 25), of which there are two each frame (two-field interlaced system), is a long-duration, continuous pulse. During this blanking interval the vertical sawtooth sweep is synchronized by vertical sync pulses and the vertical sweep is retraced.

The duration of the vertical blanking pulse is set between 833 and 1330 microseconds—Federal Communications Commission minimum and maximum tolerances. There are 30 complete pictures or frames per second and 60 fields. Consequently, the vertical field rate is 60 per second and the vertical field interval is 1/60 second, or 16,667 microseconds, of which 833 to 1330 microseconds represent vertical retrace time. During this time approximately 13 to 20 horizontal lines occur upon which no picture information appears for each field. This represents the inactive 26 to 40 lines per frame, which subtract from the total 525 lines to give us the number of active lines. It is apparent that the number of active lines increases with use of the shorter duration vertical blanking interval. However, in the case of film telecasting, the film frames are projected on the camera-tube mosaic only during the vertical blanking period, and it is advantageous to use a longer vertical blanking period to increase the average illumination of the mosaic.

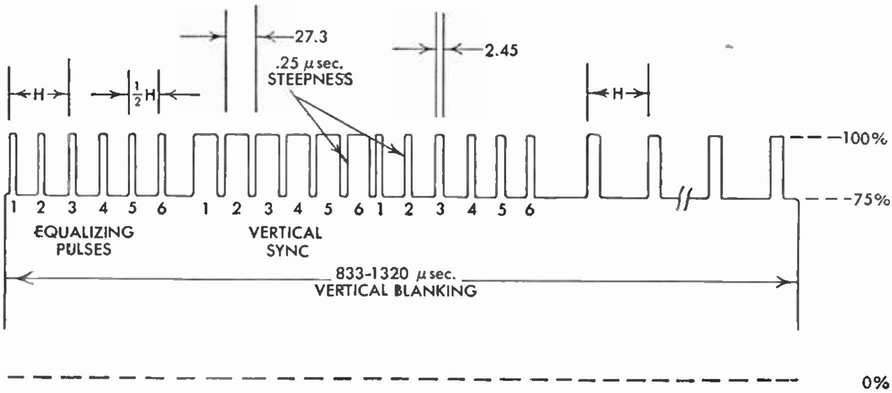


FIG. 25 Vertical Retrace Interval

24. *Vertical Sync and Equalizing Pulses*

The vertical sync pulse intervals, of which there are two each frame (interlaced scanning system), keep the vertical motion of the picture-tube scanning beam in sync with the vertical motion of the pickup-tube scanning beam. These vertical sync pulse intervals hold the pattern stationary vertically (*i.e.,* prevent flopping over). Thus, the combination of horizontal and vertical sync control holds the pattern absolutely rigid and the picture-tube beam always directed at the proper point on the scanning raster. A typical vertical sync pulse (Fig. 25) has a duration of 27.3 microseconds—actually, the vertical sync pulse interval consists of six 27.3-microsecond pulses. The brief interruptions of the vertical sync pulse interval, called a *slotted* or *serrated vertical,* prevent loss of horizontal synchronism during the vertical interval, still producing a vertical sync interval equivalent to 190 microseconds. Immediately before and after the actual vertical sync pulses, a series of six equalizing pulses (Fig. 25) are transmitted. These very short-duration pulses insure equidistant spacing between scanning lines, permit transmission of identical vertical intervals between frames and between fields, and prevent loss of horizontal synchronism during vertical retrace intervals. Here again, after the completion of the first or second fields (end of second field is also end of frame and therefore the vertical retrace at the end of a frame is called the *vertical retrace between frames,* and the vertical retrace between first and second fields is called the *vertical retrace between fields*), the arrival of a blanking pulse drives the picture-tube control grid to the blanking level. A short time later the equalizing pulses arrive and then the vertical sync pulses. At the conclusion of the vertical sync and equalizing pulses, the blanking pulse continues to hold the fluorescent screen blanked until the sweep has had sufficient time to return to the top of the raster and is ready to begin a new field.

It is important to note that the horizontal is synchronized not only on the leading edges of the horizontal sync pulses but also on the leading edges of the vertical and equalizing sync pulses. Therefore, alternate leading edges of the vertical sync and equalizing pulses are also 63.5 microseconds apart. If these leading edges were not present, the horizontal would operate without synchronization during the longer duration vertical sync interval, producing interlace instability. The horizontal synchronizes on the odd-numbered vertical sync and equalizing pulses between fields, and on the even-numbered between frames.

25. *Signal Standards*

Some of the more important facts and figures of the composite signal are listed here:

- 6-megacycle channel
- 4:3 aspect ratio
- 525 lines—500 active and 25 inactive lines interlaced to 485 active and 40 inactive interlaced
- scanning—30 frames per second and 60 fields per second
- 15,750 lines per 30 frames (30×525)—line rate
- one frame—33,334 microseconds
- one field—16,667 microseconds
- one horizontal sweep cycle—63.5 microseconds
- horizontal retrace or horizontal blanking interval—10.16 microseconds to 11.4 microseconds
- horizontal trace—53.34 microseconds
- horizontal sync pulse—5.08 to 5.68 microseconds
- vertical blanking interval—833 to 1,330 microseconds per field; 1,666 to 2,640 microseconds per frame
- vertical sync pulse interval—190.5 microseconds
- vertical sync pulse—27.3 microseconds
- equalizing pulse—2.54 microseconds
- spacing between vertical pulses—4.44 microseconds
- spacing between equalizing pulses—29.2 microseconds
- spacing between leading edges vertical pulses—31.75 microseconds

spacing between leading edges equalizing pulses—31.75 microseconds

leading edges horizontal, vertical, and equalizing sync pulses—0.254 microsecond

leading edges vertical blanking—6.35 microseconds

The time durations of other components of the composite signal can be calculated in terms of H or V in Fig. 23.

H—Time from start of one line to start of next, or 63.5 microseconds

V—Time from start of one field to start of next, or 16,667 microseconds

26. *Signal Sequence*

Inasmuch as it is difficult mentally to assemble and retain all the details concerning the composite signal, the following sequential analysis of the composite signal will serve as a ready reference. The signal sequence, keyed to the numerals of Fig. 23, will be given from the start of one frame to the start of the next.

1. Start of picture signal at top left-hand corner of scanning raster.
2. Picture or video signal for one active scanning line. Light distribution along first line of televised scene.
3. Horizontal blanking pulse at completion of first line. Blanks out screen to make retrace invisible.
4. Horizontal sync pulse which initiates the sweep retrace—controls horizontal sweep circuits.
5. Horizontal blanking continues, allowing sufficient time for the sweep to retrace.
6. Second line of video signal begins—actually, this is the number three scanning line because of interlaced scanning system.
7. All the odd-numbered scanning lines are covered—consecutive odd-numbered lines interrupted by the horizontal blanking intervals.
8. Odd-numbered lines scanning completed at bottom right-hand corner of scanning raster. (Last active lines of field on left side of waveform A, Fig. 23.) Only a few of the horizontal cycles are shown on the drawing.
9. Vertical blanking pulse at completion of first field. Blanks out screen to make vertical retrace between fields invisible.
10. A series of six equalizing pulses.
11. Vertical sync pulse interval which initiates the sweep retrace—controls vertical sweep circuits.
12. Series of six more equalizing pulses and continuation of blanking to allow sufficient time for sweep retrace.

13. Horizontal pulses continue throughout vertical blanking to prevent loss of horizontal control during the vertical retrace.

14. Retrace returns sweep from lower right-hand corner of raster to top center (drawing B).

15. Vertical blanking continues, ending at the top at a half-line point (drawing B).

16. First horizontal blanking interval of second field.

17. Even-numbered lines are scanned until picture information for the frame is completed at lower center of the scanning raster.

18. Vertical retrace between frames begins.

19. Sweep is returned from bottom center to top left of scanning raster, completing one frame.

20. Start of first scanning line of next frame.

In the above sequence it was assumed that the vertical trace interval occurred for a period equivalent to time of an integral number of lines and the retrace interval corresponded to time of a finite number of lines plus a half-line. If the trace time contained the odd half-line, the scanning of the raster would be somewhat different. Again, the scan might start at top left (point 1) for odd-line scanning, and retrace for first field would begin from bottom center and return to top center before start of second field. This second field would end at bottom right from which it would retrace to top left. On our standard waveform, scan would start at point 1 and field blanking at point 18. Second field scan begins at point 15 and frame blanking at point 8. New frame unblanking again occurs at point 19.

Whether or not the retrace is an integral line length or contains an odd line depends on what portion of the vertical sync pulse interval initiates the vertical retrace, as will be understood more definitely in your study of Chap. 7. In fact, it is very possible that the retrace can be initiated from most any point along the bottom of the raster, start of field and frame retrace always being separated by a half-line along this last horizontal scan.

27. *Sweep Timing*

To review again exactly what occurs at the picture tube, refer to the drawing of Fig. 26, which shows us the active and inactive portions of the sweep cycle. During the active horizontal scan, the beam is unblanked and is deflected left to right across the screen by the slow linear rise of the horizontal sawtooth applied to the deflection plates or coils. At the end of the interval during which picture information is present on the signal, a horizontal blanking pulse appears on the grid of the picture tube, which cuts off the beam. However, if the beam were not cut off, it would continue to move to the right a short distance (shaded area on right side of drawing) because the trace portion of the horizontal sawtooth continues on for a short interval after the beam has been blanked. The horizontal sync pulse arrives at the horizontal sweep system of the receiver approximately 1.2 microseconds after the beam has been blanked (time of front porch). This sync pulse initiates the start of the snap-back of the horizontal sawtooth. The fast retrace of the sawtooth reshapes the deflection field, and when the beam is unblanked at conclusion of the horizontal blanking pulse it will begin a new scan on the left side.

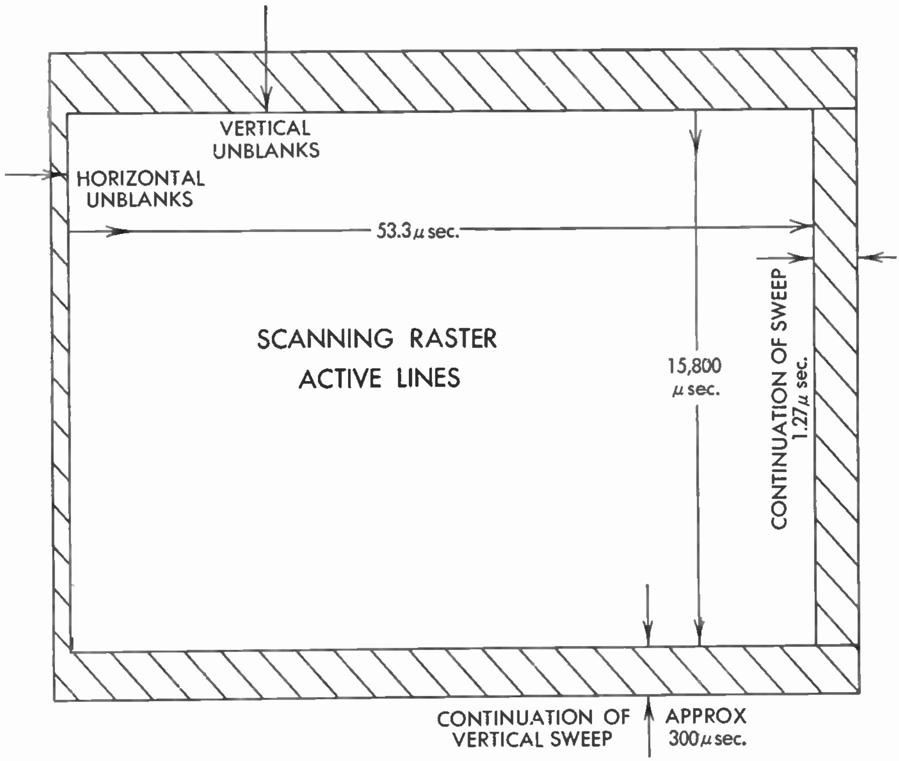


Fig. 26 Sweep Timing

The shaded area at the left represents a very short interval during which the beam remains blanked during the rise of the sawtooth, depending on the sharpness or speed of the horizontal sawtooth retrace.

Likewise, the vertical sawtooth continues to fall after the beam has been blanked by the vertical blanking pulse because the vertical retrace is not initiated in the vertical deflection system of the receiver until the fourth or fifth vertical sync pulse. Some of the inactive lines represent shaded area above the raster, depending upon speed of vertical retrace.

28. *Television Glossary*

A glossary of television terms is presented in the following pages. It will review many of the television terms introduced to you and add a few more. As you study succeeding chapters, it will serve as a reference glossary and give you a quick interpretation of television technical terms. The first group of terms is a collection of Federal Communications Commission definitions which apply to commercial television broadcast stations.

A. GENERAL

Television broadcast station—a station in the television broadcast band transmitting simultaneous visual and aural signals intended to be received by the general public.

Television broadcast band—those frequencies in the band extending from 44 to 216 megacycles which are assignable to television broadcast stations. These frequencies are 54 to 72 megacycles (channels 2 through 4), 76 to 88 megacycles (channels 5 and 6), and 174 to 216 megacycles (channels 7 through 13).

Television channel—a band of frequencies 6 megacycles wide in the television broadcast band and designated either by number or by the extreme lower and upper frequencies.

Television transmission standards—the standards which determine the characteristics of the television signal as radiated by a television broadcast station.

Standard television signal—a signal which conforms with the television transmission standards.

Television transmitter—the radio transmitter or transmitters for the transmission of both visual and aural signals.

Antenna field gain—the ratio of the effective free-space field intensity produced at 1 mile in the horizontal plane, expressed in millivolts per meter for 1-kilowatt antenna input power to 137.6 millivolts per meter.

Free space field intensity—the field intensity that would exist at a point in the absence of waves reflected from the earth or other reflecting objects.

Polarization—the direction of the electric vector as radiated from the transmitting antenna.

Effective radiated power—the product of the antenna power (transmitter-output power less transmission-line loss) times (1) the antenna power gain, or (2) the antenna field gain squared.

Service area—the service resulting from an assigned effective radiated power and antenna height above average terrain.

Antenna height above terrain—the average of the antenna heights above the terrain from 2 to 10 miles from the antenna. (In general, a different antenna height will be determined by each direction from the antenna.)

The average of these various heights is considered the antenna height above average terrain.

B. VISUAL TRANSMITTER

Visual transmitter—the radio equipment for the transmission of the visual signal only.

Amplitude modulation (AM)—a system of modulation in which the envelope of the transmitter wave contains a component similar to the waveform of the signal to be transmitted.

Aspect ratio—the numerical ratio of the frame width to frame height, as transmitted.

Black level—the amplitude of the modulating signal corresponding to the scanning of a black area in the transmitted picture.

Color transmission—the transmission of television signals which can be reproduced with different color values.

Field frequency—the number of times per second the frame area is fractionally scanned in the interlaced scanning.

Frame—one complete picture.

Frame frequency—the number of times per second the picture area is completely scanned.

Interlaced scanning—a scanning process in which successively scanned lines are spaced an integral number of line widths and in which the adjacent lines are scanned during successive cycles of the field frequency scanning.

Monochrome transmission—the transmission of television signals which can be reproduced in gradations of a single color only.

Negative transmission—transmission whereby a decrease in initial light intensity causes an increase in the transmitted power.

Positive transmission—transmission whereby an increase in initial light intensity causes an increase in the transmitted power.

Progressive scanning—a scanning process in which scanning lines trace one dimension substantially parallel to a side of the frame and in which successively traced lines are adjacent.

Scanning—the process of analyzing successively, according to a predetermined method, the light value of picture elements constituting the total picture area.

Scanning line—single, continuous, narrow strip containing high lights, shadows, and half tones, which is determined by the process of scanning.

Synchronization—the maintaining of one operation in step with another.

Vestigial sideband transmission—a system of transmission wherein one of the generated sidebands is partially attenuated at the transmitter and radiated only in part.

Visual frequency—the frequency of the signal resulting from television scanning.

Visual transmitter power—the peak power output when transmitting a standard television signal.

Peak power—the power over a radio frequency cycle corresponding in amplitude to synchronizing peaks.

C. AURAL TRANSMITTER

Aural transmitter—the radio equipment for the transmission of the aural signal only.

Center frequency—the average frequency of the emitted wave when modulated by a sinusoidal signal; the frequency of the emitted wave without modulation.

Frequency modulation—a system of modulation where the instantaneous radio frequency varies in proportion to the instantaneous amplitude of the modulating signal (amplitude of modulating signal to be measured after pre-emphasis, if used) and the instantaneous radio frequency is independent of the frequency of the modulating signal.

Frequency swing—the instantaneous departure of the frequency of the emitted wave from the center frequency resulting from modulation.

Percentage modulation—as applied to frequency modulation, the ratio of the actual frequency swing to the frequency swing defined as *100-per cent modulation* expressed in percentage. For the aural transmitter of television broadcast stations, a frequency swing of 25 kilocycles is defined as 100-per cent modulation.

The next group, arranged in alphabetical order, will define and expand, for added clarity, the many television technical terms.

Aspect ratio—ratio of length to width of the scanning raster. Present standard is 4:3.

Average brightness—average illumination of the elements of one line, frame, or televised scene.

Average-brightness level—a d-c component of average brightness as represented by a fixed voltage level. This average level shifts as the average brightness of the scene changes.

Back porch—that portion of the flat top of the blanking pulse between trailing edge of the sync pulse and its own trailing edge.

Black level—voltage level at which picture tube blacks out. In a properly adjusted system, black and blanking levels are coincident.

Blacker-than-black region—super-sync voltage range between blanking and sync tip levels.

Blanking level—a constant d-c level represented by the voltage level of the flat tops of the blanking pulses. When the television signal reaches this level, the picture-tube screen should black out.

Blanking pedestal—flat-top portion of blanking pulses upon which the sync pulse is mounted.

Blanking pulse—a pulse used to blank or black out retrace lines at the receiver. Vertical pulses are long duration; horizontal, shorter.

Camera dolly—movable carriage upon which television camera is mounted to follow changing scene conveniently.

Camera pre-amp—video amplifier which increases amplitude of weak signal from pickup tube.

Cathode follower—a circuit which acts as an impedance transformer with excellent frequency response.

Composite synchronizing waveform—that portion of the television signal which has to do with the synchronization of the television system.

Composite television signal—the entire signal, consisting of video signal and synchronizing waveforms.

Constant impedance network—a network which presents a constant impedance and, therefore, no frequency discrimination over a wide band of frequencies.

Contrast—range of light gradations between the darkest and brightest portions of a scene.

Counter circuit—a frequency dividing circuit; for example, with a 5:1 counter, a 300-cycle input signal will produce a 60-cycle repetition rate output signal.

D-c amplifier—an amplifier in which the d-c as well as the a-c component of a signal is transferred between stages.

D-c restoration—a means of re-establishing a d-c level after it has been lost by transfer of signal by a capacitor.

Delay circuit—a means of delaying one pulse with respect to another.

Dipole—a resonant antenna which is an electrical half wave in length.

Electron gun—elements in a cathode-ray tube which form a highly concentrated, thin stream of electrons.

Flicker—annoying light fluctuation caused by not cutting illumination off and on rapidly enough or cutting it off for too long a period.

Front porch—that portion of the flat top of the blanking pulse between its leading edge and the leading edge of the sync pulse.

Iconoscope—a television pickup tube. It converts light energy to an equivalent electrical representation.

Image dissector—a television pickup tube.

Image orthicon—a television pickup tube.

Instantaneous brightness—brightness of a single element of the picture.

Inter-sync separation—a method of segregating the basic sync components.

Keying pulses—a series of pulses which control the insertion of various pulse types into the composite synchronizing waveform.

Kinescope—a television picture tube.

Leading edge—sharp change in voltage at beginning of a pulse.

Linearity—proper proportioning of image at all points on the fluorescent screen.

Line and combining amplifiers—amplifiers in which sync and blanking are combined with picture to form composite television signal.

Loading—reduction of circuit Q with shunt resistance to obtain a broader bandpass characteristic

Mike boom—movable microphone rod which can be moved about over the heads of the artists and out of sight to pick up speech and music.

Mosaic—photosensitive surface of pickup tube, which is scanned by an electron beam that releases picture signal from it.

Narrowing circuit—a method of reducing the width of a pulse.

Pairing—improper spacing of scanning lines causing loss of resolution. Lines are paired instead of equidistantly spaced.

Peaking coil—a small inductor used to improve the high-frequency response of a video amplifier.

Persistence of vision—ability of the human eye to retain an impression of illumination after the light source has been removed.

Picture tube—a television receiving tube. It converts electrical signal into a pictorial reproduction.

Pulse—a rectangular squared wave used as a synchronizing waveform.

Reproduced image—reproduction of televised scene as it appears on fluorescent screen of picture tube.

Resolution—clarity and definition of a picture as set by the number of picture elements.

Sawtooth—comparable to an asymmetrical sawblade. It is a gradual, linear rise of voltage and faster decay. It is used in circuits which deflect the scanning beams.

Sideband filter—filter which absorbs portion of a sideband not transmitted.

Scanning raster—that portion of screen swept by scanning beam. It represents the illuminated area on the face of the picture tube.

Serrated pulse—a pulse used to synchronize scanning. Instead of a long, continuous pulse, however, it is interrupted for short intervals.

Shading regulation—an electronic means of removing spurious dark spots which appear in the image.

Sweep circuits—circuits which generate sweep voltages that cause electron beam to scan the pickup and picture-tube rasters.

Sync generator—circuits which generate the composite sync and blanking waveform.

Sync pulse—a pulse used in synchronizing receiver scanning with transmitter scanning.

Sync separation—a method of obtaining synchronizing pulses from the composite signal at the receiver.

Sync tip level—the very top, flat portion of the sync pulse. Sync tip level represents a fixed d-c level.

Televised image—televised scene as it is focused on photosensitive surface of pickup tube by a lens system.

Televised object—scene which is being televised.

Trailing edge—sharp change in voltage at ending of a pulse.

Vestigial sideband transmission—partial suppression of one sideband to reduce channel width required to transmit a picture of a given definition.

Video amplifier—an amplifier which amplifies video frequencies. It is to be compared with an audio amplifier which amplifies audio frequencies.

Video frequencies—those frequencies which convey picture information. Video frequencies are comparable to audio frequencies except that they entail a much broader band of frequencies.

Video signal—that portion of a television signal which consists of only the actual picture information.

White level—voltage which represents peak illumination.

Wide-band amplifier—an amplifier which amplifies a wide band of frequencies.

QUESTIONS

1. Name all components of the composite television signal.
2. What is function of each component of the composite signal?
3. What is picture-to-sync ratio?
4. At what percentage of peak amplitude is the blanking level?
5. At what percentage is the black level? Maximum white?
6. During what percentage of total frame time is the beam blanked?
7. In how many microseconds must the horizontal sync pulse rise from 1/10 to 9/10 peak amplitude? In how many microseconds must the vertical sync and equalizing pulses do the same?
8. During what percentage of an active horizontal-line interval is picture information transmitted?
9. What is microsecond duration between leading edges of adjacent vertical sync pulses?
10. What is repetition rate of vertical sync pulses?
11. What is repetition rate of individual vertical sync- and equalizing-pulse groups?
12. What is repetition rate of horizontal sync pulses?
13. How many lines are transmitted per field?
14. How many line intervals occur during the period the six equalizing, six vertical sync, six equalizing pulses are transmitted?
15. What is the duration of a single equalizing pulse?
16. In the composite waveform of Fig. 23, how many microseconds does V represent?
17. In the same drawing, how many microseconds does H represent?
18. Go through, from memory, the signal sequence from beginning to end of frame.
19. What is meant by negative transmission?
20. Why must a slotted vertical pulse be used?

Chapter 3

GENERAL OPERATION OF THE TELEVISION SYSTEM

29. *Television as an Entirety*

In this chapter it is your task to follow the feeble picture signal generated by the camera tube through its many processes until it finally reproduces the camera image on the fluorescent screen of the picture tube. In following its path, the characteristics of the many circuits it passes are discussed. Consequently, an understanding of at least the purpose of the major components of the television system will be gained. Later, in the study of specific circuit detail, this understanding of the system as an entirety will be valuable. It is much easier to grasp circuit detail if we know what is to be expected of the specific circuit under discussion and how it fits into the entirety.

30. *Camera Tubes*

The camera tube can be compared to the microphone of a conventional broadcast station with the exception that the output of a camera tube is many, many times weaker than the output of even the most insensitive microphone. The two most widely used camera tubes today are the iconoscope and image orthicon. The iconoscope, because of its excellent resolution and need for intense lighting, is ideal for film projection and some studio application. For remote telecasts, the image orthicon receives the nod because of its compactness and ability to cope with an extremely wide light range.

ICONOSCOPE

The iconoscope (Fig. 27) consists of electron gun, deflection system, and light-to-signal conversion elements. The electron gun has the same essential elements as any cathode-ray-tube gun but is designed to concentrate the beam to a much finer point on the photosensitive mosaic. Since it is not necessary to strike the mosaic with a high-velocity, large-current beam, as in the picture tube, the second anode potential is in the vicinity of only 1,000 volts. The deflection system for the iconoscope uses deflection coils. The sawtooth currents, which pass through these coils, are of proper frequencies and amplitudes to cause the electron beam to scan the photosensitive mosaic, upon which the image has been focused, in accordance with the standard interlaced scanning pattern. It is important to note that the lens inverts the image on the mosaic, and scanning of necessity starts at bottom.

The light-to-signal conversion elements consist of the light-sensitive mosaic, the signal plate, the thin mica strip which separates mosaic and signal plate, and a collector ring. Light distribution of the scene to be televised is gathered by an external lens system and focused on the photosensitive mosaic. The

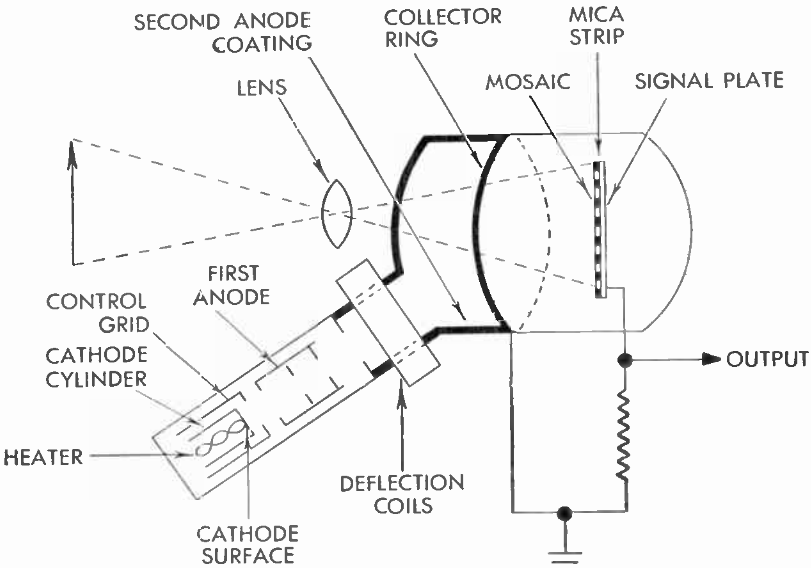


FIG. 27 Iconoscope

mosaic is then scanned line after line by the iconoscope beam releasing electrons in accordance with the photoelectric charges. These electrons reach the collector ring and constitute the picture-signal output of the iconoscope. Return path for the picture signal is through the capacity existing between mosaic and signal plate—thin mica sheet serving as the dielectric.

The mosaic is formed by depositing sensitized silver globules on a thin mica sheet. These globules, though deposited very close together, are insulated from each other by the mica. Consequently, there is no serious exchange of charges between globules, each globule maintaining its charge in accordance with the intensity of the light striking it. Each globule is less than 0.001 inch in diameter.

Since the scanning beam has a spot size or diameter of approximately 0.007 inch, a large number of globules are being scanned at any one instant. The area covered on the mosaic at any one instant by the scanning beam is called an *element*. The total number of elements is a measure of the detail transmitted. At present, the iconoscope has capabilities of scanning more elements on the mosaic than the bandpass of the system or the spot size of the picture-tube beam can adequately convey. The total number of picture elements for the modern television system is better than 200,000.

The other side of the mica strip is covered with a conducting coating of colloidal graphite, which forms the signal plate. Thus, each silver globule is effectively capacitively coupled to the external circuit, with the mica serving as the dielectric and the signal plate as the other capacitor plate. Since each globule functions as a photocathode or tiny photoelectric element, a charge appears on each effective capacitor in accordance with the light focused on it. It is this charge that determines the amplitude of the signal transferred to the external circuit by way of the collector ring and signal plate at the time the globule is scanned by the electron beam from the iconoscope gun. An unusual feature is that the amplitude of the signal output is higher for a dark spot than a bright spot.

The first point to realize in studying the operation of the iconoscope is that the electron beam, as it arrives from the electron gun, is not a part of the output circuit. Neither does the electron stream serve as a return path for the picture signal. The sole purpose of the electron gun and deflection system, therefore, is to supply a concentrated source of electrons which scan the mosaic in accordance with the standard interlaced scanning pattern.

The actual signal circuit consists of the mosaic, signal plate, and collector ring. This circuit is closed for each picture element at the time the beam strikes it, knocking off secondary electrons which flow to the collector ring. Thus, a circuit is closed from mosaic to collector and through the external circuit back to signal plate and mosaic. Any variation in the current through this circuit appears across the output load resistor and on the grid of the first amplifier. When there is no light focused on the mosaic, there is essentially no photoelectron emission. However, secondary current still exists when the mosaic is scanned because the impact of the beam knocks off secondary electrons, which flow toward the positive collector ring. In fact, when there are no photoelectron charges, the greatest number of secondary electrons are knocked off and the greater becomes the absolute value of the current.

Now, at the time a scene is focused on the mosaic, each light-sensitive globule emits photoelectrons in accordance with the light striking it. The more intense the light, the greater the number of electrons emitted. These electrons are displaced from the element, causing the element to assume a positive charge. The photo-emitted electrons themselves are not emitted with sufficient velocity to get over to the collector. However, when the electron beam strikes this element, the number of electrons it knocks off is reduced in proportion to the positive charge assumed by the element, which in turn is proportional to photoelectron displacement and light intensity. This reduction in secondary emission and, consequently, secondary current flow to collector is understandable when we consider that the more positive any mosaic element becomes the less attraction the collector has for the electrons remaining on the element.

Inasmuch as the charges on the globules differ in proportion to the amount of light focused on them, there is a similar but inverse variation in the number of secondary electrons knocked off and in the amplitude of the output current. Thus, as the beam scans across these dissimilar charges, the output current varies inversely, producing an alternating picture signal across the output load, which is a relative electric variation of the light distribution on the mosaic.

A disadvantage of the iconoscope is the generation of a spurious dark-spot signal. Actually, less than 25 per cent of the electrons removed from an element reach the collector. The remainder redistribute themselves indiscriminately over the surface of the mosaic. When an area that has been blanketed with secondary electrons is scanned, a dark-spot signal appears in the output. When an iconoscope camera is used, it is necessary to use a special shading generator to generate an opposing signal, which cancels the spurious dark-spot signal.

IMAGE ORTHICON

The image orthicon has exceptional sensitivity and permits televising a scene in very low illumination. The resolution of the image orthicon is somewhat inferior to that of an iconoscope, although a special studio-type image orthicon has improved resolution at a sacrifice in sensitivity. Sensitivity of the image orthicon solves many of the major illumination difficulties in television, and its versatility permits coverage of various types of programs having widely separated illumination characteristics. Its outstanding advantages are:

1. Ability to cover all scenes of visual interest, particularly those under low-lighting conditions.
2. Improved sensitivity, permitting greater depth of field and inclusion of background that might otherwise be blurred.
3. Improved stability, which protects images from interference due to exploding photoflash bulbs and other bursts of brilliant light.
4. Smaller size of tube, facilitating use of small telephoto lens.
5. Type of design, which permits construction of portable lightweight camera equipment.
6. Improved gain control system to permit unvarying transmission, despite wide fluctuations of light and shadow.

The image orthicon has three main sections:

1. Electron image section to amplify the photoelectric current.
2. A low-velocity type of scanning section to scan the target in accordance with the standard interlaced pattern. No spurious dark-spot signal is generated.
3. An electron multiplier section to increase in amplitude the relatively weak picture signal.

A functional drawing of the image orthicon (Fig. 28) divides the tube into its three major sections. Light from the televised scene is picked up by an optical lens system and focused on the photosensitive surface immediately behind the face of the tube. The surface elements emit electrons in proportion to the strength of the light striking them. These electrons, accelerated by a positive voltage and held on a parallel course by an electromagnetic field, flow from the back of the photosensitive surface to a target. Secondary emission from the target, caused by the impact of the electrons, leaves on the target a pattern of varying positive charges (charge on each element dependent on the number of electrons striking it), which corresponds to the pattern of light from the televised scene.

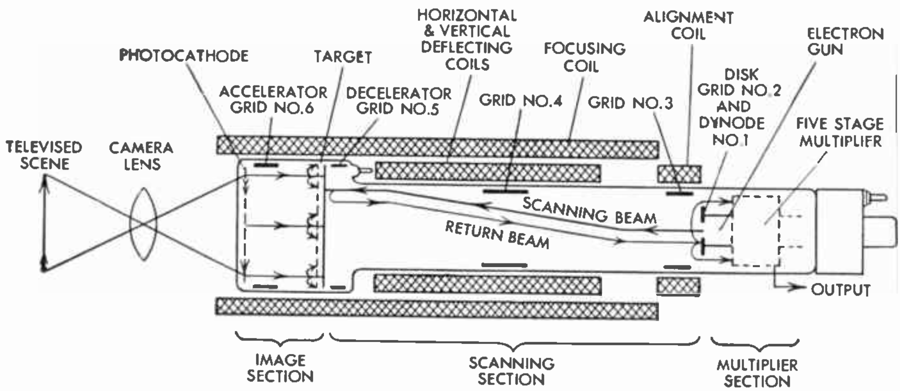


FIG. 28 Image Orthicon

The back of the target is scanned by a beam of electrons (standard interlaced system for commercial telecast application) generated by an electron gun in the base of the tube. Electrons in this beam are slowed down to have them stop just short of the target and return to the base of the tube except when they approach an element which has a positive charge on it. When this occurs, the beam will deposit enough electrons on the back of the target to neutralize the positive charge, after which it will again fall short of the target and turn back toward the gun. It is apparent, therefore, that the beam is modulated, as it scans, by the varying positive charges on the target which are representative of the light pattern of the televised scene.

The returning beam, with picture information imposed upon it by the varying losses of electrons left behind on the target, is directed toward the first of a series of dynodes (electron multipliers) near the base of the tube. This multiplying process continues, with the strength of the signal increasing at each dynode, until it reaches the signal anode and is conveyed to an external circuit.

31. *Camera Pre-Amplifier*

The camera pre-amplifier is a part of the camera and as such is mounted very near the camera tube. The camera unit is somewhat similar to a motion

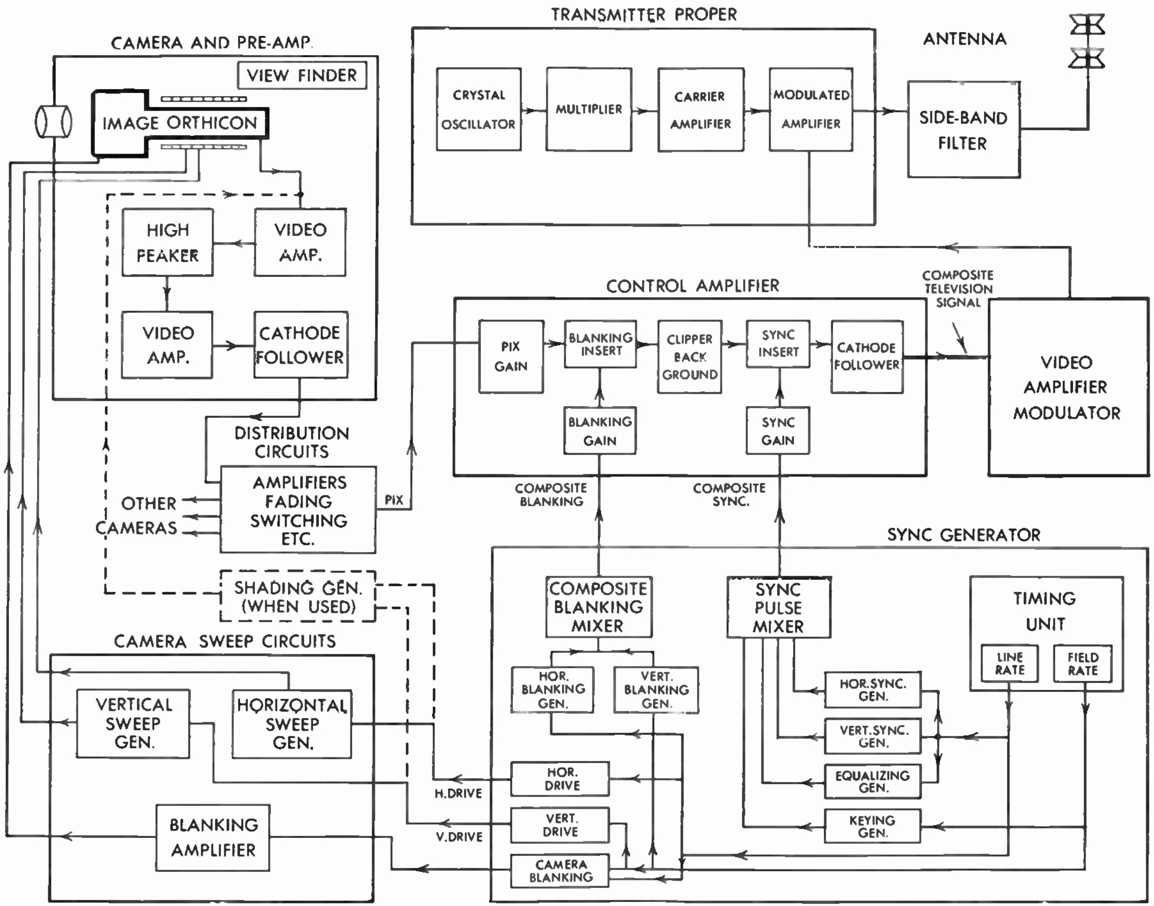


FIG. 29 Functional Block Diagram of Television Transmitter

picture camera, and the studio version is mounted on a dolly for ease in following a scene. The remote version of the television camera is mounted on a tripod. The camera pre-amp is a part of the camera and must be positioned near the camera tube as it is not possible to convey the feeble output of the camera tube any distance because of attenuation and noise pickup.

This pre-amplifier must amplify the extremely weak output of the pickup tube to a level (approximately ¼ volt peak to peak) satisfactory for transfer of the signal through a coaxial cable to the distribution amplifiers. At the same time, it must keep this very feeble signal above the tube and circuit noises and pass uniformly all frequency components of the picture signal (approximately 30 cycles to 5 megacycles per second). Performing all these tasks simultaneously leads to a number of complications. Consequently, some compromises are made. For example, one stage is designed to bring the signal out of the noise with little regard to frequency response while another stage may consider frequency response but have little regard for gain. Still another stage may just afford an efficient means of coupling. Therefore, the video pre-amplifier is designed to have the most gain and best signal-to-noise ratio commensurate with its extended frequency response.

A typical stage layout for a camera pre-amp is shown in the functional block diagram of a transmitter, Fig. 29. It is a characteristic of the output circuit of a camera tube to sacrifice the very high-frequency components of the signal in order to get the middle and low range of frequencies above the circuit noise level. When the signal is so weak, circuit noises such as thermal agitation and tube noises cannot be neglected. After the signal has been built up to a higher level by the video amplifier, this initial attenuation of highs must be compensated for in a special stage called a *high-peaker*. The high-peaker amplifies the high-frequency components of the signal more than the lows, and therefore the frequency response is equalized.

The signal is further amplified and matched to a coaxial transmission line through a cathode follower. Cathode followers are the transformers of the television system. Needless to say, no physical transformer can pass the very wide band of frequencies of which the picture signal is composed, and a vacuum tube, properly connected, serves as a transformer (Fig. 30). A vacuum tube so connected has no gain but effectively matches a high-impedance signal to a low-impedance line without a loss in high-frequency response. Cathode followers are used extensively throughout the television station. For theory and mathematics refer to Chap. 14.

32. *Distribution and Control Amplifiers*

The distribution amplifier has a function similar to that of the studio console in a broadcasting station which mixes the signals from the various microphones. The video distribution amplifier mixes the signals from the various cameras and permits cameras to be switched in and out or faded in and out and permits picture from one camera to be superimposed on picture from another. It adds to the versatility of the camera system. Monitoring facilities are also incorporated to permit observation of the scene picked up on each camera although only one is feeding the transmitter.

The control or mixing amplifier is an amplifying, combining, and control amplifier. It is here that all the components of the signal converge to construct the composite television signal. It is in this amplifier that the relative amplitudes of all signal components are properly set.

A block diagram of one type of mixing amplifier is shown in Fig. 29. Picture signal from the distribution amplifier is first increased in amplitude by a video amplifier stage and then is combined with blanking pulses. These blanking pulses are inserted at the proper polarity with respect to the light and dark polarity of the picture signal; it is necessary that the dark portions of the televised scene be nearest the flat top of the blanking pulses.

A simplified schematic of a typical combining stage is shown in Fig. 31. The picture signal appears across both plate resistors with the blanking pulses inserted across the lower resistor. The resultant signal is then applied to a so-called "clipper stage" which clips and levels off the top of the blanking upon which stray signals might appear. The amount of clipping is generally made variable so that there can be a precise adjustment of relative amplitudes between blanking and picture to set properly the background brightness.

The combined picture and blanking are amplified again and passed on to a second combining stage in which the sync pulses are added. Here again the sync pulses must be of proper polarity to appear on top of the pedestal formed by the flat top of the blanking. They must also be of proper amplitude to be the top 25 per cent of the composite signal. The composite sync and composite blanking signals both originate at the sync generator-shaping circuit (Fig. 29).

After the composite signal has been formed, it is coupled, again by means of a cathode follower and coaxial line, to the video amplifiers in the modulator.

33. *Sync Generator*

The sync generator of the television station is the timing unit of the system. In addition to serving as the basic timing unit, it generates, shapes, and arranges in proper sequence all the pulses necessary in the control of the 525-line interlaced system.

The timing unit itself (Fig. 29) generates a control frequency of 15,750 cycles per second (line rate) or 31,500 cycles (double-line rate) and another control frequency of 60 cycles per second (field rate). These timing frequencies control the generation of the various horizontal and vertical pulses. As shown, the line-rate frequency controls the generation of all three sync pulses—horizontal, vertical, and equalizing. This might at first be confusing because we know the vertical synchronization occurs only sixty times per second. However, when the vertical sync and equalizing pulses do occur, they come at a rate of

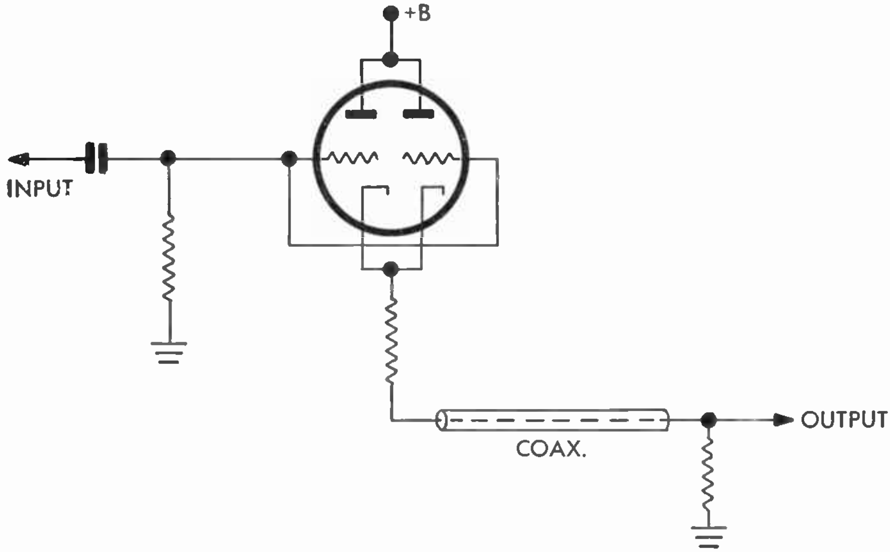


Fig. 30 Cathode Follower

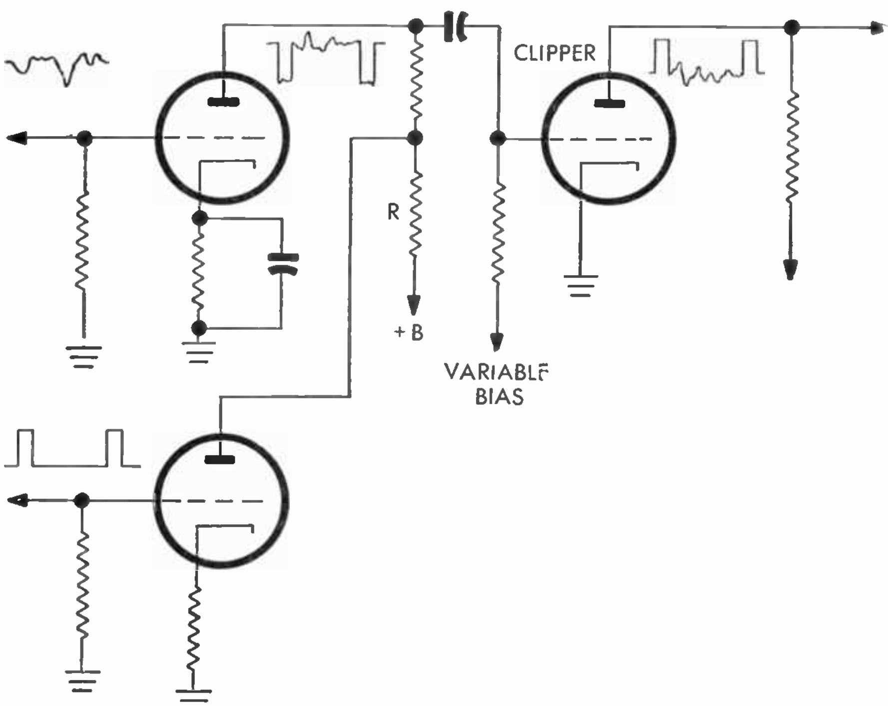


Fig. 31 Pulse Insertion System

31,500 per second, or double-line rate. Thus, their base frequency is double the line rate. To make sure these double-line rate pulses are used only every 1/60 second, a group of keying pulses are generated under control of the field rate control frequency from the timing unit. These keying pulses arrange the sync pulses in proper sequence and form the composite sync signal, which is applied to the control amplifier.

The timing unit also controls the generation of the horizontal blanking and long-duration vertical blanking pulses which are combined and fed as a composite blanking signal to the control amplifier.

Another function of the sync generator is to generate and shape the camera blanking and sync pulses (called *horizontal* and *vertical drive pulses*) which synchronize the camera-sweep system. Thus, at the same instant that a sync pulse is formed for transmission to the remote receiver, a similarly timed sync pulse is formed to control the camera sweep. This, then, is the method used to insure a tightly locked television system. In summation, the sync generator forms and shapes the following pulses:

1. Horizontal sync pulse, which is inserted in the composite signal to control the receiver horizontal sweep.
2. Horizontal driving pulse to control, at the same instant, the camera horizontal sweep.
3. Vertical sync pulses, which are inserted in the composite signal to control the receiver vertical sweep.
4. Vertical driving pulse to control, at the same instant, the camera vertical sweep.
5. Horizontal blanking pulse, which is inserted in the composite signal to black out the receiver screen during horizontal retrace.
6. Vertical blanking pulse, which is inserted in the composite signal to black out the receiver screen during vertical retrace.
7. Horizontal and vertical blanking pulses to blank the camera-tube beam during retrace intervals.
8. Equalizing pulses, which are inserted in the composite signal to produce uniform vertical firing and prevent line-pairing at the receiver.
9. Keying pulses, which are used within the sync generator to insert the various pulses into the composite signal in proper sequence.

34. *Camera Sweep Circuits*

The sweep circuit of the camera has the same basic stages as the deflection system of the receiver—horizontal and vertical sawtooth generators plus horizontal and vertical amplifiers. These circuits generate the deflection currents which move the camera-tube beam through the same standard interlace scanning pattern the picture-tube beam follows. They both move in synchronism because camera and picture tube are locked in by similarly timed pulses.

Camera sweep circuit also consists of a blanking amplifier which increases the amplitude of the composite blanking pulse from the sync generator before it is applied to the blanking grid of the camera tube. In the iconoscope sweep system it is also necessary to use a special correction stage to compensate for the greater distance from the gun to the top of the mosaic than from the gun to the bottom of the mosaic.

A shading generator is at times a part of the camera sweep system although for studio work it is generally an external unit to itself. The shading generator, again under control of horizontal and vertical drive pulses from the sync generator, generates periodic waves (sawtooth, parabola, and sine) of the line and field rate which are injected into the picture signal at the pre-amp to compensate for spurious dark-spot signal. The shading operator varies the controls of the shading generator until the dark spots disappear from the image he watches on the monitor.

35. *Modulator and High-Frequency Transmitter*

After the composite television signal leaves the control amplifier (Fig. 29), it passes to the video amplifiers of the modulator, which increase the amplitude of the signal to a level sufficient to modulate the transmitter. An unusual requirement of the modulator section is that it must amplify a single-polarity signal the average content of which varies with average brightness of the televised scene. A representative line with low average brightness and another

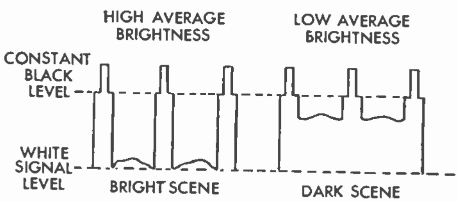


FIG. 32 Composite Signal with High and Low Average Brightness

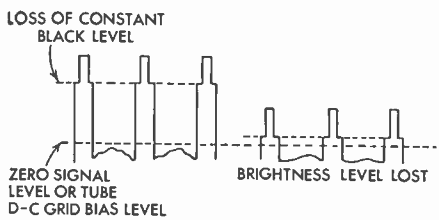


FIG. 33 Loss of Level When Coupling with a Capacitor

with high average brightness are shown in Fig. 32. It is the task of the modulator section to sustain these levels. In particular, the sync tip and blanking levels must be held constant regardless of the variations in picture content and average brightness. Thus, the modulator consists of d-c amplifiers and/or d-c restorers which sustain the levels of the single polarity signal. If such stages were not employed and conventional interstage coupling used, the signals would arrange themselves about an axis of symmetry (plus and minus about zero), and all the levels would shift with changes in scene brightness (Fig. 33), and average signal brightness would remain fixed.

The transmitter proper is very similar to any high-frequency transmitter and consists of crystal oscillator, multipliers, and amplifiers which build the signal up to rated power at the correct carrier frequency. One of the final stages is modulated by the composite signal from the modulator. Generally it is the final stage although some television transmitters use class-B linear amplifiers following the modulated amplifier.

The type of modulation used in television transmitters is conventional grid, cathode, or plate modulation. But the means of applying the modulation to the modulated stage differs. Here again, the coupling from modulator to modulated stage differs because of the necessity for transferring the d-c component of the composite signal. Thus, in television, direct-coupled grid, cathode, or plate modulation is used.

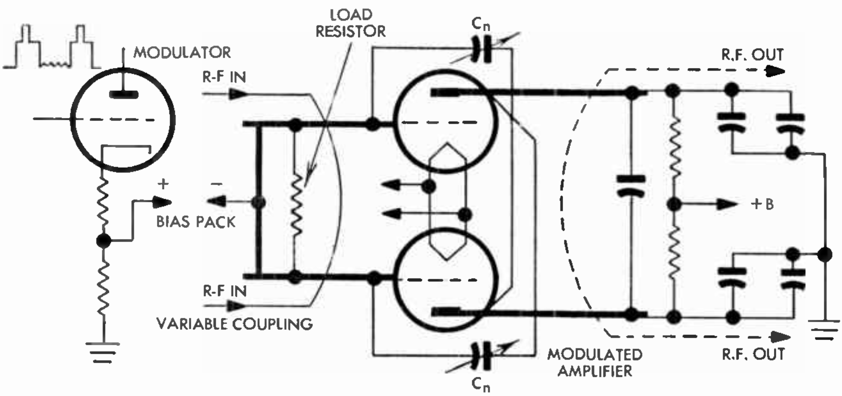


FIG. 34 Direct-Coupled Grid Modulation

A simplified version of a direct-coupled grid-modulated stage for high-powered installations is shown in Fig. 34. In this system the modulator stage is a cathode follower which is coupled through a coupling bias pack to the grids of the modulated amplifier. The actual bias on the modulated-amplifier grids is contributed largely by the bias pack and the current flowing through the cathode resistor. Cathode current varies with the picture signal. Any change in this cathode current, d-c change or a-c, is immediately felt as a change in grid voltage in the modulated amplifier, effectively modulating the r-f output. Since a definite cathode current represents the blanking level, the grid voltage that occurs at this level is also constant and represents a certain r-f power output. Inasmuch as modulator and modulated amplifier are direct-coupled, whenever the composite signal reaches the blanking level there will be this amount of grid voltage regardless of the extent of the variations of the picture signal between blanking pulses.

Since the frequency components of the picture modulation are at times in excess of 4 megacycles, a number of special considerations must be given to the design of the modulated amplifier because it must have a linear output over approximately an 8-megacycle band (both sidebands). A high-frequency component of a certain voltage must produce the same r-f output as a low-frequency of the same voltage. These considerations are:

1. To permit proper adjustment of modulation, r-f excitation, and bandpass, variable coupling exists between modulated amplifier and preceding r-f stage.

2. Bandpass is flattened by a resistor which shunts the tuned circuit (parallel lines) of the modulated amplifier grids. In high-power installations this resistor is often water-cooled because of the high power it must dissipate. Bandpass is also affected by r-f coupling adjustment.

3. Bandpass is affected by loading of the modulated-amplifier plate circuit (variable coupling) and flatness of antenna or linear amplifier when used.

The modulation of the r-f carrier of the commercial television station is termed *negative transmission* because the darker the televised scene, the higher the power output of the transmitter becomes. To obtain this characteristic, the polarity of the modulation must be correct. With grid and plate modulation, the more positive the modulating signal, the greater the power output; with cathode modulation, the more negative the modulating signal, the greater the power output. Consequently, the modulating composite signal must be positive in polarity for grid and plate modulation and negative for cathode modulation.

Some of the characteristics of negative transmission as associated with d-c modulation are:

1. Sync tips always represent a fixed carrier output with an instantaneous modulation close to 100 per cent.

2. Blanking level represents a fixed instantaneous modulation percentage near 75 per cent; it may only vary (plus or minus) 2.5 per cent (assuming sync tip is 100 per cent).

3. Average modulation percentage with no picture signal (only sync and blanking) averages around 80 to 85 per cent (black signal).

4. As soon as picture modulation is applied, the average modulation percentage falls—the higher the average brightness of the scene, the lower the average modulation.

5. The instantaneous picture modulation varies between the 15-per cent (or less) and 75-per cent levels. Instantaneous modulation percentage of 75 per cent represents a very dark spot; instantaneous modulation near 15 per cent, a very bright spot.

From the above characteristics, we observe that blanking and sync represent higher modulation levels than picture signal. The greater radiation during sync and blanking intervals develops, at the receiver, a signal sufficient in amplitude to drive the picture-tube control grid to a negative level which cuts off the fluorescent screen. When the instantaneous modulation percentage is below 75 per cent, during transmission of picture signal, the screen is illuminated—instantaneous illumination being inversely proportional to modulation strength.

36. *Sideband Suppression and Antenna System*

Partial sideband suppression is employed to utilize the allocated channel more fully, as explained, for it permits the transmission of higher frequency components of the picture signal. This in turn permits transmission of a higher definition picture. If we use channel 2, 54 to 60 megacycles, as an example, we find the flat portion of the picture response extends from 54.5 megacycles to 59.25, with the picture carrier located on 55.25 megacycles. Thus the high-frequency sideband is 4 megacycles wide, extending from 55.25 to 59.25 megacycles; the low-frequency sideband is ¾ megacycle wide, extending from 54.5 to 55.25 megacycles. It is, therefore, at 54.5 megacycles that we begin suppressing the low-frequency sideband.

SIDEBAND FILTER

A simplified schematic of a sideband suppression system is shown in Fig. 35. It has the following characteristics:

1. A constant impedance or load is presented to the power amplifier at all times, regardless of the frequency of the modulating signal.

2. Energy is either radiated by the antenna or absorbed by the dissipating resistors. High-frequency sideband, carrier, and part of low-frequency sideband are radiated by antenna; remainder of low-frequency sideband is very largely dissipated by the resistor. In high-powered installations this resistor is water-cooled or is very long, presenting a large radiating surface to the air.

3. A series resonant circuit, *L2* and *C2* at 53 megacycles, shunts the antenna resistance. At the resonant frequency, this appears as a short circuit and prevents frequencies in the vicinity of 53 megacycles, beginning at 54.5 from being radiated by the antenna.

4. A series resonant circuit, *L3* and *C3* at 55 megacycles, shunts the dissipating resistor, preventing the carrier and radiated sidebands from being lost across this resistor. At the same time it presents the proper impedance to the power amplifier to accept and dissipate the unwanted sideband in the form of heat.

5. The input capacitor *C1* and inductor *L1* are of proper value to maintain a constant load on the power amplifier. Therefore, the r-f amplifier generates carrier and both sidebands just as though the sidebands were not present.

6. Other sections of filter sharpen the high-frequency cutoff and remove all traces of signal at the adjacent channel sound frequency of the next highest channel.

Although the tuned circuits in the simplified schematic are shown as having lumped constants, the actual circuits consist of sections of coaxial lines which are, of course, much more efficient at these frequencies. By properly choosing the length of these lines, they can be made to have the same characteristics as tuned circuits.

LOW-LEVEL MODULATION

Low-level modulation (Fig. 36) is also used in the design of some TV transmitters. Modulation occurs in a low-level stage in which only a low-amplitude video signal is necessary to fully modulate the class C stage. This expedient simplifies modulator design. The broad-band modulated signal is now amplified by class B r-f stages which, of course, must now amplify sidebands and carrier—their broadness reducing their efficiency a substantial amount.

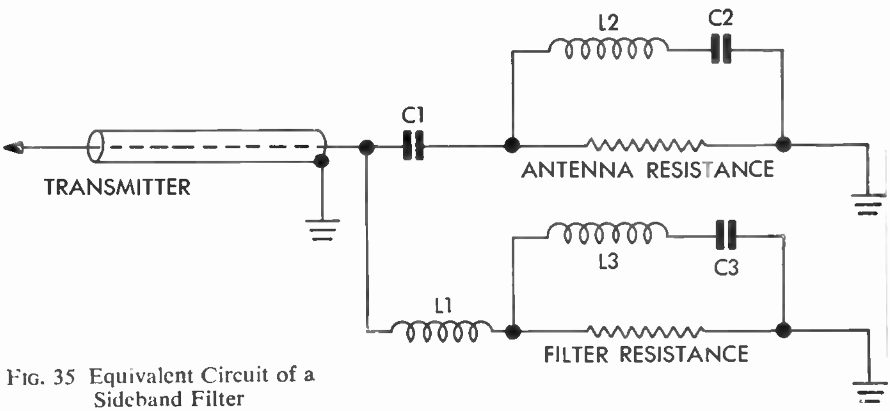


Fig. 35 Equivalent Circuit of a Sideband Filter

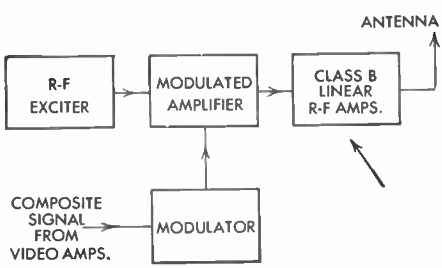


Fig. 36 Block Diagram, Low-Level Modulation TV Transmitter

When low-level modulation is used, a sideband filter is not required because the undesired frequencies can be removed by detuning the class B stages. Thus, in each individual stage the suppressed sideband components are attenuated, reducing these frequencies to an insignificant level at the antenna feed point.

TRANSMITTING ANTENNA

The transmitting antenna of the television picture transmitter is generally a turnstile arrangement which has reasonably uniform horizontal radiation (*i.e.*, emits a signal which radiates with equal strength in all directions horizontally). Thus, a uniform coverage is obtainable.

It is a well-known fact that the radiation from a single dipole is broadside to the antenna; consequently, one dipole would not serve as a good transmitting antenna. If we take two dipoles and mount them at right angles to each other, a more uniform horizontal radiation can be obtained. However, to obtain this uniform radiation, it is necessary that the dipoles be fed 90 degrees out of phase (current delivered to one dipole is maximum at the same instant minimum current is being delivered to the other). A simple method to obtain this 90-degree shift is to make the transmission line feeding one dipole one-quarter wave longer than that which feeds the other dipole.

Another factor which must be considered in the operation of a television antenna is the bandwidth. That is, the antenna must have uniform radiation over a wide band of frequencies in order to radiate the very-high-frequency components (frequencies at outer extremity of passband) and lower-frequency components with the same efficiency. Consequently, the ordinary, straight dipole is not satisfactory, for it is rather sharply tuned at the frequency for which it is cut and falls off rapidly on the sidebands. Instead, a specially shaped antenna, which has a larger effective radiating surface, is used. This type of antenna is broad-band and radiates efficiently and uniformly over the entire channel.

Still another point to be considered is vertical directivity (angle with which signal is radiated with respect to the surface of the earth). Inasmuch as the frequencies used in commercial telecasting are definitely line-of-sight, the vertical angle of radiation should be held down (30 degrees and under). Energy radiated at higher angles penetrates the upper atmosphere and does not return to earth. It represents lost power which might well be used at lower angles to improve signal strength. To improve vertical radiation at low angles, sections of antenna are stacked vertically to increase antenna gain and keep radiation at low vertical angles or the antenna is mounted one-half wave above a good ground surface. (This may be actual ground or a metallic surface, such as the top of a building, which is efficiently grounded.)

37. *Receiver Characteristics*

The television receiver and its antenna pick up both the picture and sound signals. Thus the antenna and r-f section of the receiver must have a reasonably linear response to a 6-megacycle band of frequencies. Picture and sound signals are separated in the input of the i-f system of the receiver, the sound going into a conventional FM channel. The picture is amplified by its own i-f channel to a level suitable for application to the video detector. Detected signal is increased in level by a video amplifier and conveyed to the control grid of the picture tube.

A portion of the signal is also taken off the video detector to be used in the sync and sweep system of the receiver. This signal is segregated into vertical and horizontal components (Fig. 37), and controls the generation of the horizontal and vertical sawtooth voltages. Deflection voltages are increased in amplitude and applied to the deflection system of the picture tube.

Receiving antenna is directional toward the transmitter site and has a broad, flat bandpass (width of channels to be received). If possible the antenna should also have a certain amount of directivity in the vertical plane (low angles with respect to horizon) to improve sensitivity and reduce pickup of noise signals coming in at higher angles.

38. *R-F Section*

The r-f section of the receiver must also be broad band and have a uniform sensitivity over approximately 6 megacycles. In most receivers, an r-f amplifier is used between antenna and mixer to step up the signal before it reaches the mixer and to isolate antenna and mixer. Of course, in all these broad-band circuits, efficiency is relatively low and proper choice of parts and values is imperative to obtain satisfactory results. Miniature tubes are used extensively because of their greater efficiency for wide-band high-frequency service. The mixer stage generally uses a high-frequency pentode, and occasionally you find a receiver with a triode mixer because of its lower noise level. In most cases, a separate oscillator is used to permit an improvement in the conversion action and signal-to-noise ratio of the mixing process. Local oscillator frequency is on the high-frequency side of the picture and sound carriers.

Band switching is generally used in changing channels—the switch being calibrated according to channel numbers. A band-switching arrangement permits use of optimum value parts on all channels. Actual fine tuning is accomplished by varying the local oscillator frequency over a limited range. Of course, the r-f section has its customary trimmers and padders for alignment. While alignment is moderately involved, it is not required too often because of the wide bandpass and more rigidly designed television r-f sections.

The receivers of today are designed to deliver a clean, noise-free picture when 500 microvolts are delivered to the input. They will give satisfactory performance on signals as weak as 150 microvolts.

Two operating controls are a part of the r-f section—*channel-selector switch* and *fine-tuning control*. The selector switch throws in the proper inductors and/or capacitors for the channel to be received; fine tuning varies the local oscillator frequency for proper centering of the signals on the i-f bandpass.

39. *I-F Systems*

The i-f system of the television receiver increases the amplitude level of the mixer output. I-f amplification does much to establish sensitivity and selectivity of the television receiver. There are two basic i-f systems—dual channel, Fig. 37, and intercarrier, Fig. 37a.

In the dual channel i-f system, the picture and sound components are separated at the output of the mixer or in one of the earlier i-f stages. In a typical receiver a sound carrier component can exist on a 21.25-megacycle fre-

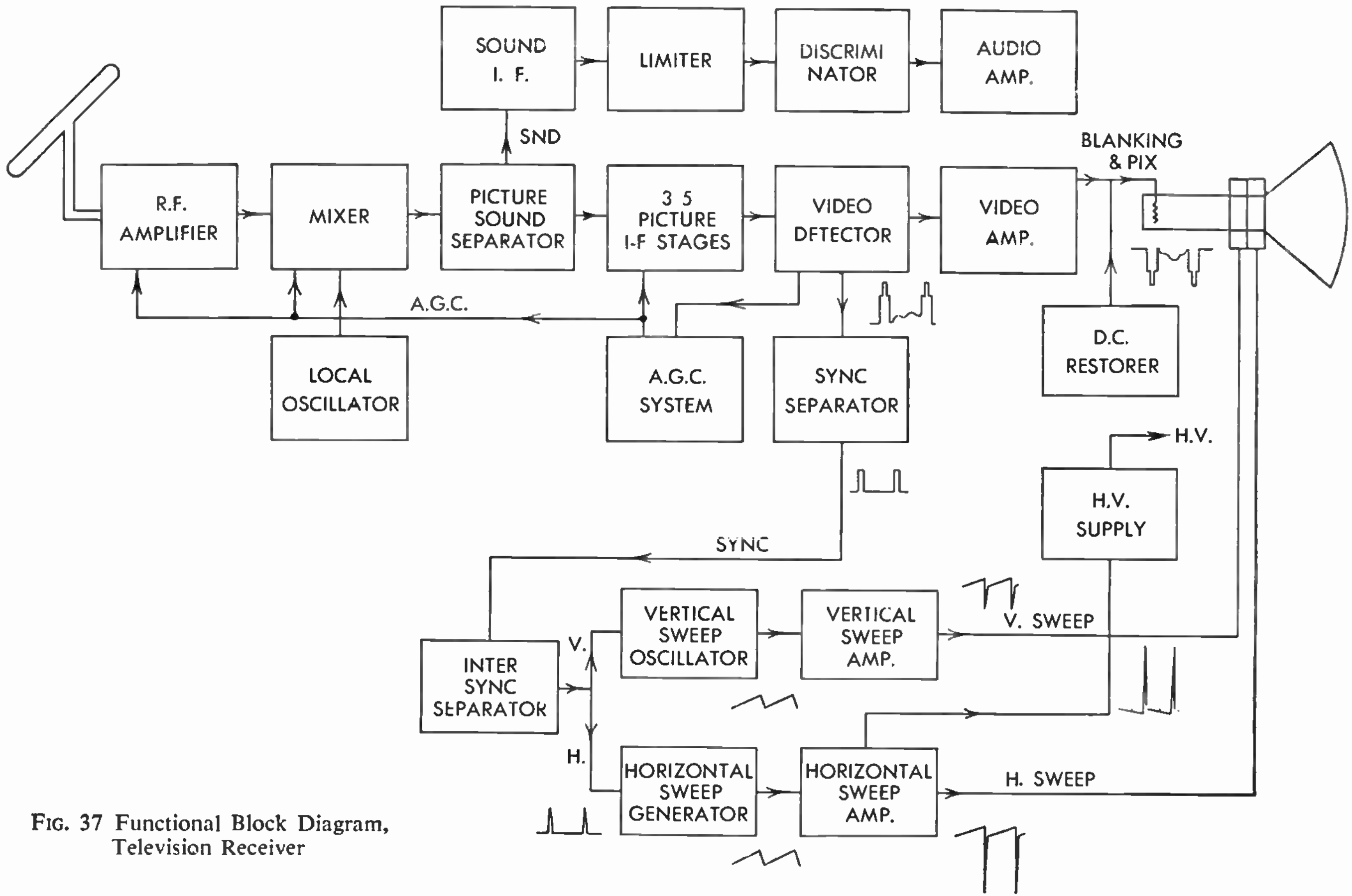


FIG. 37 Functional Block Diagram,
Television Receiver

quency and a picture component on a 25.75-megacycle frequency. Thus if we have two separate i-f systems—one resonant to the picture carrier and sidebands but not to the sound and the second resonant to the sound carrier and sidebands but not to the picture—a means of separating picture and sound becomes apparent. Consequently, the commercial television receiver has only single antenna, r-f section, and local oscillator. Picture and sound carriers appear at the plate of the mixer and are separated in the i-f system or video amplifier.

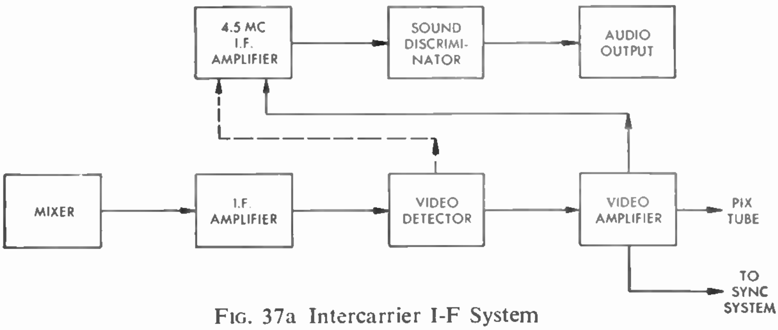


FIG. 37a Intercarrier I-F System

The picture i-f system is again a wideband amplifier, and although every attempt is made to obtain high gain, the strict bandwidth requirements mean sufficient gain is obtained only with the use of four to five i-f stages. The i-f amplifier will deliver a 3- to 10-volt signal to a video detector for the normal ranges of received signal strength. The detector output will give a 2- to 6-volt-peak detected signal.

The over-all response curve of the i-f system is reasonably flat over a 3-megacycle band. Gain near the picture-carrier frequency is less than on the lower frequency range of the bandpass. This tapered response compensates for the partial sideband suppression which occurs at the transmitter. A few other special circuits are used in the i-f system to reject the associated sound carrier, adjacent channel picture carrier, and adjacent channel sound carrier. This prevents the appearance of interference bars on picture. In some receivers one operating control, or contrast control, is a part of the i-f amplifier. It varies the gain of the amplifier and, therefore, indirectly the actual peak-signal voltage applied to the grid of the picture tube.

In the intercarrier receiver both picture and sound carriers are amplified in a single channel. Picture and sound components are separated at the video detector or in the video amplifier. In the amplification of both carriers in a single channel, one carrier (picture) is made to dominate greatly the other (sound). Consequently, the weaker carrier assumes the characteristics of a sideband of the stronger—a sideband separated by 4.5 megacycles from the carrier (separation of 4.5 megacycles between the transmitted picture and sound signals from station).

The sound-carrier sideband appears as a 4.5-megacycle component at the output of the video detector. A resonant circuit at the output of the video detector or in the video amplifier can remove this sound carrier and apply it to a 4.5-megacycle-sound i-f channel. This same resonant circuit and similar ones also act as traps and prevent the appearance of the 4.5-megacycle component on the picture. They also prevent picture components from reaching the sound channel and producing audio buzz. The sound channel contains an amplifier, a discriminator, and an audio output.

40. *Video Amplifier*

After the picture signal is detected by the second detector, it is stepped up in amplitude by the video amplifier before it is impressed on the control grid of the picture tube. The picture signal, as it is detected, is of a single polarity and similar to the signal as it originally appeared at the transmitter. In most cases the receiver video amplifier is not direct-coupled throughout, and the levels of the single polarity signal are not sustained. Consequently, the video amplifier often contains a d-c restorer which re-establishes the sync tip and blanking levels. As a result, the sync tip and blanking levels remain constant while the average bias on the picture-tube grid varies with the average brightness of the scene, as it should.

The gain of the video amplifier is sufficient to raise the peak video signal up to approximately 60 to 120 volts for the picture tube. One operating control, *brightness,* is associated with the video amplifier. It sets the d-c bias on the picture-tube grid and, therefore, the average illumination of the fluorescent screen. In some few receivers the contrast control is also in the video amplifier and varies the gain of the video amplifier instead of the i-f system.

41. *Sync and Sweep System*

The sync and sweep system, under control of the received sync pulses, generates the deflection voltages used to sweep the beam across and down the fluorescent screen in synchronism with the camera-tube beam. First, the sync is clipped from the composite signal by the sync separator (Fig. 37). Then the intersync separator divides it into horizontal and vertical components. The sharp leading edges of the sync pulses (horizontal, vertical sync, and equalizing) represent horizontal sync components; the long-duration flat top of the vertical sync pulses form the vertical sync component.

Each sync component synchronizes the respective horizontal or vertical sawtooth-forming oscillator or system. The properly timed sawtooth voltages are now amplified and applied to the deflection coils or plates of the picture tube. When deflection coils are used, a sharp transient voltage is generated inherently in the horizontal coils; this is often rectified and used as a source of high voltage for the picture tube.

Six operating controls—*horizontal* and *vertical hold, horizontal* and *vertical linearity, height,* and *width*—are associated with the sync and sweep system. The hold controls are simply frequency controls which bring the oscillators within the range of control of the sync pulses. Once the sync pulses grab on, the horizontal will lock in on 15,750; the vertical, on 60. The height and width controls set the peak amplitude of the sawtooth and therefore determine the width-and-height area of the fluorescent screen which the beam scans. The linearity controls ensure a straight-line sawtooth and, therefore, prevent crowding and distortion of the picture as it is painted on the fluorescent screen by the electron beam, element by element and line after line.

QUESTIONS

1. What are the essential differences between iconoscope and image orthicon?
2. Briefly describe operation of iconoscope.
3. Briefly describe operation of image orthicon.
4. What is the purpose of high peaker?
5. What is the function of each pulse generated by the sync generator?
6. Explain in detail just how camera-tube and picture-tube beams are synchronized together.
7. What unusual characteristics of the composite television signal require use of d-c amplifiers or d-c restorers?
8. Describe action of sideband filter.
9. List the functions of the distribution and control amplifiers.
10. Draw, from memory, a complete block diagram of the transmitter.
11. What must be the frequency of the local oscillator on channel 6 to produce a 25.75-megacycle picture and a 21.25-megacycle sound?
12. Explain just how it is possible to use one r-f section and one local oscillator to receive two signals simultaneously.
13. Why must the response of the picture i-f system be so broad?
14. What are functions of sync separator?
15. What are functions of receiver video amplifier?
16. What is function of intersync separator?
17. Why is it necessary to have both a horizontal and vertical sawtooth generator?
18. Does a vertical sync pulse also contain a component of horizontal sync?
19. If a 3-megacycle component of modulation is released from the camera tube, what i-f frequency will represent it in the picture i-f channel of the receiver?
20. Draw, from memory, the complete block diagram of the receiver.

Chapter 4

R-F AND I-F SYSTEMS

42. *Wide-Band Amplification*

In the amplification of the television r-f or i-f signal it is necessary to amplify a very wide band of frequencies because frequency components of modulation in the television system are at times in excess of 4 megacycles. The r-f section of the television receiver, on each channel, must amplify linearly a 6-megacycle band of frequencies. The i-f amplifier of the receiver must also amplify a band of frequencies almost as great. Sharply resonant circuits, although they represent high Q and efficiency, cannot be used because they amplify only a narrow band of frequencies about the resonant frequency. In wide-band service, Q and stage gain are not the only considerations. Instead, we try to obtain the best gain at the required bandwidth. In fact, we must sacrifice gain to obtain linear amplification over a wide band of frequencies.

The gain of a wide-band stage can be closely approximated by the simple formula: Gain $= g_m R_L$. The resistive component of the load (Fig. 38) or the R in the formula is, in many cases, set by an actual physical resistor which is shunted across the output tuned circuit to reduce its Q. A lower Q, of course, means that the tuned circuit is not sharply resonant and passes a band of frequencies. However, the actual peak gain of this stage is reduced in the presence of this shunt or load resistor. At times this resistive component of the load also includes the resistance of the amplifier tube in shunt with the loading resistor in the case of a low-resistance tube. More often if a low Q-tuned circuit is used, the resistive component (at times a reflected component) of the tuned resonant circuit also forms a part of the resistive component of the load. However, inasmuch as the actual physical loading resistor shunted across the tuned circuit is so low, a very practical approximation of the stage gain can be obtained by multiplying the g_m times the value of the loading resistor.

The value of this resistive component, which must shunt the output tuned circuit of a wide-band amplifier, per given band width can be calculated from the following simple formula: $R = X_r f_o / 2f_\Delta$ in which X_r represents a reactive branch of the tuned circuit at the resonant frequency, f_o is the resonant frequency of the tuned circuit, and f_Δ is the bandwidth or change in frequency.

It is apparent, therefore, that the resistive component of the load can be made higher, and therefore the stage gain is larger, when the reactance of inductor and capacitor is high and when the ratio of the resonant circuit frequency to the bandwidth is low. Thus, the greater the bandwidth required, the lower the value of the resistor, and therefore the more we load the tuned circuit and the more we reduce the stage gain (Fig. 38). To obtain an appreciable signal with such a low value of load resistor, it is necessary that the tubes have a high mutual conductance. The mutual conductance, or g_m, of a vacuum tube is a measure of how efficiently the tube converts a small change in grid voltage to a large change in plate current. It is the large change in plate current for a

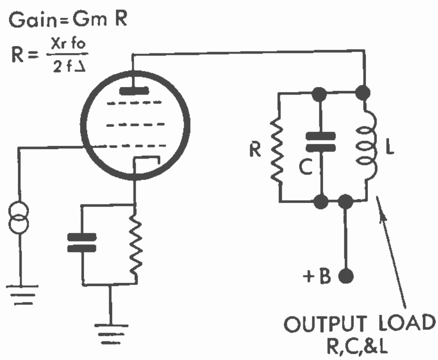


FIG. 38 Wide-Band Amplifier

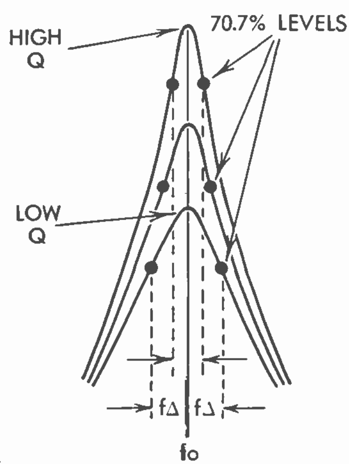


FIG. 39 Circuit Q and Bandwidth

given grid signal that we require in a wide-band stage to develop an appreciable voltage variation across a low value of plate-load resistance. Thus, the only tube that is really practical in wide-band service is a high-g_m tube such as the 6AC7, the 6AK5, and the 6AG5. These tubes also have low input and output capacities, another feature of a wide-band amplifier necessary to obtain the best gain at a prescribed bandwidth.

To obtain the best gain at the required bandwidth, it is also necessary to keep the reactive branch of the load high. The reactive component of the load, of course, is dependent on the L-to-C ratio of the tuned circuits. If the inductance is high and the capacity low, as in a high L-to-C-ratio tuned circuit, the inductive and the capacitive reactance is high. At the resonant frequency, of course, these two reactances are equal and act in shunt to produce a resistive component, or X_r in the formula, which is high when the inductive reactance and the capacitive reactance are high. Thus, the tuned circuits of the wide-band stages are designed with a high L-to-C ratio because the more inductance and the less capacity we have per given resonant frequency, the higher the X_r of the tuned circuit.

The L-to-C ratio of the wide-band stage is made high by keeping the capacity low, using the following arrangements:

1. Tubes such as the 6AC7 and the newer miniature-type tubes are used because their input and output capacities are so very low per g_m.
2. Wiring capacity and parts capacity to ground are held to an absolute minimum.
3. In many receivers, no physical capacitors are used to tune the tuned circuits; instead, the inductor of the tuned circuit resonates with the total distributed capacity of the circuit at the required frequency. The tuned circuits are adjusted to the exact frequency by means of movable cores in the inductors.

Another factor which increases the permissible value of the R for a given bandwidth is the ratio of the tuned-circuit frequency to the bandwidth. It is apparent that the permissible value of R is higher when the bandwidth is low. Likewise, it is higher when the tuned-circuit resonant frequency is high in comparison to the bandwidth. Therefore, it appears expedient to increase the frequency of the wide-band stage, or in the case of an i-f system to increase the i-f frequency, in order to obtain a greater gain for given bandwidth. This is exactly what has been done in the newer model receivers, which has increased their i-f frequency to the 40- to 45-megacycle region and higher from the 20- to 25-megacycle region. However, to take full advantage of the increase in the basic i-f frequency, the reactive component of the tuned circuits must also be high. In the prewar television receiver it was not feasible to increase the i-f frequency because of the limitations of the tubes and tuned circuits. Before the development of low-capacity tubes and efficient high-frequency resonant circuits, the L-to-C ratio decreased too much when the resonant frequency of the i-f system was increased. If the L-to-C ratio falls, of course, the X_r in the formula decreases. Consequently, there was no rise in the permissible R because, as the f_0/f_Δ increased, the X_r decreased. However, with modern, efficient design and the use of miniature tubes, the X_r is held near the same value when the frequency of the i-f is raised.

When double-tuned transformers or bandpass T-load circuits are used in the i-f stage, the gain formulas become much more complex. However, for each individual tuned circuit it is possible to use the simple formula given above to find the load resistor value. In summation, the following general characteristics apply to all types of wide-band stages:

1. Low-capacity, high-g_m tubes.
2. High L-to-C-ratio tuned circuits.
3. Resistor loading to increase the band of frequencies amplified linearly.
4. Some means of obtaining broad bandwidth by means of overcoupling of tuned transformers or the use of stagger tuning.

For mathematical presentation of wide-band amplification, refer to Chap. 14.

43. *Overcoupling*

The basic method of obtaining overcoupling between two tuned transformers, and a resultant broader bandwidth, is to place the two coils in very close proximity to each other. When this is done, the double-hump response characteristic (Fig. 40) is obtained. The separation between the peaks of the double-hump characteristic and, therefore, the bandwidth of the bandpass, is determined by the degree of coupling between the two windings. The closer the windings, the farther the two humps are spread. However, there is a practical limit to the degree of coupling because, as the double hump spreads, the amplitude goes down and the valley in the center becomes too deep. Nevertheless, this system of overcoupling is used extensively because even a small amount of overcoupling widens the band of frequencies passed by the tuned circuits.

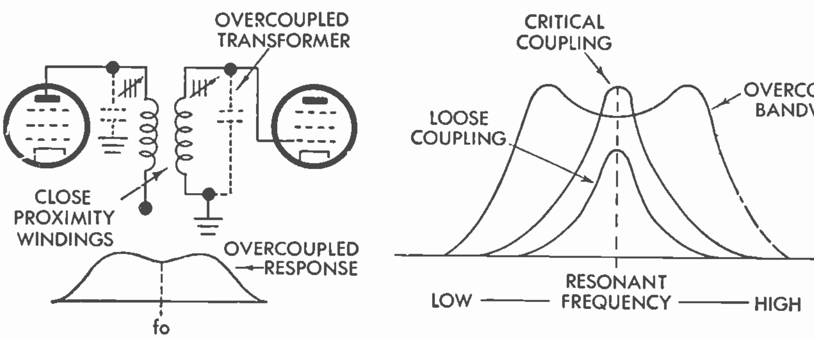


FIG. 40 Overcoupled Transformer

In addition to obtaining overcoupling by close magnetic coupling of the windings of the tuned transformer, other methods of overcoupling (Fig. 41) are used. The methods shown use a mutual element capacitor, inductor, or tuned circuit to overcouple between the primary and secondary windings. It is not now necessary that the magnetic lines of the primary link the secondary. In many cases the primary and secondary windings are mounted in separate shields. In circuit *A* the high side of the two windings are linked by a small trimmer capacitor. With this method a very broad bandpass is obtained, and the extent of the bandpass can be controlled by the value of the capacitor. A typical value for this capacitor would be somewhere between 2 and 10 micromicrofarads. In circuit *B* a common inductor links the two windings. A blocking capacitor in this case is necessary to prevent transfer of the plate voltage.

In circuit *C* an actual tuned circuit (not tuned to the resonant frequency of the transformer) is used as a means of mutual overcoupling. The advantage of this particular type of overcoupling is that the tuned coupling circuit serves two functions—the first, as the means of overcoupling; and the second, as a tuned resonant circuit that presents a maximum impedance to some undesired frequency, preventing the transfer of this frequency between primary and secondary windings. In this application it is called a *tuned trap,* and in many television receivers it is tuned to the adjacent channel sound frequency or the associated channel sound frequency, two frequencies which are not wanted in the picture i-f system because they cause a series of bars to appear on the picture. It is interesting to note that the tuned trap acts as a mutual capacitor or a mutual inductor at the picture i-f frequency, depending on which side of the picture i-f frequency it is tuned. If it is tuned to some undesired frequency above the picture i-f frequency, it acts as an inductor because the lowest reactance element dominates in a parallel circuit. When it is tuned to some undesired frequency below the picture i-f frequency, it acts as a mutual capacitor so far as overcoupling is concerned. In other television receivers a bandpass T transformer is used to couple between the plate of one stage and the grid of the next. This type of bandpass or overcoupling system is shown in Fig. 42. Again, either a capacitor or an inductor can serve as the mutual overcoupling element.

In most of the overcoupling examples mentioned, no physical resonant circuit capacitor is used. The distributed circuit capacity of the stage serves as the capacitive element of the resonant circuit. A fine adjustment of the resonant circuit is obtained with a movable core.

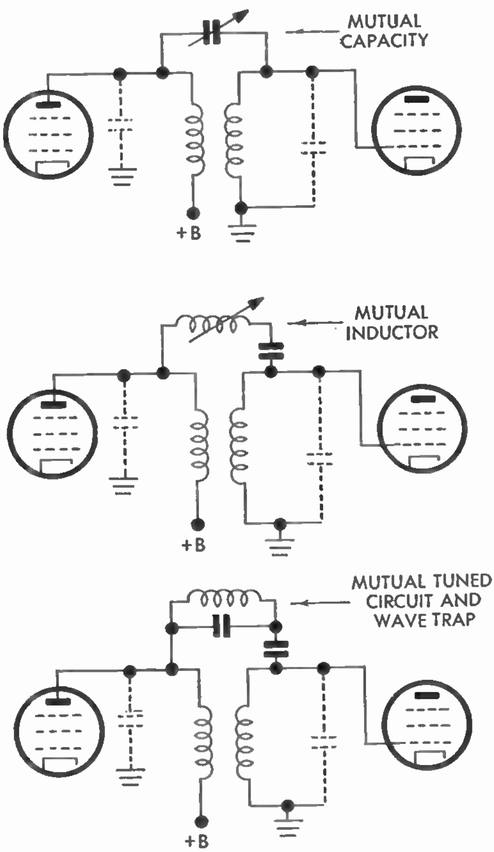


FIG. 41 Overcoupling Methods

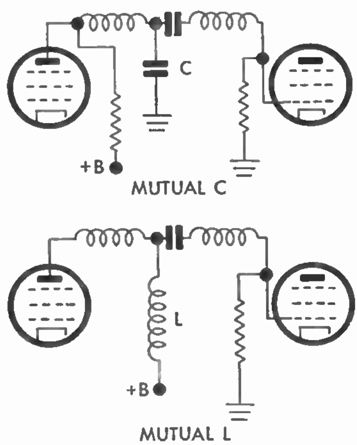


FIG. 42 Bandpass T Broad-Band Coupling

44. *Stagger Tuning*

A system called *stagger tuning* is also used to amplify linearly a broad frequency band. Basically, the stagger-tuned i-f system (Figs. 43 and 44) consists of i-f stages which are not all tuned to the same frequency. For example, in Fig. 43 only alternate stages are tuned to the same frequency.

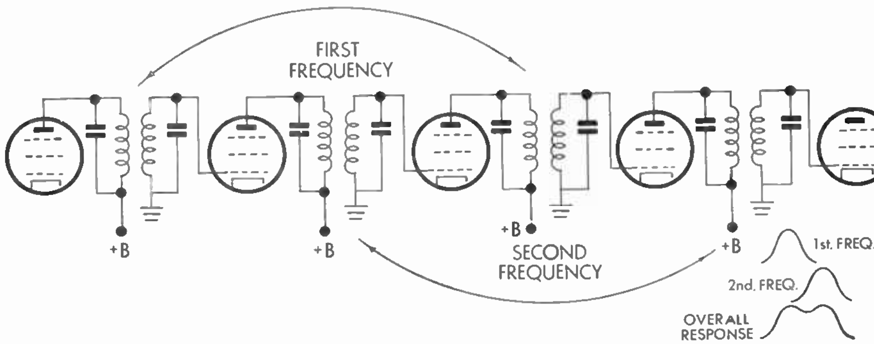


FIG. 43 Stagger-Tuned Wide-Band Amplifier

Thus, the response of an individual stage is not too broad; however, the response of a number of stages, each tuned to a different frequency, produces a broad bandpass characteristic. The over-all response characteristics of the stagger-tuned i-f system is a double-hump characteristic similar to the double-hump characteristic of an overcoupled stage. It is possible to remove the valley of the double-hump characteristic by tuning a stage or two to a frequency midway between the two resonant frequencies of the stagger-tuned i-f system. In this case the i-f system becomes a triple-tuned i-f system.

This stagger-tuning idea can be carried still further; in fact, in the RCA receivers the individual stages are each tuned to a different frequency. A simple breakdown of this type of i-f system is shown in Fig. 44. It can be seen that each stage has its own individual resonant frequency and response

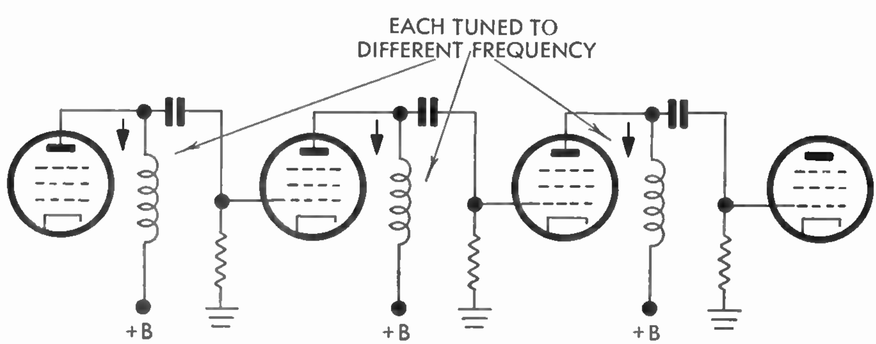


FIG. 44 Stagger-Tuned, Single-Tuned Transformer

and that the over-all response of all stages is the ideal response characteristic of a picture i-f system. Some stagger-tuned i-f systems use a double-tuned transformer (Fig. 43), while other systems use just a single-tuned transformer between i-f stages (Fig. 44).

45. *Wavetraps*

Wavetraps are used throughout the i-f system of the television receiver to reject unwanted frequencies. These traps are particularly designed to reject the many sound i-f frequencies. The sound modulation on these i-f carriers would cause a disturbing bar pattern on the picture-tube screen if they were permitted to reach the picture detector. Some locations for wave traps in the

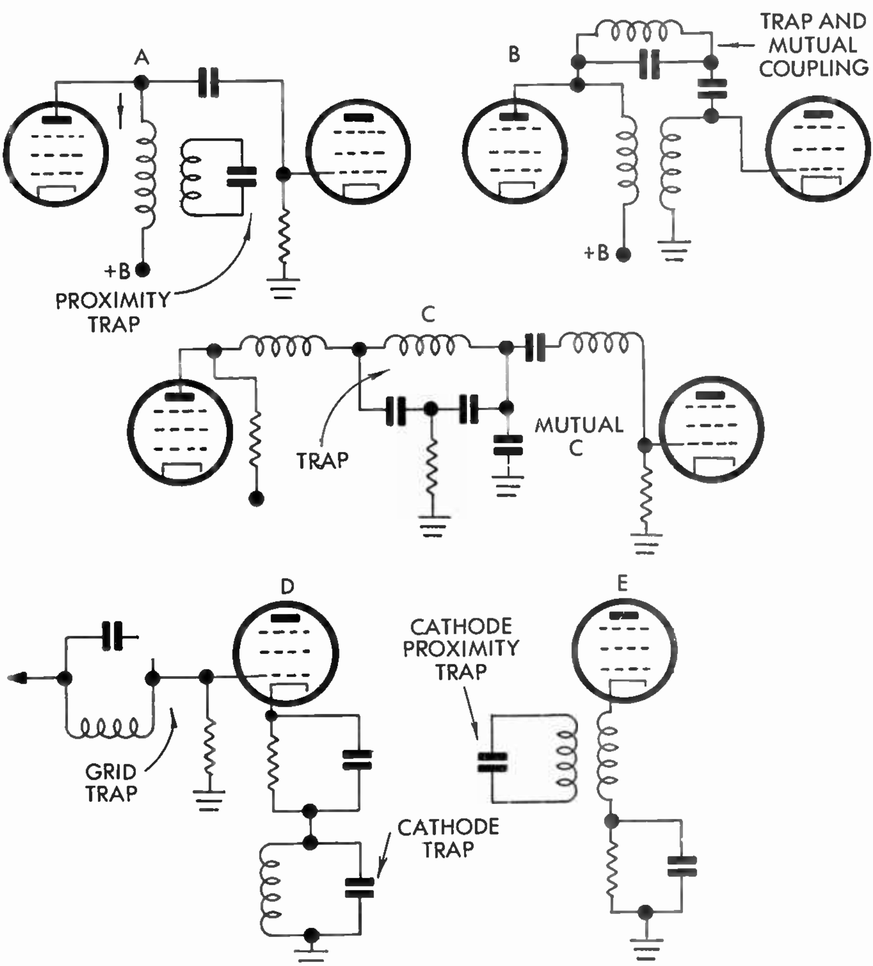


FIG. 45 Insertion of Wavetraps

picture i-f systems are shown in Fig. 45. In circuit *A*, a tuned circuit is placed in close proximity to the regular inductor of an i-f tuned transformer. This tuned trap is floating and removes energy at its resonant frequency from the regular i-f system in the same manner as an absorption wavemeter would draw energy from a tuned circuit in a transmitter. In this case, however, the trap is placed very close so that it absorbs a good amount of the undesired frequencies which might be present in the regular tuned circuit of the i-f stage. This is called a *proximity wavetrap*.

In circuit *B* the tuned transformer is placed between the high sides of the regular primary and secondary of the tuned transformer between i-f stages. In this position it serves as a trap to the frequency to which it is tuned and, at the same time, serves as a mutual coupling element at the frequency of the picture i-f system. Inasmuch as it is a parallel resonant circuit or a maximum impedance resonant circuit at the frequency at which it is tuned, no signal of that frequency is transferred between the primary and secondary windings.

A wavetrap can also be inserted into the bandpass T method of coupling, as shown in circuit *C*. Other parallel resonant circuits can be inserted at various other points in the picture i-f systems to reject the unwanted frequencies. For example, a small trap can be inserted in series with the grid lead, as shown in circuit *D*. Here again, it is a parallel resonant circuit in series with the signal path and, therefore, rejects the frequency to which it is tuned. The same kind of trap can be inserted in the cathode circuit of a tube, as shown in circuit *E*. This serves as a very efficient means of rejecting unwanted frequencies because the amount of rejection is less dependent on the Q of the resonant circuit. Actually, in this type of connection, a degenerative voltage is developed across the parallel circuit in the cathode which almost completely nullifies the grid signal. Thus, the effective grid signal is very much reduced at the frequency to which the parallel circuit is resonant. At other frequencies the impedance of this circuit is, of course, very low, and the cathode of the tube is effectively returned to ground. Undesired energy and undesired frequencies can also be absorbed from a cathode coil by means of a proximity trap, as shown in circuit *F*.

46. *Miniature Tubes for Television*

The miniature tube is synonymous with economical, efficient television-receiver production. It not only permits reduction in size of the television chassis, but its operating characteristics permit broad bandwidth and improved amplification because of higher mutual conductance and low interelectrode capacities.

The figure of merit of a vacuum tube used in wide-band service is the ratio of its mutual conductance to the sum of the tube's input and output capacities. The mutual conductance, or g_m, of a tube is the ratio of a change in plate current to a given change in grid voltage, or: $g_m = dIp/dEg$. Consequently the g_m of a tube is the measure of how effectively a tube can convert a small change in grid voltage to a large change in plate current. Of course, the larger the plate-current change per given grid signal, the greater the output voltage- or plate-voltage variation. Actually, in wide-band application, the output voltage is the product of the plate-current change times the load impedance. This load impedance is largely the resistance of the tuned-circuit load resistor in the case of a wide-band r-f or i-f amplifier.

In an r-f or i-f amplifier in wide-band service, mutual conductance and interelectrode capacity are important. The gain of such a stage (Fig. 46) is determined mainly by the value of the shunt load resistor, which shunts the tuned circuit to broaden its bandwidth. The approximation of stage gain for such a stage, as discussed before, is g_m times the load resistance. The value of the load resistor per given bandwidth is determined by the inductance-to-capacity ratio of the tuned circuit which it shunts. Inasmuch as the tube interelectrode capacity is a part of this tuned circuit capacity, the L-to-C ratio becomes higher when the interelectrode capacity is low. Here again, the value of the shunt load resistance is set by the interelectrode capacity, and its value becomes higher as the interelectrode capacity is lowered. Thus, the tube with high g_m and low capacities gives us a greater gain per given bandwidth.

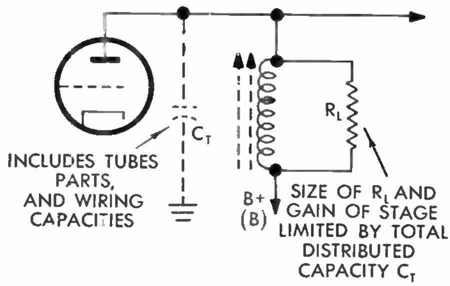


FIG. 46 Output Circuit of Wide-Band R-F or I-F Amplifier

MINIATURE TUBE DATA CHART

Tube Type	Capacitances gp	gk	pk	Voltages E_p	E_{sg}	E_c	g_m
6J4	4	5.5	0.24	150	—	$R_k = 100$	12,000
6J6	1.6	2.2	4	100	—	$R_k = 50$	5.300
6AK5	0.02	4	2	180	120	$R_k = 200$	5.100
6AG5	0.025	6.5	1.8	100	100	$R_k = 100$	4,750
6BA6	0.0035	5.5	5	100	100	$R_k = 68$	4,300
6C4	1.6	1.8	1.3	250	—	−8.5	2,200

The physically small miniature tube has these characteristics—high g_m and low electrode capacities. (Refer to the accompanying chart.) Those characteristics of a miniature tube which make it an efficient, low-capacity, wide-band amplifier are:

1. Physically small elements.
2. An efficient cathode and close spacing between control grid and cathode.
3. Small surface area and short leads along with ample spacing between leads coming from the individual elements of the tube.

Small physical surface means there is less capacity between elements. At the same time, the control grid is closer to the cathode and exerts more control over the electron stream. Therefore, the mutual conductance is higher. Although the capacity between control grid and cathode increases as spacing is decreased, the mutual conductance increases as the square of the decrease in spacing. Consequently, the figure of merit of the tube is higher because figure of merit is the ratio of conductance to the interelectrode capacity. The small physical size of the element keeps the interelectrode capacity well down beneath that of the conventional tube.

The miniature tube is also a very efficient tube at high frequency because of its high-gain-to-low-noise characteristics. It is an inherent quality of high-frequency receivers that the dominant noise is the tube noise generated by the first r-f amplifier or mixer. This noise is minimized in a high g_m tube because the larger plate current variations bring the signal above the noise. Thus the signal-to-noise ratio in a wide-band television receiver is very much higher when miniature tubes are used. In the prewar television receiver it was impossible to obtain an appreciable gain because of the broad band of frequencies which had to be amplified. In the new receivers, higher-g_m miniature tubes replace the high-g_m metal tubes, such as the 6AC7, giving a substantial increase in gain over a wide band of frequencies.

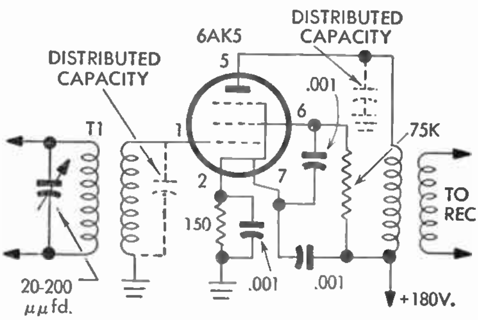


FIG. 47 Typical R-F Stage, T1 Overcoupled

A typical r-f stage is shown in Fig. 47. Such a stage could also be an addition to a present television receiver, mounted on the television chassis or at some remote point. Its small size makes it convenient for mounting on a small subchassis which can be attached to the main chassis of the receiver. It is also possible to mount this r-f box close to the antenna termination where it is free from noise and can amplify the weak signal being received before it is conveyed through coaxial cable to the main section of the receiver. In constructing such an amplifier it is necessary to keep the r-f leads short and the wiring capacity at a minimum to take full advantage of the low-capacity characteristics of the tube.

