Basic Television

Bernard Grob

THIRD EDITION

3RD EDITION

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PRINCIPLES AND SERVICING

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Basic Television: PRINCIPLES AND SERVICING

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To Ruth and Harriet Deborah

Preface

This book presents a comprehensive course in black-and-white and color television for technicians and servicemen. The practical explanations of television principles and receiver circuits are planned for the beginning student in television. In addition to the troubleshooting principles helpful in repairing receivers, each circuit section is analyzed as an application of basic electronic principles that apply to all types of electronic equipment. This method of presentation, based on experience in teaching television and electronics for many years, has proved very successful in helping students.

The book is designed as a text for television courses that follow a course in radio or electronics fundamentals. Similarly, Chapters 24 and 25 on color television can be used for a short course in color television receivers after the students have learned the essentials of black-and-white television.

The explanations are essentially nonmathematical, but quantitative work with some simple algebra is included where it helps to illustrate specific examples for applications of important fundamentals. Topics analyzed quantitively are wide-band amplifiers applied to practical design of video amplifiers, self-induced voltage applied to horizontal deflection circuits, resonance applied to i-f amplifiers and *RC* time constant applied to deflection oscillators.

On the practical side, all examples are from typical receivers; many photos are used to show actual components, including printed-wiring boards; oscilloscope photos illustrate typical waveshapes; and common troubles are shown by photos of the kinescope screen. Each chapter on receiver circuits concludes with essential troubleshooting principles useful for localizing troubles. More details with techniques in use of the oscilloscope, meters, and signal generators are given in Chapter 23 on Receiver Servicing. Tables 23 · 2 to 23 · 5 summarize nearly all the common troubles in television receivers.

Summary of contents. The third edition, like the first two, is organized to present the basic material first, thus allowing continuity from topic to topic. At the start, the discussion of the fundamental principles of television, Chapters 1 to 5, explains the video signal, the scanning procedure, and how the signal is transmitted. Then the general requirements of television receivers and details of picture tubes are described in Chapters 6 and 7. With this basis, a detailed analysis of each type of receiver circuit is given in separate chapters. Circuits for video, synchronization, and deflection are analyzed first because they represent the most important parts of television circuitry. These circuits are emphasized by two chapters dealing with video amplifiers, two chapters on synchronization, and three chapters on vertical and horizontal deflection.

For each section of the receiver, the general requirements of a basic circuit are explained and illustrated by typical commercial schematic diagrams. The reason why a circuit operates as it does is explained in this manner. Also, the television receiver can then be analyzed as an application of the fundamental circuits important in all fields of electronics. Since the sound is an FM signal, the theory of FM and receiver requirements are included.

Reorganization of topics. Most of the material has been rewritten to make the third edition completely up-to-date. The important subjects of color television, synchronizing circuits, and deflection circuits are emphasized with additional chapters. The fundamentals of electronic circuits are applied to the analysis of television receivers. Specific examples are:

1. The coverage of color television has been expanded to two chapters. The overall system, described in Chapter 24, now contains material on $R-Y$, $G-Y$, and $B-Y$ color video signals. The new Chapter 25, Color Television Receivers, explains kinescope setup with purity adjustments and convergence techniques. More details of color receiver circuits are included, with additional details on the operation of synchronous demodulators.

2. The topic of deflection output circuits has been expanded with one chapter devoted to vertical deflection and another to horizontal deflection. Thus the specific requirements of each can be analyzed more thoroughly. A quantitative analysis of horizontal output circuits shows exactly how much flyback high voltage and boosted B+ is produced.

3. A separate chapter, Horizontal AFC Circuits, contains a detailed analysis of the four main types.

4. The coverage of *RC* and *RL* circuits, including time constant and transient response, is in the Appendix. Specific formulas for *RC* charge and discharge, using common logarithms, are given here so that calculations can be made to determine exact deflection waveshapes and oscillator frequency.

5. Deflection waveshapes are analyzed more definitely in terms of average d-c level and peak-to-peak a-c component. This method provides concrete examples, which clarify the principles of how the waveforms are generated and shaped.

The theory of FM has been combined with FM receiver circuits. More details of operation for the quadrature-grid detector are in Chapter 22, The FM Sound Signal. Similarly, the electron scanning beam has been incorporated in Chapter 7, Picture Tubes, where the theory has direct application. As a result, the student can progress more quickly from the general requirements in the first five chapters to the practical aspects of receivers, picture tubes, and circuits in the remaining chapters.

New topics. The growing importance of color television, UHF tuners, remote tuners, world-wide television broadcasting, and transistor circuits are the major new developments; all are covered in the third edition. Specific new features are:

1. Each chapter on receiver circuits has a section at the end that describes how transistors can be used instead of vacuum tubes.
2. UHF tuners are included in Chapter 20, The R-F Tuner, because new TV receivers must include provision for receiving both VHF and UHF channels.
3. A new section on remote tuners is also included in Chapter 20. This section explains how supersonic waves are used for selecting channels from a remote position.
4. A detailed treatment of color television receivers and tricolor kinescopes is presented in Chapter 25.
5. With the advent of Telstar and world-wide television broadcasting, the television standards of foreign countries gain new interest. These are listed in Appendix C.