To keep the L-to-C ratio and the gain high, it is necessary that the only capacity we have in the grid-tuned circuit and plate-tuned circuit is the capacity of the tube and of the wiring. Thus, no physical capacitor shunts the tuned circuits. The tuned circuits are tuned to resonance by changing the inductance of the coil. The resonant frequency is set by the number of turns of wire on the coil and, over a limited range, by the spacing of the turns.

If the response of the r-f stage is too sharp, as indicated by ability to receive the picture but not the sound, or vice versa, or as indicated by a loss of horizontal resolution, the response of the r-f stage can be broadened at the expense of gain by shunting the tuned circuits with resistors. The bandpass of the r-f stage can be broadened and the gain of the stage sustained by means of a double-tuned input transformer (Fig. 47). The primary of the tuned transformer is shunted by a physical capacitor, which lowers the L-to-C ratio and the impedance of the primary in order to match the antenna properly to the input circuit of the r-f stage. The broad bandpass characteristic is obtained by overcoupling the primary and secondary windings of the input transformer, producing a double-humped characteristic, the bandwidth of which can be controlled by the degree of overcoupling between primary and secondary windings. Overcoupling is obtained by bringing these primary and secondary windings in very close proximity to each other, nearly touching each other.

Many miniature tubes have two cathode connections (pins 2 and 7 of the tube socket in Fig. 47) to reduce the degenerative effects of cathode-lead inductance at high frequencies. Screen and plate by-pass capacitors are returned directly to pin 7; cathode bias and grid circuit, to pin 2.

The miniature tube is also an efficient mixer and i-f amplifier tube. The high conversion transconductance of the miniature tube produces a signal with a high signal-to-noise ratio.

The miniature tube can also be used as a local oscillator in the television receiver because of its high mutual conductance and its high-frequency stability. Miniature tubes can be used throughout the television receiver—in sound channels, video detectors, video amplifiers, sync circuits, and sweep circuits. It is only in the power stages that it is advisable to use a conventional tube. It is apparent that the miniature tube reduces the physical size of the television chassis and standardizes tube types. Very few receivers will be manufactured without their quota of miniature tubes.

47. *Grounded-Grid and Cathode-Coupled Amplifiers*

Two tubes suitable for wide television application are the miniature triodes 6J4 and the dual triodes 6J6. The 6J4 miniature triode is a grounded-grid amplifier, and the 6J6 consists of two high-g_m triodes. It is an admitted fact that the advantage of the pentode in the amplification of i-f and r-f frequencies is its high plate impedance, which means it will amplify a small signal voltage to a much greater extent. However, in wide-band amplification of i-f and r-f frequencies, the plate load impedance itself must be lowered in order to pass the wide band of frequency; consequently, the advantages of the pentode have been nullified. Thus, it is possible in wide-band service to use a triode with a high g_m to give us approximately the same gain as a normal pentode would. An added advantage of using a triode in this type of service is that its inherent noise is much lower because of the absence of additional grids. Another advantage of the triode in this service is its inherent lower impedance, which means that less severe external loading is necessary to cover a band of frequencies. In fact, the triode amplifier is linear over a substantial band of frequencies without any external loading at all. This is particularly the case when a cathode-coupled arrangement is used.

A typical grounded-grid r-f amplifier is shown in Fig. 48. Note that the grid is grounded, that the cathode and the plate are above ground, and that the stage uses a tuned input circuit as well as a tuned plate circuit. Normally a

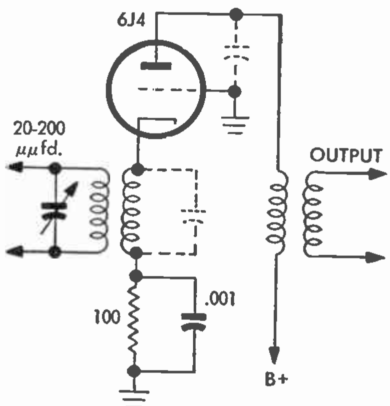


FIG. 48 Grounded-Grid R-F Amplifier

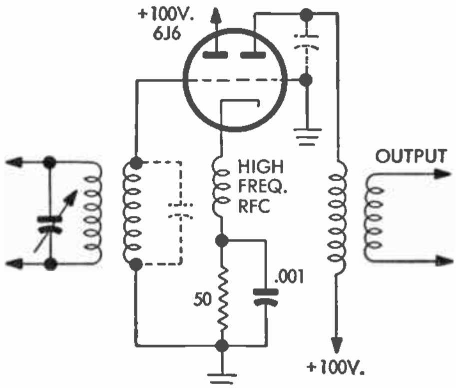


FIG. 49 Cathode-Coupled R-F Amplifier

tuned-grid tuned-plate stage would oscillate; however, in a grounded-grid stage the grid is grounded and acts as a shield between input and output circuits. Any feedback due to plate-cathode capacitance is not in proper phase to produce oscillation. The advantages of the grounded-grid amplifier are as follows:

1. Low impedance and an inherently broader bandpass characteristic.
2. Low electrode capacities, permitting high L-to-C ratio and greater gain tuned circuits.
3. Low tube noises and a better signal-to-noise ratio.

The cathode-coupled amplifier (Fig. 49) uses the two sections of the 6J6 dual triode—first section is a cathode follower; next section, a grounded-grid amplifier. The gain of this stage, although it consists of two triodes, is equivalent to the gain of a good single-pentode stage. The number of component parts are approximately the same; however, the cathode-coupled stage has the following advantages:

1. Wide bandpass characteristics because of the inherently low impedance triode.
2. Low noise characteristics, because internal noise is low in a triode.
3. Low input and output capacities and consequent high L-to-C tuned circuits.

4. Grounded-grid connection to minimize tendency to oscillate. The coupling between the two triodes is due to common impedance of the cathode inductor, which has a substantial reactance at the frequencies to be passed.

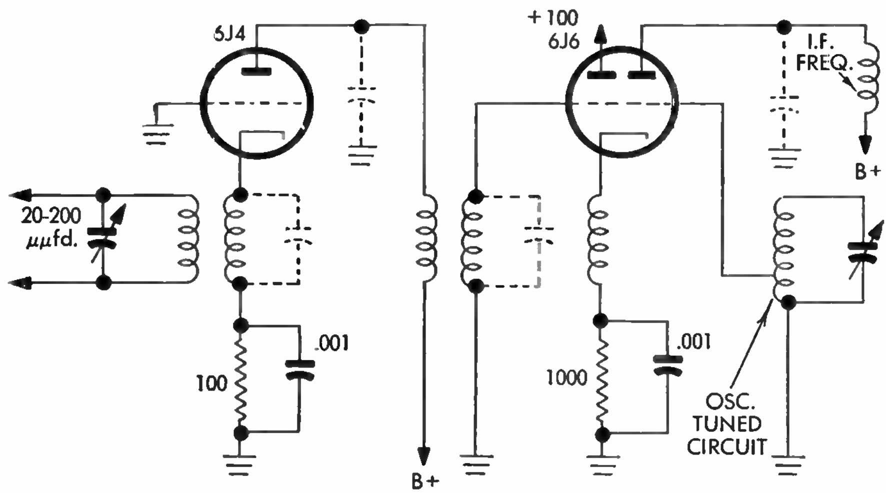


FIG. 50 Cathode-Coupled Mixer and Oscillator

Another application for the cathode-coupled stage is shown in Fig. 50. In this circuit, which shows a grounded-grid r-f amplifier and a cathode-coupled stage following it, the cathode-coupled stage acts as a mixer. The local oscillator signal is injected by connecting a portion of the oscillator coil between the grid and ground of the second section of cathode-coupled stage. Normally the grid is tied directly to the ground.

These special miniature tubes perform well at frequencies up to 500 megacycles and higher. Consequently they will see wide application in television relay and high-frequency transmission. At these frequencies, and for that matter at lower frequencies too, it is well to utilize a push-pull arrangement (Fig. 51) to lower effective capacity and form a balanced high-frequency circuit.

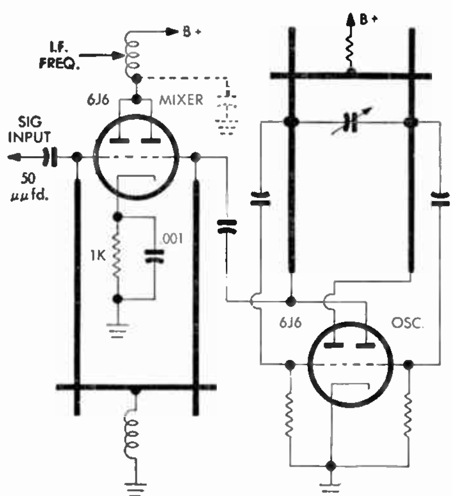


FIG. 51 Push-Pull Mixer-Oscillator

Many tuned circuits used in high-frequency television circuits are so-called linear tank circuits made of an effective section of a transmission line. For example, a quarter wavelength of transmission line shorted at the end will act as a parallel resonant circuit and can be attached as such to a vacuum tube circuit at its open end. One advantage of this type of a tuned circuit (Fig. 51) is that it can be made to resonate at the desired frequency with the distributed circuit capacity forming a high *Q*, high L-to-C ratio circuit. It can be designed to be a low-reactance tuned circuit and minimize the effects of lead inductance and circuit capacity. It also lends itself to switching systems as its resonant frequency can be changed readily by simply moving the position of the short at end of line (shorter the effective length of line between tube and short, the higher the resonant frequency). It is not only adaptable to push-pull circuits but is equally effective as a tank circuit for single-ended stages, replacing the usual form of coil and capacitor. Linear tuned circuit is not always in the form of straight rods but can be in form of a loop with a moving or switched shorting bar or made of small coils added incrementally to change frequency to some lower value. Concentric tank circuits are also adaptable.

47a. *Cascode Amplifier*

The development of the cascode amplifier and of special tubes, such as the 6BQ7 and the 6BK7, evolved from some of the special needs of the input circuits or the r-f tuner of the television receiver. A most important consideration is the signal-to-noise ratio of the tuner; assuming optimum design of the remainder of the receiver, this ratio and the useful sensitivity of the receiver are set at the r-f amplifier stage of the tuner. When a triode amplifier, which has the lowest noise figure, is used in the r-f input stage, the sensitivity of the tuner will permit reception of a television signal of a weak level with a low percentage of noise or snow in the picture. However, in television service, the triode r-f stage has a number of disadvantages when used in the usual grounded-cathode or grounded-grid amplifier connection. The triode r-f amplifier is more subject to interference and lack of stability than is the pentode. It also has a higher capacity between input and output circuits; consequently, the local oscillator signal from the tuner can leak through to the antenna system and cause interference on neighborhood receivers. In addition, the input impedance of the r-f stage varies substantially with changes of bias and, over the wide band of frequencies necessary, changes the termination characteristics presented to the antenna system. The higher impedance input of the pentode permits some input circuit-gain and better selectivity, which makes the input system less subject to interfering signals.

The cascode amplifier offers the advantages of pentode operation with the use of triode amplifiers and their resultant low noise figure. A basic cascode amplifier, Fig. 51a, consists of two triodes with the plate of the first triode direct-coupled to the cathode of the second. In effect, the first triode acts as a constant impedance input transformer, matching the incoming signal to the cathode circuit of the grounded-grid second stage. The second stage of the cascode system provides the bulk of the amplifier gain while the first triode establishes the signal-to-noise ratio of the received signal. The plate circuit of the first triode is shunted by the very low impedance (a few hundred ohms) of the cathode input of the grounded-grid amplifier. Consequently, the signal level can be never any higher than the amplitude of the incoming signal at the higher impedance input circuit of the first triode. Thus the signal amplitude is not sufficiently high to cause oscillation or instability. Nevertheless, some form of neutralization is often used, not so much to protect the stability of the cascode amplifier as to obtain a more uniform and lower noise-input circuit over the very wide range of television frequencies. Although the total distributed capacity contributed by the output capacity of the first stage, the wiring capacity, and the input capacity of the second stage is very, very low, its reactance can drop to a degenerative value on the high television frequencies. Thus there is a degenerative phase shift, and a signal can be fed back into the input circuit and affect its impedance characteristic adversely. This defect can be eliminated by feeding an out-of-phase component from the plate circuit of the second triode to cancel feedback at the input grid.

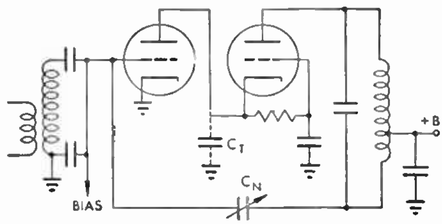


FIG. 51a. Basic Cathode Circuit

One advantage of the pentode stage is the shielding action of the screen grid which holds the capacity between grid and plate, or input and output circuits, to a minimum. Consequently, there is less opportunity to feed signal in either direction. When there is an easy path between input and output circuits, interfering signals can feed directly through the r-f stage and into the remainder of the receiver. Likewise, the local oscillator-signal component can find its way out into the antenna system. The cascode amplifier has the same low admittance qualities as the pentode stage, because the input-to-output capacities of the two triodes are effectively in series, thus reducing the total effective capacity. In addition, there is a low impedance to ground between stages which also has a shunting influence on spurious signals traveling in either direction.

Still another consideration in the design of a television tuner is the range of signal amplitudes it must be made to accommodate. A tuner must be able to receive, without overload and distortion, a very strong signal when the receiving site is near the transmitter station. Conversely, in a distant location, it is required to pick up an extremely weak signal and give full benefit of its low-noise factor to its reception. To enable a tuner to utilize signals of such great range in strength, a system of automatic biasing is employed in the usual television receiver. This bias, as a function of signal strength, applies a negative voltage to the r-f amplifier stage, reduces the gain for reception of a strong signal, and increases gain to permit reception of a very weak signal. During the process of changing bias it is essential that the input impedance presented to the antenna system be held essentially constant in order that it not mismatch the antenna system and create loss of resolution and/or reflections. Again the low impedance plate circuit of the first section and the small amount of neutralization keep the input impedance reasonably constant for wide variations in input stage bias. In addition, the direct-coupled connection between the two triode sections that are effectively in series with the plate supply causes the plate voltage of the first section to change with a shifting bias. Therefore, cutoff bias, itself, varies as a function of signal strength, and there is less danger of overloading and of resultant cross-modulation with reception of strong television signals. Thus the cascode amplifier has the same influence as a remote cutoff operating condition has, but without the usual change in input impedance that occurs with remote cutoff operation.

Special tubes such as the 6BQ7, 6BK7, and 6BZ7 have been constructed in such ways as to enhance the advantages of cascode amplification. Each of these tubes has been designed with minimum interelectrode capacity and with the grid mounted as near as possible to the cathode circuit in order to produce a high mutual conductance and, therefore, a peak gain from low impedance circuits. To keep a low admittance figure between input and output circuits two triodes have been mounted in a single envelope with excellent shielding between the two triode sections. However, the very nearness of the two triodes means that a single short jumper can connect one triode to the other, that is, from the plate of the first section to the cathode of the second section, and consequently, the distributed capacity is held to an absolute minimum. In fact, with this type of construction and the very low distributed capacity to ground, no neutralization is required for low television-band operation. Special construction of the control grid holds the plate-to-cathode capacitance to an average value of only 0.135 micromicrofarad. The low input capacity and the lack of degenerative feedback in cascode amplifier result in the input conductance of the r-f stage being held at a very low value. Consequently, some antenna circuit gain can be obtained, a voltage step-up existing between the low impedance antenna termination and the few thousand ohms input impedance of the grounded-grid section (high impedance input to the grid of the first triode). A typical RCA-developed cascode amplifier using the 6BQ7 is illustrated in Fig. 51b. This cascode amplifier consists of the tuned antenna input transformer with a special shield between windings, which minimizes direct capacity feed-through and reduces the possibility of interfering signals entering the tuner. The secondary is shunted by a loading resistor in order to keep the input impedance a bit more constant with changes in frequency and d-c bias. The compact construction permitted by use of a 6BQ7 makes it possible to avoid employment of external neutralization for low-band operation. On the high band even a very small amount of shunting-distributed capacity could affect the noise factor adversely by reflecting a degenerative component to the input grid or by adding some of the noise content of the second triode to the first one. This problem is overcome by employment of an effective series-resonant circuit consisting of inductor L and total distributed capacity C_t. This resonant circuit is tuned near the center of the high band and, consequently, sets up an extremely low impedance from the inter-stage coupling line to ground. Therefore, even on the high band, neutralization can be eliminated, the series-resonant circuit holding the coupling system to an extremely low high-band impedance.

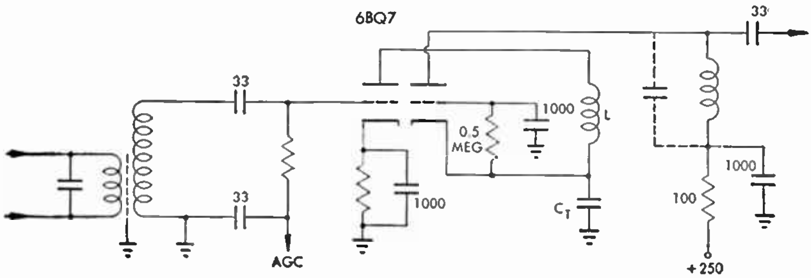


FIG. 51b Cascode Amplifier with No Neutralization

The first triode section employs cathode bias, while the second or grounded-grid amplifier section uses a small amount of contact bias that changes slightly with signal level but permits maximum gain-setting of the bias for grounded-grid amplification. However, the series connection develops a high signal voltage across capacitor C_t, which is the actual applied signal voltage for the grounded-grid amplifier stage. The low impedance input holds this voltage to a value no greater than the input grid voltage, thus eliminating the possibility of oscillation. A special cascode-bias modification has been developed by Philco for better accommodation of the wide range of possible signal strength to be received in a given locality. In fact, with the spread of television into new areas, practically every community is expected to receive a very strong local signal and, at the same time, weak long-distance signals. It was found in the usual cascode circuit, with the a-g-c bias applied to the first triode grid, that upon reception of a very strong signal there was still the possibility of overload and cross-modulation. This results from the fact that the first triode has very little gain; consequently, a change in its bias does not appreciably affect the over-all gain of the cascode amplifier. However, by changing the bias of the second tube, which is really the gain-stage, a better range of gain-

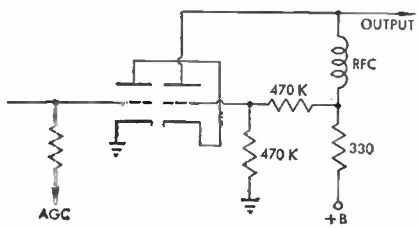


FIG. 51c Philco Bias Modification

shift can be obtained and can accommodate a wider range of input signal levels. In the circuit, Fig. 51c, the a-g-c voltage is applied to the first grid; however, a special bleeder-bias network has been connected to the grid of the second section; thus, the voltage from grid to ground is held constant, regardless of any changing amplifier gain and a-g-c bias. However, a change in the a-g-c bias on the first grid immediately changes the plate voltage of the first section and thus the cathode voltage of the second section. With the grid voltage of this second section being held constant, the appreciable change in the cathode voltage of the second section then causes a great change in the gain of the second stage or grounded-grid amplifier. Thus the change in a-g-c bias has a much more decided influence on the over-all gain of the amplifier and permits reception of a very weak signal with peak sensitivity and, at the same time, reception of a very strong signal that requires only a low gain if overload and possible distortion of the picture are to be prevented.

48. *Receiver R-F Sections*

With the advent of miniature tubes, the r-f section of the television receiver, consisting of r-f amplifier, mixer, and oscillator, has become a compact, efficient assembly. The entire r-f section of the new receivers is mounted on a highly stable turret assembly or multisection selector switch which serves as a band-switching unit for the 12 television channels. All circuit components are a part of the assembly except signal, heater, and supply voltage lines. Needless to say, the small size of the miniature tubes makes the unit all the more compact. On this unit are generally three switch wafers or tuned sections which switch the r-f amplifier, mixer, and oscillator inductors channel by channel. Other systems employ continuous tuning, permeability tuning, or printed-circuit technique.

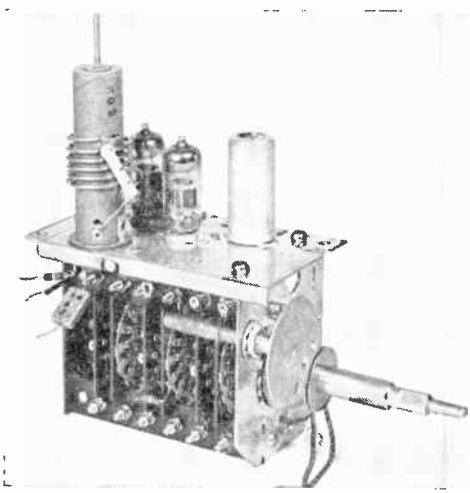


FIG. 52 RCA Selector Tuner

The r-f selector tuner of an RCA model is shown in Fig. 52; a Farnsworth version, in Fig. 53. Schematic diagrams of same units are given in Figs. 54 and 55. The RCA unit is a push-pull r-f amplifier, mixer, and oscillator arrangement using transmission-line sections as parallel resonant circuits. A grounded-grid r-f amplifier, pentode mixer, and triode oscillator are used in the Farnsworth assembly.

One of the most ingenious tuning systems has been designed by Paul Weir of the DuMont Laboratories. This system consists of a cathode input stage, mixer, and local oscillator and is continuously tunable from 44 to 216 megacycles. It covers television, FM, and other bands. The inductuner, as used for television, consists of a three-section variable inductance arrangement mounted in a die-cast housing. The three coils are mounted on a ball-bearing

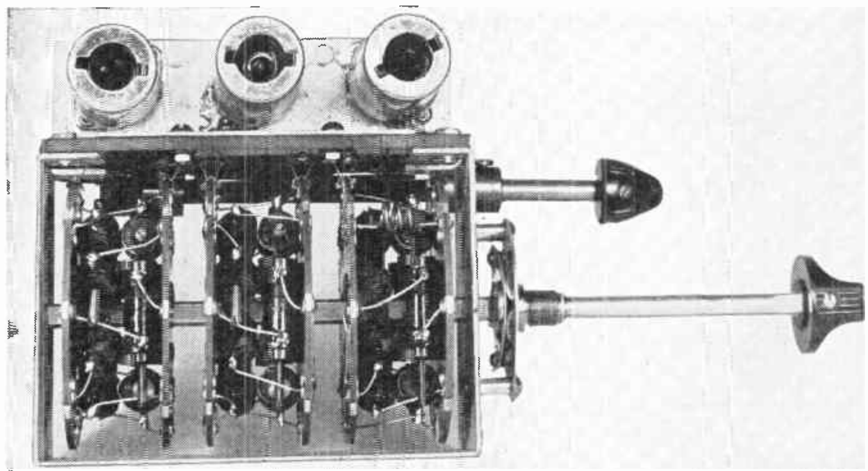


FIG. 53 Farnsworth R-F Tuner

shaft and trolley arrangement (Fig. 56). The coil windings move through a trolley contact which divides each coil into a used and unused portion. An unusual and advantageous feature of the inductuner is that the *Q* of the tuned circuits increases toward the high-frequency end. A schematic diagram of an inductuner r-f assembly, as incorporated in a DuMont receiver, is also shown (Fig. 57).

A simple, basic schematic of one section of an inductuner is shown in Fig. 58. Coil *L1* is a variable inductor, and the contact effectively moves from the low-frequency to the high-frequency end of the coil. The contact is shorted to the low-frequency end of the coil, keeping unused section of the coil resonant at a very high frequency. The high-frequency limit of the tuner is set by the fixed end inductor *L1* and the circuit capacity needs. Consequently, the tuned circuit can be designed for maximum efficiency and proper bandwidth at the high-frequency end of the bandpass. At this high-frequency end of the desired bandpass, the top resonant frequency of the unused portion of the coil is still high and does not interfere with the operation of the tuned circuit. Thus, on this account, the resonant circuit of the inductuner is varied by means of a tap along the coil, and the unused portion of the coil does not affect the characteristics of the tuned circuit.

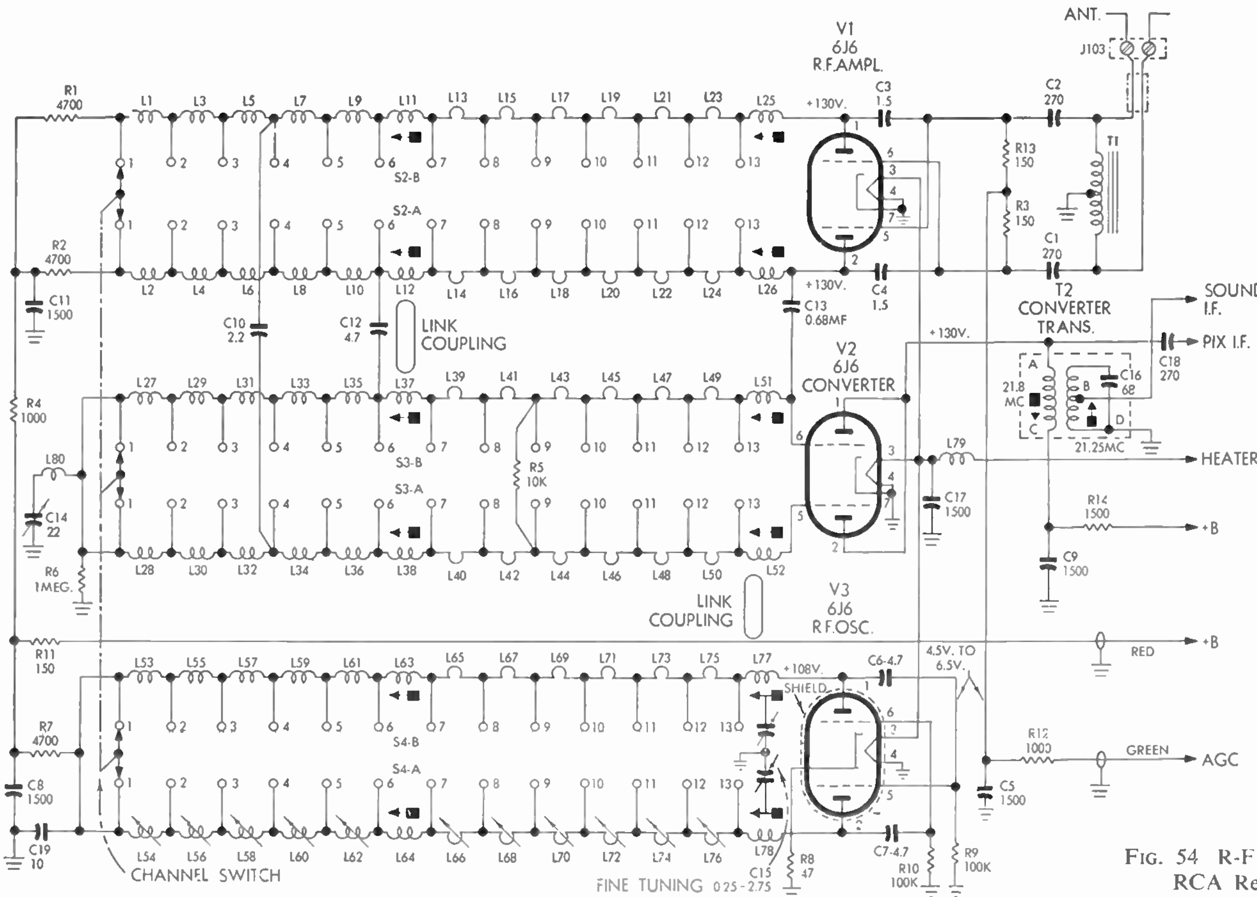


FIG. 54 R-F Section, RCA Receiver

It is also possible to construct an overcoupled transformer arrangement, as shown in Fig. 59. This overcoupled transformer consists of two variable coils and, of course, the two fixed end coils. To obtain the proper degree of overcoupling over the bandwidth desired, it is sometimes necessary to use a combination of inductive and capacitive mutual coupling between the tuned circuits. Consequently, as shown, overcoupling occurs between the two end inductors and also by means of the common capacitor $C3$.

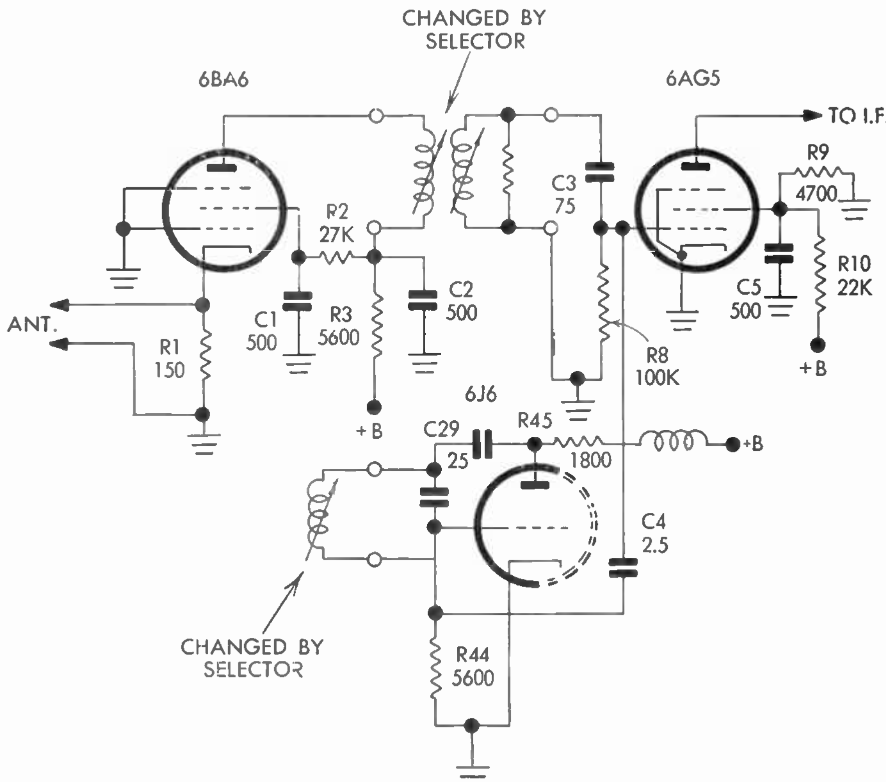


FIG. 55 R-F Section, Farnsworth Receiver

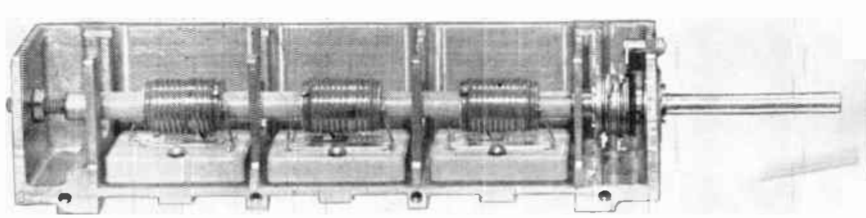


FIG. 56 Mallory Inductuner

Still other receivers such as Silvertone use an entirely separate high- and low-band 3-stage tuner to obtain good efficiency at widely separated frequencies. In Bendix tuners a single but separate high- and low-band resonant circuit is used at each position a tuned circuit is needed (antenna input, mixer, and local oscillator). Individual channel tuning is accomplished by

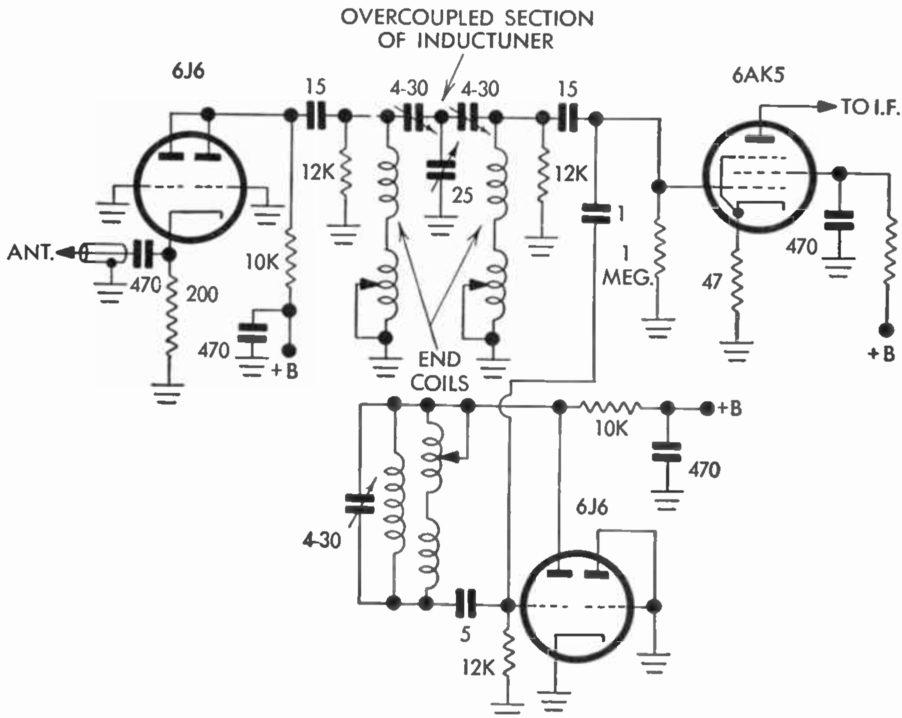


FIG. 57 DuMont Inductuner Circuit

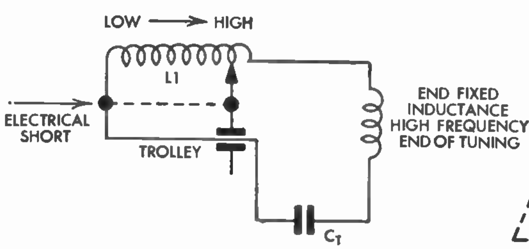


FIG. 58 Basic Inductuner Circuit

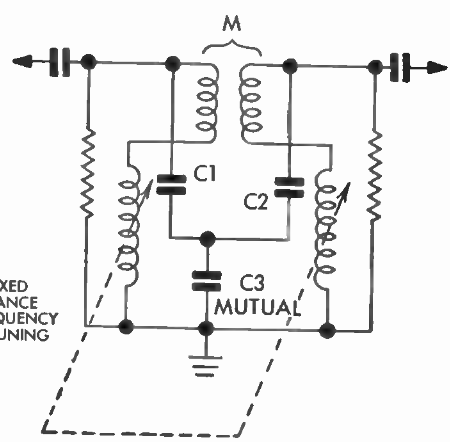


FIG. 59 Overcoupled Inductuner

position of a permeability tuning sleeve with relation to the main inductors. The advantage of such a system is no moving or switched contacts except when switching between high- and low-band stations.

49. *R-F Amplifier and Mixer-Oscillator Characteristics*

The three major functions of the r-f amplifier of the r-f section are: (1) to increase receiver sensitivity; (2) to improve signal-to-noise ratio; and (3) to prevent local oscillator radiation. A high-gain r-f stage increases receiver sensitivity; however, it must be remembered that in television bandwidth is an equally important consideration, and tubes and circuits must be chosen which produce an acceptable gain over a broad band of frequencies.

One source of interference when television receivers are grouped is local oscillator radiation off the receiver antennas. These locally generated signals feed into receivers set on other channels and destroy the resolution of the picture. The presence of an efficient r-f stage isolates to a limited extent the local oscillator from the antenna system.

The mixer-oscillator system of the present television receiver is carefully designed to convert r-f to i-f with peak efficiency. When the antenna and transmission line are properly installed in a reasonably noise-free location, the primary source of noise is the r-f amplifier or the mixer oscillator. An appreciable noise in the television receiver is the mixer-conversion noise. But the greater the signal on the grid of the mixer and the more efficiently the converter utilizes this signal, the more the signal will rise above the noise.

Conversion conductance is a measure of how effective a tube is in converting a grid-voltage change at signal frequency to a plate-current change at the i-f frequency. This conversion conductance is always less than the actual mutual conductance of the tube. Nevertheless, conversion efficiency can be made ample by holding up the L-to-C ratio of the grid-input and plate-output tuned circuits and by injecting the local oscillations from a separate local oscillator directly into the signal control grid of the mixer. In many television receivers triodes are used as mixers because of their lower noise characteristics.

Some of the features of mixer-oscillator systems for television are:

1. Bandswitching. Inasmuch as proper choice of parts value and the mechanical arrangement of the tuned circuits are so very critical in obtaining the most gain at the required bandwidth, one set of operating constants will suffice as an efficient combination at only one frequency. Consequently changing channels is done by means of bandswitching; often an entire new set of constants is substituted for each channel. Thus the r-f section operates at optimum efficiency on each channel. This system, of course, affords a very convenient method—push buttons or rotary switches—for the televiewer to change channels. In some receivers, a high-efficiency continuous-tuning system is used which tunes over 12 television channels in one or two bands.

2. A small variable capacitor is used for fine tuning, varying the local oscillator frequency over a limited range (less than 1 megacycle). Inasmuch as the r-f amplifier and mixer stages are very broadly tuned, it is necessary to vary only the local oscillator frequency to set the picture- and sound-carrier frequencies at the proper point on the i-f bandpass characteristic.

3. Push-pull amplifier mixer and oscillator stages are sometimes used because of their high-frequency efficiency. These stages employ linear tank circuits which can be conveniently tuned by shorting bar arrangement.

4. In many r-f transformer circuits the resonant circuits are tuned by varying the resonant circuit inductance with a movable metal core. This method does not require variable capacitors, which would add capacity to the circuit and lower the L-to-C ratio.

5. Use of tubes which have a high conversion conductance, particularly miniature tubes with low capacities. Separate oscillator tubes used with grid injection of mixer tube to raise conversion efficiency (more i-f signal per given signal on mixer grid).

6. Use of special resonant circuits and filters to reduce sensitivity of r-f sections to i-f and other low frequencies. These filters and resonant circuits insure maximum transfer of desired signals and maximum rejection of undesired signals. Thus, high-gain i-f systems can be used without danger of feedback or interference.

7. Relatively low-value grid resistors or other loading resistors are often used to load the tuned circuits somewhat and to flatten the bandpass characteristic.

The local oscillator tube also has high mutual conductance and low capacities to oscillate efficiently and with stability at the high frequencies necessary. Tubes such as the 6J6, 6C4, 6J5, and 7A4 are used as oscillators and are connected in modified Hartley or ultra-audion circuits. When switching channels, the local oscillator must also be changed. This is done in most receivers by adding or subtracting sections of inductance from the tuned circuit. In some few receivers it is accomplished by switching actual trimmer capacitors. Each oscillator-tuned circuit, corresponding to a certain position of channel switch, is tuned by means of a movable core or small trimmer capacitor. A small variable capacitor called a *fine-tuning control* is used for precise setting of the local oscillator frequency on each channel. By means of this control, which changes only the oscillator frequency, the picture and sound carriers are properly located on the i-f band-pass characteristics. Thus the exact i-f frequencies are produced in the mixer output. It is not necessary to vary the other tuned circuits for this fine adjustment, because they are inherently broad with respect to the small frequency variation of the local oscillator necessary to put the i-f picture and sound carriers on frequency. The i-f frequencies of television receivers are in the 20-to-50-megacycle or higher frequency regions.

Local oscillator is tuned to the high-frequency side of the picture and sound carriers. For example, when set on channel 4 many commercial receiver local oscillators are tuned to 93 megacycles. Two difference frequency components exist:

Local oscillator frequency, 93 megacycles, minus picture-carrier frequency, 67¼ megacycles, equals a picture i-f frequency of 25¾ megacycles.

Local oscillator frequency, 93 megacycles, minus sound-carrier frequency, 71¾ megacycles, equals a sound i-f frequency of 21¼ megacycles.

It is apparent, therefore, that the mixer of the television receiver functions as a dual mixer because it acts as a mixer for the picture and sound signals. Consequently, because of this dual mixing and the spacing between picture- and sound-carrier frequencies, two i-f frequencies are produced with a single local oscillator.

50. *Antenna Matching Input Circuits*

To obtain the highest gain and the best signal-to-noise ratio, it is necessary to match the antenna and transmission line system precisely to the input of the r-f amplifier or, in the case in which the antenna excites the mixer, to the mixer grid. Consequently, the input transformer should match the high impedance of the input circuit of the tube. One simple method of matching antenna to grid-input circuit is shown in A, Fig. 60. Here the antenna is

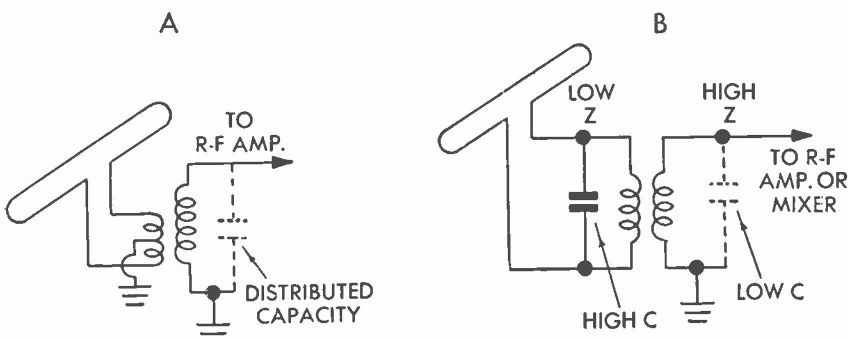


FIG. 60 Antenna Matching Methods

coupled to a low-impedance portion of the grid-input circuit through a few turns of wire located near the ground end of the tuned circuit. A higher gain method is shown in B, which utilizes a tuned primary and tuned secondary. In this system the impedance match is a function of the L-to-C ratios of the primary and secondary tuned circuits. The secondary has a high L-to-C ratio and, therefore, a high impedance to match the grid circuit of the tube, while the primary has a low L-to-C ratio and, consequently, a low impedance, which matches the antenna and transmission-line system. Thus, the secondary is resonated to the proper frequency by means of the distributed capacity of the circuit, which, of course, is very low. The primary has an actual capacitor shunted across it. This capacitor has a substantial capacity and lowers the L-to-C ratio of the primary. In switching bands, both the primary and secondary tuned circuits are replaced by other sets of primary and secondary tuned circuits.

A third method is employed in other receivers (Fig. 61). This system has still higher gain and consists of two mutual couplings. First, the small antenna winding is coupled to the primary tuned circuit by means of a few turns located at the low impedance end of the primary tuned circuit. Primary and

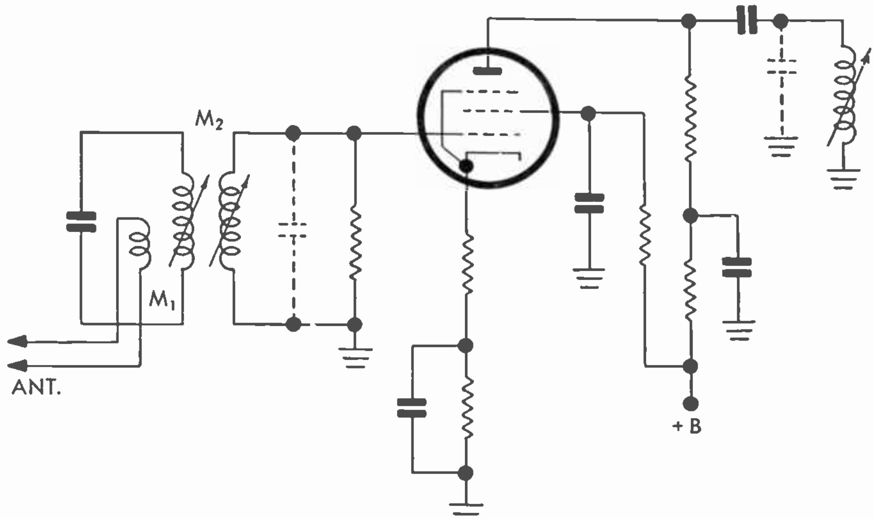


FIG. 61 Pentode R-F Amplifier

secondary windings are then overcoupled to obtain the double-humped broadband characteristic. Consequently, the coupling between the untuned antenna winding and the primary constitutes the impedance match, and the coupling between the primary and secondary windings is a control of the bandwidth of the tuned circuits and is overcoupled.

In the RCA, Farnsworth, and DuMont r-f assemblies, the antenna system is terminated across actual resistors. The RCA input termination (Fig. 54) uses two 150-ohm resistors in series; the Farnsworth receiver uses a single 150-ohm cathode resistor (Fig. 55) shunted to 90 ohms because of its presence in the grid and plate circuits of the grounded-grid stage.

51. *Tuner Features*

The responsibilities of the television tuner are many, amplification and conversion being the two most obvious, Fig. 62. A tuner must have the required bandwidth if all components of the incoming signal are to be amplified uniformly. At the same time, the edges of the bandpass must fall off as rapidly as possible, thus reducing interference from adjacent channels, Fig. 63. A satisfactory tuner is also expected to have many additional features, such as high signal-to-noise ratio, good image rejection, freedom from second harmonic mixing, weak local oscillator radiation, a minimum of direct i-f feed-through, and positive durable switching.

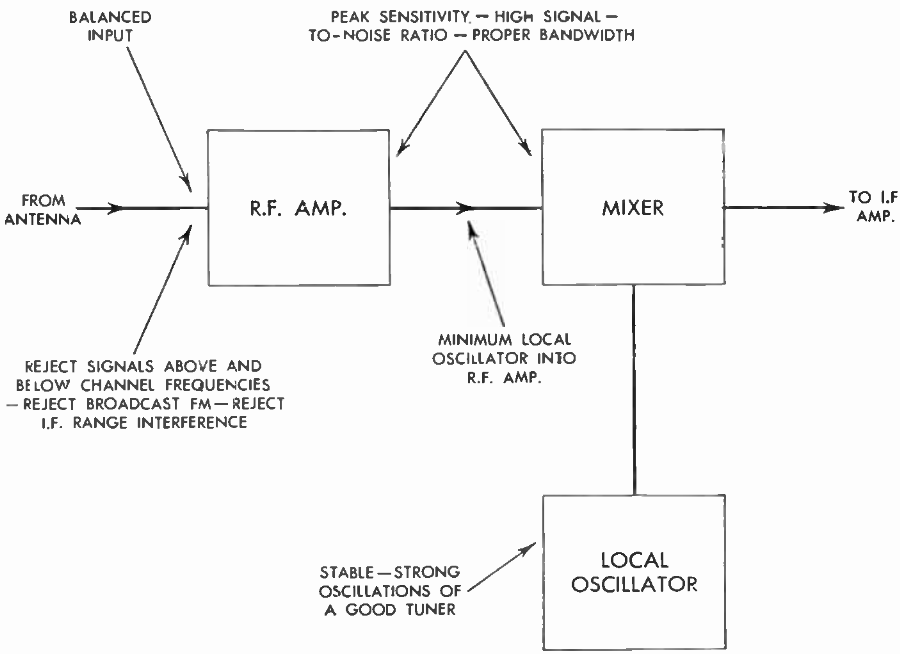


Fig. 62 Features of a Good Tuner

The higher the gain of the tuner per required bandwidth and noise content, the more sensitive the tuner is said to be. Not only should the tuner have an essentially constant gain over each channel (constant gain means a uniform bandwidth), but gain should not vary appreciably from one channel to another. In many tuners, however, one can expect to find less high-band sensitivity. Gains average between 20 and 30 over the television channels (voltage delivered to i-f input, as compared to signal applied to input antenna terminals of tuner).

If true picture resolution and proper sound level are to be retained, the bandwidth must extend over the 6-megacycle channel. In order to render an image truthfully, the response must be essentially flat over this 6-megacycle span, no more than a 30-per cent dip or valley being permitted at the center of the response curve. With an ideal response the gain would fall off abruptly at the "end" of the channel. However, practical design and the cost factor do not permit such a sharp cut-off, and the response is permitted to taper off.

Such a slow decline in response means the tuner has greater sensitivity to adjacent channel signals.

A tuner must add an absolute minimum of noise to an incoming signal. A low noise level indicates a good signal that will produce a clean picture with a motionless background. Even a weak signal, if it is not ridden with input circuit noise, can produce a usable picture. When the tuner's noise level is high the background of the picture becomes shaky and unstable; signal-to-noise ratio is, therefore, a very important factor in fringe-area reception.

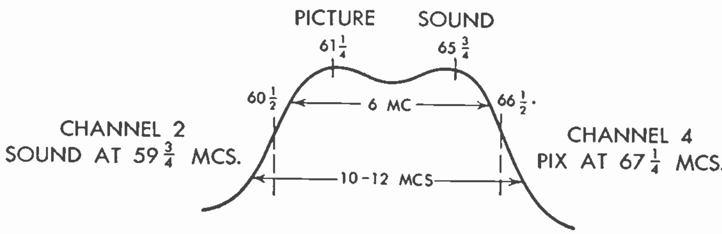


FIG. 63 Channel *3* Response of a Tuner

Interference rejection is a trying problem when a unit functions with a wide acceptance band and little selectivity. Tuners must be planned and designed carefully if they are to minimize stray signal pick-up. The frequency ranges at which a receiver is particularly subject to interference are as follows:

1. On channel
2. Adjacent channel's sound and picture
3. Image interference, consisting of local oscillator plus signal frequencies, local oscillator plus i-f frequency, and twice local oscillator plus and minus i-f frequency
4. Direct i-f feed-through
5. Local oscillator radiation

RESPONSE CURVE AND ITS SIGNIFICANCE

The response curve discloses much about the performance of a tuner on each channel. The curves are not identical on each channel as might be expected, because certain departures from the ideal flat curve are permitted in the interest of economical design. There are specific limits to these departures that can be checked on response curve of each individual channel. The variations are in the form of bandwidth, tilt, and center valley or dip, Figs. 64a and 64b; as the receiver is switched from channel to channel, the response curve changes slightly but must not be permitted to exceed specific limits.

Each of the individual tuned circuits does not have a linear flat response. Instead, each contributes a certain segment of the desired and over-all response. Several possible arrangements are illustrated and can be classified in a number of basic approaches. The most common arrangement is that of a single-tuned input circuit and an overcoupled double-tuned circuit between r-f plate and grid of mixer. The over-all response is formed by the combination of the two curves—input tuned circuit having more influence on the center of the curve, overcoupled resonant circuits on the sides and skirts of the over-all response. Usually a special bandwidth control is present in the overcoupled transformer to permit optimum adjustment per bandwidth and steepest skirts. The sharper input response adds selectivity and permits establishment of a higher signal-to-noise ratio at the r-f amplifier input circuit, allowing a better noise factor for a given over-all bandwidth. If two r-f stages are used, the individual resonant circuits can be staggered in order to give the desired over-all response, and on occasion, an input tuned circuit is used that is either broadly resonant or overcoupled.

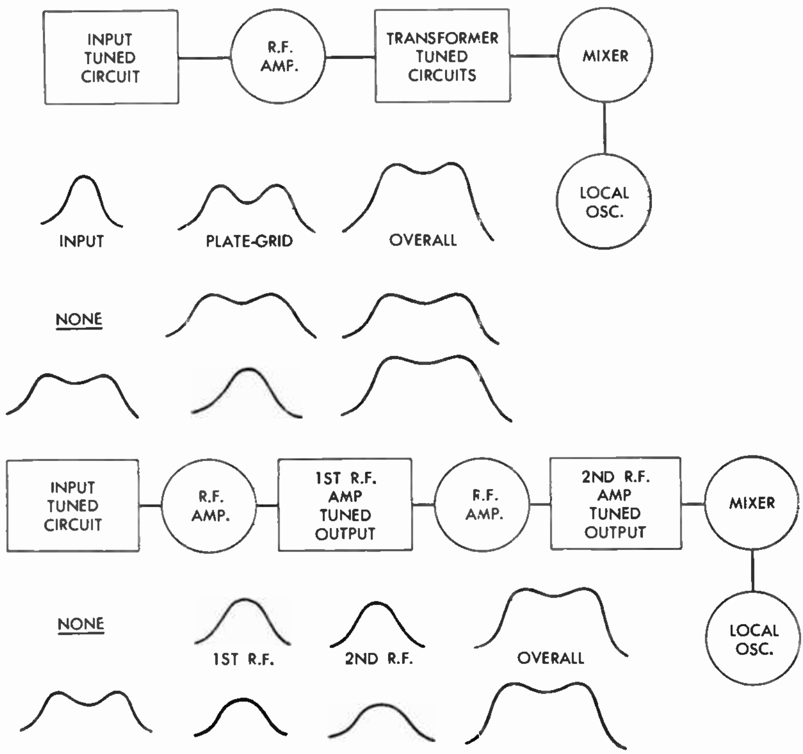


FIG. 64a

Usually the bandwidth of a tuner on a given channel is a function of the amount of overcoupling between the r-f plate and mixer grid-tuned circuits. A specific circuit element, or the proximity of two tuned circuits, sets this bandwidth individually on the high- and low-band channel groups. However, at the low-frequency end of either band the ratio of carrier frequency to bandwidth is smallest; for a fixed setting of any bandwidth control, it means a narrower bandwidth is obtained at the low-frequency end than at the high-frequency end of the specific channel group; thus it is more difficult to obtain necessary bandwidth at the low end of channel groups (channels *2, 3, 7,* and *8*). A practical limit would place carriers no lower than 80 per cent levels on the response drop-off (the sound carrier of certain intercarrier models can be below this level) as per the first curve of Fig. 64b. At high-end channels (*6, 12, 13*) the bandwidth tends to be too broad, widening the noise- and interference-acceptance bandwidth. The 50-per cent down points should certainly not be separated by more than 10 or 12 megacycles as per second curve.

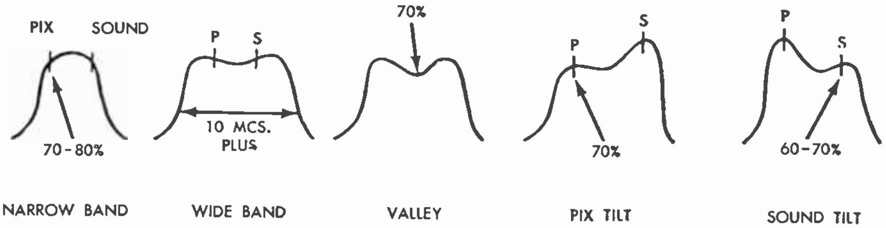


Fig. 64b Tuner Response Limits

It is difficult to obtain optimum loading on all channels, because resistive damping is also a function of bandwidth and frequency of operation. Thus, response curves are not flat on all channels, and practical bandwidth-setting to cover such a wide range of frequencies often places a valley between over-coupled humps. It is recommended that this valley dip no lower than a 70 per cent level. Still another response defect, because of the inability to retain optimum loading and tuning on each channel, is the tilt shown on the latter drawings. Tilt must not be permitted to drop below a 70-per cent level for either carrier, Fig. 64b.

52. *Commercial Tuners*

The RCA input elevator transformer consists of two 150-ohm sections of effective transmission lines, formed by interlacing two windings on a coil-form, Fig. 64c. Impedance of 150 ohms is obtained with proper spacing and diameter of wires. For a 300-ohm termination, the one coil is attached to each leg

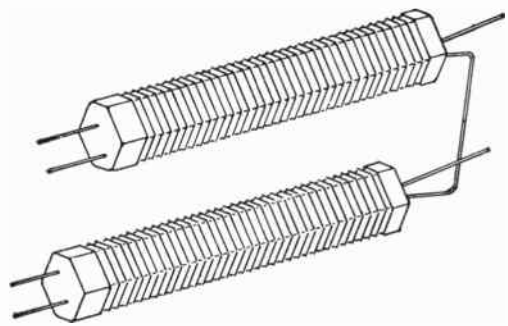


Fig. 64c Coil Plan of RCA Elevator Transformer

of the line—the two sections being in series across the line to form a 300-ohm termination and to act as a balance-to-unbalance transformer, Fig. 64d. The resulting advantages of the transformer-type windings for these lines are the more uniform impedance over a wide range of channels and the ability to shunt unbalanced and spurious signals on the line which are attempting to penetrate into the tuner. The desired resonant and balanced signals find a proper termination and pass unimpeded into the tuner (coils act as a continuation of the line), while unbalanced signals meet a high impedance and are shunted to the ground by the transformer-action.

In the 75-ohm position, the two transmission-line sections (coils) of 150 ohms are connected in parallel to form a 75-ohm termination for a coaxial-type transmission line.

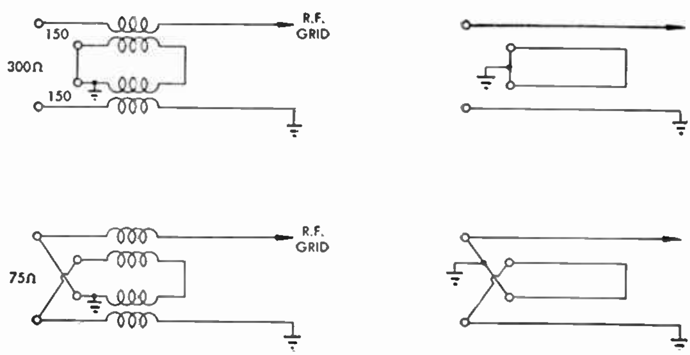


Fig. 64d Elevator Transformer Action

The RCA tuner, shown schematically in Fig. 64e, uses a cascode r-f amplifier and a triode oscillator and pentode mixer. To obtain satisfactory sensitivity and bandwidth, four separate resonant circuits are used in the tuner, as demonstrated in the simplified schematic of Fig. 64f. There is a tuned-grid circuit for the first triode section of the amplifier, a tuned-plate circuit for the second triode of the r-f stage, a mixer resonant circuit, and a local oscillator-tuned circuit. The inductances that comprise these resonant circuits are mounted around the periphery of the various selector-switch sections. As the selector switch is turned from the high end to the low end of the VHF band, inductances are progressively added to the original inductance, each addition lowering the frequency of resonance. At the same time, proper selectivity and bandwidth must be retained as the frequency changes progress from the high to the low end of the VHF spectrum.

The antenna system is attached to the elevator transformer input circuit, which also includes a special high-pass filter that rejects possible lower-frequency interference (particularly interference in the i-f frequency range). It is significant that special parallel resonant traps form a maximum impedance in series with the signal path, rejecting in particular the sound i-f frequency of 41¼ megacycles and the picture i-f frequency of 45¾ megacycles. Consequently, possible feedback from the video i-f strip that could cause oscillation or regeneration is eliminated at this point. If such interference did exist it would appear most often as a pattern of thin vertical bars or repeats on the picture. Interference is also possible from the harmonics of certain frequencies in the i-f range (particularly harmonics that originate in the non-linear detector

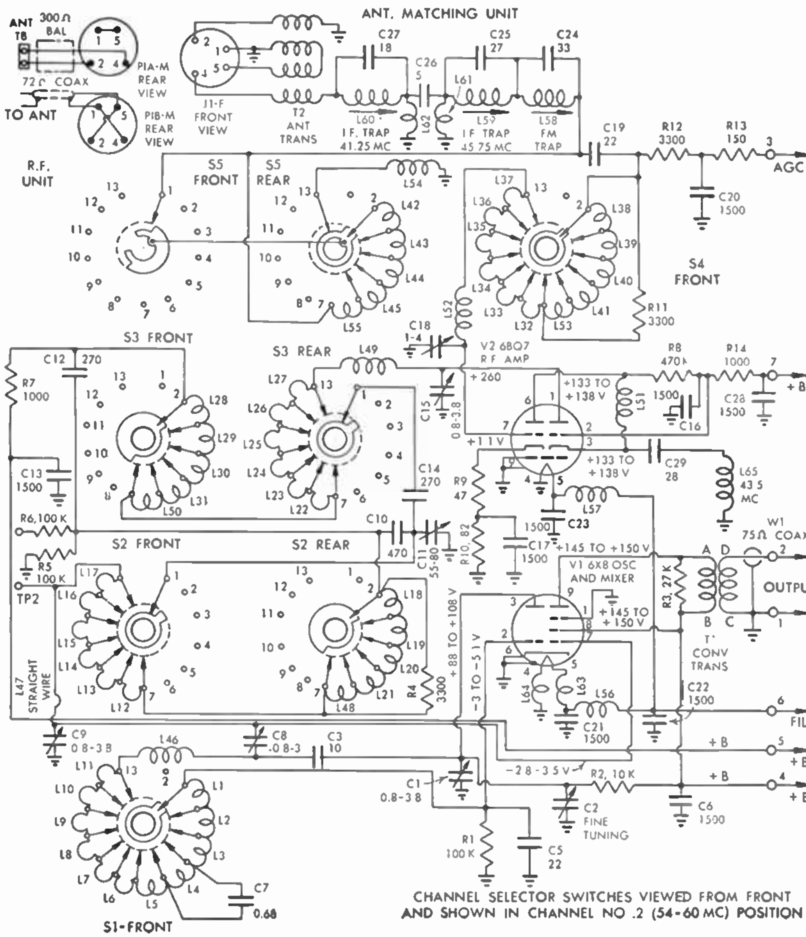


FIG. 64c RCA Cascode Tuner

stages, such as the video detector or the FM detector), which can also feed back into the input circuit and set up spurious beats. An FM trap is also included in order to minimize possible interference from local FM-broadcast stations. The inductors associated with these traps are adjustable, which permits them to be peaked to exact frequencies. Likewise, inductors *L61* and *L62* can be set accurately to maintain the proper response curve, thus enabling the filter to obtain maximum rejection in the i-f range and, at the same time, to offer proper sensitivity to channel *2* and *3* signals. A single input inductor *L54* keeps the input impedance characteristics reasonably constant over the entire high band. However, to keep impedance characteristics more consistent, incremental coils are added to *L54* as the channel selector is rotated from channel *7* back to channel *2*. The signal from the antenna-matching unit is conveyed through capacitor *C19* to the series-tuned input resonant circuit of the r-f amplifier stage. Proper bandwidth on the low band is obtained by using shunting capacitor *R11*. The plate of the first section of the r-f stage is direct-coupled through inductor *L51* to the cathode of the grounded-grid second section. Also to obtain maximum rejection of i-f range interfering signals, a special series trap, tunable with *L65*, is included. No direct feedback is employed in the cascode amplifier, but inductor *L51*, together with distributed circuit capacity, minimizes phase shift; the partially degenerative cathode circuit of the input r-f stage keeps the input impedance to the cascode amplifier essentially constant over the necessary frequency range.

A double-tuned resonant circuit couples the r-f amplifier plate to the mixer grid, with a mutual capacity element controlling bandwidth and selectivity—capacitor *C11* on the high band and the series combination of *C10* and *C11* on the low-band channels. An a-g-c bias is applied to the grid of the first section of the r-f amplifier. In the group of resonant circuits discussed, the alignment capacitor *C18* has the greatest influence on the maximum amplitude and sensitivity of the response curve; capacitor *C15* tunes the r-f amplifier plate circuit and affects the frequency of the bandpass more noticeably than do the other resonant circuits. Capacitor *C19* tunes the mixer grid circuit and influences the tilt of the curve, while capacitor *C11* is the bandwidth adjustment and primarily affects the bandwidth of the tuner-response curve. The above controls affect the response of the tuner, in particular on the high-band channels. A group of tunable inductors, *L48, L50*, and *L53*, control the bandwidth and sensitivity of the tuner response on the low-band channels.

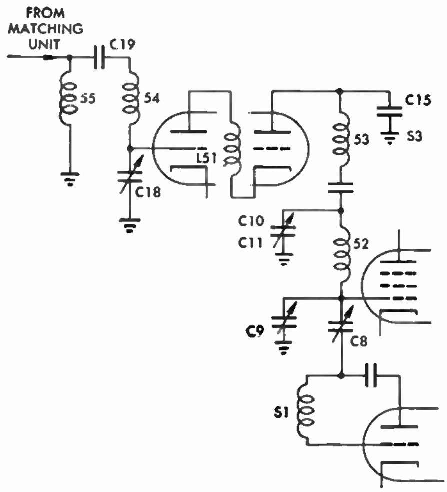


FIG. 64f Simplified Plan of Tuner

A combination pentode mixer and triode oscillator has been made possible by the development of the 6X8 converter tube. The pentode mixer provides increased conversion-gain and, because of the low interelectrode capacity between plate and grid circuit, minimizes local oscillator feed-through into the i-f strip of the receiver. An ultra-audion local oscillator circuit is employed with various small capacitors that must be aligned to obtain proper tracking over the entire VHF band. The difference frequency, after mixing action, is developed across the resonant circuit at the plate of the mixer and is link-coupled to the input circuit of the video i-f amplifier.

The Philco tapered line input is similar in some respects to the elevator-type transformer. However, the interlaced windings are not spaced equidistantly on each coil but have a progressively greater separation as they approach the tuner side. Consequently, the impedance of the effective line section rises by a factor of approximately two, there being a voltage step-up to the grid of the r-f amplifier stage, Fig. 64g. Again the coiled line sections are tied in series to match 300-ohm line and in parallel for 75-ohm line.

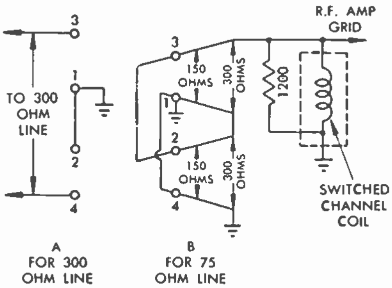


FIG. 64g Philco Tapered-Line Input

There are two basic tapered lines. One functions as a continuation of each leg of the transmission line and transfers maximum signal to tuner input. A second method, Fig. 64h, attaches the tuner input circuit to the windings of each coil which have no direct connection to the antenna terminals. Consequently, desired signal is transferred by transformer action. Some slight transfer-loss exists, but there is more thorough rejection of unbalanced spurious signals on the line.

The Philco tuner employs a cascode input stage using a 6BZ7, Fig. 64h, which is a further improvement on the cascode type; it has a higher G_m and a lower noise factor in addition to very thorough shielding between triodes and consequent higher impedance between input and output of cascode amplifier. To obtain the best noise factor and the least loading by transit time, a small capacitor *C505* neutralizes the input stage. It is connected from the plate of the first section to the ground side of the input resonant circuit and, along with *C504*, develops the necessary out-of-phase voltage to cancel the feedback voltage that is coupled through the interelectrode capacities of the amplifier tube. Inductor *L512* and the input capacity of the grounded-grid amplifier (second section of the cascode amplifier) are series resonant in the high band and, consequently, minimize the neutralization problem. In addition, the series-resonant condition assures that a high signal voltage is developed across the input capacitor of the grounded-grid stage, thus permitting maximum transfer of signal level. A-G-C bias is applied to the grid of the first section and a constant bleeder bias to the grid of the second section to insure maximum a-g-c action. This bias point is filtered for r-f variations as well as for power supply ripple.

The cascode amplifier output is coupled to the grid of the mixer via a double-tuned transformer, the bandwidth of which is controlled by various small capacitors that are switched into the circuit from channel to channel and which are connected across the high side of the resonant circuits. Incremental inductances are switched into the circuit by the turret progressively when the receiver is switched from channel *13* to channel *2*. There are four such incremental coil combinations, one for the cascode amplifier, two for the interstage coupling system, and a final one for the local oscillator switching.

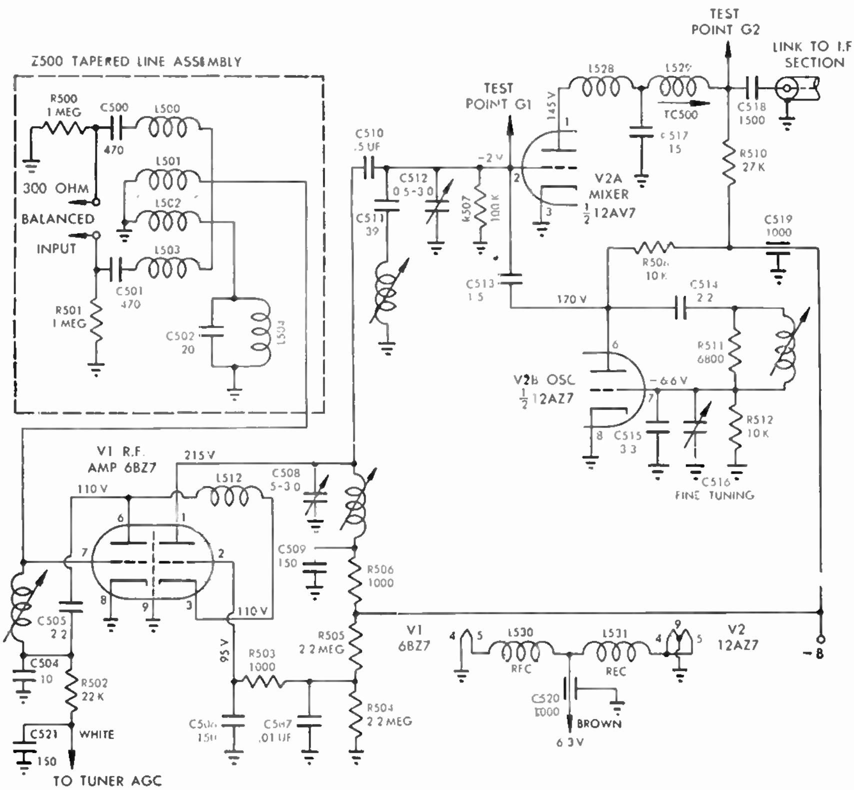


FIG. 64h Philco Cascode Tuner

A dual triode is used for the mixer local-oscillator combination. The circuit employs grid injection from the ultra-audion local oscillator via the small capacitor *C513*. There are convenient test points at the grid of the mixer and at the plate circuit output of the mixer for alignment purposes. A low-pass filter type of tuned circuit is used at the mixer output, a resonant circuit being formed by capacitor *C517*, inductor *L529*, and the input capacity of the first intermediate-frequency amplifier stage. This technique permits low-impedance coupling through a coaxial cable between major sections of the receiver, in this case, coupling the mixer output to the input of the i-f amplifier. Maximum signal is transferred, regardless of low-impedance path, because of the series current and maximum voltage developed across the input capacity of the first i-f amplifier stage.

The General Electric tuner consists of a triode first r-f stage, a pentode second r-f stage, and a dual triode mixer oscillator, Fig. 64i. The 6AB4 is connected as a grounded-grid amplifier and has an excellent signal-to-noise ratio

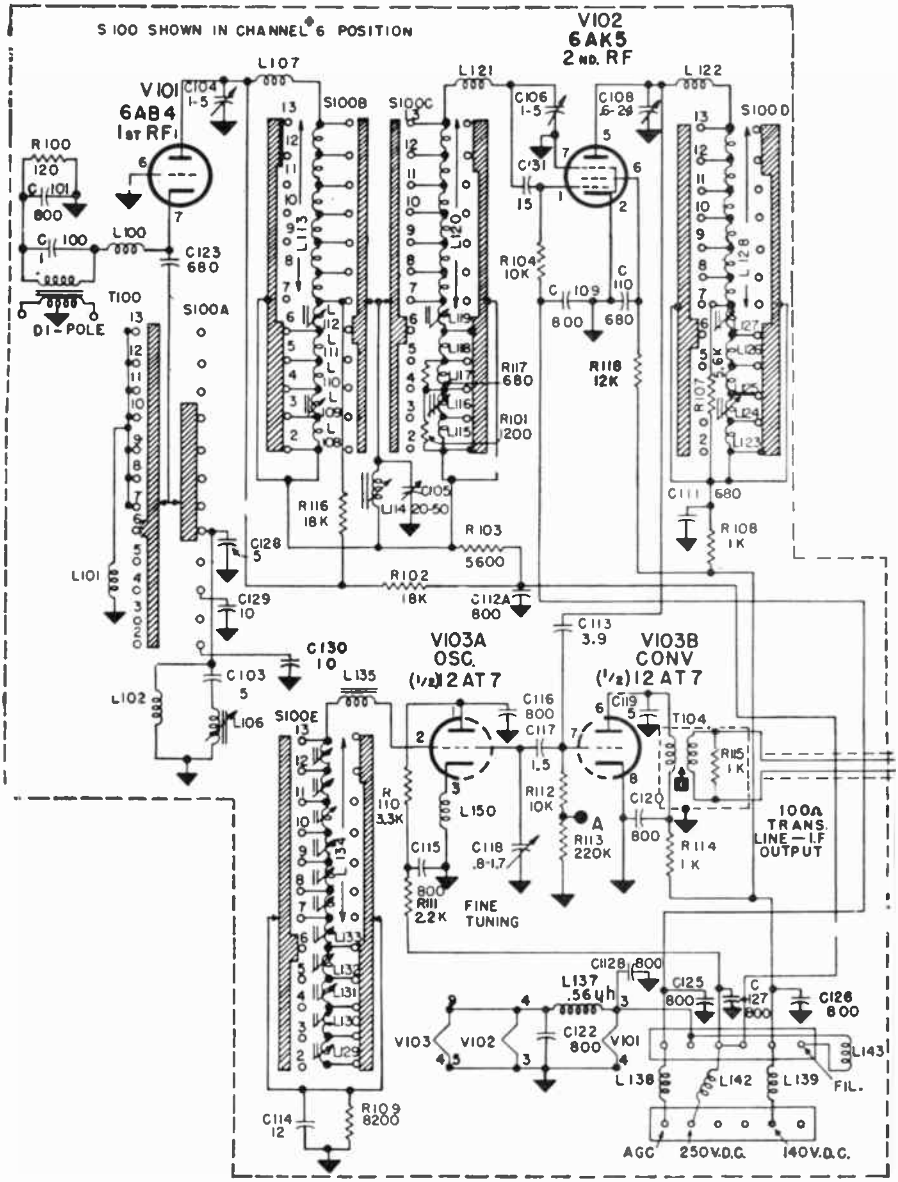


FIG. 64i General Electric Tuner

in high-frequency operation. The antenna system is matched through an input transformer to the cathode circuit of the grounded-grid stage with the low-pass input combination of *C100, L100,* and effective series capacity from cathode to ground; a suitable match is hereby established to the antenna system from the cathode of the grounded-grid amplifier. On the low-frequency channels, sections of inductance are added progressively to the low side of *C123* to retain proper impedance-matching over the low-band channels. The combination of capacitor *C103* and inductors *L106* and *L102* is tuned to the i-f frequency range and rejects signals attempting to enter the tuner in the 45-megacycle i-f frequency range. This precaution is necessary for low-band channel operation, because the desired signal-frequency range comes nearer and nearer to the i-f frequency range.

A double-tuned and overcoupled transformer, consisting of incrementive coil sections, conveys the signal between the first and second r-f stages. Bandwidth on the low band is controlled by the mutual inductor *L114,* on the high band by the mutual capacitor *C105.* The resonant transformer is properly loaded by the degree of coupling and by grid resistor *R104.* A single-tuned resonant circuit couples the plate circuit of the second r-f stage to the grid of the mixer. Local-oscillator injection arrives at the same mixer grid via small capacitor *C117* from the grid of the local oscillator. The local oscillator is a conventional ultra-audion high-frequency type, using the second section of the 12AT7 tube in a cathode feedback arrangement. The plate of the oscillator tube is at r-f ground potential because of capacitor *C116,* while feedback is accomplished with inductor *L150* between cathode and ground. The i-f frequency output of the tuner is link-coupled to the first resonant circuit of the video i-f strip.

Most tuners include separate high-band and low-band alignment-adjustment. On the high band, capacitors *C104, C106,* and *C108* adjust resonance and gain of the two circuits while capacitor *C105* controls the bandwidth. On the low band, inductors *L112, L119,* and *L127* control resonant tuning while inductor *L114* sets the bandwidth. In addition controls *L109, L116,* and *L124* improve resonant conditions at the very low end of the low band, channels *2* and *3.* For local-oscillator alignment separate inductors are tuned on each of the individual channels, with the fine-tuning capacitor *C118* set at the middle of its range.

The Standard Coil tuner, Fig. 64j, is a turret type that does not use an incremental coil system. Instead it uses separate individual coils for the r-f amplifier, mixer, and local-oscillator resonant circuits on each channel, and as the channel selector is rotated from one channel to another a different set of coils is switched into the circuit for each. The Standard Coil tuner uses a cascode r-f amplifier and a 6J6 mixer-oscillator combination. A single set of alignment adjustments is provided, and these are adjusted with the tuner set on channel *12.* Capacitor *C13* controls the amplitude of the pattern and capacitor *C3* the center frequency. The combination of capacitors *C3* and *C6*

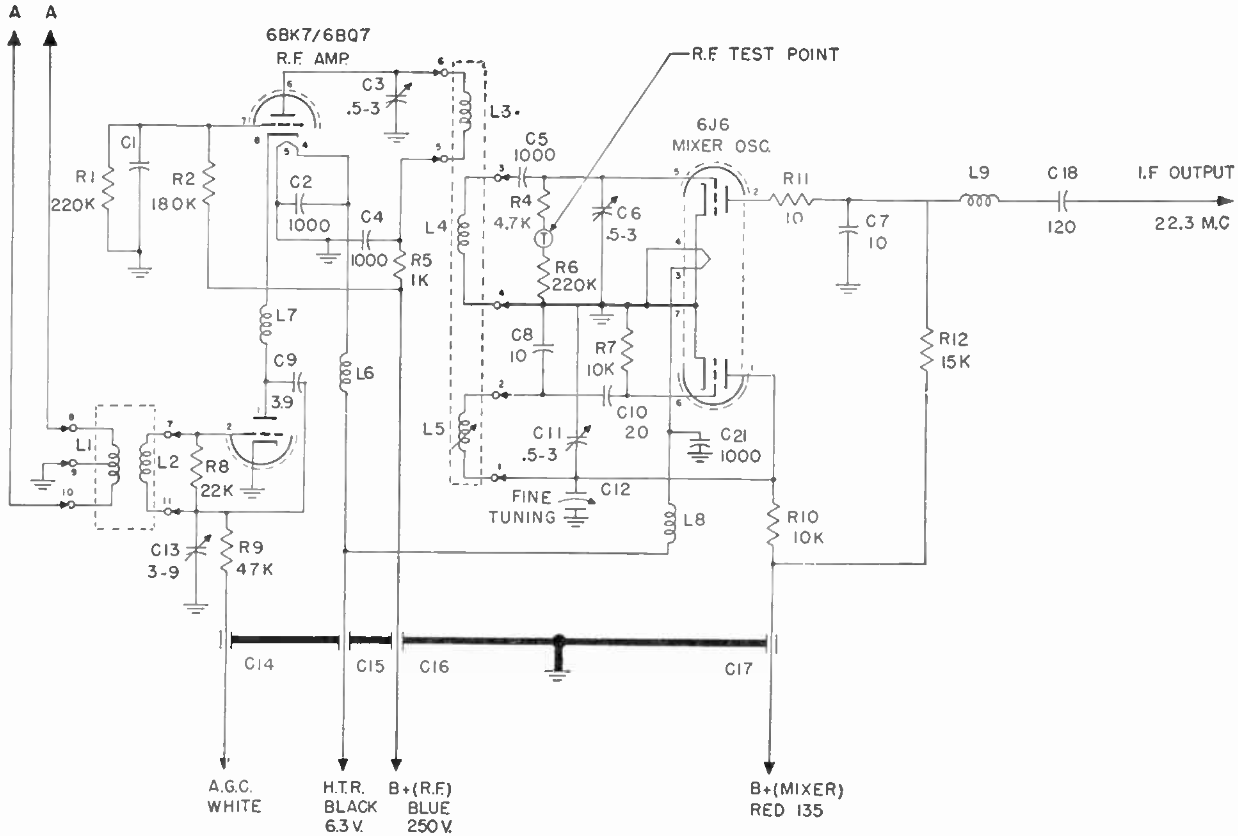


FIG. 64j Standard Coil Tuner

determines the over-all response and bandwidth of the alignment curve. There is no bandwidth adjustment provided as such, because the bandwidth on each individual channel is controlled in production with the mutual relationship between inductors *L3* and *L4* on each channel strip. A separate local-oscillator alignment-adjustment is provided for each channel of the tuner. In this type of tuner, it is possible to remove any one of the VHF strips for an unused channel in an area and to substitute a UHF converter-strip to permit reception of a UHF channel.

53. *Automatic Frequency Control*

This innovation in the television receiver, aimed at better r-f stability and elimination of the fine-tuning control, is automatic frequency control of the local oscillator. In this system (Fig. 65) a reactance tube driven by the output of the sound discriminator holds the local oscillator precisely on frequency. Consequently, the sound does not drift in and out, and there is no critical

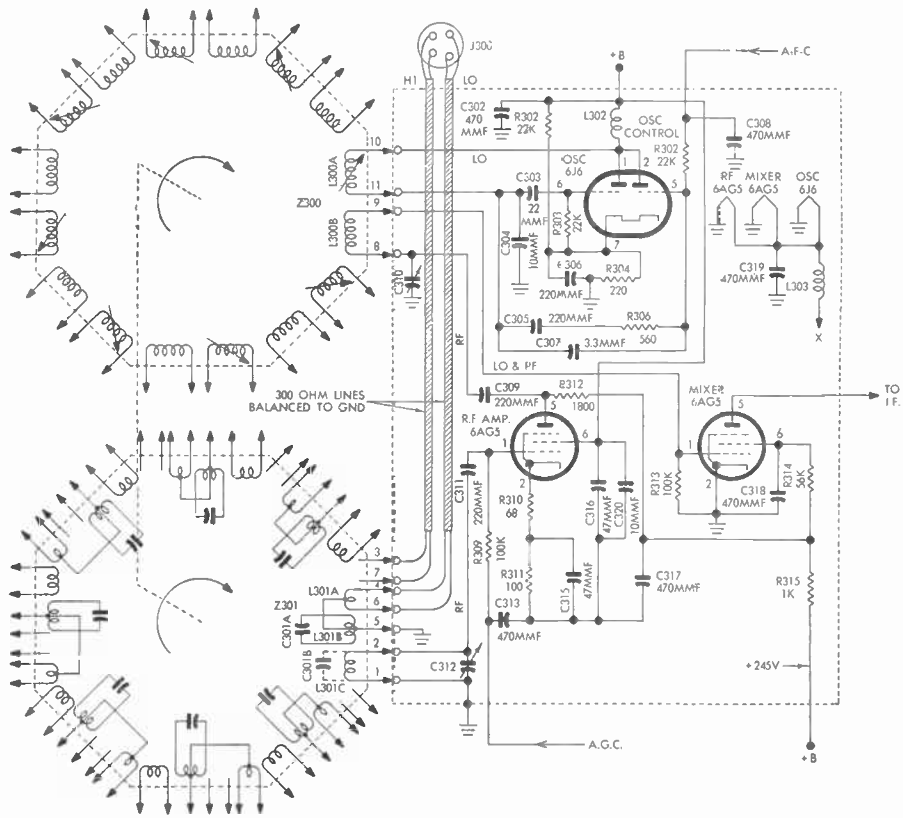


FIG. 65 Philco Turret Tuner

adjustment of the fine-tuning control. In fact, there is no need for such a control at all.

In the Philco system, shown in Fig. 66, a 6J6 miniature is used as a combination local oscillator and a-f-c reactance tube. If for some reason, such as heating or shift in supply voltage, the local oscillator attempts to drift in frequency, the center frequency of the sound i-f will also drift. Normally this would cause the sound to cut off or distort. When a-f-c control is used, another

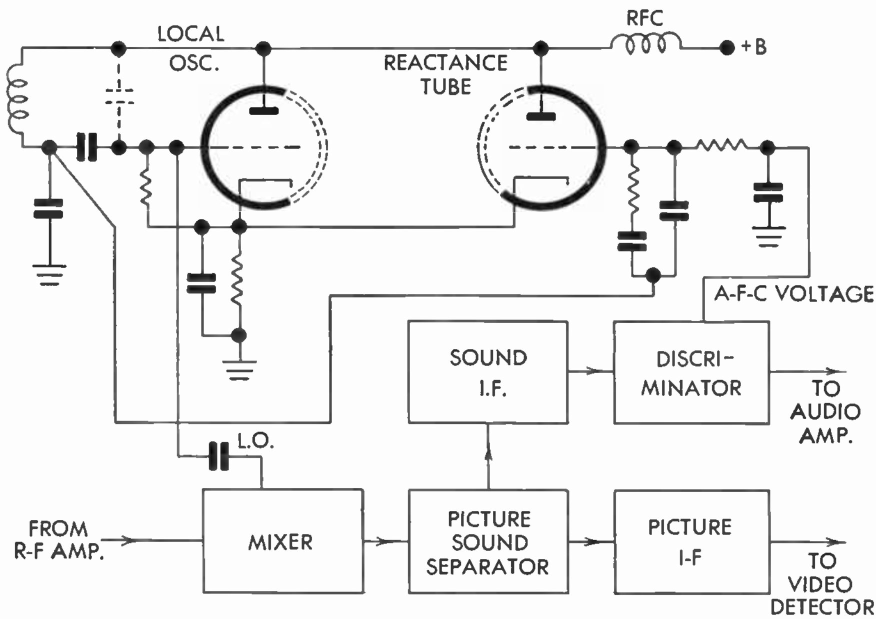


Fig. 66 Philco A-F-C Control of Local Oscillator

action begins when the sound i-f center frequency changes. For example, we know that at any time the center frequency drifts, the d-c component of discriminator voltage also changes. An a-f-c system utilizes this change in discriminator output. Actually, as the discriminator voltage changes with center-frequency drift, the bias on the reactance tube also varies.

The reactance tube in this case shunts more or less reactance across the oscillator-tuned circuit in accordance with its grid bias; therefore, the oscillator's resonant frequency shifts. The cycle of events is as follows:

1. Local oscillator drifts.
2. Sound i-f center frequency drifts.
3. D-C component of discriminator voltage changes.
4. Reactance tube grid bias changes.
5. Shunt reactance changes.
6. Local oscillator restored to its original frequency.

Inasmuch as only one local oscillator is used and it is held on frequency by the narrow-band sound system, it is also held precisely on frequency so far as the broader picture i-f system is concerned.

54. *General Characteristics of Television I-F System*

The i-f system of the television receiver is complex because of the many functions it must perform and the precautions which must be taken to prevent interference. The primary function of the i-f system, of course, is to amplify the picture signal at the output of the mixer to a level which will produce

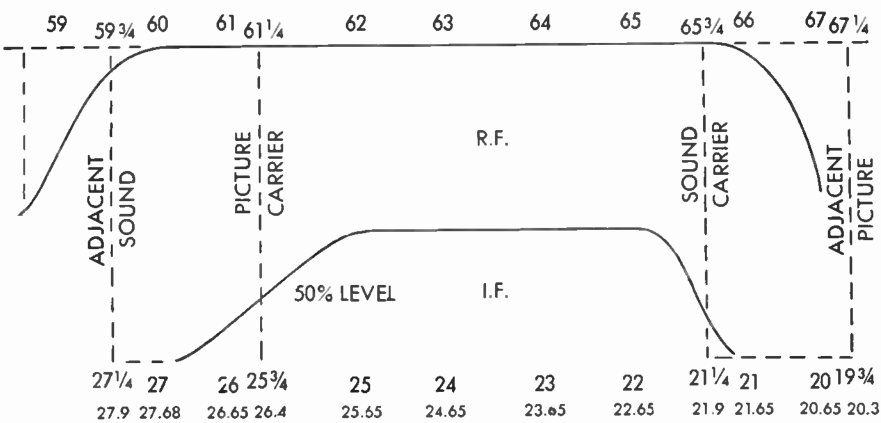


FIG. 67 Ideal R-F and I-F Bandpass. Local Oscillator on 87 or 87.65 Megacycles

a substantial detected output. It must linearly amplify a band of frequencies from 4 to 5 megacycles wide to take full advantage of the high-frequency components of modulation. The other major functions of the television i-f system are:

1. A means to separate the sound and picture i-f components is necessary.

2. Special-tuned circuits are necessary in the picture i-f systems to prevent interference by the associated channel sound, the adjacent channel sound, and in some cases the adjacent channel picture carrier.

3. The picture i-f carrier frequency must be detuned to the 50- to 60-per cent level to compensate for the vestigial sideband method of picture transmission. The ideal response for a television receiver at the r-f and i-f ranges is shown in Fig. 67. The i-f response at two i-f frequency ranges is indicated with channel *3* taken as a typical example. On channel *3* the picture-carrier frequency is 61¼ megacycles and the sound-carrier frequency is 65¾ megacycles. The local oscillator frequency of the television receiver is always higher than the frequency of the incoming signals; therefore, with standard i-f frequencies of 21¼ for sound and 25¾ for picture, the local oscillator frequency is on 87 megacycles. Thus, if we trace the dotted line on the figure from the

picture-carrier frequency of 61¼, we can observe how the i-f spectrum is arranged on the drawing. Notice particularly that the i-f picture-carrier frequency is detuned by 50 to 60 per cent on the i-f response characteristics.

RESONANT CHARACTERISTICS

The two most important factors to remember in analyzing stages with resonant circuits are whether the resonant circuit is in series or parallel with the signal path and whether it is a series or parallel resonant circuit. The simple schematics of Fig. 68 demonstrate the basic resonant characteristics. In drawing A, a parallel resonant circuit is inserted in series between the generator (vacuum tube can be considered as a generator) and the load (which might well be the grid resistor of a following stage). Since it is a basic fact that a parallel resonant circuit has a maximum impedance at resonance, it

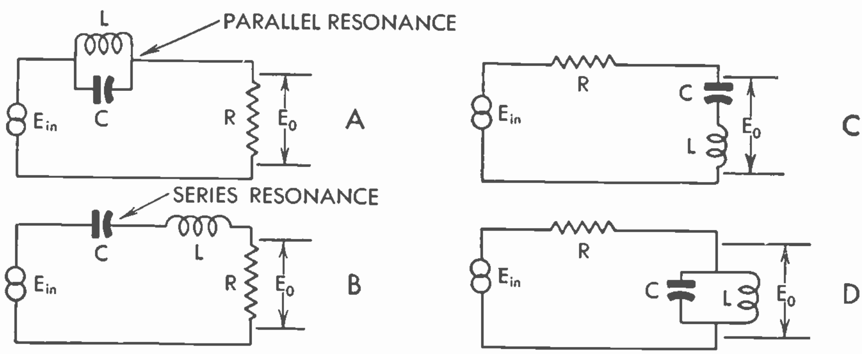


FIG. 68 Fundamentals of Resonant Circuits

opposes the transfer of a signal of the resonant frequency from the generator to the output resistor. However. at frequencies which depart from resonance, the impedance of the resonant circuit falls off and more of the signal reaches the output resistor. In drawing B we have the same arrangement with the exception that the tuned circuit is series resonant. Inasmuch as a series resonant circuit offers practically no impedance to the resonant frequency, an opposite transfer characteristic exists. Here the tuned circuit offers no opposition to a signal of the resonant frequency, it appearing almost in its entirety across the output load resistor. However, at frequencies which depart from resonance, the impedance of the series resonant circuit rises and opposes the transfer of these frequencies to the output resistor.

In drawings C and D two other conditions are represented in which the tuned circuits are the load (plate- and grid-tuned circuits of r-f and i-f amplifiers) for the signal, and the resistance *R* is in series with the signal path to the load (representative of the plate resistance of a vacuum tube). In this arrangement no signal appears across the output when a series resonant (drawing C) circuit is used and a resonant frequency is applied. At frequencies which depart from resonance, the impedance of the series-tuned circuit rises and output voltage appears across this impedance. When the load is a parallel resonant circuit (drawing D), its high resonant impedance causes maximum output of the resonant frequency and a decreasing output as the frequency departs from resonance.

The following facts concerning resonant circuits should be understood and retained:

(a) Parallel resonant circuit has maximum impedance at resonance, developing maximum voltage across it. It draws minimum current from the source.

(b) Series resonant circuit has minimum impedance at resonance; minimum voltage is developed across it. It draws maximum current from the source.

(c) The resonant circuit can either serve as a load across which an output voltage appears or can be inserted in series with a signal path to oppose a signal transfer.

(d) A parallel resonant circuit, as a load, develops maximum voltage at resonance; as a series insertion, prevents the transfer of the resonant signal.

(e) A series resonant circuit, as a load, prevents development of the resonant frequency; as a series insertion, it prevents transfer of signals not of the resonant frequency.

SIDEBAND CORRECTION

To conserve space in the television channel spectrum and to make transmission of a better defined picture with only a 6-megacycle channel, a system called *partial sideband suppression* is used. As shown in Fig. 67, the high-frequency sideband of the received signal is flat over approximately a 4-megacycle range while the low-frequency sideband is flat over only a ¾-megacycle range. A 4-megacycle component of modulation, nevertheless, is transmitted by the system with only a 6-megacycle channel. Normally, with symmetrical sideband transmission, to transmit a 4-megacycle modulation component would require an 8-megacycle channel. It is apparent that a component of modulation of less than ¾ megacycle will have a high sideband and a low sideband while a component of modulation greater than ¾ megacycle will have only a high-frequency sideband. Thus, the video detector in the television receiver is excited by two sidebands for components of modulation less than ¾ megacycle and by only one sideband for components of modulation in excess of ¾ megacycle. To compensate for this nonlinear detection, it is necessary that we start tapering the i-f gain at about a ¾-megacycle point on the high-frequency side (actually this will be the low-frequency side of the i-f spectrum signal). Thus, for a fixed amplitude-modulation component, the detected signal will have the same amplitude for a low-frequency component of modulation as for a high-frequency component of modulation. To produce the properly tapered i-f gain, it is necessary that the picture i-f carrier be down about 50 per cent of peak amplitude.

Inasmuch as the picture i-f bandpass is very broad, it is very susceptible to interference from signals with frequencies in the bandpass or near the supposed ends of the bandpass. For example, the high-frequency limit of the received r-f signal produces an i-f sideband component of 21.75 megacycles, which is only ½ megacycle away from the i-f sound-carrier frequency. The i-f system must be capable of amplifying the 21.75-megacycle sideband component and have a much lower response to the 21.25-megacycle sound which would put an interfering bar modulation pattern on the screen. To minimize this sound-carrier signal, a high-impedance wavetrap must be used to absorb and reduce sound signal in the picture i-f system.

Another source of interference in the television i-f system is the signal which originates at the adjacent channel sound-carrier frequency. For example, if your receiver is set on channel *3*, the local oscillator is tuned to 87 megacycles. Now, if there is a station telecasting on channel *2*, its sound-carrier frequency is of course located at 59¾ megacycles, which is only ¼ megacycle away from the frequency limit of the channel to which you are tuned. Thus, if your receiver is a bit sensitive to 59¾ megacycles, this adjacent sound carrier will enter your receiver, beat with your local oscillator frequency (tuned to 87 megacycles), and produce an i-f signal in the plate circuit of the mixer. This undesired i-f frequency will be, of course, 87 megacycles minus 59¾ megacycles, or 27¼ megacycles. As shown in Fig. 67, 27¼ megacycles is very close to the end of the high-frequency limit of the i-f bandpass. Consequently, if the i-f system is the least bit sensitive to 27¼ megacycles, modulation from this adjacent sound signal will reach the video detector. It is necessary, therefore, that we also have wavetraps in the picture i-f system tuned to 27¼ megacycles to block these frequencies from the video detector.

Occasionally there is interference from the adjacent channel picture signal on the channel immediately above the channel to which you are tuned. Again, if you are tuned to channel *3*, your oscillator is tuned to 87 megacycles; and if your receiver is the least bit sensitive to channel *4* when tuned to channel *3*, a picture-carrier frequency of 67¼ megacycles will enter your receiver and beat with your local oscillator to produce an i-f frequency of 19¾ megacycles. It is necessary in some receivers to employ a 19¾-megacycle trap to reject this adjacent channel picture signal. In summation, the three traps in the picture i-f system are tuned to the sound i-f frequency, adjacent channel sound i-f frequency, and adjacent channel picture i-f frequency.

55. *Picture-Sound Separation*

In the dual-channel receiver tuned circuits are used to separate the sound and picture i-f carriers and sidebands. The sound i-f is, of course, very narrow with respect to the picture i-f channel; consequently, a rather sharply tuned circuit can be used to remove the sound from the output of the mixer or one of the i-f stages. A number of picture and sound separating systems are shown in Fig. 69. In the RCA receiver, which uses a single-tuned transformer and stagger-tuned i-f system, the sound is taken off in the converter transformer. The sound is taken off in a sharply resonant circuit tuned to the sound-carrier frequency and coupled near the regular plate load inductor of the mixer. This proximity-tuned circuit is used to absorb the sound-carrier frequency and sidebands which are present in the output of the mixer.

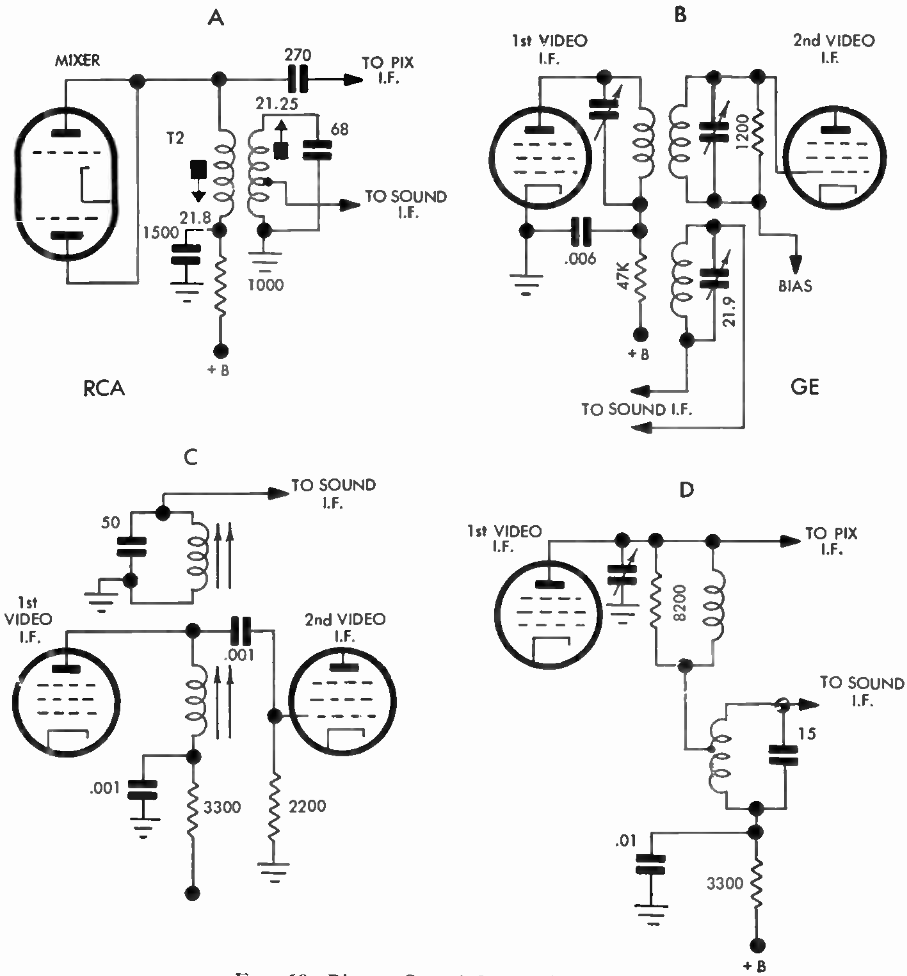


FIG. 69 Picture-Sound Separating Systems

In the General Electric receiver (drawing B) sound is again picked up with a proximity-tuned circuit. In this case the sound signal is withdrawn from the secondary of a double-tuned i-f transformer between the first and second picture i-f stages. Of course, in all cases when the sound is removed, it is again further amplified in a sound i-f system. However, the sound i-f bandpass is not very broad, and only one or two stages are necessary before the signal can be applied to the limiter and discriminator of the sound channel. The circuit of drawing C also uses a proximity-tuned circuit which is coupled to the single-tuned transformer between the first and second picture i-f stages. In some television receivers two tuned circuits are present in the plate circuit of the first i-f stage. These two tuned circuits are effectively in series, one of them tuned to the picture-carrier frequency and the other to the sound-carrier frequency. They present a maximum impedance to the frequency to which they are tuned and a lower impedance to the other frequencies. Thus, in one tuned circuit picture frequency dominates and in the other, sound.

It is to be remembered that in all the picture and sound separation systems we do not completely remove one signal from the other. What is done in the first step in a picture-sound separation is to make either the picture or the sound dominate in a given tuned circuit. Thus, in the sound take-off tuned circuit the sound dominates the picture; and in the picture-tuned circuit from which the sound was absorbed the picture dominates the sound. In subsequent stages, this ratio is raised and the picture dominates more and more until we reach a point where the picture is the only signal present in the picture i-f stage. Likewise, in just a few sharply tuned circuits, the sound dominates the picture in the sound i-f channel.

56. *Commercial Receiver I-F Systems*

The picture i-f system of the RCA receiver, shown in Fig. 70, employs a stagger-tuned i-f system with single-tuned transformers. Four miniature-tube i-f stages are used employing 6AG5's. Each tuned circuit is resonant to a slightly different frequency producing the characteristics shown in Fig. 71. The composite response of each of the individual tuned circuits produces the desired over-all response characteristics. It can be seen that the converter transformer *T2* and the plate-tuned circuit of the first picture i-f are tuned near the extremities of the bandpass of the i-f system. The remainder of the tuned transformers are tuned at intermediate frequencies, producing the desired over-all characteristics. The sound i-f take-off is a tuned circuit placed close to the plate circuit of the converter transformer. The first sound trap, which is an adjacent channel sound trap, absorbs this frequency from the primary of transformer *T103*. A proximity trap associated with transformer *T104* absorbs the adjacent channel picture signal from the i-f system. The associated channel sound trap is a proximity trap in the cathode circuit of the fourth picture i-f tube. It absorbs the undesired sound signal from the primary winding of transformer *T105*, which in conjunction with *C131* series resonates at 23.4 megacycles. Thus, so far as the desired signal is concerned, the cathode is at absolute ground potential, and a maximum 23.4 megacycle signal is developed across the output of the stage. The gain of the i-f system is varied by means of a grid-bias control which varies the grid bias on the i-f stages.

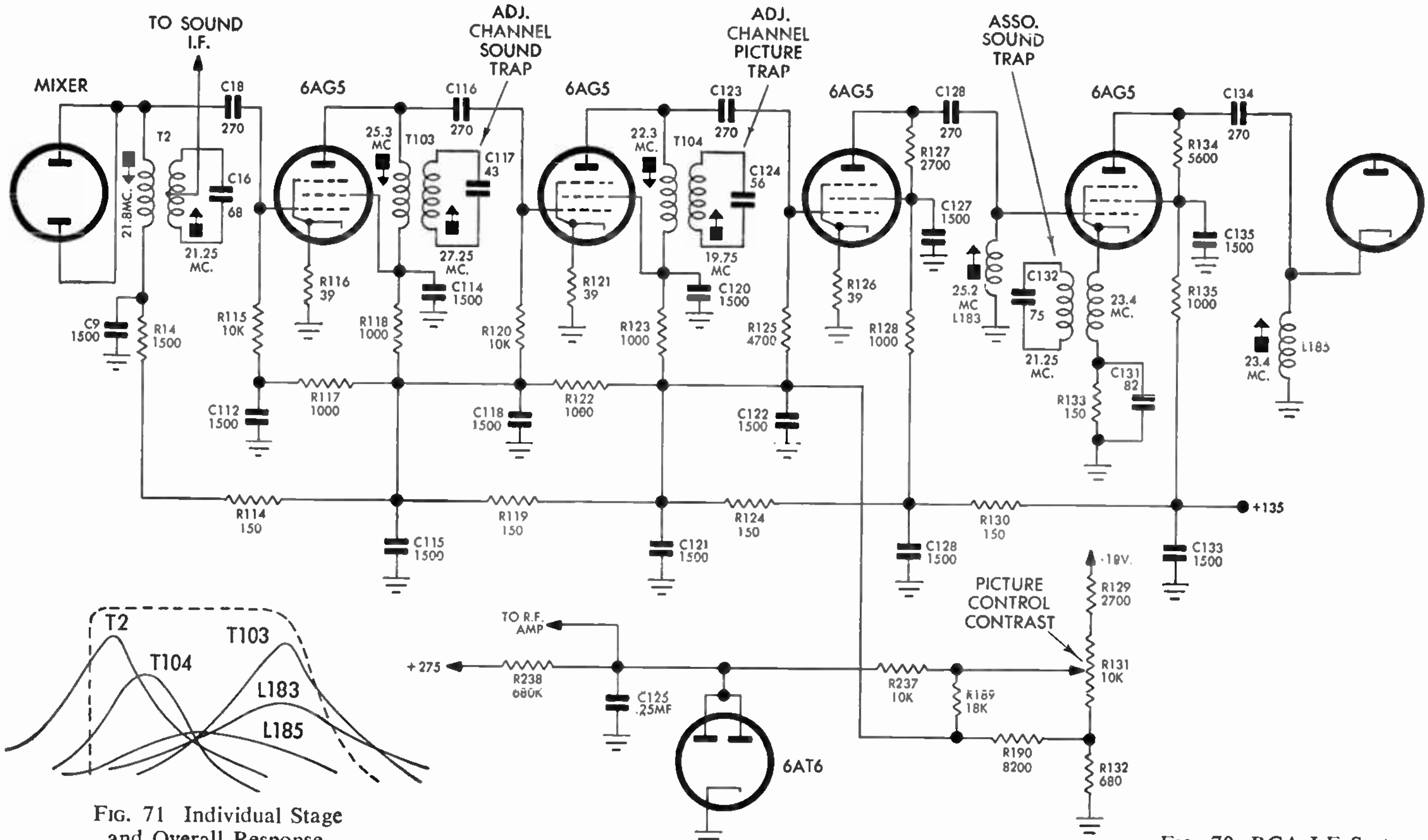


FIG. 71 Individual Stage and Overall Response

FIG. 70 RCA I-F System

The picture control varies the grid bias on these tubes and, therefore, the gain of the i-f system. It controls peak-to-peak amplitude of the detector signal and, therefore, the contrast of the picture signal as it is applied eventually to the grid of the picture tube.

Each of the tuned transformers is loaded by a grid or plate resistor which broadens the response of the tuned circuit for the required bandwidth. To reduce hum pickup in the i-f system and to reduce i-f frequency loss and tendency to oscillate, the heater circuits are returned very effectively to ground by means of a filter capacitor and choke. This combination is used in the heater circuits in all of the i-f tubes. The screens and plates are also returned to ground through the proper decoupling resistors and capacitors.

A novel bias-control system is used in the i-f amplifier which permits the bias of the i-f amplifier to be varied with the picture control but does not affect the bias of the r-f amplifier until there has been a considerable change in the setting of the picture-control potentiometer of the i-f system. It can be seen that the bias for the r-f stage is mainly controlled by the diode current of the 6AT6 tube, and it is not until the arm is well down on the picture control potentiometer that the negative voltage is applied to this plate of considerable amplitude to cut the diode off and apply bias to the r-f amplifier.

The i-f system of a General Electric receiver is shown in Fig. 72. In this receiver the picture i-f carrier frequency is 26.4 megacycles and the sound i-f carrier is 21.9 megacycles. The i-f transformers consist of the conventional double-tuned close-coupled transformers to give the best gain at the required bandwidth. An adjacent channel sound trap is placed in close proximity to the secondary of the first transformer *T1*, and an associated channel sound trap is placed in close proximity to the secondary of transformers *T2*, *T3*, and *T4*. The adjacent channel sound trap is tuned to 27.9 megacycles; the associated channel sound trap is tuned to 21.9 megacycles. A pretuned trap is also present in the cathode circuit of the second video i-f tube and is used to broaden the attenuation around the associated sound channel frequency to prevent serious overshoot or a sharp characteristic. A novel contrast-control arrangement consists of a potentiometer and a diode. The diode rectifies 6.3 alternating voltage, which passes current through the contrast potentiometer and another series resistor to ground. The 6.3 volts is applied to the cathode of the diode and a negative d-c voltage appears across the diode load, which is the contrast potentiometer.

The tuned transformers are, of course, loaded with shunt resistors to broaden the frequency response, and the plate and screen circuits are all properly decoupled and filtered.

In other receiver models by General Electric a single-winding i-f transformer is used (Fig. 72a), primary and secondary segments resonating (slug tuned) at correct frequency with circuit capacity. Resonant circuits are loaded by low-value plate and grid resistors.

The video i-f systems of the DuMont receiver series use bandpass T cou-

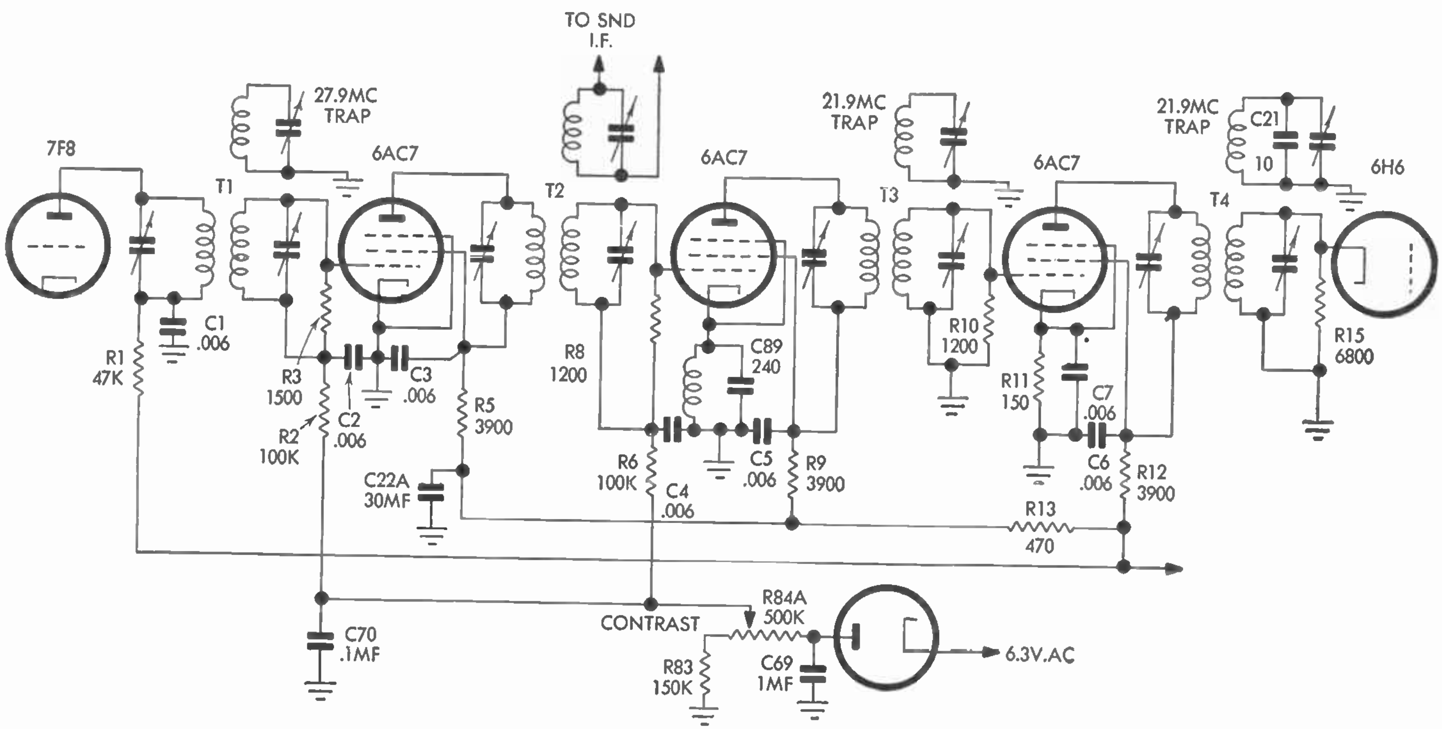


FIG. 72 GE I-F System

pling between stages (Fig. 73). The primary and secondary tuned circuits are adjusted with movable cores and resonate with the distributed circuit capacity at the proper i-f frequency. A mutual inductor overcouples between resonant circuits. For example, between the mixer and first i-f stage inductor, *L5* is the primary, *L6* the secondary, and *L7* the mutual coupling inductor. Sound i-f traps are located between first and second i-f stages and second and third i-f stages. These traps are series-tuned to the undesired sound i-f and contain a parallel resonant circuit to prevent loss of video i-f components.

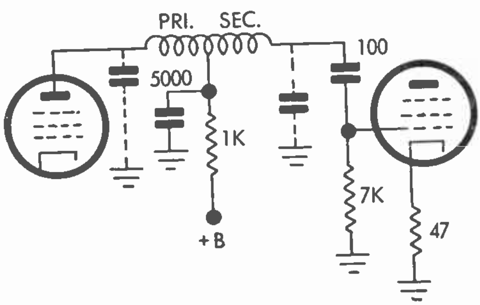


FIG. 72a Single-Winding I-F Transformer

The contrast control varies grid bias of the first and second i-f stages and is supplied from a separate 20-volt bias source.

In the alignment of the television i-f system, it is absolutely essential that the technician understand the i-f system and the function of the various components and various controls in this system. Do not attempt the alignment of any i-f system unless you thoroughly understand the system you are working with or unless you have precise alignment instructions necessary for the receiver. The i-f systems of most television receivers require no alignment at all or very infrequent alignment. It is the best policy not to change any control unless you know that changing that control is absolutely necessary and unless you know absolutely what can be expected after it is changed.

56a. *Intercarrier I-F System*

A special i-f and video system called "intercarrier" was developed by General Electric. In this system, Fig. 73a, the picture and the sound are carried through the picture i-f and video amplifier up to the grid of the picture tube. Only one i-f system is necessary. The sound is generally removed by a tuned circuit located between the video output and the grid of the picture tube, although in some intercarrier designs the sound is separated as early as or before the video detector. The sound is removed at a frequency of 4½ megacycles and passed to a conventional limiter-discriminator and audio system, after amplification by its own 4½-megacycle sound i-f amplifier.

Actually, the 4½-megacycle frequency to which the sound i-f tunes is the megacycle separation between the picture and the sound carriers. In other words, in transmitting the picture and sound signals from the station a continuous 4½-megacycle separation is maintained between their carriers. However, the sound signal from the station is frequency-modulated, and thus its frequency relation shifts in frequency with relation to the picture carrier as a

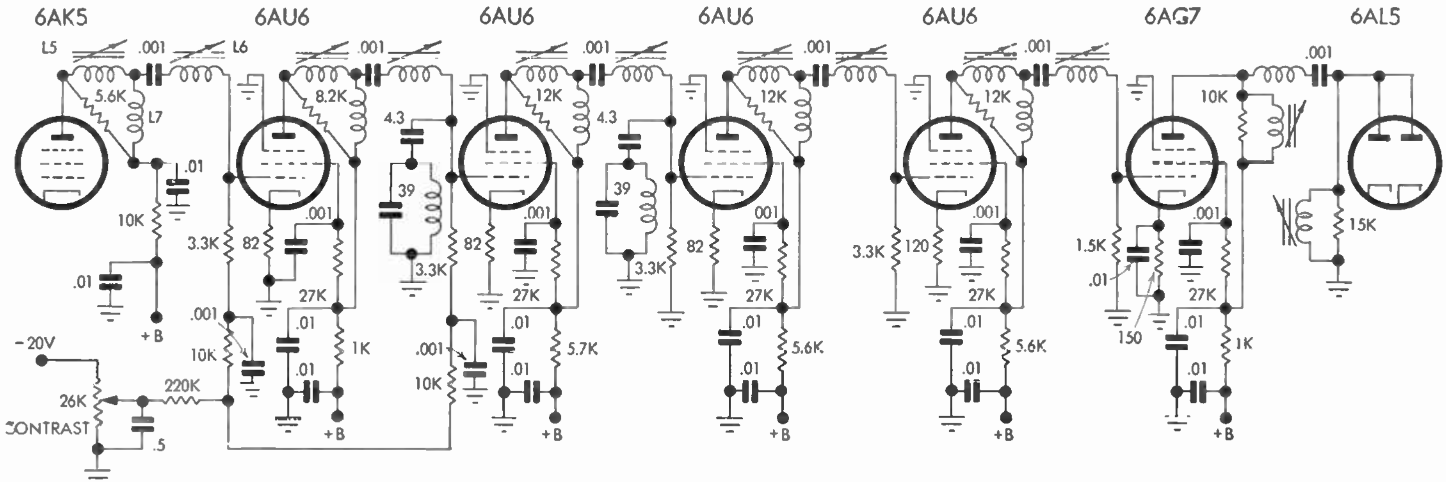


FIG. 73 DuMont I-F System

function of the audio variations to be transmitted. Consequently, the 4½-megacycle separation between the two carriers also deviates in accordance with the 25-kilocycle deviation of the sound carrier. If the picture carrier is made to dominate the sound carrier as the former's wide band of frequencies is passed through the i-f system, the frequency-modulated sound carrier looks just like another sideband component of modulation to the picture carrier and the video detector—a component that frequency-modulates with the original

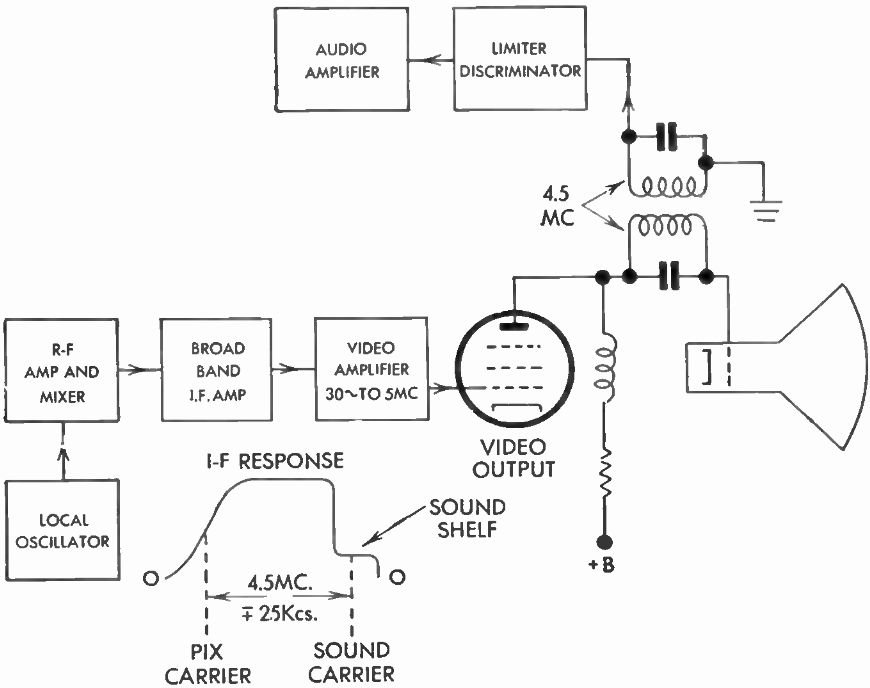


FIG. 73a GE Intercarrier System

audio information. To make certain that the picture carrier always dominates at the video detector, the i-f system is detuned greatly at the sound carrier; consequently the sound carrier appearing on a shelf on the response curve, Fig. 73a, is at a level less than 5 per cent of the amplitude of the picture carrier. Inasmuch as the sound carrier is frequency-modulated, the megacycle separation between it and the picture carrier is not always 4½ megacycles but deviates 4½ megacycles plus or minus 25 kilocycles. Consequently, there is a frequency-modulated 4½-megacycle center frequency in the parallel resonant circuit that is removed and later detected as the sound signal. The parallel resonant circuit in series with the picture signal-path also prevents the sound from appearing on the grid of the picture tube. The presence of the 4½-megacycle sideband component in the picture signal would cause a series of very fine vertical lines to appear in the picture from left to right across the entire screen. With the intercarrier system oscillator tuning is not critical, because the sound i-f frequency is dependent upon the megacycle separation between picture and sound carrier, and this is independent of the setting of the local oscillator. Inasmuch as the picture i-f system is so very broad, the setting of the local oscillator is not so critical as in the dual channel type of i-f system where it must set the sound component within narrow limits on its i-f bandpass.

Thus the sound presentation need not depend on any precise setting of the local-oscillator frequency, and as a result there is less adjustment of the fine-tuning control necessary with an intercarrier receiver. Likewise the customer can always tune for the best picture and need not worry about peak sound-setting as much as he would with a dual channel i-f system. Microphonics, oscillator-drift, and power hum are not as injurious to the intercarrier sound.

There are, however, a number of defects that can cause a particularly disturbing hum, referred to as "intercarrier buzz" in the intercarrier receiver. A number of basic intercarrier defects can cause this disturbance:

1. If the sound carrier becomes too strong in relation to the picture carrier and is modulated by picture components, these components will reproduce in the sound channel as an audio-frequency buzz.

2. Generally, any amplitude-modulation of the sound carrier (4½-megacycle carrier) by the picture carrier or sidebands can be removed by the action of a well-designed limiter in the 4.5-megacycle channel. However, under certain conditions variations in the picture carrier of a frequency or phase nature can produce frequency- or phase-modulation of the 4½-megacycle carrier, and this will eventually reproduce as noise in the sound channel. Consequently, greater care must be taken to prevent frequency- or phase-modulation of the picture carrier at the source of signal.

3. Any interruption of the picture carrier at the source also causes a loss of the 4½-megacycle component and a disturbance in the sound channel. For example, if the picture carrier is interrupted at a specific low-frequency rate, this same rate will cause the interruption of the sound carrier at the receiver and a resultant detected noise at the sound output. For example, if the video signal is allowed to swing into the white far enough to produce less than 5-per cent modulation of the picture carrier, it is possible that the picture carrier will be interrupted at this rate and cause a similar interruption in the sound signal. This defect often becomes apparent to the viewer as a change in the intercarrier buzz level when there is a change-over made between television cameras of a particular program. Minimum modulation of the picture carrier should be limited to no less than 10 per cent.

4. These defects can be accented by improper design of the intercarrier i-f

system, improper alignment of the i-f system, and improper adjustment or drift of resonant traps and sound take-off circuits.

5. A 4.5-megacycle sound channel must be aligned carefully, particularly the limiter and discriminator.

6. When the sound is taken off at the output of the video amplifier, the video amplifier must have a linear amplitude-response from sync tip to white level in order to prevent picture components from modulating the 4.5-megacycle sound-signal component. Thus video amplifier design is more critical if buzz level is to be held at a minimum. The intercarrier system in its basic form requires fewer stages and fewer component parts and apparently has a definite economic advantage. However, if full-picture bandwidth is to be obtained and buzz level held at a minimum, the added design criticalness of the i-f system, video amplifier, and sound channel causes such economic advantage to be very questionable.

In an intercarrier i-f system, the picture i-f carrier can be considered to act as a local-oscillator injection voltage and the sound i-f carrier as the very weak signal voltage. When these two signals of considerable difference in amplitude are mixed at the video detector, the resultant signal will possess the amplitude and modulation of the weaker component. In this case, the second detector produces a 4.5-megacycle-beat signal free essentially of the AM modulation contained on the picture carrier but still carrying its frequency-modulated audio component. Thus the video i-f system must handle the two relative signal levels with care to prevent noise and high hum level.

56b. *Commercial Intercarrier I-F Systems*

A double-tuned link-coupled transformer conveys the signal between the mixer of the RCA VHF tuner and the grid of the first i-f amplifier, Fig. 73b. Associated with this transformer is a single trap circuit located in the grid-transformer can that is tuned to 39¾ megacycles, the adjacent channel's picture-carrier frequency. The 45-megacycle i-f system is free of many of the interference-producing combinations prevalent in the 25-megacycle range. In addition, local-oscillator radiation is eliminated in the television band. However, it is possible to have the fundamental i-f frequencies and their harmonics, as generated in non-linear circuits of the i-f system, find their way back to the input system of the VHF tuner and cause interference with incoming signals.

A bandpass-filter combination transfers the signal between the first and second i-f amplifier stages, whose resonant circuits are housed in separate shields; over-coupling is obtained with mutual inductor and capacitors—two series resonant traps and capacitor *C225*. One of these traps is the associated channel sound frequency of 41¼ megacycles, and it is always set critically to obtain the proper amplitude relationship between the picture and sound carriers passing through the intercarrier i-f amplifier. The second series resonant trap, located in the second can, is tuned to 47¼ megacycles, the adjacent

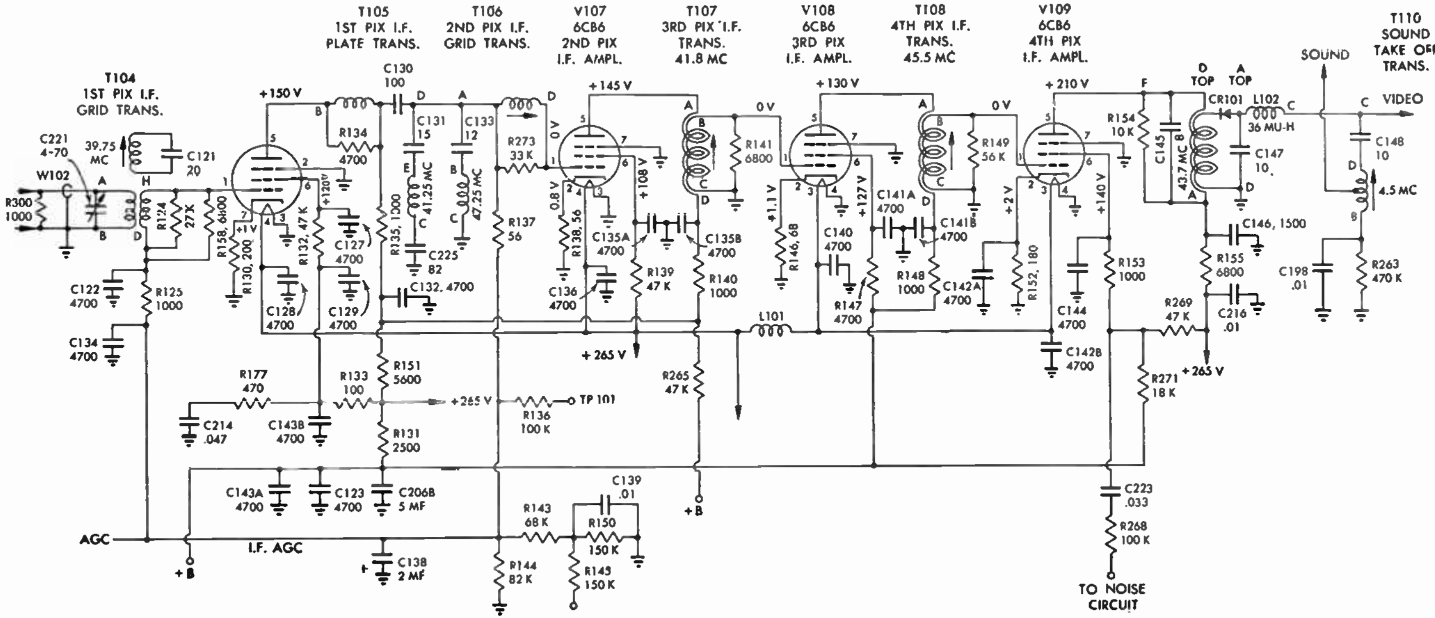


FIG. 73b RCA Intercarrier I-F System

channel sound frequency. Thus the various trap-circuits permit proper adjustment of carrier levels and minimize the possibility of adjacent channel spill-over. Notice in particular that the earlier i-f amplifier stages employ unbypassed cathode circuits to minimize the possibility of oscillation and to permit more stable operation. At the same time the unbypassed arrangement reduces "Miller Effect" and, consequently, prevents a change in the resonant frequency of the tuned circuits with a-g-c bias. All supply lines are thoroughly filtered to prevent degeneration and possible oscillation in the i-f system.

The third, fourth, and fifth i-f stages are stagger-tuned and employ bi-filar coils as portions of the resonant circuits. The third i-f stage tunes to the low-frequency side of the i-f bandpass and, therefore, is important in establishing fine picture detail. The fourth i-f stage tunes near the picture i-f carrier, while the fifth tunes near the center of the bandpass. Over-all bandwidth of the i-f amplifier is approximately 3.8 megacycles, with the various resonant circuits properly loaded with resistors in order to establish correct bandwidth.

In the bi-filar winding the two wires are simultaneously wound parallel and beside each other from the start of the winding to the finish. Primary and secondary windings are therefore interlaced with respect to each other. This method of winding, as compared with the use of two separate windings spaced on a coil form, provides a coefficient of coupling that approaches unity-coupling with a broadband characteristic. The bi-filar transformer is tuned by a threaded powered iron core adjustable from top or bottom of the i-f can. The bi-filar winding eliminates the need for a blocking capacitor and reduces the influence of impulse noises on bias and on operating conditions in the i-f stages. The secondary winding provides a low-resistance grid circuit and prevents interference and noise impulses from charging up any long time-constant grid circuits, which would adversely affect bias or create noise-modulation of the video i-f signal.

A germanium crystal diode is used as a video detector and is mounted on the top terminals of the fifth i-f transformer (the shield of the transformer preventing harmonic radiation from the non-linear detector circuits). The output of the video detector is coupled through a low-pass video filter to the grid of the video amplifier stage. The sound is removed from a special series resonant circuit housed in its own shield. This series resonant circuit, consisting of capacitor *C148* and tunable inductor, acts as a sound take-off and, at the same time, has a shunting effect on the 4.5-megacycle component, preventing it from appearing at the grid of the video amplifier stage. The series resonant circuit which can be sharply tuned also minimizes the possibility of inter-carrier buzz, in which the low frequency sync components of the video signal reach the sound i-f system. Sound is removed from a single element of a series combination, and consequently, a reasonably strong sound level is delivered into the 4.5-megacycle sound i-f system.

In the Philco i-f system, Fig. 73c, a low-impedance length of cable connects the output of the tuner with the input of the i-f system. A simplified schematic of the coupling system, as shown in Fig. 73d, indicates that the circuit is basically a low-pass filter arrangement. Thus higher-frequency sources of interference are rejected from the i-f system. In addition, the location of

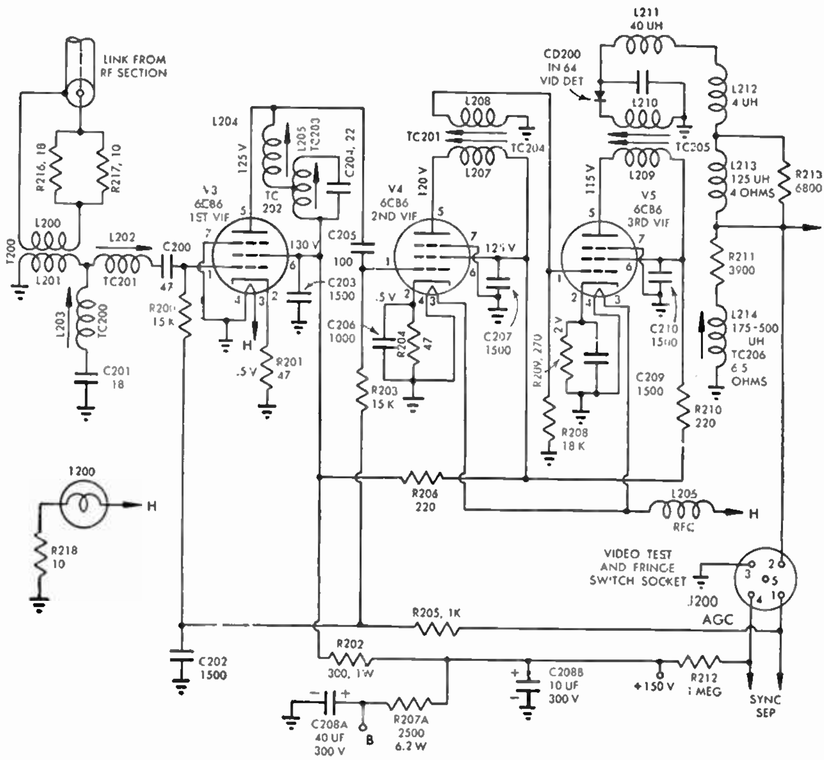


FIG. 73c Philco Intercarrier I-F

the cable at a low-impedance point of the coupling system prevents the cable capacity from shunting the i-f component to ground and permits a reasonable length of coaxial cable, not subject to interference pick-up. The cable terminates in a low-impedance coupling-transformer to the input system of the i-f amplifier for added insolation and selectivity. A series resonant trap, consisting of inductor *TC200* and capacitor *C201,* tunes out the adjacent channel sound-interference.

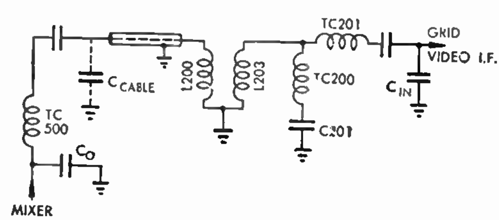


FIG. 73d Philco Tuner to I-F Coupling

A parallel absorption trap, Fig. 73c, consisting of inductor *L205* and capacitor *C204,* is tuned to the adjacent channel picture carrier. Double-tuned and over-coupled transformers transfer the signals between succeeding stages, finally delivering the i-f signals to the crystal video detector. The output of the video detector, through a suitable peaking- and lowpass-filter, applies the signal to the grid of the first stage of the video amplifier of the receiver. From the plate of this video amplifier stage a parallel resonant circuit removes the 4.5-megacycle sound component and applies it to the sound i-f amplifier. The elaborate filter network at the

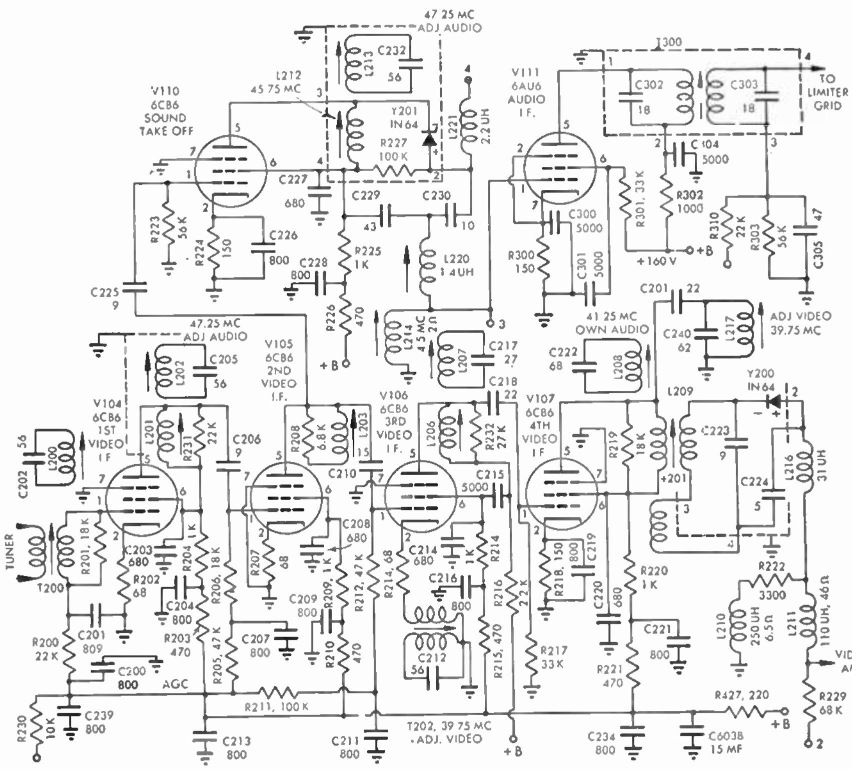


FIG. 73e GE Intercarrier I-F

output of video detector prevents the i-f frequency range signals from entering the video amplifier system, minimizes the harmonic radiation at the video detector circuit, and properly peaks the video bandpass. An adjustable peaking coil *L214* is available for properly setting the high-end response of the video system.

A special socket and fringe-local switch are included to permit more favorable operation of the a-g-c system and sync separator in the fringe areas. This system will be discussed in detail in conjunction with the paragraph on a-g-c operation, paragraph 70.

To minimize intercarrier buzz and to prevent the appearance of sound on the picture, General Electric, Fig. 73e, uses what is practically a dual channel i-f system in its intercarrier receiver. Sound and picture i-f components are amplified at wide-band level through the initial two stages of the video i-f system. However, signal paths divide at the plate circuit of *V105*, sound going to a special sound take-off amplifier that is resonated particularly and sharply to the sound i-f carrier frequency while picture through capacitor *C210* is applied to the grid of the third video i-f stage. A special crystal detector at the output of the sound take-off stage demodulates the 45¾-megacycle sound component in order to obtain an output frequency of 4.5 megacycles which is applied to the grid of the first audio i-f amplifier. The output of this amplifier is sharply resonant to 4.5 megacycles, and the signal passes in a conventional manner to a limiter and a ratio detector.

A special resonant circuit rejects the adjacent channel audio component by absorption in the plate circuit of the sound take-off amplifier. The video-carrier component passes through two more video stages in order to build up the video i-f level to proper amplitude for application to the video crystal detector *Y200*. In the latter two stages, special 41¼-megacycle associated-channel audio-traps are employed to prevent the appearance of audio bars in the picture. A series adjacent-channel video-trap is connected to the plate output circuit of the fourth video i-f stage.

This stage is also link-coupled to the crystal detector circuit which is completely shielded to minimize harmonic radiation. In fact, both crystal detector circuits are carefully shielded in the higher-frequency i-f systems to prevent harmonics generated in the non-linear circuits from reaching the input system of the VHF tuner and causing beat interference. Two adjacent-channel audio-traps are located in the input and output circuits of the first video i-f stage which receives its signal via link-coupling from the mixer output. A degenerative trap is located in the cathode circuit of the third video i-f amplifier and rejects the adjacent-channel video interference.

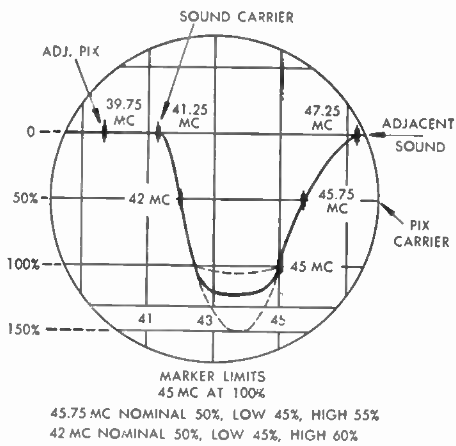


FIG. 73f GE Alignment Curve

Again the initial stages employ unbypassed cathode circuits to assure alignment stability and prevent a-g-c bias changes from shifting the desired response curve. Desired over-all response curve, Fig. 73f, shows how precisely possible sources of interference can be de-emphasized and the desired signal made to dominate with full bandwidth and fidelity. The intercarrier buzz problem is minimized with the early and separate sound take-off and detector.

QUESTIONS

1. What factors determine the value of resistor shunted across a tuned circuit in a wide-band amplifier?
2. What is relation between bandwidth, stage gain, and size of loading resistor?
3. Differentiate between single-tuned and double-tuned i-f transformer.
4. Differentiate between overcoupled and stagger-tuned i-f system.
5. State the general characteristics of an efficient wide-band amplifier.
6. What means are used to obtain overcoupling?
7. Why are wavetraps necessary in the picture i-f system?
8. Show location of wavetraps in the picture i-f system.
9. What are the particular advantages of miniature tubes in television circuits?
10. Draw the schematic and explain the characteristics of a grounded-grid amplifier.
11. Draw the schematic and explain the characteristics of a cathode-coupled stage.
12. Draw the schematic and explain the characteristics of the r-f amplifier which uses a section of transmission line as a parallel resonant circuit.
13. What is the importance of g_m and tube capacities in wide-band service?
14. Describe a typical television receiver r-f section.
15. What are the advantages of the inductuner?
16. What are the characteristics of an efficient r-f amplifier?
17. What methods are used to terminate the antenna system properly?
18. What are the general characteristics of an efficient mixer-oscillator combination?
19. Describe a typical commercial mixer-oscillator combination.
20. Describe briefly the operation of an a-f-c system for television.
21. If sound i-f frequency is 21.75 megacycles, what are the following frequencies: local oscillator on channel 5; picture i-f carrier; associated channel sound trap; adjacent channel sound and picture traps?
22. Why is the picture i-f carrier detuned?
23. How are picture and sound i-f carriers separated?
24. Describe a typical commercial i-f system.
25. How is the peak amplitude of the detected signal varied?

Chapter 5

VIDEO AMPLIFIER SYSTEMS

57. *Video Amplification*

The video amplifier is a resistance-capacitor coupled amplifier with necessary refinements to extend its high- and low-frequency response. It must amplify linearly a band of frequencies extending from 30 cycles to 3 or 4 megacycles. To do this, a video amplifier has a number of features which distinguish it from the conventional audio amplifier:

1. Low value of plate resistor
2. Tubes with high g_m and low capacities
3. High-frequency compensating circuits
4. Careful positioning and choice of parts to prevent capacity loss
5. Large coupling capacitors and grid resistors
6. Large cathode capacitors
7. Properly decoupled plate and screen-supply voltage lines
8. External grid-bias source and
9. Low-frequency compensating circuits

The initial four of these features maintain good high-frequency response, the remainder good low-frequency response.

58. *Frequency and Phase Response*

HIGH-FREQUENCY LOSS

The one factor which causes degeneration of the *high frequencies* is the distributed circuit capacity (Fig. 74), which is the sum of input and output capacities of the tubes, wiring capacity, and parts capacity to ground. The larger this capacity is, the more serious the high-frequency degeneration becomes. To keep this distributed circuit capacity down, low-capacity tubes are used and every precaution is taken to hold wiring and parts capacity at a minimum.

It is a fact that the output voltage of a resistance-coupled amplifier is the product of the current variation times the value of the output resistor. If we were to shunt this output resistor with another resistor of approximately the same value, the output voltage would decrease accordingly (same or somewhat higher plate current times a smaller value resistor). Likewise, a reactance shunted across the output resistor, if it is approximately the same or smaller in value than the resistor, will cause a decrease in output. The only difference is that capacitive reactance varies with frequency—in fact, decreases as frequency increases. Therefore, the reactance of the distributed circuit capacity may be very high over the low and middle range of frequencies and have little or no effect on the value of the plate resistance. As frequency increases, however, this reactance progressively decreases until it falls to a value equal to the value of the resistor; as frequency increases still higher, the reactance becomes smaller than the value of the resistor. This means that the value of the plate impedance is no longer equal to the value of the resistor but has become progressively lower and lower and the output less and less for a given change in plate current. In summation, it is the decreasing reactance of this distributed shunt capacity with higher frequencies which lowers the effective plate resistance and, therefore, the stage gain at high frequencies.

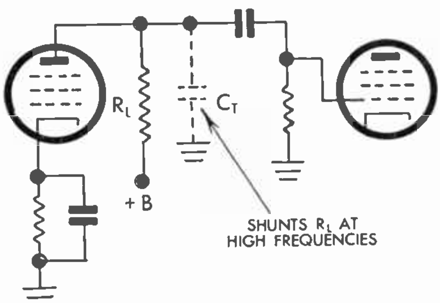


FIG. 74 Loss of Highs through Shunt Capacity

There are two ways to combat this degeneration. One method, of course, is to keep the value of the capacity low and, therefore, the reactance relatively higher at all frequencies. A second method is to lower the value of the plate resistor. This means that the reactance will have to drop to a much lower value before it begins to shunt the resistor seriously. However, the lower value resistor means less gain at even the middle- and low-frequency ranges. Truthfully, we have equalized the frequency response not by increasing the high-frequency gain but by lowering the gain at the middle- and low-frequency ranges.

Consequently, in video amplifiers, low-value plate resistors are used to broaden response at a sacrifice in gain. The gain of a video amplifier is many times less than that of an audio amplifier for this reason. To obtain the most gain possible under these circumstances it is important to use a high-g_m tube. In other words, to obtain a substantial output across a small-value plate resistor, the current variations should be great. A *high-g_m tube* produces large plate-current variations per given change in grid-signal voltage.

LOW-FREQUENCY LOSS

Low-frequency response is limited by the interstage-coupling combination and cathode-bias combinations. Here, it is the rising reactance of the capacitors which causes the loss of lows. In the case of the interstage-coupling capacitor, its reactance does not shunt the grid resistor of the next stage but is in series with it. Its reactance is large at low frequency (Fig. 75) in comparison to the grid resistance, and a greater portion of the low-frequency signal appears across the capacitor instead of on the grid of the next tube. Thus, it is apparent that the capacitor and resistor act as a voltage divider and, to obtain sufficient voltage on the grid of the tube, it is necessary that the capacitive reactance be small in comparison to the grid resistance.

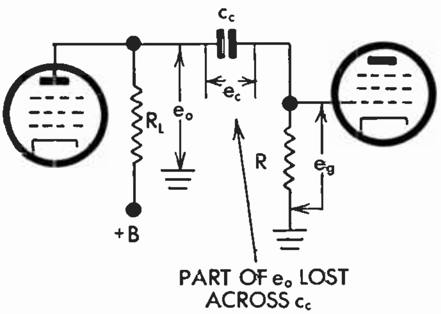


FIG. 75 Loss of Lows across Coupling Capacitor

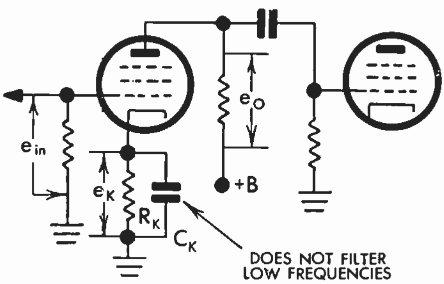


FIG. 76 Loss of Low Frequencies by Cathode Degeneration

To prevent low-frequency degeneration, therefore, the video stage uses large-value coupling capacitors (low reactance) and large grid resistors so that the low frequency signal appears across the grid resistor in almost its entirety. However, there are limitations to the value to which these parts can be raised. For example, the size of the grid resistor cannot be higher than the maximum set by the tube manufacturer to prevent excess gas current. The capacitor cannot be too large because of low-frequency disturbances, its greater d-c leakage and its large physical size and added capacity to ground, which affects the high-frequency response.

Another factor which causes low-frequency degeneration is insufficient cathode by-pass capacity (Fig. 76). The cathode capacitor normally acts as a filter and prevents any a-c variation from appearing between cathode and ground. However, to be an effective filter, its reactance at signal frequency should be less than 1/5 to 1/10 the value of the cathode resistor. At low frequencies the reactance of this capacitor begins to climb until it reaches a value equal to, or greater than, the value of the cathode resistor, permitting an a-c signal variation to appear across the cathode circuit. This variation is in such phase that it opposes the grid signal, thereby reducing the effectiveness of the grid signal and, in turn, the amplitude of the plate output. Therefore, it is necessary to use large cathode capacitors if we are to prevent excessive low-frequency degeneration. However, we are again hampered by physical size and construction of large capacitors. Many video amplifiers employ an external grid-bias source with grounded cathodes, thus eliminating one source of low-frequency degeneration completely. All plate and screen circuits are decoupled effectively (large by-pass capacitors) to prevent feedback and low-frequency oscillations through common power-supply impedance.

PHASE RESPONSE

Phase distortion is still another consideration in the design of a video amplifier. Fortunately, in a video amplifier with just a few stages, such as the video amplifier of a home television receiver, the maintenance of a linear frequency response also keeps the phase response in order. Phase distortion associated with poor frequency response seriously impairs picture sharpness. The reason it is noticeable is because the scanning beam moves at such high velocity across the screen that if one frequency component of a television signal is delayed with respect to a different frequency component, as it is with phase distortion present, the presentation begins to blur. This we can understand when we consider that signal components are not arriving on the control grid of the picture tube in the same precise order they were picked up at the pickup tube. Instead, some are delayed with respect to others because of their frequency. Thus, if two different frequency components of a signal were originally transmitted with their peak amplitudes occurring at the same instant, phase distortion (actually a nonlinear time delay—that is, one frequency is delayed with respect to another) would cause one to reach peak amplitude before the other at the picture-tube control grid. Therefore, the two components would be slightly displaced on the fluorescent screen, causing the picture to blur.

Phase distortion can also alter the waveform of a rectangular pulse, for the pulse is composed of a fundamental and a series of harmonic frequencies. If the high-frequency components of the pulse are delayed, the pulse will round off and lose its sharp characteristic. Actually, a very fast change or sharp vertical line in a picture is represented by a signal which has the appearance of a pulse, and if this pulse is distorted, the reproduced image at the receiver will not appear exactly as the camera-tube image. Occasionally, a video amplifier is found that has a peaked high-frequency response. The phase distortion present with this improper frequency response develops a sharp high-frequency voltage or transient which causes a thin white line to follow directly behind a thin black line of picture, or a thin black line to follow every sharp white line of picture.

Phase distortion at low frequencies can also impair picture quality. At low frequencies, phase distortion causes the low-frequency components to lead while at high frequencies the phase distortion causes a lag of the high-frequency components. The low-frequency components of the picture signal set the background brightness, and if they are distorted, there will be a change in background brightness from top to bottom of the screen. If the low-frequency phase shift is severe, white tails will appear to the right of black characters, sometimes extending all the way across the image.

59. *High-Frequency Compensation*

To permit use of a larger value plate resistor and, therefore, a greater gain per stage, small inductors are inserted at various points in the plate circuit of video amplifiers to hold up the plate-load impedance at the high frequencies. These inductors have an increasing reactance with frequency to balance the decreasing reactance of the distributed shunt capacity. Thus, the combination of the inductor, called a *peaking coil,* and the distributed capacity acts as a broad, partially resonant circuit at the high-frequency end of the frequency band to be passed. There are three common types of high-frequency compensation: shunt peaking, series peaking, and combination series-shunt peaking. In the shunt-peaking circuit, the peaking coil is inserted in series with the plate resistor (Fig. 77). The series impedance of the plate resistor and peaking coil rises with frequency while the reactance of the distributed capacity *C* declines with frequency. A second type of peaking, series peaking (Fig. 78), has the inductor inserted in the coupling path between plate and the grid of the next tube. This method permits a still larger plate resistor because it effectively isolates the distributed capacities of the succeeding stage-grid circuit from the plate circuit of the compensated stage. Thus, the distributed capacity shunting the plate-load resistor is less, and the value of the plate resistor can safely be increased. It is important to note that the increase in gain is not directly the result of the series-peaking coil but rather the effective isolation of the input capacity of the next stage from the compensated stage. This means that a larger value plate resistor can be used with consequent increase in gain.

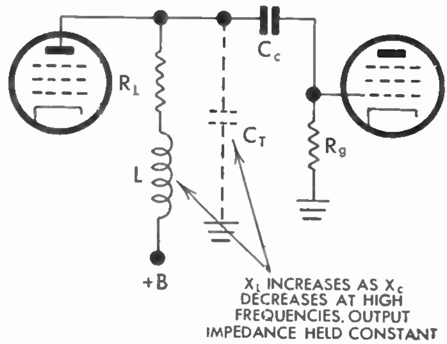


FIG. 77 Shunt Peaking

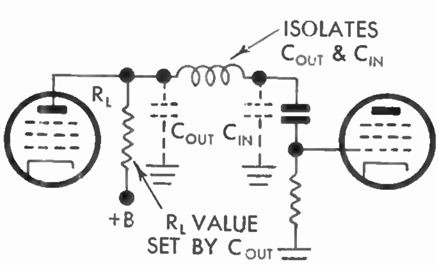


FIG. 78 Series Peaking

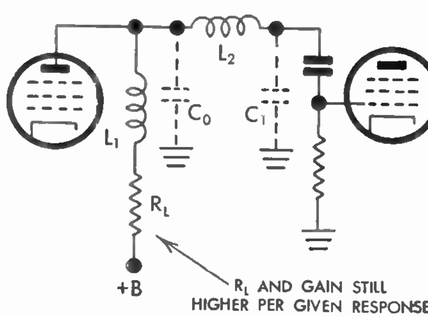


FIG. 79 Series-Shunt Peaking

Series peaking along with input and output capacities form a low-pass filter which passes all frequencies linearly up to high-frequency limit. A still further improvement in gain is attained by using a combination of series and shunt, as shown in Fig. 79.

60. *Low-Frequency Compensation*

Special circuits are also necessary to support the low-frequency response. The low-frequency amplitude lost across the coupling capacitors and cathode capacitors is re-established by increasing the plate impedance to the low frequencies. Thus, a relatively small capacitor, such as C_F in Fig. 80, which, of course, has an increasing reactance as the frequency falls, adds to the plate

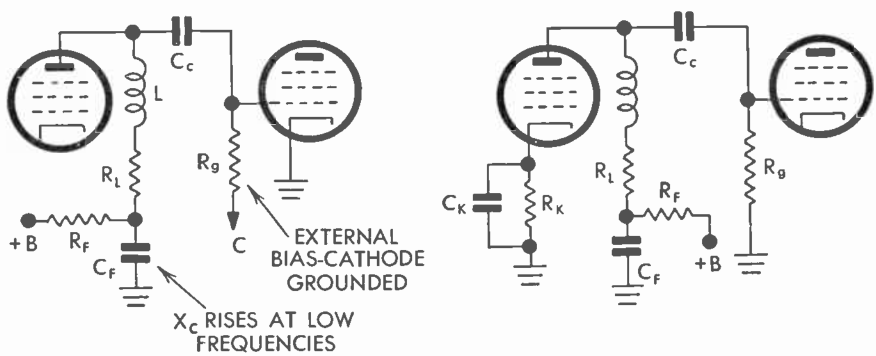


FIG. 80 Low-Frequency Compensation

resistance at low frequencies. Thus, so far as low frequencies are concerned, the stage has an increasing gain to compensate for low-frequency degeneration in other parts of the stage. External bias is used on the stage of drawing A, Fig. 80. When cathode bias is used (drawing B), a similar compensating system can be used to compensate for cathode degenerative losses. For a mathematical presentation of video amplification, refer to Chap. 14.

61. *Video Detection*

The video detector removes the picture modulation from the picture i-f carrier. This modulation consists of picture, sync, and blanking and is in the form of the composite television signal formed at the transmitter. The composite signal, of course, is a single-polarity signal, and the diode detector must be connected to develop a signal of the proper polarity for utilization at the picture tube. The video detector must be designed to detect the highest frequency component of modulation and at the same time reject the lowest frequency component of the i-f spectrum.

It is necessary that we have a negative-going composite signal applied to the grid of the picture tube. Thus, the darkest portion of the televised scene and the blanking level of this signal occur at a point where the instantaneous grid voltage is far enough negative to cause the picture to be dark, or to be cut off. During the brighter portion of the scene there is less negative voltage applied to the grid of the picture tube, and the screen is illuminated. Since the detected output of the video detector is the composite television signal, the polarity of the diode detector output must be such that the polarity of the composite signal, when it reaches the control grid of the picture tube, is negative. If there

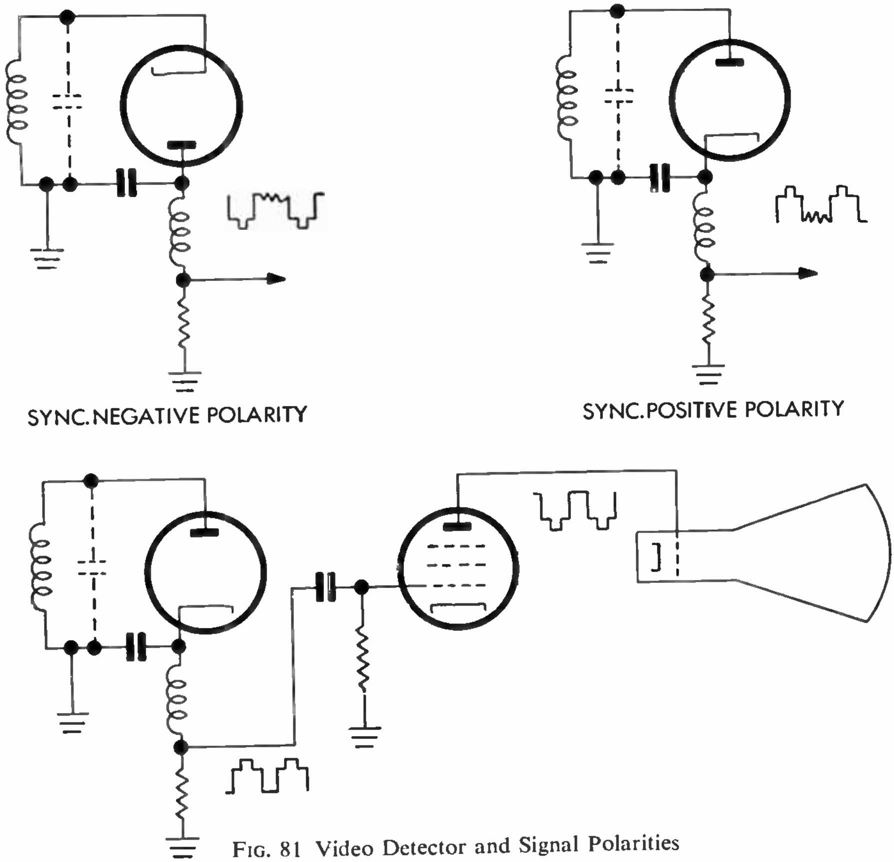


FIG. 81 Video Detector and Signal Polarities

is an even number of video stages between the detector and grid of the picture tube, the detector output must be negative; for an odd number of stages, positive. If the signal is applied to the cathode of the picture tube a positive signal must be applied to the cathode of the picture tube. This necessitates a positive detected signal when an even number of stages are used, and a negative signal when an odd number of stages are used. Two typical detectors, one of each polarity, are shown in Fig. 81. When the plate of the diode is connected above i-f signal ground, a positive polarity signal develops across the diode load resistor. When the cathode is above i-f signal ground, a negative signal appears across the diode load resistor. The latter is the preferred connection because the tube capacity shunted across the output diode load resistor (plate-to-ground capacity) is less.

DETECTOR CHARACTERISTICS

It is an inherent characteristic of the diode detector to develop a stronger signal at the middle range of frequency than at the higher frequency ranges because of the degenerative effects of shunt capacity across the diode load resistor. For detection of audio frequencies alone, this defect is not serious, and the size of a typical diode load resistor varies from ½ megohm to 5 megohms. However, in television, where we detect video frequencies as high as 4 megacycles, it is necessary to drastically lower the size of the diode load resistor to prevent degeneration at these frequencies in comparison to the middle range of frequencies. Thus, the decreasing reactance of the distributed capacity at high frequencies can not seriously reduce the value of the output load resistance.

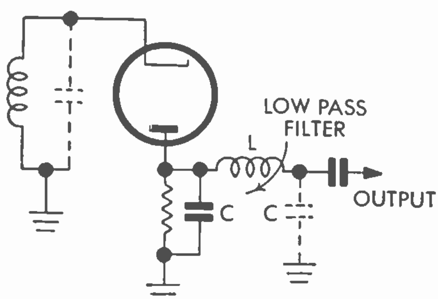


FIG. 82 Bandpass Filter in Output of Video Detector

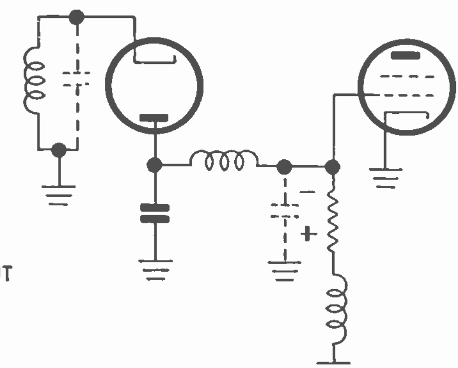


FIG. 83 Direct-Coupling from Video Detector

Typical video diode load resistors vary from 2,000 to 5,000 ohms. The efficiency of the diode detector is seriously reduced when the value of the diode load resistor is decreased to such a low value. Consequently, the amplitude of the developed signal is much lower, and it is necessary that the i-f amplifier have a much higher gain than the i-f system used in broadcast receiver service. It is apparent, therefore, that a multitube i-f system is necessary for picture reception not only because of the bandwidth of the i-f bandpass required but also because of the need for a much higher amplitude signal as applied to the video detector.

It is not only necessary for the video detector to detect the high-frequency components of modulation up to 4 megacycles, but the detector output circuit must also filter the i-f frequencies. Inasmuch as the lowest frequency i-f component is not very much higher in frequency than the highest frequency component of modulation, it is necessary to use a low-pass filter which passes the high-frequency components of modulation but has sufficient attenuation to remove the lower frequency components of the i-f frequency spectrum. Such a low-pass output circuit is shown in Fig. 82. Observation of the diagram shows that the filter is a simple low-pass filter, which passes components of frequency up to 4 megacycles but after that point begins to reject frequencies and causes all the i-f components to be filtered out. In the case of the older television receivers, this bandpass filter had to be critically designed because the lowest frequency component of i-f signal was at approximately 8½ megacycles, which is only double the highest frequency component of modulation at 4 megacycles. With the present i-f frequencies in the 25- to 50-megacycle range, this low-pass filter must not be quite as critical to remove i-f frequencies and harmonics.

D-C COUPLING

In many television receivers the output of the video detector is direct-coupled to the first video amplifier. Inasmuch as the output of the detector is a single-polarity signal with a d-c component of signal, it is not lost when the signal is direct-coupled to the succeeding video amplifier. If coupled through a capacitor d-c level is lost with changes in average brightness. A typical d-c coupled stage is shown in Fig. 83. In this circuit the grid of the video amplifier is connected directly to the top of the diode load resistor. Consequently, the no-signal current which flows through the diode resistor because of contact potential sets the no-signal bias of the grid of the video amplifier. This bias is, of course, negative. When signal is applied to the video detector, the output of the diode load resistor swings negative in accordance with the instantaneous amplitude of the i-f signal. The darker the scene transmitted, the further negative the voltage across the diode resistor becomes, and, consequently, the further negative the grid of the first video amplifier is driven. However, the applied signal can only make the output signal swing negative; consequently, the grid signal will only swing negative with respect to this fixed d-c component of bias. This constitutes direct coupling to the video amplifier.

Observation of the diagram shows a coil connected between the low side of the diode resistor and ground. This inductor is a shunt-peaking coil and boosts the output impedance of the diode circuit at high frequencies. Actually, it forms a partially resonant circuit with the distributed circuit capacity counteracting the decreasing reactance of the shunt capacity, which is so effective in degenerating the high frequencies. Thus, most diode detectors, in addition to the series bandpass filter inductor, also employ the shunt-peaking coil to boost further the high-frequency components of modulation.

62. *Typical Video-Detector Circuits*

A typical direct-coupled negative-polarity video detector is represented by the General Electric television receiver (Fig. 84). The i-f signal is applied between cathode and ground of the detector and develops a negative polarity signal across resistor *R14*, a 1,500-ohm diode load resistor. The bandpass filter of the detector consists of capacitor *C20*, inductor *L5*, and the input capacity of the video amplifier. Grid bias on the video amplifier is represented by the

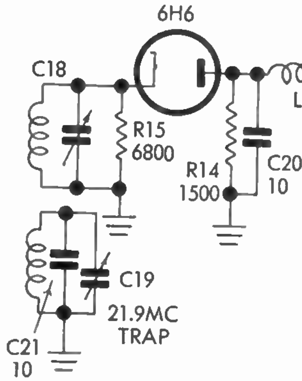


FIG. 84 GE Direct-Coupled Video Detector

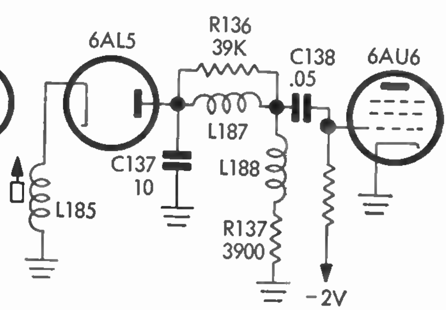


FIG. 85 RCA Video Detector

very small voltage drop across *R14* caused by the static diode current. Resistor *R15* is the load for the video-detector tuned circuit. A proximity trap tuned to 21.9 megacycles removes any trace of sound from the picture i-f signal.

The video detector of the RCA receiver (Fig. 85) also develops a negatively polarized signal across the output resistor, *R137*. However, this output is

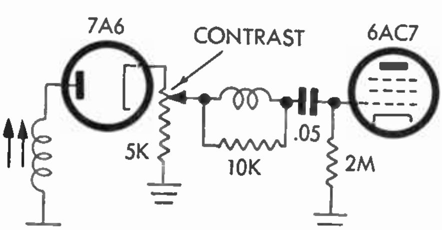


FIG. 86 Video Detector and Contrast Control

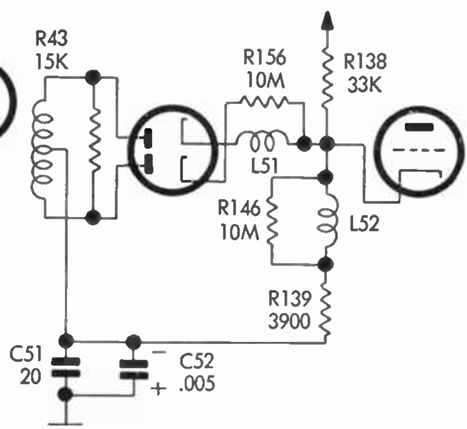


FIG. 87 RCA Push-Pull Detector

capacitively coupled to the grid of the first video amplifier, which receives its negative d-c component of bias from an external source. The lost d-c component of signal, when signal is coupled through the capacitor, must be reinserted at a later point by a d-c restorer. The output circuit of the RCA detector contains a bandpass filter consisting of *C137*, *L187*, and the input capacity of the following stage plus a shunt-peaking coil, *L188*. A resistor, *R136*, shunts the series coil of the output circuit to load this coil and prevent self-oscillation and transients. The input circuit of the diode detector consists of variable inductor *L185*, which is resonated with the input capacity of the diode to the proper i-f frequency.

The i-f signal is applied between plate and ground of the video detector of the television receiver (Fig. 86). Consequently, a positively polarized signal appears across the 5,000-ohm diode load potentiometer. In this receiver the potentiometer is the contrast control of the receiver and varies the amplitude of the picture signal applied to the grid of the video amplifier.

To develop a higher amplitude detected signal, an older RCA receiver used a push-pull detector, as shown in Fig. 87. The i-f signal is applied in push-pull across the diode plates—one diode conducting on one alternation of the i-f signal and the second diode on the other. A positive polarity signal is developed across the diode resistor *R139* and is direct-coupled to the cathode of the following stage. A low-pass filter and shunt peaking are used in the output circuit. The diode detector does not have a direct d-c path to ground to permit application of external bias to the following stage.

63. *Functions of Receiver Video Amplifier*

The video amplifier (one to three stages) increases the amplitude of the detected signal to a level sufficient to swing the picture-tube control grid over its full range. Two other functions of the video amplifier are to apply a properly polarized signal to the picture-tube control grid and to establish the average brightness of the picture signal. Properly polarized signal, of course, is accomplished by considering the number of video stages and properly choosing the diode detector polarity to produce a negatively polarized signal on the picture-tube grid. The establishment of correct average brightness is performed by a d-c restorer, which will be discussed in the succeeding paragraph.

A video amplifier is shown in Fig. 88. The many special features of this amplifier are numerically keyed with the diagram.

1. Polarity of diode is such that a negatively polarized signal appears across the diode load resistor and, after transfer through two stages, is also negative on the control grid of picture tube.
2. Low-value diode load resistor reduces amplitude of detected signal but preserves high-frequency response up to the highest modulation component.
3. Diode-peaking coil, which assists in maintaining the highs and aids in filtering out the i-f frequency.
4. High-value coupling capacitor to prevent degeneration of the lows.
5. Contrast adjustment which controls amplitude of signal developed across output of video amplifier (controlled degeneration) and, consequently, amplitude of signal as it appears on picture-tube control grid.

6. Adequate screen filtering to prevent oscillations or degeneration in video amplifier.

7. Shunt peaking to prevent loss of highs.

8. Low-value plate resistor to prevent loss of highs.

9. Resistor-capacitor combination with a long time constant for d-c restorer action.

10. Cathode grounded for d-c restorer action.

11. Shunt-peaking coil to prevent loss of highs.

12. Low-value plate resistor to prevent loss of highs.

13. D-c coupling between video output and picture-tube grid to prevent loss of d-c level established by d-c restorer action.

14. Brightness control to set d-c bias on picture tube for proper illumination of the fluorescent screen.

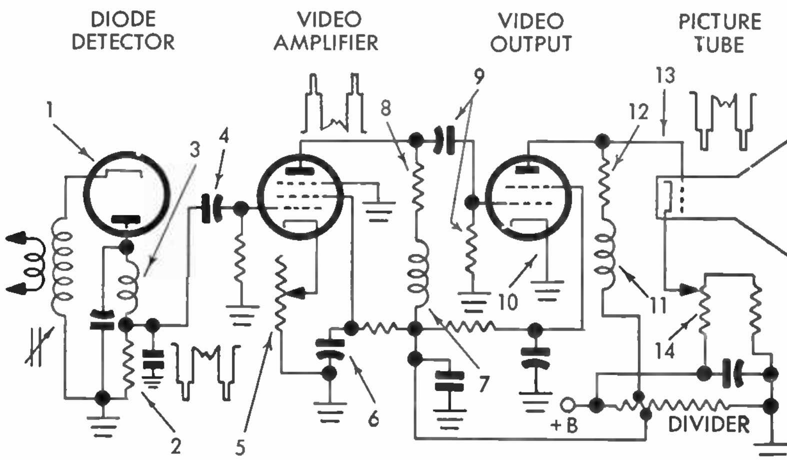


FIG. 88 Typical Video Amplifier

64. *D-C Restoration*

We have discussed contrast and brightness in a physical sense; now we will discuss them with respect to the construction of the composite signal. Each of the drawings of Fig. 89 represents the signal distribution along a few lines of a televised scene, with the signals inverted as they would actually appear on the control grid of the picture tube.

Drawing A shows a scene with a high-contrast range (illumination from very dark to very bright), with point *A* representing an extremely dark point on the line (almost at black level) and point *B* an extremely brilliant point. Furthermore, it represents a very abrupt and sharp change from dark to bright at this point.

Drawing B shows part of a scene in which almost the entire line is at a high negative-voltage level (near the blanking level), which maintains a high negative voltage on the grid, producing a correspondingly dark scene. Although the average brightness is low, there can be instantaneous bright spots along the line, such as point C, which represents a very brilliant spot.

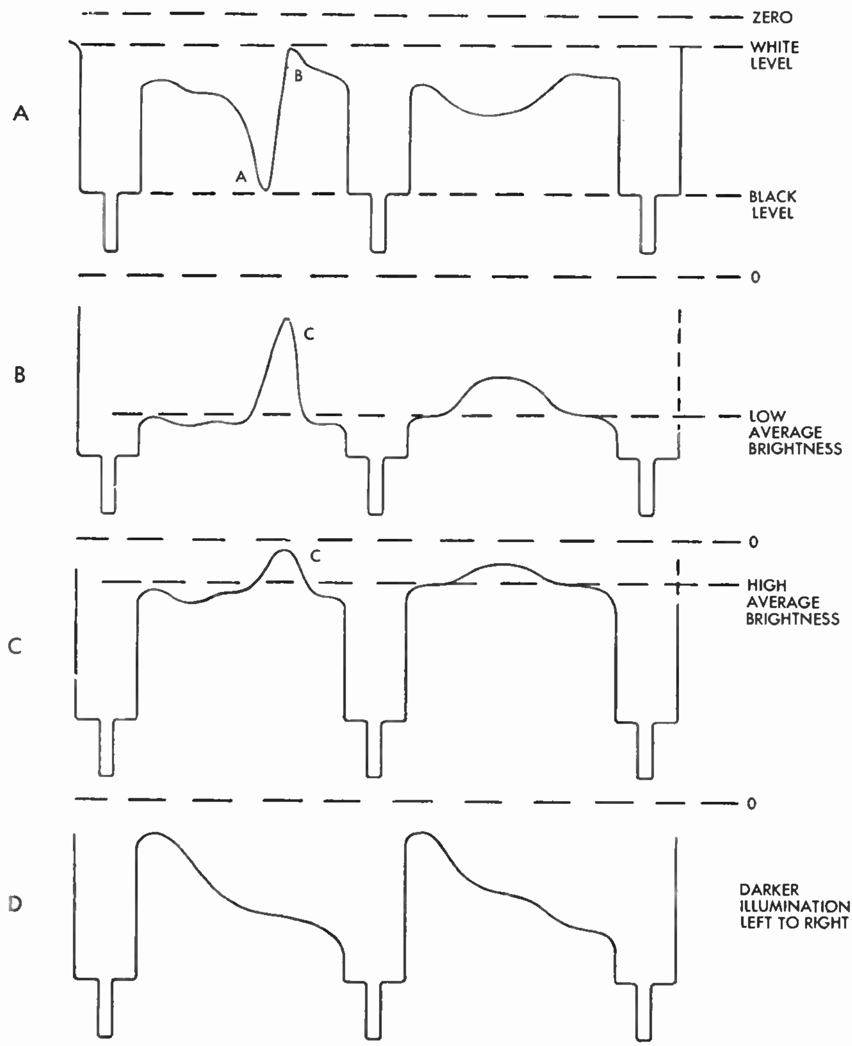


FIG. 89 Distribution of Picture Information

The relative light distribution along the line represented in drawing C is almost the same as the relative light distribution along the line of drawing B. However, the average brightness is higher, for the entire signal is at a lower negative-voltage level and the average illumination of the screen is correspondingly greater.

Drawing D shows a gradual decline in screen illumination as the beam moves from left to right across the screen, scanning one line.

As stated, each drawing represents light distribution along one line. If the same average brightness, as shown on any one drawing, is maintained throughout the total number of active lines for one frame, the average brightness of that frame is set. When the average brightness of the scene televised changes, the background level or brightness of the reproduced picture follows (difference between drawings B and C). Likewise, if the average brightness of the scene changes from top to bottom, the background brightness follows. Now the average picture-signal level (axis of symmetry of the video signal) represents the average illumination or background brightness of the scene. As shown, this d-c level rides up and down as scene brightness changes. (Compare drawings B and C.) It is a function of the video amplifier, and more specifically the d-c restorer, to cause the average bias on the picture-tube grid to vary with this shift in level, thereby changing the average illumination of the picture tube in accordance with average brightness of the scene. Special circuits are required to maintain the d-c levels of this single-polarity signal because each time it is capacitor-coupled, the signal arranges itself as an average about a zero axis (or bias axis of succeeding stage) without regard to the average-brightness level of the scene.

The d-c restorer associated with television receivers re-establishes the d-c blanking level of the composite signal as a constant after it has been lost through capacitive coupling in the video amplifier. Actually, it is a d-c reinserter re-establishing the d-c brightness component of the composite signal in accordance with its initial separation from the blanking level. If it were possible to develop a large enough signal across the diode detector load resistor to excite the control grid of the picture tube, no d-c restorer would be necessary. This is true because the signal output across the diode load resistor has a fixed blanking level which conforms to the level held by the original modulation at the transmitter. All that would be necessary, then, if sufficient amplitude were available, would be to direct-couple it to the picture-tube grid. Whenever capacitive coupling is used in the amplification process, this level is lost and must once again be reinserted at the video-output stage.

Briefly, the need for maintaining a fixed blanking level is to have the picture-tube control grid reach the blackout level for each blanking pulse, regardless of the character and average content of the picture signal. Thus, changes in average brightness of transmitted scene in no way affect the blanking of the fluorescent screen. Instead the *average* bias on the picture-tube grid varies up and down with changes in picture brightness, as it should (referred to as the *d-c brightness component*), while the blanking level of the signal remains fixed. If this system were not employed, the picture-tube grid bias and, therefore, average illumination would remain constant, and the blanking level would vary up and down.

Shift in the blanking level of the signal is caused by the charge and discharge

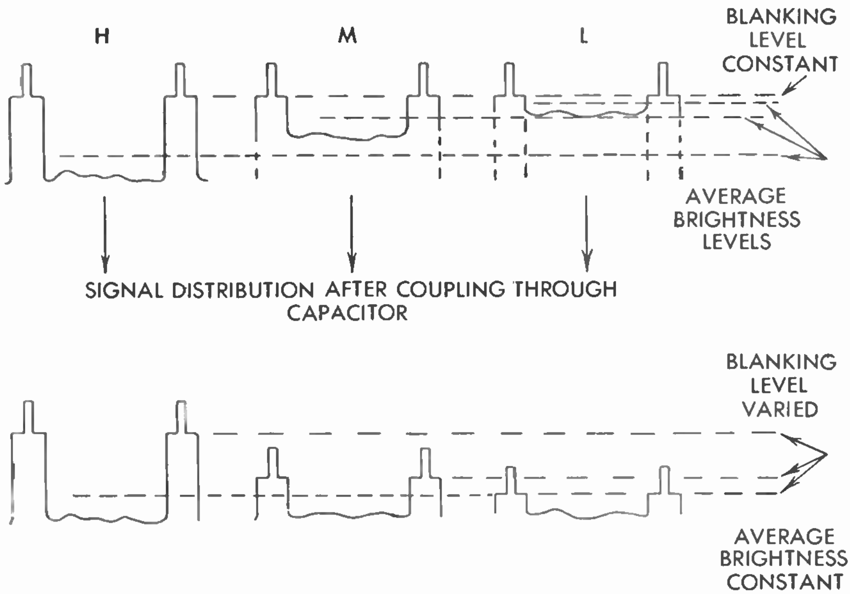


FIG. 90 Loss of Average Brightness and Blanking Levels through Capacitor Coupling

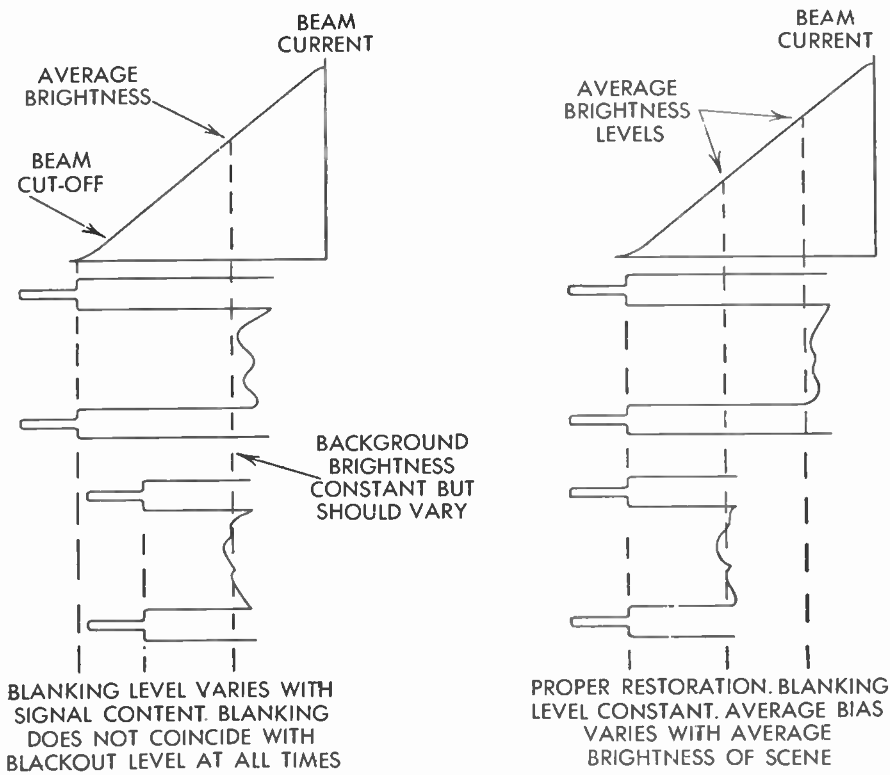


FIG. 91 Signal Distribution on Grid of Picture Tube with and without Proper Restoration

characteristics of any capacitor-resistor coupling combination between stages. This defect is demonstrated in Fig. 90, which shows the effects of the capacitor on the blanking level. In drawing B we have a signal with a low average brightness which, after it averages itself about zero signal level (as it would when being transferred through a capacitor) or negative-grid bias level on the succeeding tube, has a very low amplitude blanking level. If the signal has high average brightness (drawing A), the blanking level is only displaced a small amount and is of relatively higher amplitude. How this affects picture-tube operation is shown in Fig. 91. In drawing A we see that the shift in blanking level reduces the amplitude of the blanking level so much, in the case of a darker scene, that it does not supply a sufficient grid-voltage level to blank out the fluorescent screen. Furthermore, the average brightness of the screen (d-c brightness component) remains unchanged, while it should actually shift because a darker scene is being transmitted. The improvement in light distribution when using a d-c restoration system is demonstrated in drawing B. Notice that now the blanking level is held constant whether the scene is dark or bright while the average level of the picture signal shifts, which is just what should occur to change the average brightness of the scene.

A typical d-c restorer system is shown in Fig. 92. The video-output stage is operated without bias when no signal is applied (cathode grounded). As soon as signal is applied the control grid draws current, and this current flowing develops an average bias voltage, which is sustained by the capacitor *C*. Now, the amplitude of this bias is set by the peak-grid current drawn during the sync-tip interval (functions similar to a peak vacuum-tube voltmeter). Since the amplitude of the sync tip has varied with average picture content up to this stage, the peak grid current and developed bias vary correspondingly, being high for a high-amplitude sync tip and less for a low-amplitude sync tip. Just sufficient bias is added or subtracted by change in peak grid current to hold the blanking level essentially constant while the average brightness component varies. This is the desirable condition.

After the d-c level has been reinserted into the composite signal, this level must be retained up to the grid of the picture tube. Thus, if the level is re-established in the video-output stage, this stage must be direct-coupled to the control grid of the picture tube. Therefore, we can consider the final stage a d-c amplifier in the sense that it is direct-coupled to the following tube.

In the case of the video-output stage of Fig. 92, it is direct-coupled to the picture-tube grid. Since the d-c blanking level is held constant on the grid of the video-output tube, it draws a fixed amount of plate current through the plate resistor whenever the blanking signal is applied. Therefore, blanking at the grid of the picture tube represents a fixed voltage level. The average plate current of the video-output stage varies with scene brightness, and in turn the average picture-tube bias (direct-coupled) follows.

To summarize d-c restoration, the following step-by-step sequence should be studied in conjunction with Fig. 92:

1. The low and high average-brightness signals, as applied to the control grid of the d-c restorer video-output tube, have different sync-tip levels. The high average-brightness signal has the greater peak amplitude.

2. The high average-brightness signal will draw the highest peak grid current through low resistance path (grid-cathode conducting path) and will put the highest negative charge on C_c. The time constant of R_gC_c is long enough to hold this charge between sync tips (grid nonconducting), and the only time grid current is drawn is during the sync tip. During each sync tip the negative

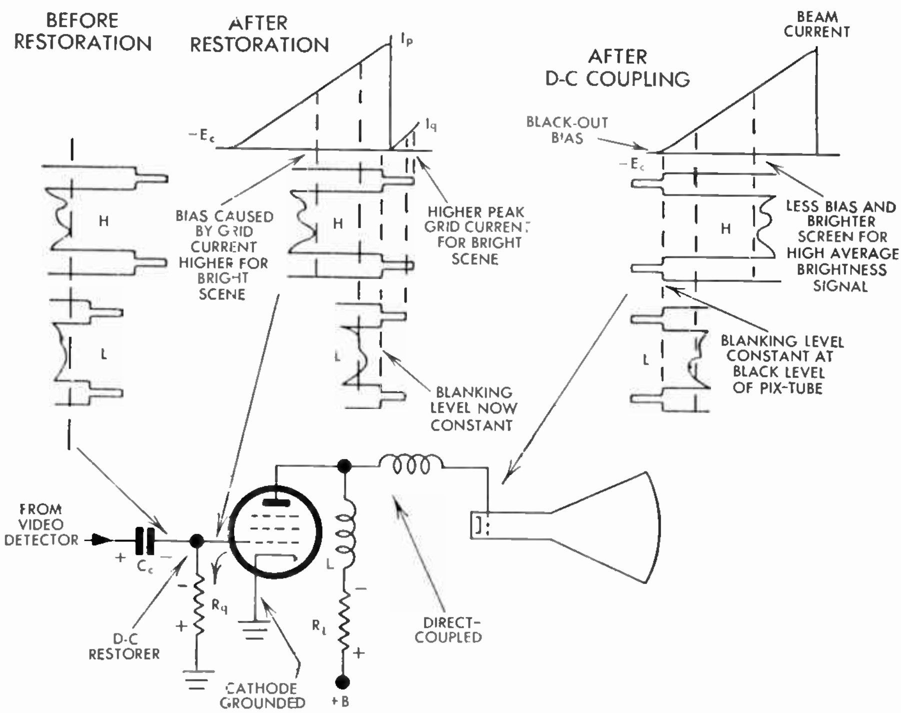


FIG. 92 Grid-Rectifier D-C Restorer and Waveforms

charge on C_c is replaced to its peak value—only a very minute discharge occurs between sync tips. It so happens that a higher peak charge is placed on C_c during the transmission of a high average-brightness scene because of the greater peak amplitude of the sync tip. Thus, the average bias is more for a high average-brightness scene while the blanking level for both signals is at the same level. For a more complete discussion and mathematical presentation of time constants, refer to Chap. 14.

3. With a higher average bias (higher average-brightness scene), the average plate current is less for the brighter scene. Inasmuch as the blanking level represents a fixed grid-voltage level, the instantaneous plate current drawn during the transmission of blanking is a constant

4. When the average plate current is least, as for a high average-brightness scene, the d-c component of plate voltage of the video-output tube is highest (less negative-voltage drop across R_L) and, therefore, there is the *least* bias (direct-coupling) on the grid of the picture tube, and the average brightness of the fluorescent screen is at its highest, as it should be, for the high average-brightness scene. Again, the plate voltage of the video-output tube during blanking is always the same and places an instantaneous grid-voltage level on the picture tube sufficient to black out the screen.

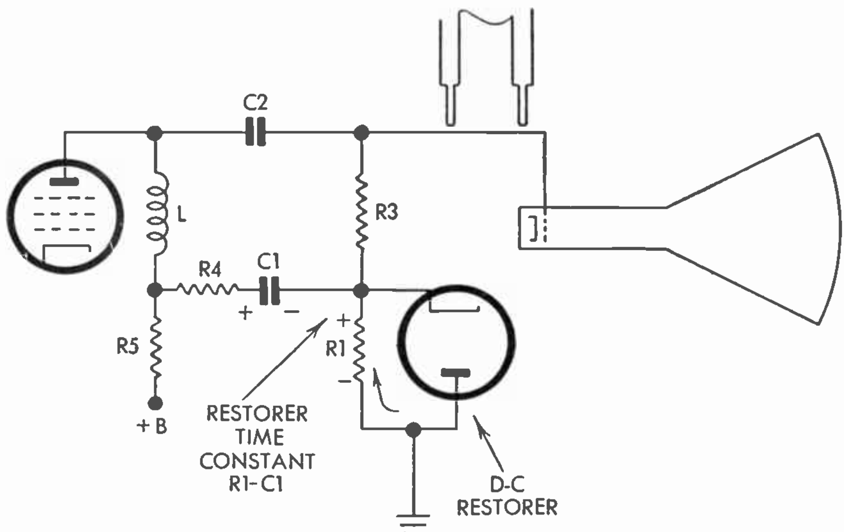


FIG. 93 Diode-Type D-C Restorer

A very common d-c restorer is the diode type, shown in Fig. 93, which is connected in the grid circuit of the picture tube. When this type is used, no direct-coupling is required and the video-output tube is capacitor-coupled to the picture tube. In this particular circuit, a diode is connected between grid and ground through an isolating resistor *R3*. A negatively polarized composite signal is applied to its cathode through isolating resistor *R4* and capacitor *C1*. The peak diode current drawn during the sync tip (drives cathode furthest negative) develops a *positive* voltage which charges capacitor *C1* through conducting diode. Again, the *RC* time constant is long enough to hold this charge between sync tips. The charge, of course, reduces the bias on the grid of the picture tube. The higher the average brightness, the higher the peak diode current and the higher the positive charge on *C1*. Thus, there is less bias on the grid of the picture tube, and background illumination of the screen is at its highest, as it should be.

The a-c component of signal is capacitor-coupled from the plate of the video output to picture-tube grid. Resistor *R3* prevents shunting of the high-frequency components of the picture signal by the capacity of the diode. Resistor *R4* prevents the conducting diode from shunting the plate resistor of the video-output tube.

Two advantages of the diode type of restorer, as compared to the grid rectifier type, are that it improves linearity as to brightness changes and also permits operation of video-output tube with higher gain. With the grid-rectifier type, output tube is zero-biased (no signal), and screen voltage must be kept down to prevent excessive plate current. Lower screen voltage means less gain.

D-C AMPLIFICATION

If it is possible to carry the d-c component of the television signal from the output of the video detector to the grid of the picture tube, no system of d-c restoration is necessary. A so-called d-c or direct-coupled amplifier can convey such a d-c signal component. In such a system the d-c blocking coupling capacitor is not used because about any such capacitor the signal would arrange itself as an average (same number of electrons displaced in and out of capacitor) and d-c brightness level is lost.

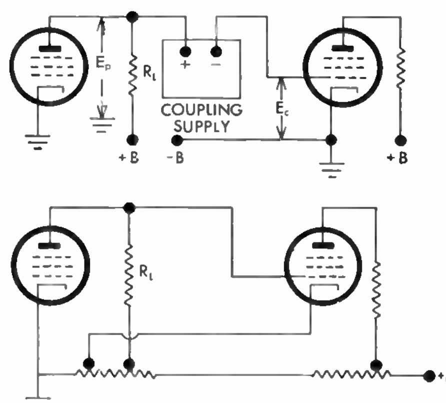


Fig. 93a Basic D-C Amplification

The problem, then, in d-c amplifiers is to devise a system which permits direct coupling between stages and does not apply excessive potential to tube grids or demand too high a supply potential. In the first system shown in Fig. 93a an isolated voltage supply is inserted between plate and grid which must have a high enough negative potential to oppose the d-c plate voltage of the preceding stage and apply proper grid bias to next tube. Thus the d-c component of grid bias is the series algebraic sum of the coupling pack voltage, negative voltage drop across R_L, and supply voltage (sum of latter two is tube plate voltage, of course).

The a-c variations of the television signal on grid of first tube cause similar variations in the voltage drop across the plate load resistor. Since it is a part of the series combination which supplies grid voltage to next stage these same variations are present between grid and cathode of next tube. They are not attenuated because only a d-c voltage drop exists across the plus B supply and the coupling supply. It is also evident that if the d-c component of current in the plate load resistor changes, its being a part of the same series chain would cause a similar change in the d-c bias on the grid of the next tube. This is exactly what occurs if the average brightness component of the television signal shifts. For example if the average brightness of the scene decreases for the positive-going composite signal applied to the grid of first tube the average d-c brightness level of the signal rises toward the blanking level and average plate current of the tube rises correspondingly. This change in d-c voltage drop across the plate resistor (a result of the change in average brightness) is immediately conveyed to the grid of the next tube as a change in d-c component of bias, thus effectively transferring a d-c voltage change.

In the system shown in the second drawing the plate is tied directly to grid of next tube and cathode is returned to a voltage point on the power supply bleeder system which will keep grid bias at proper negative level. This means the plate voltage on the next tube must be substantially higher if proper plate voltage is to be obtained and at same time overcome plus potential on its cathode. Thus although this system does not require the separate isolated coupling supply for each stage the supply potential required is very high and becomes higher and higher as more d-c coupled stages are added. The amplifier operation is same as previous discussion, d-c and a-c voltage drop across plate resistors being transferred to grids of succeeding tubes. Such a system is quite applicable to video amplifiers of television receivers because at the most two video amplifier stages are used and the necessary potentials can readily be obtained from the negative and positive potential low-voltage supplies.

65. *Crystal Diodes in Video Section*

The new crystal diodes are well adapted to a number of functions in the television receiver. Crystal diodes have excellent signal-to-noise ratio and low capacity, which are particularly advantageous when the diodes are employed as

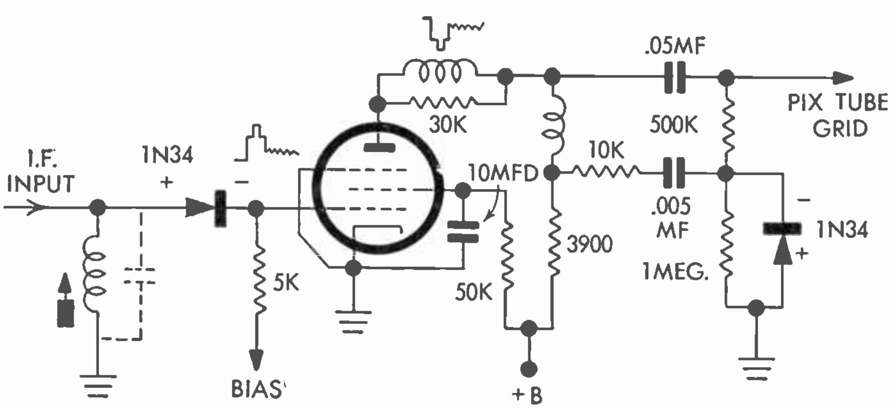


FIG. 94 Crystal Diodes in Video Section

detectors in a wide-band, high-frequency, i-f system (Fig. 94). The tuned circuit associated with the video detector, therefore, can be designed with a high L-to-C ratio. Crystal diode also has less loading effect on preceding i-f stage.

A crystal diode can also be used as a d-c restorer—advantage is again taken of its low capacity and light loading at high frequencies. In addition to its electrical advantages the crystal diode is economical, small in size, and requires no filament wiring.

66. *Picture-Tube Control-Grid Circuit*

The picture-signal circuit or control grid to cathode circuit has, when d-c coupling is used, four distinct voltages applied to it. These voltages are: alternating component of picture signal, d-c component of picture signal (average negative voltage drop across video-output load resistor), plate voltage of video-output tube, and picture-tube cathode bias supply. When direct coupling is

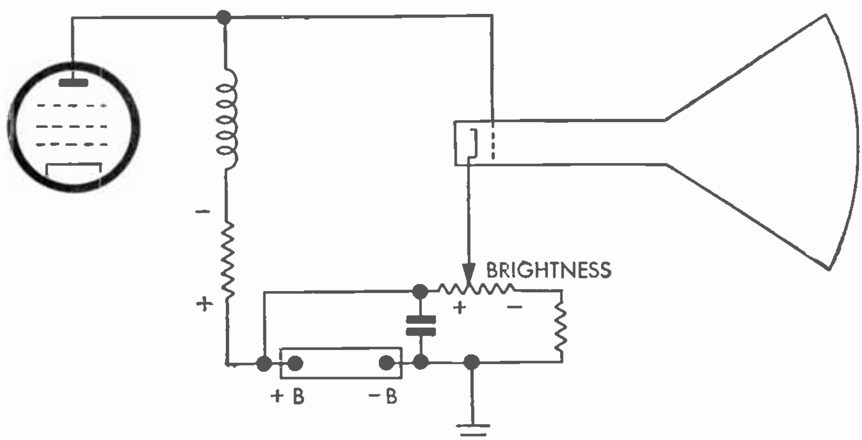


Fig. 95 Signal and Bias System of Picture Tube with D-C Coupling

used, the positive plate voltage of the output tube (Fig. 95) is applied to the grid of the picture tube. If a negative bias is to be supplied to the grid of the picture tube, it means a more positive voltage must be applied to the cathode to overcome the positive grid voltage, making the voltage from picture-tube grid to cathode negative by the proper amount.

The cathode voltage is made adjustable by means of the potentiometer. It sets the level of picture-tube brightness and is adjusted to proper degree of screen illumination for comfortable viewing.

When a diode d-c restorer is used ahead of the picture tube (Fig. 96), a much lower positive voltage is applied to the cathode, as there is no need to overcome a high, positive grid voltage. The only positive voltage applied to the grid is the d-c component of voltage developed by the conducting diode d-c restorer. Brightness control is again a potentiometer in the cathode circuit.

The a-c component of picture signal applied to the grid of the picture tube must have sufficient amplitude to drive the grid between beam cutoff, or black level, which occurs during blanking and black portions of scene, and peak illumination during the high lights of the scene. If the signal is weak, the contrast range (gradations of white to black between brightest and darkest portions of scene) is confined to a narrow range concentrated at some point in the spectrum dependent on the setting of the brightness control. Generally, the brightness will be turned up, in which case the picture will lack blacks and

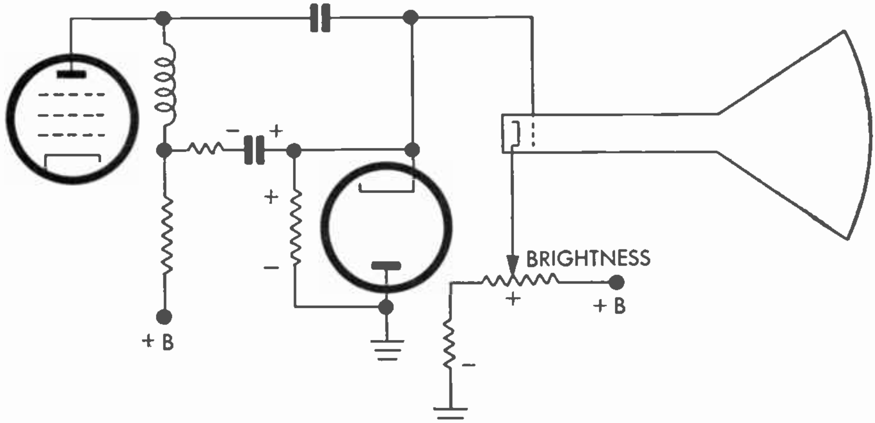


FIG. 96 Signal and Bias System with Diode-Restorer Circuit

retrace lines will be apparent. If the brightness is now retarded to remove retrace lines, the presentation will lack whites—white areas now having a gray, washed-out appearance because the signal cannot swing the grid to high-illumination levels.

If picture signal is too strong, the tube is driven too far positive and too far beyond cutoff. As a result, dark grays will reproduce black and light grays will reproduce white, thus destroying the tonal range of the picture. The high lights will also bloom and cause loss of resolution because of the intense beam and inertia of the fluorescent screen.

67. *Typical Video-Amplifier Circuits*

The General Electric video amplifier (Fig. 97) is direct-coupled from detector to the cathode of the picture tube. In this video amplifier, no d-c restorer is necessary because the output of the video detector is a single-polarity signal (with blanking and sync-tip levels constant), which is direct-coupled to cathode of picture tube. The negative-polarity composite signal which is developed across the diode load resistor, *R14,* is direct-coupled to the 6AC7 video amplifier. In fact, the diode load resistor also serves as the grid resistor of the video amplifier. Whenever modulation is detected, a negative voltage appears across *R14* and, therefore, on the grid of the video amplifier. The i-f band-pass filter consists of capacitor *C20,* inductor *L5,* and the input capacity of the video-output tube. The plate of the video-output tube is direct-coupled to the cathode of the picture tube. Inasmuch as the signal is negative on the grid of the video amplifier, it is a positive composite signal which is direct-coupled to the cathode of the picture tube. In effect, it is the same as applying a negative polarized signal to the grid of the picture tube. Series-shunt peaking is used in the output stage, and a portion of the signal developed across the plate load

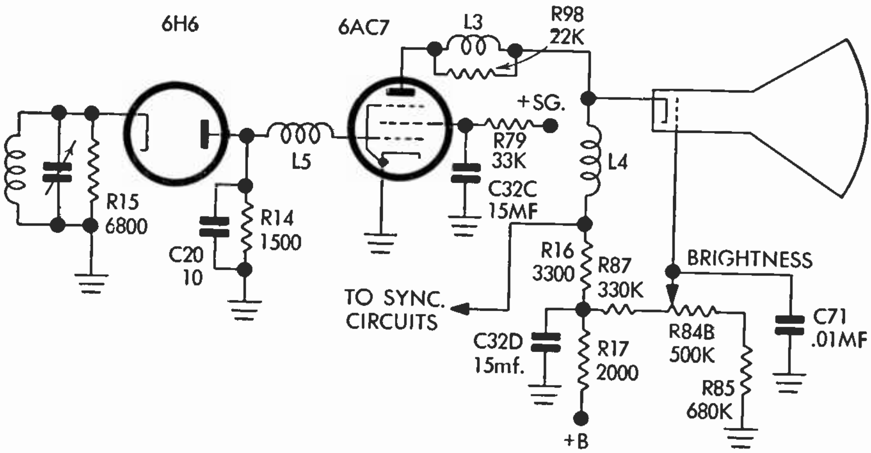


FIG. 97 GE Video Amplifier

resistor, *R16*, is also coupled into the sync circuits of the receiver. A positive voltage is applied to the grid of the picture tube through a bleeder network and the brightness potentiometer, *R84B*. The positive voltage applied to the grid of the picture tube is itself negative with respect to the positive voltage direct-coupled to the cathode of the picture tube by a proper amount to set the bias for the picture tube.

The video amplifier of the RCA television receiver (Fig. 98) consists of two video-amplifier stages. Capacitive coupling is used between the video-detector and the video-amplifier stages with a d-c restorer tube located in the grid circuit of the picture tube. This is the same type of restorer circuit discussed in the previous paragraph. A portion of the composite signal also appears across resistor *R150* and is coupled into the sync circuits. The cathode of the picture tube is grounded, and negative voltage is applied through the brightness potentiometer and the decoupling resistor, *R151*, to the grid of the picture tube. The cathode of the first video-amplifier tube is grounded and a negative voltage is applied to the grid, setting the bias for each tube. The high-frequency components of the signal are sustained by a combination series-shunt peaking system. A negative composite signal is developed across the video-detector load resistor, and it appears positive in the plate circuit of the first video amplifier and negative in the plate circuit of the second video amplifier. Consequently, a negative signal is applied to the grid of the picture tube.

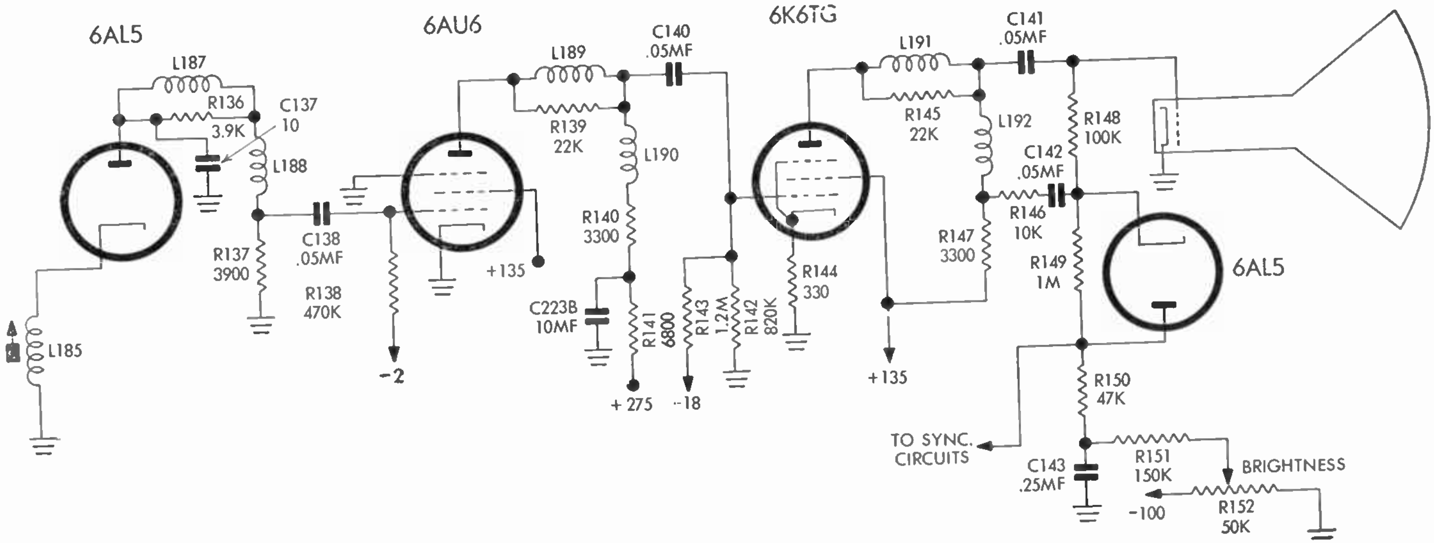


FIG. 98 RCA Video Amplifier

67a. *Additional Responsibilities of the Video Amplifier*

The video amplifier of the modern television receiver has numerous responsibilities. It must have not only a satisfactory frequency and phase response in its own right but is often planned in such a way as to correct for frequency and phase deficiencies that occur in earlier portions of the receiver. Additional functions of the video system are to increase level of video signal after detection and to retain adequate relative levels of amplification of light and dark portions of the scene. A video amplifier must be able to accommodate weak and strong signals without overload, to reduce impulse noises for all signal levels, and to retain proper average brightness.