Teaching aids. Many drawings and photos of simplified circuits, typical waveshapes, and actual components illustrate the principles with concrete examples that make the discussion easier to understand. Tables and graphs are used to summarize useful information and emphasize important comparisons. Most of the text has been rewritten in shorter sentences and paragraphs to make this edition easier to read.

On the basis of successful experience with self-testing material, each chapter now concludes with:

1. Self-examination questions, including multiple-choice, true-false, fill-in, and matching questions. Answers are given at the back of the book.
2. More thorough and specific essay questions for review, including functions of components and troubleshooting questions for practical problems, plus graphical analysis for a better understanding of electronic principles.
3. A separate group of numerical problems for practice in quantitative analysis. Most of these problems deal with the fundamentals of electronic circuits applied to television receivers. Examples include *RC* time constant, decibels, induced voltages, resonance, and a-c impedance. Answers to the odd-numbered problems are found at the back of the book.

These self-testing exercises are helpful in organizing class work and reviewing the most important features of each topic. In addition, the book is suitable for self-study, since the student can check his own progress. The Bibliography at the end of the book lists helpful references for more information on television, transistors, and electronic circuits.

Credits. The schematic diagrams and photographs have been made available by many organizations, as noted in each illustration. This courtesy is gratefully acknowledged. Especially helpful are the color photos in Plates I to VI, which are from the *RCA Color Television Picto-O-Guide*, by John R. Meagher, published by the RCA Electron Tube Division. Photographs have also been used from the Television Servicing Course of the Home Study School of RCA Institutes, Inc.

For the final credit, it is a pleasure to thank my wife Ruth for her excellent work in typing the manuscript.

Bernard Grob

Contents

Chapter 1 The television system

Television means "to see at a distance." Our practical television system is a method of transmitting and receiving a visual scene in motion by means of radio broadcasting. The sound associated with the scene is transmitted at the same time, to provide a complete sight and sound reproduction at the receiver of the televised program.

Although the end result required is a motion picture, television is basically a system for reproducing a still picture such as a snapshot (Fig. 1 · 1*a*). Many of these are shown one after the other in rapid sequence during each second, to give the illusion of motion. Therefore, the first requirement of the television system is that it be capable of transmitting and receiving a simple still picture.

1 · 1 Picture elements

A still picture is fundamentally an arrangement of many small dark and light areas. In a photographic print, fine grains of silver provide the differences in light and shade needed to reproduce the image. When a picture is printed from a photoengraving there are many small black printed dots in the reproduction, which form the image. Looking at the magnified view in Fig. 1 · 1*b*, we can see that the printed picture is composed of small elementary areas of black and white. This basic structure of a picture is evident in newspaper photographs. If they are examined closely, the dots will be seen because the picture elements are relatively large.

Each small area of light or shade is called a *picture element.* All the elements contain the visual information in the scene. If they are transmitted and reproduced in the same degree of light or shade as the original and in proper position, the picture will be reproduced.

As an example, suppose that we want to transmit an image of the black

(a)

(b)

Fig. 1 · 1a
A still picture.

Fig. 1 · 1b
Magnified view of still picture.

cross on a white background shown at the left in Fig. 1 · 2 to the right side of the figure. The picture is divided into the elementary areas of black and white shown. Picture elements in the background are white, while the elements forming the cross are black. When each picture element is transmitted to the right side of the figure and reproduced in the original position with its shade of black or white, the original image is duplicated.

1 · 2 Transmitting and receiving the picture information

In order to transmit and reproduce the visual information corresponding to a picture element, the television system requires a camera tube and an image-reproducing tube. The camera tube is a photoelectric tube that produces electric signal corresponding to the visual information in a picture element. The image-reproducing tube is the picture tube, or *kinescope*, in the receiver. The picture tube converts the signal voltage back into a

Fig. 1 · 2 Reproducing a picture by duplicating its picture elements.

visual image on its screen, which duplicates the original picture elements. If we compare this with a system for sound, the camera tube corresponds to the microphone at the broadcast station and the picture tube corresponds to the loudspeaker at the receiver.

A camera tube, such as the iconoscope in Fig. 1 · 3, can be used to convert the light image into electric signal. The three main types of camera tubes are the iconoscope, image orthicon, and vidicon, which are described in detail in Chap. 2. The iconoscope is an older type, of relatively simple construction. In Fig. 1 · 3, note the image plate about 3 by 4 in. consisting of many small globules of photosensitive material coated on the surface.