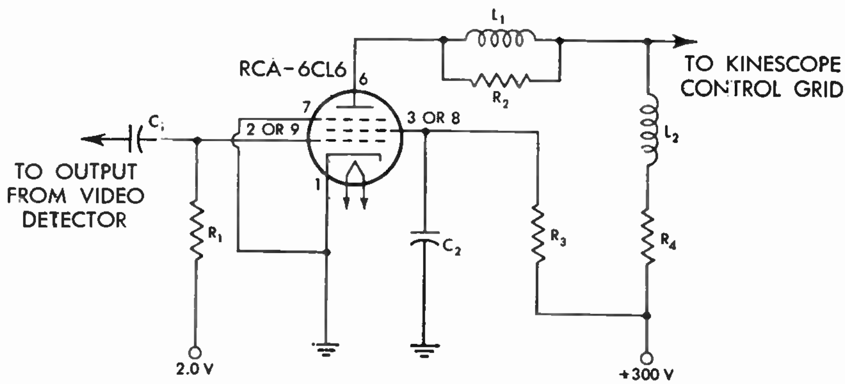


FIG. 98a RCA Video Amplifier

The amplitude of the signal at output of the video detector must be increased in level for satisfactory excitation at the control grid of picture tube. The large-screen picture tube requires a high signal amplitude (better than 100-volt peak) in order to attain full brightness range between brightest bright and darkest dark. Thus the video amplifier must have a gain of some 35 to 40 times. A two-stage video amplifier is often required to provide suitable signal level.

A new video amplifier of high *gm* and high gain has been developed for video-amplifier use. This tube, RCA 6CL6, permits a single-stage gain of some 40 to 45 times and thus enough amplification to allow for the employment of only a single video-amplifier stage in a modern television receiver. A typical video stage using this tube, Fig. 98a, has a bandwidth of 4 megacycles and sufficient gain to amplify a 3-volt signal at video detector to some 130 peak volts for delivery to picture-tube grid.

AMPLIFICATION OF DARK AND BRIGHT RANGES

In the composite video signal, Fig. 98b, the dark portions of the scene crowd around the blanking level while bright components are concentrated at the opposite voltage end of the signal. In video-amplifier design the white portions of the scene must not be compressed but should be amplified in the same proportion as is the dark region. Such a defect (compression) produces a non-linear amplification of half-tone levels. Consequently, brightness levels merge in the white regions, and half-tone gradations are lost. This defect is further emphasized by the fact that the average televiewer sets his contrast too high.

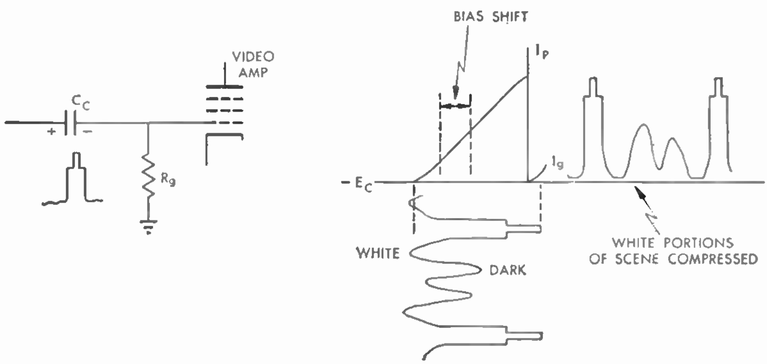


FIG. 98b Compression of Whites by Bias Shift on Bright Scene

In the design of a video amplifier some of the advantages gained by linear amplification or by even somewhat limited compression of darks are:

1. Half-tone rendition appears more realistic.

2. High-brightness gradations are emphasized to compensate to a degree for the fact that the eye is less sensitive to changes in white than to changes in black.

3. Some emphasis of whites improves the apparent picture-clarity and -crispness, because white gradations are sharper. Noise impulses swinging largely toward black are de-emphasized—a boon to fringe-area reception.

Generally, white compression is caused by operation on non-linear portions of the video-amplifier tube-transfer characteristics. This condition is at times difficult to overcome if the amplifier is to give peak results on a weak signal and yet not overload or compress on a very strong signal. It will certainly become more and more of a consideration as more stations come on the air, making local and fringe signals available in most cities.

White compression is most often caused by bias shifts that occur with changes in average brightness when positive sync pulses reach up and draw grid current and charge grid capacitor negatively. This charge sets the bias too far negative, and the whites of the picture lie in the compressed area of the tube transfer characteristic. Noise pulses can create the same disturbance, driving whites into compression and making the noise more apparent on the picture because the dark range is emphasized.

Two basic approaches, Fig. 98c, to prevention of white compression are the use of a single-stage video amplifier with proper signal polarities and proper circuits to minimize bias shifts in detrimental directions. In a single-stage video amplifier, a sync-negative signal can be taken from the video detector and applied to the grid of the video amplifier. In this arrangement the sync is negative-going and does not draw grid current, which would upset bias relations. At the same time the noise impulses are driven beyond cut-off and clipped in order not to affect biasing. The positive-going signal at the plate of a video amplifier requires that the cathode of a picture tube be excited by the video signal in order to obtain correct picture-phase.

In a two-stage video amplifier the use of d-c coupling and/or low-impedance bleeder bias for grids and screens can minimize the disturbance. Low-impedance bias sources, either by tube current-regulation or by d-c voltage changes that are a function of received signal levels, can change video-stage operating potentials and transfer characteristics, thereby holding whites on linear portions of the curve regardless of signal level and composition. Direct coupling eliminates the long time-constants that are charged by noise and cause sync-pulse sweeps into the grid current range.

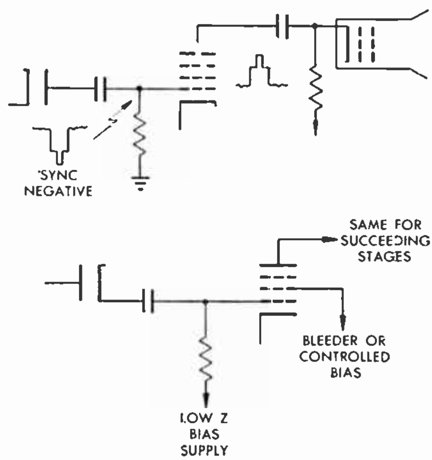


FIG. 98c Reduction of White Compression

SIGNAL LEVELS AND NOISE REJECTION

The modern television receiver is expected to give peak performance on weak or strong signals. A receiver must be able to handle a strong signal without sync clipping or serious compression of either whites or darks and with good noise-rejection. The very same video amplifier must be able to give full amplification to a weak signal and, at the same time, offer thorough noise-rejection to this low-level signal. Fortunately, a good a-g-c system in the receiver helps to equalize to a degree the signal levels at the output of the video detector.

In the utilization of a strong signal, it must be made to fit the video-amplifier transfer characteristic in order to prevent sync swinging far enough negative to clip (as per example of Fig. 98d) and upsetting the stable synchronization of receiver. At the same time it must carry near enough to cut-off to enable noise bursts to be clipped off by cut-off action. In addition, the darks and whites must not be allowed to swing into the compressed areas if true tone-gradation of the picture is to be retained.

When a weak signal is received, it must also fit the transfer characteristic in much the same manner and with the same noise-rejection action. Under this condition, however, we have a much weaker signal and grid swing between cut-off, and the top of the transfer must be less if noise pulses are to be removed by cut-off action. Cut-off must occur earlier and at a less negative grid voltage. If the tube were operated with an early cut-off for a strong signal, however, the sync would be clipped. Consequently, a means for changing video tube-transfer is expedient for peak performance on weak and strong signals.

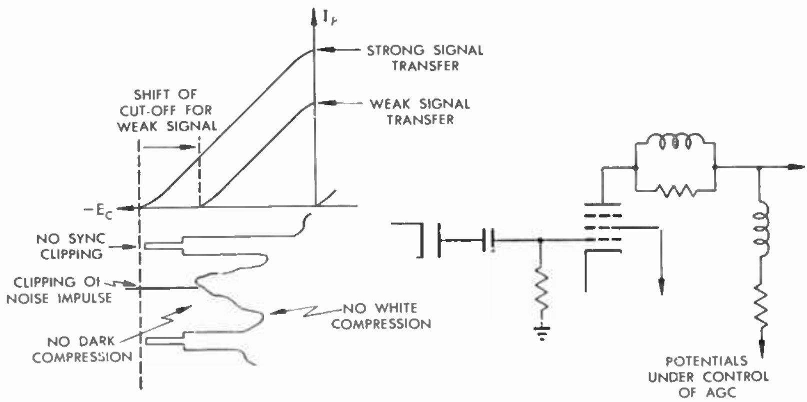


FIG. 98d Signal Level and Noise Clipping

One method employed is the use of the a-g-c system to regulate one or two video-amplifier potentials, permitting transfer to expand and compress as a function of signal level. For example, by changing the plate and/or screen voltage on a given video-amplifier tube, the bias of cut-off can be moved up or down and is thus able to meet the amplitude needs of the signal to be received.

VIDEO-AMPLIFIER CONTROLLED RESPONSE

There are so many points in a television system where bandwidth can be harmed that definite trends toward a receiver with an adjustable video amplifier are evident. Some of the many factors that influence bandwidth are:

1. Transmitted signal. There are variations from station to station in regard to phase shift and frequency response.
2. Conveyance of signals along coaxial or microwave links.
3. Receiving antenna and transmission-line response. Antenna orientation.
4. Tuner response and alignment.
5. I-F amplifier response and alignment. Response restrictions of an intercarrier system.

The video-amplifier response can be altered by use of an adjustable peaking coil or by controlled cathode frequency-degeneration.

The Philco video amplifier, using a high gain 12BY7, is excited by a sync-negative composite video signal from a crystal video detector. Consequently, white compression is absent. Positive-going signal at the plate of the video stage is applied to the cathode of the picture tube, Fig. 98e.

A series-shunt peaking arrangement, with means of control of series or shunt elements, exists at the output of the video detector. This plan permits response-correction in the video amplifier to compensate for average deficiencies at some other section of the transmission system.

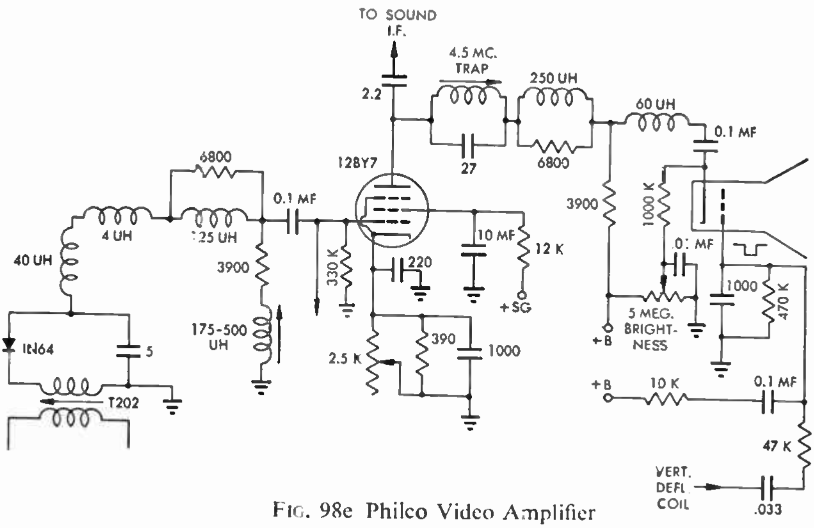


FIG. 98e Philco Video Amplifier

Contrast is controlled in the cathode circuit of a video amplifier by means of bias and degeneration control. Effective cathode resistance is variable—from near 390 ohms to zero. Resistance increase is fast initially and then slows to permit smooth contrast-control for weak or strong signals.

Retrace blanking is formed for application to the grid of the picture tube by the shaping of a pulse from the sawtooth across the vertical deflection coils. In developing the blanking pulse the sawtooth must be reduced in amplitude and then be differentiated to emphasize the sharp retrace portion. At the same time the sharp transient spike must be suppressed. Thus the network is a combination of attenuator and differentiator.

A typical RCA video amplifier, Fig. 99, receives a negative sync signal from the output of the video detector and is direct-coupled to the cathode of the picture tube. A frequency-compensated contrast control is connected in the series path between the plate of the video amplifier and the cathode of the picture tube. With contrast control R167 on maximum, the signal is coupled directly to the cathode without any attenuation. As the arm of the control is moved down the potentiometer progressively, the signal is attenuated, less signal being applied to the cathode of the picture tube. However, it is necessary to attenuate both low- and high-frequency components in like amount, and consequently, small capacitors also form a part of the attenuator circuit. As more and more resistance is added, with attenuation becoming greater, there is also a progressive increase in reactance due to the series combination of capacitors having less effective capacity. Consequently, lows and highs are attenuated by a like amount. If the capacitor-divider arrangement were not employed, the higher frequency components would be attenuated more severely than the lower ones, because of the influence of the input capacity at the cathode of the picture tube.

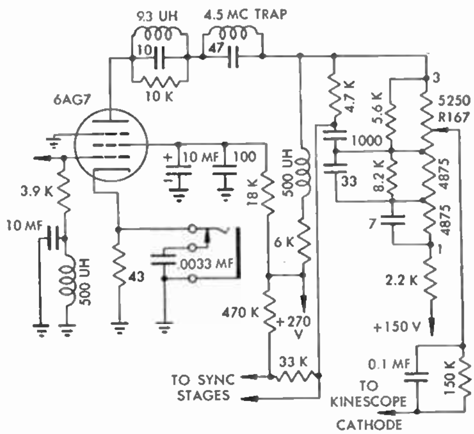


FIG. 99 RCA Commercial Video Amplifier

One of the difficulties of a direct-coupled video-amplifier system has been that the average plate current of the video-output tube changes over too great a range, and consequently, the over-all brightness of the picture changes too much with changes in average brightness and signal level. To confine this variation to a more realistic value, the network of resistors produces a voltage shift at terminal 1 of the contrast control which counterbalances the change in plate voltage of the video-output tube as present at terminal 3. Thus the effective change in cathode voltage at the picture tube is less than the change of plate voltage at the video-output tube, and the average picture brightness does not vary over as great a range. However, the direct-coupling assures a truer rendition of the picture's average brightness from top to bottom of a given scene, a quality so often lacking in a system that does not use any form of d-c restoration or coupling.

68. *Automatic Gain Control (A-G-C) Systems*

To take care of variation in signal strength, many television receivers use an a-g-c (or automatic-brightness-control) system. The a-g-c circuit varies the gain of the i-f amplifiers of the receiver and keeps the peak amplitude of the i-f signal delivered to the video detector essentially constant with substantial variations in received signal strength. The a-g-c system of a television receiver is comparable with the a-v-c system of a broadcast receiver, with the exception that control voltage is set by the peak amplitude of the received signals or the sync tip, while in the a-v-c system of a broadcast receiver the control voltage is proportional to the average strength of the received signal. In the television system, we know that the average signal strength varies with the average brightness of the scene which is transferred. This variation in average signal strength sets the average background brightness of the television picture tube, and if the a-g-c system were to operate with changes in average brightness, this relation would be disturbed. It is true that fading also causes the average signal strength to vary and would, therefore, affect the background brightness of our scene at the wrong time.

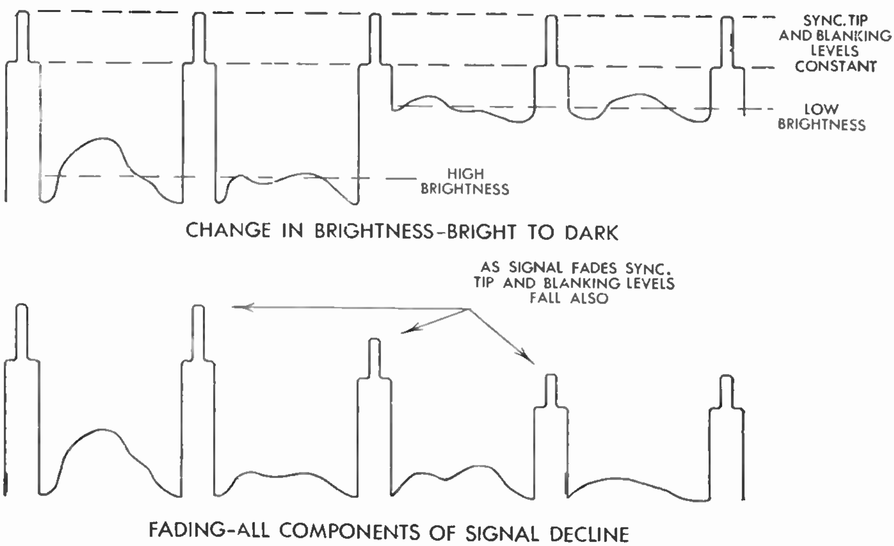


FIG. 100 Change in Signal Distribution with Changes in Average Brightness and Fading

There is one important difference between the change in average signal strength caused by fading and the change in average signal strength caused by a change in scene brightness. The peak amplitude of the television signal (blanking level and sync-tip levels) does not vary when the brightness of the televised scene changes because the blanking level and sync-tip level are held constant at the 75 per cent and 100 per cent levels, respectively, of the transmitted composite signal. The amplitude of the received signal during the blanking and sync-tip intervals is a constant, regardless of the variation in scene brightness. However, when fading is present, it affects not only the picture portion of the signal, but also causes a loss in amplitude of the blanking and sync-tip levels. This variation in the peak amplitude of the television signal by fading is the variation which is used to operate our a-g-c system in a television receiver. Therefore, the negative a-g-c voltage is proportional to the peak amplitude of the received sync tip, which, of course, is affected only by fading and not by changes in the televised-scene brightness. This point is clearly demonstrated in Fig. 100.

A simple a-g-c system is shown in Fig. 101. In this circuit, a portion of the i-f signal present at the video detector tuned circuit is coupled through a large capacitor to the plate of a diode. The i-f signal applied to this diode causes it to conduct current which charges up the large capacitor. Inasmuch as the sync tip is the highest amplitude point of our signal, the diode conducts the most current during this interval; therefore, the sync tip determines the final charge

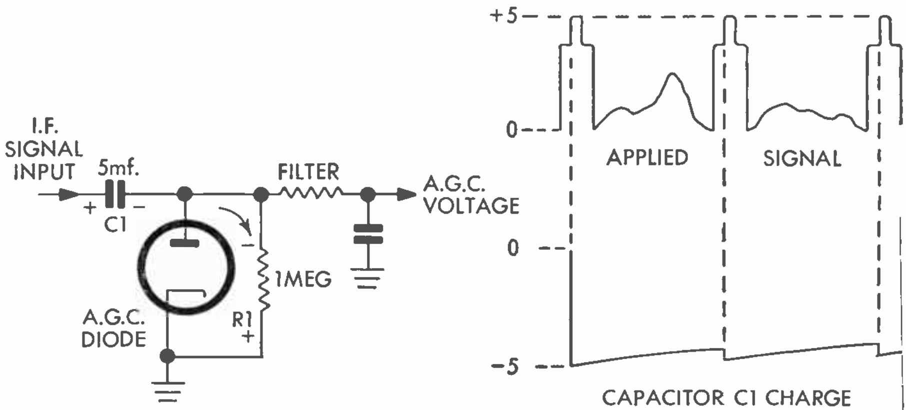


FIG. 101 Simple A-G-C System

placed on the capacitor (the final charge placed on capacitor is equal in amplitude to the peak signal voltage). Now the time constant of the RC is so very long that the peak charge placed on the capacitor holds almost at its full value between sync-tip intervals. (Refer to waveforms in Fig. 101.) Inasmuch as this charge is essentially constant and its peak amplitude dependent on the peak amplitude of the received signal, it serves as a very convenient a-g-c voltage.

If a signal fades, the amplitude of the sync tip will decrease, and in turn the peak current drawn during the sync tip will be less. Therefore, the charge on the capacitor will be slightly smaller. Again, it will be equal in amplitude to the sync-tip voltage level. Nevertheless, the final charge will not be so great as it was previously; consequently, the a-g-c voltage will be less negative. With a less negative a-g-c voltage, the i-f amplifier gain will be slightly higher. This, in turn, will amplify the i-f signal a bit more, and the strength of the i-f signal applied to the video detector will be essentially at the same level as it was previously. In summation, the function of our a-g-c system is to apply an i-f signal to the video detector which is essentially constant for substantial variation in received signal strength. This is accomplished by rectifying an a-g-c voltage which is proportional to the amplitude of the i-f signal during the sync tip.

69. *Typical A-G-C Systems*

A more elaborate a-g-c system is used in some RCA television receivers (Fig. 102). This receiver uses five tube sections to make up the a-g-c system. First, the output of the fourth picture i-f amplifier is applied to the plate of the a-g-c detector. A long time-constant circuit, *C240* and *R251*, charges up to the peak amplitude of the sync tip. The voltage which appears on the cathode of the a-g-c detector is, therefore, proportional to the amplitude of the signal sync tip, as in the case of the previous a-g-c system, with the exception that a positive instead of negative charge appears across capacitor *C240*.

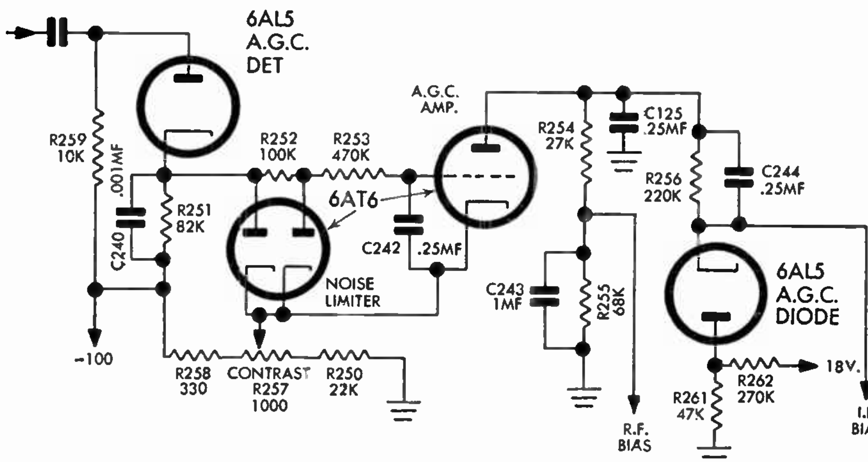


FIG. 102 RCA A-G-C System

One disadvantage of the simple a-g-c system is that noise pulses will also charge up the capacitor and operate the a-g-c system. Thus, if strong noises are received, not only do they charge up the a-g-c system, but this additional charge further reduces the amplification of the i-f system and reduces the actual signal strength. Not only is the noise present but also the signal strength of the composite television signal is reduced during the time noise is operating the a-g-c system. To prevent this defect in the RCA's a-g-c system, a noise-limiter diode circuit immediately follows the a-g-c rectifier. A sharp noise will cause the two diodes to conduct, which will shunt capacitor *C240* with a low resistance and prevent the capacitor from being charged by the sharp noise impulses. In addition, the large resistor, *R253*, and the large capacitor, *C242*, in the grid circuit of the a-g-c amplifier, will also effectively filter sharp noise impulses. Consequently, the voltage on the grid of the a-g-c amplifier will be an essentially constant voltage depending on the average charge placed on capacitor *C240*. Of course, this average charge is dependent upon the peak level of the continuously received sync tips.

The cathode of the a-g-c amplifier tube is biased negative with respect to its plate, which is at ground or near ground potential. The grid of the a-g-c amplifier is biased by two voltages, one of them the voltage of the negative supply and the other the positive charge built up on capacitor *C240*. When the positive voltage across *C240* is sufficient to overcome the negative voltage from the negative supply (depending on setting of picture control or contrast control of receiver), the a-g-c amplifier will conduct. The a-g-c amplifier conducts and causes a voltage drop across *R254* and *R255*, which makes the plate negative with respect to ground. Therefore, negative bias will be applied to the r-f bias line and the i-f bias line. The amplitude of this negative voltage will be

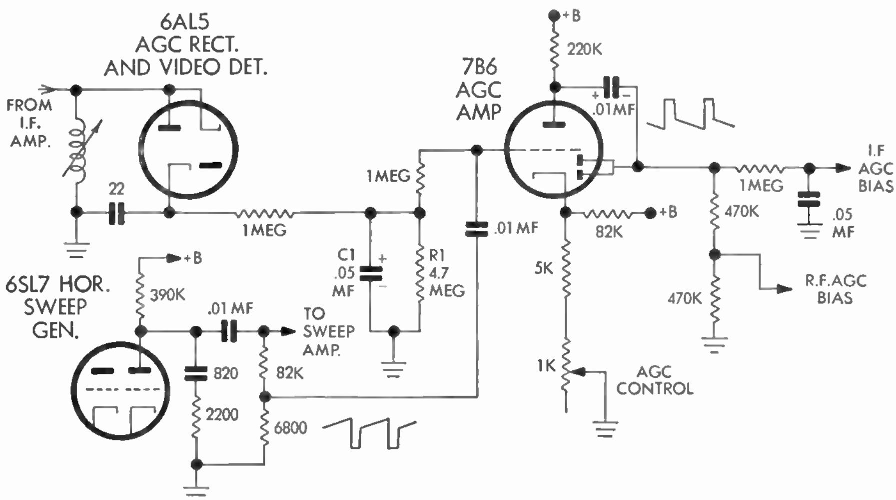


FIG. 103 Philco A-G-C System

proportional to the positive charge on capacitor *C240*, and the positive charge on the capacitor is dependent on the peak amplitude of the received signal. If the setting of the picture control is changed, the a-g-c amplifier grid bias will also change and, therefore, the negative voltage on the r-f and i-f bias line will vary. This is a very convenient method of changing the gain of the i-f amplifier and, therefore, the peak-to-peak amplitude of the detected signal.

In the reception of the television signal, signal-to-noise ratio is an important consideration, and if the gain of the receiver is to be changed, it is better to change the i-f gain and permit the r-f stages to operate at maximum gain because of the improved signal-to-noise ratio. If the signal, however, is exceptionally strong, it is possible to overload the first i-f stage, in which case the r-f gain must be reduced. The a-g-c system shown does just this—when the signal is of medium signal strength, the i-f gain is affected much more than the r-f gain, and the signal-to-noise ratio is sustained. However, when an exceptionally strong signal is received, the r-f gain is affected more than the i-f gain. This shift of control is performed with the a-g-c diode which, after its cathode reaches a certain negative value, conducts and prevents the bias from rising on the i-f bias line while it continues to rise at a more rapid rate on the r-f bias line. Thus, on a very strong signal, the current drawn through resistors *R255* and *R254* is the current of the a-g-c amplifier tube plus the current of the a-g-c diode, and consequently the bias on the r-f line rises more rapidly.

A unique a-g-c system, not readily susceptible to noise operation, is used by Philco (Fig. 103). The system consists of an a-g-c rectifier with an extremely long time constant, *C1* and *R1,* which rectifies d-c potential proportional to the average peak of the sync tip. If there is fading of the signal, the average level of the sync tips decreases, and the charge on *C1* becomes less positive.

This charge, which is proportional to sync-tip level, sets a positive bias on the grid of the a-g-c amplifier (grid held negative by additional plus bias applied to cathode), which determines the gain of the amplifier. To the grid of the a-g-c amplifier a portion of the generated and modified horizontal sawtooth is also applied, and the amount this peaked sawtooth is amplified is determined by the bias on *C1*. The positive peak of the plate-circuit waveform, applied to the diode plates, charges the a-g-c filter system.

Thus, when signal fades, the charge on *C1* decreases and the gain of a-g-c amplifier is reduced. This reduction causes less amplification of grid-modified sawtooth and the peak of plate waveform is less. A lower peak means less rectified diode current and a less negative a-g-c bias, which increases gain of r-f and i-f amplifiers.

70. *Keyed A-G-C Systems*

A further improvement in the a-g-c action in a television receiver can be accomplished by the use of a so-called "keyed a-g-c system" that acts only during the sync-pulse interval and is immune from noise impulses that occur between sync pulses. In such a system a sharp pulse is generally developed from the horizontal output circuit and applied to an a-g-c tube, permitting it to conduct only when the feed-back pulse is present.

In the RCA system (Fig. 104), a d-c bias is taken from the cathode of the sync separator and, through a proper control, is applied to the grid of the a-g-c tube. This voltage, of course, is a function of the amplitude of the sync pulse applied to the sync-separator tube and, consequently, has a value that is a function of received signal strength only. The setting of the contrast control does not affect the sync-pulse amplitude, because sync is removed from the video amplifier of this model of receiver, previous to the contrast control itself. A sharp pulse from the horizontal output circuit is applied to the plate of the a-g-c tube via a suitable capacitor-voltage divider and differentiating system. Consequently, this sharp pulse represents the only voltage applied to the a-g-c tube plate, thus permitting it to conduct on only its positive sharp spike (derived and coincident with the retrace period of the horizontal sweep system). Thus the triggering of the a-g-c tube is coincident with the arrival of the horizontal sync pulses. Likewise, the amplitude of the d-c voltage applied to the grid of the a-g-c tube is a function of the peak amplitude of the arriving horizontal sync pulses and, consequently, will change with the average change in amplitude of the arriving sync pulses, as a result of any fading or signal disturbances that affect the strength of the signal arriving at the input of the tuner.

When the a-g-c tube conducts and current flows through resistors *R1, R2, R3,* and *R4,* a negative voltage is developed in the a-g-c system and is a function of the grid voltage of the a-g-c tube. This negative charge is held between sync pulses by the capacitors in the a-g-c lines. It is to be noticed that the time constants are not excessively long and that the a-g-c system has the ability to follow rather abrupt changes in signal strength, such as might come from aircraft flutter. This type of fading occurs very quickly, as a reflected signal from a plane adds to or subtracts from the direct signal at the receiver input.

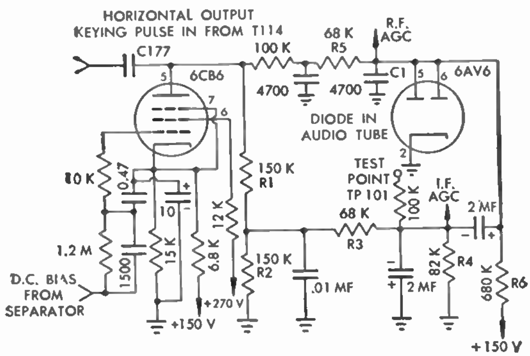


FIG. 104 RCA Keyed A-G-C Circuit

The first and second i-f stages are biased continually by the a-g-c system, regardless of incoming signal level. However, the manual preset a-g-c control permits adjustment of this bias range as a function of received signal strength for a given locality. It is to be noted that by varying the a-g-c control the d-c component of bias on the grid of the a-g-c tube can be regulated. This in turn sets the plate-current flow and the negative voltage level developed in the a-g-c system. In general, an a-g-c system of this sort is adjusted to offer maximum sensitivity to the weakest signal; at the same time, it must not overload with the reception of the strongest signal in the area. Insofar as the r-f stage of the tuner is concerned, the a-g-c bias is not applied until the incoming signal level exceeds a specified value. This arrangement permits the r-f amplifier stage to operate with the lowest possible noise factor while receiving weak fringe-area signals. Only after the signal has a level which might possibly cause overload, if it were not controlled, is a-g-c voltage applied to the r-f stage. The r-f and i-f bias ranges, as a function of received signal strength, are shown in chart form in Fig. 104a. Notice that bias is not applied to the r-f line until the incoming signal exceeds 500 microvolts.

Notice, also, that in the r-f bias line the two diode electrodes of the first audio amplifier shunt the junction point of resistor *R5* and capacitor *C1* to ground. This same junction, through the large resistor *R6*, is returned to a positive voltage source. Consequently, the diodes are conducting and place a shunt from the above mentioned point to ground, and therefore, no bias is applied to the tuner section via the r-f a-g-c line. However, if the received signal strength is high and a high negative voltage is developed by the a-g-c

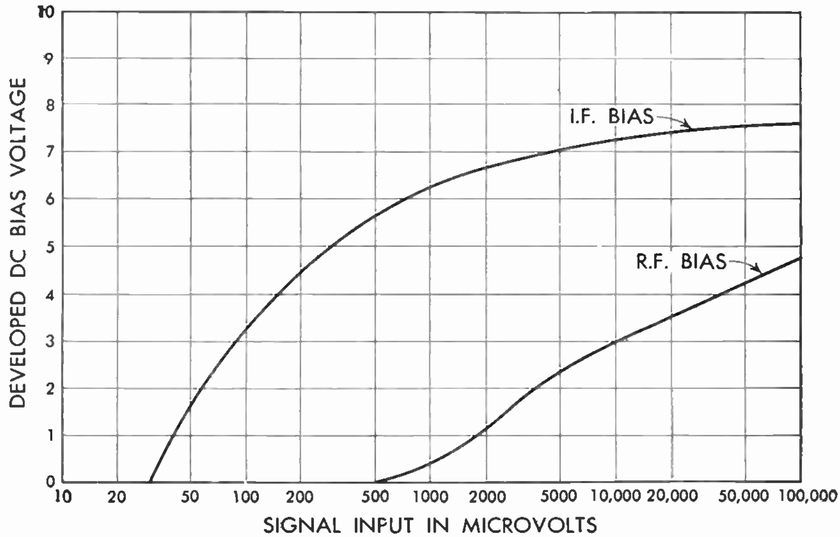


Fig. 104a A-G-C Bias Action

amplifier and appears across *C1* eventually, the diode can be made nonconducting. This occurs when the negative voltage at the junction point exceeds the positive voltage from the divider to plus B. When the diodes are nonconducting, this negative bias is applied to the r-f stage and continues to increase with an increase in incoming signal strength. This arrangement prevents overloading of the input to the i-f amplifier system with the reception of very strong signals.

In the Philco a-g-c system the necessary voltages are developed in the grid circuit of the sync separator, Fig. 105. Here, too, the peak grid-current flow is a function of the peak amplitude of the sync pulses arriving from the sync amplifier. The resultant grid charge is a function of peak amplitude and varies according to any fading or variation in this peak level. Suitable time constants are charged with this rectified grid-current flow and, via the divider network, are applied to either the i-f amplifier grid or the grid of the r-f stage in the tuner. In this system, too, it is necessary to prevent biasing of the tuner until the incoming signal rises above a prescribed level. Otherwise, the tuner would have a poor signal-to-noise ratio for reception of a weak fringe-area signal.

Again a diode and a large series resistor to plus B prevent the application of bias to the tuner until the incoming signal is sufficiently strong to put a high enough negative charge on capacitor *C1* to effect cut-off of the diode and prevent its operation. At this level a negative voltage goes out through the output filter to the tuner a-g-c input circuit. When the incoming signal is weak, the plate of the diode is held positive by the bleeder network and shunts the output of the a-g-c circuit.

The i-f amplifier receives a-g-c voltage continuously in order to stabilize signal level. However, for operating in levels where the signals are very weak a special fringe-local switch is included. On the fringe position an additional one-megohm resistor is shunted from the a-g-c output to ground. Consequently, the a-g-c bias is reduced, and the i-f amplifiers are permitted to operate with maximum sensitivity for the reception of very weak signals. Some a-g-c action is still retained, but it is made to operate over a range more suitable for fringe-area reception and for variations in a fringe-area signal.

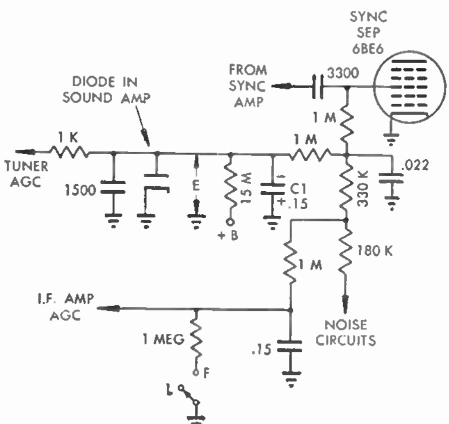


Fig. 105 Philco Keyed A-G-C System

A unique double-delayed a-g-c system is employed by Zenith, Fig. 105a. In this system, too, the r-f amplifier of the tuner does not receive any a-g-c bias until the incoming signal exceeds approximately 500 microvolts. Thus, optimum signal-to-noise ratio is retained for best fringe-area reception. The i-f amplifier, however, receives a-g-c bias for both weak and strong signals. In the a-g-c system employed, the keying pulse is derived at the input to the horizontal discharge tube in the form of a positive pulse that drives the a-g-c tube into conduction during the retrace time of the horizontal deflection. The video signal is direct-coupled from the first video-amplifier stage and is applied to the grid of the a-g-c tube, causing conduction-flow in proportion to the peak amplitude of the applied sync pulse. The video signal is direct-coupled to this grid, and therefore, the sync tip remains essentially constant in its peak amplitude with relation to picture variations and average brightness. However, the peak amplitude of the sync pulse does vary as a function of signal strength, and therefore, peak current-flow is a function of signal amplitude, setting the peak amplitude of the a-g-c current which can flow only during the horizontal retrace period when the positive pulse is being applied to the plate of the a-g-c tube. Thus the current-flow through resistor *R4* and the bias voltages established on the various a-g-c filter capacitors are strictly functions of the peak amplitude of a-g-c tube current which is, as explained, in turn a function of received signal strength. Inasmuch as this current-flow occurs only during the horizontal sync interval and because the noise is clipped off the top of the sync pulse, the charge is strictly a function of signal strength and is little influenced by noise that occurs *between sync pulses* or *during sync pulses.*

The a-g-c tube itself operates near ground potential, as can be determined by observing the methods used to supply electrode voltages. For example, cathode voltage is determined by a bleeder network to plus B as well as by the amount of current flow. However, the direct connection between cathode and plate via resistor *R4* keeps the difference of potential between plate and cathode small. The grid receives its bias through a divider network, consisting of resistors *R2* and *R3* from the plate supply of the first video-amplifier tube.

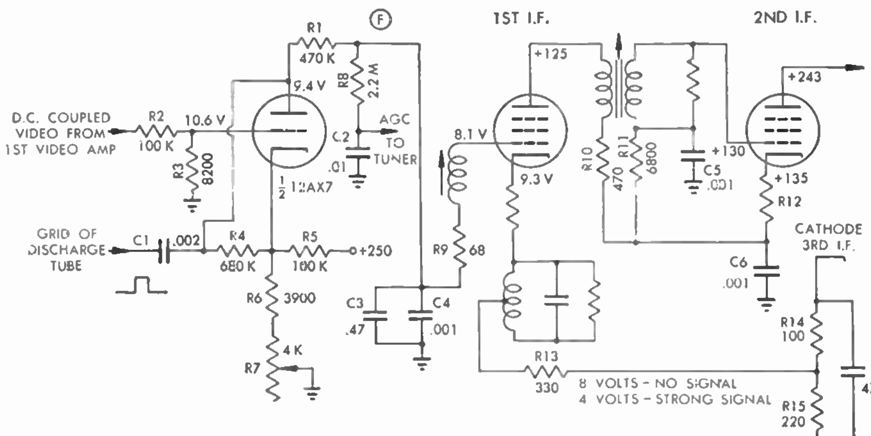


FIG. 105a Zenith Keyed A-G-C System

Thus the a-g-c tube is biased slightly negative (grid to cathode), and the plate-to-ground potential can vary as a function of the amount of plate-current flow; since this potential can be either negative or positive with respect to ground, point *F* is biased negatively or positively as a function of received signal strength. This arrangement permits delayed a-g-c action in the a-g-c line. When a weak signal is received the a-g-c current is so low that the voltage drop across resistor *R4* holds the plate potential between approximately plus 8 and 9 volts, and point *F* is at a positive potential. Inasmuch as the grid of the r-f amplifier of the tuner employs contact bias (grid current-flow through large-value resistor *R8*), the positive potential at point *F* contributes no bias to the r-f stage. Therefore, the r-f stage operates with the very best signal-to-noise ratio for reception of a weak signal.

The cathode of the first i-f stage, however, is supplied with a positive voltage from the cathode circuit of the third i-f tube, and this positive voltage is high enough to match the positive voltage applied to the grid of the first i-f tube from the a-g-c line in order to set normal i-f stage bias. Consequently, the bias on the first i-f stage does vary as a function of amplitude of the incoming weak signal. Inasmuch as the first and second i-f stages are effectively in series, in relation to the plate supply source, a change in the bias of the first i-f stage will also change the bias of the second i-f stage. Consequently, the a-g-c bias change influences both i-f stages and causes a-g-c action with reception of either weak or strong signals. A-G-C action is, of course, advisable even on the reception of a weak signal because the average brightness of the picture is held at a more constant level with impulse noise interference, aircraft flutter, and other signal level variations. A reasonably short time constant is used in the i-f bias line to permit a-g-c action on reasonably fast changes of received signal level.

The range and speed of a-g-c action are also aided by the special cathode-biasing system used for the first i-f and third i-f cathode circuits. The cathode bias for the first i-f stage is derived from a voltage-divider network in the cathode circuit of the third i-f. The voltage at the junction of the two cathode resistors is also a function of signal level, being approximately 8 volts for no signal and declining to 4 volts with the reception of a strong signal. This voltage is applied to the cathode of the first i-f stage and causes the cathode voltage to rise or fall as a function of received signal strength; thus a-g-c activity is conservative in the reception of weak signals but becomes more and more pronounced and active as stronger and stronger signals are received.

When a strong signal is received, the a-g-c tube draws a substantially higher plate current because of the rise in peak amplitude of the sync pulse, developing a negative voltage with respect to ground at point *F*. This negative voltage through resistor *R8* is now applied to the grid of the r-f amplifier stage and permits a-g-c action in the tuner when very strong signals are received. Do understand that when point *F* is positive the contact potential dominates and no a-g-c action occurs in the tuner, but when point *F* becomes negative, a-g-c bias dominates and adds negative bias to the r-f amplifier grid. The same negative bias also reaches the grid of the first i-f stage and drives that grid in the negative direction, reducing stage gain. However, as signal level increases, the cathode of the first i-f stage becomes less positive and follows to a degree the change in grid bias. However, as the signal becomes stronger and the negative grid bias increases, the cathode voltage does not follow in the same proportion. When very strong signals are received the grid bias of the first i-f stage and, indirectly, the bias on the second i-f stage are driven far negative. This controlled action plus the bias supplied to the tuner prevent overloading of the i-f system with the reception of a strong local signal and, at the same time, permit peak sensitivity for the reception of a fringe-area signal in the same location. A-G-C action, as a function of the strongest and weakest signals to be received at a given location, is set at an optimum value with the cathode potentiometer in the a-g-c circuit. This control, which is a part of the bleeder network to the supply voltage, is set at an optimum level to allow proper a-g-c action for the reception of a weak signal and no overloading with the reception of a strong local signal.

QUESTIONS

1. Explain how high-frequency components of a signal are lost through distributed circuit capacity.
2. What circuit elements cause loss of low frequencies?
3. Explain how low-frequency components of a signal are lost.
4. Explain how a shunt-peaking coil improves the high-frequency response.
5. Differentiate between series and shunt peaking.
6. Explain action of low-frequency compensating circuit.
7. Of what importance is phase response in the television system?
8. How is it possible to obtain either a positive or negative output from the video detector?
9. What are the characteristics of a commercial video detector?
10. Why is the blanking level constant regardless of picture content at the output of the video detector?
11. Why is d-c restoration necessary?
12. Why isn't d-c restoration necessary in some receivers?
13. Explain operation of grid-rectifier type of d-c restorer.
14. Explain operation of diode-type restorer in grid circuit of picture tube.
15. Describe a commercial video amplifier.
16. How is correct polarity signal applied to picture tube?
17. Explain function and location of *contrast* and *brightness* controls.
18. What voltages determine picture-tube grid bias when video output tube is direct-coupled to picture-tube grid?
19. Does the picture-tube grid bias vary or remain constant with changes in scene brightness?
20. How is the blanking level held constant when scene brightness changes?
21. Where is the sync tip utilized in the video amplifier?
22. How does a-g-c and a-v-c differ?
23. Why is noise a consideration in an a-g-c system?
24. Explain operation of a commercial a-g-c system.
25. Explain operation of intercarrier system.

Chapter 6

TELEVISION PICTURE TUBES

71. *Basic Construction*

The television picture tube is very similar to a test oscilloscope tube and consists of an electron gun which forms and converges electron emission into a narrow, small-diameter beam of electrons, a fluorescent screen which illuminates under impact of this small-diameter beam, and, last of all, a deflection system which moves the beam to various positions on the surface of the fluorescent screen. The electron gun of the picture tube consists of a cathode which emits the electrons and a number of cylindrical elements which converge the electrons into a pin-point beam of electrons. The electron gun is really an electronic lens and functions in the same manner as an optical lens system, which gathers light rays and focuses them into a central point of light.

To obtain a well-resolved scene, it is necessary that this stream of electrons be concentrated into an extremely tiny spot at the surface of the fluorescent screen, in the same manner as an optical projection system focuses on the viewing screen. The beam of electrons is focused by an electrostatic field within the electron gun or by means of a magnetic field which is generated by an external focusing coil. Inasmuch as the picture is not presented instantaneously on the fluorescent screen, as in an optical projection system, a means is provided for moving the beam across and down the screen to paint the picture, element by element. Again, the beam is made to scan by means of a changing electrostatic field between two deflection plates or a changing magnetic field generated by external deflection coils. The fluorescent screen is a chemical deposit on the inner surface of the glass at the front of the tube. Each element of the fluorescent screen glows when it is struck by the electron beam. The degree of illumination depends on the velocity and number of electrons which strike it. Elements remain illuminated for a finite time, depending on their chemical composition, after the passage of the beam. This persistency of the screen, plus the ability of the eye to retain an impression of light after the source has been removed, causes us to see the picture as an entirety and not as a progressive series of light variations.

The functions of the various electrodes of the picture tube (Fig. 106) are as follows:

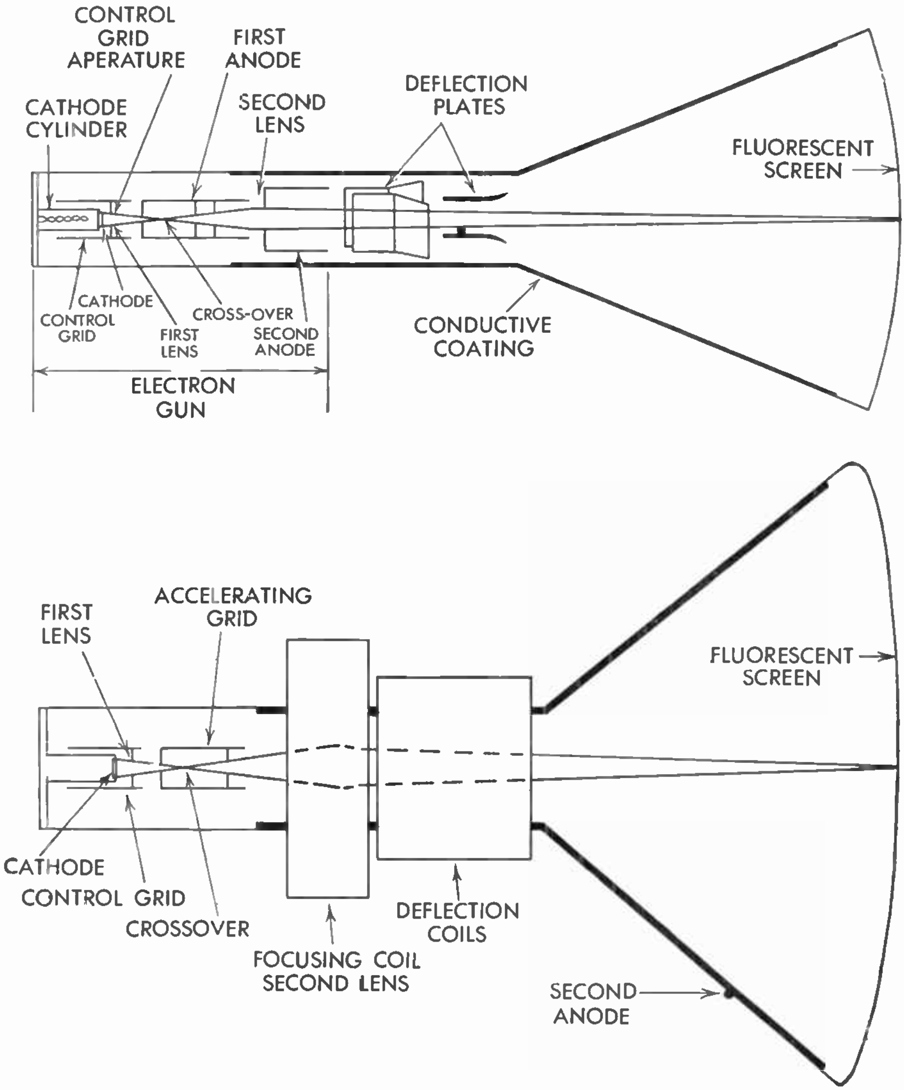


FIG. 106 Electric and Magnetic Focusing and Deflection for Picture Tubes

Cathode. An indirectly heated cathode is a source of electrons for the picture tube.

Control grid. The control grid of the picture tube has a function which is similar to the control grid of a conventional vacuum tube. It controls the number of electrons which speed toward the fluorescent screen. However, the control grid of a picture-tube electron gun is not a wire mesh, as in a vacuum tube, but a cylindrical, metallic electrode with a small opening at the end for passage of electrons. The number of electrons which pass through the opening is determined by the instantaneous potential of the electrode, the flow of electrons varying with the rise and fall of the signal, if any, applied to the control grid. In the picture tube the average intensity of the presentation on the screen is determined by the d-c bias which exists between control grid and cathode.

First anode. The first anode, another cylindrical element with narrow apertures, further concentrates the beam and, by reason of its higher potential, draws the electrons away from the control grid and speeds them toward the fluorescent screen. Actually, the first anode is called the *focusing anode* because the difference of potential which exists between it and the second anode, and to a certain extent the potential difference between it and the control grid, concentrates and focuses the electron beam to a fine pin point at the fluorescent screen. To permit fine adjustment, the first anode potential is variable over a limited range.

Second anode. The second anode is the final cylindrical electrode; it gives the final acceleration to the electrons. In fact, the final velocity is determined principally by the second anode, and only to a small extent by the other electrodes. It causes the electrons to hit the fluorescent screen with sufficient impact to make it fluoresce, and at a sufficiently perpendicular angle to prevent grazing at the outer extremities of the screen. The second anode extends upward toward the fluorescent screen in the form of a conductive coating on the inner glass walls of the picture tube. This extension of the second anode, sometimes called a *collector anode,* collects the secondary electrons knocked off the fluorescent screen by the electron beam.

Horizontal deflection plates. The horizontal deflection plates, mounted upright on the left and right sides of beam path, reflect the electron beam horizontally in accordance with the difference of potential existing between the two plates.

Vertical deflection plates. The vertical deflection plates, mounted horizontally above and beneath beam path, deflect the electron beam vertically in accordance with the difference of potential between the two plates.

Fluorescent screen. The fluorescent screen, a chemical coating called a *phosphor,* is on the inner glass surface of the picture tube. It is insulated from the collector-anode coating. The chemical composition of the coating causes the screen to illuminate in accordance with the velocity and strength of an electron beam which strikes it. The intensity of the illumination is dependent mainly upon the grid bias (number of electrons striking it) and the second anode potential (velocity of electrons striking it). The average illumination of the screen is varied by changing the d-c grid bias; instantaneous variation of the screen illumination can be caused by a varying grid signal.

Magnetic focusing. In a magnetically focused picture tube (Fig. 106*B*) the first anode is replaced by an accelerating grid which gives the initial acceleration of the beam electrons. The beam is focused by a magnetic field generated by an external focusing coil. The second anode in this type of picture tube is not always a part of the main section of the cylindrical electron gun. Instead, it is in the form of a coating on the inner surface of the glass extending from the neck of the tube up toward the fluorescent screen.

Magnetic deflection coils. In a picture tube using magnetic deflection, the deflection plates are replaced by external deflection coils. A pair of horizontal and vertical deflection coils generate the magnetic fields which penetrate the glass and deflect the beam horizontally or vertically.

72. *Electron Emission*

In television picture and camera tubes three major types of electron emission are utilized. These are thermionic, photoelectric, and secondary emissions. In the image-orthicon camera tube, all three types of emission are employed, while in the picture tube two of these types, thermionic and secondary emission, are present.

Electron emission can occur off most types of metal; however, there are only a few pure metals which have appreciable and a practical amount of emission. More efficient emitters are obtained by depositing a thin oxide or other type of coating on the pure metal.

At the surface of a metal there exists a tension which prevents the transfer of the electron in the metal to those of the other medium. This is a so-called "potential barrier" which prevents the break-up of the metal and its dispersal into another medium. In the case of a gas, this potential barrier is weak and easily overcome, causing the gas to mix freely. Emission occurs from the metal when a force is generated which is sufficient to overcome the potential barrier at the surface of the metal. This barrier normally prevents escape of free-moving electrons from the metal. If the electrons are given increased energy they can overcome and escape the barrier when they reach the metal surface. The barrier is overcome by heat, incident light, or the impact of other electrons.

THERMIONIC EMISSION

Thermionic emission occurs when the metal is heated to an extremely high temperature, thereby overcoming the potential barrier. Inasmuch as appreciable emission is obtained only at extreme temperatures, it is necessary to mount the emitter in an evacuated glass or metal bulb to prevent oxidation of the metal. This is our vacuum tube. The three types of thermionic emitters are: certain pure metals (tungsten is the most widely used of this type); composite surfaces, such as thorium on tungsten; and oxide-coated emitters. The tungsten and thorium-tungsten heaters are widely used in transmitter tubes; oxide-coated cathodes are used in small tubes and television pickup and receiving tubes.

The oxide-coated cathode is of a rather complicated chemical composition consisting of a nickel base and a combination barium- and strontium-oxide coating. The oxide-coated cathode is capable of very efficient emission at relatively low temperatures and, consequently, it is ideal for small, low-current tubes. In high-voltage transmitting tubes the thoriated-tungsten type has preference because of its high-current capabilities, though at a reduced efficiency. Oxide-coated cathodes cannot be used with high anode potentials. The typical cathode system of a picture tube (Fig. 106) consists of a tungsten wire heater surrounded by a nickel shield. The actual cathode or emitting surface is located at the end of the nickel cylinder. The oxide-coated emitting surface is in the form of a small disc in the case of a picture tube because only with a small area emitting surface is it possible to converge the emitted electrons into a small-diameter beam of electrons.

One of the defects of thermionic emission is its random emission of electrons. This very slight variation in the number of emitted electrons introduces a noise component into the output. Fortunately, in most applications of the vacuum tube it is a very tiny variation as compared to normal signal levels. However, we know now that at very high frequencies and with extremely weak signal levels this random emission of electrons introduces a dominant noise signal into the receiver. This is also a dominant noise in camera pre-amp tubes because they, too, are extremely sensitive to a weak signal and must amplify signals no greater than a few hundred microvolts.

PHOTOELECTRIC EMISSION

A second method of overcoming the potential barrier of certain chemical compositions is by means of incident light waves. This is a system used to free electrons from the photocathode of the image orthicon and from the mosaic of the iconoscope. Emission of electrons occurs when the light rays impart additional energy to the free-moving electrons of the metal. Electrons now have sufficient energy and momentum to overcome the potential barrier at the surface.

Most photoelectric surfaces consist of a silver base, or silver-coated surface, with caesium interspersed with the silver particles. A film of photosensitive caesium oxide covers this layer.

SECONDARY EMISSION

The final type of emission to be considered is secondary emission of electrons. This type of emission depends on collision between incident electrons and free-moving electrons of the metal. An incoming electron knocks off one or more electrons from the surface. The actual number of electrons removed is dependent on the velocity and number of incident electrons, and on the type of surface.

Secondary-emission surfaces studied in connection with camera tubes are the mosaic of the iconoscope and the target of the image orthicon. The multiplier of the image orthicon also consists of a series of electrodes which, when struck by incident electrons, emit a greater number of secondary electrons which move on to the next electrode. The fluorescent screen of the picture tube also has secondary emission characteristics which must be considered, especially in the construction of projection television picture tube.

73. *Electron Guns*

The electron gun of the picture tube guides and concentrates electrons emitted from the cathode. Actually, the electron gun is an electronic lens system, consisting of a first and second lens, which takes the electrons emitted from the cathode and focuses them to a fine point on the fluorescent screen. The potential gradients between electrodes form an electronic lens, shown in Fig. 107, which can be compared with the conventional optical lens system. Just as an optical lens takes the parallel rays of incident light and concentrates them into a fine focal point, the electronic lens takes the incident stream of electrons and concentrates them into a fine crossover point. This electronic crossover point is then focused on the fluorescent screen by a second electronic lens, just as the second optical lens system focuses the rays of light on some remote point.

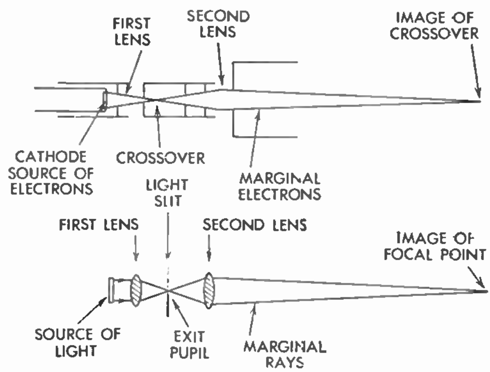


FIG. 107 Comparison of Electronic and Optical Lens Systems

The picture-tube gun (Fig. 106) consists of a heater, cathode emitter, and control grid, which form the first lens of the electronic lens system. The first and second anodes of the electrostatically focused picture tube form the second lens. If magnetic focusing is used the second lens consists of an external coil surrounding the neck of the tube, which generates a magnetic field that penetrates the glass and focuses the electron beam. The correct focus is obtained by properly positioning the magnetic focus coil and by controlling the intensity of the magnetic field which the coil generates.

FIRST LENS SYSTEM

The picture-tube gun generally consists of a tungsten heater and a nickel cylinder at the end of which is located the cathode emitter, consisting of a barium or strontium oxide. The area of the cathode emitter is chosen to have the proper emission at required spot size. The larger the cathode area the higher is the beam current and the more difficult it becomes to concentrate the beam into a tiny spot. Practically, an optimum value of cathode area must be chosen according to the beam current necessary and the spot size required. The picture-tube spot must be fairly large and the beam current reasonably high to properly excite the fluorescent screen. Picture-tube cathode area is larger than that of camera tubes. So far as the iconoscope and image-orthicon beam is concerned, we are primarily interested in obtaining a small spot size and, consequently, the cathode area must be reduced.

The control grid is a cylindrical element with a small aperture through which the electron beam passes. This aperture assists in holding the spot size down by stopping those electrons which are diverted from the beam. The picture-tube control grid also has a skirt on it which extends past the aperture toward the following anode. This skirt increases beam current and improves the control characteristic producing a change in beam current and illumination, which is reasonably linear with applied control grid signal. The various electron-gun electrodes are constructed of nickel tubing or a nickel bronze which is nonmagnetic.

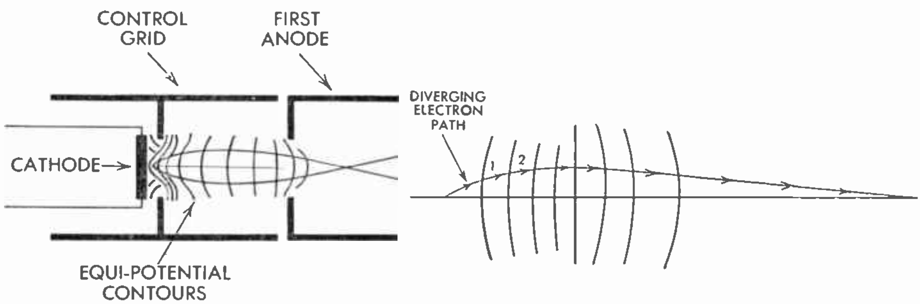


FIG. 108 Voltage Gradients of First Lens

First anode, control grid, and cathode are the first lens of the electron gun. Its effect on the beam is first to convert it into a small spot called the *crossover*, which is considerably smaller in diameter than the cathode emitting surface. This spot, or crossover, is then imaged or focused by the second lens onto the screen.

A better understanding of the operation of the electronic lens can be obtained by closely studying Fig. 108. Actually, the portion of the electron gun between the cathode and the first anode represents the lens. It is the changing potential gradient between the electrodes which causes any electrons that leave the beam to take a parabolic path back to the center.

The curved contours are lines of equal voltage which become progressively higher in amplitude as we approach the first anode. Contours are not straight but misshapen, in accordance with the physical construction of the electrode and their apertures plus the ratio of the voltage between the electrodes, in this case, first anode, control grid, and cathode. It is apparent that all electrons moving down the center of the electron gun will continue in that path because of the equal potential attraction on all sides of the center by the electrodes when the beam is passing through them. It can be seen that the voltage-gradient contours between electrodes are also essentially straight at the center. Those electrons which are emitted at other than the center of the cathode surface, or are emitted from the cathode at an angle which is not perpendicular, are converged back to the center of the beam by the lens system. Actually there are two forces attracting these electrons. These are the actual attraction and acceleration caused by the second anode potential and also the attraction of the walls of the electrode through which the electron is passing. For now, you see, the electron is situated in some portion of the area between the cylindrical wall at which there is not an equal attraction on all sides of the electron, the electron, of course, wanting to move toward the most positive side. Consequently, if there were no lens system to pull these electrons back into the beam they would eventually alight on one of the cylindrical electrodes.

As is to be noticed in Figs. 108 and 109, the contour lines on one side of the space between electrodes fold over in one direction and in just the opposite direction on the other side. Thus an electron entering the contour field on the left and moving in a direction divergent from the center attempts to move always toward the nearest, most positive point. If the electron did not have an initial velocity and direction, it would take a perpendicular path between contour lines and be rapidly drawn to center axis because of contour fold. Initial motion and influence of focusing field (infinite number of contours could be plotted) bend the electron path around (second drawing of Fig. 108) until the electron is moving along a path parallel to center axis, and finally moving toward center axis. The electron, however, must not reach center until the crossover point in case of first lens, and not until it reaches fluorescent screen in case of second lens. To make certain the electron reaches center at the proper distance from gun and simultaneously with other electrons, the contours on the other side of the space between electrodes have opposite fold to keep the electron moving at the proper angle so it will reach center at the correct distance.

The electrons pass through the separation between the electrodes of the first lens where forces exerted by the electric field (forces represented by the constant-voltage contours of the figure) accelerate them in the direction of the highest potential points. As shown, these forces cause the beam to converge on a point near the first anode. This point of convergence, the crossover point, is much smaller in diameter than the cathode surface itself. It is this crossover point which the second lens focuses on the fluorescent screen to produce a much smaller spot than if the cathode surface spot was focused on the fluorescent screen. The small aperture in the control-grid electrode further confines the beam to a small spot diameter by removing those electrons which are too divergent from the center of the beam. Inasmuch as the beam electrons are most dense at the center these few electrons which strike the defining aperture and are lost do not seriously reduce the intensity of the beam. Their removal, however, considerably reduces the beam diameter.

SECOND LENS SYSTEM

The second lens of the gun images the crossover on the fluorescent screen. In the case of the picture tube which is focused electrostatically (Fig. 109) the first and second anodes form the second lens. The voltage ratio between the first and second anodes determines the distance at which the crossover is focused. Consequently, the first anode voltage is variable by means of the so-called "focus control" of the television receiver, and is adjusted until the picture is brought to sharp focus on the fluorescent screen. When magnetic focusing is used (Fig. 109) the current through the coil and the position of the coil on the neck of the tube determine proper focusing of the picture.

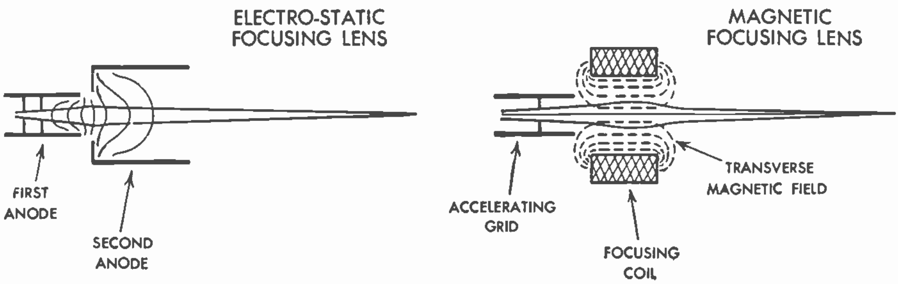


FIG. 109 Second Lens Systems

Again, after leaving the crossover point, the electrons on the outer portions of the beam again diverge (mutual repulsion between electrons) and move toward the first and second anode wall. It is necessary, therefore, to bring these divergent electrons back into center and have them join the center of the beam at the fluorescent screen. It is the function of the field distribution between the second and first anodes again to bring the divergent electrons back into the center of the beam. As shown in Fig. 109, a set of equal potential contours can be set up for this lens showing how the divergent electrons, because they are pulled on by two forces, can be once more brought into the center of the beam at screen. The first anode of the electron gun (Fig. 106) contains three apertures which further confine the beam by removing those electrons which are too divergent from the dense portion of the electron stream. The first aperture of the first anode has a very small diameter while the latter two apertures are somewhat wider. This is necessary because the beam, in passing through the first anode, is diverging until it reaches the second lens. The presence of the apertures in the electron gun also prevents, to a certain extent, an electronic lens defect called *spherical aberration*. This is the same defect encountered in a conventional light lens system in which the light rays passing through the outer portion of the lens do not come to focus at exactly the same point as those light rays which pass through the center portion of the lens. The removal of those electrons which pass through the outer fringes of the electronic lenses by the defining apertures absorbs these electrons which would cause defocusing of the spot at the fluorescent screen. The second anode in many guns also contains a single aperture.

MAGNETIC FOCUSING

When magnetic focusing is used, the divergence of the electrons after the crossover is acted upon by magnetic lines which penetrate the glass from the external coils. These magnetic lines cause the diverging electrons to take a helical path back to the center of the beam, the convergence point again

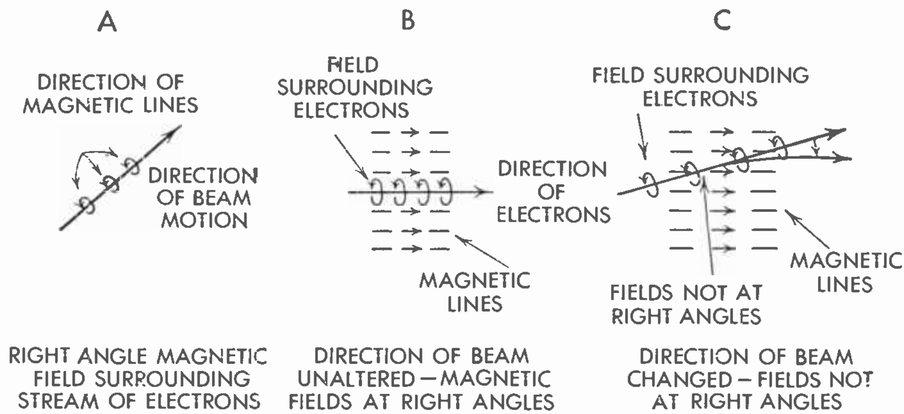


FIG. 110 Magnetic Fields and Beam Motion

occurring at the plane of the fluorescent screen. In a magnetically focused picture tube the second anode is the conducting coating on the inner surface of the glass near the fluorescent screen. The first anode is replaced by a so-called "accelerating grid" which is the first attraction for the electrons leaving the cathode emitter.

An understanding of the action of the magnetic lens can be grasped quickly when the following basic magnetic laws are understood. When electron flow exists in a long, straight conductor, a magnetic field is established around the conductor. Direction of these magnetic lines looking toward the direction of electron flow is counterclockwise (drawing A of Fig. 110). Thus, if there is electron motion from A to B, magnetic lines are as shown. This very same rule applies to a narrow beam of electrons which are being accelerated by an anode. Therefore, in drawing A, the direction of the magnetic lines surrounding the electron beam is indicated by the small arrow. When *two magnetic fields* are *perpendicular* to each other there is no reaction between them. Thus, if there is a magnetic field parallel to a direction of electron motion (drawing B of Fig. 110), there is no interaction between this magnetic field and the electron stream because the field surrounding the electron stream is at right angles to the magnetic field. If there are any electrons in the beam which are not parallel to the direction of flow, the magnetic field generated by the external coil will act upon them.

When a conductor carrying current is in a magnetic field, there is a force exerted on the conductor attempting to move it at right angles to both the direction of current flow and the lines of force of the magnetic field. This rule also applies to an electron stream, the actual direction of the stream shifting. This principle will be discussed in more detail in connection with magnetic deflection of the electron beam.

These magnetic principles are used in the focusing of a beam of electrons by a magnetic lens. In this task we rely mostly on the characteristic that there is no reaction between perpendicular magnetic fields, but as soon as the angles depart from 90 degrees there is reaction. The focus coil is mounted on the neck of the tube in a position which produces a longitudinal magnetic field—that is, lines of magnetic force parallel to the electron beam. This means the magnetic field surrounding the electron beam is perpendicular to the focus field and there is no interaction. Thus, those electrons which move in a straight line through the electron gun toward the fluorescent screen are not affected by the focusing field.

However, when any electron departs from the straight-line path (Fig. 109), as many of the electrons do after they leave the crossover point, their surrounding field is no longer perpendicular to the focusing field and the motion of these electrons is affected. These electrons now move in an arc and finally make a complete arc back to the center of the beam at the fluorescent screen. The spot at which all electrons converge is the focus point, and is made to be precisely at the fluorescent screen by proper setting of the current through the focus coil and proper positioning of the focus coil on the neck of the tube.

Were it not for the original velocity imparted to the electron, the divergent electron would make a complete circle (electron motion at right angles to its surrounding field and to the transverse focusing field) and return to the center. However, forward attraction by the second-anode potential causes it to follow a so-called "helical path," which is the result of a forward attraction and the circular motion. Actually, in the magnetically focused picture tube this complete helix is never made because the focusing field is confined to the relatively small width of the focusing coil, and its magnetic field only gives a twist to the divergent electrons as they pass through its magnetic field. Thus the focusing field just sets the divergent electrons on the proper path and they continue to move in this direction after they leave the field of influence of the focusing coil, finally meeting the center of the electron stream at the fluorescent screen.

ELECTRON BEAM SIZE

The required diameter of the scanning beam as it strikes the scanned surface is a function of the height of the scanning raster and the number of active lines. The quotient of raster height and active lines is the line width. The diameter of the scanning beam is made somewhat smaller than the line width to prevent overlapping of lines. In effect, therefore, the beam diameter must become small to preserve the resolution of the picture, and when the number of active lines is increased the beam diameter must be decreased. A large-diameter beam, however, is advantageous from the standpoint of picture brightness because the brightness varies as the density and size of the beam.

The factors which influence the beam size in the electron gun are:

Emitting area. A larger emitting surface produces a stronger beam but also a larger spot. In some types a curved cathode surface is used to obtain a small spot and still a relatively high beam current.

Electrode apertures. The various electrode apertures, because of their small diameter, limit beam size to the dense center portion of the electron stream. The control-grid aperture and first aperture of the first anode limit the crossover diameter, and the second and third aperture of the first anode prevent the more divergent electrons coming out of the crossover from entering the second anode. The final aperture of the first anode, and occasionally an aperture associated with the second anode, prevents random secondary electrons coming off the earlier apertures from passing on toward the screen.

Crossover. The purpose of the crossover in the electron gun is to take the electrons emitted from the cathode and group them into a fine, concentrated area of electrons.

Anode potential. Under most circumstances when high accelerating voltages are used it is to obtain additional brightness and a larger raster. Consequently, the beam current is higher and the scanning spot diameter larger. For a specific picture tube, however, a decrease in the accelerating voltage below the rated value causes a loss of resolution and an increase in beam diameter. Because the electrons are not drawn out of the crossover and off the cathode with peak velocity, space-charge effects increase the diameter of the beam.

IMPROVED ELECTRON GUNS

One disadvantage of the gun discussed in connection with Fig. 106 is that the first anode forms a part of both lenses. In conjunction with the cathode and the control grid it is a part of the first lens and in conjunction with the second anode it forms the second lens of the electron gun. As a result, changing the setting of the control-grid voltage or brightness of the scene on the fluorescent screen also affects the operation of the second lens. Actually, the changing of the control-grid bias moves the crossover point, and the first-anode voltage has to be readjusted to focus the newly positioned crossover point on the fluorescent screen. Likewise, changing the first-anode voltage affects the brightness of the scene because of its effect on the beam current. Actually, the accelerating action of the first anode on the beam is varied and the number of electrons pulled out of the crossover is affected.

To prevent this interaction between the two electronic lenses, another electrode was inserted into the electron gun (Fig. 111). This new element, called an *accelerating grid,* is at second anode potential, and is positioned up near the control grid. Thus the first lens consists of accelerating grid, control grid, and cathode; the second lens is formed by the first and second anode. This arrangement permits the crossover point to be formed with a higher voltage, producing a smaller scanning spot and making the crossover less affected by space charge effects when the control-grid bias is varied. Because the accel-

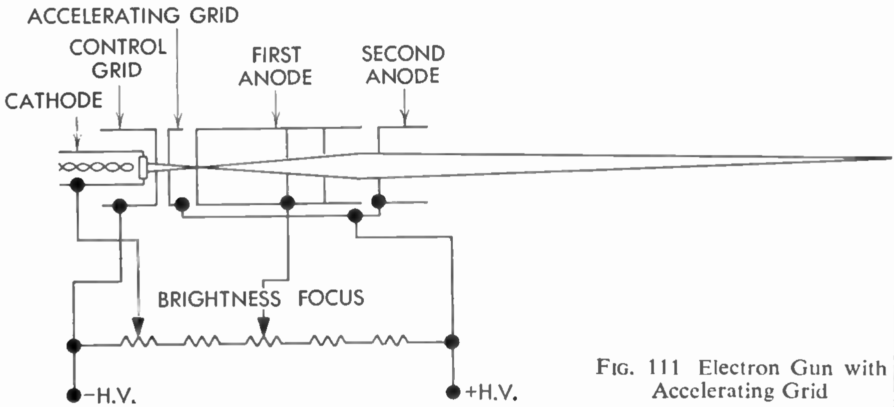


FIG. 111 Electron Gun with Accelerating Grid

erating grid is at second-anode voltage, its spacing to the control grid can be increased over that of the conventional control grid's first-anode spacing, with the result that a narrower beam is generated.

Another improvement in the electrostatically focused electron gun is the so-called "zero-first-anode-current gun" (Fig. 112). This new type of gun, employed in the newer picture tubes, has an accelerating electrode which has

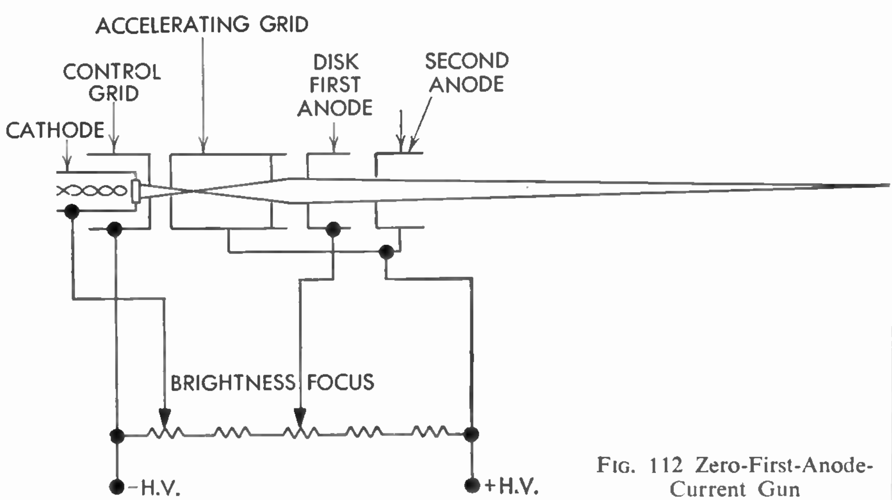


FIG. 112 Zero-First-Anode-Current Gun

been lengthened to carry apertures. The first anode has been shortened to a disc which is used only in connection with the second anode for focusing. The gun has better focusing characteristics, and has an added advantage in that no current is required at the focusing connection on the high-voltage power-supply bleeder. The actual concentration and masking of the beam is performed by the apertures of this accelerating electrode, which again is at the same potential as the second anode. The first anode is strictly a focusing anode and, inasmuch as it draws zero current, focusing is fixed and not affected by power-circuit regulation. Neither does its potential vary with beam intensity (brightness) which, in the other gun types, causes a variation in electron flow to the first anode and its associated bleeder resistor. The focus ratio of first-anode to second-anode voltages is determined, therefore, only by the bleeder tap instead of being altered by first-anode current which, heretofore, varied with beam current and, consequently, upset the voltage ratio. This is evident when we consider that with an elongated first anode having three apertures more electrons strike the first-anode apertures and wall when the beam current is high and there are more divergent electrons. With the first anode limited to disc size and the apertures transferred to the accelerating grid, there is no first-anode current.

74. *Fluorescent Screen*

The fluorescent screen of the picture tube glows in accordance with the variations of the electron beam which strikes it. The chemical composition of the fluorescent screen has the ability to absorb energy from the impinging electrons and cause subsequent re-emission of this energy as visible or near-visible radiation. This type of illumination is often called *cold light* because there is little heat lost in the energy transfer. Inasmuch as more energy is involved in the transfer when an element is under impact of higher velocity electrons or more electrons, the illumination is brighter. When this emission is terminated rapidly after the source of energy (electron beam) is removed, it is termed *fluorescent*. When emission continues after the passage of the beam (longer than 10^{-8} second), it is termed *phosphorescence*. Since light emission continues after the passage of the beam in the television picture tube, it is more truly a phosphorescent screen, and the chemical composition which causes this illumination under impact of the electrons is called a *phosphor*.

The ideal phosphor for the fluorescent screen would be one which had a uniform light emission over the entire visual spectrum, producing a true white light. White light and half-tone gradations of white light are acceptable because of our familiarity with motion pictures and black-and-white photography. In addition, the contrast range between white and black is greater than that for any other colors, and it is more acceptable to the human eye because it approaches the quality of diffused daylight. No practical phosphor has a uniform emission characteristic over the visual spectrum. A sensation of white may be obtained, however, by proper choice of the chemicals which make up the compound, most often consisting of a blue-emitting and yellow-emitting phosphor mixed in the proper proportion to produce an apparent white light.

A very important consideration in the choice of a phosphor for the television picture tube is its persistence. The persistence actually is the decay time of the phosphor, or how long the element will remain illuminated after the passage of the cathode ray beam. The persistence of the phosphor may neither be too long nor too short. If the element remains illuminated too long after the passage of the beam, although this does increase the average brightness of the illumination, rapid motion in the televised scene will cause the action to blur or have a series of long tails after it, because the elements have not had time to die out to a low value before the beam rescans. Consequently, for television application it is necessary that the phosphor decay to at least 1 per cent of its full brightness in a frame time, or in 1/30th second. If the screen persistence is too short, the light decays too rapidly and reduces the average brightness of the scene, causing flicker to be much more apparent. Actually, if the element illumination decays rapidly and the element is under excitation of the impacting electrons for only a short time during the total frame time, the element will be in darkness for a greater portion of the frame time and the apparent flicker will be much higher. Therefore, in television application an optimum choice of persistence must be used, which is termed a *medium persistence screen.* In practice, most of the phosphors used on the television picture tubes of today decay to 1 per cent of their original illumination in considerably less than 5,000 microseconds. The decay time is faster than we might expect from the standpoint of brightness and apparent flicker, but it is necessary to follow the rapid changes of scene and action now associated with television transmission.

In summary, the characteristics of the phosphor used on the fluorescent screen of television picture tubes are as follows:

1. *A chemically stable phosphor.* A phosphor is necessary which will not readily decompose and will have stability and long life under electron bombardment.

2. *A color pleasing and acceptable to the human eye.* A white-appearing phosphor or a white phosphor with a blue or yellow tint is employed.

3. *Ability to produce high, instantaneous brightness under impact of short duration excitation.* It is necessary that the brightness of an illuminated element rise quickly with the impact of the electron beam, which strikes the element only for a very short interval during each frame time. This requirement is dependent on the efficiency of the phosphor or the degree of illumination per beam excitation. Again this requirement is met with the use of a medium or fast persistence screen because this type of fast-acting phosphor has the highest efficiency.

4. *Correct persistence screen.* The persistence of the phosphor must be fast

enough to follow rapid motion correctly, avoiding loss of contrast and blurring due to a lengthy decay interval.

5. *Brightness.* Screen brightness must be adequate and flicker-free.

75. *Electrostatic Deflection*

The function of the deflection system of the television picture tube is to move the point of impact of the electron beam along the fluorescent screen in accordance with the standard interlaced scanning pattern. The deflection system is, therefore, located between the electron gun and the fluorescent screen and, as the beam passes through the deflection plates or deflection coils, it is deflected in accordance with the strength of the deflection field. In electrostatic deflection (Fig. 106) the two pairs of deflection plates, vertical and horizontal, are mounted within the glass envelope of the tube. The difference of potential between the two horizontal plates will make the beam strike the fluorescent screen to the right or left of center; the difference of potential between the two vertical plates will cause the beam to strike the screen above or beneath the center. Whether it is to the right or left or above or beneath the center depends on polarity and extent of the difference of potential between plates.

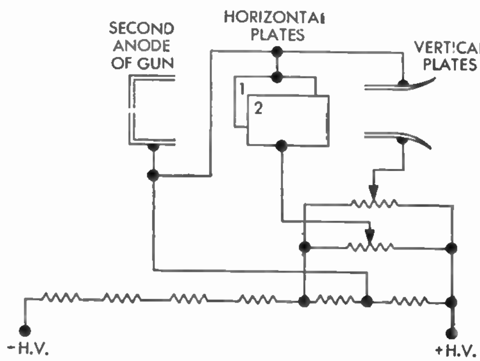


FIG. 113 Simple Deflection System as Used for Centering

If no difference of potential exists between the deflection plates of the properly centered gun, the electron beam will strike the exact center of the fluorescent screen. If a difference of potential exists between either pair of plates, the beam will be deflected as it passes through the plates in the direction of the most positive plate. The deflection system, as customarily used with a small test oscilloscope, is shown in Fig. 113; the deflection plates are always at or near second-anode potential to prevent deceleration of the beam as it passes between the deflection plates. When second-anode potential is applied to the four plates, the beam passes through their center and strikes the center of the fluorescent screen. When one of the plates is at a greater or lesser potential than the second anode, as can be adjusted with the so-called "centering potentiometer," the beam will deflect in the direction of the most positive plate. For example, when deflection plate *2* is made negative with respect to the second anode, the beam will deflect toward deflection plate *1*. To go a step further we can apply a signal to plate *2*, as shown in Fig. 114. Assuming this signal to be a linear sawtooth waveform, we find that the potential of plate *2* is at first negative with respect to the second-anode voltage, and then as the sawtooth rises it reaches the same potential as the second anode, and then goes further positive. Consequently, when the sawtooth first starts to rise, plate *2* is negative and the beam is deflected toward plate *1*, striking the fluorescent screen on the left looking toward the screen from the gun. As the sawtooth voltage increases in amplitude the difference of potential between plate *1* and plate *2* becomes progressively less and, therefore, the beam is not deflected to the left as far. In the mid-point of the linear rise of the sawtooth the two plates have the same potential and the beam is striking the center of the fluorescent screen. As the sawtooth voltage causes plate *2* to swing plus with respect to the second-anode voltage, the beam is now deflected toward plate *2* and, therefore, strikes the screen to the right of the center point. When the time of retrace of the sawtooth is reached, the much faster change in voltage snaps the beam back to the left side from the right side. Thus, you see by causing the potential of plate *2* to rise and fall about the second-anode potential, the beam has been moved from left to right across the screen. Likewise, if we were to apply a sawtooth signal to the vertical plates we could make the beam move up and down on the fluorescent screen.

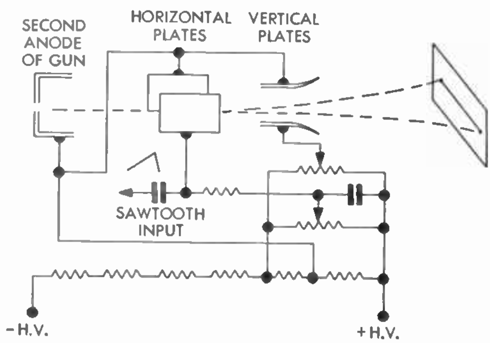


Fig. 114 Horizontal Deflection by Sawtooth Voltage

A more satisfactory method of deflecting the beam is by means of a balanced deflection system, as shown in Fig. 115, which eliminates a distortion

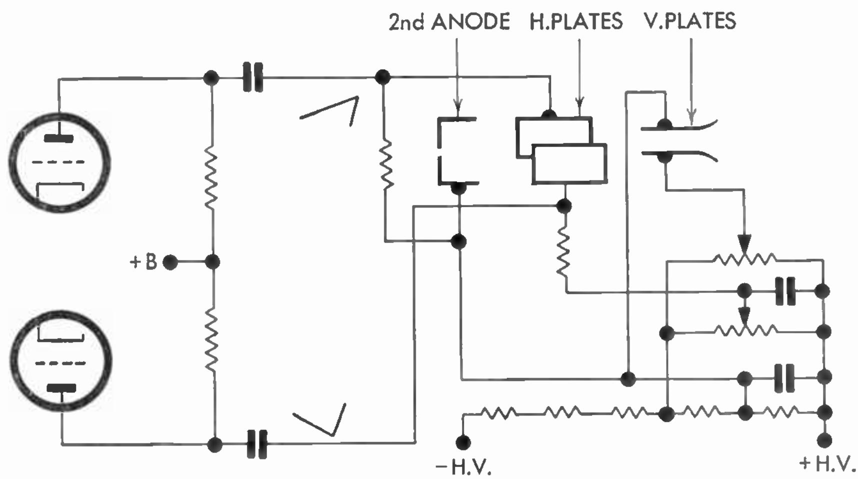


Fig. 115 Balanced Deflection Circuit

inherent in the unbalanced method of Fig. 114. The sawtooth voltage on one plate and a strictly d-c potential on the other does not cause an ideal field distribution between the two plates for all amplitude levels of the sawtooth voltage, particularly at the extremities of the sawtooth voltage when the beam is at the far left or far right. This change in the form of the equal potential distribution lines causes defocusing of the beam and nonlinearity, particularly in those regions in which the beam passes very close to the deflection plates (when the beam is deflected to the far left or right or to the very top or bottom of the scanning raster). To prevent this nonuniformity a balanced deflection circuit is used, as shown in Fig. 115. Here a sawtooth voltage is applied to one plate and an equal but opposite polarity sawtooth to the other to produce a more uniform deflection field for all amounts of deflection.

Thus, as one sawtooth voltage rises on the one plate it has more attraction for the electrons, while at the same time the sawtooth voltage is falling on the opposite plate, causing it to repel the beam electrons. The push-pull action produces a still greater deflection of the beam and a uniform deflection field.

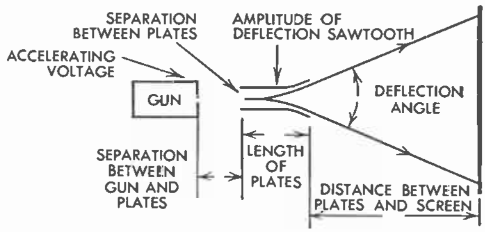


FIG. 116 Factors Which Influence Deflection Sensitivity and Angle

The factors which influence the deflection of the beam with an electrostatic deflection system (Fig. 116) are as follows:

Beam velocity. The higher the velocity with which the electrons are accelerated through the deflection system, the stronger the electrostatic deflection field required to produce a given amount of deflection. An increase in second-anode voltage, therefore, as required for a larger diameter picture tube, means the deflection field must be stronger for given amount of deflection because the second anode decides the final beam velocity (varies approximately as square of the second-anode voltage).

For a given electrostatically deflected picture tube an increase in anode voltage increases the beam current, decreases the spot size, because of less space charge effect, and improves the efficiency of the fluorescent screen. The increase in second-anode voltage, however, requires a higher voltage power supply and, for a given amount of deflection, a higher amplitude deflection voltage to obtain a stronger deflection field.

Length of deflection plates. To obtain a greater amount of deflection for a given deflection voltage, it is possible to lengthen the plates of the deflection system to keep the beam under the influence of the deflection field a longer period of time. Doing so, however, lengthens the tube physically and limits the angle to which the beam can be deflected, before the beam is deflected at an angle which causes its interception by the end of the deflection plates. Thus to obtain a large scanning raster the tube must necessarily be long.

Separation of plates. The deflection plates can also be moved nearer each other to obtain a stronger field and a greater deflection per given change in deflection voltage. However, again moving the deflection plates nearer to each other means that the angle of deflection is limited because again the beam can be deflected only through a small angle before it is intercepted by the end of the deflection plate. In most picture tubes, as shown in Fig. 106, the plates are moved relatively nearer to each other, and the ends of the deflection plates flared to prevent them from intercepting the beam electrons which are deflected through a wide angle. This construction means that the beam deflection can be improved by having the plates nearer to each other, and at the same time a reasonably wider angle of deflection can be obtained.

Distance from deflection system to fluorescent screen. Again, the greater the separation between the electron-gun deflection system and the fluorescent screen the smaller the angle through which the beam has to be bent to produce a scanning raster of given dimensions. However, too great a separation between the deflection system and the fluorescent screen increases the physical length of the tube and reduces the beam illumination, because of the added mutual repulsion and defocusing of the beam electrons in traveling over a longer distance. A smaller spot and higher illumination can be obtained by decreasing the distance between the screen and deflection system. To deflect the beam over the entire area of scanning raster now requires a stronger deflection field and a wider angle of deflection.

The strength of the deflection field necessary for a given deflection area or size of scanning raster is dependent on a number of factors. The strength of the deflection field and, therefore, output of the deflection amplifiers, is directly related to the size of scanning raster. The deflection field must be made stronger if the beam velocity is increased, the separation of the deflection plates increased, or the angle of deflection made wider. A weaker field is necessary for a given amount of deflection when the plate size is greater or the separation between the deflection system and fluorescent screen is lengthened.

Defocusing of the beam. In the deflection system there is always some defocusing of the electron beam present because of the close proximity of the deflection plates, particularly when the beam is being deflected through a wide angle and the beam electrons pass close to the deflection plate. The resultant defocusing of the beam and consequent loss of resolution is caused by two defects. First, inasmuch as the beam diameter is a finite value, those electrons nearest to the positive deflection plate are accelerated more than those which are on the other side of the beam and, consequently, the spot, instead of being small and circular, is pulled out or elongated, particularly when the beam is passing near the deflection plates. This first defect causes the beam to be defocused at the outer extremity of the scanning raster. The second defect is caused by the average increase of acceleration of the beam (two right-angle forces) when the beam is deflected near the deflection plates.

This means the focal length of the second lens has been changed to a certain degree producing the defocusing action.

Another potential defect is the unequal spacing between the gun and the outer edges of the scanning raster, and between the gun and the center of the scanning raster. This unequal separation causes the outer edges and the center to have different focal points. This defect can be overcome to a certain extent by curving the face of the tube and the fluorescent screen. Curving the surface of the fluorescent screen the proper amount will cause the outer edges of the screen to be the same distance from the gun as the center. While this expedient may cut the electronic distortion an optical distortion is now introduced because the televiewer is not looking at a flat screen. Consequently, with electrostatic deflection it is impossible to have the center of the picture properly focused at the same time the outer edges are precisely focused, resulting in a loss of resolution at one point or the other.

DEFLECTION SENSITIVITY

Deflection sensitivity is a measure of how far along the scanning raster a beam is deflected for a given change in deflection voltage. The deflection sensitivity is measured vertically and horizontally per given change in vertical and horizontal deflection voltage.

Deflection sensitivity is generally measured in terms of volts per inch deflection for a given second-anode voltage. For example, the deflection sensitivity of a specific pair of deflection plates for 3,000 volts second-anode voltage is 100 volts per inch. If it were necessary to deflect this beam over 4 inches it would require a peak-to-peak voltage change of 400 volts. Inasmuch as balanced deflection amplifiers are used in television, it would require a sawtooth voltage of 200 volts peak amplitude from each section of the deflection amplifier stage, or a total voltage change of 400 volts peak, 200 volts positive on one plate and 200 volts negative on the opposite deflection plate. In most television picture-tube circuits the pair of deflection plates nearer the electron gun is generally used for horizontal deflection, and, because their separation from the fluorescent screen is greater than that of the vertical deflection plates and the fluorescent screen, the sensitivity is better. Actually, if the 100 volts per inch previously mentioned was the deflection sensitivity of the vertical deflection plates, the sensitivity of the horizontal deflection plates might be 80 volts per inch.

The deflection sensitivity also varies with the second-anode voltage. Naturally, the deflection sensitivity becomes poorer as the second-anode voltage is increased. For example, for a given tube, if 1,000 volts is the second-anode voltage the deflection sensitivity might be 30 volts per inch, but if second-anode voltage is increased to 3,000 volts the deflection sensitivity will now be less and will require 90 volts per inch deflection. Thus, in tube-rating lists you will often find deflection sensitivity given in terms of volts per inch per thousand volts of second-anode voltage.

BEAM DEFLECTION ANGLE

The deflection angle is the limit of deflection of the electron beam. For example, if the deflection angle is listed at 50 degrees it means the maximum number of degrees the beam is deflected from the center is 25 degrees on each side of the center line, or a total angle of 50 degrees. The higher the deflection angle can be made the shorter the tube can be made to give a scanning raster of a definite size. If the deflection angle is small, as it is in electrostatically deflected tubes, the tube must be lengthened in order to increase the width and height of the scanning raster. When magnetic deflection is used, it is possible to deflect the beam at sharper angles. Therefore, the deflection angle itself is greater, and for a given-size scanning raster the distance between deflection system and fluorescent screen can be made shorter and, consequently, the physical length of the tube shorter.

The following factors again influence wide-angle deflection. Each factor is, however, again limited by need of some other feature or desirable characteristic of the tube.

1. *Beam velocity.* The lower the beam velocity and second-anode voltage, the easier it is to deflect the beam through a wide angle. The limitation again is loss of brightness, poor efficiency, and increased spot size.

2. *Size of deflection plate.* If the size of the deflection plate is increased and the strength of the deflecting electrostatic field increased, the beam can be made to deflect over a wider angle without striking the end of the deflection plate. The disadvantages of decreasing the length of the plate below an optimum value are: the considerable increase in deflection voltage (poor deflection sensitivity), and the more serious loss of resolution at the outer portion of the scanning raster because of beam defocusing when the beam angle is wide.

3. *Separation of plates.* The beam can be made to deflect through a wider angle without intercepting the edges of the deflected plate when the plates are well separated. Doing so again increases the strength of the deflection field required and causes beam defocusing at the outer portions of the scanning raster.

4. *Distance from deflection system to fluorescent screen.* The position of the fluorescent screen has no effect on the absolute value of the beam deflection angle. However, for a given beam angle the size of the scanning raster increases as the fluorescent screen is moved away from the deflection system. It is well to point out again that this is of no particular advantage because as the screen is separated from the deflecting system, the brightness decreases and the spot size increases. The ideal picture tube, therefore, would be one with a wide deflection angle and close spacing between fluorescent screen and deflection system to give us high brightness, small spot size, and a high-efficiency fluorescent screen. Magnetic deflection of the beam is more in

accordance with these requirements and is, therefore, most widely used for television picture tubes.

76. *Magnetic Deflection*

Magnetic deflection is the most common type of deflection used for picture tubes. Only a few picture tubes (7 inch) use electrostatic deflection. Magnetic deflection is adaptable to television picture tubes because for a given screen area the tube can be made shorter and size of spot smaller in diameter with consequent improvement in resolution.

A necessary step in understanding magnetic deflection is to know exactly what occurs when it is stated that the electron is deflected at right angles to both the magnetic field and the forward motion of the electron. If an electron is moving in the direction indicated by arrow *1* (Fig. 117) and a magnetic force is exerted with its magnetic lines indicated by arrow *2*, a force will be exerted on the electron *A* which will attempt to move the electron in the direction of arrow *3*. The three dimensional perspective afforded by the cube shows us that this motion is at right angles to both electron motion and the magnetic lines of force. Actually, the motion of the electron is not so abrupt—it does not stop dead and make a sharp left turn as soon as it comes under the influence of the vertical magnetic field. Rather, the electron does not lose its forward motion (acceleration given to it by the second anode, in case of a television picture tube) and is deflected only in proportion to the strength of the magnetic field. Consequently, the motion of the electron is circular as long as it is under the influence of the magnetic field. The motion can be depicted by the series of blocks (drawing B), which shows the influence of the magnetic field in causing the electrons to bear left. Once the electrons have left the magnetic field they continue along their forward straight path and, in case of a television tube, strike the screen left of center.

The influence of the magnetic field on the electron beam of a picture tube is shown in drawing C. Here we have an upright magnetic field set up by the two horizontally mounted deflection coils. Between *A* and *B* the beam is moving in a straight line toward the center of the fluorescent screen. However, at point *B* it comes under influence of the magnetic field and the beam is pulled to the left between *B* and *C*—the extent of the deflection, however, is dependent on the strength of the field. Between *C* and *D* the beam again moves in a straight line, striking the screen to the left of center. If the lines of magnetic force are reversed (easily done by changing the direction of current flow through the coils) the beam is deflected to the right of center. Likewise, if another pair of deflection coils are mounted vertically, their magnetic field can be used to deflect the beam vertically.

The magnetic rule which applies particularly to magnetic deflection says that when a conductor carrying current is in a magnetic field there is a force exerted on the conductor attempting to move it at right angles to both the direction of the current flow and the lines of force of the magnetic field. This rule also applies to an electron stream, the actual direction of the stream shifting. It is the magnetic field surrounding the beam which is the contributing factor in the deflection of the beam. To prove this, observe the deflection system of drawing B, Fig. 117, looking from the fluorescent screen as shown in drawing A, Fig. 118. In this drawing, the electron stream is coming off the paper toward you, and thus the magnetic field surrounding the deflection

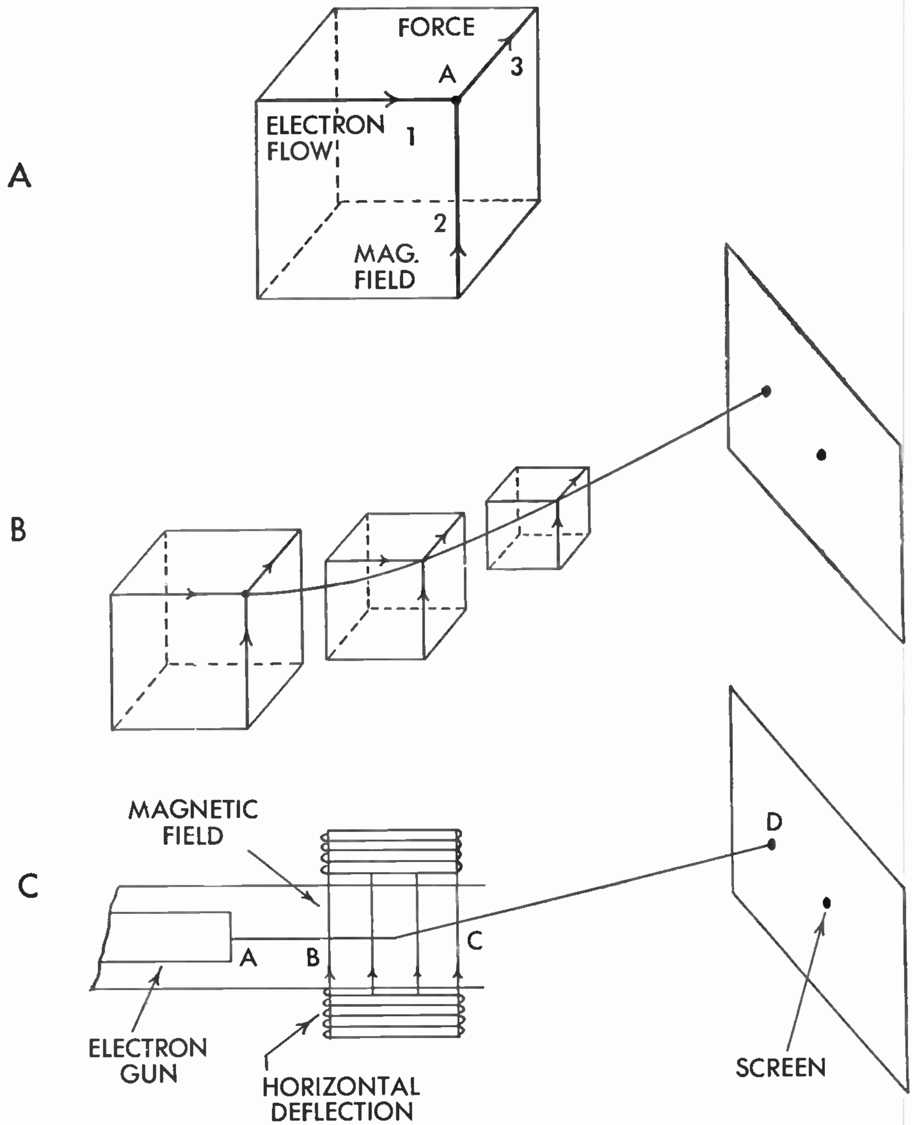


FIG. 117 Principles of Magnetic Deflection

coils and those surrounding the electron stream oppose each other on the right side, resulting in some cancellation. On the left side the two fields are additive. Thus, there is a greater pressure on the left and the beam is forced to the right (beam is forced to the left so far as the perspective of drawing C, Fig. 117, is concerned, for there we were observing from the electron gun end of the picture tube). The magnetic deflection field has been reversed in the next drawing, causing cancellation on the left side and a conse-

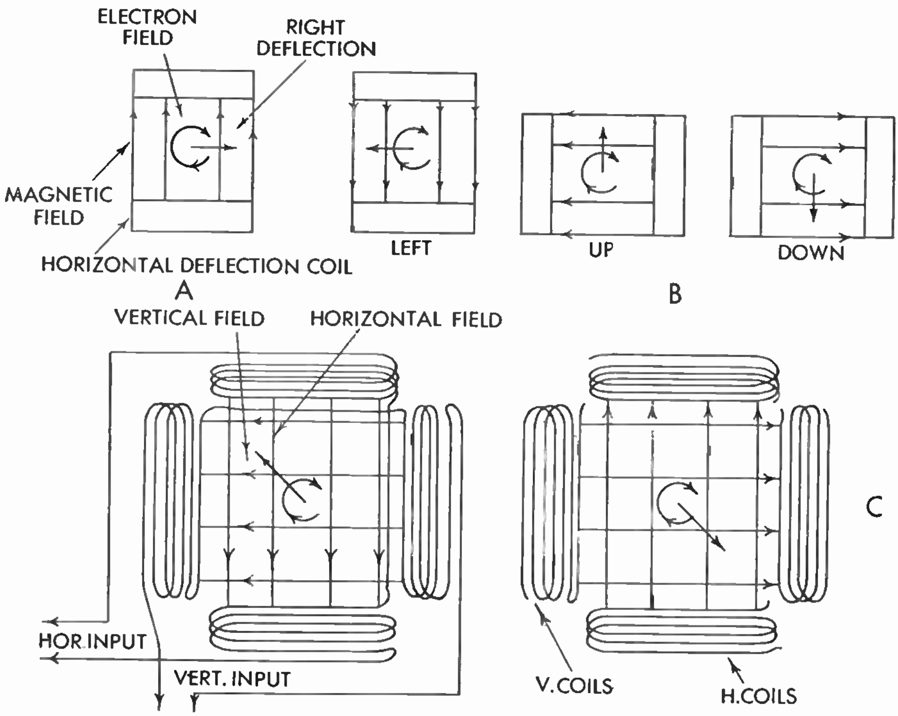


FIG. 118 Direction of Deflection

quent left deflection. In the remaining two drawings deflection of the beam by vertical deflection coils is indicated. In the actual picture tube there are two pairs of deflection coils, horizontal and vertical, and, as shown in drawing C by changing the strength and the polarity of the magnetic fields relative to each other, it is possible to move the beam to any position between the deflection coils. Consequently, with the proper application of sawtooth currents to the deflection coils, it is possible to form a scanning raster just as in the case of electrostatic deflection.

In actual practice, the coils are not mounted in a square rectangular manner as indicated, but instead are curved to fit the neck of the tube. In fact, horizontal and vertical coils almost entirely overlap each other to produce uniform magnetic fields. Inasmuch as the windings and resultant fields are perpendicular, there is no appreciable interaction and cross-coupling between the windings. Generally it is the horizontal deflection coils which are mounted nearest to the glass envelope and, therefore, nearest to the electron beam, because it is more difficult and requires more power to generate the much faster changing horizontal deflection fields for television. There are factors similar to those associated with electrostatic-deflection sensitivity which influence magnetic-deflection sensitivity. These factors are:

Beam velocity. The higher the second-anode voltage and beam velocity the stronger the deflection field required for a given amount of deflection. With magnetic deflection, deflection is inversely proportional to the square root of the velocity. Therefore, increasing the second-anode voltage and beam velocity does not decrease the deflection sensitivity at as fast a rate as it does for electrostatic deflection, for which the deflection sensitivity is inversely proportional to the second-anode voltage. Consequently, magnetic deflection is adaptable to picture tubes which require a high second-anode voltage needed for large-screen picture tubes.

Size of deflection area. Again, as in the case of electrostatic deflection, the amount of deflection is dependent on the length of time the electron stream is under the influence of the deflection field.

Separation between deflection coils. Deflection coils for magnetic deflection are mounted externally and, therefore, the beam which is internal to the tube can be deflected over a relatively wide angle before it would meet any obstruction. Consequently, with magnetic deflection the beam can be deflected over a relatively wider angle than with electrostatic deflection. This means that the actual scanning raster can be moved closer to the deflection system for a given raster size than it can be for electrostatic deflection.

Distance from scanning raster to deflection coils. The greater the separation between scanning raster and deflection system the wider is the area over which the beam can be deflected. However, one advantage of magnetic deflection is that the beam itself can be deflected over a wider deflection angle. Consequently, for a given scanning area the scanning raster can be positioned nearer to the deflection system. In addition, with magnetic deflection, the actual strength of the deflection field must not be increased at as rapid a rate with second-anode voltage; therefore, the strength of the deflection field required for wide-angle deflection of a high-velocity beam is not out of proportion.

Defocusing of beam. Again, as with electrostatic deflection when the beam is deflected at its most divergent angle from the center, the spot tends to spread because of the different velocities of the electron which make up the beam. This defocusing at the extremities, however, is not nearly so severe with magnetic deflection. This is understandable again when we consider that the deflection is not nearly so dependent on electron velocity as it is with electrostatic deflection, because magnetic-deflection sensitivity varies inversely with the square root of the velocity; with electrostatic deflection, deflection sensitivity varies inversely as the velocity.

In summation, magnetic deflection is preferred because of the shorter length of the tube for a given scanning area and second-anode voltage. We realize, of course, that with the higher second-anode voltage and the nearness of the scanning raster to the electron gun, the smaller the spot size and the better the resolution of the picture tube. Likewise, the unequal deflection of the various electrons which make up a cross section of the beam and have differing velocities is not as pronounced with magnetic deflection, because the deflection is not as readily affected by differing velocities of the electrons which make up the beam.

77. *Ion Trap Circuits*

Ion spot defect in the television picture tube produces a dark spot at the center of the screen. This dark spot or dark stain, which often appears at the center of the picture tube, is due to the decomposition of the chemical compound of the screen under impact of ions. The free release of electrons from the cathode also permits the release of ions from the cathode. These ions have the same charge as the electrons, but have a much greater mass. Consequently, they strike the fluorescent screen with a much greater force than the electrons, and although they are arriving at the same velocity their added weight is instrumental in causing decomposition of the chemical compound of the screen.

When electrostatic deflection is used these ions are dispersed over the entire surface of the screen in accordance with the deflection of the electrons and, therefore, do not cause severe damage to the screen. However, with magnetic deflection, the ions are deflected only at a small angle by the magnetic deflection fields. They continue to strike the screen at one spot in the center area and eventually break that area down. This is understandable when we consider that with electrostatic deflection, deflection is caused by an attraction for the charge of the electron. Inasmuch as the ion charge is the same as the electron charge it has the same attraction to the deflection field. With magnetic deflection, however, there is a magnetic force which shifts the position of the entire electron stream and, of course, the amount of shift is dependent on the mass of the objects to be displaced. The mass of the ion is much greater than that of the electron and the electron is displaced through a much greater angle than the ion.

A number of innovations are used in picture tubes to attract the ions and prevent them from striking the fluorescent screen. The three most common ion trap or bender circuits used in magnetic deflection tubes are shown in Fig. 119. In the system shown in the first drawing an asymmetrical electron gun is used in conjunction with an external ion trap or bending coil. As the electrons and ions of the beam pass to the accelerating anode, they are attracted by the lower extension of this anode; therefore, ions and electrons are both deflected toward this portion of the accelerating anode. As the electrons and ions move toward the wall of this accelerating anode, they come under the influence of a magnetic field generated by the external coils. The magnetic field is of the proper direction and intensity to return the electrons to the center of the gun structure; however, the magnetic field does not appreciably influence the ions of the beam, and they strike the accelerating grid and are dissipated. Thus the ions and electrons are both made to diverge from the center of the gun structure and only the electrons that have a much lower mass are readily deflected by a magnetic field and returned to the center line. The ions again, because of their greater mass, cannot be deflected by the magnetic field, and they strike the walls of the attracting accelerating grid.

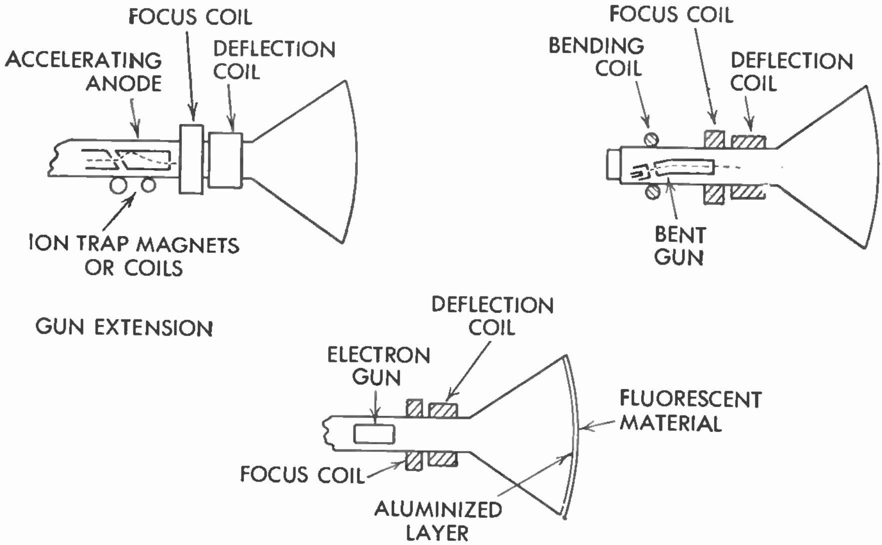


FIG. 119 Ion Spot Removal

The electrons must be deflected at two points, once just before the electrons reach accelerating anode (point *1*) and a second time to pull the electrons back in the center (point *2*) and moving in a straight path toward screen. The deflection angle at point *1* is greater and, consequently, magnetic field must be the stronger. Thus the larger coil or stronger magnet, if fixed magnets are used, is nearer the picture-tube base.

Another means of removing ions from the beam is termed the *bent-gun method*. In this system the actual cylindrical elements of the gun are mounted at an angle with respect to the center line of the tube. An external bending coil or magnet must also be used with this method. In this system, when the electrons and the ions emerge from the bent gun they move toward a collector wall of the next cylindrical element. As the electrons and ions leave the gun they come under the influence of a magnetic field which is of the proper strength and polarity to return the electrons to the center line of the picture-tube structure. However, the ions again are not influenced by this magnetic field, and they strike the collecting wall to be dissipated. Actually, the same basic theory is applied in both of these types of ion traps. The difference is that in the first type an asymmetrical gun was used, while in the second type an angled gun was used.

In the third type an aluminized screen is used which removes the ions from the beam. The actual aluminized screen is extremely thin and will readily transfer electrons, causing them to strike the fluorescent material. However, the ions, which are of greater mass and dimension, will not pene-

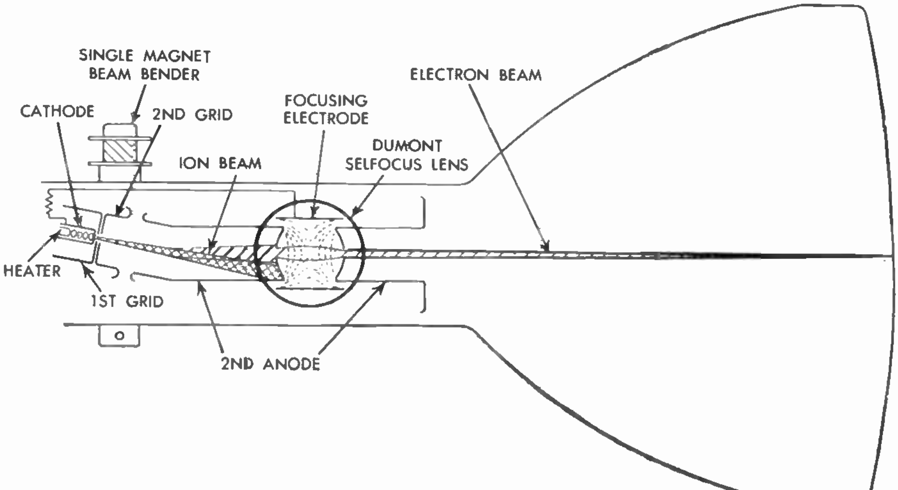


FIG. 119a DuMont Self-Focus Gun

trate the aluminized layer and, therefore, they are dissipated by it and do not reach the fluorescent screen of the picture tube. This is by far the simplest method of removing the ion because it does not require the construction of an elaborate electron gun, nor does it require an external magnetic field to remove the ions.

DuMont has developed a self-focus picture tube that uses electrostatic focus with no external focusing circuit. The base plan of this tube is the same as the standard base for magnetically focused types. No focus coil is required while a standard-deflection yoke and a single-magnet beam-bender are employed. The picture is in focus when the ion trap is set at maximum brightness.

An important advantage of the self-focus tube is that it remains in focus for wide variations in its electrode voltages and for appreciable shifts in the a-c line voltage. In fact, self-focus properties mean that no external focusing control is necessary when this type of tube is used.

In a conventional electrostatic gun the focusing is a function of the voltage ratio between the first and second anodes. It is also influenced by the position of the electron crossover, which is a function of cathode, grid, and first anode voltages. Consequently, any change in the absolute electrode voltages or any shift in the second-to-first anode ratio upsets the true focus. In the self-focus gun, Fig. 119a, the second anode is split, and the focusing electrode is mounted around the split. The focusing electrode is tied back to the cathode through an internal resistor with no external focusing connection required. An external connection and potential are needed for some newer types. The position of the focusing electrode and the width of the second anode-split bring electrons to focus at the fluorescent screen. It is important to realize that the voltage ratio between the two sections of the second anode is always constant and unity, and likewise, the focusing-electrode potential is set at the cathode and is also a constant. Therefore, because variations in high voltage and potential at other electrodes have a lesser influence on proper focus, better focus-regulation exists.

ALUMINIZED FLUORESCENT SCREEN

The aluminized or metal-backed fluorescent screen has a number of advantages, the most important being increase in brightness and improvement in apparent contrast. The aluminized screen consists of a thin metallic aluminum layer deposited on the side of the fluorescent screen nearest the electron gun. This screen reflects the light normally reflected back off the fluorescent screen to the glass walls of the tube and causes it to be concentrated forward, increasing the screen brightness.

The three factors which most influence the brightness of the fluorescent screen, particularly in the case of high illumination projection tubes, are:

1. An increase of second-anode voltage, thereby increasing the current of the beam without sacrificing spot size.
2. Improvement of the luminous material and the attainment of a better conversion efficiency from beam impact to light.
3. Sufficient utilization of all light emitted from the fluorescent material.

The latter requirement is met by the metal-backed screen because it concentrates forward all light normally lost through back radiation and diffuse dispersal from the glass walls of the tubes.

The particular advantages of the metal-backed screen, therefore, are:

Increased brightness and improved contrast. In the older tubes, approximately 50 per cent of the light generated was emitted toward the electron gun and another 15 to 25 per cent was lost by internal reflection on the glass walls of the tube. With the metal-backed screen most of this previously lost light is concentrated forward where it will serve as useful increase in average brightness. There is also an increase in apparent contrast, because much of the reflected light previously diffused itself indiscriminately over the fluorescent screen and caused halation and loss of contrast. The removal of this dispersed light permits the televiewer to observe the true contrast of the picture on the screen.

Elimination of the ion spot. The aluminized layer in back of the screen also serves as an effective trap for the ions emitted from the cathode because, having a greater mass than the electrons, they cannot penetrate the layer. Consequently, only the electrons can pass through the layer onto the fluorescent screen. The thinness of the layer permits the transfer of electrons without loss.

Elimination of secondary-emission restriction. To produce the high illumination necessary for large television tubes, it is necessary to increase the second-anode voltage and, therefore, the beam current. This means the electrons are arriving at the fluorescent screen in quantities. Heretofore it was a characteristic of the fluorescent screen that the secondary-emission characteristics of the screen were not ample enough to drain the electrons off at a fast enough rate. Consequently, the elements of the fluorescent screen assumed an equilibrium negative potential value with respect to second anode which retarded the arrival of further electrons and slowed the beam velocity. In effect, this means that the second-anode attraction was not as great as the second-anode voltage implied, and the velocity of the electrons as they arrived at the screen was not sufficiently high to produce the added illumination required. This so-called "sticking" of the fluorescent screen caused the screen to reach a certain brightness, which could not be increased with any further rise in second-anode potential. The presence of the metal-backed screen, which quickly drains off these secondary electrons, eliminates this restriction. Therefore, with a metal-backed screen it is possible to increase the second-anode voltage substantially and with this increase also obtain a substantial increase in illumination.

78. *Commercial Picture Tubes*

A listing of the most common picture tube types and their characteristics is given in Fig. 120. Some of the tubes have an external conducting coating in addition to the internal second-anode coating. A high-voltage filter capacitor of approximately 500 micromicrofarads is formed by the two coatings (external coating is grounded) and the tube's glass wall which serves as the dielectric.

The three most common base connection types are also shown in the chart, Fig. 120. The first base-type is generally used for picture tubes that use electrostatic deflection and focus. The second base-connection is generally employed by tubes using electrostatic focus and magnetic deflection or the DuMont self-focus picture tube. Magnetic deflection and magnetic-focus picture tubes, as well as the older self-focus picture tube, generally use the third base-diagram. Either single or dual magnets are used for the ion trap as a function of the electron-gun design. In general, those tubes which have an asymmetrical electron gun use the double ion trap, while with a bent or tilted gun-construction, a single magnet is used.

Types of Picture Tubes	*Focus*	*Deflection*	*Deflection Angle*
7″ Round Glass	*E*	*E* or *M*	50°
10″ Round Glass	*E* or *M*	*M*	50°
12½″ Round Glass	*E* or *M*	*M*	40°–50°
12½″ Round Metal	*E* or *M*	*M*	54°
14″ Rectangular Glass	*E* or *M*	*M*	70°
15″ Round Glass	*M*	*M*	50°–57°
16″ Round Glass	*E* or *M*	*M*	50°–70°
16″ Rectangular Glass	*M*	*M*	70°
16″ Round Metal	*M*	*M*	70°
17″ Rectangular Glass	*E* or *M*	*M*	70°
17″ Rectangular Metal	*E* or *M*	*M*	70°
19″ Round Glass	*M*	*M*	66°
19″ Rectangular Glass	*M*	*M*	70°
19″ Round Metal	*M*	*M*	66°
20″ Round Glass	*M*	*M*	54°
20″ Rectangular Glass	*E* or *M*	*M*	70°
21″ Rectangular Glass	*E* or *M*	*M*	70°
21″ Rectangular Metal	*E* or *M*	*M*	70°
24″ Round Metal	*E* or *M*	*M*	70°
27″ Rectangular Glass	*M*	*M*	90°
27″ Rectangular Glass	*M*	*M*	90°
30″ Round Metal	*M*	*M*	90°

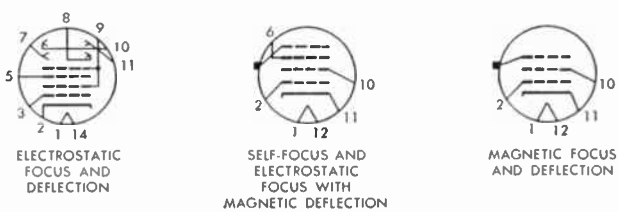


FIG. 120 Picture-Tube Chart

It is important to realize that (for a picture tube of a given size) the wider the deflection angle, the shorter the over-all length of the tube can be for a given raster size. Accordingly the wider angle picture tubes are shorter in physical length. Some picture tubes employ gray filter-glass faceplates to reduce room reflections and glare, while still others use a frosted faceplate for increased brightness, apparent contrast range, and minimum reflection.

79. *Picture-Tube Signal and Voltage Circuits*

A number of a-c signals and direct voltages must be applied to the electrodes of the picture tube. The three a-c signals are picture and blanking to the control grid, and vertical and horizontal sawtooth waves to the respective deflection systems. A variable d-c bias voltage must also be applied to the control grid, and the proper potentials must be applied to the elements of the electron gun. In addition, it is necessary to have a d-c component of deflection voltage or current to center the entire picture on the scanning raster properly.

The various signals, voltages, and currents necessary to operate a magnetically deflected picture tube properly are indicated in Fig. 121. Signal voltage is, of course, applied to the control grid of the picture tube to modulate the electron stream in accordance with the light variations and to shut off this electron beam during the retrace intervals of the scanning process. Sawtooth signal current in the horizontal and vertical deflection coils produces the changing magnetic field, which causes the beam to scan across and down the screen. The electrode voltages for the cylindrical elements of the electron

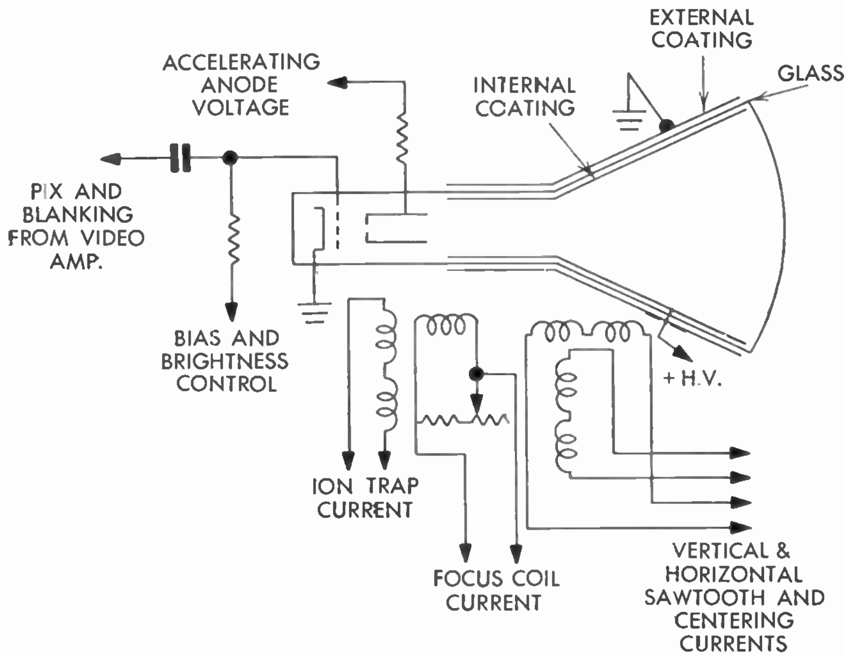


FIG. 121 Picture-Tube Circuit, Magnetic Deflection and Focusing

gun are generally taken off a bleeder divider across a high-voltage power supply. The grid is the most negative electrode of the gun. It is either at the low-potential end of the bleeder system or a separate low-voltage power-supply source is used to set the grid negative with respect to the cathode. A d-c component of current also passes through the deflection coils to center the electron stream properly. In addition the magnetic field generated by the focusing coil and by the ion trap coil must be formed with a component of d-c current generally taken off the low-voltage power supply. As can be seen, proper operation of the picture-tube circuit is dependent on a number of signals and voltages which must be properly set in level and amplitude to obtain a satisfactory picture.

When electrostatic deflection is used (Fig. 122), a similar picture and blanking voltage must be applied to the control grid, and sawtooth voltages of the proper horizontal and vertical repetition rate must be applied to the deflection plates. The electrode potentials for the electron gun and the deflection plates are generally taken off a bleeder across a high-voltage power supply. One deflection plate is always at second-anode potential, and its companion plate can be raised above or below the second-anode voltage by means of the centering control; consequently, the picture as a complete unit can be moved right or left or up and down to properly center the picture on the face of the tube. Generally, the d-c component of grid voltage, which sets

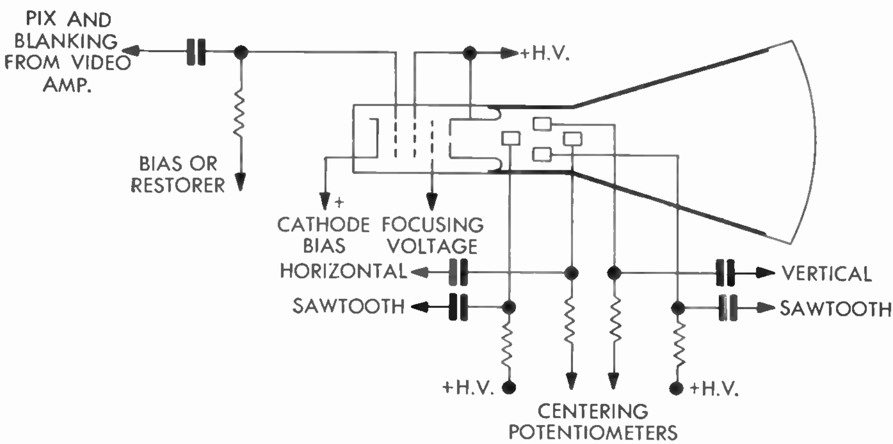


FIG. 122 Picture-Tube Circuit, Electric Deflection and Focusing

the average brightness, is obtained from a divider system across the low-voltage power supply. The divider system applies positive voltage to the cathode of the picture tube. So far as grid signal is concerned, it is the picture-signal voltage variation. Inasmuch as little power is required in picture-tube circuits, component parts can be made at a reasonable cost and without excess weight. To generate a magnetic field of sufficient strength to deflect the beam its full sweep horizontally requires some power. The horizontal current required sometimes approaches ½ ampere.

80. *High-Voltage Systems*

A high accelerating potential is necessary to generate the high-velocity scanning beam of the cathode-ray picture tube. Although the potential necessary is in the thousands of volts, the actual current drawn at this high voltage is relatively small. Thus, it is possible to construct a satisfactory high-voltage supply for picture-tube operation with relatively small component parts which do not add excessively to the weight and area of the receiver chassis. A number of systems are being used presently in the modern television receiver. There are only a few television receivers today that use a high-voltage transformer excited by the 110-volt primary-line voltage. Most of the receivers rectify the peak amplitudes of some oscillating or transient voltage as their source of high-voltage d-c potential. This system has been found very practical because the very minute current drawn does not appreciably load the transient voltage source, or in any way adversely affect the operation of the circuit to which the transient is associated.

The current requirement of the average picture tube is extremely low, and the very peak instantaneous current is less than 1 milliampere. The low current requirement means that the size of the component parts can be kept small, and a simple filter system, consisting generally of only a resistor and a pair of capacitors, very effectively filters the rectified voltage. If the time constant of the filter capacitors and resistor is kept sufficiently long to prevent any appreciable discharge of the capacitor between conducting periods of the high-voltage rectifier, a well-filtered high-voltage source is obtained.

When electrostatic-deflection picture tubes are used, a series of resistors is connected across the high-voltage source and the various electrodes of the picture tube are tapped off at appropriate points to obtain their correct potentials. The current drawn by the picture tube and the current passing through the bleeder resistor represent the load placed on the high-voltage supply. In practice, the values of the resistors in the bleeder network are relatively high, and bleeder current as a result is always less than 1 milliampere. In the receivers in which a picture tube is used that is focused and deflected magnetically, there is often no bleeder system across the high-voltage supply, high-voltage supply furnishing only the second-anode potential. Proper potentials for the accelerating grid and other low-voltage electrodes of the picture tube are obtained from the low-voltage supply. As a result there is an extremely light load on the high-voltage supply system because of the absence of a bleeder network, plus the fact that the only potential required is that which is applied to the second anode of the picture tube.

To prevent focusing instability and variation of correct focusing point with changes in brightness setting, it is necessary that the high-voltage supply system have good regulation. Good regulation, of course, can be obtained by keeping a constant load on the high-voltage supply which does not vary appreciably with the changes in the picture-tube beam current. This can be accomplished by using a bleeder network which draws appreciable current from the high-voltage system. Of course, the greater the current drawn the higher the ripple voltage becomes and stricter filter requirements are necessary. If magnetic deflection and focusing are used there is little interaction between focusing and high-voltage systems. High-voltage regulation requirements are not as strict. In the transient and oscillating voltage system the base ripple frequency is much higher than when a 60-cycle high-voltage transformer is used and, consequently, it is filtered with greater ease by relatively small value capacitors. This expedient also makes the high-voltage supply less hazardous because the small capacitors cannot accumulate as high a charge as the larger capacitors needed to filter the low-frequency, 60- or 120-cycle ripple.

The transient voltage supply is an economical and effective means of obtaining the 8,000 to 10,000 volts required for optimum operation of the television picture tube. Decidedly fewer massive component parts are required. The filtering requirements are not as strict because of the higher ripple frequency. This means of obtaining the high voltage also presents a safety feature as far as the operation of the picture tube is concerned, because if there is a failure of horizontal sweep voltages the high-voltage second anode also fails at the same time and an extremely bright spot does not appear on the fluorescent screen.

If the high-voltage system is to be operated at peak efficiency it is necessary that the load on the system be kept extremely light and, consequently, dielectric losses and arcing losses must be kept at a minimum. Thus, high-tension cables are used to convey the high-voltage signals to the picture tube, and proper spacing must be allowed between the high-voltage points and ground as well as other surrounding objects. Tube sockets and other mountings also must be made of good dielectric material to prevent stray losses. To prevent the occurrence of corona and sparking, it is necessary that the conducting wires as well as the connection points contain no sharp point or surface. Rounded corona shields are used at many connection points in the high-voltage system to distribute the high potential over a large surface and prevent arc-over from any one concentrated point. As the potential requirements become increasingly higher the corona and spark-discharge problems become more acute. This applies particularly to projection television receivers which use voltages as high as 15,000 to 90,000 volts on the picture tube.

FLYBACK HIGH-VOLTAGE SYSTEM

A transient or flyback voltage supply is a part of the horizontal-deflection output circuit and utilizes a transient voltage generated in the horizontal-deflection coils to supply the high-voltage potential for the picture tube. A typical transient voltage supply is shown in Fig. 123. During the horizontal trace period a sawtooth of current is building up in the deflection coil and, therefore, a field is built up around the deflection coil. During the retrace period the deflection output tube is generally cut off, and the abrupt removal of plate current causes the magnetic field to collapse and a high-amplitude single-alternation transient voltage to appear across the deflection coil. Were it not for the damping tube used in the secondary of the output transformer this transient voltage would produce a series of damped oscillations, occurring at the resonant frequency of the deflection coil and the distributed circuit capacity. This one alternation, however, has a peak amplitude in the thousands of volts and can be used, if properly rectified, to develop a high direct voltage.

The purpose of the horizontal sweep output transformer is to present the proper loading for the tube from the low impedance of the deflection coils. In so doing, the transformer-turns ratio from primary to secondary is step-down. This means that going from secondary to primary the transformer is actually step-up, and the transient voltage developed in the secondary is further increased in amplitude by the step-up of the transformer. Therefore, the high-voltage pulse is still greater in amplitude in the primary of the output transformer. A higher boost in amplitude is also obtained by using an extended primary winding, so there is a greater step-up between secondary and primary, the plate being tapped on at some point along the primary where the proper impedance match is made to the deflection coil.

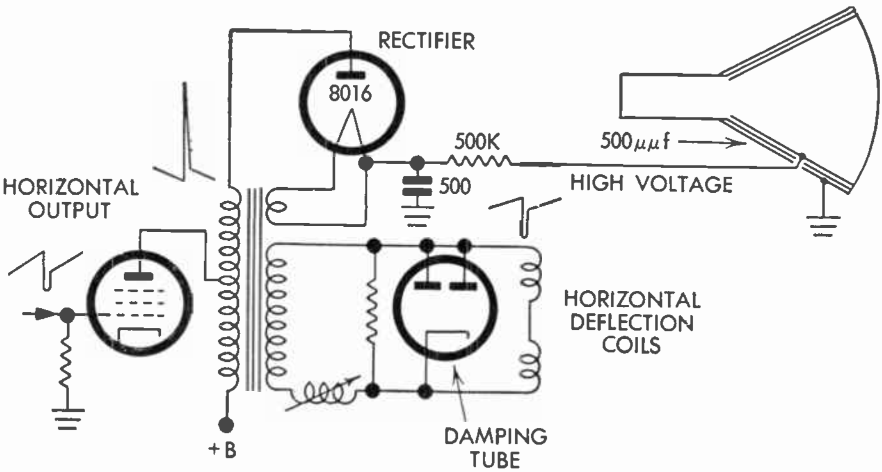


FIG. 123 Flyback High-Voltage Supply

The top of the primary winding is attached to the plate of a high-voltage rectifier tube which rectifies the transient voltage, which is now filtered to produce an essentially constant high-voltage d-c potential. It is important to note that the transient voltage itself is extremely short in duration, and occurs during a portion of the horizontal retrace period. The actual spacing between pulses is relatively long. Therefore, if a d-c voltage is to be obtained the charge on the filter capacitor must be held during this interval. The actual frequency of any ripple voltage that might be apparent is the repetition rate of the horizontal sweep system (approximately 15,750), because one transient pulse is developed for every sweep period. It is apparent, therefore, that the ripple frequency is relatively high, and rather small value capacitors can be used to do the filtering, a typical size being 500 micromicrofarads. The time constant of the filter capacitors and resistor is sufficiently long to prevent any serious discharge of the filter capacitors during the interval between transient pulses. Fortunately, it is no difficult task to do this because the actual current required from the voltage source is extremely small and, therefore, the resistive component of the discharge circuit is extremely high. Actually, as shown on the schematic, the second filter capacitor of the high-voltage system is the capacity that exists between the second-anode coating on the inside of the picture tube and the grounded outside coating, which is external to the glass of the picture tube, the glass serving as a dielectric. The effective capacity between coatings is approximately 500 micromicrofarads. As considerable charge can be retained on this capacity and, although it is not dangerous, it is often surprising and causes one to drop and injure the picture tube. It is wise always to ground the second anode momentarily to the outside coating before picking up the tube. The heater potential for the high-voltage rectifier is obtained by means of a small, few-turn pickup coil which is also wound on the core of the horizontal output transformer. The energy induced into the winding by the fast-changing magnetic fields is sufficient to excite the heater of the high-voltage rectifier tube. The pickup loop is, of course, properly insulated for high-tension voltages.

TRANSFORMER-TYPE HIGH-VOLTAGE SUPPLIES

Two transformer supplies excited by the 60-cycle power mains are shown in Figs. 124 and 125. If any amount of current were necessary at the extremely high voltages required, the transformer would be too massive. Fortunately,

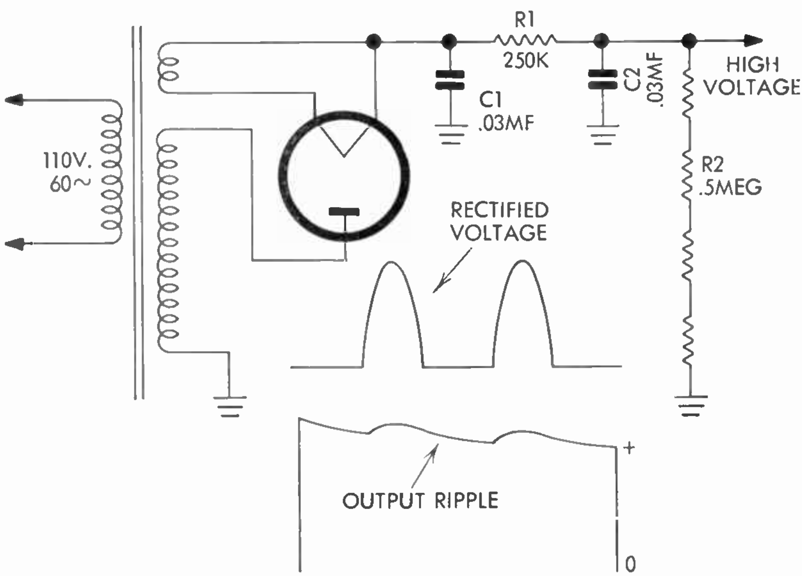


FIG. 124 Transformer High-Voltage Supply

only a small current is necessary, and although it is a high-voltage transformer, it need not be excessively bulky. Consequently, the windings are constructed of small-diameter wire and if, for any reason, a partial short is placed across the transformer, the windings open up. Many receivers use a safety resistor to limit the current, such as resistor *R1* in Fig. 125, which prevents excessive secondary current in case of a partial or complete short across the high voltage output of the power supply. Be extremely cautious when working on the high-voltage supply which is excited by the 60-cycle power mains because the current capabilities of such a system are very high. In addition, an appreciable charge is stored in the filter capacitors because they must be of a large capacity to filter the low-frequency, 120- or 60-cycle ripple frequency effectively.

The simple high-voltage supply shown in Fig. 124 consists of a half-wave rectifier and a simple *R-C* filter. When so little current is drawn from the supply, half-wave rectification and simple filtering are all that is necessary to remove the ripple frequency. Actually, the ripple component looks similar to a sawtooth wave, as shown in Fig. 124, because the capacitors *C1* and *C2* charge when the rectifier conducts and discharge when the rectifier is nonconducting or doing the opposite alternation of the applied 60-cycle sine wave. Every attempt is made to keep this ripple under 1 per cent of the d-c output voltage by properly choosing a long-time constant *R-C* filter. This filter does not permit the capacitor to discharge appreciably during the nonconducting alternation of the rectifier, in accordance with the current required. The time constant of the *R-C* filter must be high in comparison with the period of the wave which it must filter. Of course, a given time constant can be formed with a large *R* and a small *C* or a large *C* and a small *R*. The choice of *R* and *C* depends on the current required and the regulation expected. For example, if the current requirements are extremely light, the *R* of the time-constant circuit is very large and for a given time constant a relatively small value of *C* is required. If the current requirements are high or if better regulation is desired with a change in the brightness adjustment, a heavier load must be placed on the high-voltage power supply. This is accomplished by reducing the value of the bleeder resistor or the *R* of the time constant. In this arrangement, of course, a larger *C* must be used to produce an equivalent time constant, and the current capabilities of the entire system must be somewhat higher.

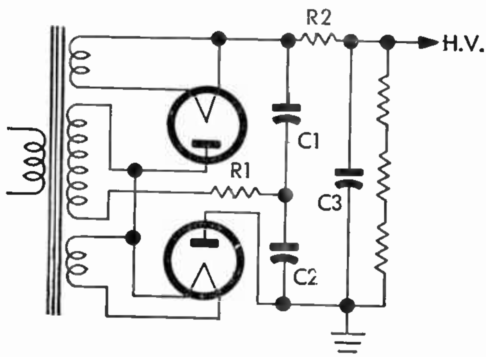


FIG. 125 Voltage-Doubler Supply

The power-supply filter of Fig. 124 is called a double-section filter because of the addition of an input capacitor *C1* and a series resistor *R1*. In this filter system the initial peak charge is placed on *C1*, and the discharge voltage, which is mainly developed across *R2*, is also filtered by capacitor *C2*. Consequently, the proper amount of ripple filtering can be obtained with the use of component parts of smaller value and size. The system of Fig. 125 uses a voltage-doubler output system in which capacitors *C1* and *C2* are charged on opposite alternations of the input sine wave. Consequently, in this circuit the ripple frequency is 120 cycles and the components of the *R-C* filter can be made smaller in value. Thus, it is ideal when very high potentials are desired. An extremely high voltage can be obtained which is double the sine wave peak output of the secondary of the transformer.

OSCILLATOR HIGH-VOLTAGE SUPPLY

Another safe and practical means of generating the high voltage required for picture-tube operation is the use of a low-frequency oscillator and the voltage step-up characteristics of a tuned transformer. Such a basic-oscillator

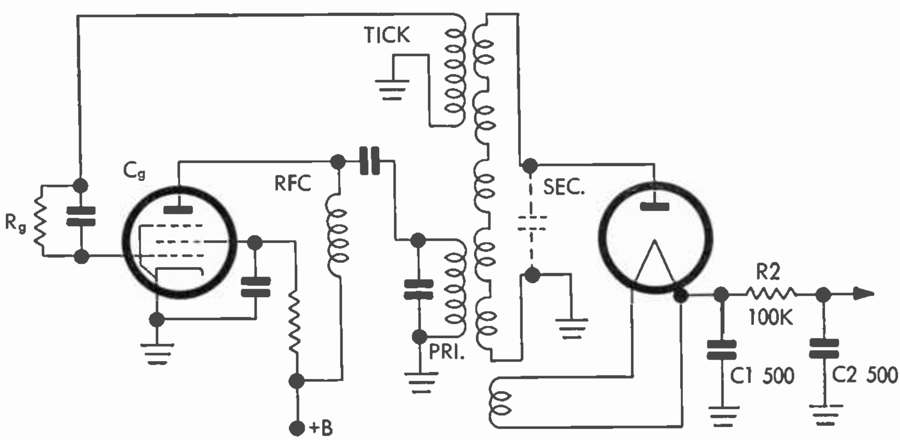


FIG. 126 Oscillator High-Voltage Supply

high-voltage supply is shown in Fig. 126. In this type of supply a beam-power tube or a power pentode is used to generate strong oscillations. For best stability and a constant output over a substantial range in frequencies a feed-back system with a tickler winding is used as the oscillating method.

The resonant frequency of the oscillation is set by the primary of the transformer; and feedback to the grid is through the tickler winding, grid bias being developed by the grid current flow through R_g and sustained by capacitor C_g. The primary and secondary windings are generally overcoupled to obtain a broad bandwidth and, in case of frequency drift due to heating or change in supply voltages, the output will remain essentially constant in spite of the frequency change. A high-voltage half-wave rectifier is placed across the secondary of the transformer and rectifies the resonant voltage developed across the secondary. Again, a simple *R-C* filter is used to remove the ripple frequency. The oscillators are designed to operate generally in the range somewhere between 50 kilocycles and 300 kilocycles. Consequently, with the relatively high frequency used, very little filtering is requiring. The oscillating high-voltage supply is again a very safe high-voltage system because of the small size of the filter capacitors, and the fact that any load placed on the output of the high voltage will cause the oscillator to stop operating or decrease the output voltage to a safe value.

To obtain a substantial voltage amplification in the tuned transformer it is necessary that the secondary be lightly loaded and therefore have a high impedance as compared to the impedance of the primary winding. Fortunately, the television picture tube again draws only a very small current and does not contribute an appreciable load to the secondary of the tuned transformer. Other factors that ensure a high voltage across the secondary are the proper choice of winding for the transformer and the proper choice of wire. In many of these high-voltage oscillating transformers Litz wire is used because of its inherent low resistance and ability to obtain high inductance with small-diameter wire at low resistance. Consequently, it is possible to construct a winding which has an extremely high inductance by using many turns of wire, and at the same time the resistance and consequent power loss will be low if the wire used has a very low resistance. Of course, the smaller the wire the closer it can be wound and, again, the more inductance the winding will have. It is evident, therefore, that Litz wire is ideal because of its small diameter, high inductance and low resistance. Inasmuch as the impedance of the secondary varies as the square root of the inductance over the capacitance, every attempt is made to keep the distributed capacity at a minimum and the inductance high. Therefore, the secondary is resonant to the proper frequency with the distributed circuit capacity and no physical capacitor is shunted across the secondary. In addition, a rectifier tube and component parts are chosen to have a minimum of capacity. Again, as with the transient voltage supply, the heater potential for the high-voltage rectifier is obtained from a small winding mounted on the transformer.

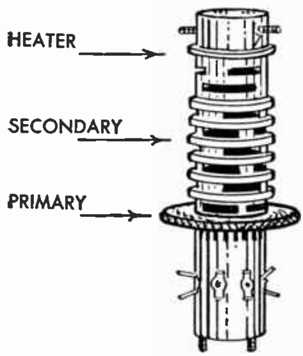


FIG. 127 R-F Transformer Assembly

The secondary winding of the transformer is generally a pie winding of five to six pies and, in some cases, even the primary is constructed in a pie-winding arrangement. The layout of a typical pie-wound coil is shown in Fig. 127. The secondary is made up of a series of five pie windings; each pie winding is universally wound with five to 10 turns per layer. The separation between pies is generally a little greater than the pie width. Too great a pie width increases the distributed capacity; too narrow a pie width reduces the mechanical stability of the winding structure.

To analyze the advantages of a pie-wound secondary, assume a pie-wound secondary consisting of five pies and having a total output voltage of 5,000 volts. This means that the voltage developed across each pie is 1/5 of 5,000, or 1,000 volts. Consequently, each pie can be designed for a 1,000-volt difference of potential. If the entire secondary was universally wound, that is, layer upon layer, the last turns of the winding would be reasonably close to the beginning turns; consequently, it would have to be designed for a breakdown potential of 5,000 volts. Inasmuch as the start and finish of the pie-wound secondary are very well separated, only the breakdown of the individual pie must be considered. Likewise, the distributed capacity is not felt between the early windings of the secondary and the final windings but just between the windings of the individual pies. The distributed capacity of each pie adding in series causes a lower distributed capacity to exist between the start of the secondary winding and its termination.

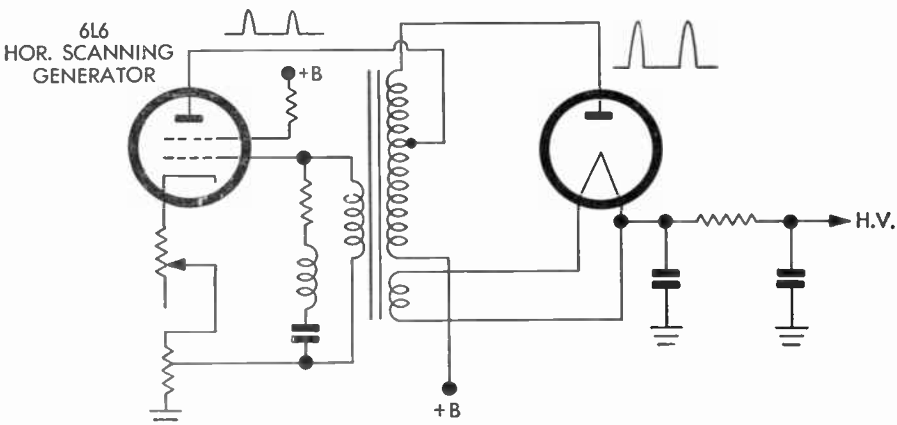


FIG. 128 Beam-Relaxor High-Voltage Supply

In a number of the higher voltage supplies the heaters of the high-voltage rectifiers obtain their current from a special resonant circuit. Instead of having a heater winding as a part of the oscillator transformer, a resonant circuit is associated with each heater-cathode circuit and the bursts of rectifier current flow set it into oscillation. The resultant current is taken off in autotransformer fashion to supply heater excitation.

An ingenious high-voltage supply is that associated with the beam-relaxor horizontal-scanning generator developed by Farnsworth. In this circuit (Fig. 128) no sawtooth oscillator is required; the deflection amplifier functions as an oscillator. The beam-relaxor oscillator develops a short, positive pulse in the plate circuit, which is again stepped up in amplitude by the autotransformer arrangement of the secondary and supplied to the plate of a high-voltage rectifier. A simple *R-C* filter is again used to remove ripple.

PULSE-TYPE SUPPLY

Another system for generating a high transient voltage which can be rectified and used as a high-voltage d-c potential is the pulse-type high-voltage supply (Fig. 129) used by DuMont. In this supply the grid of the pulse amplifier is driven by a short, positive pulse such as might be obtained from the grid of a blocking-tube oscillator. This sharp pulse drives the pulse amplifier near saturation and then sharply drops, causing a sharp change in plate current. The change in plate current develops a high-voltage transient in the inductor which is stepped up to a high peak voltage by the autotransformer connection. The top of the winding is again connected to a high-voltage rectifier which obtains its low-current heater voltage from the same autotransformer by mutual coupling. A simple *R-C* filter is again used in the output circuit.

The DuMont high-voltage generator (Fig. 129) uses a blocking tube oscillator triggered by a pulse from the horizontal-sweep output system. This sharp negative pulse overcomes minus bias on the blocking tube and causes it to oscillate each time it arrives. The sharp grid voltage is then applied to the beam-power amplifier, driving grid first near saturation and then sharply to cutoff. This sharp change in plate current through *T2* generates a high-voltage transient which is stepped up by autotransformer action and applied to rectifier. If driving sync pulse from horizontal-sweep system is absent, high voltage is not generated and picture-tube screen is protected.

A novel control system is used to vary the screen voltage of the amplifier to keep a constant output with changes in power-circuit voltages. For example, if output voltage attempts to fall, grid voltage on second 6SN7 section falls and plate current decreases. This means a higher plate voltage and, therefore, a higher voltage applied to the screen of the amplifier, which permits a greater plate-current change and a boost in output transient, compensating for original loss. A VR tube in the cathode of the regulator triode serves as fixed grid-voltage level.

High-voltage d-c can be varied over a limited range by controlling voltage amplitude at the grid of the amplifier. Also screen voltage is not applied to amplifier until remainder of circuit is warmed up, preventing an intense spot on the picture-tube screen.

In summation, there are a number of methods which can be used to generate a high-voltage low-current source, and although the d-c potential is in the thousands of volts these supplies are relatively safe—with the exception of the transformer type driven by excitation from the 60-cycle power mains. However, the trend is away from the transformer type toward the safety of the transient-voltage and oscillating-voltage supply. All of these higher frequency high-voltage systems are practical and efficient but must be well shielded to prevent interaction with other circuits associated with the television receiver. Fortunately, some interaction can be tolerated because the

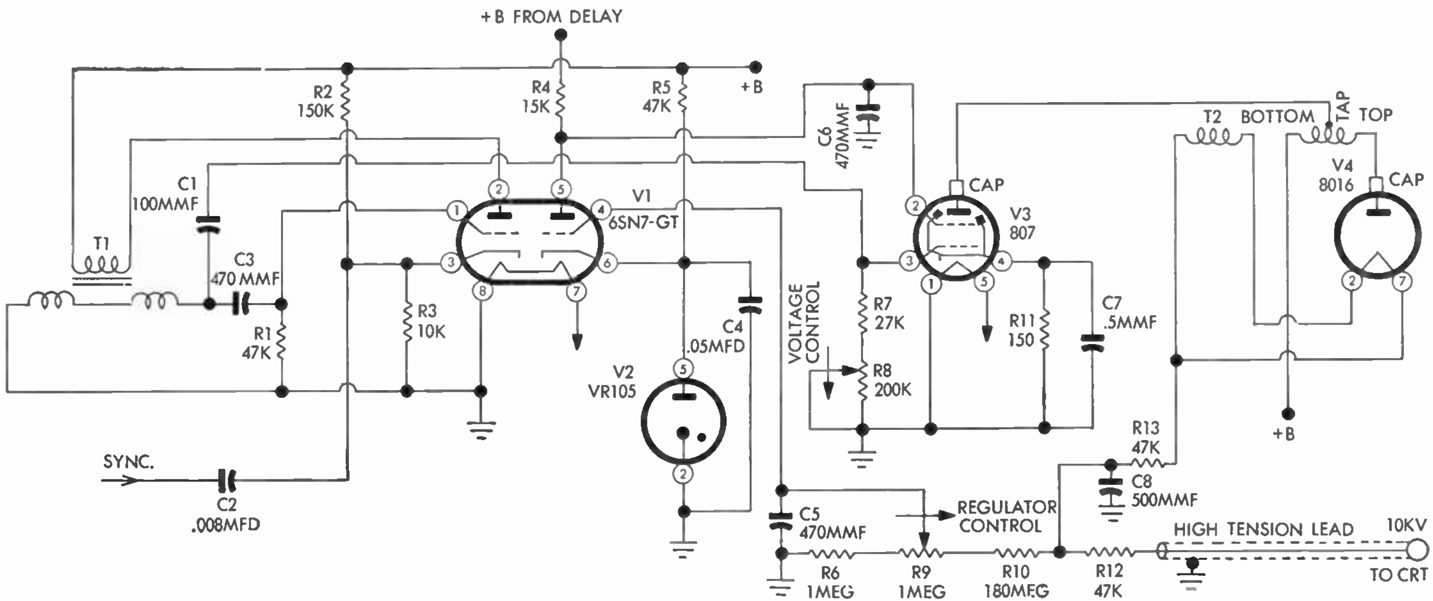


FIG. 129 DuMont Pulsed High-Voltage Supply

generation of the high-voltage pulse is during the retrace of the horizontal sawtooth when the picture beam is normally cut off.

81. *Commercial Receiver Signal and Voltage Circuits*

A commercial picture-tube circuit as used in the RCA television receiver for a small television tube is shown in Fig. 130. This schematic shows the many operating voltages and signals required to properly present the picture on the television screen. To the grid of the picture tube is applied the negative-going composite television signal from the video-amplifier output utbe. This negative-going single-polarity signal is augmented with two d-c voltages. One of these voltages is contributed by the d-c restorer circuit which maintains a steady charge on capacitor *C1* because of the small current discharged through resistor *R2*. A d-c component of grid bias is also taken off the brightness control *R3*, which is variable to properly set the negative voltage applied to the grid of the picture tube. The correct d-c potential for the accelerating grid of the picture tube is obtained from the low-voltage power supply.

The anode voltage for the picture tube of 9,000 volts is obtained from a transient flyback high-voltage supply. The negative transient developed across the horizontal-deflection coil during the horizontal retrace is stepped up in amplitude by transformer action and inverted to apply a short-interval high-voltage positive pulse to the plate of the 8016 high-voltage rectifier. This rectifier, which requires only approximately a 50-milliampere current for the heater, obtains it from the same horizontal-output transformer by means of a few-turn pickup, coupled near the high-voltage windings. The high voltage pulse is rectified and filtered and serves as the second anode potential of the picture tube.

A sawtooth of current is present in the horizontal-deflection coil as developed by the horizontal-output tube. Likewise, a vertical sawtooth of current is developed in the vertical-deflection coil by the vertical-output tube. In addition to the horizontal- and vertical-sawtooth currents a d-c component of current must be passed through each set of coils to properly center the picture as a whole on the fluorescent screen. These d-c components of current are taken off two 20-ohm potentiometers which are in series with the B-supply voltage between the 300- and 275-volt points. By controlling the amount and direction of current flow with the centering controls the picture can be properly centered horizontally and vertically.

It is also necessary to pass the proper d-c current through the ion trap coil and the focus coil. These currents are obtained off a series of bleeder resistors across the minus 105 negative-voltage supply. The current through the focusing coil is controlled by means of a focusing potentiometer so it can be adjusted to obtain the sharpest picture. The two ion trap coils also obtain their current from the same bleeder network. It is evident that the proper operation of the picture tube is dependent on a number of signals and voltages;

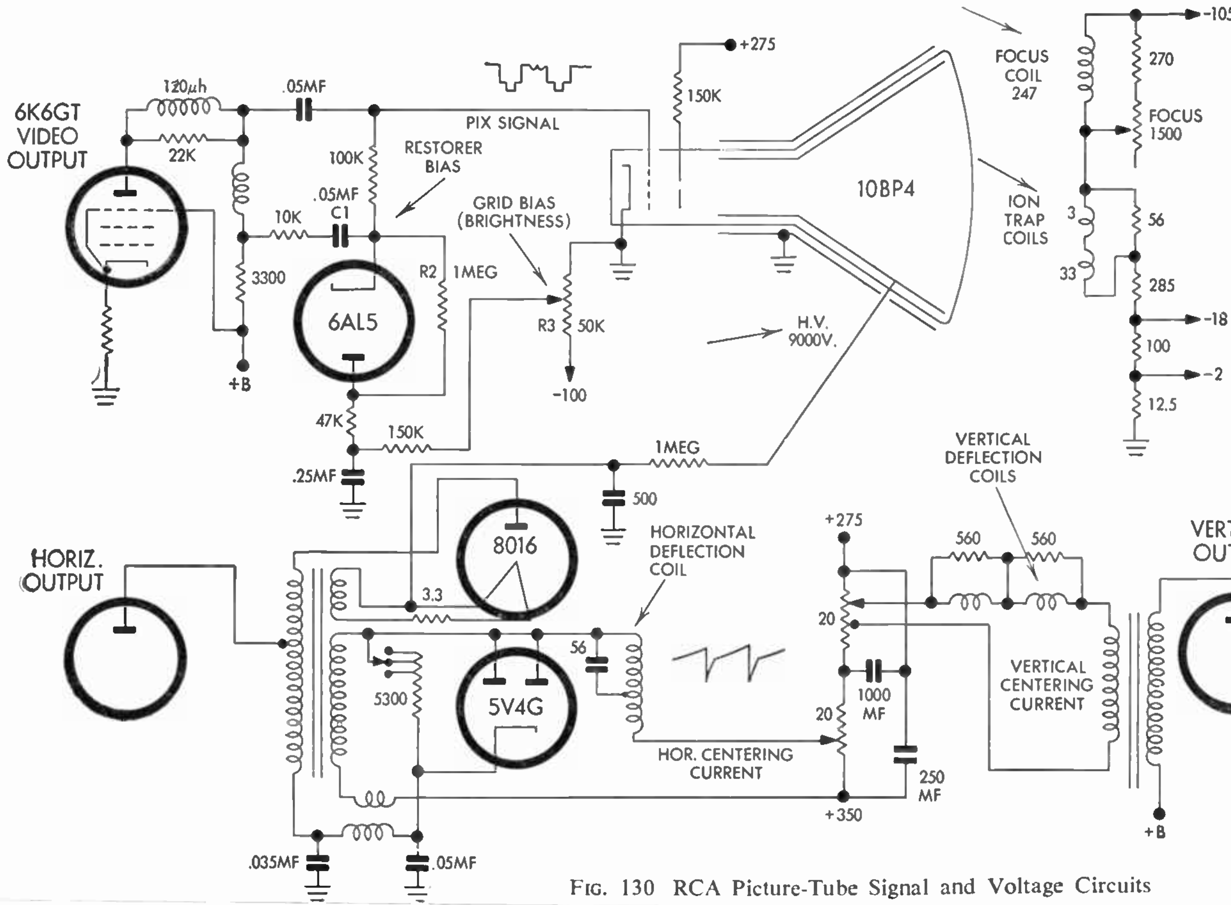


FIG. 130 RCA Picture-Tube Signal and Voltage Circuits

and what is more, troubles can, in most cases, be isolated by observation of the picture on the screen. The nature of the malfunction directs the television technician with a thorough knowledge of circuit operations toward the defective circuit.

There are three basic magnetic-deflection picture-tube circuits as a function of type of picture tube. There are circuits for picture tubes that employ magnetic focusing and magnetic deflection, electrostatic focusing and magnetic deflection, and the DuMont-type self-focus picture tube. It is possible to apply the video signal to the grid or cathode of the picture tube and to supply it by direct coupling or capacitive coupling. Each basic system has its own methods of application for signals and supply potentials.

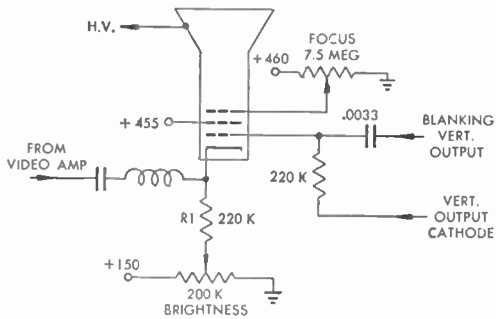


FIG. 131 Zenith Picture-Tube Circuit

A typical picture-tube circuit using electrostatic focus is that of the Zenith method, illustrated in Fig. 131. The positive-sync video signal is applied through a proper peaking system to the cathode of the picture tube and appears across resistor *R1*. The low side of the cathode resistor is attached to the brightness control which is returned to a positive supply-source. A small positive bias is applied to the control grid of the gun along with a vertical blanking pulse derived from the vertical output tube of the television receiver. This pulse has been properly shaped to drive the gun to cut-off and to prevent the appearance of vertical retrace lines on the picture. A focus control permits regulation of the focus-electrode potential and is a preset adjustment. A typical picture-tube circuit for a tube using magnetic deflection and magnetic focusing (external focus coil) is illustrated in Fig. 132. In this circuit the video is applied through a suitable peaking system to the grid of the picture tube, while brightness is regulated with a controlled positive potential applied to the cathode of the gun. In addition, both vertical and horizontal blanking pulses (positive in polarity) are applied to the cathode to blank out both vertical and horizontal retrace lines. Horizontal blanking is obtained by prop-

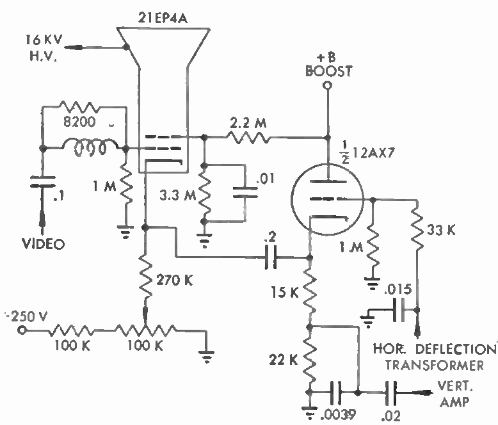


FIG. 132 GE Picture-Tube Circuit

erly shaping a pulse taken from the horizontal deflection-transformer and applying to the grid of the cathode follower. The vertical blanking pulse, derived from the output of the vertical amplifier, is applied to the cathode circuit of the triode, where it is combined with the horizontal blanking, both pulses being applied to the cathode of the picture tube. The accelerating grid receives its supply potential from a voltage-divider circuit to the plus-B boost-voltage supply, which also applies a high positive potential to the plate of the cathode-follower blanking tube.

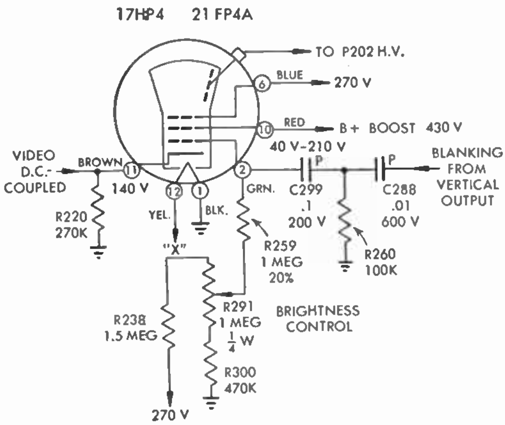


FIG. 133 DuMont Self-Focus Circuit

A DuMont self-focus picture-tube circuit, Fig. 133, receives a direct-coupled signal at its cathode from the preceding video amplifier. Of course, at this point there is not only the a-c video signal, but a direct-coupled d-c positive voltage from the plate circuit of the video amplifier. To obtain suitable operating bias for the electron gun and to have a negative potential of proper level established between grid and cathode of the gun, it is necessary to return the control grid of the gun to a positive supply point. Brightness of the picture can be adjusted through regulation of this positive potential applied to the grid of the gun. In addition, through the usual differentiating circuit *C288* and *R260*, a vertical retrace blanking pulse is developed and applied to the same grid. In any direct-coupled system it is necessary to establish the proper d-c voltage for the proper negative difference of potential between the grid and cathode of the gun.

In the newer type of self-focus picture tubes it is necessary to apply a positive potential to the focusing electrode, while in the older self-focus tubes the focusing electrode was returned internally to the cathode of the gun.

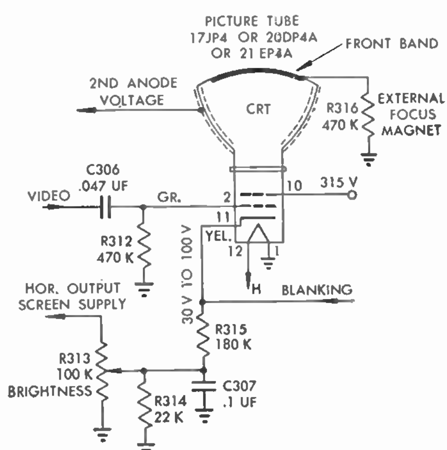


FIG. 133a Philco Picture-Tube Arrangement

In the Philco picture-tube circuit, Fig. 133a, the video signal is applied to the grid of the gun while d-c brightness control is established in the cathode circuit. In this particular circuit an external focusing magnet, instead of a coil, is employed. Blanking is applied as a positive pulse to the cathode of the gun. It is a fact that in most picture-tube circuits there is some interaction between brightness setting and picture width. In the usual picture-tube circuit an adjustment of picture-tube brightness also influences the width of the picture because of the change in gun biasing and beam content. In this particular circuit an adjustment of the brightness control also changes the d-c potential applied to the screen grid of the horizontal output tube, and consequently, there is a corresponding deflection adjustment that holds the width of the raster constant, despite changes in brightness setting.

QUESTIONS

1. What is the crossover point and its significance in the electron gun?
2. What factors determine picture-tube spot size?
3. Explain magnetic focusing.
4. Explain magnetic deflection.
5. What is the function of first and second lenses of an electron gun?
6. What is deflection sensitivity?
7. Describe methods of removing ions.
8. Compare at length the characteristics of magnetic- and electrostatic-deflection picture tubes.
9. What is the function of accelerating grid?
10. Explain the zero-first-anode-current gun.
11. What are the characteristics of a satisfactory fluorescent screen?
12. What is a balanced deflection system?
13. What is deflection angle and factors which influence it?
14. What are the advantages of metal-backed screen?
15. Draw schematic of and describe complete picture-tube circuit using magnetic deflection and focusing.
16. Draw schematic of and describe complete picture-tube circuit using electrostatic deflection and focusing.
17. Name and tell purpose of all voltages and signals applied to picture tube.
18. Explain each of a number of methods of obtaining high voltage for a picture tube.
19. What is a pie winding and what are its characteristics?
20. Describe and compare a number of commercial picture-tube circuits.

Chapter 7

SYNC AND INTER-SYNC SYSTEMS

82. *Pulse Techniques*

Television picture information is not transmitted continuously, but is interrupted at intervals for insertion of the rectangular sync and blanking pulses. These pulses, which are being transmitted approximately 20 per cent of the total frame time, are the pulses which lock the transmitting- and receiving-tube scanning beams together and prevent the appearance of spurious signals on the television image during the synchronizing period.

We have traced the path of a television signal as it originates at the transmitter and as it terminates at the grid of the picture tube in the receiver. We have studied the action of the picture tube when its control grid is excited by the picture signal, and also observed what occurs to the picture-tube beam during the transmission of the rectangular blanking pulses. However, we have not discussed at length the uses of the sync pulses at the receiver. Actually before the sync pulses can be used to synchronize the horizontal- and vertical-sweep generators, two operations must be performed. First, the sync pulses as an entirety must be segregated from the composite television signal, and then vertical and horizontal components of the sync must be shaped for proper application to the sweep oscillators and to prevent interaction between vertical and horizontal components. A composite signal from which the sync can be extracted is generally taken off in the video output stage or at some earlier stage in the video amplifier. In some receivers a separate sync detector is employed. The picture and blanking components of the composite signal are then removed, leaving only the composite synchronizing signal. The synchronizing signal itself then is shaped into vertical and horizontal components.

Synchronization of the horizontal and vertical sweep of the television receiver is performed by rectangular pulses similar to the blanking pulses, but of shorter duration. Instead of using the synchronizing pulses in the same form in which they are received, these pulses are shaped in accordance with their application as vertical or horizontal synchronization. The three synchronizing pulses—equalizing, horizontal, and vertical—are respectively two and one-half, five, and approximately twenty-five microseconds in duration. Actually the vertical sync block consists of six spaced vertical sync pulses to form a total duration of approximately 190 microseconds. The leading edges of the horizontal and vertical sync pulses synchronize the horizontal sweep; the pulse width or duration of the vertical sync block locks in vertical sweep. The leading edges of the equalizing pulses also synchronize the horizontal oscillator and sustain the rigidity of the interlace, as will be explained.

RECTANGULAR PULSES

The rectangular sync pulse (Fig. 134) consists of a base frequency and a number of low-order and high-order harmonics, which are vested in the leading and trailing edges and flat top of these pulses. These high harmonics constitute fast changes in voltage per unit time while the low-order harmonics

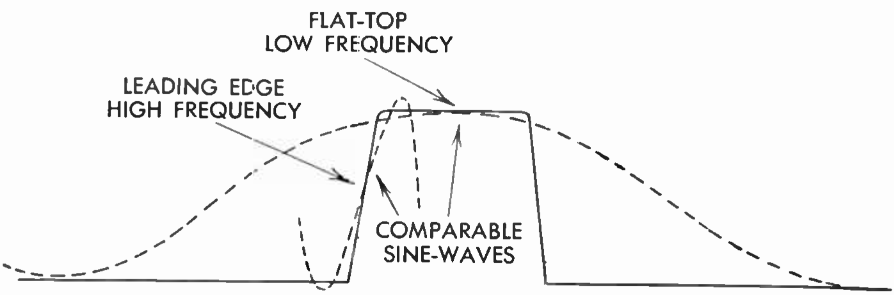


FIG. 134 High- and Low-Frequency Characteristics of a Pulse

are vested in the duration or flat tops of the pulse, which represent a slow change in voltage per unit time. For example, the leading and trailing edges of the sync pulses—horizontal, equalizing, and vertical—must rise from minimum to maximum in approximately ¼ microsecond and must fall from maximum to minimum in the same time. A comparison with a sine wave tells us that the frequency of any sine wave which must rise from minimum to maximum in ¼ microsecond is a 2-megacycle wave. The period of a 2-megacycle wave is ½ microsecond, and therefore this 2-megacycle wave rises from minimum to maximum in ¼ microsecond. This approximation tells us the frequency-response requirements of any stage which must pass the leading and trailing edges of the television synchronizing pulses.

The flat top of a pulse represents a low frequency because it is a sustained voltage level which must be held constant for the duration of the pulse. Therefore, the duration of the pulse determines the low-frequency-response requirements because the longer the pulse the longer the time interval the voltage must be held at a level value. To sustain an absolutely flat, flat top the period of a comparable sine wave would have to be many times longer than the duration of the pulse because the pulse itself is a constant voltage level. But the comparing sine wave rises and falls sinusoidally and maintains its peak value for only a small percentage of the total period of the sine wave. It is evident, therefore, that the frequency-response requirements of any stages which must pass the pulse with fidelity are dependent on the steepness of the leading edge of the pulse for its high-frequency limit and the duration of the pulse for its low-frequency limit. It is also this very characteristic of a pulse which permits segregation of components of the pulse into horizontal and vertical sync.

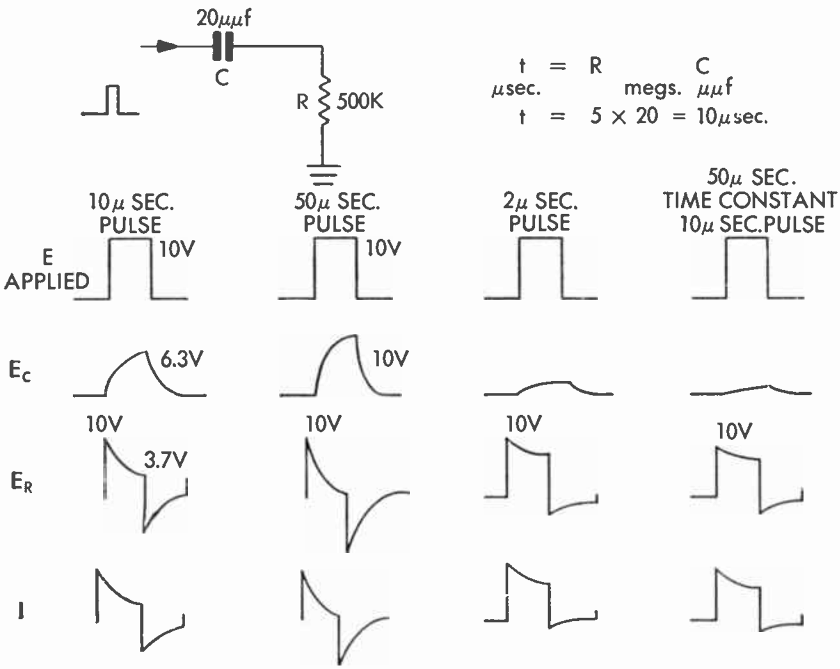


Fig. 135 Time Constant and Coupling Pulse

The more exact method by which to interpret the response of a circuit to a pulse is in relation to the time constant of such a circuit. In this interpretation the pulse is considered to be a d-c voltage which is cut on and off at specific intervals for specific durations. In a simple *R-C* combination (Fig. 135) the time required to put a 63 per cent charge on the capacitor is termed the *time constant* (exponential rise of voltage on the capacitor, charging toward the applied-voltage level). The time constant in seconds is equal to the product of the resistance in ohms and the capacity in farads. For practical use in television circuits time constant is most often in microseconds, and is equal to the product of the resistance in megohms and the capacity in micromicrofarads. Thus, if a 10-microsecond pulse is applied to a circuit with a 10-microsecond time constant the capacitor will charge to 63 per cent of the applied pulse amplitude at the end of the 10-microsecond pulse (Fig. 135). If the *pulse* is applied for five time constants or is five times longer in duration than the time constant of the circuit, the charge on the capacitor at the end of 50 microseconds will be almost the peak amplitude of the applied pulse. If the time constant of the circuit is made five times longer than the pulse duration, the charge on the capacitor at the end of the pulse will be relatively low.

The instant the pulse is applied and before the capacitor begins charging, the voltage across the resistor is maximum and equal to the peak amplitude of the pulse. At this time, the current flow is maximum because there is a current circulating around the circuit which will attempt to charge the capacitor to the peak value of the applied voltage. As the charge builds up on the capacitor this potential subtracts from the potential of the applied voltage and the current in the circuit therefore is reduced. The voltage drop across the resistor, as shown in Fig. 135, is reduced a corresponding amount. Therefore, as the charge builds up on the capacitor, the current flow gradually

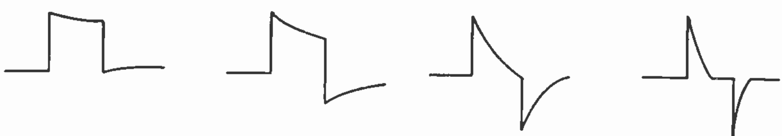


FIG. 136 Poor Low-Frequency Response and Pulse Distortion

reduces and voltage drop across the resistor decreases. Thus, if a 10-volt pulse is applied, at the end of the first time constant, 6.3 of the original 10 volts will appear on the capacitor and 3.7 across the resistor. At the end of six time constants the capacitor will have been fully charged so far as any practical utilization is concerned (theoretically, it never reaches full charge). Therefore, the current flow will be zero and the voltage drop across the resistor zero also. If the pulse is applied for a duration much shorter than the time constant, or if the time constant is increased to have it much larger than the pulse duration, the charge on the capacitor will be very slight and consequently the voltage across the resistor will not decrease an appreciable amount. It is evident, therefore, if the capacitor and resistor we are talking about were the coupling condenser and grid resistor of an interstage coupling circuit, the pulse would appear across the resistor with the best fidelity when the time constant of the circuit is much greater than the duration of the pulse.

Actually, to maintain an absolutely linear flat top on our pulse it is necessary that the time constant of the grid-capacitor, grid-resistor combination be almost a hundred times the duration of the pulse. Likewise, the longer the duration of the pulse the larger the time constants become in order to maintain this ratio.

A pulse is often used to check the frequency response of video amplifier because the extent of the taper of the flat top is an indication of the low-frequency response of the amplifier. The poorer the low-frequency response the more decided the pulse is distorted. Pulse distortion as the low-frequency response becomes progressively poorer is shown in Fig. 136.

TIME CONSTANT AND THE LEADING EDGES OF A PULSE

We have studied that the presence of the distributed capacity across the plate-load resistor of our video stages causes a loss of high frequencies. Likewise, this very same capacity across the plate load will cause distortion of the leading edges of a pulse because the leading edges of the pulse are made up of high-frequency components. Again a better understanding can be obtained thinking in terms of time constants. Thus if a pulse is applied to a parallel resistor-capacitor combination, and the leading edge of that pulse

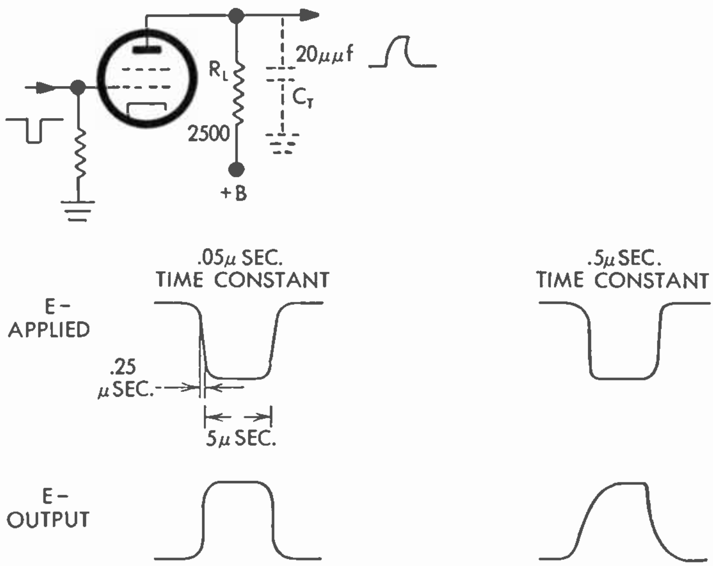


FIG. 137 Time Constant and Pulse Leading Edge

rises from minimum to maximum in ¼ microsecond, it is necessary that the distributed capacity be capable of charging up to that level in the same amount of time. If the time constant of the plate resistor and distributed-capacity combination is long, the capacitor will not assume the maximum charge in the ¼ microsecond allotted, and will continue to charge during the flat top portion of a pulse. Consequently, the pulse will be distorted (Fig. 137) and this steepness of the leading edge will be lost. If the time constant is made very short in comparison to the rise of the leading edge it will have no difficulty following the sharp rise of the voltage at the beginning of the pulse. Thus, to sustain the fidelity of the leading and trailing edges of all pulses it is necessary that the shunt time constant be short in comparison to the rate at which the pulse climbs. If this is not so, the lag imposed by the long time constant of the shunt-R-C combination will cause the steepness of the leading edge to be lost, and the pulse will fold over. To keep this time constant short the size of the plate resistor must be relatively low and the distributed capacity must be kept at a minimum—the same requirements imposed on a video amplifier.

A squared pulse, therefore, can also be used to check the high-frequency response of a video amplifier because a loss in high-frequency response means a long time-constant-shunt combination, which means the steepness of the pulse will be lost in accordance with the extent of the high-frequency loss. Figure 138 shows us the extent of the pulse distortion as the high-frequency response becomes progressively poorer.

FIG. 138 Poor High-Frequency Response and Pulse Distortion

83. *Sync Clipping*

Before the sync pulses are shaped for application to the horizontal- and vertical-oscillator circuits the sync as an entirety must be separated from the composite television signal. The sync separator or clipper must separate that portion of the composite television signal above the blanking or black level. The three signals present above the blanking level are the horizontal sync, the equalizing sync block, and the vertical sync block. The leading edges of all the sync pulses are used to keep the horizontal sweep in synchronism, and the vertical sync block is used to synchronize the vertical sweep. The actual separation occurs in a circuit which holds all portions of the signal below the blanking level at cutoff. Most sync separators are signal-biased; that is, the current drawn in the input circuit at the peak of the sync tip supplies the bias for the stage. This current and resultant bias is sufficient to hold all portions of the signal below the blanking level at a cutoff value.

Three typical sync separators are shown in Fig. 139, using a diode, triode, and pentode tube. In the diode separator illustrated, a positive-going composite signal is applied to the plate of the diode. The peak of the sync tip draws maximum diode current, and therefore capacitor C is charged to peak value during the sync tip interval of the composite signal (capacitor charges quickly because of the short time constant formed by C and low resistance of conducting diode). The charge placed on C by the peak diode current flowing through R is sufficient to prevent conduction of the tube until the signal amplitude is near the blanking level. Consequently, the only portion of the signal which is developed across the diode-load resistor occurs when the diode is conducting. The diode conducts only for signal amplitudes which are higher in level than the blanking level. Only the sync pulse itself is higher in level.

The charge placed on C during the sync tip interval by the peak diode current is held at a constant value by the long time constants of R and C.

Actually, the capacitor is recharged to peak value approximately every 60 microseconds, and the time constant of *R* and *C* must be sufficiently long to hold this charge level between recharging intervals. Thus the time constant of *R* and *C* is much longer than the interval between sync tips, and is in most cases in excess of 25,000 microseconds.

In some cases, external bias is also supplied to the diode to make certain that the diode will clip off at the blanking level and will not remain cut off until the peak of the sync tip. As shown in Fig. 140, sync separator can be wired to accept either a positive- or a negative-going composite signal just by changing the electrode to which the composite signal is applied. An interesting fact about the charge and discharge cycle of capacitor *C* is made evident. During the charge of capacitor *C* the diode is conducting, and con-

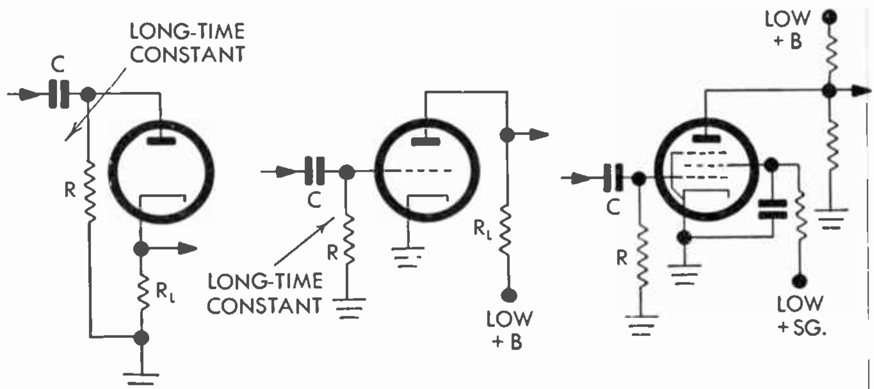


FIG. 139 Typical Sync Separators

sequently the time constant of the conducting diode's d-c resistance and the capacitor *C* determines the fast rate at which the capacitor reaches peak value. During the discharge cycle the diode is cut off and the discharge is through the large resistor *R*. Consequently, the capacitor quickly charges during the sync tip interval because the time constant is short. It is able to sustain a fixed charge during the remainder of the line interval because of the long time constant of *C* along with *R*. Thus we find the diode sync separator very similar to the d-c restorer, with the exception that the d-c restorer conducts only at the very peak of the sync tip while the diode separator conducts between the blanking and the sync-tip levels.

It would be possible to operate a diode separator with external bias alone by making certain that the bias applied would not permit the diode to conduct until the amplitude of the blanking level was overcome. This means that a certain level signal would have to be applied at all times. The advantage of the signal-bias separator is that it is self-compensating, over a limited range, to changes in amplitude of the incoming signal. In this respect it is again similar to a d-c restorer because the actual bias placed on capacitor *C* is again dependent on the peak amplitude of the sync tip; when the signal level decreases both the sync tip and the blanking level of the signal decrease in amplitude. If our signal has decreased in amplitude it means that the externally biased diode would not conduct until the signal amplitude is in excess of the blanking level; consequently, a portion of our sync amplitude would be removed. However, with signal bias the decrease in initial amplitude of the sync tip causes less of a charge to appear on the capacitor *C*, consequently less amplitude is necessary to reach the conducting level of the diode. Once again, conduction of the diode will occur at the blanking level and the entire sync pulse will appear in the output.

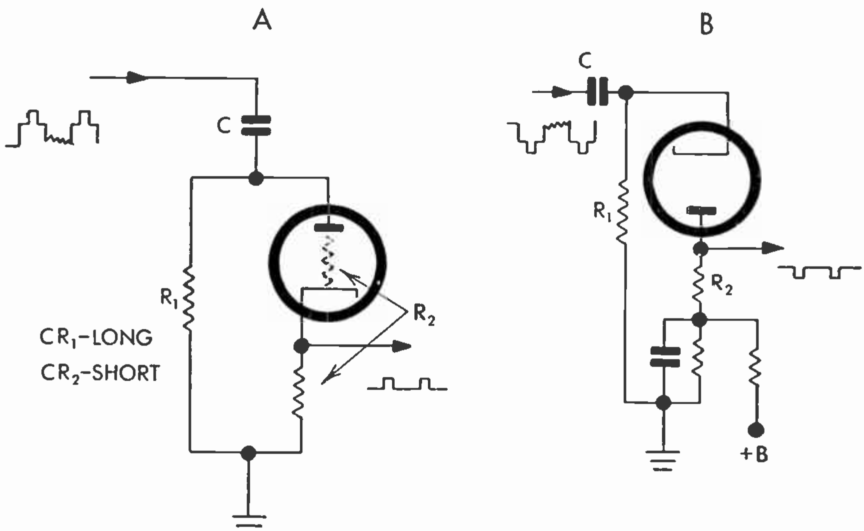


FIG. 140 Diode Separators—Positive and Negative

A typical triode separator is shown in the second drawing of Fig. 139. This separator draws grid current during the peak of the sync tip, and this current charges capacitor *C* to a level which permits conduction of the tube only when the applied signal is higher in amplitude than the blanking level. Inasmuch as the tube conducts only on signal amplitudes greater than the blanking level, only the composite sync signal will appear in the plate circuit. The remainder of the signal is removed by cutoff. Again it is the time constant of *R* and *C* which holds this bias level constant at the grid of the sync separator. Charge is established each line interval by the sync tip. If there is a change in signal level there is also a change in the peak grid current drawn and a change in the bias level. This change in bias level compensates for any change in signal amplitude, and once more conduction level will appear at the blanking level of the signal for a wide range of signal amplitudes.

Again a better understanding of sync separator and d-c restorer operation can be obtained by comparing the two actions as presented in Fig. 141. In drawing A, we observe the grid draws sufficient current to set the average bias of the stage the proper amount beyond cutoff to have cutoff bias level appear at the blanking level of the signal. When the signal amplitude is less, the grid current drawn and the bias are correspondingly lower, and again the blanking level appears at cutoff. In the grid-type d-c restorer, drawing B, the average bias set on the stage by the signal-biased grid is at some point

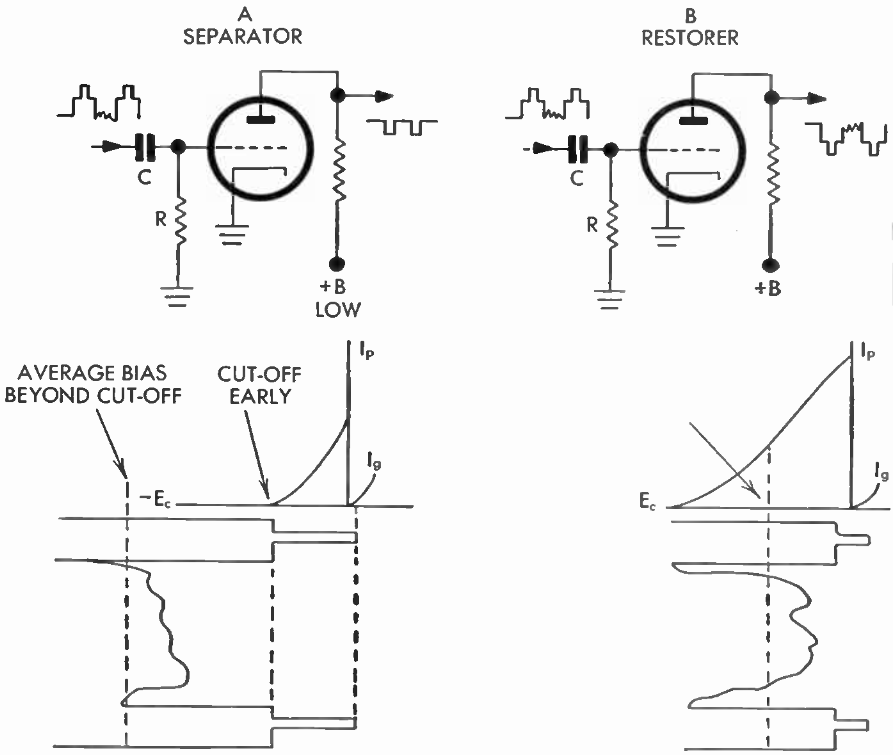


FIG. 141 Triode Sync Separator and Grid-Type Restorer

on the linear transfer of the tube. This places the picture content of the composite signal on the linear portion of the characteristic of the tube and keeps the blanking level essentially constant near the zero bias level of the tube. With a change in signal level, peak grid current changes, and the bias of the tube changes the correct amount to set the blanking level at a point near zero bias again. Thus the major difference between sync separation and d-c restoration is the point at which the bias is set—in one case beyond cutoff and in the second case on the linear portion of the characteristics of the tube. A number of factors decide what this bias level will be—signal amplitude, grid current characteristics of the tube, time constant, cutoff bias for the tube, and plate voltage on the tube. If a tube is to be used as a separator, it is necessary to have cutoff occur early. Consequently, the sync separator, as compared to the d-c restorer, operates with a lower plate voltage, which means cutoff will occur at a lower bias and, in addition to that, substantial grid current flow is more readily obtained.

A signal-biased pentode, cathode-grounded and with low screen and plate voltages, can also be used as a sync separator. Again it is the grid current drawn during the sync tip which sets the bias of the tube at the proper level to have all portions of the signal below the blanking level beyond cutoff. Advantages of the pentode-type separator are that good clipping will occur over an appreciable signal-amplitude range and the output of the pentode separator will be essentially constant in amplitude over this range of signal amplitudes. Actually, the sync pulses which appear in the plate circuit are a result of a grid-voltage variation between cutoff and a positive grid voltage at which grid limiting occurs. Thus, if the plate voltage is held constant as it is with a bleeder network from B plus (third drawing of Fig. 139) the sync-signal amplitude in the plate circuit will be constant in spite of the variation in peak signal amplitude at the grid of the separator. Thus with appreciable change in signal amplitudes by fading or switching from one channel to another, the amplitude of the sync at the output of the separator will remain essentially constant.

Composite signal for application to the sync separator can be taken off the video section of the receiver at a number of points. In the older receivers it was customary to take the portion of the composite signal to be applied to the sync separator off the video detector. The composite signal can also be taken off one of the video amplifiers or, as is most often done in the modern receiver, from the coupling circuit between the video-output tube and the picture-tube grid. At this point, the composite signal is high in amplitude, and much of the noise greater in amplitude than the sync tip has been clipped off, improving the sync-to-noise ratio of the receiver. If the applied signal to the video amplifier is too great in amplitude, the sync pulse itself is severely limited, and consequently sync stability is affected. However, if a signal of this amplitude is permitted to reach the grid of the picture tube the contrast range is often destroyed and there is no particular significance to the loss of sync stability. If the contrast is properly adjusted a composite signal with the proper sync level is applied to the sync separator circuit.

84. *Differentiation and Integration*

The output of the sync separator (composite sync signal consisting of horizontal, equalizing, and vertical sync pulses) is diverted into two paths. One channel makes use of the leading edges of the pulses to synchronize the horizontal deflection system, while the second channel uses the long duration vertical sync pulses to synchronize the vertical oscillator. So far as television application is concerned, the circuit which utilizes the leading edges of the sync pulses is called a *differentiating circuit,* and consists of a simple high-pass filter which readily passes on to the output the high-frequency short pulses or leading- and trailing-edge components of the sync pulses. The integrating circuit is a simple low-pass filter which is not capable of following the leading edges of the pulses but which builds up a charge on the capacitor in accordance with the duration of the pulse. The length of the flat top of the pulse or its duration is, of course, a low-frequency component. Thus, by employing circuits which respond to the high- or low-frequency components of a pulse, we obtain a method of segregating horizontal- and vertical-synchronization components.

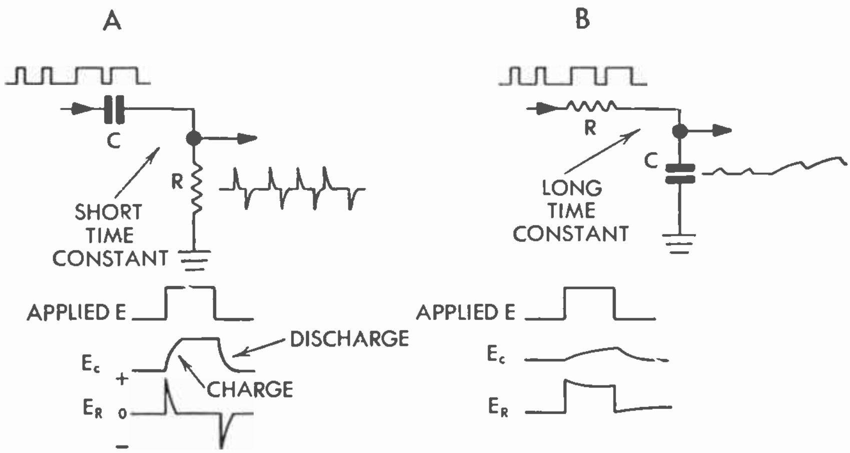


FIG. 142 Differentiating and Integrating Circuits

Differentiating circuits generally consist of a short time constant *R-C* combination (Fig. 142, drawing A). In this combination the time constant is short in comparison to the pulse duration and therefore the capacitor will charge hurriedly. At the instant the pulse is applied the voltage across the resistor will rise with it to maximum amplitude of the pulse. However, the capacitor will charge rapidly through the short time constant presented and will quickly charge to the peak pulse amplitude, at which level no further current flows and the voltage across the resistor has dropped just as quickly to zero. Consequently the voltage across the resistor is a sharp spike of voltage, while the charge on the capacitor is the same level as the peak amplitude of the applied pulse. This distribution, maximum charge on *C* and zero voltage across *R*, will continue for the duration of the pulse. At the termination of the pulse the applied voltage quickly drops back to zero and the capacitor quickly discharges. At the termination, therefore, the voltage drop across the resistor will be maximum negative and exactly the same value as the full-charge level of the capacitor (sum of voltage drops across the *C* and *R* will now equal zero). It is evident the voltage across R is now negative since the electron flow is now in the opposite direction because the charge is leaking off capacitor C. As this charge rapidly falls off C through the short time constant, the voltage across R becomes progressively less negative and falls toward zero at the same rate the capacitor charge is falling toward zero (always maintaining an algebraic sum of zero). Thus at the termination of the pulse, a sharp negative spike appears across the differentiating resistor. Inasmuch as the leading-edge spike is the timing voltage of the horizontal oscillator, the negative spike generated at the termination can be clipped off and removed. If for some applications a negative-going leading-edge spike is necessary, it is only required that the applied pulse to the differentiating circuit be negative- instead of positive-going.

The integrating circuit is a long time-constant R-C combination with the integrated vertical synchronization taken off the integrating capacitor (drawing B of Fig. 142). When a pulse is applied to an integrating circuit, the capacitor charges very slowly during the entire duration of the pulse. Consequently, the time constant of the integrating combination is much longer than the duration of the pulse. In a typical television receiver the differentiating-circuit time constant may be at some value between less than 1 microsecond and 10 microseconds; the integrating-circuit time constant may be from 500 to thousands of microseconds. This is extremely long. Only a charge equal to an extremely low percentage of the total pulse amplitude appears on the capacitor. The vertical sync pulses are of long duration and short spacing, and during the vertical sync bloc interval it is possible to build an appreciable charge on the integrating capacitor.

A simple inter-sync separating system along with the action of the differentiating and integrating circuit on both the horizontal and vertical sync pulses is demonstrated in Fig. 143. In the drawing waveform *1* represents the short-duration horizontal sync pulses which produce across the differentiating resistor a series of spiked voltages as shown in waveform *2*. Likewise, the leading and trailing edges of the vertical sync pulses (waveform *4*), although they are of longer duration, also produce equal amplitude spikes across the differentiating resistor, as shown in waveform *5*. These pulses, however, occur at a double rate, which will be discussed subsequently. Thus, the spikes generated during the vertical sync-pulse block prevent loss of horizontal synchronization during the vertical retrace intervals.

When the horizontal sync pulse is applied to the integrating circuit, it is of such a short duration that only a very tiny charge appears on the capacitor and is of no consequence (waveform *3*). During the vertical sync block, however, the pulses are of longer duration and place an appreciable charge on the capacitor. Furthermore, the charge of one pulse is not permitted to leak off the capacitor before it is again reinforced by the second pulse, and so on for the entire six pulses of the vertical sync block. Thus, each pulse adds a charge to the capacitor and the accumulative effect of each charge produces an ample

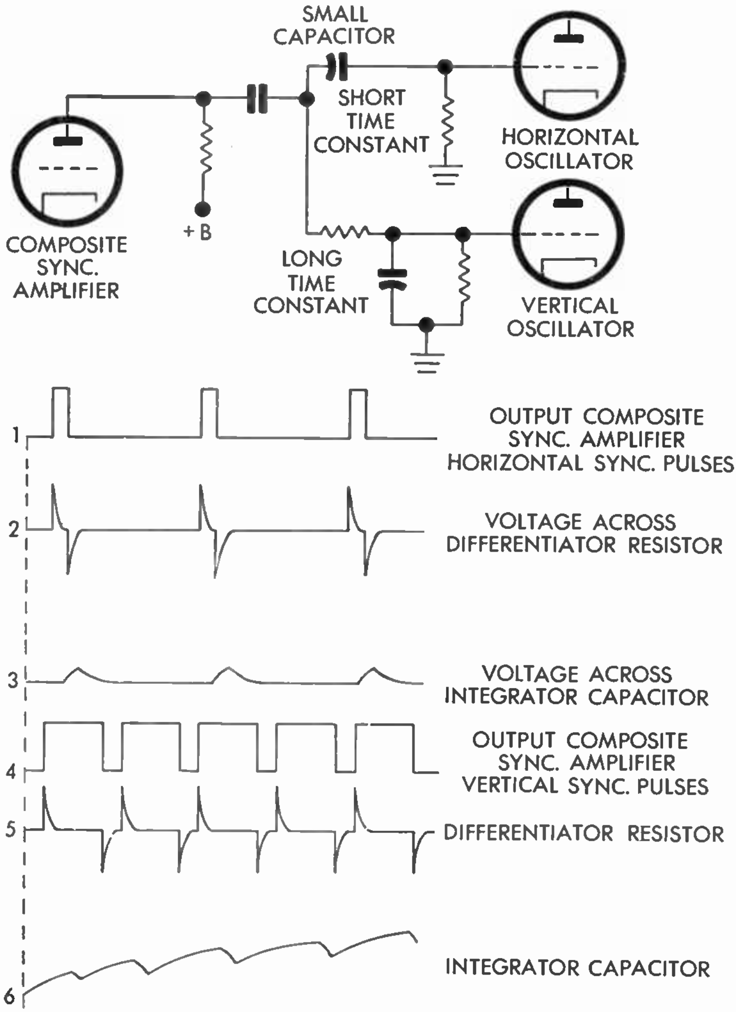


FIG. 143 Simple Inter-Sync Separation

voltage across the integrating capacitor (waveform 6). This integrated or step voltage is used to synchronize the vertical oscillator. Inasmuch as the six vertical sync pulses occur only once every 1/60 second, the vertical oscillator is fired at a repetition rate of 60 per second. The horizontal leading edges, however, occur once for each line of the picture, and therefore the horizontal is locked at a rate of 15,750.

85. *Horizontal Synchronization*

The horizontal sweep is synchronized by leading edges of the sync pulses which occur at intervals of 63½ microseconds. These leading edges occur not only during the scanning of the active horizontal lines, but also occur during the vertical retrace interval because it is necessary to maintain tight synchronism of the horizontal sweep even during the time that the vertical is retracing. This rigid requirement is needed to preserve the rigidity of the interlace system because it is necessary that the even-numbered lines fit precisely at the mid-point between the odd-numbered lines to have an interlaced high-resolution system. If we were for an instant to lose horizontal synchronization and then re-establish it after the start of each new field, the discontinuities present would cause unstable horizontal synchronism at the top of the picture, until the horizontal sweep would again fall in with the arriving horizontal sync pulses. When this condition does exist it appears as a displacement of lines to the right at top of picture and/or a loss of interlace part way down the picture. Often it occurs in receivers with poor low-frequency response, which causes differentiation of the vertical blanking pulse at some point and consequent tilt in the blanking flat top. Blanking tilt drops the horizontal sync pulses that follow behind the vertical sync interval below the horizontal firing level. Thus for some time during the vertical blanking period and top portion of picture, the horizontal is running unsynchronized.

Thus it is necessary even during the vertical-retrace interval to generate leading edges to maintain synchronism of the horizontal. Thus, instead of using a long, continuous, vertical sync pulse which would serve just as well as far as vertical integration and synchronization is concerned, the vertical sync block is broken up into a series of pulses (called *serrated vertical sync pulses*) which contain leading edges for maintenance of horizontal lock-in. Spacing between *alternate leading edges* of the vertical sync pulse and equalizing sync pulses is also 63½ microseconds.

Horizontal synchronization during active line intervals and during vertical-retrace period between fields and frame is shown in Fig. 144. If the top drawing represents the vertical retrace between fields and the lower one, retrace between frames, it is evident that the end of a field occurs at bottom right and the leading edge of the first equalizing pulse is one full line away from the last horizontal sync pulses. Thus, between fields, the horizontal locks in on the odd-numbered equalizing and vertical sync pulses (top drawing, Fig. 144). The horizontal synchronizes between frames on the even-numbered equalizing and vertical sync pulses. We also know that with interlace scanning at the end of a frame, the beam is retraced from the bottom center of the scanning raster; therefore the time interval between the last horizontal pulse at the end of the frame and the first equalizing is 31.75, or one-half the line interval. Therefore, if it were necessary to synchronize on the odd-numbered equalizing and vertical sync pulses as before. the horizontal oscillator would naturally shift phase for an instant. It is necessary to insert equalizing and vertical sync pulses halfway between the odd-numbered pulses mentioned previously. Consequently, the second equalizing-pulse leading edge is one line interval away from the last horizontal sync pulse at the end of a frame. There-

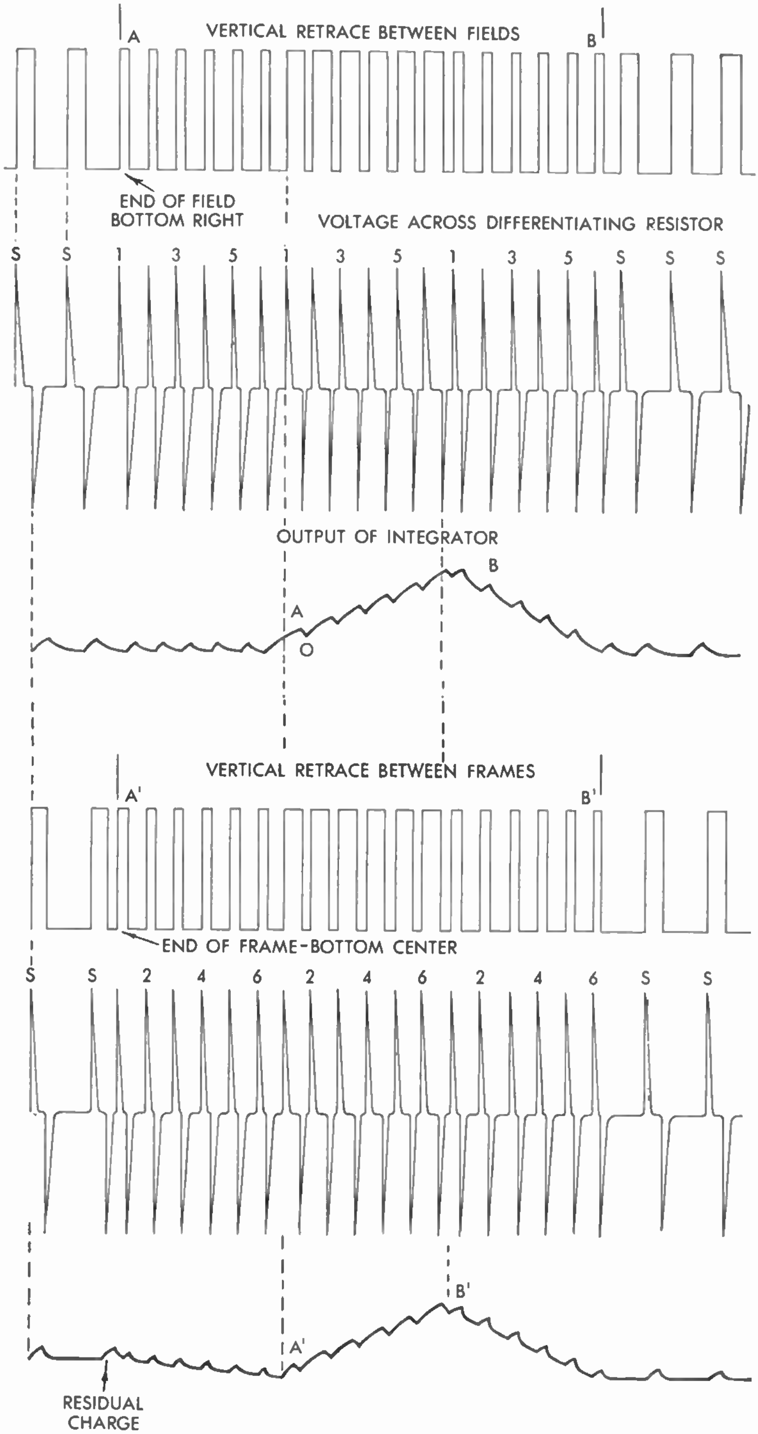


FIG. 144 Horizontal and Vertical Synchronization between Fields and Frames

fore, during the retrace interval between frames the horizontal is synchronized on the even-numbered equalizing and vertical sync pulses. Thus the double-line rate-equalizing and vertical sync pulses are necessary to maintain horizontal synchronism between fields and between frames.

It is also evident from Fig. 144 that the actual vertical-retrace interval between fields and between frames is absolutely identical, starting at the same instant for each field or frame and ending at the same instant. Thus the generating, formation, and stabilization of the vertical-retrace signals at the transmitter is simplified. In summation, note the following characteristics of Fig. 144:

1. During the vertical-retrace interval, the horizontal is synchronized by the leading edges of the equalizing and vertical sync pulses.

2. Between fields the horizontal is synchronized by the leading edges of the odd-numbered equalizing and vertical pulses. This occurs because the vertical retrace between fields has been assumed to begin at the bottom right, and there is a full horizontal-line interval between the last horizontal pulse and the first equalizing pulse.

3. Between individual frames of the picture the horizontal is synchronized by the leading edges of the even-numbered equalizing and vertical sync pulses. This occurs because the start of the vertical retrace between frames occurs at the bottom center of the scanning raster, and consequently there is only one-half line interval between the last horizontal sync pulse and the first equalizing pulse.

There is, however, a full line interval between the last horizontal sync pulse and the second equalizing pulse. The insertion of this mid-point equalizing and vertical sync pulse steps up the repetition rate of the equalizing and vertical sync pulses to 31,500.

It is evident that although the series of equalizing and vertical sync pulses occur only every 1/60 second at the instant they do occur, the six pulses of each group are occurring at the rate of 31,500 pulses per second. Horizontal synchronization between fields is not affected by the even-numbered pulses; and between frames it is not affected by the odd-numbered sync pulses. They do not adversely affect the control of the oscillator, because they occur at a time when they cannot affect the synchronization or timing of the horizontal oscillator. Thus the arrangement and duration of the vertical retrace and its component pulses are identical from the leading edge of the first equalizing pulse to the trailing edge of the last, in spite of the half-line interlace arrangement. Horizontal synchronism is not lost because of the double frequency rate of the equalizing and vertical pulses.

86. *Vertical Synchronization and Equalizing Pulses*

The vertical sweep is synchronized by the cumulative charge that is built up on the integrating capacitor during the 190-microsecond vertical sync block. Inasmuch as this block is transmitted once only every 1/60 second, the vertical is synchronized at 60 cycles per second. The vertical sync pulses are of long duration and very short spacing. They differ in that respect from the horizontal and equalizing pulses, which are short duration pulses and long interval spacing. Therefore, the vertical sync pulses during the vertical sync block build up an appreciable charge on the integrating capacitor (Fig. 144). Each pulse adds an appreciable charge to the capacitor, and although a small percentage is lost during the slot in the vertical, the actual spacing is so small that the accumulative charge builds up until it reaches an amplitude at the fifth or sixth pulse sufficient to trigger the vertical sweep circuit of the receiver.

It is imperative for the vertical to fire at the same instant for each cycle when an interlaced scanning system is used. If the vertical between frames fires a bit ahead of or after its normal time, compared to the vertical firing between fields, the interlace is lost. If there is any displacement of time whatsoever, the even-numbered lines will not fall exactly midway between the odd-numbered lines, causing pairing of lines and loss of picture resolution. The purpose of the equalizing pulses is to make absolutely certain that the vertical synchronizes at the same time between fields as it does between frames. We realize that between frames in our example the picture ends at the bottom center, and between fields at the bottom right. Consequently, at the time it ends between frames at the bottom center there may be present a small residual charge on the integrator capacitor from the previous horizontal sync pulse only a half-line preceding. Thus, at the instant the first vertical sync pulse arrives (assuming no equalizing pulses are present) there would be already present on the capacitor a small charge, and therefore the accumulative charge on the capacitor would build up the firing level a bit ahead of its normal firing cycle. This is evident when we consider that between fields the charge placed on the integrating capacitor by the last horizontal sync pulse had a full line to discharge before the arrival of the first vertical sync pulse, while between frames capacitor discharges for only a half-line interval. To prevent this unsymmetrical firing and to drain off any residual charge on the integrating capacitor, the equalizing pulses have been inserted into the composite television signal. These equalizing pulses are of very short duration and extremely long spacing, which permits a long discharge time and only a very small charge time. Thus, any residual charge present on the integrating capacitor at the start of the vertical retrace interval is slowly drained off during the equalizing pulse block. Therefore, between fields or between frames the charge on the integrating capacitor is the same at the instant the vertical sync pulse arrives; and the vertical oscillator will be fired at the same instant between fields as it is between frames. After the firing of the vertical oscillator the trailing set of equalizing pulses discharge the integrator capacitor back to zero or some fixed level again before the start of the horizontal scanning intervals. During the horizontal intervals the integrator charges from, and discharges to, the base level. The trailing equalizing pulses are needed to discharge the integrator more than it charges in order to bring the charge down to the base line.

It is advantageous to mention at this time that in our discussions we have considered the field retrace as starting from bottom right and frame retrace from bottom center. This is true when the vertical oscillator is triggered by an even-numbered vertical sync pulse (generally 4 or 6). However, it is quite possible to have field retrace start from bottom center and frame from bottom right if the vertical were to synchronize on one of the odd-numbered vertical sync pulses (3 or 5). In fact, retrace can begin from any point along the last line dependent on time of vertical firing—retrace start between fields and frames, however, is always a half-line apart.

In summation, notice the following characteristics of vertical synchronization:

1. During horizontal-line intervals, notice integrator has time to discharge to base level.
2. Vertical sync block made of long-duration short-interval-spacing pulses which build up accumulative charge to fire vertical oscillator.
3. Short-duration long-interval spacing of equalizing pulses drains any residual charge off capacitor before start of vertical sync block.
4. Vertical is triggered at same time between fields and frames. Amplitudes and durations between AB and A′B′ are identical.

87. *Basic Sync Control Systems*

In the basic television sync sweep system (Fig. 145), the differentiated horizontal sync and the integrated vertical sync are applied to the respective horizontal and vertical sweep oscillators which generate the sweep sawtooth voltages. These sync signals, available at the output of the inter-sync separator, are applied directly to the sawtooth sweep generators. The sharp horizontal sync pulse is used to synchronize the horizontal sweep oscillator and, therefore, controls the generation of the horizontal scanning lines. The vertical sync pulses control generation of the lower frequency vertical sawtooth which produces field scanning. If sharp pulse or random noises also ride along with the sync pulses the noise pulses are also capable of triggering the sweep oscillator if they are of sufficient amplitude and occur at the proper time. If noise pulses trigger the horizontal oscillator at a time before or after the normal horizontal sync pulse, the new scanning line is displaced with respect to the line previously scanned. Consequently, triggering of the sweep oscillator by the noise impulses causes a displacement of the horizontal line, and therefore a displacement and distortion of the reproduced image. The presence of noise in the vertical sync and sweep system causes the vertical oscillator to trigger ahead of or later than its normal cycle, and causes the entire frame to be displaced with respect to previous one. It is evident, therefore, that severe noise in the sync and sweep system of the television receiver causes loss of picture stability because of the displacement of the scanning line or displacement of the framing. The vertical sync system of the television receiver is not as adversely affected by noises as the horizontal because of the presence of the equalizing pulses and the fact that the vertical fires on an integrated wave. The vertical sync firing is accomplished by slowly building up a charge on the integrating network for the duration of the vertical sync pulse interval. Inasmuch as noises are short high-frequency bursts they do not appreciably charge the integrating capacitor and are effectively shunted. Thus the noise must be substantial before it affects vertical synchronization. Horizontal synchronization is readily disturbed by even a small amount of noise, especially sharp-impulse electrical noises such as ignition and sparking.

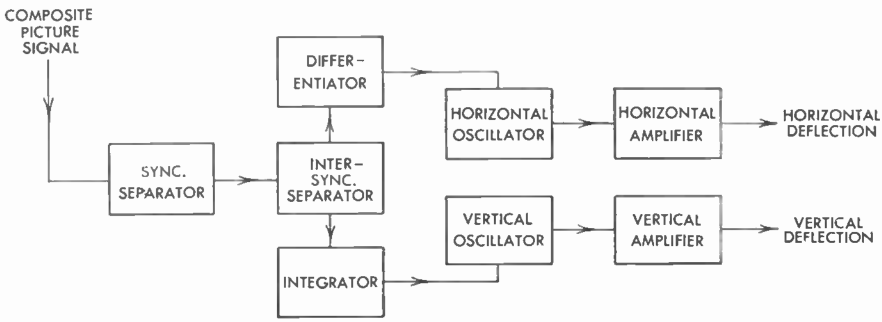


FIG. 145 Basic Sync and Deflection System

In the modern television receiver horizontal-sync control systems are employed to counteract the noise susceptibility of the horizontal synchronizing system. A control system makes the triggering of the horizontal oscillator independent of instantaneous noise impulses. It controls the horizontal oscillator because of the presence of the sync pulses as a repeating signal. Thus the horizontal is fired because sync pulses occur as a repeating signal, and not because they have sufficient amplitude to trigger the horizontal oscillator directly. Thus the triggering of the oscillator is more dependent on the repetition rate of the pulses and their phase position than on their amplitude. Random or impulse noises have little effect on horizontal synchronization because they are not cycled at the line rate as the horizontal sync pulses are.

Two basic systems are used to lock in the sawtooth frequency with the repetition rate of the horizontal sync pulses. In the first system (Fig. 146), a sine-wave oscillator is used to generate a sine wave of the line rate (15,750) which is compared with the incoming sync pulse in a discriminator. The phase discriminator produces a d-c component of output voltage whenever the phase or frequency of the sine-wave oscillator drifts away from the repetition rate and phase of the incoming sync pulses. This d-c component is applied to the grid of a reactance tube, the controlled reactive current of which keeps the horizontal oscillator at the same frequency and phase as the rate of arrival of the sync pulses. Thus the synchronization of the sine-wave oscillator is more dependent on the rate of arrival of the sync pulses than on their amplitude; consequently, bursts of noise which occur do not have any control over the sine-wave oscillator.

The frequency of the sine-wave oscillator is dependent on the average frequency of the sync pulses. Synchronization of the horizontal oscillator line by line has nothing to do with the amplitude of the sync pulses. The sine-wave output of the oscillator is clipped and differentiated and then applied to the

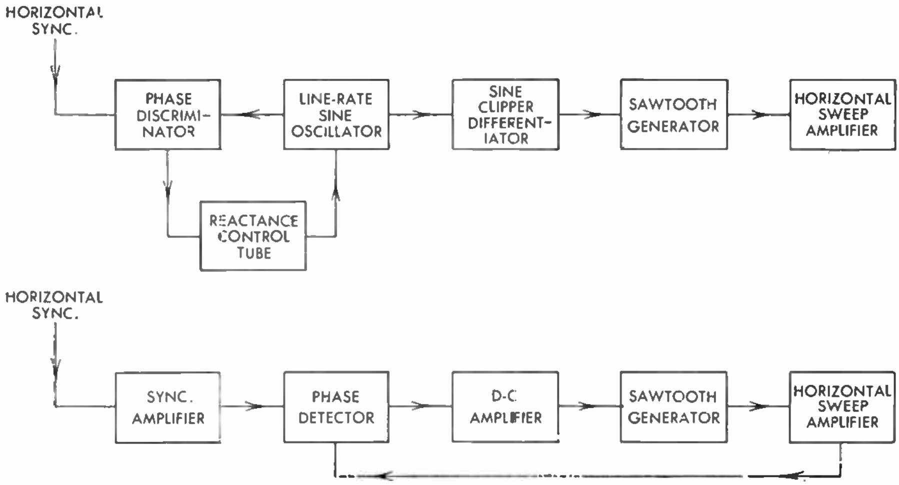


FIG. 146 Horizontal Sync Control Systems

sawtooth generator. In summation, the generation of the sawtooth is under control of the frequency of the sine-wave oscillator which, in turn, is dependent on the repetition rate of the arriving sync pulses and independent of the noise impulses that ride along with the received sync pulses.

A second method to make horizontal synchronization independent of instantaneous noise impulses is shown in the lower drawing. In this system a phase discriminator is also used which compares the frequency and phase of the incoming sync pulse with a sawtooth voltage fed back from the sweep amplifier. If the fed-back sawtooth voltage drifts in phase or frequency, caused by a change in frequency of the sawtooth generator, a d-c component of output is presented by the phase discriminator to a succeeding d-c amplifier which supplies d-c grid bias to the sawtooth generator. Inasmuch as the d-c bias on the sawtooth generator also determines its frequency, a change in the d-c component presented by the d-c amplifier causes the frequency of the generator to shift to a point where it will once more compare with the phase and frequency of the incoming sync pulse. Again, the frequency of the sawtooth generator is independent of bursts of noise, because these bursts of noise never reach the sawtooth generator. Instead, the frequency of the sawtooth generator is again dependent on the average frequency or repetition rate of the incoming sync pulses. If the sawtooth oscillator attempts to wander, the output of the discriminator will bring it back to the correct oscillating frequency. In summation, the frequency of the sawtooth oscillator is determined by its d-c component of grid bias, and if the frequency and phase of the sawtooth wave do not compare with the incoming sync pulse, a d-c component of output from the phase discriminator changes the d-c bias of the sawtooth generator and re-establishes the frequency at the proper point.

88. *Commercial Horizontal Sync Control Systems*

The basic RCA sync-lock circuit is shown in Fig. 147. In this circuit a Hartley oscillator is used to generate a sine wave of approximately 15,750 cycles per second, the screen of the pentode serving as plate of oscillator. A high-amplitude oscillator grid signal produces a clipped sine wave at the plate output. The frequency of the oscillator is determined by the constants of its tank circuit and presence of the reactance tube across it, which pulls a reactive or quadrature current through the oscillator tank circuit. The current amplitude varies with the grid bias on the reactance tube. Thus the frequency of the sine-wave oscillator over a limited range is determined by the grid bias on the reactance tube (for complete discussion of the reactance tube, refer to

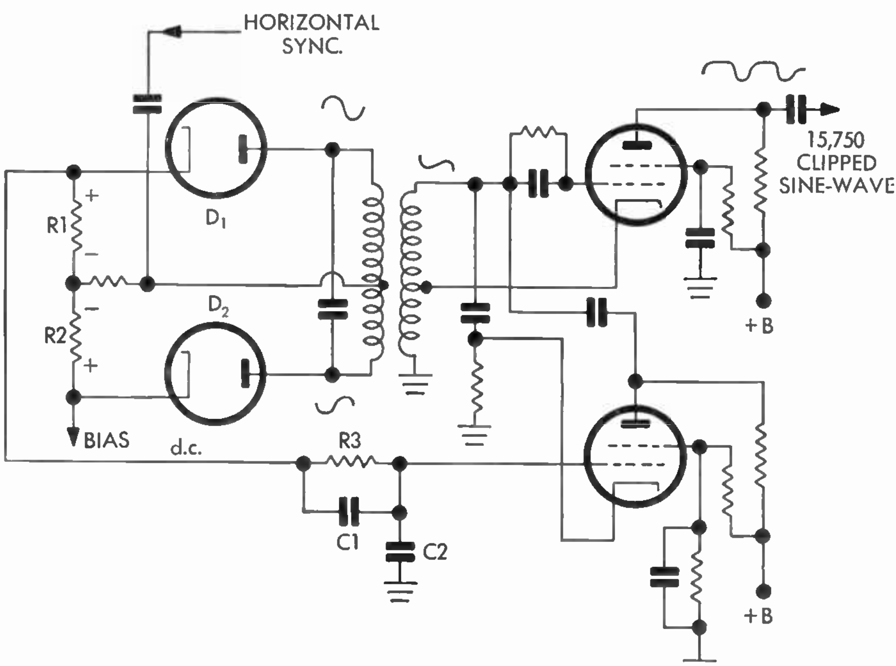


Fig. 147 RCA Sync-Lock System

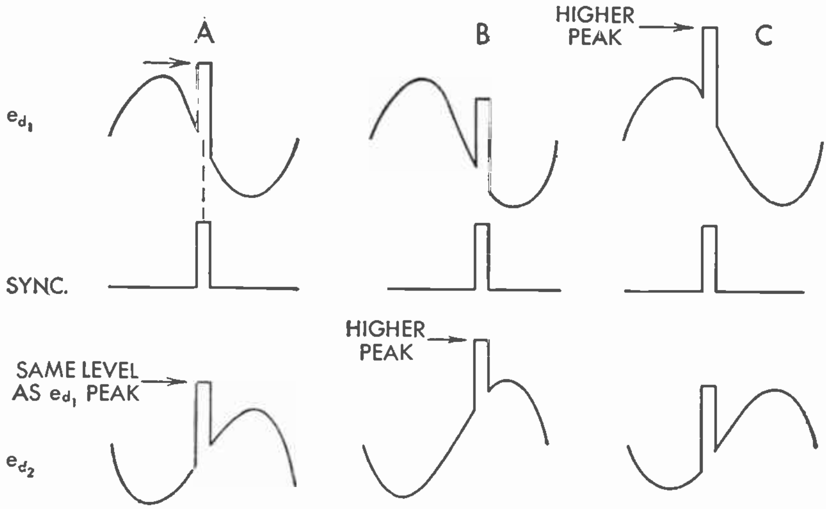


FIG. 148 Sync-Lock Waveforms

Chap. 9). The bias on the grid of the reactance tube is, in part, a d-c component of bias from the discriminator output circuit which varies with the phase relation between the inductively coupled sine waves and the arriving sync pulses.

The operation of the phase discriminator is explained with reference to Fig. 148. Out-of-phase sine waves are fed to the plates of the discriminator diode and sync pulses are applied positive-going to both plates of the diode. Consequently, the sync pulse adds algebraically to the sine wave and causes a peak diode current to flow in accordance with the peak difference of potential existing between the plate and cathode of an individual diode. When the proper phase and frequency relations exist between sine-wave oscillator frequency and frequency of the incoming sync pulse, the sync pulse occurs at the zero point of sine wave applied to each diode. Consequently, the peak diode current drawn by each diode will be equal, and inasmuch as the diodes pull current through the load resistors *R1* and *R2* in opposite directions, the output voltage will be zero and no bias is contributed by the discriminator to the reactance tube. However, if the frequency of the sine-wave oscillator were to drift and the phase relation between the incoming sync pulse and the sine wave are as shown in drawing B of Fig. 148, the sync pulse would add to the sine wave applied to the bottom diode on its positive alternation and would add to the sine wave applied to the top diode on its negative alternation. Consequently, the peak diode current drawn by the bottom diode would be greater than the peak diode current of the top diode. If this were to occur over a period of time a d-c component of negative voltage would appear at the output of the discriminator and, therefore, across capacitor *C2* and on the grid of the reactance tube.

The reactance tube in turn would draw a lower plate current and would shift the frequency of the sine-wave oscillator to a point at which it would once more be coincident with the sync pulse at the zero level of the sine wave. Likewise, if the drift of phase was such that the sync pulse would appear at the instant shown in drawing C, the top diode would draw the peak diode current and, consequently, the d-c component of discriminator output would be positive and a positive voltage would be applied to the grid of the reactance tube. This, in turn, would shift the frequency of the sine-wave oscillator in the opposite direction, again correcting for the original drift in frequency and once more causing the sync pulse to occur at the zero point of the sine-wave cycle. Thus any drift of the sine-wave oscillator (slow in comparison to the rate of arrival of the sync pulses) causes a d-c output to appear at the discriminator, which, applied to the reactance tube, causes the frequency of the sine-wave oscillator to return to the proper point.

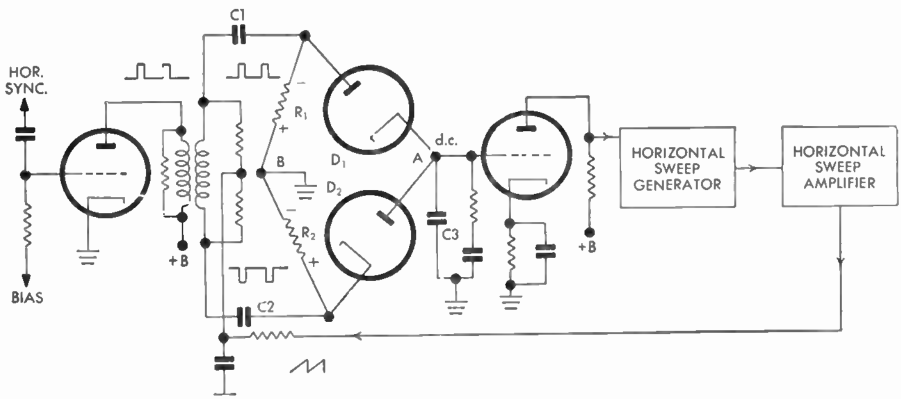


FIG. 149 GE Sync Control System

The presence of noise along with the incoming sync pulse will have little effect on the operation of the discriminator, because the rate of arrival of the noise impulses over an average period of time would not be regular enough to cause operation of the phase discriminator. The regularity of arrival of sync pulses and their fixed position with respect to a certain section of the sine wave over a lengthy time interval causes operation of the discriminator and consequent correction for any frequency drift of the oscillator.

It is necessary that the system respond reasonably fast to a change in phase of the incoming sync pulse from the transmitter. The speed with which the system operates is determined by the time constant of *C1*, *C2*, and *R3*. If the time constant is long the speed with which the circuit operates is slow and the system is insensitive to rapid changes. The ratio of *C1* and *C2* also attenuates rapid changes in level such as are presented by the vertical synchronizing pulses or bursts of interference. If the transmitted synchronizing pulses themselves vary in phase the time constant and capacity ratio must be made lower in order to respond quickly to the transmitted variation. If the time constant is lowered it does make the circuit a bit more susceptible to noise and interference.

A second automatic sync-control system as used by General Electric is shown in Fig. 149. In this system, the incoming sync pulse is compared to a sawtooth waveform fed back from the horizontal sawtooth amplifier circuit. The d-c component of output of the phase detector is applied to a d-c amplifier in the grid circuit of which any variations are filtered out. A d-c bias component is applied to the sawtooth generator from this amplifier. If there is any

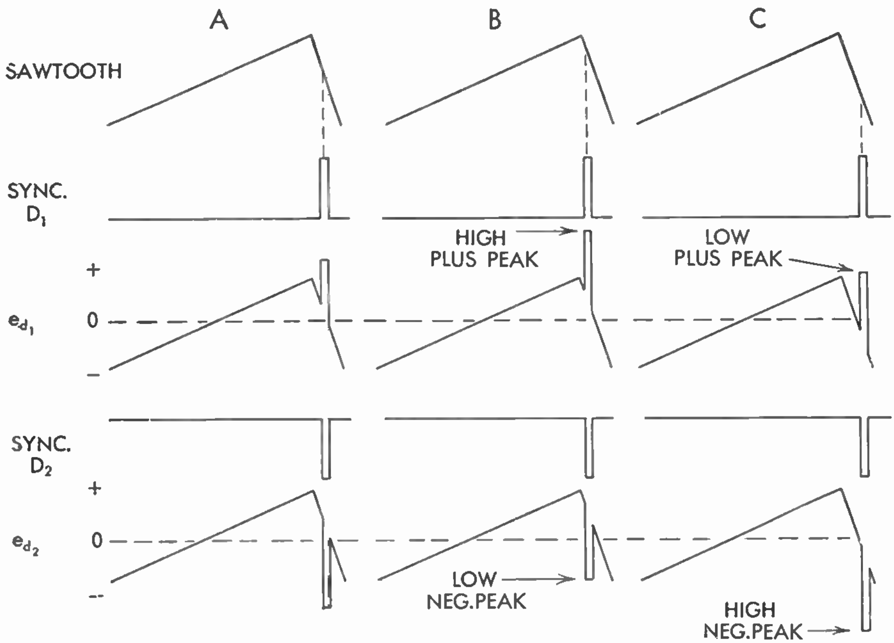


FIG. 150 Sync Control Waveforms

variation in the average phase or frequency of the sawtooth wave with respect to the incoming sync pulses, this d-c component of bias on the sawtooth generator will change, causing the frequency to shift a corresponding amount. The sync pulse is applied to the phase detector with the proper polarity to make both diodes conduct. The polarity of the fed-back sawtooth can be positive or negative, and sawtooth appears at same polarity on the top plate and lower cathode. If the phase and frequency of the incoming sync and the sawtooth wave compare favorably the sync pulse adds to the sawtooth at the center of the sawtooth retrace. Consequently, the peak current drawn by each diode is the same (refer to the waveform of Fig. 150, drawing A). The diodes are peak rectifiers and draw current only during the peak of sync tip (action is similar to a diode-type d-c restorer). If the peak current drawn by each diode is the same, there will be no d-c component in output (both ends of the bridge unbalanced so far as the output is concerned). During the interval between sync tips the diodes are held cut off by the long time constant *C1, R1* and *C2, R2*. The average charge on both capacitors is identical when the sync pulses are coincident with the zero point of the sawtooth retrace.

When the phase of the sawtooth is displaced with respect to the arrival of the sync pulses, the sync pulses occur on the positive side of zero for one sawtooth and on the negative side of zero for the other sawtooth. Consequently (Fig. 150, drawing B) the peak positive voltage applied to one diode during a sync tip will exceed the peak negative voltage applied to the other. Therefore, the average charge of *C1* will differ from the average charge on *C2*, and a d-c component will appear between A and B which is actually across capacitor *C3*. If the phase displacement is in the opposite direction the other capacitor will receive the greater negative charge. Inasmuch as one capacitor retains a positive charge and the other a negative charge, the polarity of the d-c component across *C3* will reverse in accordance with the direction of the phase displacement.

The charge on the filter network is the bias on the grid of the d-c amplifier. A change in this d-c component will shift the frequency of the sawtooth oscillator which is direct-coupled to the plate of the d-c amplifier and receives its bias from that amplifier. The time constant of the filter is long and will not respond to fast changes in voltage. Although a burst of noise may at times be great enough to drive the diode into conduction between the sync pulse interval, the change occurs so quickly that the filter does not follow because of its long time constant. Thus the filter very effectively removes noise synchronization and changes its voltage only over a long period of time when the average frequency of the sawtooth and the incoming sync pulse differ. Again synchronization is set by the average frequency of the incoming sync pulse and is little affected by amplitude of noise interference. It is again important that the filter networks have the proper time constant. If the time constant is too long, the system cannot follow a shift in phase of the transmitted sync pulses and will produce displaced or scalloped lines down the screen. A too short time constant permits the system to follow rather sudden and short shifts in d-c level, vertical-sync interval often causing the entire pattern to vibrate horizontally.

A less elaborate automatic sync control system by GE (Fig. 150a) uses a locked blocking oscillator as a sweep generator. Frequency of this oscillator is under control of a d-c potential applied to its control grid, d-c level being a function of the phase of the incoming sync as compared to a sweep waveform fed back from the horizontal sweep output. Actually the waveform fed back to the grid of tube *V1* is a modified sawtooth formed by combining the sawtooth sweep output of the blocking oscillator *V2* with a negative transient from the horizontal output transformer. This resultant waveform is combined with the arriving sync pulses at the grid of *V1*.

Only when the sync pulse is present along with the modified sawtooth is there sufficient peak signal amplitude to overcome the normally negative bias (contributed by the blocking oscillator grid circuit through a 3.3 megohm resistor) on the grid of $V1$, permitting it to conduct. The length of time it conducts and therefore the level of the charge placed on the cathode capacitors

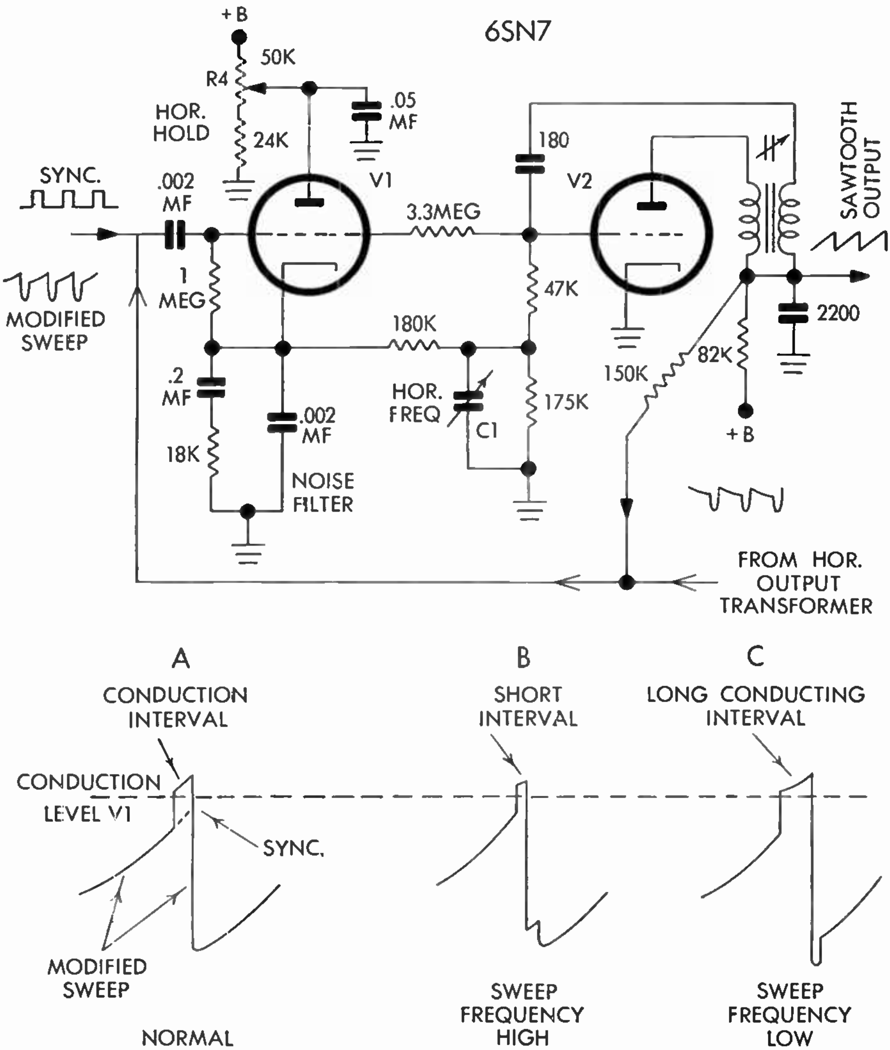


FIG. 150a Another GE Sync System

depends on the time position of the sync pulse with respect to the modified waveform. The charge on these capacitors is divided across the 180K and 175K resistors, and the voltage across latter is also present on the grid of the blocking oscillator and any variation in this potential causes a shift in the oscillator frequency. Again the charge is a function of the rate of arrival of the sync pulses and the circuit is relatively insensitive to sporadic noise bursts. Noise filter permits only a slow change in charge such as that encountered when the arriving sync pulses slowly drift in phase or blocking oscillator wanders away from correct frequency or phase.

What happens at the grid of *V1* is shown in the waveforms. In waveform *A* the sync pulse is above cutoff the proper length of time to cause a charge across the cathode capacitors of the proper voltage to hold the blocking oscillator on the correct frequency. In waveform *B* sweep frequency is a bit high and a part of the sync pulse drops below cutoff with the sharp decline of the modified waveform. Thus the tube does not conduct for as long an interval, the charge on cathode capacitors drops, *V2* grid voltage decreases, and frequency of blocking oscillator decreases to the same frequency as the arriving sync pulses. In the case of waveform *C* sweep frequency is low and sync pulse is above cutoff for a lengthy interval, causing an increase in the voltage across *V2* grid resistor and a correcting increase in the frequency of the blocking oscillator.

The free-running frequency of the blocking oscillator is adjusted with the blocking oscillator transformer control and the frequency control *C1* which regulates the time constant in the grid circuit. The hold control is in the plate circuit of *V1* and determines the level of plate current flow for *V1* and in turn the cathode charge and frequency of the blocking oscillator.

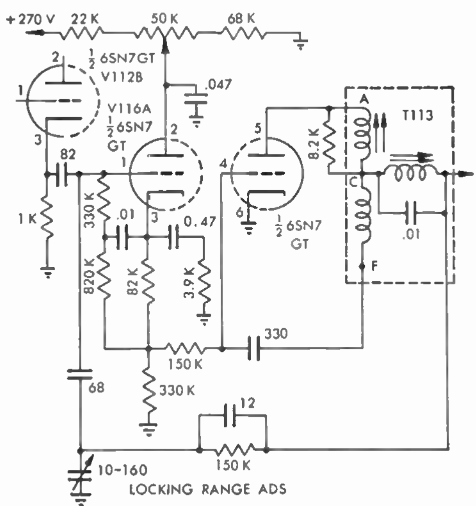


FIG. 150b RCA Synchroguide

This method of horizontal control is often referred to as a pulse-width system because it is the duration of the pulse above the conduction level that determines the charge and the d-c voltage that biases the blocking-tube oscillator. Just how long the pulse width is depends on its phase relation with respect to the apex of the sawtooth wave. The pulse duration becomes longer when the pulse comes ahead of the top of the sawtooth wave, and duration becomes much shorter when the pulse occurs after the apex of the sawtooth. Normal operation occurs when approximately one-half of the normal sync-pulse duration occurs atop the conduction level. The actual phase comparison is made between the pulse and the apex of the sawtooth, while the very sharp spike derived from the horizontal output circuit insures a fast drop-off of the pulse after the sawtooth apex. This sharp downward slope immediately following the peak preserves stability and precision in the sync-control system.

In a newer synchroguide circuit, Fig. 150b, only the sawtooth and pulse are used in the pulse-width control system, the sharp spike having been abandoned. However, to produce a fast drop-off again after the apex of the sawtooth and to obtain adequate stability the sawtooth portion is properly modified. An approximate line-rate resonant circuit is placed in series with the horizontal oscillator plate supply to add stability to the blocking oscillator and to sharpen the drop-off time. This tuned circuit sharpens the approach of the waveform of the blocking oscillator toward the conduction level, and consequently, a far greater change in conditions of operation must occur before the triggering of the oscillator can be affected harmfully. The introduction of the sine-wave is apparent when observing the waveform at point *C*, Figs. 150b and 150c. The waveform as it actually appears at the grid of the control tube shows the sawtooth formation and the sync-pulse interval as well as the fast drop-off at the conclusion of the comparison period.

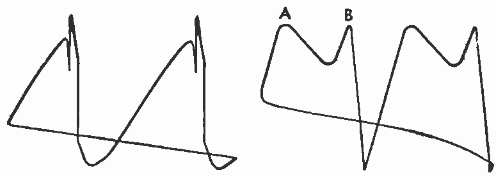


FIG. 150c Synchroguide Waveforms

A core in the oscillator transformer permits a coarse adjustment in frequency, while the fine adjustment is made by controlling the d-c current flow in the control tube with the plate-circuit potentiometer. Proper stability is obtained with the correct adjustment of the core in the stabilizing resonant circuit. Lock-in range is controlled by properly regulating the peak amplitude of the composite waveform at the grid of the control tube. In this circuit, as in the previous one, this control tube is held at cut-off with the grid bias of the oscillator tube except during the sync-pulse interval. The time constant in the cathode circuit of the control tube is chosen critically in order to prevent hunting and to permit the oscillator to follow the line-to-line changes of the incoming sync pulses with precision. Also included is a long integrating time constant to filter low-frequency disturbances properly.

Another a-g-c sync system, Fig. 150d, uses a triode phase-detector, a reactance control tube, and a sine-wave oscillator. The phase-detector triode acts as a dual diode with the grid and cathode acting as one diode and the plate and cathode as the second diode. A horizontal sawtooth is taken from the sawtooth-forming circuit and appears across resistors *R353* and *R352* with the same polarity. The sync pulse, however, is applied to the junction of these two resistors and develops a negative pulse across *R352* and a positive one across the top resistor. The relative position of the sync pulses on the sawtooth retrace (above or below the sawtooth's center line) determine the dominating current flow, whether it be maximum current down through resistor *R353* or maximum up through resistor *R352*. Dominant current flow develops the appropriate charge on capacitor *C375* via the large-value resistor *R355*.

Consequently, the grid bias on the horizontal reactance tube is a function of the phase relationship between incoming sync and the fed-back sawtooth wave, as in the usual phase-detector type of a-f-c control.

A Hartley oscillator operates at approximately 15,750 cycles, and its resonant circuit consists of an inductor and a shunt capacitor *C361* plus the network consisting of capacitors *C359* and *C353* and resistors *R362* and *R363*. The frequency adjustment is resistor *R363* which changes the influence or effective reactance contributed by capacitors *C359* and *C353*. Any change of

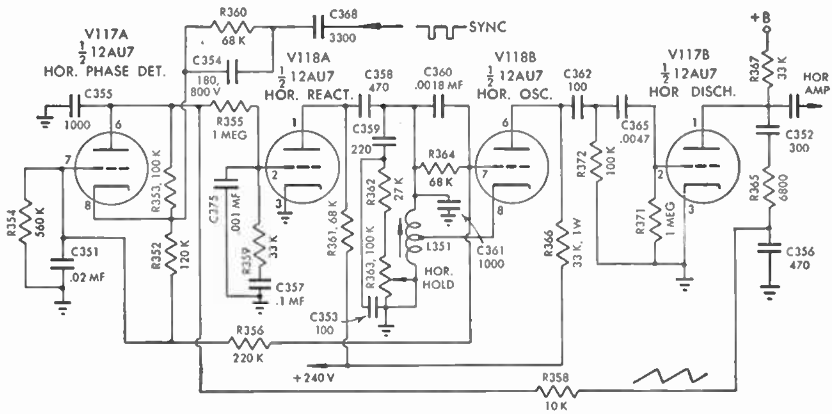


FIG. 150d Phase-Detector AFC (GE)

effective reactance in this network changes the resonant frequency of the tank circuit over a limited range. The moderately long time constant contributed by capacitor *C360* and resistor *R364* maintains the horizontal oscillator at cut-off for a greater portion of the sine-wave cycle developed in the resonant circuit. Consequently, plate-current flow is in bursts, and an appropriate plate waveform that can be used in the generation of the sawtooth waveform is developed.

Another *R-C* combination that influences the resonant frequency of the tuned circuit is that of capacitor *C358* and the plate resistance of the reactance-control tube. Actually, the capacitor represents the reactive element, and its effectiveness is influenced by the value of the plate resistance contributed by the control tube. Thus the reactance operation is not quite the same as that associated with the usual reactance circuit. Here the reactance tube itself does not contribute any reactive component but merely influences the reactive effect contributed by the small capacitor. The plate resistance of the tube, however, is a function of the d-c voltage (phase-detector control voltage) present on its grid; as this component changes the plate resistance of the triode, the reactive component contributed by capacitor *C358*, and, in turn, the resonant frequency of the oscillator tuned circuit also change. Therefore, as in most a-f-c systems, the relative phase relation between the sync pulse and the fed-back sawtooth influences the d-c voltage applied to the reactance tube and eventually brings the frequency and phase of the oscillator to center sync pulse on the retrace portion of the sawtooth wave that has been derived from the oscillator signal.

The positive portion of the waveform developed at the plate of the oscillator triggers the discharge tube and thereby initiates the retrace portion of the horizontal deflection sawtooth. The generation of the deflection waveform will be treated in detail in Chapter 8. The integrator and low-pass input network at the grid of the reactance control tube prevents noise impulses from affecting the bias circuit. A portion of the negative bias developed by the oscillator tube is applied through isolating resistor *R356* to bias the elements of the phase detector properly and thus obtain a balanced current flow for the acting diode sections of the tube.

89. *Sync and Inter-Sync Systems*

Commercial sync systems differ widely in circuits and methods. The basic function of all types is, first, removal of the sync and then utilization of the horizontal and vertical components of that sync. Five representative systems will be presented in this section. An RCA sync system is shown in Fig. 151, and consists of six tubes. A negative-going composite signal is taken off the input circuit to the picture tube and supplied to the grid of the 6SK7 sync amplifier. The 6SK7 tube is a remote cutoff tube and is used to prevent the negative-going sync tip from being cut off for wide variations of signal strength. The composite sync output of the last video amplifier varies with signal strength and setting of the contrast control. A high-amplitude composite signal is applied to the grid of the succeeding tube, which is the sync separator and uses a combination of signal bias and external bias to make certain that the cutoff occurs approximately at the blanking level of the composite signal. The negative-going sync in the output of the separator is applied to a second sync amplifier and clipper which levels off both the positive and negative portions of the composite sync signal.

The output of the final sync amplifier divides into two paths—one path is to the vertical integrating circuit and second path to the discriminator of the horizontal sync control system. The integrating circuit consists of a triple-section *R-C* combination which permits build-up of a reasonably high amplitude-integrated voltage and also is effective in shunting all high-frequency components of the signal and noise interference to ground. Leading edges of the composite sync signal are coupled through a small capacitor to the center point of the discriminator transformer. The sine wave from the horizontal oscillator is inductively coupled to the discriminator through the same transformer. Actually, the discriminator transformer is tuned slightly off the resonant frequency of the horizontal oscillator to prevent severe loading and instability. Inasmuch as it is a phase discriminator and not a frequency discriminator, the fact that the secondary is tuned off resonance is unimportant.

The phase-discriminator output, when there is a phase displacement between the sine wave and the sync pulse, adds to the minus 2-volt external bias and supplies grid bias to the reactance tube. This control or reactance tube causes a reactive plate current to flow in the oscillator tank circuits. The amplitude of this current flow determines over a limited range the frequency of the Hartley oscillator. A manual adjustment of the grid resistance in the grid

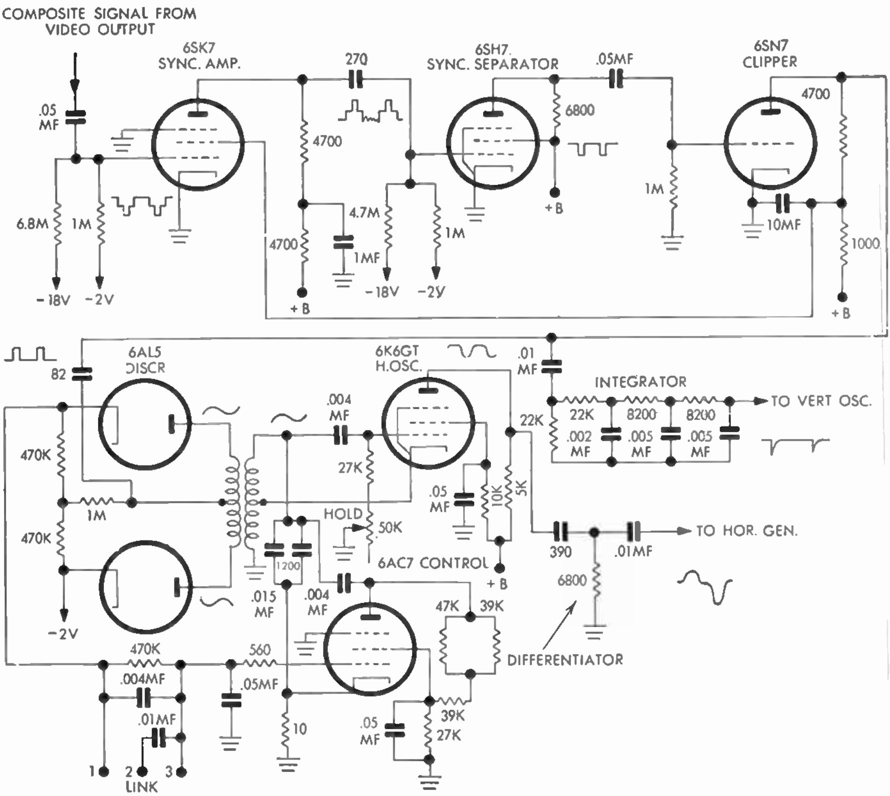


FIG. 151 RCA Sync and Inter-Sync System

circuit of the oscillator also permits a change in frequency over a limited range. Actually, the screen grid of the 6K6GT oscillator serves as the plate for the oscillating section. Sufficient grid variation is obtained to produce an actual clipped wave at the true plate output of the 6K6. This squared-off wave is then applied to a differentiating circuit which, in turn, excites the sawtooth generator of the horizontal sweep system.

A schematic of a General Electric sync system with horizontal sync control is shown in Fig. 152. In this receiver negative-going composite signal is again taken off the video-output tube and is coupled through a capacitor and isolating resistor (isolate capacity of the sync separator stage from the high-frequency video-output circuit) to the grid of a triode sync separator. The separator has its cathode grounded and is supplied with a low supply voltage to keep the composite sync output constant in amplitude for reasonable variations of peak signal amplitude at the plate of the video-output tube. The negative-going composite sync at the output of the separator is fed into two paths, one going through a large-series resistor and shunt-capacitor integrating circuit which develops an integrated wave on the grid of the vertical sync

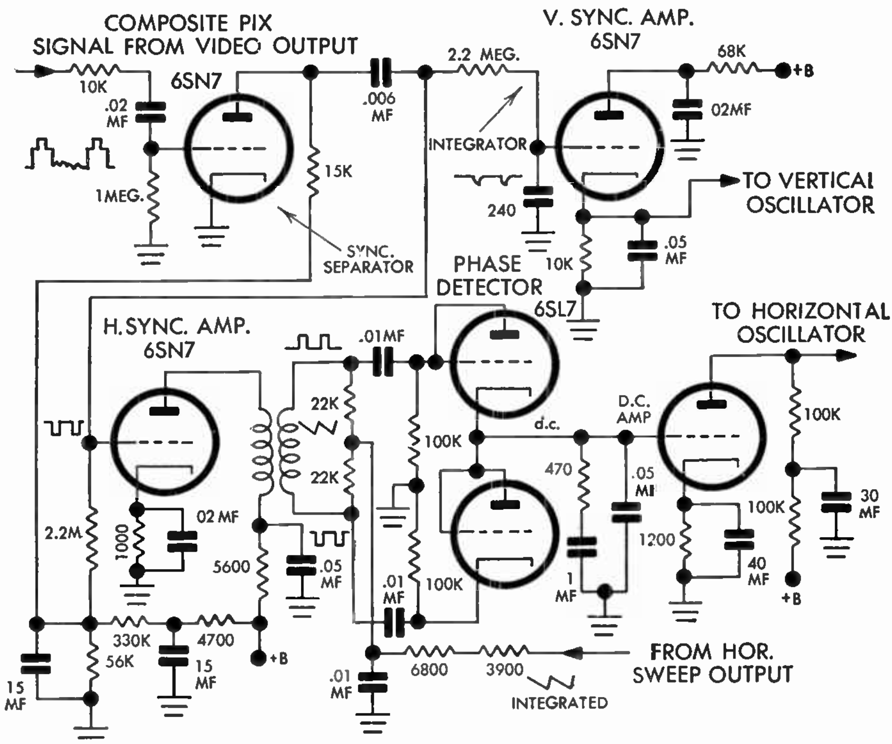


FIG. 152 GE Sync and Inter-Sync System

amplifier. An integrated negative pulse is taken off the cathode and applied to the first section grid of the vertical multivibrator. The three shunt capacitors in the vertical sync amplifier all help to integrate the wave and to shunt off high-frequency components and noise interference.

Horizontal sync pulses are amplified by the sync amplifier and produce positive pulses in the plate circuit. The positive pulses are coupled through the transformer to the phase discriminator circuits and drive the plates of the top section positive and cathode of the bottom section negative. To the midpoint of a resistor divider, sawtooth voltage from the secondary of the horizontal-output transformer is applied with the proper polarity to supply a sawtooth to the top diode and a same polarity sawtooth to cathode of the lower diode. When phase displacement exists between the horizontal sync pulse and the mid-point of the sawtooth retrace, a d-c component of voltage is developed across the filter network. Actually the filter capacitor consists of a 1-microfarad capacitor and a 0.05-microfarad capacitor. The presence of both capacitors, particularly the large one, ensures filtering of extremely low frequencies and prevents any instantaneous shift in phase from affecting the operation of the saw-tooth generator. The second capacitor is necessary to

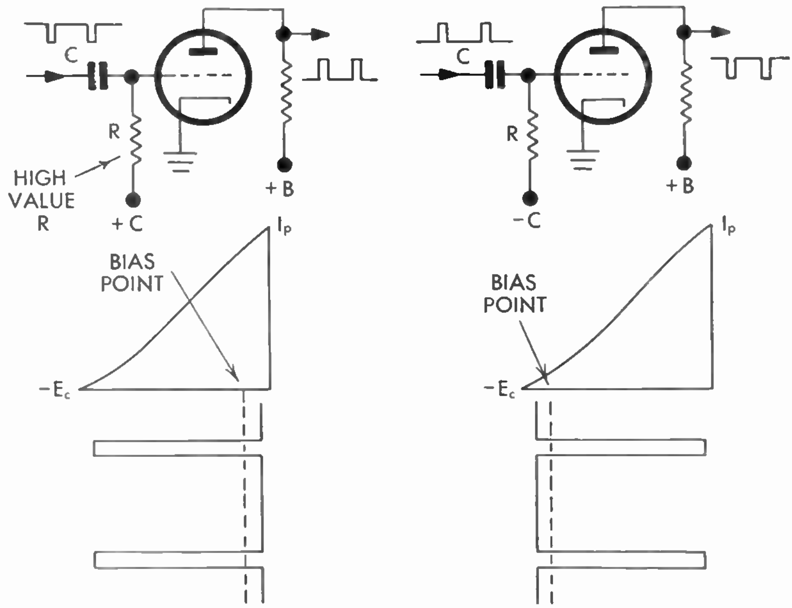


FIG. 153 Biasing a Pulse Stage

present a low reactance at higher frequencies where the large capacitor may become inductive or system break into self oscillation. Very careful filtering therefore prevents instantaneous changes from affecting synchronization, and only the average shift in phase over a lengthy period of time causes the d-c component of charge on the capacitor to vary. The capacitor charge determines the grid bias of the d-c amplifier, and inasmuch as its plate is direct-coupled to the grid of the horizontal sawtooth generator, it also determines the grid bias on this tube, and therefore its frequency over a limited range.

It is interesting to note at this point that positive grid bias is applied to grid of vertical sync amplifier and also to the grid of horizontal sync amplifier. This expedient keeps the grid near zero bias level when no signal is applied. Inasmuch as the applied signal is negative-going in polarity the tube is never driven positive by the applied signal. Stage is capable of receiving a high-amplitude signal, because the signal can swing over the range from zero bias point to cutoff if it swings on one side of the bias point, as the negative-going composite sync does. The positive bias also counteracts the high negative self-bias which occurs when a 10,000-ohm cathode resistor is used in the cathode circuit of the vertical sync amplifier.

This system of biasing grids at one end or the other of the transfer characteristic is used extensively in pulse circuits, or any circuit, for that matter, in which a single-polarity signal is to be passed. It permits use of the full sweep of grid signal, still keeping the signal on the reasonably linear portion of the tube characteristic. This factor is demonstrated in Fig. 153 for a negative-going sync pulse applied to a stage which is biased near to zero, and for a positive pulse applied to a stage which is biased near cutoff. This expedient permits application of a signal twice as high in amplitude as could be applied if the tube were biased at the mid-point of its linear characteristic as is done for sinusoidal amplification. Base line of pulse signal is also kept level by slight overdrive.

90. *Commercial Sync Systems*

In summary, the responsibilities of the sync system of the television receiver are separation of the sync from the composite video signal (often referred to as sync stripping), amplification of sync components to a suitable level for synchronization, segregation of the horizontal and vertical components of synchronization, and, finally, application of individual sync signals to the proper sweep-generating circuits. In the sync-amplification process, circuits are incorporated in order to keep the sync pulses at a constant amplitude, regardless of variations in incoming signal strength, presence of noise in the sync signal, or variations in picture content and brightness. A level sync line assures positive and continuous synchronization of the sweep oscillators. Another responsibility of the sync amplifier-separator system is to clip off noise impulses, thus preventing erroneous synchronization of sweep-generating circuits.

In a typical RCA sync-amplifier system, Fig. 154, there are separate horizontal and vertical sync-separator and sync-amplifier channels. This dual channel-arrangement permits choice of optimum time constant for favoring the horizontal or vertical sync components that pass through. At the same time, the horizontal and vertical components are separated from each other and do not interact on each other adversely. Sync separation in both horizontal and vertical channels is accomplished by developing positive cathode bias and permitting the tubes to conduct only during the positive sync-pulse intervals. Positive sync pulses are obtained from the plate circuit of the video amplifier and are direct-coupled in order to have part of the grid bias of the sync-separator tubes contributed by the video-amplifier plate circuit. Positive grid bias and positive cathode bias are of proper value to supply just sufficient negative grid bias to permit conduction above the blanking level but not during the picture-information intervals. All portions of the composite video signal below the blanking level are below cut-off bias on the two separator stages.

The vertical sync pulses are applied through resistor *R1* to the grid of tube V113A. Resistor *R1* and the input capacity to the stage act as an integrating circuit, removing horizontal sync components. A high cathode bias is developed by a voltage-divider to plus-B supply and is held constant by capacitor *C1*. The very long cathode-circuit time constant and the large-value plate-load resistance filters thoroughly the higher frequency horizontal components; only the low-frequency vertical information, which is developed at the plate circuit of the separator, is applied as a negative integrated pulse to the grid of the vertical sync amplifier. The vertical sync amplifier is biased near saturation

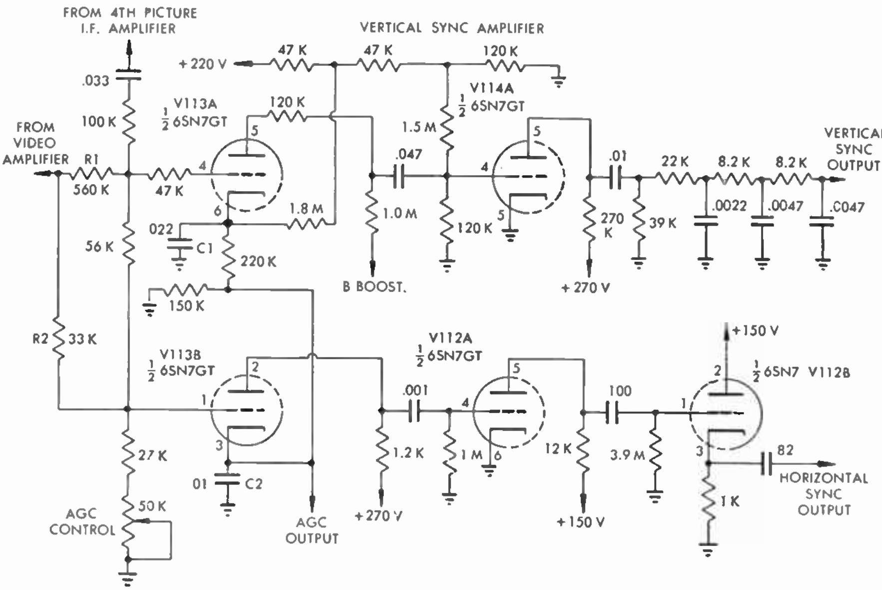


FIG. 154 RCA Sync-Separator Circuits

because the cathode is grounded, and the grid is returned to a positive supply voltage source through a large-value grid resistor. Consequently, the noises present during the intervals between vertical sync pulses are leveled off by plate-current saturation, and at the same time, the vertical sync pulse that swings negatively away from the saturation bias-point is strong enough in amplitude to reach cut-off. Therefore, noise pulses present on the vertical sync pulse are also removed by cut-off action. A positive vertical sync pulse is developed at the plate of the vertical sync amplifier and is applied to a triple-section integrator circuit before being applied to the vertical sawtooth-forming generator. The triple integrator circuit helps to attain positive interlace and, at the same time, through the series of r-c filters effectively shunts to ground any higher frequency noise impulses or horizontal sync-component signals that might pass through to the vertical oscillator and cause erroneous or jittery synchronization. The multiple-section integrator system makes certain that the start of the integrator voltage build-up at the sawtooth generator begins at a definite voltage amplitude between fields and between frames, assuring absolutely uniform spacing between vertical trigger times and preventing any pairing of lines that would be caused by faulty triggering of the vertical oscillator. It is also far more difficult for higher frequency noise pulses to penetrate and cause premature triggering of the vertical generator.

Horizontal sync pulses, through resistor *R2*, are applied to the grid of the lower sync separator. Again the separator is biased beyond cut-off because of positive potential developed at the cathode and positive voltage supplied to the grid from the video amplifier. Thus the separator conducts only during the positive sync tips of the horizontal interval. The cathode time constant (particularly capacitor *C2*) is sufficiently short to allow capacitor *C2* to discharge some during the long time interval between arriving sync pulses. Thus the tube always conducts with the arrival of the horizontal sync pulses. However, the time constant is long enough to enable the capacitor to remain charged and to prevent substantial conduction during the vertical sync pulse interval (longer duration pulses and shorter spacing); therefore, a vertical pulse is not present in the plate circuit of the horizontal sync separator. Negative sync pulses are applied to the grid of the horizontal sync amplifier, which is also biased near saturation by virtue of the cathode return to ground and the long time-constant grid circuit that is biased by the applied signal. Only during the negative sync pulse, itself, is there appreciable change in plate current, the negative sync pulse driving the tube toward cut-off and developing a positive sync pulse at the plate circuit. Again this saturation bias system removes noise impulses and video information that might exist between the horizontal sync pulses.

The swing of the negative sync pulse at the grid of the first horizontal sync amplifier also sweeps below cut-off and removes impulse noises that exist on the flat-top of the horizontal sync pulse, itself. The positive sync pulse in the plate circuit is applied to a second sync-pulse clipper stage to clean the pedestal and sync tip further before its application to the cathode-follower output-arrangement and then to the input of the horizontal sync-control system. A suitable grid time constant is chosen (small capacitors and large-value resistors) in order to permit a rapid charge of the capacitor but a very long discharge. Thus the bias is able to follow absolute sync-pulse amplitude without danger of horizontal sync distortion in the cathode-follower arrangement.

It is possible to obtain a-g-c voltage from the cathode circuit of the sync separator because the d-c charge developed across capacitor *C2* is a function of the peak amplitude of the sync pulse applied to the grid. A-G-C voltage is actually the sum of the d-c voltages present (bleeder circuit plus the peak voltage developed by the amount of plate-current flow during the sync-pulse intervals). It is possible to regulate the a-g-c voltage with the a-g-c control in the grid circuit of the separator. A variation of this control changes the d-c bias applied to the grid of the separator because of its bleeder function with the d-c plate voltage direct-coupled from the video amplifier. Thus, as the value of the effective resistance from grid to ground is reduced the positive grid bias also is reduced. Consequently, plate-current flow and the d-c charge on capacitor *C2* change accordingly.

A special noise-suicide circuit is associated with the sync separator because of its association with the screen voltage-supply circuit of the fourth i-f amplifier (refer to Fig. 73b). In this amplifier stage the screen voltage is not well-regulated or filtered and, consequently, will change sharply when the control grid is driven sharply positive. The amplifier bias-voltage, in conjunction with a-g-c action, is sufficient to keep the control grid from going positive with the reception of normal television signal level. However, very strong noise pulses (that do exist in fringe areas with relation to the arriving weak signal) are able to drive the control grid positive; screen grid voltage then drops to a low value and develops instantaneously a negative pulse which is applied to the grid of the vertical sync separator. Of course, the same noise impulse passes through the receiver normally and will appear eventually as a positive noise impulse at the plate of the video amplifier (the same plate from which the sync signal is removed before it (signal) is applied to the separator tubes). However, the noise impulse at this point is of positive polarity and will be cancelled in its effectiveness by the negative impulse that has been taken from the screen circuit of the i-f amplifier and applied to the very same sync-separator grid.

In a typical Philco sync system, Fig. 155, the composite video signal with sync negative is taken from the input of the video amplifier and applied to the grid of the sync-amplifier clipper. This stage is biased with a bleeder network at its cathode, and consequently, it is biased to permit clipping as well as amplification. The initial clipping action at the blanking level and at the sync tip begins in this stage. The positive composite signal at the plate circuit is coupled through a suitable noise-rejection r-c combination to the grid of the sync separator. This coupling arrangement permits a rather fast time constant for high-frequency components of sync, permitting positive grid-swing and clipping of the sync tips and preventing noise impulses from setting a charge that would hold over a long period of time and effect sync clipping and separation adversely. At the same time an effective and longer time constant (*R605-R609-C601*) is maintained for the lower frequency vertical components of synchronization. This grid is signal-biased by grid-current flow that permits the sync separator to conduct only during the higher amplitude sync-pulse intervals (low-value screen voltage and low-value plate voltage obtained by the bleeder network to plus-B consisting of resistors *R612* and *R615*). The same grid-circuit charge can also be used for a-g-c voltage, since its absolute value varies with the strength of the applied composite video signal. Thus, as the composite video signal increases in amplitude, the charge becomes more negative, and a higher negative potential is applied to the a-g-c line.

Impulse noise-rejection is obtained in this circuit by also supplying a second and negative composite signal to the control grid of the sync separator. This signal is obtained from the video-detector output and applied through capacitor *C605*. It is to be noted that when sharp noise impulses manage to leak through into the sync system, they will appear with negative polarity at the control grid and with positive polarity at the signal grid of the sync separator; consequently, there will be noise-cancellation activity. Negative composite sync is applied to the grid of the sync-inverter clipper. Positive vertical sync is applied to the integrating circuits in the vertical sawtooth-generating stage. Horizontal sync pulses, of equal amplitude but opposite polarity, are derived from the plate and cathode circuits of the inverter for application to a phase-comparing horizontal sync-control circuit.

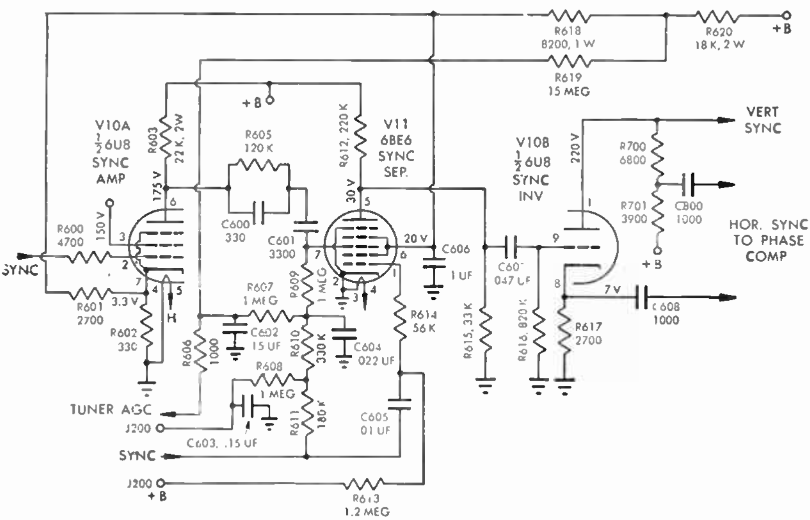


FIG. 155 Philco Sync System

In this system noise impulses that occur immediately before, after, or between the arriving horizontal sync pulses (if they are greater in amplitude than the sync pulses) are effectively removed by noise-cancelling circuit. Also, if sync pulse and noise pulse occur simultaneously, even the sync pulse is removed, because the negative sweep of the combined signals on the control grid causes plate-current cut-off. However, this occasional loss of an actual sync pulse is less objectionable than the presence of the noise impulse which can erroneously fire the sawtooth oscillators. The flywheel action of the horizontal sync-control system and the slow integration of the vertical sweep system are not adversely affected by the occasional loss of a sync pulse.

Another sync system developed by Philco, Fig. 156, also uses a noise-cancellation circuit as well as a special stage that prevents the loss of any sync pulse when noise and sync pulse occur simultaneously. Positive composite video is applied to both the grid of the noise-inverter *V14B* and to the grid of the triode sync separator. The output of the noise-inverter is also applied to the grid of the sync separator through capacitor *C602*. The noise-inverter is normally biased beyond cut-off by the bleeder network to plus-B in its cathode circuit. Thus the noise-inverter does not conduct until the noise impulse exceeds in amplitude the sync tip. A strong noise impulse at the grid of the noise-inverter will develop a negative noise pulse at its plate circuit. At this same point, of course, a positive noise impulse exists because of the presence

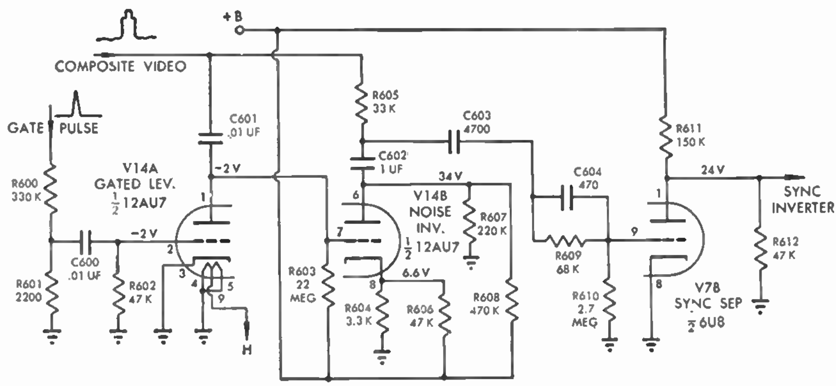


FIG. 156 Noise-Inverter and Sync Separator

of the composite video signal. However, the noise impulses are of opposite polarity and cancel before the composite signal is applied to the grid of the sync separator. Thus positive noise impulses greater in amplitude than the sync tip are eliminated, and erroneous synchronization is avoided. Of course, we realize that negative noise impulses (which occur less frequently) are also clipped off, because all components of signal information below the blanking level are removed by cut-off action at the sync separator.

The noise-inverter is prevented from conducting during the sync-pulse interval by the gated level tube. This tube is normally biased to cut-off by grid-circuit charge. However, a portion of the gate pulse from the horizontal deflection system is applied to its grid, and consequently, the level tube conducts during the horizontal sync-pulse interval. Current-flow through resistor *R603* develops sufficient negative bias at the grid of the noise-inverter to prevent its conduction during the horizontal sync-pulse interval. Thus, the noise-inverter circuit will never cause any cancellation of the horizontal pulse itself while that pulse is applied to the grid of the sync separator.

QUESTIONS

1. What constitutes the low- and high-frequency components of a pulse?
2. What are the steps involved in the preparation of the composite television signal for synchronization of the individual sweep systems?

3. Explain the action of a series *R-C* combination on the fidelity of a pulse when time constant is much shorter than duration. Much longer than duration.
4. Explain action of parallel *R-C* combination on same pulse when time constant is longer than pulse slope. Much shorter than pulse.
5. Explain significance of differentiation and integration in the television system.
6. What considerations must be made in choosing proper time constant for passing or differentiating a pulse?
7. What effect does distributed circuit capacity have on fidelity of a pulse?
8. What is function of sync separator?
9. What significance does time constant have in a signal-biased separator?
10. Why is a signal-biased separator preferable to external bias?
11. Explain functions of inter-sync separator.
12. Is horizontal sweep ever synchronized during vertical sync intervals?
13. Why do serrated sync pulses occur at double-line rate?
14. Explain in detail the function of equalizing pulses.
15. How does vertical retrace interval between fields differ from retrace interval between frames?
16. Explain importance of time constants in the inter-sync separator.
17. What are the functions of sync clipper and amplifier stages?
18. How does a sync control system reduce noise synchronization?
19. Explain in detail the operation of a sync control system.
20. Describe a typical commercial sync and inter-sync system.

Chapter 8

SWEEP SYSTEMS

91. *Sawtooth Generation*

The sawtooth generator and amplifier forms and shapes the sawtooth waves which are used to move the beam back and forth across, and slowly down the screen, and then back again. These sawtooth waves (one horizontal and one vertical) are generated continuously whether sync pulses are applied or not. When sync pulses are applied, however, the generators are locked in at the precise frequency of the sync pulses. Nevertheless, the generators are in continuous operation and although no signal is being received a scanning raster is present on the face of the picture tube. This scanning raster is not stable and retrace lines are seen streaking back and forth on the screen. When sync pulses are applied the scanning raster becomes rigid and does not streak and flicker. The instability of the scanning raster without received sync pulses is caused by the shifting of frequency and phase of the scanning oscillators. It is a function of the sync pulses to keep these receiver sweep generators locked in frequency and phase with similar oscillators at the transmitting stations which are generating the deflection voltages used for the camera-tube sweep systems.

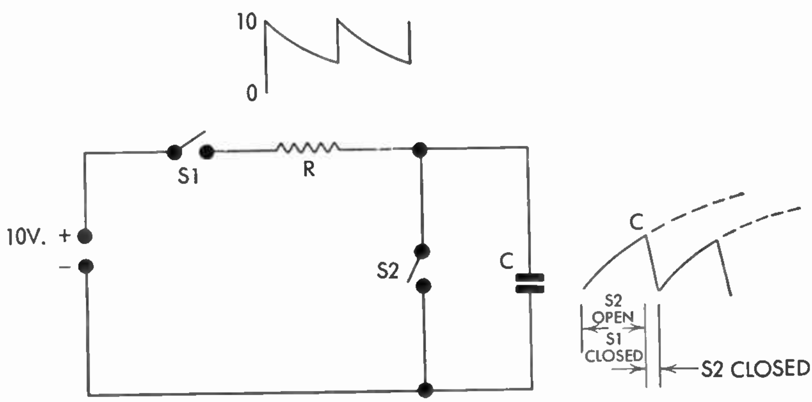


FIG. 157 Simple Sawtooth Generator

There are a number of sawtooth generating systems, all of which depend on the exponential charge of a capacitor which over the initial portion of its charging cycle has an essentially linear voltage rise. Formation of a sawtooth can be conveniently explained with the simple circuit of Fig. 157. At the instant that switch *S1* is closed (switch *S2* remains open), capacitor *C* begins to charge through resistor *R*. Capacitor *C* attempts to charge to the level of the 10 volts applied, the rate of charge depending on the time constant of *R* and *C*. The charging cycle of this capacitor, however (Fig. 158) is not linear. Therefore if we utilize the full charging cycle to generate a sawtooth voltage, the beam, in the case of horizontal motion, would slow down as it approached the right-hand side of the raster. This gradual decrease in rate of voltage rise

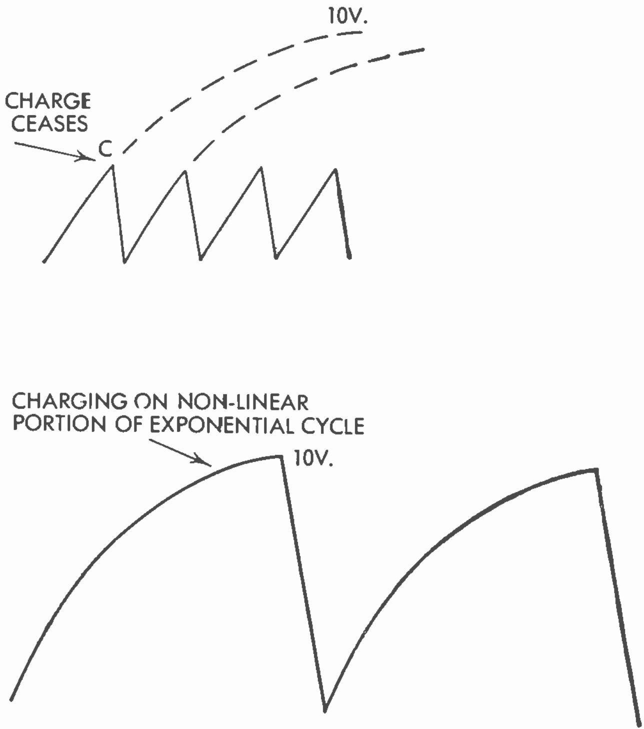


FIG. 158 Generation of a Sawtooth

and consequent decrease in beam velocity would cause the presentation of the picture elements on the right-hand side to be crowded. This is called *non-linearity*. Likewise, if the vertical sawtooth were to round off at the apex. the presentation of the lines at the bottom of the screen would be crowded, and any scene or persons presented on the screen would appear out of proportion. To present a true likeness of any scene it is absolutely necessary that the beam velocities remain absolutely constant as the beam scans across the screen, and as it covers line after line down the screen. To prevent the generation of a rounded sawtooth the charge on the capacitor is terminated and discharged before it reaches the nonlinear section of the exponential charging cycle. The basic method employed is again demonstrated simply in Fig. 157.

At the instant that switch *S1* is closed, capacitor *C* begins to charge toward 10 volts through resistor *R*. Now after it has charged to level *C* (Fig. 158), switch *S2* is closed and switch *S1* opens, causing the capacitor to quickly discharge. After the charge is quickly removed, switch *2* again opens and switch *1* closes and the capacitor begins to charge again. It is apparent through the action of the switches that only the linear portion of the capacitor has been used and that the sawtooth voltage is essentially linear as compared to the type of waveform which would be obtained if the capacitor were permitted to charge near 10 volts. The frequency of the generated sawtooth is, of course, determined by the time constant of the resistor and capacitor or by how long it takes the capacitor to charge to level *C* (point at which *S2* closes). The longer the time constant (higher resistor value and larger capacitor) the more time is required for the capacitor to charge up to level *C*, and consequently the lower the frequency of the generated sawtooth.

92. *Sawtooth Discharge Tubes*

In actual practice a vacuum tube functions as switch *S2*. When the vacuum tube is conducting, the capacitor discharges; when it is nonconducting, the capacitor charges. A typical discharge-tube circuit (Fig. 159) consists of a plate-circuit *R-C* combination and a driving grid pulse of short duration. At the

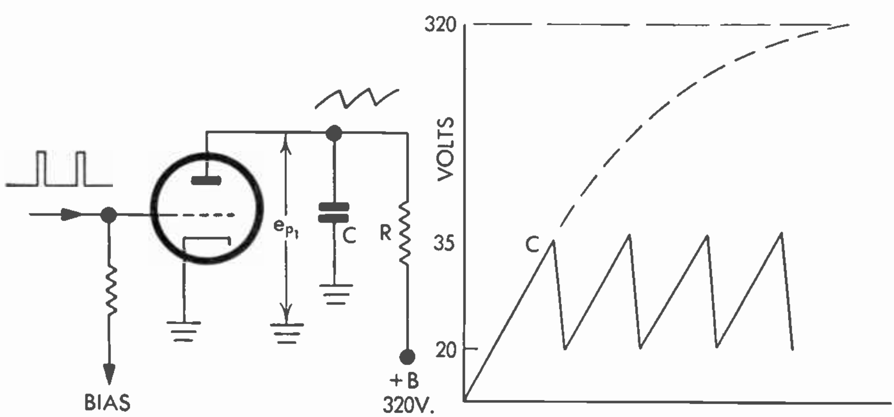


Fig. 159 Discharge-Tube Sawtooth Generator

instant the circuit is placed in operation, assume the tube is nonconducting and, therefore, capacitor *C* will charge through resistor *R* toward the *B* supply voltage level. After a specific charge has accumulated on capacitor *C*, a pulse strikes the grid of the discharge tube and causes it to conduct. Conducting tube is now discharging capacitor *C* quickly because the time constant formed by *C* and the conducting tube is very short in comparison to the long time constant of *R* and *C*. The time interval between pulses and the time constant of R and C are so chosen that the capacitor will charge only over the linear portion of its charging cycle after which a pulse will arrive at the grid of the discharge tube and cause the capacitor to be drained of charge.

Actually if the charge and sawtooth is to be sufficiently linear the capacitor may be permitted to charge to only 5 per cent of the applied supply voltage. The rapidity of the discharge is set by the plate resistance of the tube and the value of the capacitor. It is desirable to have a very fast discharge because the discharge portion of the sawtooth is the retrace portion of the deflection sweep which must occur quickly if the next active line is to be scanned a short time later. It is interesting to note that after the first sawtooth cycle the capacitor never discharges to zero because of the plate resistance of the tube—a certain plate voltage always existing between plate and cathode of the tube, low as it may be, when the tube is conducting. In the example the conducting plate voltage was 20 volts and, therefore, on the rise of the sawtooth the charge always begins at 20 plate volts and rises to 35 volts during the interval between grid pulses. The frequency of the generated sawtooth is set by the rate of arrival of the grid driving pulses. In the RCA receivers which use horizontal sync-lock control this is the exact manner in which the horizontal sawtooth is formed. The rate of arrival of the pulses at the grid of the discharge tube determines the frequency of the generated sawtooth. Inasmuch as the arriving pulses initiate the discharge of the capacitor, they also determine the phase of the horizontal retrace because it is during the pulse that the actual retrace is initiated.

In other systems, an oscillator generates the grid driving pulses for the discharge tube and the oscillator itself is synchronized by the incoming received sync pulses. A number of these oscillators will be discussed in the following paragraph.

93. *Sawtooth Oscillators*

BLOCKING-TUBE OSCILLATOR AND DISCHARGE TUBE

One of the most widely used sawtooth oscillators is the blocking tube (Fig. 160). In this type of sawtooth oscillator, properly phased feedback is obtained by means of a transformer. This feedback cuts the blocking tube on and off, the trace of sawtooth forming when the blocking tube is cut off and the retrace occurring when the blocking tube conducts. The blocking tube can be used to generate the fast horizontal sawtooth or the much slower vertical sawtooth. The primary difference in the two circuits is the value of the time constant and construction of blocking-tube transformer. The inductance of the windings of the vertical blocking-oscillator transformer is, of course, higher than the inductance of the horizontal transformer, which must have a minimum of distributed capacity to properly pass the high-frequency horizontal sawtooth and its harmonic components. Thus the horizontal transformer has maximum

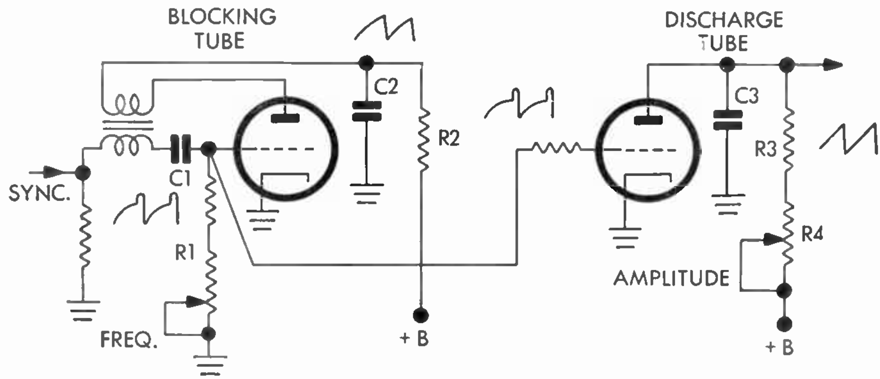


FIG. 160 Blocking-Tube Oscillator and Discharge Tube

transfer at frequencies above 15,750 while the lower frequency vertical transformer is peaked at frequencies near 1,000 cycles and less.

To study the sequence of operation of such an oscillator, assume that the grid has gone slightly positive. A positive-swinging grid (Fig. 161) causes the plate current to increase and the plate voltage to decrease. The negative drop across the transformer primary is transferred to the secondary with proper polarity to reinforce the original positive swing of the grid. This feedback cycle rapidly drives the grid positive until it reaches a level where a further increase in grid voltage causes no further change in plate current. At this point there is no transformer feedback and the grid voltage begins to slide back rapidly, causing plate current to decrease and plate voltage to increase. This new plate voltage change, now of opposite polarity across the transformer primary, causes a negative voltage to appear on the grid of the tube, initiating a new feedback cycle of opposite polarity.

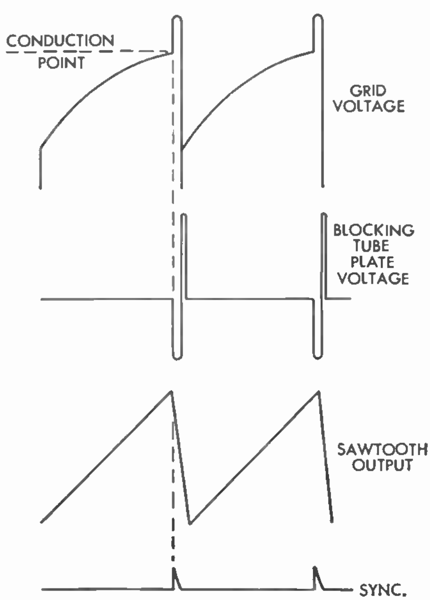


FIG. 161 Blocking-Tube Waveforms

The grid is now driven well beyond cutoff by the large plate-voltage swing and the fast (short time constant) collapsing field of the transformer when the tube is cut off. A large negative charge remains on capacitor *C1* (result of feedback and grid current flow), which now starts to leak off through resistor *R1* while the tube remains cut off. The capacitor now discharges, and current through the resistor continues to decrease until grid again reaches the level at which the tube will begin to conduct. Time required for the charge to leak off determines the period of the sawtooth cycle, and consequently its frequency. Thus the frequency of the sawtooth is dependent on the time constant of *C1* and *R1*—the longer the time constant, the lower the frequency.

In some applications the actual sawtooth voltage can be taken off across capacitor *C2*. At the time the tube is cut off, this capacitor charges through resistor *R2*, generating the sweep trace. When the tube conducts, the capacitor rapidly discharges via the tube generating the retrace portion of the sawtooth. However, in many cases a special discharge tube, as shown in Fig. 160, is used to generate the actual sawtooth. This discharge tube is direct-coupled to the blocking-tube grid. At the same time the blocking-tube grid goes sharply positive, the discharge-tube grid does likewise; when the blocking-tube grid is cut off, the discharge grid is also cut off. When the discharge-tube grid is cut off, capacitor *C3* in the plate circuit of the discharge tube slowly charges through resistors *R3* and *R4*; when it conducts, the capacitor is discharged. The use of the discharge tube permits generation of an extremely linear sweep plus a fast retrace because the time constant can be made exceptionally long. This permits the capacitor to charge over only the most linear portion of the charging cycle. If the capacitor *C2* and resistor *R2* were to have such a long time constant (high-value resistor and large capacitor), the plate voltage on the blocking tube would fall to an extremely low value, output stability being adversely affected.

The blocking tube is synchronized by a positive pulse (as will be explained) applied to its grid, which drives the grid to the conduction point each time it occurs. Thus a retrace (discharge of sawtooth capacitor) occurs each time a sync pulse is applied.

CATHODE-COUPLED MULTIVIBRATOR

In the multivibrator circuit (Fig. 162) feed-back action between two tubes (generally a duo-triode) permits one tube to conduct and then the other. When the second tube is not conducting, capacitor *C1* charges through resistors *R1* and *R2*, generating the sawtooth trace. When the second tube conducts, the capacitor discharges rapidly through the low resistance presented through the conducting tube, generating the fast sawtooth retrace. It is apparent why the retrace is much faster than the trace when we observe that the capacitor discharges through the low resistance of the conducting tube while it charges through the relatively high resistance of resistors *R1* and *R2*. The first tube of the multivibrator is the feedback tube; it functions in a manner similar to the transformer discussed in connection with the blocking-tube oscillator. That is, the first tube supplies feed-back signal of the proper polarity to drive the second tube grid further positive when the grid voltage is rising, and further negative when grid voltage of number *2* tube is decreasing.

To explain the operating cycle, first assume grid *2* is swinging positive. This means that tube *2* is drawing more plate current through the common cathode resistor, *R6*, applying a negative bias to tube *1* (higher plus on cathode). This reduces tube *1* plate current and increases its plate voltage. An increase of tube *1* plate voltage is felt as an increase in tube *2* grid voltage, reinforcing the original increase in tube *2* grid voltage. If feedback occurs rapidly and practi-

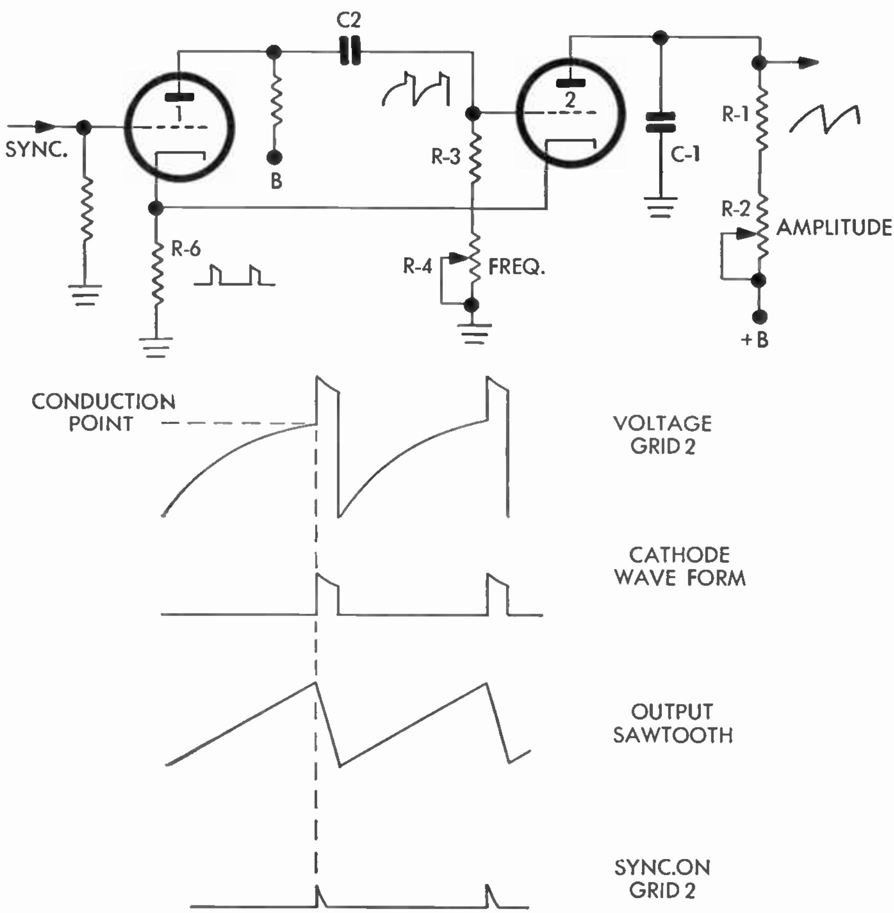


FIG. 162 Multivibrator Sawtooth Generator

cally instantaneously, it drives tube *1* toward cutoff and grid *2* positive to the point where a further increase no longer causes any further increase in tube *2* plate current. At this instant, no further feedback occurs and tube *2* grid voltage begins to slide back, reversing the feed-back cycle. This feed-back cycle continues until, aided by the amplification of tube *1*, tube *2* grid is driven way beyond cutoff by this decided drop in tube *1* plate voltage. Tube *2* is now cut off and will remain cut off until capacitor *C2* has had time to discharge through *R3* to the level at which tube *2* again conducts. Since tube *2* has been going in positive direction to reach its point of conduction, a new feedback cycle begins which drives grid *2* positive and grid *1* negative.

During the interval that tube *2* is cut off, capacitor *C1* has been charging through resistors *R1* and *R2*. At the instant tube *2* conducts, capacitor *C1* is discharged through the tube. The frequency of the sawtooth sweep generated is varied by resistor *R4*, which controls, over a limited range, the rate of discharge of capacitor *C2* and, therefore, the time interval required for it to discharge to the voltage level which will cause tube *2* to conduct. That is, the actual period of the sawtooth wave is set by the grid time constant of tube *2*, which therefore sets frequency of sawtooth. The amplitude of the sawtooth is controlled by resistor *R2* in the plate circuit. It determines to what amplitude *C1* will charge before the tube is driven to conduction by the increasing grid voltage. Thus, the cycling of the multivibrator is set by the grid time constant of tube *2*, and the amplitude or the level to which capacitor *C1* charges during the cutoff interval determines the sawtooth amplitude.

94. *Synchronization of Sawtooth Generators*

The various sync input circuits of sawtooth generators are synchronized by sharp pulses which occur in repeating sequence and initiate the retrace period of the horizontal and vertical sawtooth waves. These sync pulses are the differentiated horizontal sync pulses from the inter-sync separator and the integrated vertical sync pulses. When sawtooth oscillators such as the blocking tube or multivibrator type are used, the sync pulses are usually applied directly to the controlling grid of the sweep oscillator. If a sync control system is used, the sync pulses are used as a comparing frequency in a sync control discriminator.

SYNCHRONIZATION OF SAWTOOTH OSCILLATOR

The free-running or nonsynchronized frequency of the sawtooth oscillator is always lower than the frequency at which the oscillator is locked in by the sync pulses. It is important to understand that the oscillator locks in only when the free-running frequency is reasonably close to the frequency of arrival of the sync pulses. Thus each oscillator incorporates a frequency control to bring the free-running frequency within range of action of the sync pulses. As soon as the frequency is adjusted within this range, the applied sync pulses lock it on the frequency at which they are arriving. If you observe the scanning raster when this frequency or "hold adjustment," as it is often labeled, is made, you can see the oscillator snap into synchronization as indicated by a stationary, rigid pattern. What occurs during the synchronization is shown in the waveforms for a blocking oscillator in Fig. 163. The grid-voltage wave of the blocking oscillator is shown in drawing A. The interval between points *1* and *2* represents discharge of the grid capacitor, the tube having been previously driven beyond cutoff and the capacitor now discharging toward the conduction point of the tube. As soon as the negative charge has leaked off to this level, the tube conducts and plate current flow initiates the feedback cycle, which drives the grid rapidly positive to grid limiting, point *3*. Now the feedback polarity reverses and the grid is driven rapidly beyond cutoff to point *4*. Then begins a new cycle of grid-voltage change. This is the cycle of events as it occurs when the oscillator is free running. A similar cycle of events occurs for the synchronized oscillator, with the exception of a short interval before the tube reaches the conduction point during the discharge of the grid capacitor.

When the blocking-oscillator, normal, free-running grid waveform under synchronization reaches point *A*, the sync pulse hits the grid and drives the grid immediately to the conduction point, initiating the feedback cycle. The feedback interval or the normally conducting period of the relaxation oscillation forms the retrace of the sweep sawtooth. A sync pulse, therefore, drives the grid into conduction just ahead of the normal, free-running time. This occurs for every cycle, and consequently the blocking tube is locked in on the pulse frequency (drawing C) which is slightly higher than the normal, free-running frequency of the oscillator. If an attempt is made to synchronize an oscillator, the free-running frequency of which is higher than the pulse frequency, the oscillator will not sync on all sweeps. This causes instability because on some cycles the sync pulse does not occur until after the tube has reached the conduction point. However, in tuning a television receiver we need not worry about this condition, for if we tune for a stationary pattern, our free-running frequency is automatically set at the proper point.

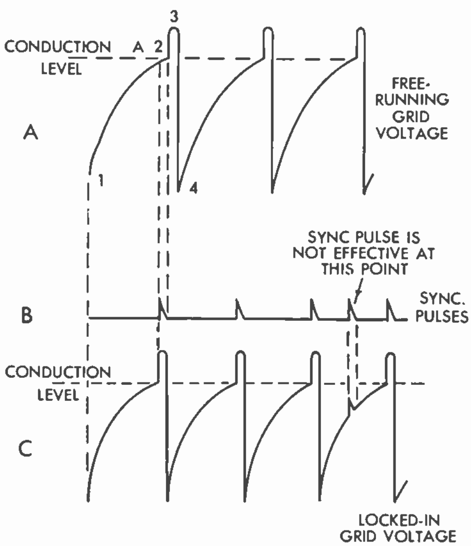


FIG. 163 Synchronization of a Blocking Tube

The sawtooth generating multivibrator is also synchronized by a positive pulse applied to its controlling grid, as shown in Fig. 162. The sync pulse itself, as applied to this multivibrator, is negative-going because it is applied to the grid of the first section of the multivibrator, which acts as a sync amplifier. Positive pulse in its plate circuit is applied as a positive sync pulse to the grid of the second section which is actually the frequency-controlling and sawtooth generating section of the multivibrator.

A disadvantage of the directly synchronized sawtooth generator is the fact that sharp bursts of noise can also synchronize or drive the sweep oscillator to its conduction level. This particularly applies if the noise is high in amplitude and occurs in a repeating sequence at time intervals spaced near to the normal conduction period of the oscillator. For example, a sharp burst of noise immediately ahead of the normal arrival time of the sync pulse will drive the generator into conduction ahead of its normal time, as shown in Fig. 164. This means the retrace of this particular line occurs before it normally should, and the start of the trace of the next line will be early. Consequently all the picture information along the next line will be displaced with respect to the picture information along the previous line. If this noise continues sporadically as the picture is scanned line after line, the actual lines themselves will be displaced with respect to each other, and a defect called *tearing out* will occur in which groups of lines are pulled to the left and others are pulled to the right, breaking

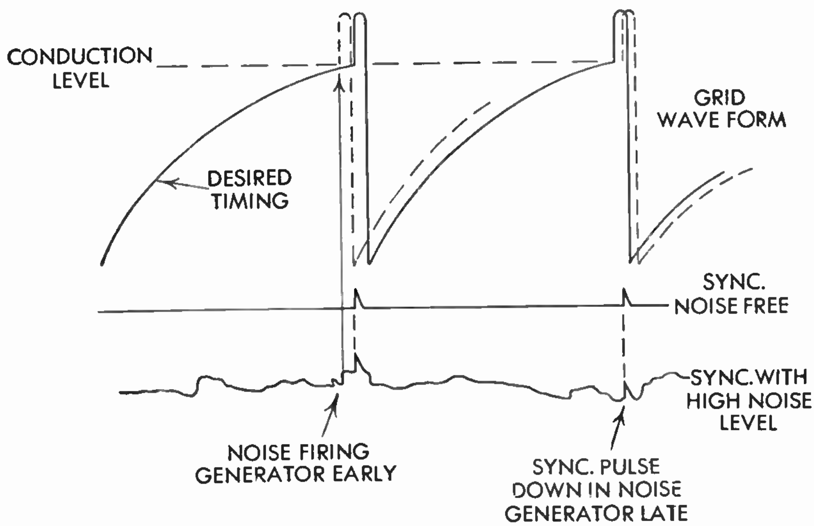


FIG. 164 Effects of Noise on Synchronization Time

down the actual picture structure. Thus, noise will cause synchronization of the oscillator before the normal time or will obliterate the incoming sync pulses themselves and take over the synchronization of the sweep oscillator. Likewise, the same noise, if it is sufficiently great in amplitude, can take over the synchronization of the vertical oscillator in which the actual fields that make up the picture will be displaced with respect to each other. This defect, called *flopping over,* will cause a motion of the entire picture vertically and the picture will move up and down on the screen.

SYNCHRONIZATION OF SYNC CONTROL SYSTEMS

In the synchronization of sync control systems, sync pulse amplitude and amplitude of noise impulses that ride along with them have little effect on the synchronization of the sweep generating circuit. The operation of these circuits is dependent on the recurrence of the sync pulses at specified intervals.

Inasmuch as noise is not likely to occur in a repeating manner with the same frequency as the arrival of sync pulses, noise will not seriously affect the synchronization. Thus, in this system the sync pulses are compared with other waveforms of the same frequency, and if the proper relation is retained between sync pulses and the other frequencies, the generated sawtooth will be the same frequency as the arriving sync pulses.

In the RCA control system, a sine wave generated by a Hartley oscillator is compared with the frequency of the arriving sync pulses in a discriminator. The d-c component at the output of the discriminator is applied to the grid of a reactance tube, the reactance tube keeping the Hartley oscillator on the same frequency as the arriving sync pulses. A second sine-wave output of the oscillator (coupled through electron stream to plate) is clipped to form a squared wave and is then differentiated and applied to a discharge tube which generates the deflection sawtooth. Thus the control of the generation of the sawtooth sweep is dependent on the arriving differentiated wave, which originates at the sine-wave oscillator. Noises which ride along with the received sync pulses are absent. The sine-wave oscillator is held precisely on frequency because it is compared with the rate of arrival of the sync pulses, noise having only a slight effect on circuit operation.

In the General Electric sync control system, a multivibrator is used to generate the sawtooth wave; however, instead of controlling the frequency of the multivibrator with sync pulses directly, the multivibrator is controlled by the d-c potential applied to one of its grids (Fig. 165). In this system, a portion of the horizontal sawtooth at the output of the horizontal sweep system is fed back to a discriminator for comparison with the arriving sync pulses. Whenever the fed back sawtooth wave and the sync pulses do not agree, a d-c component of output appears at the discriminator and is increased in level by a d-c amplifier and applied to the grid of the sawtooth multivibrator. This change in the d-c component of bias applied to the grid of the multivibrator also varies the frequency of the multivibrator until it compares with rate of arrival of the sync pulses. The multivibrator is again unaffected by noise because noise impulses do not reach the sawtooth generating system.

The waveforms associated with Fig. 165 demonstrate the manner in which the d-c potential applied to the first grid affects the frequency of the sawtooth generator. Actually, the supply potential applied to the grid of the first section of the multivibrator is the plate voltage of the d-c amplifier, and whenever this plate voltage changes, the supply potential applied to the grid also shifts. Normally, on the discharge of an *R-C* time constant in the grid circuit of the first stage of the multivibrator, the capacitor discharges to zero, as shown in waveform *A*. Although the capacitor does discharge toward zero, it never reaches the zero point because before it reaches this point at some negative voltage, the tube will again conduct and a new feedback cycle will be originated. The action of the d-c controlled multivibrator is demonstrated in the waveform of drawing B. Because the grid of the first section of this multivibrator is returned to the plate voltage of the d-c amplifier, the capacitor, after it is charged way beyond cutoff by the normal operating sequence of the multivibrator, discharges toward some B-plus value instead of toward zero as does the more conventional multivibrator. Again the capacitor never reaches this plus value because as soon as it reaches a low enough negative potential, the first section of the multivibrator conducts and a new feedback cycle is originated.

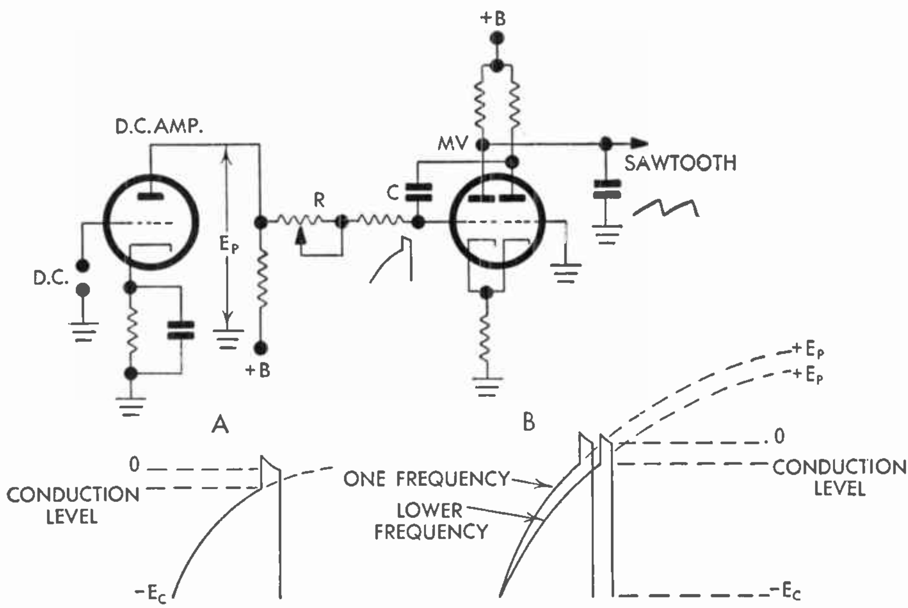


FIG. 165 D-C Control of Multivibrator Frequency

However, the rate at which the capacitor discharges and the speed with which it reaches this conduction point are controlled by the potential toward which it is discharging, as shown for two conditions in drawing B. As can be seen, the slope of the discharge is much steeper when the capacitor is discharging toward the greater B-plus value (higher value of plate voltage on the d-c amplifier as controlled by the d-c potential applied to its grid from the discriminator). If the capacitor discharges from its highest negative value to the conduction point in a shorter period of time, the period of the grid cycle is shorter and, therefore, the frequency of the multivibrator is faster. Thus, if the plate voltage on the d-c amplifier increases, frequency of the multivibrator sawtooth generator also increases. If the plate voltage of the d-c amplifier falls, the potential toward which the capacitor discharges is lower. Therefore, the rate at which it discharges between its maximum negative point and the conduction point of the tube is slower, and consequently the frequency is lower. As a result, the d-c output of the discriminator can change the frequency of the sawtooth multivibrator and make it conform with the rate of arrival of the sync pulses at the discriminator. In doing so it makes the frequency of the sawtooth dependent on the rate of arrival of the sync pulses and reasonably unaffected by noises riding along with the sync pulses.

95. *Sawtooth Frequency, Phase, Amplitude, and Linearity*

The frequency of the vertical and horizontal sawtooth waves at the receiver must conform with the frequency of similar sawtooth waves at the camera tube if the picture is to be reproduced at the receiver in the same order at which it was released at the camera tubes. If the frequency of the two sawtooth waves is not the same (drawing A, Fig. 166), the lines will not be presented in the same order they were released. For example, if the receiver sawtooth is higher in frequency than the camera-tube sawtooth, the retrace of the sawtooth at the receiver will occur before the actual line is completed. Likewise, a new scan will begin at the receiver before the blanking period is terminated. As a result, the information on the lines at the receiver, as the beam scans line after line down the screen, will be displaced line after line and an irregular pattern will be presented.

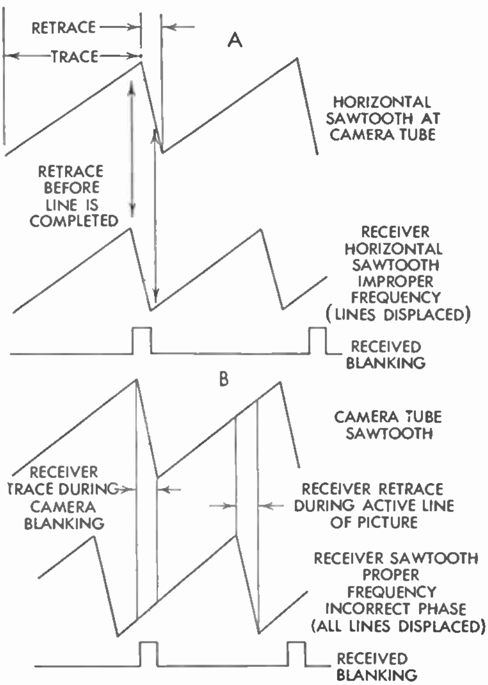


Fig. 166 Frequency and Phase of Receiver Sawtooth in Relation to Blanking

Not only is it necessary that the frequency of the camera- and receiver-tube sawtooth waves be in step, but the phase of the respective sawtooth waves must also be synchronized with respect to the picture and blanking information. For example, if the retrace of the horizontal sawtooth at the receiver differs from the retrace of the camera sweep and horizontal blanking interval, it is possible *that the transmitted blanking pulse will reach the grid of the picture tube at the same time the horizontal sawtooth is moving the beam left to right for the normal horizontal trace.* If this occurs, the horizontal blanking pulse will arrive on the grid of the picture tube and blank out the screen during the active trace of the deflection voltage, producing a black vertical bar (called a *horizontal blanking bar*). Phase displacement between the two sawtooth waves and the received blanking pulses also causes the receiver to retrace during a portion of the period that useful picture information is present on the received signal. Likewise, frequency difference or phase displacement between vertical camera sweep and received vertical blanking with respect to the receiver vertical sweep will cause overlapping pictures and a horizontal black bar called the *vertical blanking bar*. It is obvious, therefore, that it is necessary that the receiver-tube deflection fields reshape when the horizontal and vertical blanking pulses are received because they represent the time intervals during which the camera-tube beam is retracing and no picture information is being conveyed.

Fortunately, those sweep oscillators which are under direct control of the received sync pulses meet this requirement because they are triggered by the received sync pulses, triggering action initiating the deflection-field retrace period (sawtooth discharge). When automatic horizontal synchronization is used, however, it is necessary to phase the sine-wave oscillator or generating multivibrators to produce the sweep retrace during the time that the blanking pulses are being received. If this relation is not properly maintained, black blanking bars appear on the screen. A phasing adjustment can be made on the receiver (horizontal oscillator and discriminator) to remove the defects.

SWEEP WAVEFORM DEFECTS

A number of other sweep waveform defects are shown in Fig. 167, in which drawings A and B show the received blanking pulses and camera-tube sweep. If the retrace period of the deflection field of the receiver is too long, as indicated in drawing C, it means that the picture-tube beam is unblanked, and picture information is presented to the grid during intervals the deflection fields are reshaping themselves. Consequently, some picture information appears incorrectly on the screen during retrace periods. This produces a defect called *foldover*, which can occur at the left- or the right-hand side of the fluorescent screen. For example, if the receiver deflection field begins to retrace before the blanking pulse is applied to the control grid, some information is placed on the screen when the beam begins its return to the opposite side. Consequently, a bright and overlapping pattern is seen on the right-hand side of the screen. Likewise, if the retrace takes too long a period of time, picture information will again appear on the screen before the fields have been completely retraced to the left side of the screen. Thus, picture information is present on the screen while the beam is retracing the last short distance and then again as the beam swings over the same section and starts a new trace. This time there will be overlap on the left-hand side of the screen. Often foldover is caused on the right-hand side of the screen by poor frequency response in sweep system, which causes the horizontal sawtooth to fold over at the apex instead of being a sharp change from one direction to the other. This defect sends the beam on its way back to the left side before all information along the line is presented to the grid of the picture tube. Therefore, to prevent image defects on the fluorescent screen, it is important that the sawtooth wave have a fast enough retrace and that it have a sharp break between trace and retrace and again between retrace and trace of a new line.

The amplitude of the sawtooth wave (drawing D) determines the area over which the picture tube scans the fluorescent screen. In the drawing, various amplitude sawtooth waves of the same frequency have been drawn, which indicate that it is possible, when the sawtooth is too great in amplitude, that a portion of the picture will run off the edges of the screen; likewise, if the sawtooth amplitude is too low, the received picture will not cover the entire area on the fluorescent screen. Thus, the amplitude of the vertical deflection field determines the height of the picture on the screen, and the strength of the horizontal deflection field determines the width of the picture on the screen. Each must be adjusted to produce a picture which has an aspect ratio of 4:3, producing images in correct proportion.

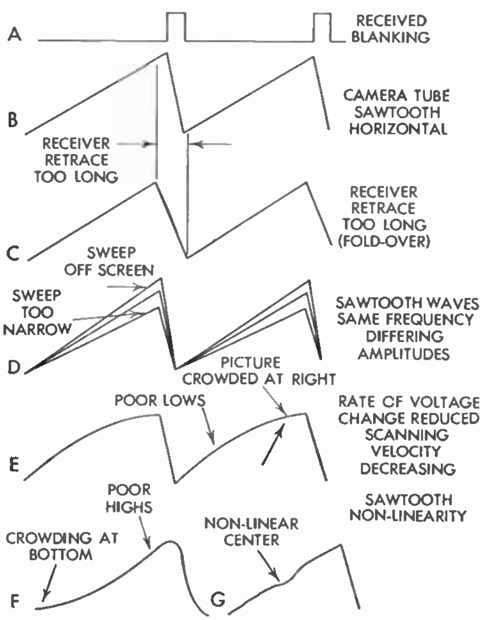


FIG. 167 Sweep Waveform Defects

Another very important characteristic of the sawtooth waveform is its linearity or whether the trace of the sawtooth wave is a straight-line function or has a decided curvature. For example, in drawing E, the waveform, as it rises toward the apex, has a lower and lower rate of voltage change; consequently, as this sawtooth causes the beam to sweep from left to right across the screen, the actual velocity of the beam begins to slow down as it approaches the right side. However, the picture information continues to arrive at the grid of the picture tube in the same order and at the same rate it was released at the camera tube; therefore, the picture information appears crowded on the right-hand side of the fluorescent screen because of the decreasing velocity of the scanning beam. This defect, called *nonlinearity,* can be overcome by making certain that the trace of the sawtooth is absolutely linear and represents a constant beam velocity from left to right across the screen. In drawing F of Fig. 167, the early part of the sawtooth trace has a slower rate of voltage change than the voltage change associated with the portion of the trace near the apex. As a result, the picture information, when presented on the grid of the picture tube, causes the image to crowd on the left side where the scanning beam is moving slowly. On the right-hand side of the beam, the information is being presented at the proper rate. Still another defect is nonlinearity at the central portion of the trace, which causes nonlinearity of the picture information at the center of the reproduced image on the screen while the left and right sides are normal.

It is apparent, therefore, that the presentation of the picture on the screen depends a lot on the fidelity of the deflection field—whether it is represented by a current sawtooth, as in magnetic deflection, or by a voltage sawtooth, as in electrostatic deflection. Likewise, it is important that the generated sawtooth at the sweep oscillator have good fidelity and that the fidelity of this deflection sawtooth be maintained as it is increased in amplitude by the deflection amplifiers.

DEFLECTION WAVEFORM CONTROL

There are four basic controls which maintain the fidelity of the deflection waveform and ensure that proper adjustments can be made to reproduce a true replica of the scene picked up by the camera tube. These controls regulate the amplitude, frequency, linearity, and phase of the deflection waveforms. The amplitude controls, or so-called "size controls," set the width and height of the scanning raster and picture on the fluorescent screen by regulating the peak amplitude of the deflection sawtooth wave. When electrostatic deflection is used, size controls are associated with the sawtooth generators and set the level toward which the sawtooth-forming capacitor charges and, therefore, set the amplitude of the sawtooth waves. If magnetic deflection is used, the vertical size control is generally at the same point, but the horizontal size control is often associated with the horizontal sweep amplifier output circuit.

A *phase control* is associated only with those systems which use automatic sync control; it makes certain that the retrace of the deflection sawtooth waveforms occurs at the time the blanking pulses are present on the grid of the picture tube and not during the intervals picture information is on the grid.

The frequency controls are always associated with the sweep-oscillator circuits and are commonly called *hold controls* because they set the frequency of the horizontal and vertical oscillator at a point near the frequency of the arriving horizontal and vertical sync pulses—the sync pulses now grabbing hold to keep the deflection waveform and frequency matched to the rate of arrival of sync pulses.

Horizontal and vertical *linearity controls* are located at various points in the sweep amplifiers to maintain the linearity of the sweep waveforms. When electrostatic deflection is used, the linearity control corrects for deficient frequency response or exponential charge of the sawtooth-forming capacitor. When magnetic deflection is used, similar linearity systems are employed, but in addition, linearity circuits are associated with the horizontal damping system to preserve the linearity of the current sawtooth in the deflection coils.

96. *Electrostatic Deflection*

The deflection amplifier increases the amplitude of the sawtooth voltage to a level which produces satisfactory deflection of the scanning beam of the picture tube. In order to obtain balanced deflection, the electrostatic amplifier is generally a phase-inverter type to obtain a positive- and negative-going sawtooth wave, the positive sawtooth applied to one deflection plate and the negative sawtooth to the other deflection plate of the pair. Actually there are two deflection amplifiers—one for the vertical and a second for the horizontal

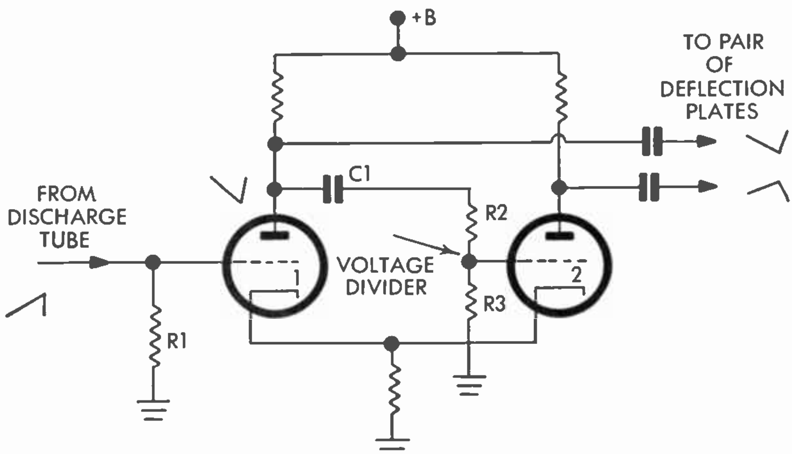


Fig. 168 Electrostatic Sweep Amplifier

sweep. In the generation and amplification of the vertical sawtooth wave, every precaution must be taken to prevent low-frequency loss and consequent distortion of the sawtooth. In the case of the horizontal sweep generator and deflection amplifier, similar precautions must be taken to prevent loss of the high-frequency components of the horizontal sawtooth which particularly affect the apex of the sawtooth where the deflection system is swinging from trace to retrace intervals.

A typical electrostatic deflection amplifier (Fig. 168) is excited by a positive sawtooth from the sawtooth generator which produces an amplified negative sawtooth in the plate circuit of the first tube. This negative sawtooth is applied to one of the deflection plates and at the same time is applied to a voltage divider which excites the grid of the second tube.

In order to produce a positive sawtooth with the same polarity in the plate circuit of this tube, it is necessary to apply the proper amplitude voltage to its grid. Thus, the negative sawtooth in the plate circuit of the first tube is reduced in amplitude by the voltage divider to develop the negative sawtooth of proper amplitude across *R3* and on the grid of the second tube. This negative sawtooth develops a positive sawtooth in the plate circuit of the second tube which has the same amplitude as the negative sawtooth in the plate circuit of the first tube, but has an opposite polarity. The positive sawtooth is now applied to the opposite deflection plate. Thus the deflection amplifier has a positive-going sawtooth applied to its input circuit and develops equal amplitude but opposite polarity sawtooth voltages in its output circuit to provide balanced excitation to the deflection plates of the picture tube. The phase-inverter circuit forms a convenient and effective means of obtaining balanced deflection voltages.

The resistor divider is the simplest means of obtaining the proper voltage division to excite the second tube. Two other means of obtaining the proper amplitude voltage to excite the second tube are shown in Fig. 169. In the first

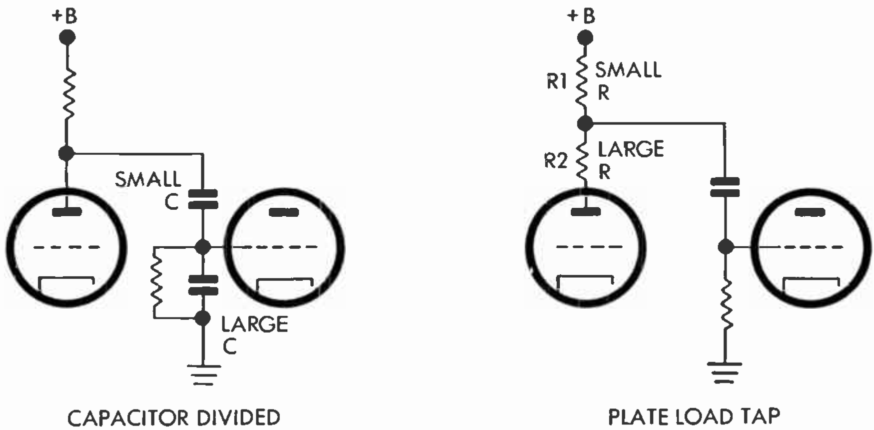


FIG. 169 Voltage Division Methods for Horizontal Sweep Amplifier

method, which is particularly applicable to the horizontal sawtooth waveform, a capacitor divider is used instead of the resistor divider. A capacitor divider has the added advantage of overcoming frequency discrimination at the fundamental frequency of the sawtooth, and its much higher frequency harmonics which are particularly high in the case of the horizontal sawtooth. If a conventional resistor divider is used, the high-frequency components of the sawtooth (near the apex) are attenuated considerably in feeding the second tube because the input capacity of the tube shunts the voltage divider resistor and changes divider ratio for the higher frequency harmonics of the sawtooth. This defect is overcome by using a capacitor divider which produces the same voltage division regardless of frequency. Therefore, the sawtooth voltage applied to the grid of the second tube will have the same shape and fidelity as the sawtooth developed in the plate circuit of the first tube. The capacitor divider is a very common circuit used in many wide-band high-frequency applications. The division of voltage always remains the same because as the frequency changes the reactance of both capacitors follow along. When a straight resistor divider is used, and any one of the resistors is shunted by a capacitor, a changing reactance of that capacitor, particularly its much lower value at higher frequency causes the voltage division to change. If a waveform is impressed on the divider which has many high-frequency components it distorts.

Another system of voltage division is shown in the second drawing in which the actual plate load resistor of the first tube is tapped, and the excitation for the second tube is taken off only a portion of the full plate load of the first tube. Therefore, it has the low amplitude value necessary to excite the second tube. Once again the output of both tubes are sawtooth waveforms of opposite polarity but equal amplitude. This method again has less high-frequency loss because the excitation for the second section is taken off a low-impedance point (across the small plate resistor *R1*) and, therefore, the capacitive loading is very much less detrimental.

SWEEP LINEARITY

Every precaution must be taken to preserve the linearity of the sawtooth voltages to produce a linear presentation of the image on the screen both horizontally and vertically. One source of nonlinearity in the generation of the sawtooth voltage is the fact that the sawtooth-forming capacitor itself (no matter how perfectly the circuit is designed) has a certain degree of nonlinearity because it charges over a portion of the exponential charging cycle which is not linear. Thus, the sawtooth voltage to some extent tends to round off as it approaches the apex.

Many deflection amplifiers, particularly those associated with the amplification of the much lower vertical sawtooth, have low-frequency degeneration because of the increasing reactance of coupling capacitors and low-frequency degeneration in cathode circuits. One source of poor low-frequency response in the electrostatic deflection amplifiers is the coupling capacitors which couple the deflection amplifiers to the deflection plate of the picture tube. On these deflection plates the second-anode potential is also applied to keep the deflection plate at a potential near to that of the second anode, and to permit deflection of the beam to the right or left of the center of the center position. This means that the coupling capacitor must withstand an unusually high voltage, and a good quality capacitor must be used. It is a fact that the higher the value of this capacitor the larger it becomes and the more leakage and tendency to break down exists. Therefore, it is advisable to use as low a value coupling capacitor as possible; but, at the same time, severe low-frequency degeneration must not be present. Actually, a certain amount of this distortion is present and produces a sawtooth voltage which appears rounded on the trace, as shown at point *B* in the waveform drawing of Fig. 170.

This exponential rise can be compensated for by means of a linearity circuit in the grid circuit of the sweep amplifier which actually causes the sawtooth applied to it to dip on the trace. The combination of the high-frequency loss in the grid circuit along with the low-frequency loss in the plate circuit produces an essentially linear sawtooth in the output. Thus, by means of a frequency discriminating circuit we have restored the fidelity of the sawtooth waveform which normally would have a rise in it because of the poor low-frequency response introduced by the low value coupling capacitor. The linearity system in the grid circuit (Fig. 170) consists of a shunt *R-C* combination with one capacitor shunted by a variable resistor which can increase or decrease the effectiveness of this capacitor. When the arm on the potentiometer of the linearity control *R3* is at the top, capacitor *C2* is shorted and substantial high-frequency loss exists to ground through capacitor *C1*. As the arm is gradually moved down the potentiometer and the resistance across *C2* increases,

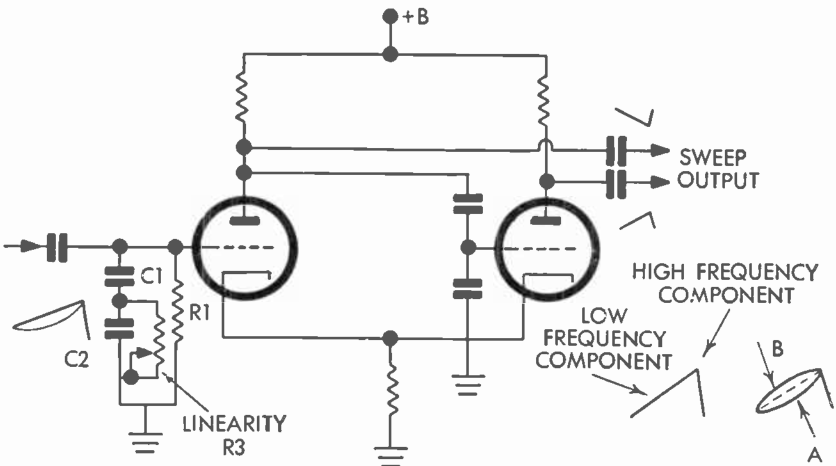


FIG. 170 Linearity Control with *R-C* Compensating Circuits

more effective capacity is introduced in series with *C1* and, therefore, the series capacity is reduced. Consequently, when the arm is pulled down the potentiometer, the capacitive loading at high frequency is presented by the reactance of *C1* and *C2* in series. Actually, the linearity circuit is nothing but a high-frequency-filter combination which filters out some of the high-frequency components of the sawtooth and, therefore, dips the trace to compensate for the poor low-frequency response in the plate circuit which causes rounding of the trace.

A complete deflection amplifier and deflection system is shown in Fig. 171, consisting of a vertical and horizontal amplifier driving the respective deflection plates of a picture tube. The slow-rising vertical sawtooth is applied to the input circuit of the top phase-inverter deflection amplifier, and its output developed across *R3* and *R4* drives the vertical deflection plates. Proper amplitude division for the lower frequency vertical sawtooth is obtained with a simple resistor voltage divider, *R1* and *R2*, which divides the output of the first tube for proper excitation of the grid of the second tube. To couple the deflection voltages to the vertical plate, capacitor *C1* and *C2* must be of relatively high value (0.05 microfarad or higher). In many circuits it is necessary to use a linearity-control network in the grid circuit of the vertical amplifier, to prevent low-frequency degeneration of sawtooth.

The horizontal sawtooth is applied to the lower sweep amplifier. It develops equal-amplitude opposite-polarity sawtooth voltages across *R5* and *R6* which are capacitively coupled to the horizontal deflection plate. In the horizontal

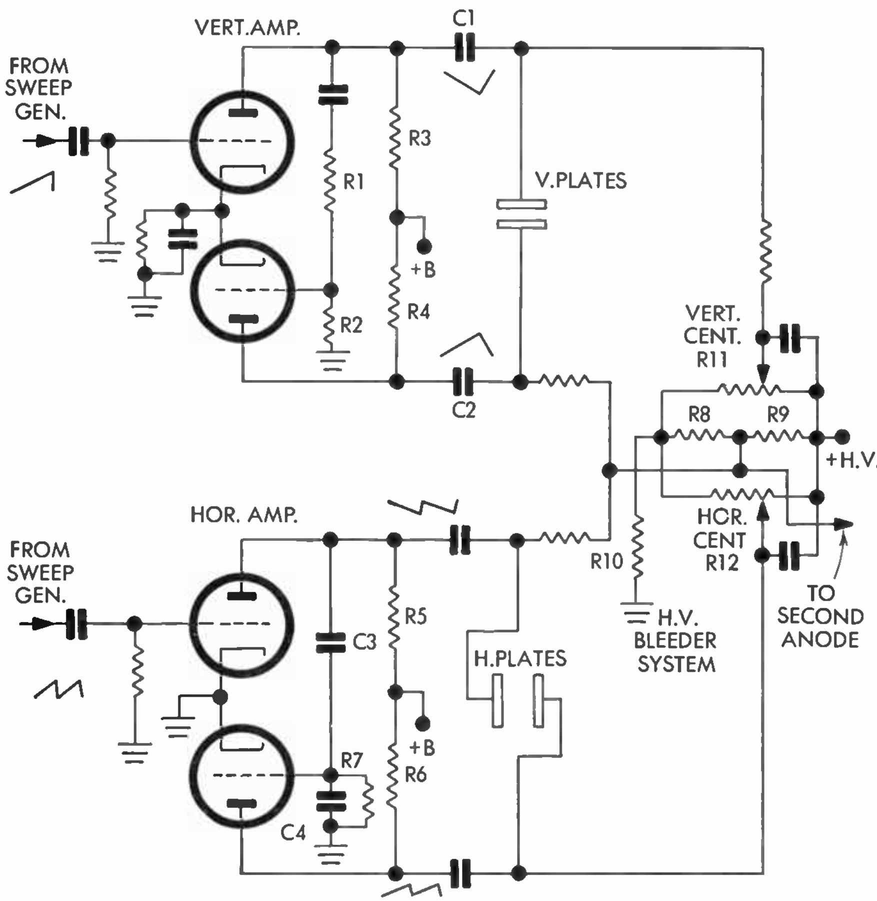


FIG. 171 Horizontal and Vertical Sweep Amplifiers

amplifier, because of the higher frequency harmonics of the sawtooth wave, a capacitive divider is used to reduce the amplitude of the sawtooth for proper excitation of the grid of the second tube. This divider consists of capacitors *C3* and *C4* and permits application of a sawtooth voltage to the grid of the second section which has the same fidelity and appearance as the sawtooth voltage present in the plate circuit of the first tube. Thus the sawtooth voltages when applied to the deflection plates have the same appearance and amplitude.

The method used to center the scanning raster on the fluorescent screen properly is also shown in Fig. 171. Resistors *R8*, *R9*, and *R10* are a part of the high-voltage bleeder network across the high-voltage power supply, and from the junction of *R8* and *R9* the second anode of the picture tube receives its potential. In order to deflect the high-velocity beam on one side or the other of the center point it is necessary that the average potential of deflection plates vary above or below the second-anode potential. Consequently, the second-anode potential from the junction of *R8* and *R9* is fed directly to one deflection plate of each pair, and the opposite deflection plate of each pair can be varied to one side or the other of the second-anode potential. Consequently, the scanning raster as an entirety can be moved above or below or to the right or left of the center condition. For example, with the vertical-centering potentiometer *R11* the top vertical plate can be supplied with a voltage greater or less than the second anode in accordance with the correct d-c potential necessary to properly center the picture on the screen vertically. The horizontal-centering potentiometer *R12* permits application of a potential to the right deflection plate, which is greater or less than the second-anode potential in order to move the entire scanning raster right or left the proper amount to center the image on the screen horizontally.

In summary, the electrostatic deflection amplifiers take the relatively weak sawtooth output of the discharge tube or sawtooth generator and amplify it to the few hundred volts necessary to properly deflect the beam horizontally and vertically to fill the useful screen area. They do so without disturbing the fidelity and linearity of the sawtooth voltage and without disturbing the trace-to-retrace ratio necessary to reshape the deflection field fast enough to prevent loss of useful picture information arriving at the grid of the picture tube.

97. *Magnetic Deflection Amplifiers*

Magnetic deflection amplifiers are used to produce a sawtooth current in the deflection coils which generate the magnetic field that penetrates the glass envelope of the tube and causes deflection of the electron beam. Magnetic fields are used to deflect the beam both horizontally and vertically. However, considerably more power is needed to produce horizontal deflection because of the greater number of times (higher frequency) the beam must be deflected over a period of time as compared to the vertical motion of the beam. Thus, the power required for horizontal deflection is often a hundred to two hundred times greater than the necessary vertical deflection power. In addition, at the higher frequency horizontal rate there are always added circuit and transformer losses.

A typical magnetic sweep amplifier (Fig. 172) is somewhat similar to the audio-output stage which drives the loudspeaker of a conventional radio receiver. The vertical sweep amplifier particularly is very similar to such an audio-output stage because the output transformer simply matches the resistance of the deflection-coil circuit to the output tube. A magnetic deflection amplifier is single-ended instead of push-pull because it is necessary to produce a sawtooth of current in the deflection coils. There is no necessity for producing a balanced-output sawtooth voltage as needed to drive an electrostatically deflected picture tube. A sawtooth or a modified sawtooth waveform is applied to the grid of the deflection amplifier. The change in plate current it creates develops a constant voltage across the secondary of the output transformer

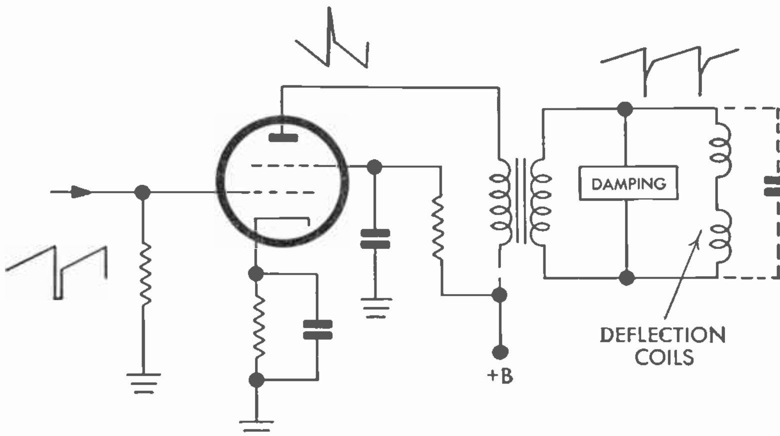


FIG. 172 Sweep Amplifier for Magnetic Deflection

and, therefore, a sawtooth of current is present in the deflection coils. A constant voltage across the secondary represents the ideal case when the deflection coil presents an essentially pure inductance to the amplifier. It is a fact that if a constant voltage is applied across an inductor the current will rise in the coil in a linear manner. In practice, however, there is always some resistance in the load circuit and definite precautions must be taken to ensure the linear rise of the sawtooth current in the deflection coil.

MODIFICATION OF DRIVING SAWTOOTH WAVEFORM

When a pure inductive load is presented to the deflection amplifier a squared voltage would produce a sawtooth of current in the coil. However, there is always a resistive component present and it is necessary that a sawtooth voltage also be present along with the squared voltage to ensure a sawtooth of current in the deflection coils. Thus, as shown in Figs. 172 and 173, the voltage applied to the grid of the deflection amplifier is a modified sawtooth consisting of a rectangular pulse and a sawtooth voltage of the proper ratio with respect to each other to produce the sawtooth of current in the output. If the deflection-yoke circuit presents a definitely resistive load to the deflection amplifier, the applied grid waveform is only a slightly modified sawtooth or, in some cases, simply a sawtooth voltage. When the load presented to the deflection amplifier is essentially inductive, the driving wave applied to the grid has a substantial squared-wave component.

Two methods of generating the necessary modified sawtooth are shown in the drawing of Fig. 174. In the first method a feed-back system modifies the sawtooth voltage applied to the grid of the deflection amplifier, adding a squared-wave component to properly modify the wave. The negative pulse

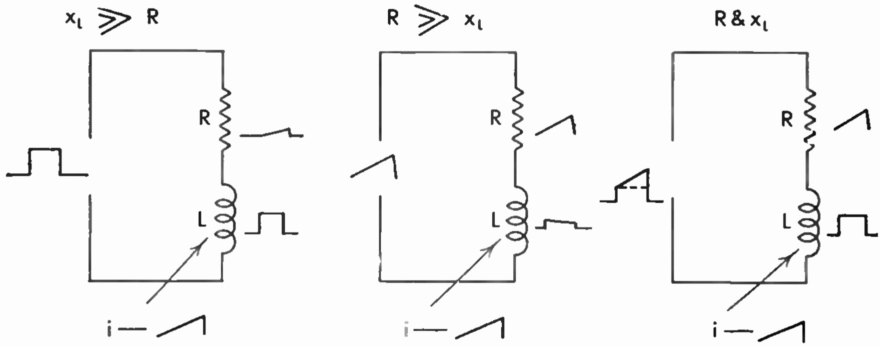


FIG. 173 Modification of Waveform to Obtain Sawtooth Current in Coil

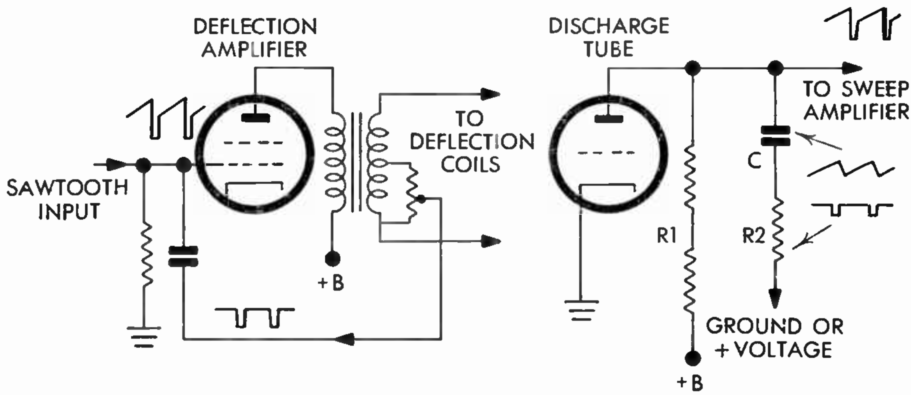


FIG. 174 Modification of Sweep Sawtooth for Magnetic Deflection

which is added to form the squared-wave component is generated in the secondary of the output circuit, the sharp negative pulse developing when the deflection field collapses into the deflection coil during the retrace period of the sawtooth wave. A pulse of the proper amplitude is obtained by tapping it off a resistor which shunts a part of the secondary winding of the transformer.

In the second system the modified waveform is formed in the output circuit of the discharge tube. Resistors *R1* and capacitor *C* form the regular sawtooth-forming *R-C* circuit of the discharge tube, and a smaller value resistor, *R2*, has been added between the normally grounded side of capacitor *C* and ground. When the discharge tube is cut off, capacitor *C* charges normally through resistor $R1$ and, of course, $R2$ which, because of its low value, does not contribute any significant voltage during the charge period. However, on the discharge the capacitor C discharges through the tube and in taking this discharge path also passes a substantial current through resistor $R2$, which, for the interval that the tube conducts, develops a negative pulse across resistor $R2$. Thus the negative pulse and sawtooth charge and discharge on the capacitor add to form the modified waveform consisting of a squared component and a sawtooth necessary to drive the grid of the sweep amplifier. Proper ratio of squared component to sawtooth can be controlled by properly choosing the value of $R2$.

Under some circumstances one end of $R2$, instead of being tied directly to ground, is returned to some positive voltage, developing a pulse with steeper sides and shorter duration across resistor $R2$. The theory behind this operation is that during the discharge of capacitor C, instead of the capacitor attempting to discharge toward zero, it will attempt to discharge toward some negative voltage (by attaching plus voltage to low side of capacitor), which means that the rate at which the capacitor does discharge towards zero will be faster and, therefore, the voltage pulse across resistor $R2$ much sharper. It is important to note that the capacitor never discharges to zero and upward toward this positive voltage because in so doing the cathode of the tube would be positive and the plate negative. Consequently, the tube cuts off as soon as the zero potential point is approached. Nevertheless, the capacitor has been discharged more quickly toward this zero point and, therefore, produces a sharper pulse across $R2$. When the discharge tube stops conducting, the sawtooth again begins to build up on C as it is charged by the electron flow through resistor $R1$.

CHARACTERISTICS OF THE DEFLECTION-OUTPUT CIRCUIT

As mentioned previously, the load on the deflection amplifier is not entirely inductive but also contains resistive components and, as a result, the sawtooth voltage applied to the grid of the deflection amplifier is actually a modified waveform and the sawtooth is present only because a resistive component is present in the load. Such a sawtooth waveform causes a rise in plate current during the trace period of the scanning cycle and is really acting as a regulated resistance (Fig. 175) to compensate for the additional resistive loss in the output circuit as the current rises in the deflection coil. Thus, during the trace interval the current continues to rise linearly in the deflection coil and builds up a substantial magnetic field by the time the beam reaches the right-hand side of the screen. At this point the supporting plate-current rise is removed and consequently the strong magnetic field surrounding the deflection coil collapses quickly as it attempts to sustain the original current flow. The rapid fall of the magnetic field causes a quick field reshaping or retrace interval.

In the generation of a sawtooth of current in the deflection coil (Fig. 173) the coil reactance and resistance as well as external resistance play a part in determing what type of applied voltage is necessary to produce a sawtooth of current. When the reactance or inductance of the circuit dominates the resistance, a sawtooth of current is obtained in the coil by applying a squared wave —the current rising in the coil for the length of time the pulse is applied. If no resistance were present in the circuit, as a theoretical possibility, the current in the coil would rise absolutely linearly. With some resistance present the current rises exponentially although the time constant of L and R can be made long enough to have the current rise in a linear manner for the duration of the applied signal.

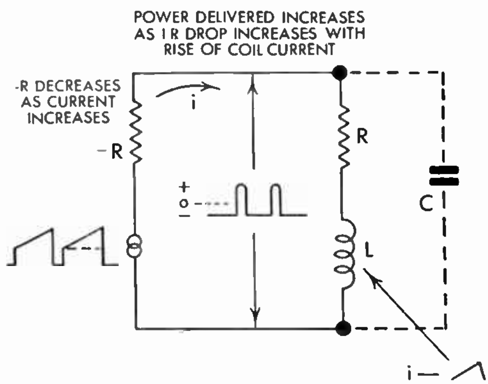


FIG. 175 Equivalent Circuit of Deflection System

If resistance dominates the inductance and inductive reactance, as in the second drawing, a sawtooth of voltage is necessary to produce a sawtooth of current in the coil because the series branch is now resistive insofar as the phase and amplitude of the applied voltage. When resistance and reactance are both present in the circuit, a combination of squared and sawtooth applied voltage is necessary if a sawtooth of current is to appear in the coil. The squared-wave component of applied modified waveform starts initial rise of current in the coil, and as current builds up the voltage drop across the resistive component increases. It is necessary that the applied voltage also rise in a sawtooth manner to overcome the voltage loss across the resistor as current through it rises. Across the inductance alone, therefore, an essentially squared-wave is present, causing a linear rise of current through the coil.

In all deflection systems, resistance and inductance are both present. Usually in a horizontal deflection system the inductance dominates, placing an inductive load on the output amplifier; in the vertical deflection system the resistive component dominates and a sawtooth waveform with some squared-wave modification is necessary to produce a sawtooth of current through the vertical deflection coil.

In an actual deflection system (Fig. 175) the output tube itself must present a regulated resistance effect to keep a sawtooth of current in an output circuit which is the combination of inductance and resistance. Thus, upon application of a positive increment of voltage to grid of deflection-output tube which acts as a generator, the increase in current produces a drop in plate voltage in the output circuit which causes the current in the coil to rise. However, as current flow increases and passes through resistance, a larger IR drop exists, which subtracts from the voltage present across the inductance and, therefore, prevents the linear rise in current in the coil. However, if the applied signal voltage to the grid of the output tube increases when this occurs, the plate current also increases and the additional plate-current flow counteracts the *IR* drop, keeping an essentially constant voltage across the deflection system which produces a sawtooth in the coil. It is evident, therefore, that the tube actually has a negative resistance effect because its value effectively decreases as the current flow increases, counteracting the *IR* drop in the resistive component of the load (resistor *R* and inductor *L* represent the reflected load in the primary of the output transformer from the deflection coil).

RETRACE TIMING

We have considered the rise of current in the deflection coil as the deflection trace occurs, and when the beam reaches the right-hand side of the screen the field must rapidly reshape to the left side to begin a new trace. Two methods are used to cause a fast retrace period. In the first method the trailing side of the sawtooth is itself fast enough to produce a quick retrace. In the second method, which is dominantly used in horizontal deflection systems, a transient oscillation is used to produce the very fast horizontal retrace.

A deflection-coil system also contains distributed capacity as well as reactance and inductance, resonating at some high frequency. At the moment the supporting voltage is removed from the secondary of the output transformer the field surrounding the deflection coil collapses quickly, producing an oscillation or a series of oscillations at the resonant frequency of the output circuit. One-half cycle of this high-frequency oscillation retraces the deflection field because it occurs so quickly. To have it occur rapidly, of course, it is necessary that the resonant frequency of the output circuit be relatively high and the period of the oscillating cycle short in comparison to the desired retrace time (time required for the deflection fields to reshape themselves to have beam start an active trace from left-hand side of the screen at instant picture tube is unblanked).

The importance of the time constant of a deflection system is demonstrated in Fig. 176. It is, of course, preferable to have a long time constant with respect to the interval during which the trace occurs, because this means there will be a linear rise of current in the coil. It means the inductance and the inductive reactance must be high and the resistance low, again showing the need for a low-resistance deflection system but one with an appreciable inductance. It must be pointed out again, however, that the distributed capacity of the deflection system resonates with the inductance to produce a half cycle of transient oscillation which rapidly retraces the field to the left side of the screen. If the inductance is too high and the distributed capacity correspondingly large, the resonant frequency will be low and, therefore, the flyback time considerably increased. Thus, a balance between the need for an appreciable inductance for the trace and the need for a small value for a fast retrace must be attained. In the case of the horizontal sweep, which is at a relatively high frequency, the deflection system presents almost a pure inductance during the trace and a resonant circuit during the retrace. The inductance must be held down to a value at which its resonant frequency with distributed capacity produce a transient oscillation, one-half period of which is no greater than the flyback time necessary to obtain complete reshaping of the deflection field.

If the time constant of the deflection system is short in comparison to the period of the trace, as is very likely to occur for the low-frequency vertical sawtooth, the current rise in the coil is not linear unless proper modification

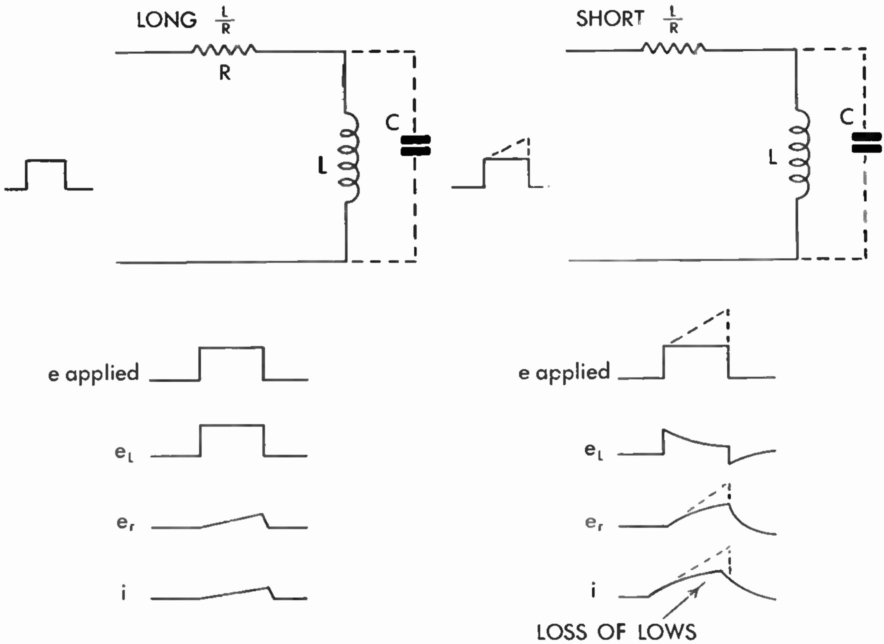


FIG. 176 Time Constant of *R-L* Combination and Its Effect on Series Current

is made to the applied voltage. At the low vertical frequencies the inductive reactance of the coil is low (in spite of the larger value of inductance used) in comparison to the resistive component in the circuit (large winding). Consequently, the time constant is short because the value of inductance may not be increased to a value which would seriously slow down the retrace time. With this short time-constant deflection system it is necessary to modify the squared pulse supplied as described previously to supply additional current flow through the coil at the point where the rise of current is no longer linear. To a smaller extent this expedient must also be used in the horizontal deflection system.

98. *Damping Systems*

Although the fast retrace is a function of the high-frequency transient, this transient can become a source of trouble. If the transient oscillation is allowed to persist after the first useful half-cycle it will appear in the trace and destroy its linearity. A damping system removes this possibility.

The two basic damping methods (Fig. 177) are resistive damping, for which damping occurs during the trace and retrace, and damping by means of a conducting tube which is active during the sawtooth trace but inactive during the retrace. Resistive damping is generally used in the vertical deflection system of the television receiver because of its simplicity and perhaps because of the substantial resistive component already present, also the fact that the retrace is not extremely rapid. Shunting the deflection coils of the vertical

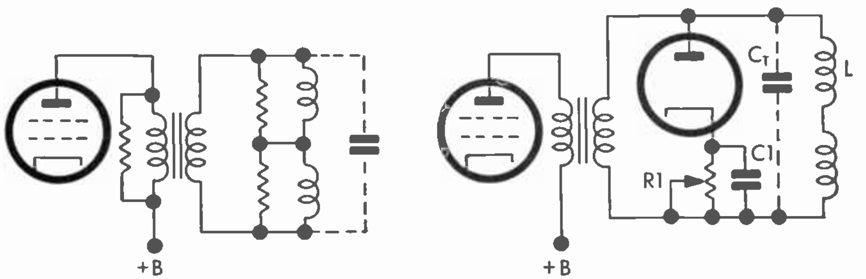


FIG. 177 Transient Voltage Damping Systems

deflection system with a low-value resistance renders the circuit aperiodic, preventing the reversal of current in the output current during the retrace period of the sawtooth. Instead, the current variation through the coil follows the sawtooth wave present in the plate circuit of the output tube during its trace and retrace intervals. The deflection system, therefore, presents a resistive load to the output tube, and the output transformer matches the resistive load of the secondary to the plate load resistance of the tube.

Retrace time of the vertical deflection system is generally dependent on the speed of the retrace portion of the sawtooth applied to the grid of the vertical output tube. If this waveform is properly modified a sawtooth of current will pass through the coil, and if the inductance of the coil is not excessively large or damping resistance too small the retrace time will not be adversely affected. However, if the inductance of the coil is too large or damping resistance too small, the long time constant of the system during the vertical retrace period will appreciably lengthen the retrace time.

In the horizontal deflection system where substantial power is necessary and the retrace time is exceedingly fast, a vacuum-tube damping system must be used. In this system the diode conducts only during the trace portion of the horizontal sweep to control the current flowing through the deflection coil. When the plate voltage of the sweep output tube swings sharply positive (horizontal output tube is cut off during the retrace period) the field about the deflection coils collapses and the secondary goes into oscillation at the resonant frequency of the inductance and distributed capacity. First negative half-cycle of this oscillation, which occurs very quickly, constitutes the *retrace period* of the deflection system. If no damping tube were present to prevent the positive sweep of this transient oscillation, a series of transient oscillations would appear on the *trace part* of the horizontal sweep. This would cause the beam to sweep back and forth a number of times at the left-hand side of the screen before it started its normal trace across the screen, and a distorted and washed-out pattern would appear on the left side. The damping tube prevents the development of the positive alternation of the very first cycle of the transient oscillation and, therefore, eliminates succeeding cycles by placing a load on the deflection coil and, therefore, a path for the transient current to flow which dissipates the energy in the form of *IR* drops.

CYCLE OF EVENTS

The action of the deflection and diode damping system is as follows:

1. When a sawtooth voltage is applied to the grid of the output tube the plate current begins to rise linearly, and this constant rate of current rise in the primary of the output transformer (during the trace portion of the horizontal sweep almost a pure inductance is reflected from the secondary) produces a constant voltage across the secondary and, therefore, a constant rise of current through the deflection coil.

2. When the grid voltage swings rapidly negative, in most cases driving the horizontal output tube beyond cutoff, the voltage which appears across secondary is removed.

3. The energy stored in the magnetic field during build-up of current is now released and the inductor attempts to keep the same amount of current flowing in the same direction.

4. The resistance of the circuit is now higher because the output tube is cut off and the positive voltage has been removed from the diode plate—diode is now nonconducting. Rapid flow of energy out of deflection coil develops a peak negative charge on the distributed capacity, which rises to peak value in accordance with the resonant frequency of *L* and *C*. The peak voltage is exceedingly high because of the rapid rate of current change.

5. This rapid decrease of current as the capacitor is charged constitutes the current flow in the deflection coil which reshapes the magnetic field. The time required to do so is set by the resonant frequency of *L* and *C* (distributed secondary capacity) and represents the retrace interval.

6. Once the capacitor has charged to full negative value the current has fallen to zero; the capacitor itself now begins to discharge through the coil to seek an equilibrium value. The inductor at first opposes the current rise, but once it is started its inertia keeps the current flowing in the same direction and a positive charge builds up on the capacitor. This constitutes oscillations (positive half cycle) which would continue until the energy is dissipated in the *IR* drop of the circuit.

7. If no loading were present the *IR* drop each cycle would be small and a

series of damped oscillations would be formed. However, when the charge polarity on the capacitor reverses the diode conducts and a low resistive path is presented to the electrons coming off the capacitor which quickly suppresses the positive alternation and any succeeding oscillations (substantial current is moved without any substantial build-up of voltage).

8. As this action occurs after the capacitor has obtained its maximum negative charge it is occurring during the active trace portion of the deflection sweep and, therefore, the conducting diode effectively removes transient oscillations from the sweep-current waveform. It is to be noted that this current flow in the diode is now in the direction to produce a current rise in the deflection coil which will assist in tracing the beam from left to right.

9. The current from the stored energy and from the conducting tube (sawtooth voltage again rising on the grid of the horizontal output tube) starts a new trace or line (produces the linear current rise in the deflection coil).

10. The inductance of the deflection coil and resistance of conducting diode present a long time constant which controls the rate of current rise in the deflection coil. Thus, the linearity of the current sawtooth can be controlled by changing the value of the cathode resistance of the damping diode and, therefore, the time constant and rate of current rise. The presence of the damping diode and its activity during the trace portion of the horizontal sweep reduces to some extent the peak power required from the deflection output tube.

99. *Magnetic Deflection System Linearity*

The linearity of the current sawtooth in a magnetic deflection system is dependent on a number of factors—frequency response, driving grid waveform, damping system, and inductance-resistance ratios in the output system. It is the resistive components (Fig. 178) of the deflection system which, if not considered, would cause nonlinearity of the sawtooth current in the deflection coil. In the low-frequency vertical sweep system the series resistance causes nonlinearity of the sawtooth waveform, because of the lowered reactance of the inductor to the lower frequency components of the sawtooth wave, causing an integrated-appearing sawtooth current to flow in the deflection coil. Actually, the period of the low-frequency components of the sawtooth are comparable with the time constant of coil inductance and the series resistance. Therefore, the voltage is not absolutely constant across the inductor (differentiated) and the rise of current in the coil is integrated.

It is apparent, therefore, that the size of the inductor should be relatively high, but cannot be too high because of the limitation it imposes on the retrace interval, as discussed earlier. The series resistance, of course, should be held at a minimum, and it is customary practice to use a triode as the vertical deflection output tube because of its lower internal resistance.

In the horizontal deflection amplifier, which must deflect the beam at a much faster rate, the inductive reactance of the coil can be made relatively high with just a small value of inductance. During the greater portion of the trace of the horizontal interval the deflection system presents almost an entirely pure inductive load to the output tube. However, at exceedingly high-frequency harmonics of the sawtooth wave the inductive reactance of the coil becomes comparable to the value of the plate resistance of the tube and, therefore, the higher frequency components of the sawtooth current are attenuated.

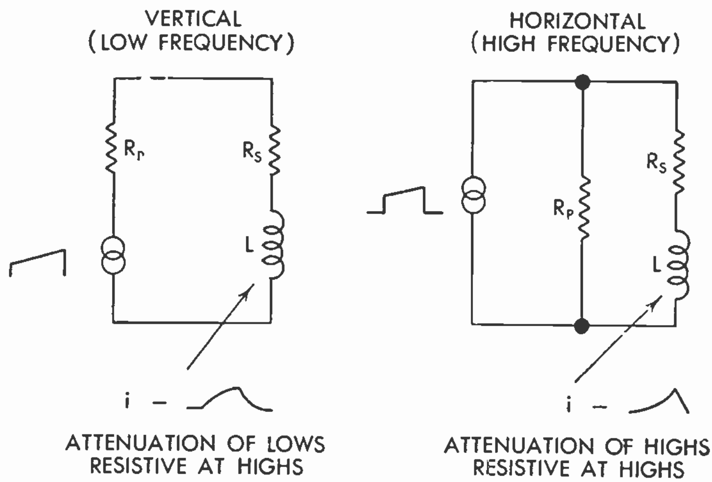


FIG. 178 Distortion of Horizontal and Vertical Sweep Currents

It is interesting to observe that as far as a sawtooth voltage is concerned the higher frequencies of that voltage affect the apex and retrace of the sawtooth wave, and if the circuit is deficient in high frequencies the apex will round off and retrace will integrate, causing a sag in the trace. When the circuit becomes deficient in low-frequency components, the retrace is passed readily; but during the trace there is some differentiation and the trace is rounded off.

So far as a sawtooth current is concerned, the absence of high-frequency response also causes the trace of the sawtooth to sag and current to build up at an accelerated pace. A deficiency in low-frequency response actually causes the sawtooth current rise to decelerate and round off the trace. Thus, the loss of low-frequency components in the vertical magnetic output system causes the beam to slow down as it approaches the bottom of the screen and, in the case of the horizontal deflection system, in which there is most likely to be a deficit of high-frequency components, the actual beam velocity will be greater on the right-hand side of the screen.

Inasmuch as the nonlinearity of the sawtooth current is caused by defective frequency response, the defect can often be compensated for by proper frequency compensating circuits. Thus, in the case of the horizontal output circuit in which there is a deficit of high-frequency components, it is customary to leave the cathode-bias circuit not by-passed so far as low-frequency components are concerned (producing degeneration at low frequencies), while the higher frequency components are adequately filtered with a small-value cathode capacitor. Therefore, the high-frequency components are held at the same level as the low-frequency components in the deflection coil circuit because of the added degeneration of the lower frequency components before they reach the output circuits. Such a system is shown in Fig. 179, along with another system in the second drawing for compensating for a deficit in low-frequency components as in the vertical output system. In this system a small capacitor is

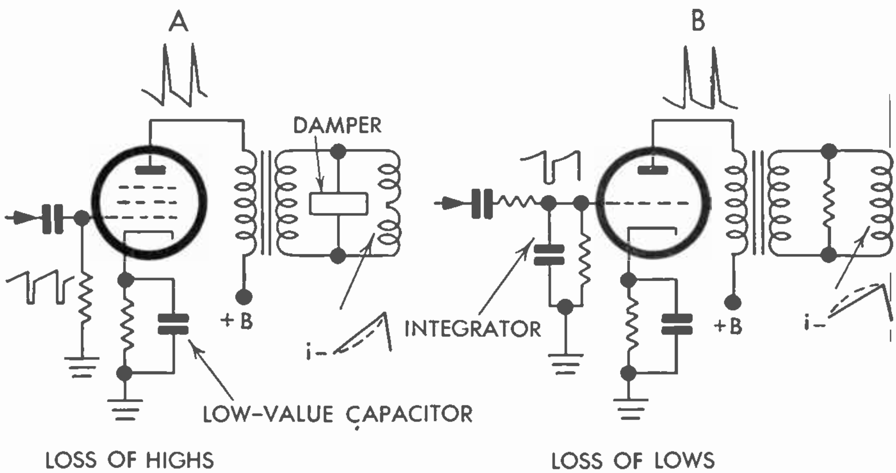


FIG. 179 Frequency Response Correction of Nonlinearity

shunted from grid to ground of the vertical output tube. Therefore, the higher frequency components of the applied sawtooth are attenuated to compensate for the subsequent attenuation of the lower frequency components in the deflection output circuits, producing a linear current sawtooth in the deflection coil.

Still another linearity control system is shown in Fig. 180, in which the linearity is controlled by properly setting the bias on the sweep amplifier output tube to a point on the transfer characteristic which is nonlinear. This nonlinearity in the transfer characteristic is used to compensate for the opposite nonlinearity in the applied sawtooth. For example, if the sawtooth were rounded off as it is applied to the grid of the sweep amplifier tube, and linearity control adjusted to place the bias at the point indicated in the drawing, the transfer characteristic is also nonlinear. It is nonlinear at voltages more negative than this bias point and would, therefore, compensate for the nonlinearity of the applied sawtooth at voltage levels less than this bias point, producing an essentially linear plate-current change and output sawtooth voltage waveform. With the same linearity method it is also possible to bias the tube at a less negative bias point which would operate the tube on the nonlinear portion of the curve which approaches the saturation point of the tube. Biasing the output tube at this point would correct for an opposite slope of the applied sawtooth waveform. Under most circumstances, however, the actual distortion of the sawtooth occurs in the deflection output system, and it is necessary to overcompensate the highs or lows in deflection amplifier tube to take care of a loss of the same frequency component in the output circuit. Thus, although it might appear that the sawtooth applied to the grid of the sweep amplifier is linear and, therefore, no linearity control system is necessary, it must be remembered that the linearity system may be correcting for a defect before it actually occurs and actual grid waveforms should be nonlinear.

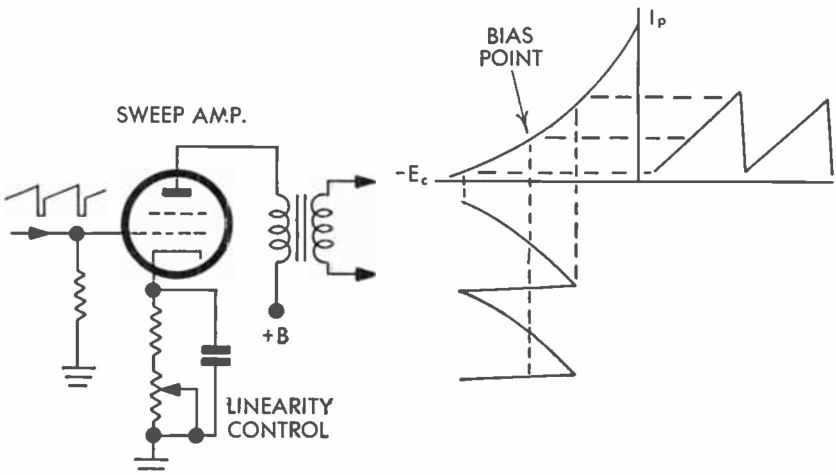


FIG. 180 Linearity Control by Biasing on Nonlinear Portion of Curve

100. *Horizontal Sweep Amplifiers*

Although the same basic features and requirements previously discussed apply to the horizontal sweep amplifier of the modern television receiver, a number of added features and functions are associated with the horizontal sweep amplifier used in the modern television receiver. A typical commercial horizontal sweep system (Fig. 181) consists of beam power output driver tube, a deflection output system with a triode damper, and a voltage supply which rectifies the deflection coil transient voltage and uses it as a source of high voltage for the picture tube. The operating cycle of such a sweep amplifier is as follows:

1. A sawtooth voltage with a sharp negative peak is applied to the grid of the deflection output tube, negative peak formed by a peaking resistor in the sweep oscillator. The trace interval of the horizontal sweep is formed during the linear rise of the sawtooth wave. However, there is no flow of plate current until the sawtooth wave reaches about mid-point of its rise.

2. An inductive load is presented to the output tube during the trace interval of the beam, left to right across the screen. The output transformer just acts as an inductance transformer, presenting the proper load to the tube from the deflection coils.

3. A constant rise of plate current in the inductive load develops a squared voltage (positive squared wave) across the secondary of the output transformer. This squared wave causes a linear increase of current in the deflection coils. However, as the deflection current builds up the *IR* drop (resistance of

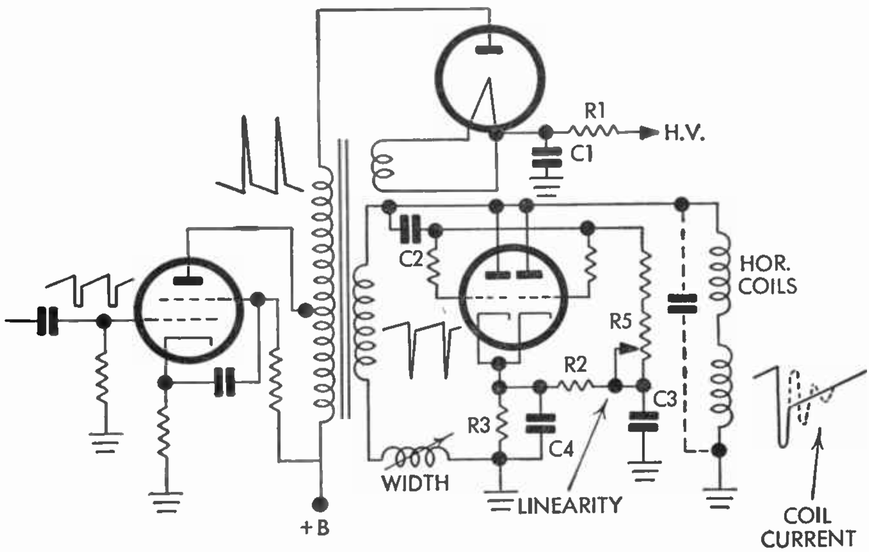


FIG. 181 Horizontal Deflection Amplifier

secondary of transformer, resistance of deflection coils, and damping resistance) prevents a true squared wave from appearing across the deflection coils, and it is necessary that the wave appearing across the secondary have a slight upward taper to keep the deflection current rise linear.

4. The rise of deflection-coil current continues until the grid waveform is driven sharply negative and cuts off the driver tube. This means there is no supporting voltage to sustain the build-up of current in the deflection coil and the stored energy within the coil is now released. Likewise, the removal of the voltage causes the damping tube to stop conducting, and the resistive component presented to the energy now released from the deflection coil is high, producing a short time-constant *RL* circuit.

5. Although the inductor attempts to keep the same current flowing for the instant, it rapidly falls to a low value through the high-resistance load charging the distributed capacity of the secondary circuit which, along with the inductance of the coil, resonates at approximately 75 kilocycles.

This very rapid release of energy in the deflection coil (collapse of the magnetic field) develops a very sharp negative voltage of 1,000 volts or more (fast current change through the high reactance of the deflection coil at 75 kilocycles) and represents the negative alternation of the transient oscillations developed when the load is removed from the secondary, and stored energy in the deflection coil excites it into oscillation. Although only one alternation of this transient voltage is permitted to occur, this sharp negative alternation rapidly reshapes the horizontal deflection field to cause the beam to start an active scan on the left-hand side of the screen. The resonant frequency of the inductance and the distributed circuit capacity must be made sufficiently high to have the negative alternation occur in a time interval which is shorter than the FCC-assigned horizontal retrace time. Thus, the half period of the transient oscillation must be less than the horizontal retrace time of approximately 8½ microseconds (sync pulse plus back porch).

6. If the transient oscillations were not suppressed after the first alternation, a series of damped oscillations would ride on the deflection-current sawtooth rise (dotted line of current waveform drawing in Fig. 181) and the beam would weave left and right as it gradually moves across the screen. It is the function of the damping tube to suppress the oscillations after the first sharp negative alternation used to retrace the beam. Thus, as the oscillation starts on its first positive alternation, the damping tube conducts, shunting a low resistance and heavy load across the deflection coils. Thus, a long time-constant circuit is presented, and the energy is released from the coil at a much slower rate.

7. This slower release of the stored energy into the damping resistance forms a part of the active trace of the picture beam because the release of current is linear (a long time constant) and in the proper direction to form the start of the magnetic field which causes the beam to scan left to right. Therefore, the stored power in the deflection coil at the end of the previous trace is used in the formation of the start of a new trace, conservation of power reducing the power requirements on the driver tube. The flow of current is, of course, linear over the initial part of the release but gradually tapers off exponentially. At that point the driving tube begins to conduct and supports the rise of current in the deflection coil. Thus, the active trace of the picture-tube beam is contributed dominantly first by the release of stored energy in the coils and then by the conduction of the output tube over the second half. This method of beam deflection is aptly called *reactance scanning*.

HIGH-VOLTAGE SYSTEM

The first collapse of the magnetic field about the deflection coil at the end of the active trace (when the pentode is driven to cutoff) causes a high-voltage negative transient to be developed in the secondary of the deflection system. At the instant it is developed, the damping tube is cut off and so is the output driver, and there is only a light load across the oscillating circuit. The peak negative transient developed in the secondary appears as an amplified positive spiked voltage across the primary. The plate of the output driver is tapped onto the primary at some point down the winding, at which there will be the proper match between deflection coil and driver-tube output. The full primary of the transformer amplifies the sharp negative transient in the secondary to an exceedingly high value of 7,000 to 12,000 volts, which is applied to the plate of the high-voltage rectifier. These sharp high-voltage bursts (one occurring each 63½ microseconds) draw bursts of diode current, which charge capacitor *C1* to almost the peak value of the positive transient.

If time constant of *C1, R1,* and resistive load presented by the picture tube is sufficiently high, there will be only an insignificant discharge off capacitor *C1* during the interval between positive transients. Each arriving transient pulse draws sufficient additional current to keep the charge on capacitor *C1* at its peak value.

101. *Triode and Diode Damping*

In the diode damper (Fig. 177) the current flow through resistor *R1* charges capacitor *C1* during the interval that the field about the deflection coil is collapsing and discharging through the conducting tube. Tube conduction or resistance is controlled by the bias placed on it by the voltage charge on capacitor *C1*; time constant of *R1* and *C1* is sufficiently long to keep a charge there between trace intervals. Thus, the linearity of the sawtooth current is controlled to a certain extent by the setting of *R1*. It does so by controlling the resistance of the current path. If the resistive path is too high the time constant becomes too short and the positive alternation of the transient is not completely suppressed. If the resistance path is too low, release of the energy from the deflection coil is slowed down. If the build-up of deflection current is slow, it causes nonlinearity in a system which depends on utilization of the power storage in the deflection system as well as excitation from the driver tube congruently. An advantage of the diode damper is its low resistance, and consequent minimum *IR* loss, although the very first part of the trace is not absolutely linear because of incomplete compression of the transient oscillation. If damping resistance is made low enough to completely suppress the oscillation, the build-up of current during the trace is retarded and linearity at the center is affected at the point where the driver tube starts to contribute current.

When a triode is used as a damper (Fig. 181) the linearity of the sawtooth current can be made almost perfect, although the higher resistance of the triode causes additional loss of stored energy across its resistance. Damper-tube bias is developed across capacitor *C4* by flow of current through *R3*. Resistor *R2* and capacitor *C3* compensate for the normal exponential current decline when the energy stored in the deflection coil is released, generating the first part of the active trace. Discharge current is linear at the start of the trace, but as the sawtooth current builds up toward the mid-point the rise becomes exponential. The resistor-capacitor combination in the grid circuit compensates for this exponential rise of current by applying to the grid of the damper tube an oppositely shaped voltage wave (causes an intentional deficit in the high-frequency component of the voltage waveform, causing it to sag), which means the plate resistance of the tube is changing at the same rate at which the current flow is becoming exponential, maintaining an essentially constant current flow out of the coil. The extent of this conductive control can be varied with resistor *R5*, the setting of which also controls the extent of the damping over the initial portion of the sweep trace.

The higher frequency transients are fed through low-value capacitor *C2* to the grid of the damper tube and develop in the plate circuit an opposing waveform which cancels out by degeneration the transient voltages present. Extent of this cancellation is also controlled by the setting of resistor *R5*, which controls the resistive component of the reactance-resistor divider formed by *C2* and the grid resistors.

In the more elaborate horizontal sweep systems (third drawing of Fig. 182) a combination of diode and triode damper tubes is used. Diode section is used to maintain an efficient output-deflection system which minimizes loss of the stored energy in the deflection coils at the end of the trace intervals. The diode damper when it is used efficiently also causes some nonlinearity of the trace during the initial portion. However, a triode damper has been inserted as a means of controlling the linearity of the sawtooth current and, therefore, the good features of both systems are used to advantage in the deflection system.

102. *Voltage-Booster System*

Many receivers are designed to re-employ the energy released into the damping system. In many cases this produces a potential of 50 to 100 volts, which can be placed in series with the driver tube plate voltage to supply additional voltage and power to the output driver. Fig. 182 shows a number of methods to derive this added energy and apply it to the driver tube, thereby adding additional power to the supply-voltage source. All of the methods shown derive energy directly from the damping tube load resistor whether it be a diode or a triode damper. In the circuit of the first drawing the voltage is taken off the *R-C* combination in the cathode circuit of the diode damper and applied to the lower portion of the primary winding of the output transformer, +B supply voltage for the driver tube being applied via the secondary winding of the output transformer.

In the second drawing a triode damper is again used and energy is developed across an *L-C* filter and applied to lower portion of primary winding of the output transformer. The resistance of the driver tube itself serves as the load resistance for the damper. The energy stored in the capacitors is delivered to the output driver when it conducts.

The third schematic shows a method using combination diode and triode damper, the voltage boost being obtained from the diode damper. The triode damper is in the circuit just as a means of controlling the linearity of the sawtooth current and, therefore, the linearity of the picture on the television screen.

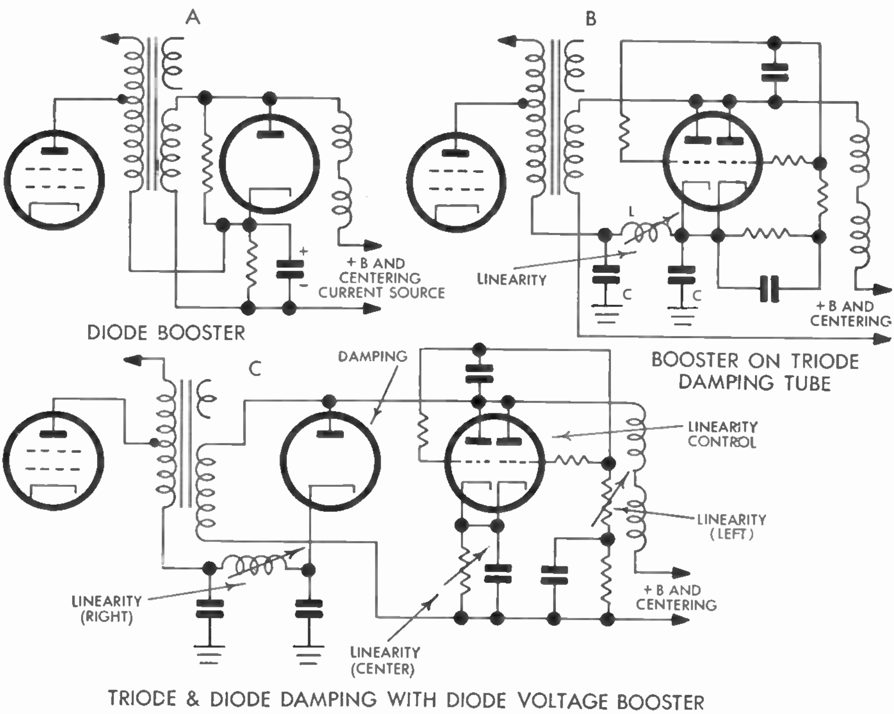


FIG. 182 Voltage Booster Systems

103. *Linearity Control Methods for Horizontal Sweep*

In a magnetic deflection amplifier, a number of adjustments affect the linearity of the reproduced picture. The linearity control methods affect the linearity of the left side, central portion, and right side of the reproduced image. The linearity control of the simple diode damper (shown in Fig. 177) determines the linearity of the initial portion of the trace interval or the left-hand side of the screen by controlling the extent to which the transients are suppressed and the rate at which stored energy is released from coil. Television-screen linearity is controlled by the inductor linearity controls in drawings B and C of Fig. 182. They control the shape of the discharge current off the capacitors as they supply energy to the output driver. The shape of the plate-voltage waveform is altered in a manner which will compensate exactly for the exponential current waveform in the deflection coil near the end of the discharge interval (at the mid-point of the sweep interval. when deflection current is being contributed by the energy stored in the deflection system and by the now-conducting driver-output tube).

The linearity of the right-hand side of the screen in a magnetic deflection system is controlled by the shape of the driving-grid waveform of the output tube. By changing the amplitude of the peaking voltage it is possible to drive the sawtooth portion of the waveform into the compressed portion (near saturation) of the tube characteristic. Consequently, the top portion of the sawtooth waveform will begin to fold over, producing a sawtooth waveform which appears rounded. The ratio of sawtooth voltage to pulse can be decreased to permit conduction of the driver tube over a longer portion of the trace interval, affecting the linearity at the central and right-hand portion of the screen. The amplitude of this sawtooth, of course, determines *IR* drop correction in the deflection-coil circuit and, therefore, linearity.

When a triode damper is used, the linearity of the left and central portions of the screen is again controlled by the cathode *R-C* combination. The linearity at the left side and the extent to which the transients are completely removed by cancellation is controlled by a linearity control placed in the grid circuit of the damping tube. A second *R-C* combination in this same grid circuit is often used to control the linearity at the central portion of the screen, where the discharge current from the deflection coil becomes exponential (Fig. 182). Again, the right-hand side of the screen and its linearity is under control of the grid-excitation waveform to the driver output tube and inductor of voltage booster circuit.

104. *Complete Magnetic Sweep Amplifier*

A complete magnetic deflection system for a picture tube (Fig. 183) consists of a vertical and horizontal deflection amplifier. The vertical amplifier generally consists of a triode or low-resistance pentode which feeds an essentially resistive load (presence of loading resistor across deflection coil, low resistance of tube, and appreciable resistance of the large winding necessary for the low-frequency vertical sweep). In addition to the sawtooth current developed in the deflection coils, as discussed in the preceding paragraph, a d-c component of current must also flow to the deflection coil to permit proper centering of the entire pattern on the screen both vertically and horizontally. This d-c component of current is generally tapped off from the low-voltage bleeder system at a very low potential point to obtain proper d-c current through the coils to center the scanning raster. The centering system, of course, must be properly filtered to prevent the sawtooth and transient voltages from feeding back into the power supply.

The vertical linearity control is in the cathode circuit of the output tube. Linearity of the sawtooth current is corrected by setting the bias at the proper point on the transfer characteristic of the tube to compensate for the resistive components present in the output load. An inductor-type linearity control is used in the horizontal deflection system to control the linearity from the point where the deflection coil and output driver are both contributing energy to the build-up of current in the deflection coil. A width control is also associated with the secondary winding of the output transformer of the horizontal deflection system to permit variation of the secondary inductance and, therefore,

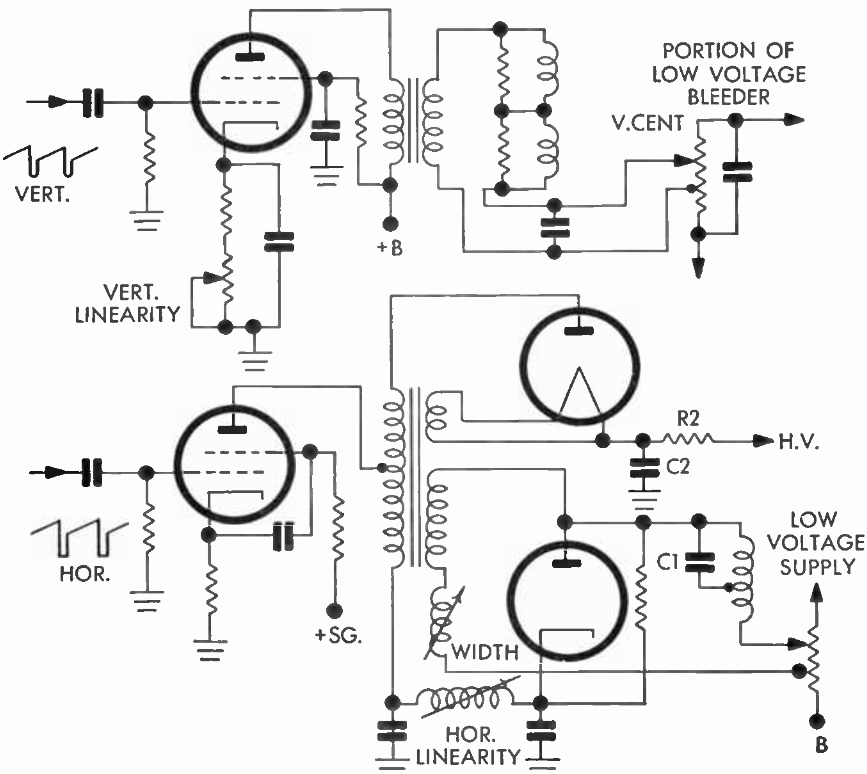


Fig. 183 Magnetic Deflection Amplifiers

amplitude of current which passes through the deflection coil (two parallel branches). The width of the vertical sweep is set by the amplitude of the driving voltage applied to the grid of the vertical output tube. However, the amplitude of the waveform applied to the grid of the horizontal output tube has more effect on the linearity of the horizontal sweep.

It is also necessary to shunt a portion of the horizontal deflection coil with capacitor *C1* to balance distributed capacity to ground from top and lower sections of the horizontal coil. Doing so permits generation of a uniform magnetic field at both ends of the deflection coil. The horizontal output circuit contains a voltage-booster system and a high-voltage power supply. The heater current for the high-voltage rectifier is also obtained from the horizontal output coil by absorption, into a few-turn pickup coil, of current flowing in the other windings. A simple *R-C* filter is used to filter the high-voltage transients which are applied to the plate of the high-voltage rectifier.

105. *Deflection for Larger Picture Tubes*

To keep the length of the larger picture tubes to a reasonable size and a practical cabinet depth, it is necessary to deflect the scanning beam over a greater deflection angle. Typical deflection angles at present are 55 degrees, 70 degrees, and 90 degrees. The problem with the greater deflection angle is to obtain uniform focus over the entire surface of the picture-tube fluorescent screen. Focus uniformity is, of course, a function of the screen size and position, the coil or yoke position on the neck of the tube, and the angle to which the beam must be deflected. The older deflection coils, because of their positioning in the yoke and around the neck of the picture tube along with a uniform layer-wound winding, produced a non-uniform deflection field, Fig. 184. This non-uniformity of magnetic field causes distortion of the scanning beam as it is deflected at sharper angles to reach the extremities of the raster and passes into the regions where the magnetic lines of force are curved. An elliptical spot results, because in passing through the curved section the central portion of the beam is pulled on more strenuously than is the outer portion, with result that the beam is pulled into an elliptical shape. Consequently, the focus becomes progressively poorer toward the outer portion of the scanning raster. Of course, the larger the raster, the more objectionable this is, because the greater deflection angle and larger screen cause the defect to be more noticeable.

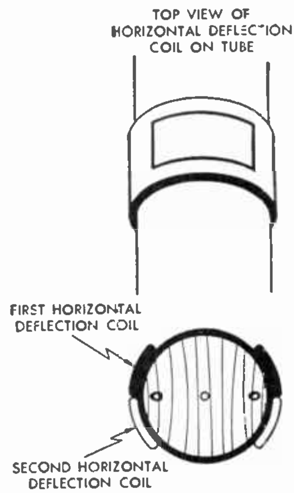


FIG. 184 Non-Uniform Magnetic Deflection Field

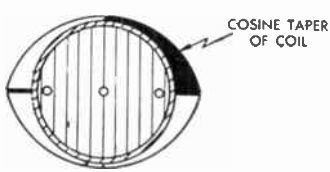


FIG. 185 Uniform Field of Cosine Yoke

This elliptical spot distortion can be minimized by special yoke construction. To obtain a uniform magnetic field, Fig. 185, it is possible to organize the windings of the deflection yoke in such a way that distribution is in proportion to the cosine of the angle between the deflection axis of the coil and the radial position of the windings on the neck of the tube. This special cosine yoke construction means that windings are concentrated in the section of the deflection coil where there was magnetic field curvature originally. Energy concentration at these points, or a greater energy concentration as the angle of deflection increases progressively, maintains a uniform magnetic field, and consequently, the scanning beam sees a uniform magnetic field regardless of deflection angle. A circular and in-focus spot is obtained over the entire surface of the raster. Cosine build up of deflection coil-winding is demonstrated in the illustration.

A second defect of the older deflection coil-winding construction is a barreling of the scanning raster, Fig. 186. Raster barreling is also caused by the non-uniform deflection field. Fortunately, this disturbance is corrected by another inherent defect of the deflection system, namely, a pincushion defect caused by the difference in path-lengths between the electron gun and the outer extremities of the raster, as compared to the distances covered when the inner sections of the raster are scanned. The greater the distance from the gun to the raster, the greater is the length over which the beam sweeps for a given deflection angle. Consequently, at the top and bottom of the raster (each of which is a greater distance from the gun than is the center of the raster) the scanning lines are longer. From the top of the raster, where the beam starts scanning, the distance from gun to raster becomes progressively smaller as the beam drops down toward the center of the raster, and therefore, each scanning line becomes progressively smaller in length. Scanning lines become progressively longer as the distance between the center and bottom of the raster is traveled. It is to be noted that the two types of raster distortion balance each other and produce a normally rectangular raster. Although the distortions cancel, the scanning beam with this type of yoke is still elliptical, causing the focusing at the sides and top and bottom of the raster to be not as good as at the center.

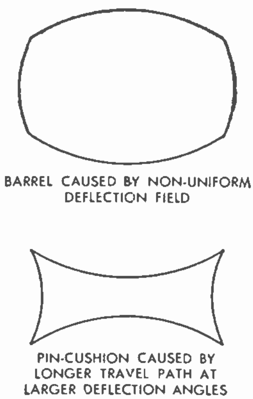


FIG. 186 Distortion of Deflection Field and Raster

The newer cosine yoke which has a uniform deflection field does not have any inherent barrel distortion. Consequently, on the larger picture tube the cosine yoke construction is likely to cause some pincushion disturbance. Thus, for many of the larger picture tubes, special pincushion magnets are mounted physically (two or four fixed magnets) on the periphery of the yoke or face plate and pull out the pincushion distortion. Still another plan for eliminating the pincushion defect is to modify further the deflection yoke-winding into a so-called cosine squared winding. This winding maintains a reasonably uniform magnetic field with some limited modification by which to correct the inherent pincushion distortion.

105a. *Commercial Sweep Systems*

A modern horizontal and vertical deflection system, Fig. 187, must be designed efficiently and contain a number of unusual circuits if it is to be able to obtain linear and efficient deflection for a large-screen picture tube. The horizontal drive pulse is supplied directly to the horizontal-sweep output tube from the preceding syncro-guide stage. To obtain a high amplitude drive voltage (modified sawtooth wave being developed across capacitor *C190*) a return is made to the plus-B boost supply through resistor *R200*. This connection permits the development of a high amplitude driving sawtooth with only a low-voltage supply source. The drive control capacitor *C181B* permits additional modifications of this waveform by means of which horizontal-output drive is adjusted to optimum value and sufficient high voltage and a linear picture from the center to the right-hand side of the screen are obtained. The horizontal-sweep output tube conducts on only the latter portion of the applied sweep waveform. A small-value cathode resistor is used as a protective device to prevent both excessive current flow by the horizontal-sweep output tube and possible damage to the deflection yoke in case of circuit failure. Additional protection is provided by the slow-blow fuse F101 in case of high current drain and possible shorts in the horizontal deflection system.

The high-amplitude pulse-sawtooth wave present at the plate circuit of the horizontal sweep tube drives the auto-transformer that develops suitable excitation both for the deflection yoke and the spike waveform necessary to derive the high anode voltage. The use of the auto-transformer instead of the conventional type is referred to as direct-drive method. In this system the horizontal deflection coils are tapped off of connections 5, 4, and 3; of a higher impedance than the earlier type of deflection coils, they reflect the proper impedance to the horizontal-output tube. The single-winding transformer can be made physically smaller than the older output transformer. Since it encounters less leakage losses, it operates with higher efficiency. Capacitive leakage and loading are minimized, thus making the very fast retrace possible, and still sufficient energy exists to make the long sweep from left to right across the screen of a large picture tube. When the deflection field collapses, the step-up auto-transformer winding develops a high-amplitude positive pulse that is applied to the plate of the *1B3* high-voltage rectifier. Rectified d-c voltage is present across capacitor *C197* and reaches a full charge of 13,200 volts for application to the anode of the 17-inch picture tube.

The transient oscillation is quickly suppressed as the negative sweep of the transient (after the useful positive pulse) is applied to the cathode of the damper tube, the damper tube then conducting and loading the deflection circuit (connection made at terminal *6*). The inertia and the long time constant of the damper circuit takes transient energy and stretches it out over a longer period of time; it may thus be used for the initial portion of the next sweep cycle, taking the beam approximately half-way across the screen for the next line. Damper conduction also develops a d-c voltage across capacitor *C196* which, when added to the plus-B voltage arriving via the horizontal linearity control *L107*, supplies a much higher B-supply voltage for both

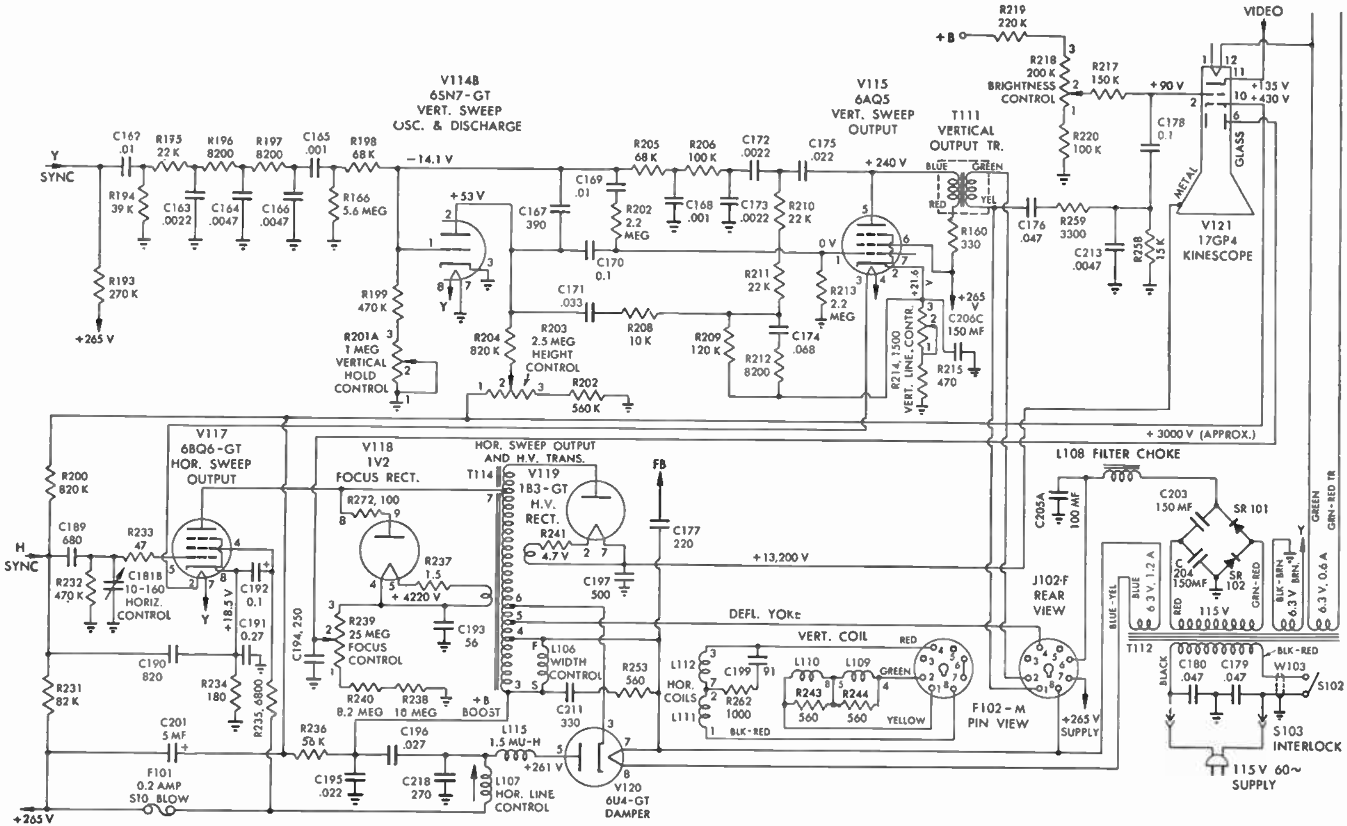


FIG. 187 RCA Sweep System

the horizontal-output tube, the horizontal sawtooth-generating circuit, and the vertical sweep circuit as well as for the accelerating grid of the picture tube. Diode conduction during the initial portion of the trace is also under control of the collapsing field of the horizontal-linearity control, and thus, the linearity of the initial and central portions of the trace can be regulated. To prevent radiation of the sweep transient and passage into the supply voltage, filter inductor *L115* and capacitor *C218* are employed. Feedback voltage is supplied to the horizontal-sync control system from terminal *4* via capacitor *C117*.

The width control *L106* is connected across a portion of the transformer and effectively shunts the horizontal-deflection coils. Its maximum inductance must be as large as possible in order to minimize loss of width at the maximum width adjustment. Likewise, its minimum inductance must drop low enough to effect the required maximum reduction in width. A blocking capacitor *C211* is incorporated to prevent a large or changing d-c current component in the horizontal coils that could cause picture de-centering. Horizontal- and vertical-deflection coils are illustrated—the latter are properly loaded with a balanced network of resistors while the horizontal coils, because of their resonant flyback function, are not shunted. However, a balancing capacitor *C199* is incorporated to balance both sides of the windings with respect to ground and to assure a more uniform magnetic field.

An important consideration in the mechanical planning of a horizontal-deflection—high-voltage generating system is lead dress. Any excessive capacitance to ground between points on the transformer increases the retrace time and reduces the high voltage obtained. In particular, the high-voltage rectifier circuit should be well isolated from ground, the plate lead should be kept as short as possible, and the tube so positioned that capacitance between its plate and chassis is at an absolute minimum. Likewise, the transformer leads should be well positioned to minimize stray coupling into the horizontal-amplifier grid circuit. Also, radiation from the horizontal-deflection system into other sections of the receiver can produce modulation of the picture signal and result in bar disturbances on the screen. Radiation from the horizontal-deflection system or feed-through into the power mains can cause interference to nearby broadcast receivers and other communication services.

Filament voltage for the high-voltage rectifier and for a special focus-voltage rectifier is obtained by coupling to the auto-transformer. The filament for the damper tube is derived from a special winding on the power transformer, which has been satisfactorily insulated to withstand the few thousand volts present at the cathode of the damper tube. A special focus-voltage rectifier *1V2* makes use of the positive pulse at the plate of the horizontal-sweep tube. A long time-constant bleeder network is employed across which the d-c high voltage is developed and retained. A focusing potentiometer permits adjustment of the focus voltage of approximately 3000 volts for application to the focusing anode of the electrostatic-focus, magnetic-deflection picture tube.

The vertical sync pulse is applied through a triple-section integrator network to the grid of the vertical-oscillator and sawtooth-forming circuit. It is to be noted that the vertical-generating and amplifier system functions as a multivibrator consisting of a triode and pentode section. The signal path from the plate of the triode to the grid of the pentode is through capacitor *C170*, and another feedback path exists from the plate of the pentode through capacitor *C175* and network back to the grid of the triode. This circuit functions as an assymetrical multivibrator with the triode conducting for a short period of time (retrace interval), while the pentode conducts over the much longer trace interval.

Frequency of the multivibrator is controlled by the time constant in the grid circuit of the triode. While this triode is non-conducting, capacitor *C171* is charged through resistor *R204* to plus-B and the height control. The sawtooth is properly spiked or modified by resistor *R208* and the feedback network to the cathode of the vertical-output tube. This network maintains control of the linearity of the picture at the bottom of the picture (the peak-current area where sweep is reaching for the bottom of the scanning raster). A positive pulse at the plate of the output pentode (the pulse exists during the cutoff period of the pentode simultaneous with vertical retrace) is coupled back to the grid of the triode and raises the grid potential prior to synchronization by the incoming vertical-sync pulse. This pulse is partially integrated to retain proper interlace by causing the vertical sync pulse to rise from the same level to the triggering potential of the triode. Linearity at the top of the picture is controlled by the special biasing network in the cathode of the vertical-output tube. This type of network, as mentioned previously, biases the tube off the central portion of the transfer curve by the proper amount needed to obtain best linearity.

The vertical-output transformer has a high primary-to-secondary turn ratio; therefore, a reasonably small current-change in the primary develops the necessary high current-change in the secondary for application to the vertical-deflection coils. This arrangement permits the use of the plus-B boost voltage and also the development of a strong deflection current with only a limited pentode plate current and, therefore, no serious current load on the boost-supply voltage source. Too severe a load on this supply can reduce anode potential and horizontal-deflection efficiency. A special network connected to the secondary of the vertical-output transformer supplies a negative pulse to the grid of the picture tube during the vertical retrace intervals. It cuts off the picture tube and prevents the appearance of retrace lines on the screen. The short time-constant network consisting of capacitor *C176* and resistors *R259* and *R258* differentiates the trace portion of the vertical cycle and prevents shading of the picture during the actual vertical motion of the scanning beam from the top to bottom of the screen. An integrating network consisting of resistor *R259* and capacitor *C213* integrates the transient oscillations that are present at the conclusion of the vertical retrace interval and prevents their appearance in the initial parts of the vertical trace.

Philco horizontal- and vertical-deflection systems are illustrated in Figs. 188 and 188a. A phase-detector horizontal sync control system is used with equal-amplitude, opposite-polarity horizontal sync pulses being applied to the top diode plate and to the bottom diode cathode. The d-c component at the output of the phase-comparison circuit is developed across capacitor *C803*

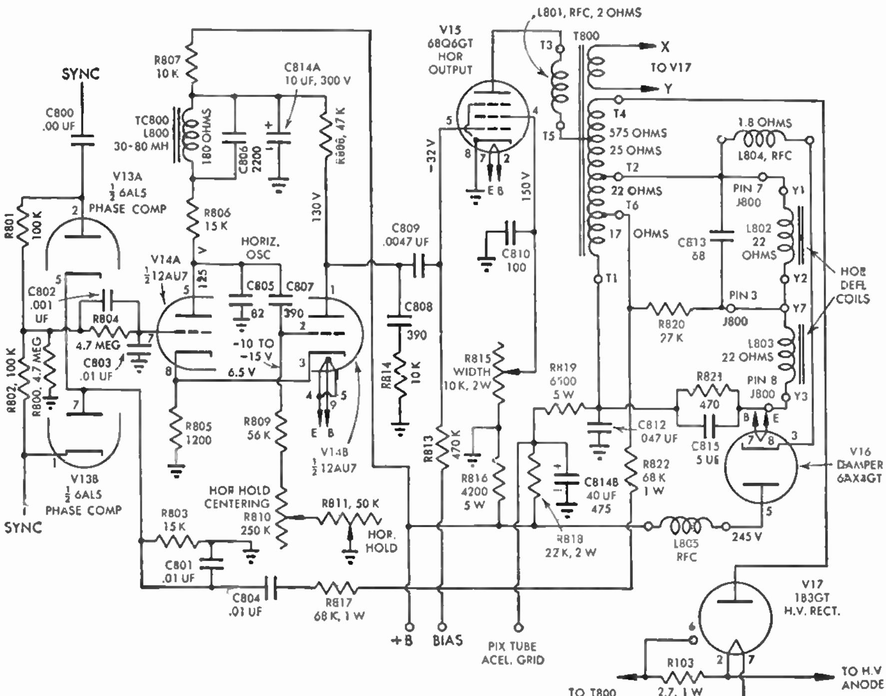


FIG. 188 Philco Horizontal Deflection

and on the grid of the multivibrator sawtooth generator. Sawtooth comparison voltage is developed from the pulse taken at tap *T6* on the horizontal-output transformer and coupled through isolating resistors *R822* and *R817* and capacitor *C804* to the top cathode and bottom plate of the comparison diode. The resistors aforementioned and capacitor *C801* function as networks which round the pulse derived from the deflection transformer to a sawtooth shape, serving as the comparison signal for the incoming opposite-polarity sync pulses. As discussed in the sync-control section the relative position of the arriving sync pulses with respect to the center of the sawtooth retrace determines the control charge developed at the grid of the multivibrator.

In the multivibrator the first section conducts for a longer interval of time than does the second section—the second section conducting only during the retrace interval and being cut off during trace. At this time the sawtooth capacitor *C808* is charged through resistor *R808* connected to plus-B. The sawtooth is properly spiked by resistor *R814.* A resonant circuit in the plate of the first triode stabilizes the multivibrator by adding a sine-wave component to the pulse developed at the plate of the first triode and coupled to the grid of the second triode. It is this pulse that drives the second triode into conduc-

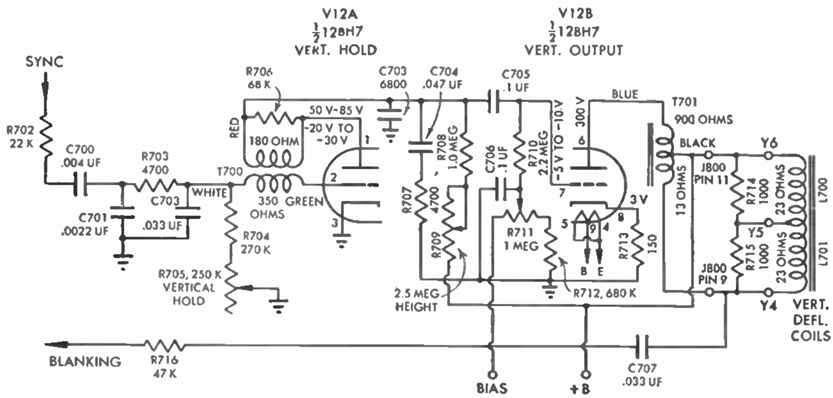


FIG. 188a Vertical Deflection System

tion and initiates or synchronizes the retrace interval. To obtain the very best synchronization this point of conduction should be held constant from line to line. Therefore, if a sharper rise of voltage is obtained near the time of conduction, synchronization of the circuit is far less influenced by any noises or impulse interference that might leak through to this grid. The sharp rise of voltage to the conduction point is contributed by the resonant sinewave developed in the stabilizing circuit.

The horizontal-output system also uses a direct-drive auto-transformer in much the same type of circuit as discussed previously. In this circuit, however, the width of the picture is controlled by adjustment of the screen voltage of the horizontal-output tube. In this method of width-control the adjustment of the screen potential also influences the boost voltage and therefore to a limited extent also influences vertical height. Thus, to a limited extent, when the picture is set for the proper aspect ratio, an adjustment of picture-width will also cause a corresponding change in picture-height, yet retain the proper aspect ratio.

The vertical-deflection system of the receiver employs a blocking-tube sawtooth generator and triode vertical-output tube. The vertical sync pulse is applied to the grid circuit of the blocking tube via a double-section integrator. The sawtooth voltage is developed across capacitor *C704* (during the cut-off time of the blocking-tube oscillator) which is charged through resistors *R708* and *R709* connected to plus B and spike resistor *R707*. Vertical linearity is controlled by a biasing system in the grid circuit of the vertical-output tube. An auto-transformer is used as the vertical output with the deflection coils tapped off at the proper ratio point. The retrace blanking pulse is also removed from the same tap.

QUESTIONS

1. Why must sweep sawtooth wave be linear?
2. How can an *R-C* combination be used to generate a sawtooth?
3. Explain operation of a blocking-tube sawtooth generator.
4. Explain operation of a multivibrator sawtooth generator.
5. What is the advantage of a discharge tube following a blocking tube?
6. How are frequency and amplitude of sawtooth waves controlled in the various generators?
7. Explain in detail synchronization of a sawtooth generator.
8. Explain synchronization process in an automatic sync system.
9. What are the effects of noise on synchronization?
10. Of what significance is the proper phase relation between sawtooth and received blanking?
11. What causes foldover?
12. What is function of peaking resistor?
13. Describe effects of poor frequency response on sawtooth wave.
14. Why is it possible to obtain a blanking bar on the screen when an automatic sync system is used?
15. Discuss characteristics of an electrostatic sweep amplifier.
16. Why must a special divider system be used between tube sections of horizontal electrostatic sweep amplifier?
17. Briefly describe linearity control methods for electrostatic sweep amplifier.
18. How is an image centered on an electrostatic tube and a magnetic deflection tube?
19. In what manner is the deflection waveform modified for magnetic deflection?
20. Why is it necessary to use a modified waveform?
21. Describe a typical vertical deflection amplifier.
22. Describe a typical horizontal deflection amplifier.
23. What is a deflection-voltage-boost system?
24. Explain action of horizontal drive control.
25. Discuss in detail a commercial sweep system.

Chapter 9

FM SOUND SYSTEM

106. *Generation of the FM Signal*

The FM signal is a *resultant wave* made up of a *center frequency* and a number of *sideband frequency pairs,* which has a constant amplitude but a frequency which deviates in accordance with the audio modulation. These sideband pairs or sideband components of modulation are: center frequency plus and minus the audio frequency, center frequency plus and minus the second harmonic of the audio frequency, center frequency plus and minus the third harmonic of the audio frequency, and so forth. The number of important sideband pairs present in the signal is dependent on the so-called "modulation index" which is determined by and equals the frequency deviation (amplitude of audio) divided by the audio frequency. The number of sideband components and their relative amplitudes, therefore, is dependent on the amplitude of the audio modulation (which sets the extent of frequency deviation) and the frequency of the audio. The frequency deviation increases with the amplitude of the audio modulation.

There are two basic methods of generating an FM signal. One is called *direct FM* because the audio modulation controls directly the frequency of the oscillator which generates the carrier frequency. A second method uses a separate sideband generator which generates sidebands in accordance with the audio modulation. These sidebands are then combined with the center frequency or carrier frequency to form the resultant frequency-modulated wave. In the direct FM system a very simple means to generate an FM wave is to have a capacitor across the oscillator-tuned circuit which varies in capacity as the audio sound pressure changes. Thus, as the capacity varies the frequency of the oscillator follows. In actual practice, however, we take advantage of the characteristics of a vacuum-tube circuit to generate the FM signal because with such a vacuum-tube circuit, properly designed, it is possible to present a capacitive or inductive load across the tuned circuit of the oscillator and, as the plate current varies in this tube, the amount of capacitive or inductive current introduced into the oscillator-tuned circuit is varied. Thus, as the grid signal on the so-called "reactance tube" varies with the audio modulation, the reactive current it introduces into the oscillator-tuned circuit also varies and causes the frequency of the oscillator to deviate accordingly.

In the generation of an amplitude-modulated wave we understand the audio frequencies cause the actual resultant wave amplitude to vary with the frequency and amplitude of the audio components. The resultant AM wave is therefore one that varies up and down in amplitude, while the resultant FM wave varies just in frequency and is constant in amplitude. In spite of

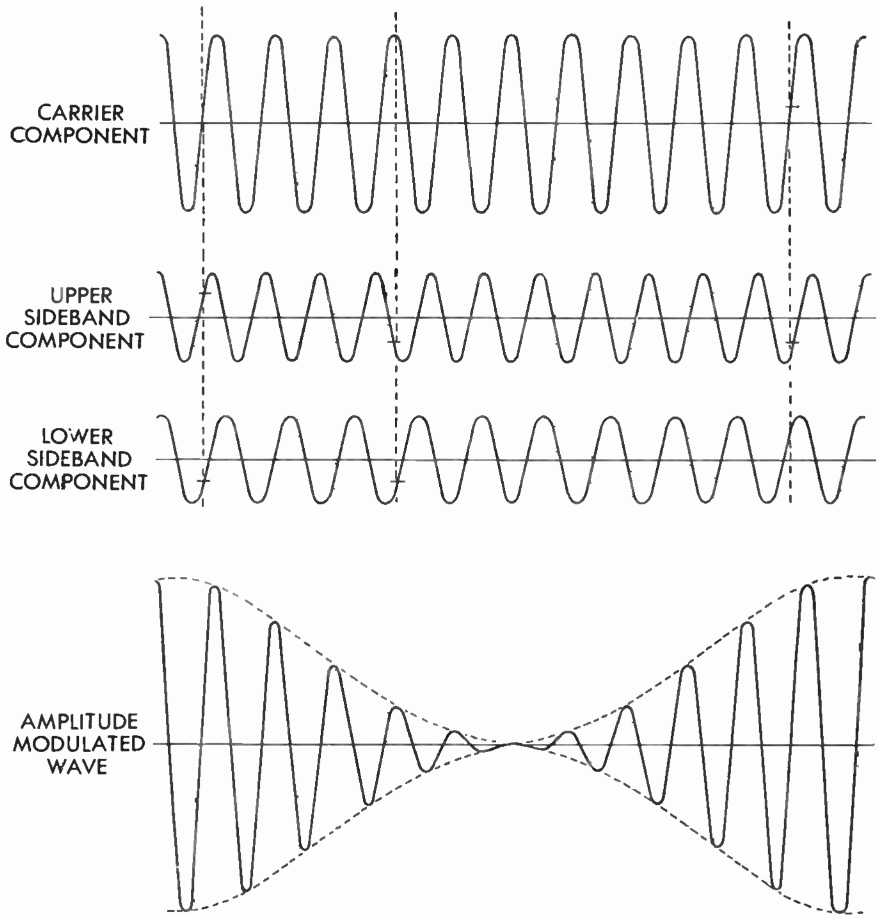


FIG. 189 Components of a 100% Amplitude-Modulated Wave

the dissimilarities in the resultant AM and FM wave, the center frequency or carrier and the sideband frequencies which make up these waves are, in many respects, quite similar. Thus, the AM wave is formed by a center frequency and two sidebands, an upper and lower, which for a given strength of audio note and frequency are all constant in amplitude but produce a resultant wave which varies in amplitude. An FM wave is also made up of a center or carrier frequency and a number of sideband pairs which, for a *given audio frequency* and *amplitude*, are all constant but produce a resultant wave which varies in frequency but is constant in amplitude.

The amplitude-modulated wave is made up of three components, the center frequency and the upper and lower sidebands. When the upper and lower sidebands are equal to one-half the amplitude of the carrier frequency a 100-per cent-modulated AM wave occurs (Fig. 189), or one for which the resultant output drops to zero on one part of the audio alternation and rises to twice the peak amplitude of the carrier alone on the other cycle of the audio frequency. The resultant FM wave always remains constant in amplitude and consists of a center frequency and a number of sideband pairs which vary in relative amplitude and number in accordance with the extent of the frequency modulation. Although the relative amplitudes of the modulation and number of sideband pairs continue to vary, the resultant is always of constant amplitude and varying frequency. It is evident if the resultant is to equal the algebraic sum of the components which form it, the center frequency itself varies in amplitude in accordance with the modulation. In amplitude modulation the center or carrier frequency is always constant in amplitude, and the sidebands just add and subtract from it to form a varying amplitude resultant. In frequency modulation, the sidebands and the center frequency vary in amplitude with modulation to form a constant amplitude resultant.

BESSEL FUNCTION CHART

Modulation Index	*Center Freq.*	*C.F. ± Audio*	*C.F. ± 2 × Audio*	*C.F. ± 3 × Audio*	*4 ×*	*5 ×*	*6 ×*	*7 ×*	*8 ×*	*9 ×*
0.0	1									
0.1	0.9975	0.0499								
0.2	0.99	0.0995								
0.3	0.9776	0.1483	0.0112							
0.5	0.9385	0.2423	0.0306							
0.7	0.8812	0.329	0.0589	0.0069						
1.0	0.7652	0.4401	0.1149	0.0196						
1.6	0.4554	0.5699	0.257	0.0725	0.0150					
2.0	0.2239	0.5767	0.3528	0.1289	0.034	0.007				
4.0	−0.3971	−0.066	0.3641	0.4302	0.2811	0.132	0.049	0.015		
6.0	0.1506	−0.2967	−0.2429	0.1148	0.3576	0.362	0.246	0.129	0.056	0.021

The complex nature of the FM wave can best be explained with a Bessel function chart. This chart compares the amplitude of the center frequency and the various sidebands which make up the wave to the amplitude of the center frequency when no audio modulation is applied. For example, if a 5,000-cycle audio note of sufficient amplitude is applied as modulation to an FM transmitter to cause a plus and minus 5-kilocycle instantaneous peak deviation about center frequency the modulation index would be 1. The Bessel function chart shows us that with a modulation index of 1 the center frequency has an amplitude of 0.7652 of the unmodulated center frequency and a first set of sideband pairs 0.4401 of the unmodulated center-frequency amplitude. The second sideband pair has an amplitude of 0.1149, and the third, 0.0196. A summation of the center-frequency amplitude and all the various sidebands would produce a constant-amplitude frequency-modulated wave which has a deviation of plus and minus 5 kilocycles about the center frequency. The various components which make up this wave are demon-

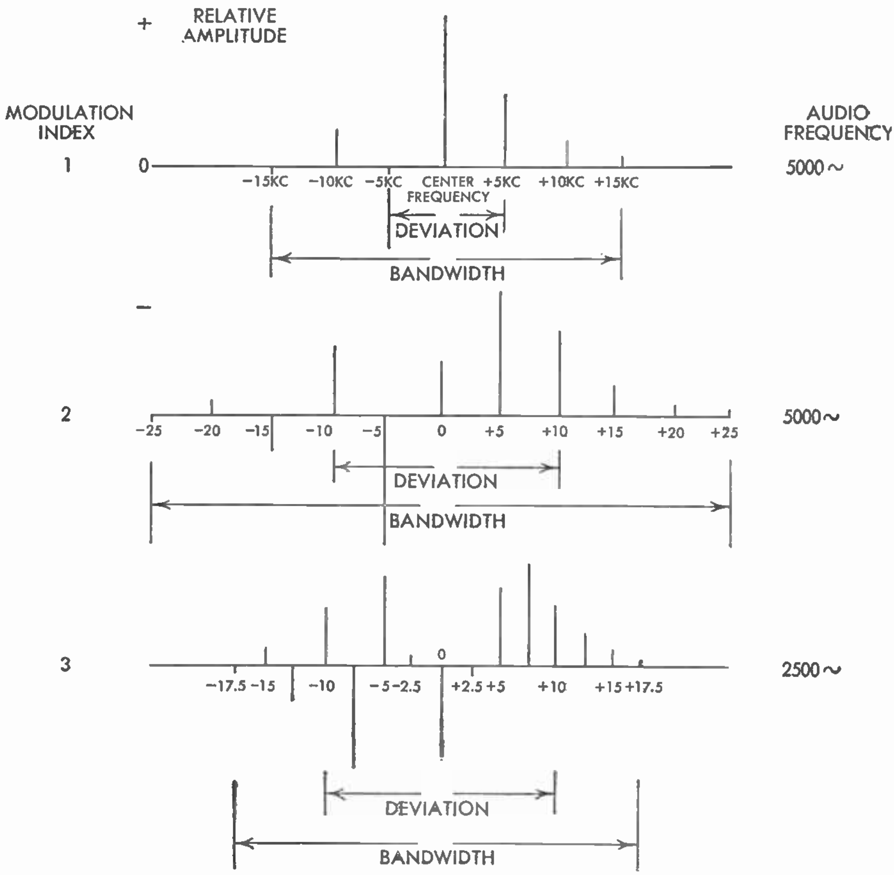


FIG. 190 Components of an FM Wave According to Modulation Index

strated in the top drawing of Fig. 190. Notice that even harmonic sidebands have same polarity; odd harmonic sidebands have opposite polarities. It can be seen that although the deviation is only plus and minus 5 kilocycles, there are sideband components at frequencies which are 15 kilocycles above and below the center frequency. Thus, the actual bandwith required to transmit this signal is not only set by the deviation but is determined by the number and frequency of the sidebands which make up the wave—in this case a total channel width of 30 kilocycles would be necessary. If the amplitude of the original 5,000 audio-cycle note is doubled the deviation also doubles, and the modulation index becomes 2, in which case the number of sideband pairs is increased as shown on the chart. With a modulation index of 2 the number of sideband pairs increases and the center frequency itself is lower in amplitude than some of the sideband components. The distribution of signal is shown in the second drawing of Fig. 190. Algebraic addition of all components forms a frequency-modulated wave with a deviation of 10 kilocycles and a bandwidth set by the highest frequency sideband pair.

If the amplitude of the audio note is left the same (sufficient to have a 10-kilocycle deviation), but the audio frequency is cut in half, to 2,500 cycles per second, the modulation index becomes 4. With a modulation index of 4, the sideband pairs increase to 7. Thus, the rise of sideband pairs with an increase in modulation index is again demonstrated. In this case the carrier or center frequency and the first pair of sidebands are actually opposite in polarity with respect to what the center frequency would be if no modulation were present. It is important to note on Fig. 190 that, although the number of sideband pairs has increased with the reduction in audio frequency, the actual bandwidth has decreased because of the shorter spacing of the harmonically related sideband frequencies. For example, the total bandwidth for a modulation index of 4 and an audio frequency of 2,500 cycles is now only plus and minus 17.5 kilocycles.

In the television FM transmitter the maximum deviation is 25 kilocycles about center frequency, and a maximum amplitude 15,000 cycle note would cause this amount of deviation when the modulation index is somewhat higher than 1.6. With a modulation index of 1.6 the number of sideband pairs is 4 and the total bandwidth four times 15,000 cycles or plus and minus 60 kilocycles. This represents the maximum channel width required to produce a 25-kilocycle deviation at the highest audio frequency. If the amplitude of the audio note is held at this maximum value and its frequency reduced, the bandwidth becomes less although the number of sideband pairs does increase. Thus, the FCC maximum tolerances are always set by the maximum deviation at the highest audio frequency to be transmitted. In actual practice the higher frequency sidebands are very low in amplitude; for example, at this modulation index of 1.6 the highest frequency sideband is only 1½ per cent. It is possible therefore to eliminate these low-level sidebands without serious injury to the fidelity of the signal. Consequently, although the chart shows a maximum bandwidth of 60 kilocycles for this highest frequency component, maximum bandwidth could well be plus and minus 40 kilocycles because of the low amplitude of the higher frequency sideband components, and also the fact that very seldom in any type of program is a maximum-amplitude high-frequency note transmitted. Generally, the high-frequency components of program material are relatively low in amplitude and would not cause maximum deviation. Again, the lower the deviation the smaller the bandwidth required to transmit the information. In the interest of oscillator drift, receiver bandwidth is 150 to 200 kilocycles.

FM AND AM COMPARISON

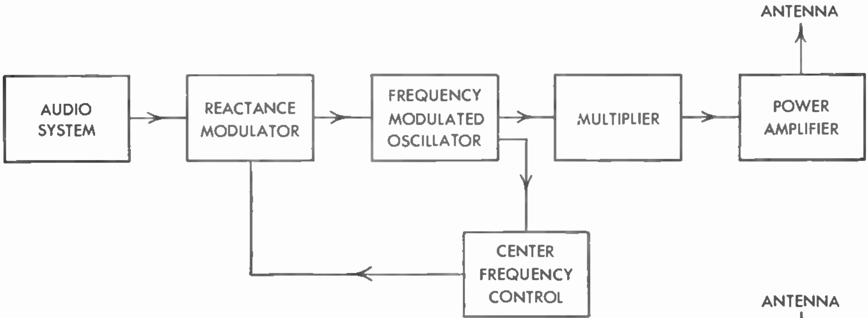
To summarize the discussion of the FM signal and its comparison to an AM signal refer to the following chart.

FM-AM COMPARISON CHART

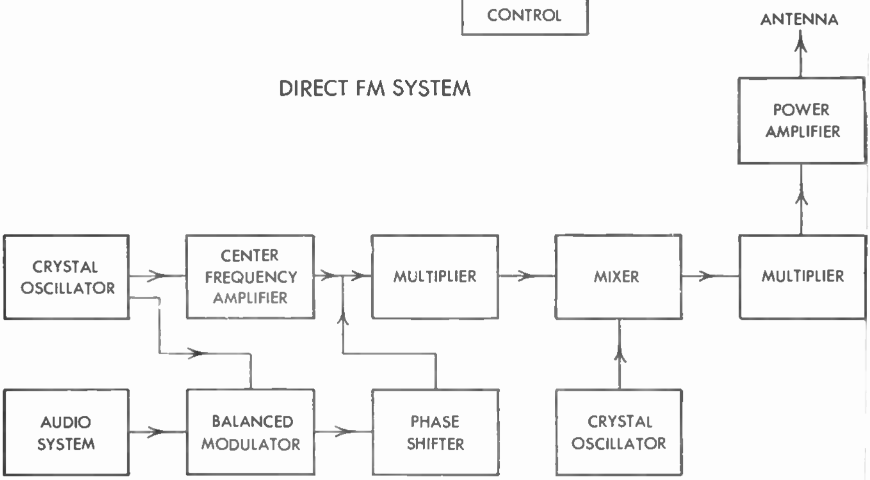
FM	AM
1. Modulation causes resultant wave to deviate in frequency.	1. Modulation causes resultant wave to vary in amplitude.
2. Audio-frequency increase causes rate of deviation to increase.	2. Audio-frequency increase causes rate of amplitude change to increase.
3. Increase in audio-frequency amplitude causes deviation to increase.	3. Increase in audio-frequency amplitude causes greater change in amplitude of resultant wave.
4. FM resultant wave constant in amplitude.	4. AM resultant wave varies in amplitude with modulation.
5. FM wave consists of center frequency and a number of sideband pairs, all harmonically related to audio frequency with respect to separation from center frequency. Number of pairs is dependent on modulation index.	5. AM wave consists of center frequency or carrier and one pair of sidebands.
6. Addition of center frequency and sidebands produces a constant amplitude resultant. Center frequency of FM wave does vary in amplitude. Only one sideband pair is of significant amplitude when modulation index is less than 0.25.	6. Center frequency or carrier remains constant in amplitude with modulation and sidebands (only one pair) add and subtract to form amplitude-varying resultant.
7. Power output of FM transmitter is constant with modulation.	7. Power output of AM transmitter varies with modulation.
8. Amplitude noises are limited. Noises from frequency modulated signal produce more interference when separated from center frequency. Signal-to-noise ratio varies as the deviation and inversely as highest audio frequency.	8. Signal-to-noise ratio decreases as the bandwidth is increased—higher audio-frequency limit.

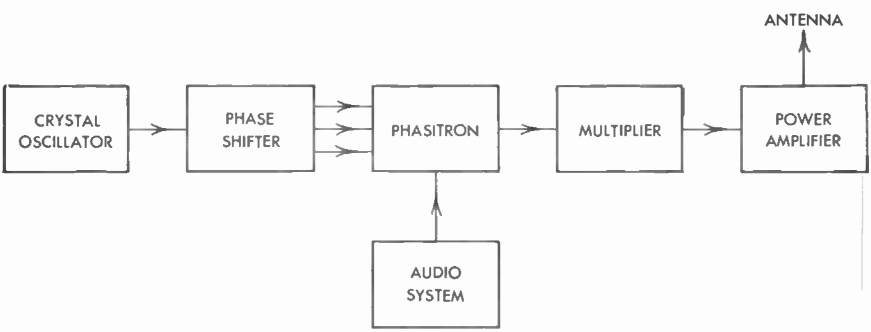
107. *General Description of the FM System*

There are three basic commercial FM-transmitter types, one direct FM type and two basic indirect FM methods. In the direct FM system (Fig. 191) the center-frequency oscillator is frequently modulated by a reactance tube the plate current of which varies with the audio frequency and amplitude. Thus, the FM transmitter consists of an audio system which steps up the amplitude of the audio which, in turn, feeds a reactance-tube modulator which frequency-modulates the self-excited center-frequency oscillator. The frequency-modulated output of the oscillator is then multiplied in frequency and deviation as well as increased in ampiltude until it is on the fundamental frequency of the station. For example, if the FM transmitter is to operate on



DIRECT FM SYSTEM





INDIRECT FM SYSTEMS (PHASE MODULATION)

FIG. 191 Commercial FM Transmission Systems channel 3 it means the sound-carrier frequency is 65¾ megacycles and the total maximum deviation plus and minus 25 kilocycles about this point. It must be again remembered that the actual bandwidth is somewhat greater than the total deviation because of the higher frequency sideband components (maximum of about plus and minus 40 kilocycles).

When phase modulation is used only a small frequency deviation can be obtained at the frequency of the crystal oscillator. A series of multipliers and mixers must be used to step up the deviation by a greater ratio than the center frequency to produce an output which is of the correct fundamental frequency and has the required amount of deviation.

The striking similarity between amplitude modulation and phase modulation with its resultant frequency modulation is demonstrated in Fig. 192. For example, if two sideband components above and below center frequency, each having 1/10 amplitude of the center frequency, are combined without shifting their phase with respect to center or carrier frequency, an amplitude-modulated wave of less than 20 per cent modulation is formed. Now, if the same two sideband components are first shifted 90 degrees in phase before recombining with the center frequency, a resultant phase-modulated wave is formed which changes with the amplitude and frequency of the audio. This shift in phase produces frequency modulation because the periods of the individual cycles change with phase variations.

To prevent amplitude modulation, and distortion greater than 1 per cent, it is necessary that each sideband amplitude be 1/10 or less than that of the center frequency, producing a maximum phase shift of only 0.2 radian. Heretofore, this radian phase shift was spoken of as *modulation index,* but now we can understand modulation index as actually representing phase shift or, the shift in phase of the resultant wave with respect to the center frequency if no frequency modulation existed. With a maximum phase shift of only 0.2 radian (modulation index 0.2) only a slight frequency deviation is obtained when the sidebands are recombined with the center frequency.

Observation of the Bessel function chart tells us that with a modulation index of 0.2 only one sideband pair is present, and each one of the sidebands has approximately 1/10 the amplitude of the center frequency, or a total sideband amplitude of 1/5 the center or carrier-frequency amplitude. These are the conditions met in the formation of the frequency-modulated wave by means of a phase-modulating system.

The small radian phase shift introduces complexity in the design of the phase-modulated or indirect FM transmitter. For example, if the crystal oscillator frequency is as low as 200 kilocycles and the maximum deviation obtainable is 10 cycles (0.2 radian shift at 50∼ or $0.2 \times 50 = 10\sim$), to reach the carrier frequency in the television channel spectrum requires a center-frequency multiplication of around 300 to 400; to obtain the FCC maximum deviation of plus and minus 25 kilocycles requires a multiplication of approximately 2,500. Thus, if a straight multiplier system were used and the deviation

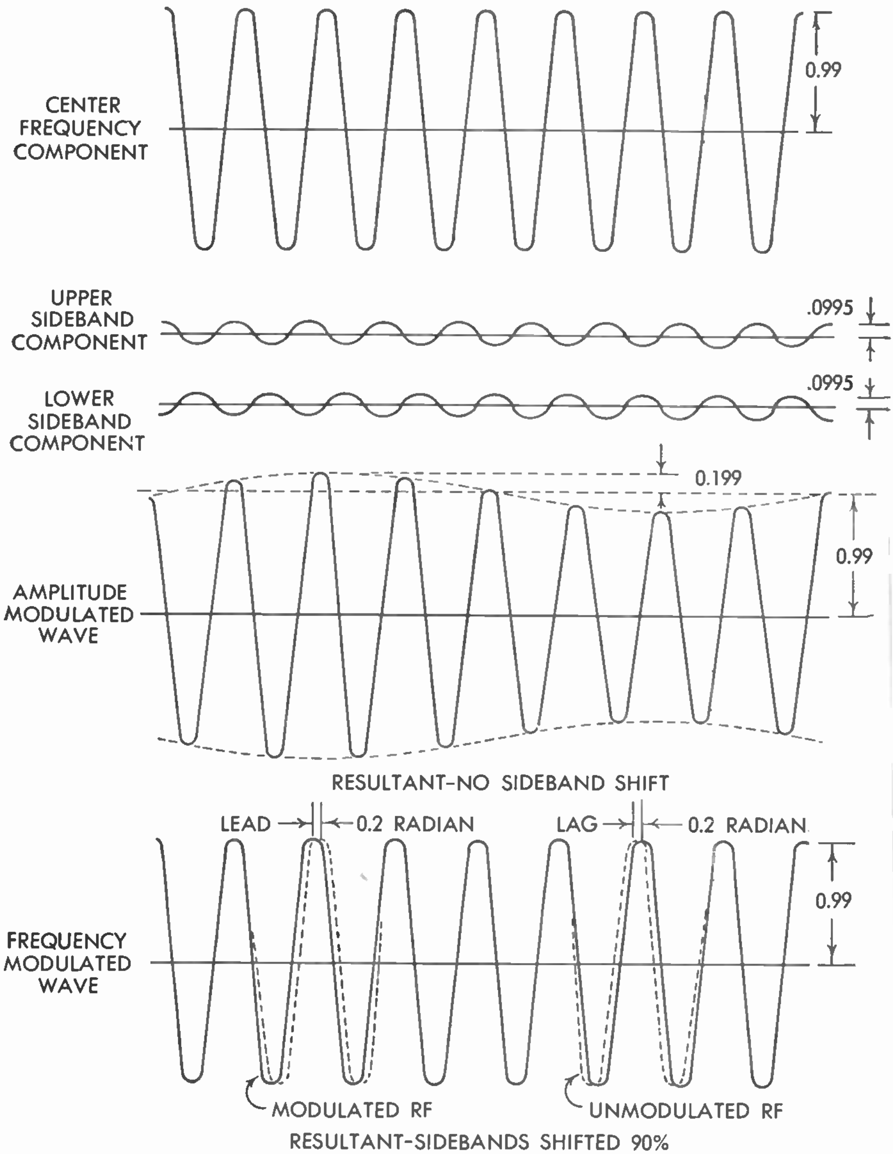


FIG. 192 Comparison of AM Wave and PM Frequency-Modulated Wave

were multiplied the proper amount, the center frequency would be too great; if the center frequency is multiplied the correct amount the deviation would be too small to meet the FCC requirements. Therefore, it is necessary to use a system which will multiply the deviation more than the center frequency. This can be done by using two multiplier circuits which have a total multiplication of 2,500, the amount necessary to bring the deviation up to plus or minus 25 kilocycles. Center frequency is also amplified a required amount by the first multiplier but then is fed into a mixing stage which, when combined with another crystal-oscillator frequency, will reduce the center frequency to some low value again. The mixer stage, although it reduces the center frequency (because the output circuit is tuned to the difference frequency between the incoming signal and the new oscillator frequency), does not have any effect on the deviation. The appreciably greater deviation and the now lower center frequency is again multiplied by the second multiplier system, and its output produces the correct center frequency at the proper deviation.

The second phase-modulating system consists of an actual tube designed to produce phase modulation. This tube, called a *phasitron,* produces the actual phase-modulated wave within the tube itself. Output of a crystal oscillator drives a phase-shifter circuit which produces three outputs at the crystal frequency but with their phases displaced 120 degrees in relation to each other. These three frequencies, when applied along with the audio frequency to the phasitron, produce a phase-modulated output which deviates in frequency a substantial amount about the original frequency of the crystal oscillator. In fact, sufficient deviation is obtained and a second mixer system is not needed. Output of the phasitron is then passed to a multiplier which increases the frequency and the deviation the proper amount to drive the power amplifier and antenna system on the sound carrier frequency of the television channel.

SOUND CHANNEL OF THE TELEVISION RECEIVER

Two basic systems are used in the television receiver to demodulate and amplify the frequency-modulated sound signal. The first method used (Fig. 193) employs two separate i-f systems, one for the picture and a second FM i-f system for the sound. Carriers of the two i-f channels are set 4½ megacycles apart, or the amount of the actual separation of the picture and sound carriers from the transmitter. Thus, the r-f system of the receiver is sensitive to both carriers and one local oscillator will mix with the two incoming signals to produce two different i-f frequencies, 4½ megacycles apart. Each i-f channel is properly tuned to accept the desired i-f signal and reject the other.

A second system, called the *intercarrier method,* uses just a single i-f channel. The r-f, mixer, and i-f systems are tuned very broadly so that the picture and the sound carriers, and sidebands are passed through the entire system and on to the video detector and amplifier. The i-f system, however, is detuned to some extent at sound carrier frequency to prevent sound carrier from dominating the video detector. Actually, the sound carrier frequency and its component sidebands appear as a 4½-megacycle component of modulation to the video detector and are present in the output circuit of video detector as a 4½-megacycle center frequency plus and minus 25 kilo-

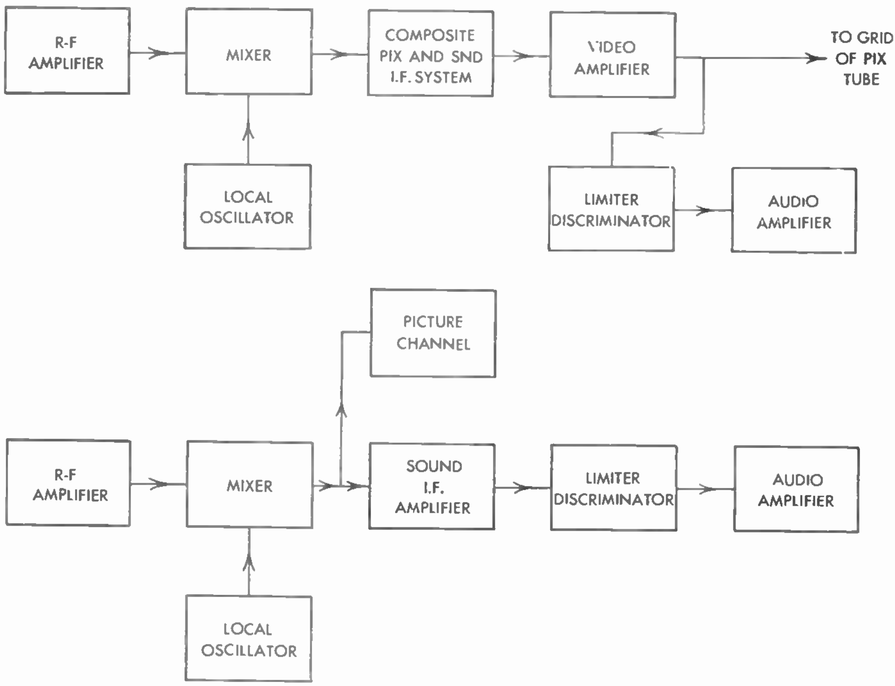


Fig. 193 Television Receiver FM Sections

cycles maximum deviation. Thus, in the output circuit of the video amplifier a tuned circuit resonant at 4½ megacycles is present to remove the sound carrier signal from the picture and prevent appearance of sound signal component on grid of the picture tube. This 4½-megacycle FM sound signal is passed to a limiter and discriminator for demodulation; sound signal is now increased in amplitude and excites a loudspeaker.

The actual FM section of the television receiver consists of a limiter and discriminator; the limiter to remove any amplitude and noise variations on the FM signal and the discriminator to detect the desired audio frequencies which are represented as frequency modulation of the sound i-f signal. In some receivers a new type of FM detector, called a *ratio detector,* is used because of added advantages. The ratio detector is itself an inherent limiter and no limiter i-f stage is necessary under most circumstances. The ratio detector is also somewhat more sensitive and produces an output with a very low hiss level as compared to the conventional discriminator, even when tuning between stations.

A high-fidelity audio system should be used, of course, to take full advantage of the characteristics of frequency modulation. A good output transformer and a larger speaker are imperative if high fidelity results and satisfactory range are to be obtained. All television receivers should be designed not only to obtain the best picture but also to take full advantage of frequency modulation in the sound section.

108. *Characteristics of the FM System*

Frequency modulation has a number of advantages as compared to amplitude modulation for sound transmission. Inasmuch as the television spectrum is inherently broad to transmit the high-frequency picture components, the little added channel needed to transmit frequency-modulated sound is insignificant. Frequency-modulated sound system substantially reduces noise pickup and interference. Interfering signal can be almost half the strength of the desired signal and still the desired signal completely dominates the output so much, undesired signal is not heard. With an AM signal the desired signal has to dominate by at least 10:1, and at times as great as 100:1, depending on the type of interference, before the desired signal is heard satisfactorily in the output. Frequency modulation means good fidelity, an improved dynamic range, and, what is most important, a more reliable noise-free coverage of the primary and fringe areas of the television station.

NOISE REJECTION

Noise in the FM channel causes amplitude and frequency modulation of the FM signal. As far as amplitude modulation is concerned, it is removed by limiting action, either by a limiter i-f stage or limiting action of the discriminator itself when a ratio detector is used. Limiting action causes the peaks and varying levels of the FM signal to be clipped off, and a constant-amplitude FM signal appears in the output of the limiter. Although this means portions of the i-f cycle are clipped off to obtain a flat level, the removal of these peaks does not in any way affect the operation of the FM signal, because the actual fidelity of the sine wave is restored by the tuned circuit which follows, plus the fact that the desired information is riding on the FM signal in the form of frequency modulation and has nothing whatsoever to do with the amplitude characteristics of the sine wave. Thus, if a signal is delivered to the receiver with sufficient strength to operate the limiter the amplitude noise components can be completely removed.

Noise interference also causes frequency modulation of the desired signal, and if that modulation compares with the deviation of the FM signal, noise signal will appear in the output. However, the actual deviation of the FM signal as caused by the desired modulation is, under most circumstances, much greater than the deviation of the same signal by noise interference. Noise interference in the FM system actually has less effect when the frequency is near to the center frequency and the beat note set up between noise and carrier is relatively low, than if the interfering signal is considerably removed from the center frequency (for example, 10 to 15 kilocycles away).

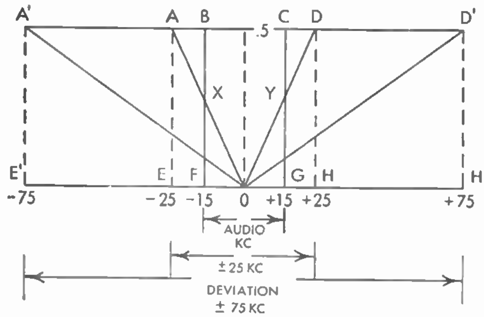


FIG. 194 Noise Characteristics of an FM System—Noise One-Half Amplitude of Signal

This condition exists because the higher the frequency of the beat note set up by the interference the more frequency deviation of the FM signal it causes. In addition, higher frequency components of the desired modulation on the FM signal are often of a lesser amplitude than the lower frequency components. Therefore, any interference of a higher signal frequency has a more decided effect on the ratio of desired signal deviation to noise deviation.

The noise characteristics of a typical FM receiver with respect to audio frequency and noise interference is shown in Fig. 194. In the noise chart it was assumed that the noise amplitude was such that a carrier-to-noise ratio of 2:1 existed. Inasmuch as an AM receiver has equal sensitivity to signal and noise over its bandpass and has equal sensitivity to noise frequencies within that spectrum, a rectangle of 0.5 peak represents receiver noise. An FM receiver has reduced sensitivity to noise as well as increased sensitivity to components of noise frequencies further separated from center frequency. A triangle can best be used to represent noise sensitivity of an FM receiver. In the FM noise triangle, noise was assumed again to be one-half the amplitude of the desired signal. In an AM receiver signal and noise are amplified by the same ratio, through the receiver, consequently the noise is represented by the rectangle *BCFG*. In the case of noise modulation of the FM signal, the noise is completely removed at the center frequency and noise inherent in an FM receiver is represented triangular in form and is caused by frequency modulation of the FM signal by noise. Inasmuch as this modulation increases with the audio frequency of the interfering noise (beat note between interfering signal and center frequency), the signal-to-noise ratio becomes decreasingly poorer as we go out toward the high-frequency limits of the received bandpass. Thus, the triangles *OXF* and *OYG* represent the noise interference in the FM signal.

A considerable improvement in signal-to-noise ratio is evident from the noise diagram. This noise could, of course, be further reduced by extending the limits of the frequency deviation of the desired signal (extend deviation to plus and minus 75 kilocycles); therefore diagonal will stretch out to *A′* and *D′*, reducing the noise triangle. However, the lower deviation of the FM sound for television is ample because it is necessary only to have sound carry and noise rejection be complete over the transmitted range of the picture signal. The smaller deviation also simplifies receiver design, as far as the FM channel is concerned, because local oscillator drift will not as readily cause interference between picture and sound i-f, and sound i-f can be somewhat narrower.

PRE-EMPHASIS AND DE-EMPHASIS

Inasmuch as it is an inherent characteristic of an FM system to have a lower signal-to-noise ratio when the interference produces a beat in the higher audio-frequency range, a de-emphasis system is always employed in the output circuit of the FM detector between detector and first audio amplifier. This network (Fig. 195) is a simple *R-C* combination which, because of the decreasing reactance of C with a rise in frequency, causes a gradual attenuation of high-frequency components. Thus severe noises which are dominant at higher audio frequencies are attenuated, improving signal-to-noise ratio.

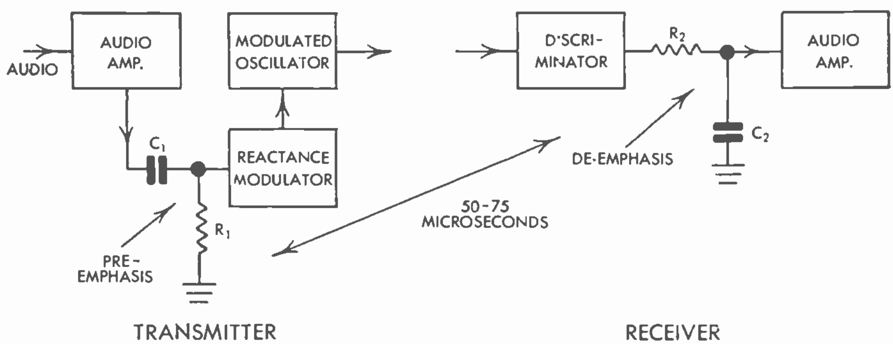


FIG. 195 Pre-Emphasis and De-Emphasis

Of course, this de-emphasis network also attenuates higher frequency signal components and frequency distortion would exist were it not pre-compensated for at the transmitter. At the transmitter a pre-emphasis network accentuates the highs to produce a greater deviation than lower frequency notes of the same amplitude. Consequently, the higher frequency components produce a stronger signal at the output of the FM detector than a lower frequency, but equal amplitude audio component. This signal when applied to the de-emphasis network takes a high-frequency loss which equalizes the response of the FM system but very decisively cuts down high-frequency noise components. The pre-emphasis and de-emphasis networks are inverse networks—one produces a gain at highs, the other a loss. If the time constants are made equal, the over-all response of the two networks is linear. It is customary to standardize the time constants at 50 to 75 microseconds.

FIDELITY AND DYNAMIC RANGE

In addition to the relatively noise-free reception offered by FM, the fidelity and range of most FM systems is superior to that of AM systems. However, it is necessary that the FM system of the receiver be designed to take full advantage of these features. In their efforts to obtain a satisfactory picture at a reasonable cost, manufacturers have been neglecting the sound system which, if properly designed, can add much to the realism of live-talent telecasts. The frequency response of the FM system should be flat from 50 to 15,000 cycles and should have a gradual taper-off beyond 15,000 to prevent transients. Fortunately, we are able to take advantage of such a wide audio range in FM if our frequency deviation is ample because signal deviation is so much greater than that created by noise interference and because high-frequency de-emphasis improves signal-to-noise ratio at higher audio frequencies.

Thus the background noise is low and signal-to-noise ratio does not increase when receiver response is broad as it does when the tone control in an AM system is switched to high. To obtain true fidelity, low-frequency response must be good and free of distortion to prevent crosstalk and consequent "boomy" bass because any distortion present will cause low-frequency components to beat together and produce a multiplicity of low-frequency notes which develop loud bass but no tonal quality.

Another excellent quality of FM is its amplitude range which permits transmission and reception of very quiet passages of music as well as full orchestra. In AM transmission quiet passages must be increased in amplitude to override the receiver noise while heavy passages are subdued to prevent overmodulation of the transmitter and consequent distortion and interference between channels. In FM, low passages will carry at true amplitude if limiting action is thorough at the receiver and noise reduction complete. Likewise, high-amplitude notes cause maximum deviation of the FM signal and do not result in removal of signal and the very serious distortion associated with AM overmodulation.

Another advantage of FM is its reliability over the service range of the TV station. Because of its noise- and interference-reduction characteristics, interfering signals must be more than half as strong as the desired signal before they can even be noticed. Therefore, the TV sound becomes a secondary problem to the conveyance of an interference- and noise-free picture.

109. *Sound I-F Amplifier and Limiter*

The sound i-f amplifier of the TV receiver is not nearly so great in bandwidth as the picture i-f system—approximately 100 to 200 kilocycles. This is, of course, much wider than the conventional AM broadcast i-f (±5 kilocycles) and about as wide as the i-f amplifiers designed for use on the commercial FM broadcast band (±75 to 150 kilocycles). All of these bandwidths

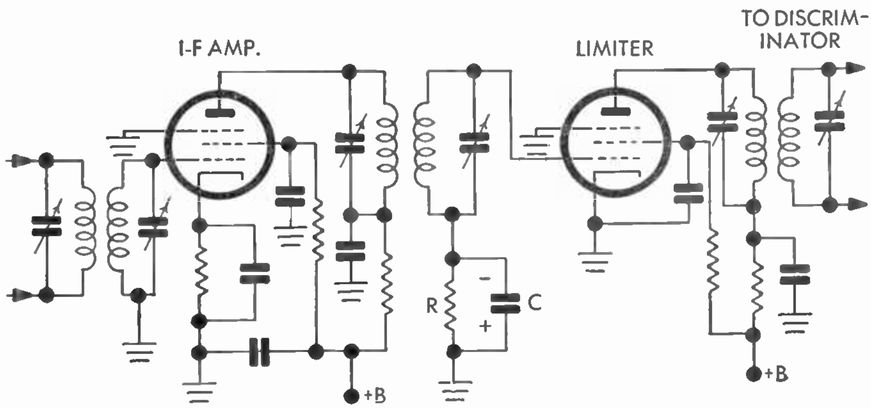


FIG. 196 Sound I-F Amplifier and Limiter

mentioned are very much narrowed in comparison to the 3½- to 4-megacycle picture i-f. Some few sound i-f stages are loaded with resistors to attain the proper bandwidth (10,000 to 100,000 ohms) although in most cases the desired bandwidth is obtained by proper coupling and loading between stages, and keeping stage Q at the proper value. For truer fidelity it is desirable not to have the response down more than 1 or 2 decibels at the 25-kilocycle points on each side of center frequency.

The limiter removes amplitude variations riding on the FM signal which have been introduced by noise and interference components. These variations are removed without disturbing the fidelity of the desired information riding on the signal in the form of frequency modulation. It must be remembered that noises are removed only when the received signal is strong enough to cause proper operation of the limiter tube. Likewise, noise causes some frequency modulation of the signal which under most circumstances is completely overshadowed by the signal but which becomes noticeable when the signal is weak or the noise particularly severe.

A typical i-f stage and limiter are shown in Fig. 196. The limiter tube operates with a reduced value of screen and plate voltage and has its cathode grounded and no external fixed bias. The low anode voltages mean the tube will cut off early and at a low value of grid voltage—thus it does not take much of a negative signal sweep to drive the grid beyond cutoff. Since no external or cathode bias is applied the stage is signal-biased by rectifier action in the grid circuit and the time constant of the grid *R-C* combination. Thus, the positive sweep can only be small before grid current flows and grid limiting sets in.

LIMITER OPERATION

In operation (Fig. 197) the positive alternation of the applied i-f wave draws grid current which charges capacitor *C* to almost the peak value of the i-f wave setting limiter bias. This charge is held between cycles of the i-f waves because the time constant is much longer than the period of the i-f

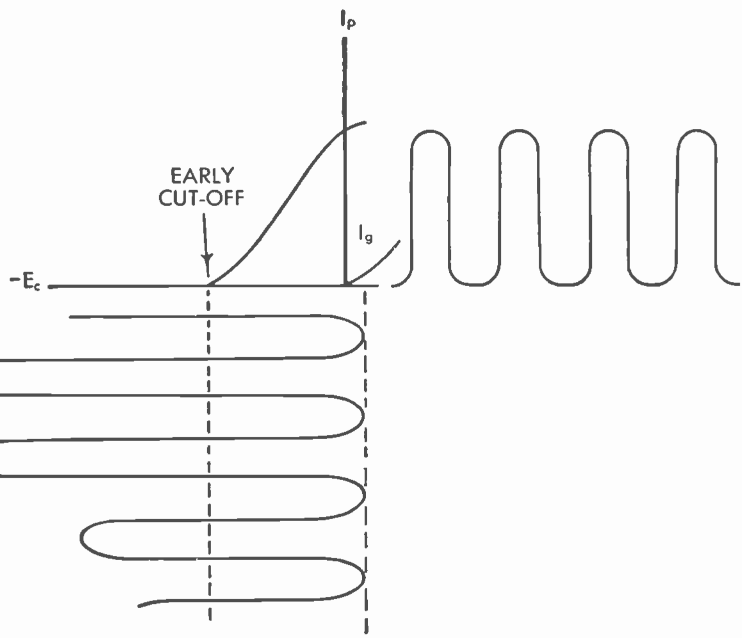


FIG. 197 Limiter Action

wave—charge is re-established during the peak of each i-f cycle. Also notice on the figure that if the cutoff bias of the tube is low, only a very small portion of the actual grid cycle causes a plate-current variation. We might at first suspect that this severe clipping of the signal would cause distortion. If the desired information were in the form of amplitude variations this would be so; however, desired information is in the form of frequency modulation which has not been disturbed. Fidelity of the sine wave is restored by the energy-storage characteristics of the plate-tuned circuit—sine waves now are of virtually constant amplitude.

Inasmuch as the grid capacitor accepts a charge proportional to the peak amplitude of the applied i-f wave, any variation in this level will change the charge on the capacitor by the same amount. Actual peak of sine waves will occur at the same grid voltage level—level at which grid current flows. Any portion of the signal which extends beyond the cutoff point will be removed and only the narrow portion of the signal between cutoff and grid limiting will pass on to the output circuit. No matter how strong the signal or how weak (so long as it is greater in amplitude than the voltage differential between cutoff and grid limiting), the same amplitude plate current change occurs in the tube. It is a function of the r-f and i-f sections of the receiver to deliver sufficient signal to the limiter to cause satisfactory limiter action over the service area of the TV station. If the signal is not great enough in amplitude to swing limiter, noise will get through to the output. Also, as the receiver is tuned off the station, noise level comes up because there is no amplitude limiting of any amount and amplitude noises do get through.

Limiting action can be improved, of course, by further reducing the voltage differential between cutoff and grid limiting. However, there is only a practical limit to which we can do this because we are also reducing the signal output of the limiter, and if carried too far there will be insufficient drive to the FM detector which follows. A more satisfactory solution is to use two cascade-limiter stages, first stage limiting severe amplitude variations while the second stage removes any residual variations which might carry through because of a weak signal applied to the first limiter.

Choice of time constant of the *R-C* combination in the grid circuit is important if effective limiting of impulse noises is to be complete. A time constant of 1 to 2 microseconds is desirable if good limiting of ignition noise is to exist. The short time constant means the grid *R-C* circuit can bias off a sharp impulse noise almost instantaneously—the charge building up negatively at almost the same rate at which the impulse voltage is swinging positive. If the time constant is short this charge will also drain off quickly and no residual charge will remain on it to upset normal limiter operation under control of the signal. If the time constant is made too short (not long enough in comparison to period of i-f cycle), incomplete amplitude limiting exists because the capacitor will not hold charge between cycles of i-f frequency.

110. *Discriminators and Audio System*

The next step in the utilization of the aural portion of TV transmission is the detection of the audio modulation and then its amplification before it is applied to the loudspeaker. The output of the limiter drives the discriminator, which is the FM detector, and converts the frequency variations of the i-f frequency to audio signal. Discriminator also produces a d-c component of output voltage which corresponds to the departure of the center frequency of the signal from the resonant frequency of the discriminator-tuned circuit. This d-c component of output is useful in operation of an a-f-c system sometimes used to keep the local oscillator on frequency.

CONVENTIONAL DISCRIMINATOR

The conventional discriminator (Fig. 198) is driven by two components of signal from the limiter and consists of two diodes connected to opposite ends of a tuned circuit. They consequently are fed out-of-phase voltages. This voltage is mutually coupled between windings of the tuned transformer. A second voltage is coupled directly through the coupling capacitor *C3* and appears across *L3* in phase with the limiter plate voltage. The mutually coupled voltage is 90 degrees out of phase with the plate or primary voltage because of the double-tuned resonant transformer.

Action of the double-tuned circuit is not difficult to understand (drawing B of Fig. 198) and is dependent on the fact that the secondary tuned circuit is a series-resonant system so far as induced energy is concerned. The phase relations of the various parameters are as follows:

a. Primary coil current i_l lags the primary voltage, by approximately 90 degrees.

b. Changing magnetic field between windings in phase with i_l induces a voltage into secondary.

c. Induced voltage e_{ind} is maximum when the magnetic lines are changing at their fastest rate—when primary current is going through zero. Induced voltage e_{ind} lags primary current by 90 degrees and primary voltage by 180 degrees.

d. Secondary current i_s is in phase with the induced e_{ind} whenever the signal frequency is the resonant frequency of the tuned circuit—current and voltage in phase in a series-resonant circuit. Whenever the induced signal voltage is not of the resonant frequency this relation does not exist. If signal frequency is higher than resonant frequency, circuit becomes inductive and current lags; if signal frequency is lower, circuit is capacitive and i_s leads induced voltage.

(1) Secondary current i_s in phase with e_{ind} at resonance and lags primary voltage e_{in} by 180 degrees.

(2) i_s lags e_{ind} when signal frequency is above resonant frequency and therefore lags e_{in} by more than 180 degrees.

(3) i_s leads e_{ind} when signal frequency is below resonant frequency and therefore lags e_{in} by less than 180 degrees.

e. Output voltage which appears across capacitor lags the secondary current by 90 degrees.

(1) At resonance output voltage e_{out} lags by 270 degrees or leads by 90 degrees input or primary voltage e_{in}.

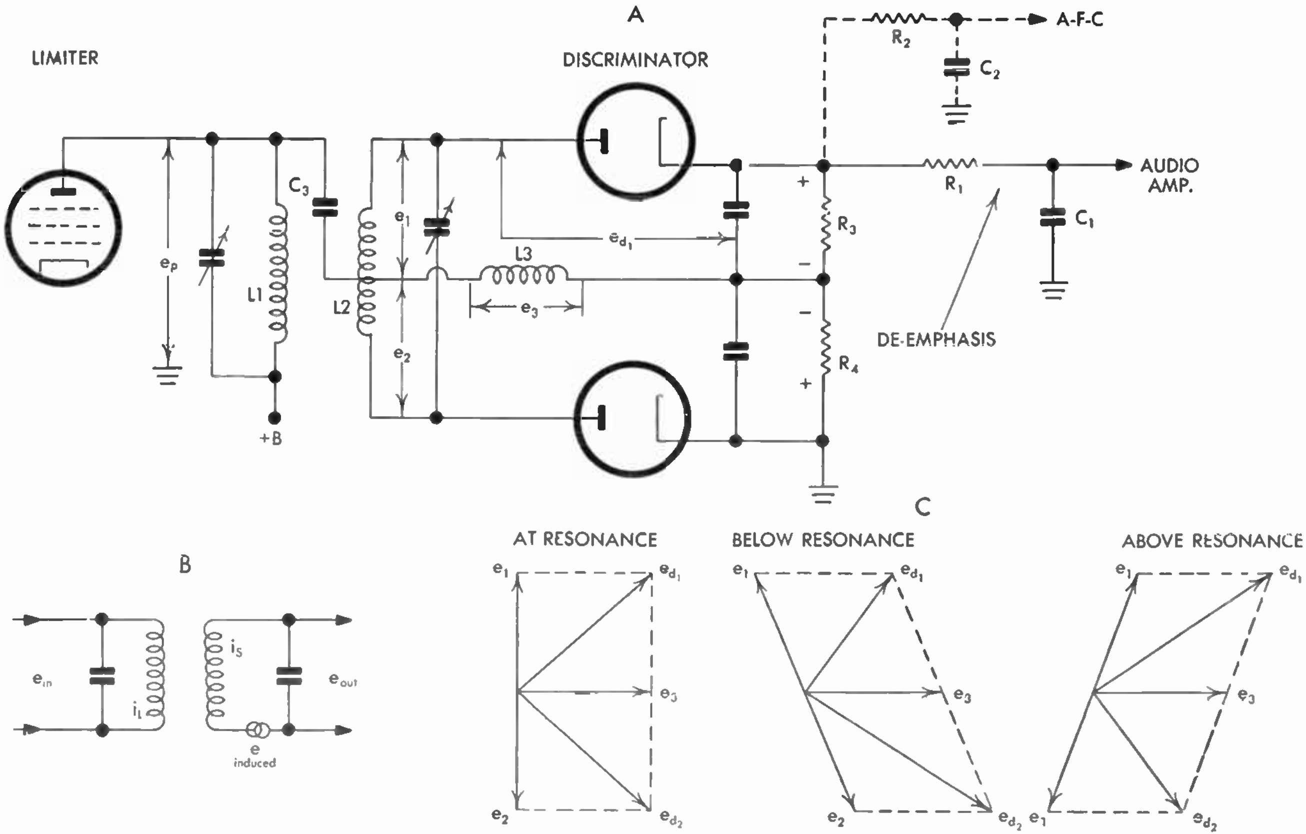


FIG. 198 Conventional Discriminator and Voltage Vectors

(2) Above resonance e_{out} leads e_{in} by less than 90 degrees.

(3) Below resonance e_{out} leads e_{in} by more than 90 degrees.

It is this phase relation between primary and secondary upon which the operation of the discriminator is dependent. Thus, the phase relation between the secondary voltage and primary or plate voltage changes as the FM signal deviates in frequency—this affects the phase of voltages e_1 and e_2 (Fig. 198, drawing A). The voltage e_3 across *L3* is a reference voltage and is always in phase with the primary voltage regardless of frequency. It is the phase relation between e_1 and e_3 and between e_2 and e_3 which causes one diode to conduct more than the other whenever the signal deviates off the center frequency. For example, when the frequency deviates above resonance the top diode conducts more than the lower one and the positive voltage dominates in the output (more current passes up through *R3* than down through *R4*). The positive voltage increases and negative voltage decreases as frequency deviates farther from center frequency (over the bandpass width of the tuned circuit). Likewise if the frequency deviates below center frequency the negative output of the discriminator dominates because the greater voltage is applied to the lower diode—becomes further negative as deviation swings away from center frequency. Thus, it is evident if the applied signal deviates about the center frequency at a sinusoidal rate, a sinusoidal output voltage appears at the output of the discriminator because output voltage follows deviation of signal applied to discriminator diodes—voltage increases as deviation extends further from center frequency. Polarity of this voltage depends on whether frequency swings above or below resonance at a given instant. At center frequency, output is zero because both diodes receive same voltage and voltages cancel across *R3* and *R4*.

VECTOR EXPLANATION

Just why one diode receives a greater voltage than the other under certain conditions and then the second a greater than the first when deviation changes can best be understood with vectors. In the first drawing of Fig. 198C, the vectors show the phase relations when the signal is at the resonant frequency. The voltage across *L3* is in phase with the limiter plate voltage and, therefore, represents the reference voltage for the vector. Output across *L2* is 90 degrees leading the limiter-plate voltage, and therefore voltage e_1 leads e_3 by 90 degrees. The voltage e_2 which is taken off the lower portion of the tuned circuit is 180 degrees out-of-phase with e_1 and lags e_3 by 90 degrees. Inasmuch as the top diode receives a voltage which is the vector sum of e_1 and e_3 and the lower diode the vector sum of e_2 and e_3 both diodes receive the same amplitude resultant voltage. Thus the diode currents are equal and voltages cancel across the output.

If the frequency of the signal is above the resonant frequency we understand the secondary voltage, or e_1, leads the primary voltage (same as e_3)

by less than 90 degrees. Inasmuch as e_2 is displaced 180 degrees with respect to e_1, e_2, lags e_3 by more than 90 degrees. If the vector is completed the greater voltage is applied to top diode and a positive voltage dominates in the output. As the frequency deviates further above center frequency the positive output voltage becomes progressively higher as demonstrated in the vectors of Fig. 199. When the frequency deviates below the center frequency, e_1 leads e_3 by more than 90 degrees and the lower diode receives the greater voltage. Some of the characteristics of the discriminator can be observed from Fig. 200. In the first drawing two typical transfer curves of a discriminator are shown, one for a high output but narrow-band discriminator and a second curve for a wider band discriminator. Bandwidth is obtained at a sacrifice in output voltage. Discriminator, of course, must be designed to be linear over

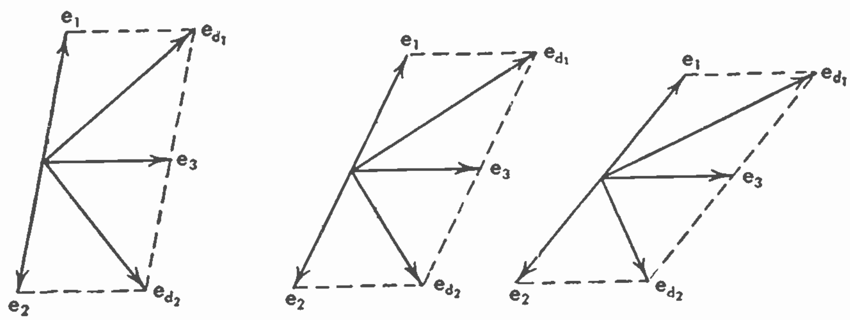


FIG. 199 Increase in Output as Frequency Departs from Center Frequency, ed_1-ed_2

the deviation range of the received signal, ±25 kilocycles or ±75 kilocycles as the case may be. This linearity must be obtained by keeping the discriminator-tuned circuits flat over this range of frequencies by proper loading, coupling, and circuit Q. If a broad bandwidth is desired the values of the diode load resistors are lower generally than if a narrow-band discriminator were being designed because of their increased loading effect on the tuned circuit.

The second drawing of Fig. 200 shows effect of frequency drift of the applied-signal center frequency, in which case not only is there a d-c component of output voltage but the transfer is no longer linear on each side of center frequency. For example, with center frequency at point *1* the applied signal has a center frequency which is higher than the resonant frequency of the discriminator-tuned circuit, and when it is applied a positive voltage appears at the output because the signal frequency is above resonance. Inasmuch as this is the resting point or average frequency about which the deviations exist, the d-c component will not be zero but some positive voltage of a value depending on how far the signal center frequency has drifted off resonance. If the center frequency were below resonant frequency of the discriminator-tuned circuit, a negative-output component would exist.

There is not only a d-c component in the output when the center frequency is not correct (which can actually be put to work in an a-f-c system), but there is also distortion of the desired a-c signal because the deviation does not have the same linear sweep on one side of center that it does on the other. For example, in the first case (point *1*) the sweep on the low-frequency side of center is linear, as it should be, but the signal has to deviate only a small amount on the high side before it runs into the curved portions of the discriminator transfer (off the flat top of the tuned-circuit response curve). It is important to understand that the very same distortion can exist if the discriminator itself is not tuned to resonance, in which case the center frequency of the signal is right but the discriminator tuning is off.

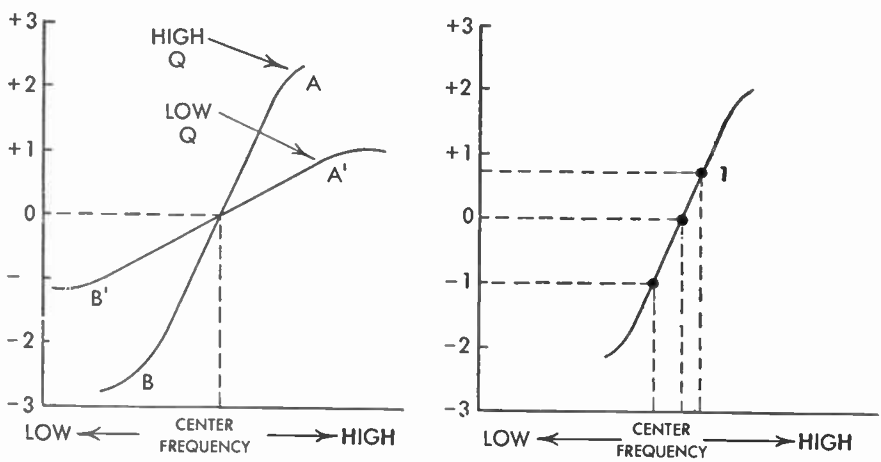


FIG. 200 Characteristics of Discriminator

RATIO DETECTOR

The ratio detector is a form of discriminator which not only detects the frequency modulation but also acts as an effective limiter, permitting the last i-f stage to operate as a gain tube instead of a limiter. In addition to this advantage the ratio detector is sensitive to weaker signals and because of its self-limiting has a low noise level even when no signal is being received.

The ratio detector is similar in schematic appearance (Fig. 201) to a conventional discriminator, with the exception that its output circuit has a stabilizing *R-C* circuit and diodes are connected in a little different manner. The input circuit and voltages e_1, e_2, and e_3 have the same relations with respect to each other as for the conventional discriminator. Thus, the voltages applied to the diodes are again a function of the phase relations between the coupled and direct voltages affecting charging of capacitors *C1* and *C2*. The direct voltage, instead of being taken off at the top of the primary winding, is taken off a low impedance point to prevent severe loading of the last i-f amplifier.

This permits a higher impedance output circuit for the i-f stage and higher gain, for remember we can now use the last stage as a gain stage instead of a limiter.

The output voltage of the ratio detector is taken off between the junction of capacitors *C1-C2* and ground instead of across the two capacitors. Actually, the output is a result of the difference of potential between the charges on the two capacitors while the sum of the two charges is always a constant.

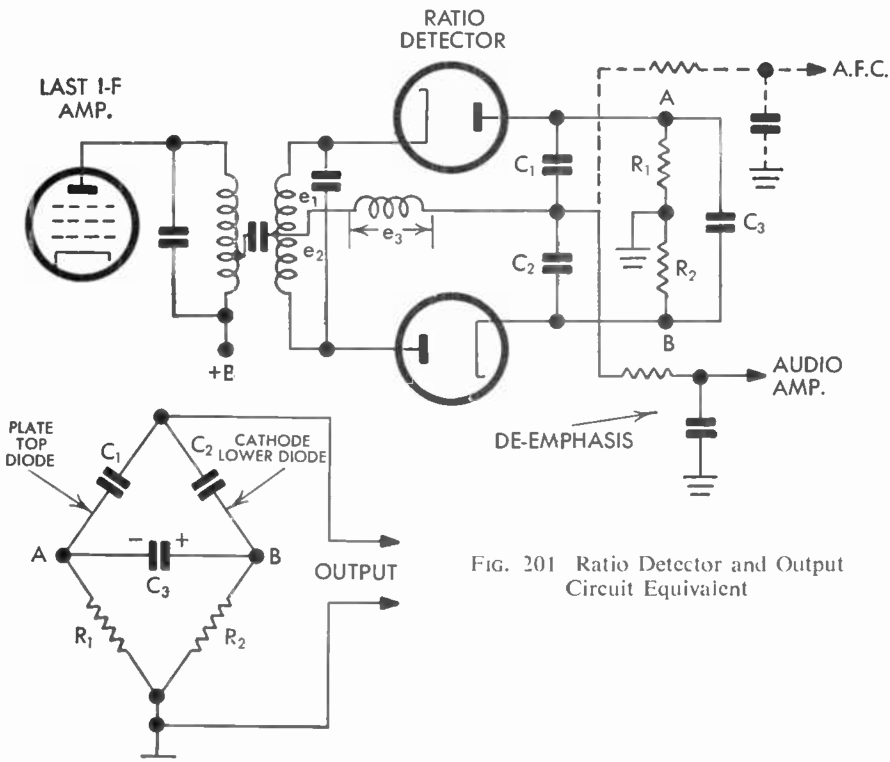


FIG. 201 Ratio Detector and Output Circuit Equivalent

The diode current (always flowing down through *R1* and *R2* because of the manner in which the diodes are connected) keeps a constant voltage (*A* to *B*) which remains constant for any amplitude variations because of the long time constant of *R1*, *R2*, and *C3*, capacitor supplying electrons when the diode current tends to fall and storing electrons when the diode current rises. However, the individual charges on *C1* and *C2* are dependent on the phase relations of the tuned-circuit voltages—difference between two capacitor voltages dependent on phase relations; sum of capacitor voltages is a constant set by diode average-current flow and energy storage of *R-C* combination. If there is any amplitude variation of the average diode current, such as might exist if amplitude variations were present, the long time-constant combinations of the stabilizer circuit would prevent any change in the voltage *A* to *B*. Time constant is chosen to be long at 30 cycles to make certain any amplitude variations will be taken care of. If there is any change in the average signal strength or in tuning from one station to another, the average diode current does change (slow rate of change in comparison to any amplitude variations) and the stabilized voltage sets itself at a different level dependent on signal strength.

It would be entirely possible to use a battery as the stabilizing voltage; however, before any diode current could flow, bias would first have to be overcome with signal amplitude, making the detector insensitive to weak signals. When the *R-C* stabilizing circuit is used, each signal, according to its strength, sets its own *A*-to-*B* fixed-voltage level—high for a strong signal, low for a weak signal. Thus, the detector is sensitive to a weak signal; in fact, it is even operating as an effective limiter for a weak signal.

Operation of the ratio detector can easily be understood by considering it a bridge circuit (drawing B of Fig. 201). In this circuit the voltage *A* to *B* is a constant but the voltages on *C1* and *C2* vary in accordance with the phase of the three voltages which are applied to the input circuit. Thus, if the voltages across *C1* and *C2* differ (as they of necessity do when the applied signal deviates off resonance) and the voltage drops across *R1* and *R2* are held constant by the stabilizing circuit, there must be a current flow through the center at the point where in the usual bridge circuit a meter is connected. Instead of a meter we connect our output system between these two points (de-emphasis network and audio amplifier) to accept the current flow.

Likewise, a d-c component of output also exists when average voltage applied to one diode is greater than that applied to the other (condition which exists when center frequency of signal is not the same as the resonant frequency of the ratio-detector tuned circuit).

AUDIO AMPLIFIER

The audio amplifier should be designed to respond to the inherent fidelity and range of an FM system. The audio, in addition to being linear with changes in amplitude, must have a reasonably linear response from 50 cycles to 15,000. Response above 15,000 must not fall off sharply but must taper off to prevent high-frequency transients—evident as sharp ringing after abrupt musical cutoffs, such as associated with staccato scores. These variations are caused by inability of amplifier to follow a very sharp voltage change (high-frequency harmonics missing), breaking it up into a series of transient oscillations. Of course, opening the audio amplifier to still higher frequencies does decrease the signal-to-noise ratio, but if the front end is properly designed the added noise should be insignificant.

Good low frequency is also a must if good fidelity is to be obtained. Often the amplitude linearity of the audio amplifier is excellent for a middle-range frequency, but for high-amplitude lower-frequency components (2,000 cycles and under) there is often distortion because of core saturation in the audio-output transformer. This distortion causes crossmodulation between low-frequency notes and produces many sum and difference components which produce many bass notes but no true fidelity. The output transformer should be large and contain enough iron to handle the low-frequency power.

111. *Automatic Frequency Control (A-F-C) System*

A-f-c systems are used in television receivers to maintain stability of local oscillators and automatic sync systems. When a-f-c is used in the receiver, there is no critical, fine-tuning adjustment to be made and no tuning adjustment is necessary when shifting between channels with the station selector. A-f-c systems compensate for drift in frequency of the high-frequency local oscillator with heating or change in operating conditions.

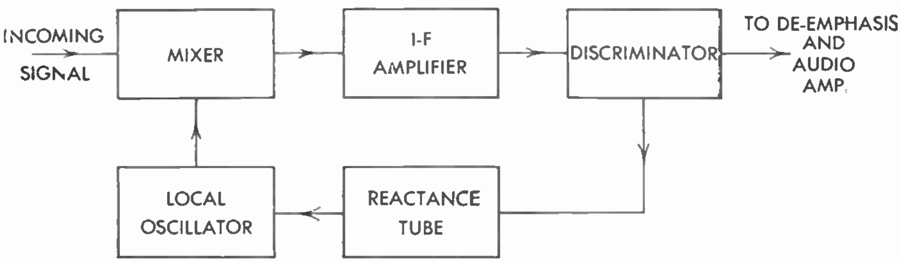


FIG. 202 Block Diagram, Automatic Frequency Control

The block diagram of Fig. 202 shows the essential components of the system. The sound section of the television receiver consists of mixer-oscillator, which operates in conjunction with both picture and sound sections; the sound i-f channel, which consists of i-f amplifiers limiter and FM detector; and reactance tube, which controls frequency of local oscillator. When the local oscillator drifts in frequency and beats against the incoming signal, an i-f frequency is produced which is not at the center of the i-f bandpass for the sound section. We have learned that when the center frequency of the signal applied to the discriminator is not at the resonant frequency of the tuned circuit, a d-c potential is produced in the output of the discriminator. A function of the reactance tube is to take this d-c potential in accordance with its polarity and amplitude and cause the frequency of the local oscillator to return to the correct value, at which point the i-f frequency present in the plate circuit of the mixer is the frequency to which the sound i-f channel is resonant. The a-f-c system operates in conjunction with the sound i-f system because of the comparatively sharper bandpass compared to the picture channel. Actually, if we keep the local oscillator on frequency in connection with the sound channel it automatically is held at the proper value to produce the correct picture i-f frequency distribution.

REACTANCE TUBES

The reactance tube introduces a reactive component of current into the oscillator tuned resonant circuit. This current is in phase with either the inductive or capacitative current (depending on reactance tube circuit) already flowing in the tank circuit, adding to one and subtracting from the other. Thus, the frequency of the oscillator changes to a point at which i_l is equal to i_c again. The amount that the frequency is varied can be controlled by changing the amplitude of the reactive current which the reactance tube introduces into the resonant circuit.

A basic reactance tube circuit (Fig. 203) consists of the oscillator, the reactance tube, and a phase-shifting network. The function of the phase-shifting network is to take the plate or tank-circuit voltage of the oscillator and shift it in phase 90 degrees before applying it to the grid of the reactance tube. Thus, the reactance tube plate current will also be shifted 90 degrees in phase with respect to the oscillator plate voltage. This plate current passes to the tank circuit and because it is reactive (shifted 90 degrees) changes the resonant frequency of the oscillator.

The sequence which brings about the change in oscillator frequency is as follows (first drawing, Fig. 204):

1. Zero phase is represented by the oscillator plate voltage e_p.

2. This voltage is applied across the series phase-shifting network of *R1* and *C1*. If this branch is dominantly resistive the series current i_s which flows is in phase with the plate voltage e_p. This relation is maintained by keeping the resistance of *R1* at least ten times greater than the reactance of *C1* at the frequency of the oscillator.

3. The series current i_s flowing in the series phase-shifting network develops a voltage across capacitor alone which lags the current that produces it by 90 degrees, causing the grid voltage of the reactance tube e_g to lag the current i_s by 90 degrees. Grid voltage e_g therefore lags e_p by 90 degrees.

4. Since reactance tube plate current is in phase with its grid voltage the plate current i_p also lags the oscillator plate voltage e_p by 90 degrees. Thus, the current introduced into the resonant circuit by the reactance tube through capacitor *C4* lags the tuned circuit voltage by 90 degrees and is therefore an inductive current causing reactance tube to act as an effective inductance.

5. The current i_l, which flows in the inductive branch of the resonant circuit, of course lags the tank voltage.

6. Likewise, the current i_c, which flows in the capacitive branch, leads the tank voltage by 90 degrees.

7. The reactive component of current introduced by the reactance tube, because it is lagging by 90 degrees, is in-phase with i_l and adds, and is 180 degrees out-of-phase with respect to i_c and subtracts. To equalize the currents (resonance) again it is necessary that the oscillator frequency increase to raise the reactance of the inductive branch and decrease the reactance of the

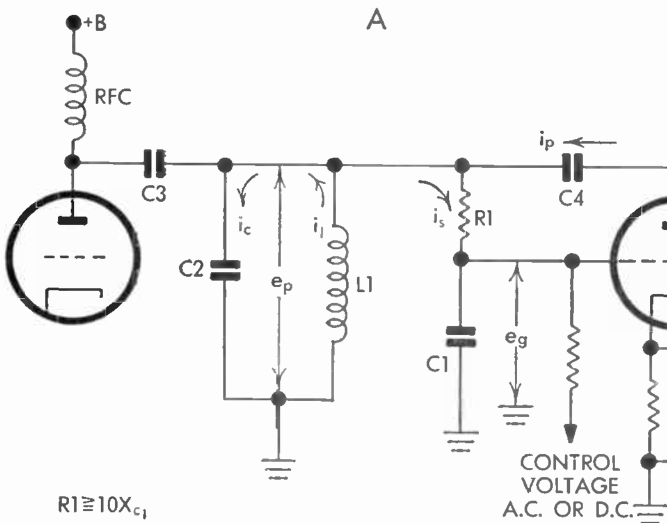


FIG. 203 Basic Reactance-Tube Circuit

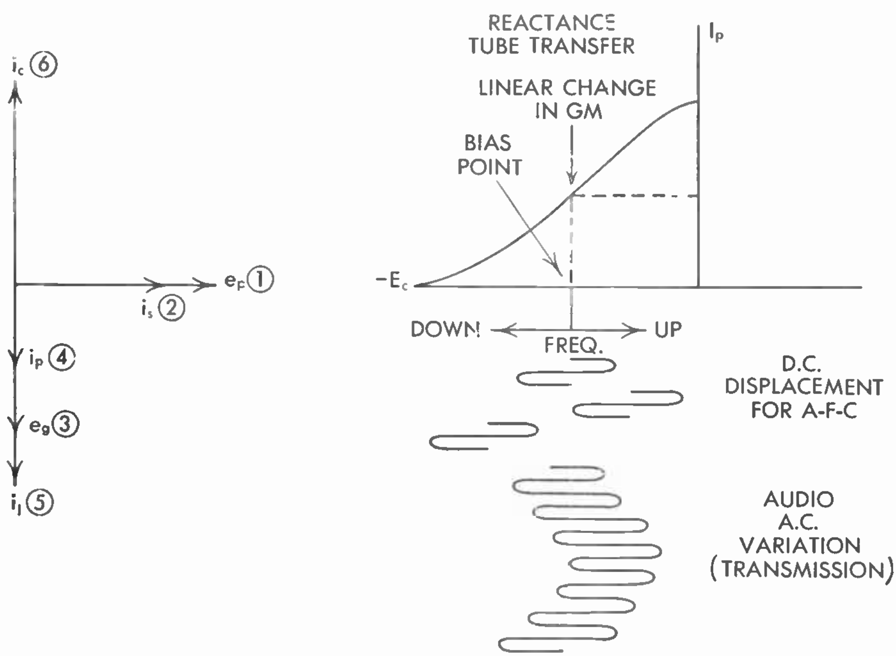


FIG. 204 Basic Reactance-Tube Relations

capacitive branch to the point at which i_l once more equals i_c. Thus, the new resonant frequency of the oscillator is higher because of the presence of the reactive current from the reactance tube.

8. The amount the oscillator shifts in frequency is dependent on the level of the reactance-tube-current (r-f component). For example, if the current is increased by raising the reactance tube grid voltage the oscillator frequency rises still further to find the point at which the tank currents will equalize. In practice the reactance tube is biased on a portion of its characteristic on each side of which a linear change in g_m is obtained. Therefore, its presence in the circuit along with the constants of the oscillator-tuned circuit sets the resting frequency of the oscillator. Now if the grid bias on the reactance tube decreases, the oscillator frequency will change in one direction (in our example it will rise) and if the grid bias is increased the frequency of the oscillator will go in the other direction (in our example it will fall). The change in reactance-tube grid bias in a typical television a-f-c system is caused by a shift in the d-c component of the discriminator output whenever the i-f center frequency drifts away from the correct value. This change in voltage changes the reactance-tube g_m and therefore its plate current, which in turn shifts the frequency of the oscillator until the oscillator frequency, beating with the incoming signal, once more produces the correct i-f frequency.

9. In a-f-c practice, only the d-c component of the reactance-tube grid bias is changed by the discriminator voltage. A reactance tube is also used to frequency-modulate the center-frequency oscillator with audio at the FM transmitter. In this case the grid bias of the reactance tube is varied at an audio rate, and consequently the frequency of the oscillator also changes at an audio rate.

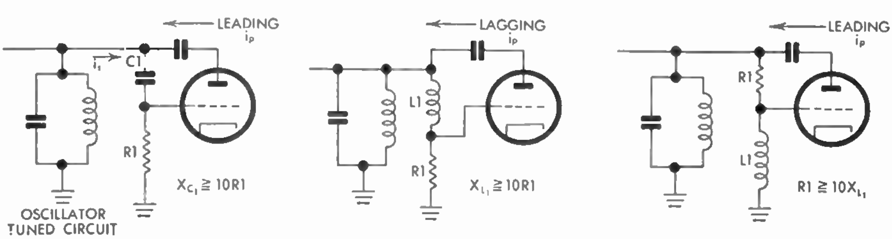


FIG. 205 Reactance-Tube Phase Shifters

There are a number of other phase-shifting combinations which can be used in the input circuit of a tube connected to operate as a reactance tube. For example, in the first drawing of Fig. 205, a series *R-C* combination is again used, but the positions of the resistor and capacitor have been reversed. In this type the reactance of *C1* at the resonant frequency of the oscillator must be at least ten times the value of the resistor. Thus, the series current leads the tank voltage and causes the reactance-tube grid voltage to do the same. Reactance tube introduces a leading 90-degree current into the tuned circuit; and therefore it presents a capacitive branch instead of an inductive one like the previous type. This means when the reactance-tube grid bias decreases the oscillator frequency will decrease because frequency varies inversely with reactance in a capacitive branch. Series *RL* combinations can also be used to obtain the proper phase shift.

In summary, a number of considerations must be made in the choice of discriminator, reactance-tube type, and oscillator frequency in the design of an a-f-c system. For example, if the local oscillator is tuned above the signal frequency and it attempts to drift higher, the i-f center frequency will also drift higher. Now if the output of the discriminator swings plus when the center frequency drifts high a reactance tube must be chosen which causes the oscillator frequency to decrease when its grid is driven plus. In this arrangement the second type of reactance tube would be satisfactory because frequency decreases as its grid voltage rises. If the local oscillator were tuned on the low-frequency side of the signal frequency, the first type of reactance tube discussed would be satisfactory because a rise in reactance-tube grid voltage causes a rise in frequency.

112. *Typical Automatic Frequency Control Systems*

A typical a-f-c system, used by Philco, is shown in Fig. 206. In the Philco receiver the sound is taken off at the output of the input i-f stage and is fed to a two-stage sound i-f amplifier. A ratio FM detector is used to remove the frequency modulation from the sound i-f carrier. Audio output of the ratio detector is coupled through the de-emphasis network to a two-stage audio amplifier. The d-c component of the discriminator output is taken off at the same point as the audio and fed through two long time-constant filter networks to the grid of the reactance- or oscillator-control tube.

In addition to the d-c voltage applied to the grid of the reactance tube from the ratio detector, the grid must also be excited by a portion of the oscillator plate voltage. Thus, a portion of the a-c oscillator voltage which appears across capacitor *C3* is coupled to the grid of the reactance tube through capacitors *C1* and *C2*. The actual a-c component of the reactance-tube grid voltage is developed across grid resistor *R2* and tube input capacity. Proper division of voltages for the high and the low band necessitates the two capacitors *C1* and *C2* and resistor *R1*. These ensure the proper level of feedback voltage and reactive plate current for both the high- and the low-band set of television channels. An a-f-c test jack, connected at the point where the d-c component of discriminator voltage is obtained, forms a convenient point from which alignment and performance checks can be made.

A reactance tube is also used in the automatic sync-lock circuit in the RCA receivers, as shown in Fig. 207. In this circuit, the phase-splitting *RC* combination consists of capacitor *C1* and resistor *R1* (which is only 10 ohms) and is connected between cathode and ground of the reactance or oscillator-con-

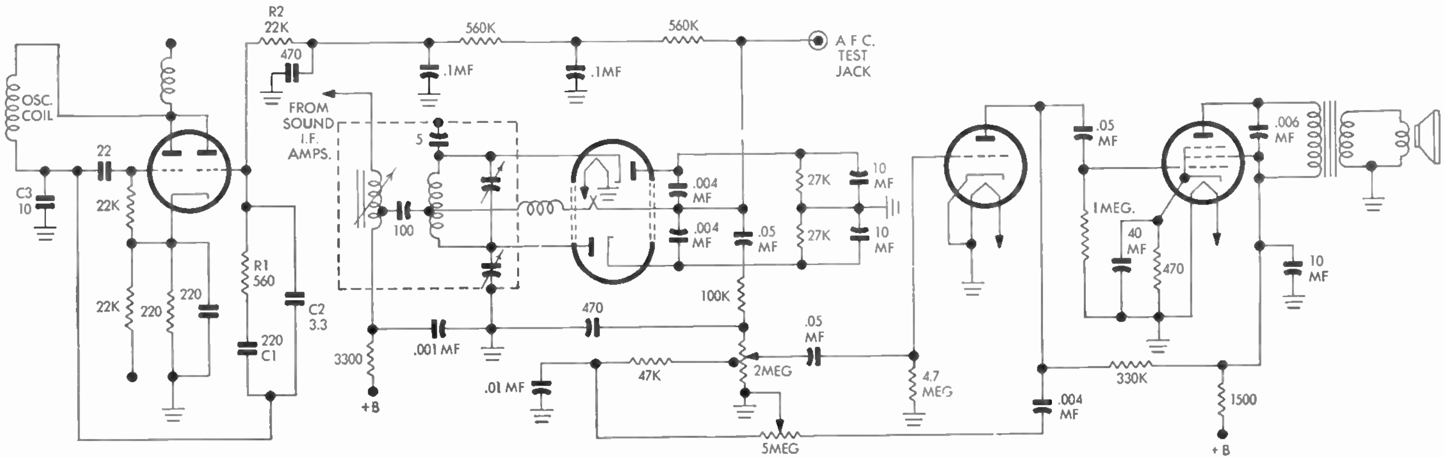


FIG. 206 Philco A-F-C and Audio System

trol tube. Thus, the oscillator voltage which produces the reactive plate current in the output of the reactance tube is applied to the cathode, and the d-c component from the sync discriminator stage is applied to the grid of the reactance tube. Isolation between two signal application points means that a very long time constant can be used in the grid circuit of the reactance tube to prevent abrupt changes in the d-c component of the discriminator output from affecting the frequency of the horizontal oscillator. In this a-f-c system the hori-

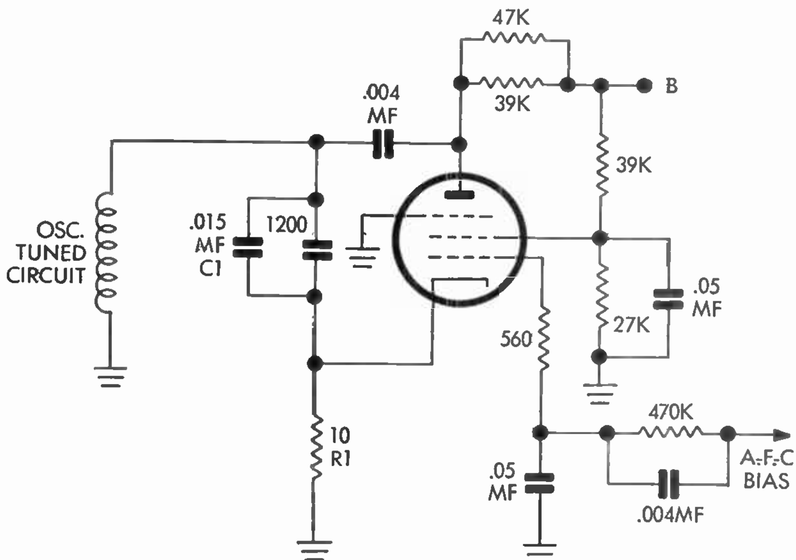


FIG. 207 Reactance Tube of RCA Sync Control

zontal oscillator is held on the correct frequency with respect to the rate of arrival of the horizontal sync pulses. Whenever there is a phase displacement between arriving sync pulses and the horizontal oscillator since wave, a d-c component appears at the output of the discriminator and, therefore, on the grid of the reactance tube in order to change the value of the reactive plate current applied by the reactance tube. Once again the oscillator is set on the correct frequency.

113. *Typical Sound Systems*

The Philco sound channel, Fig. 208, consists of a single i-f amplifier, a ratio detector, and a two-stage audio amplifier. A special discriminator test jack is provided for ease of alignment of the sound system as well as other portions of the television receiver. Ratio-detector output through a suitable de-emphasis network applies the audio signal to the grid of the first audio amplifier. A remote-volume control connection is made at *J400*, as well as a control feedback connection from the second winding on the secondary of the audio-output transformer. The feedback is of proper polarity to correct frequency response over an extended range and, at the same time, to minimize the influence of hum in the audio system. A two-stage audio amplifier, consisting of a resistance-coupled triode and a beam-power pentode, drives the loudspeaker of the television receiver.

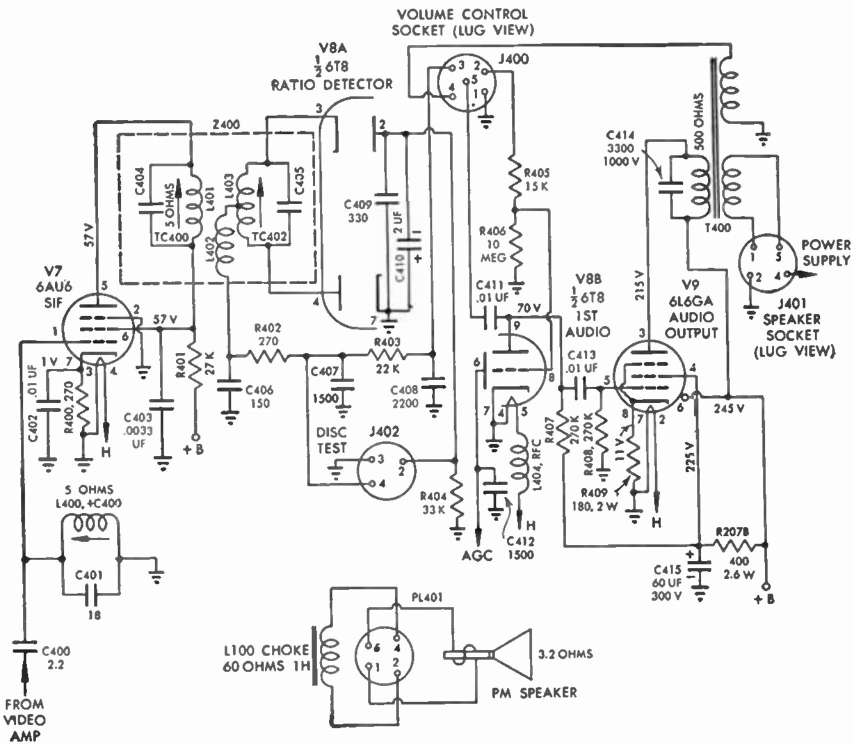


FIG. 208 Philco Sound System

Simplification of the sound system in the Westinghouse television receiver is obtained through use of a 6BN6 gated-beam tube as the FM detector and a 6BK5 beam-power pentode as the audio-output amplifier. As shown in Fig. 209, only one sound i-f amplifier stage is required to drive the 6BN6. The relatively high audio-output voltage of the 6BN6 in conjunction with the high power-sensitivity of the 6BK5 eliminates the need of an intermediate audio amplifier.

In a 6BN6 tube, Fig. 209a, the electron flow is in a vertical beam from the cathode to the plate through the various shield arrangements and control elements. The limiter (or signal) grid and the quadrature grid serve as control elements, while the accelerator and the plate are operated at positive potentials.

The positive potential on the accelerator structure attracts the electrons as they leave the cathode. Whether or not the electrons pass through both openings in the accelerator structure and move toward the plate depends on the limiter-grid voltage. If the limiter grid is more than a few volts negative, electron flow toward the plate is completely cut off, and the electrons flow to the accelerator structure. If the limiter grid is less negative than the cutoff value, the electron beam passes readily through the limiter grid and continues on

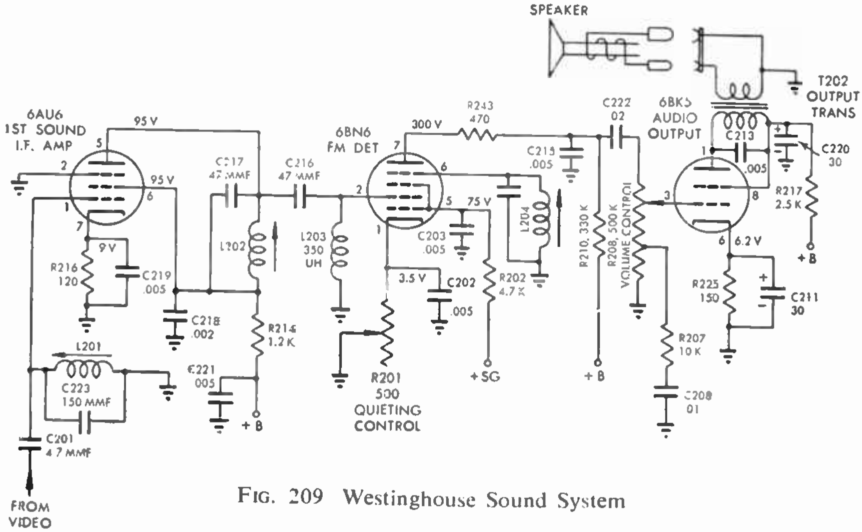


FIG. 209 Westinghouse Sound System

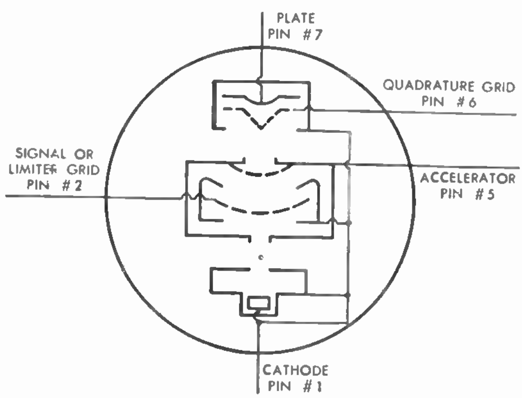


FIG. 209a Grated-Beam Tube

toward the plate. Before the electron beam reaches the plate, it encounters another control element—the quadrature grid. Again, the electron flow toward the plate can be cut off if the quadrature grid is more than a few volts negative. If this is the case, the electrons again return to the accelerator structure. If, however, the quadrature grid is not sufficiently negative to cause plate-current cutoff, the electron beam continues on and reaches the plate. Since both the limiter and quadrature grids can change the plate current from cutoff to saturation with only a small amount of voltage change on either grid, each grid can be considered as a voltage-controlled gate.

The 4.5-megacycle signal from the sound i-f amplifier is applied to the limiter grid of the 6BN6. This signal is approximately 1.25 volts root-mean-square when the weakest usable television signal is received. A signal voltage of 1.25 volts root-mean-square or greater at the limiter grid drives the tube from plate-current cutoff to plate-current saturation, thus producing a square-wave variation in beam current in the region beyond the limiter grid.

A parallel resonant circuit tuned to the center frequency of the FM signal (4.5 megacycles) is connected to the quadrature grid. Through space-charge coupling, some energy from the square-wave pulses of the electron beam transfers to the quadrature-grid circuit. When a center-frequency signal is applied to the limiter grid, the voltage developed across the quadrature-grid resonant circuit lags the input signal by approximately 90 degrees. As a result of this phase difference, the plate current is cut off by the quadrature-grid voltage during one-half of the time that the limiter grid allows electron flow. Therefore, only one-half of each square-wave pulse developed reaches the plate of the tube. When the signal swings higher in frequency than 4.5 megacycles, the quadrature resonant circuit has a reactive component, and the voltage at the quadrature grid lags the input signal by *less* than 90 degrees. Under this condition, less than half of the square-wave pulse reaches the plate, and the average value of plate current is less than when the signal is at center frequency. Correspondingly, when the signal swings lower in frequency than 4.5 megacycles, the phase difference is *more* than 90 degrees, and more than half of the square-wave pulse reaches the plate. The average plate current is then greater than when the signal is at the center frequency. From this, it is apparent that the modulation content of the FM signal is available at the plate of the tube as variations in the average plate current.

A signal voltage of 1.25 volts root-mean-square at the limiter grid is sufficient to drive the tube from plate-current cutoff to plate-current saturation. With the tube operating between cut-off and saturation, higher signal inputs cannot increase the plate current. Therefore, amplitude-limiting occurs when the signal voltage at the limiter grid exceeds 1.25 volts, and the detector is insensitive to amplitude fluctuations (AM). To obtain these conditions, the tube must be biased in such a way as to operate on the correct portion of the characteristic curve. The quieting control *R201* provides a means of adjusting the bias for maximum AM rejection.

QUESTIONS

1. Compare various signals which constitute an AM wave with those that make up an FM wave.
2. How is FM created by phase modulation?
3. Differentiate between direct and indirect FM.

4. Why is a dual-multiplier system needed in an Armstrong transmitter?
5. What advantages does FM have to offer over AM?
6. If audio frequency is decreased (amplitude the same), what happens to the number of sideband pairs?
7. If audio frequency is decreased (amplitude the same), what happens to the bandwidth?
8. If audio frequency is decreased (amplitude the same), what happens to deviation?
9. Explain significance of FM noise triangle.
10. How are amplitude noises suppressed in FM system?
11. How does pre-emphasis at transmitter improve signal-to-noise ratio at receiver?
12. Explain operation of a typical limiter.
13. Discuss importance of tuned transformer characteristics in an FM discriminator.
14. Compare in detail the theory of operation of conventional discriminator and ratio detector.
15. What are the advantages of a-f-c?
16. Explain Philco a-f-c system.
17. Discuss in detail operation of a reactance tube.
18. What is dynamic range?
19. What causes a "boomy" bass?
20. Discuss FM systems in television receivers.

INDEX
PART I*

Boldface numbers refer to entries in Volume II of *Television for Radiomen.* (Volume II begins on page 338. Part II, which is devoted to color television, follows page 662 in Volume II.)

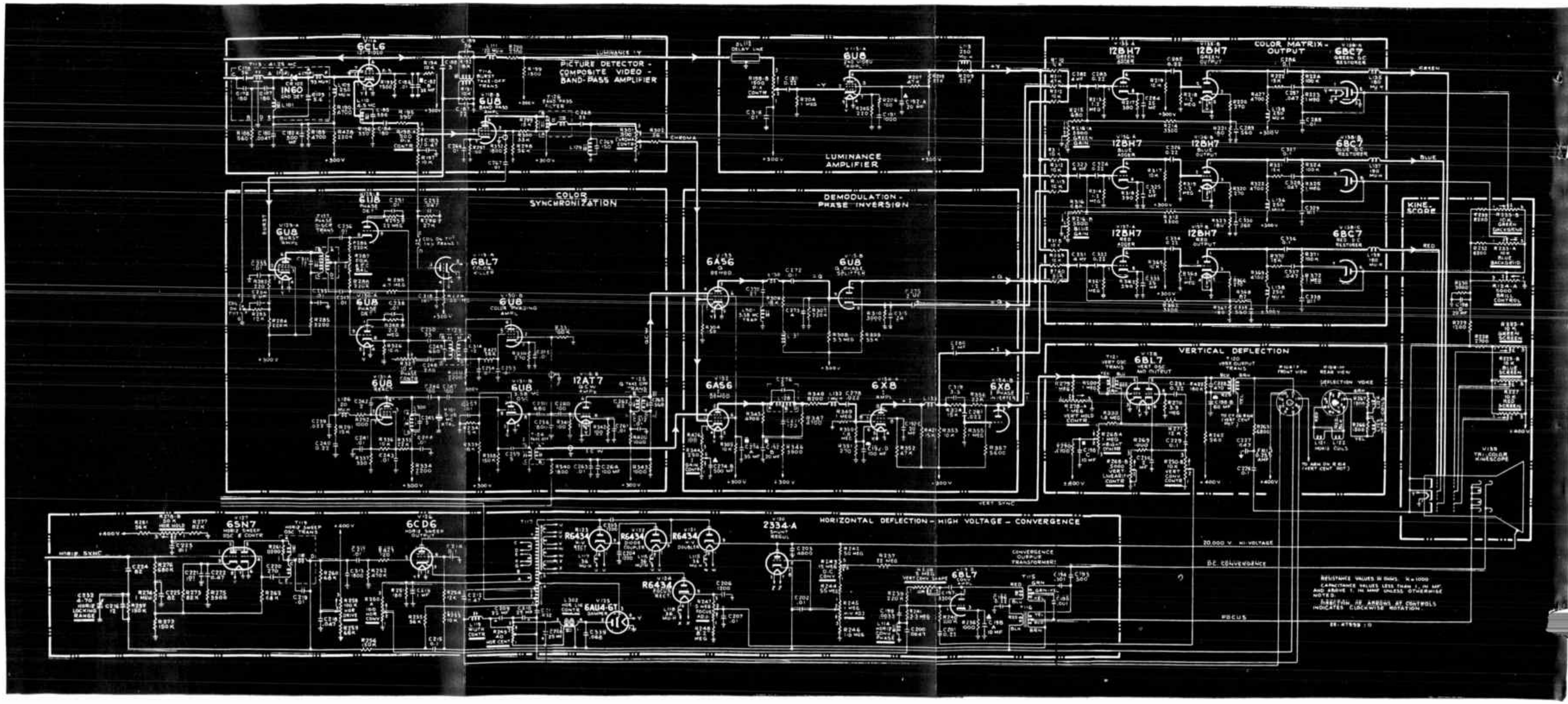
* Index to pages 1–**662** of Part I. Index to pages **1–93** of Part II follows Index to Part I.

INDEX
PART II*

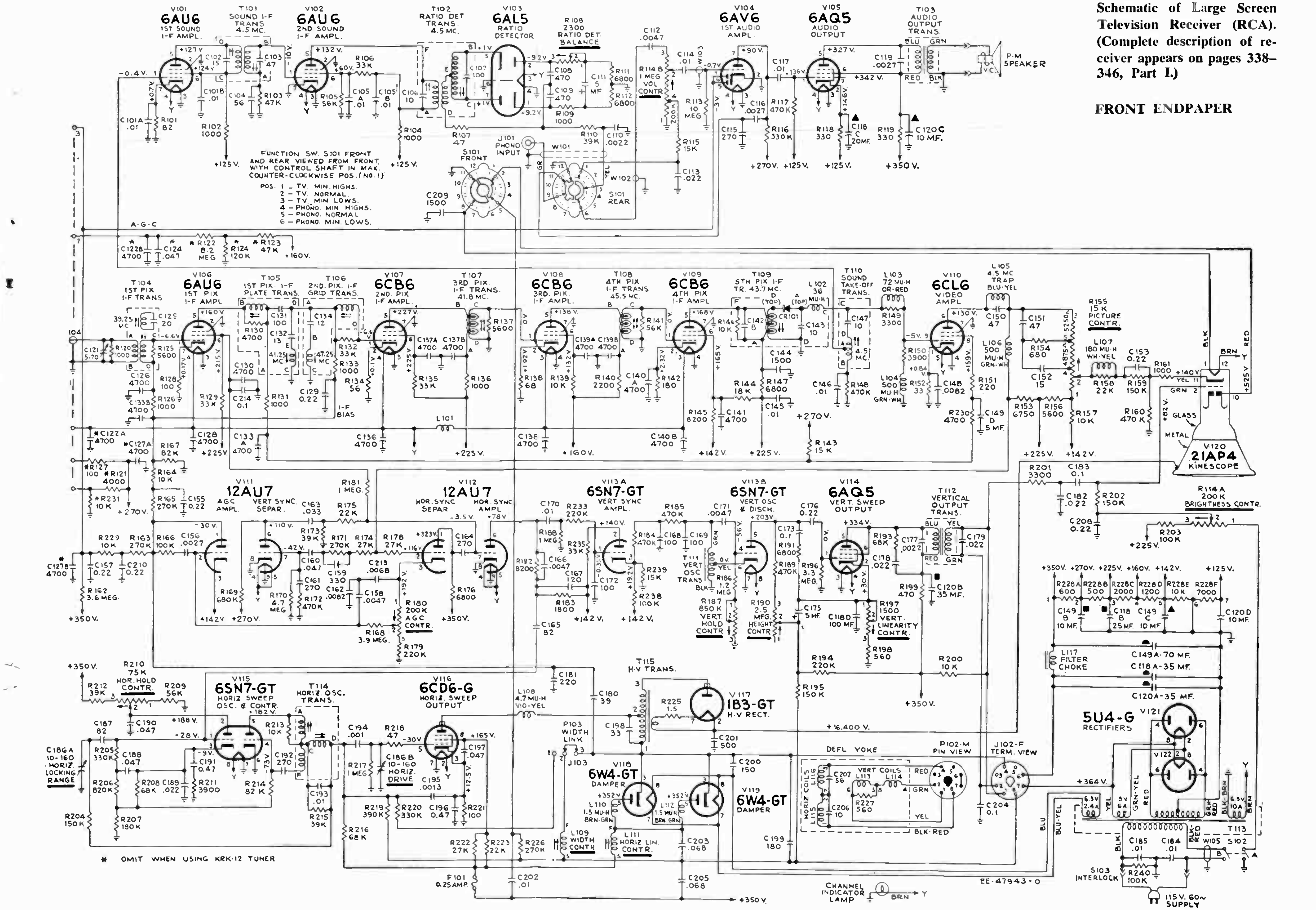
Boldface numbers refer to entries in Volume II of *Television for Radiomen.* (Part II, which is devoted to color television, follows page 662 in Volume II.)

* Index to pages 1–662 of Part I precedes index to pages 1–93 of Part II.

REAR ENDPAPER



Schematic of Color Section of a Color Receiver (RCA). (Complete description of receiver appears on pages 63–72, Part II.)



Schematic of Large Screen Television Receiver (RCA). (Complete description of receiver appears on pages 338–346, Part I.)

FRONT ENDPAPER