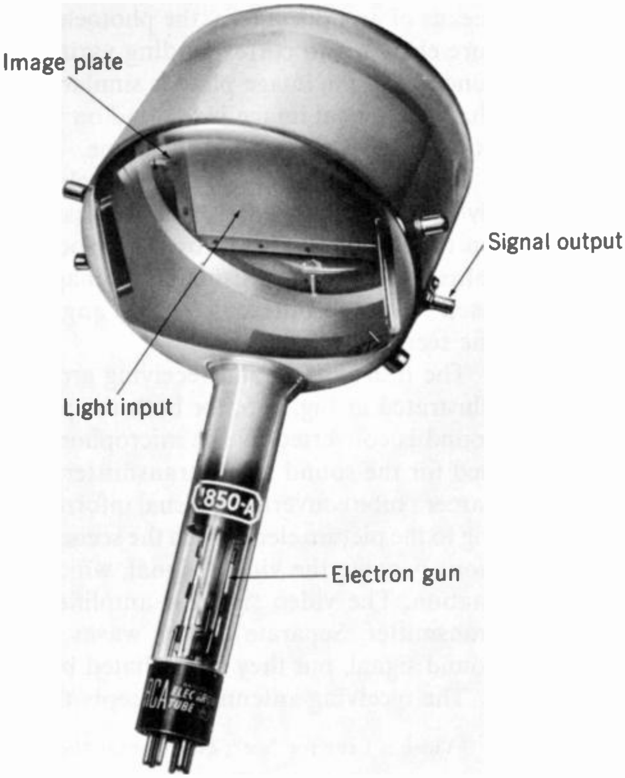


Fig. 1 · 3
Iconoscope camera tube. The electron gun produces a beam of electrons aimed at the photoelectric image plate. (RCA.)

Fig. 1 · 4 Television studio scene. (National Broadcasting Co.)

The amount of photoemission from each point depends on its light intensity. When the entire scene to be televised is focused on this image plate by means of an optical lens, the photoelectric globules can convert all the picture elements to corresponding variations in signal output. Note that the function of the image plate is similar to the film in an ordinary camera, in that the optical image is focused on it. If you can look at the image plate, you will see the image of the scene.

Figure 1 · 4 shows how several television cameras are used to televise a live show in the studio of a broadcast station. Also note the microphone on the boom overhead for the associated sound. Each camera unit contains a camera tube, usually the image orthicon type. Several cameras are used to obtain different camera angles, with close-ups and long shots of the scene.

The transmitting and receiving arrangement for the television system is illustrated in Fig. 1 · 5, for both the picture and sound signals. The desired sound is converted by the microphone to an audio signal, which is amplified for the sound signal transmitter. For transmission of the picture, the camera tube converts the visual information into electric signals corresponding to the picture elements in the scene being televised. These electrical variations become the video[1] signal, which contains the desired picture information. The video signal is amplified and coupled to the picture signal transmitter. Separate carrier waves are used for the picture signal and sound signal, but they are radiated by one transmitting antenna.

The receiving antenna intercepts the radiated picture and sound carrier

[1] Video is Latin for "see"; audio means "hear."

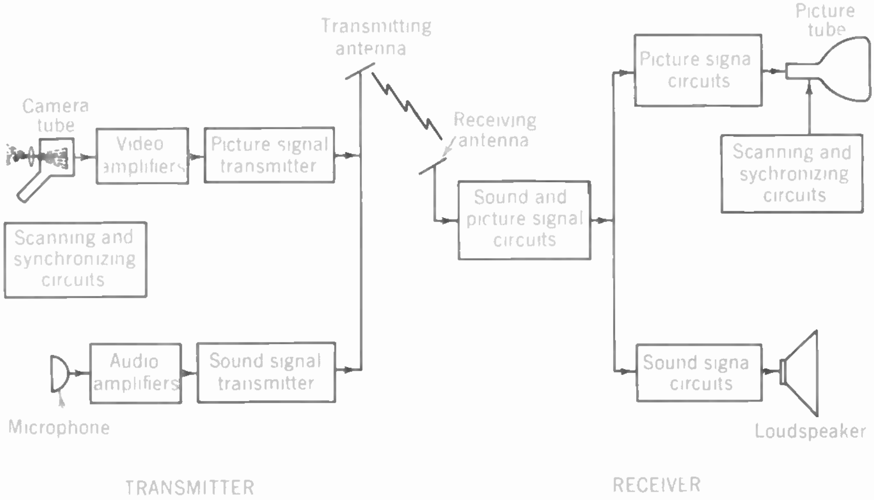


Fig. 1 · 5 Block diagram of the television system.

Fig. 1 · 6 A typical picture tube. The electron gun in narrow part of tube produces a beam of electrons that is accelerated to the phosphor screen. (Sylvania Electric Products, Inc.)

signals, which are then amplified and detected in the receiver. The detector output includes the desired video signal containing the information needed to reproduce the picture. Then the recovered video signal is amplified and coupled to an image-reproducing tube that converts the electric signal back into picture elements in the same degree of black or white.

The picture tube in Fig. 1 · 6 is very similar to the cathode-ray tube used in the oscilloscope. The glass envelope contains an electron-gun structure to produce a beam of electrons aimed at the fluorescent screen. When the electron beam strikes the screen, it emits light, and the screen becomes bright in proportion to the intensity of the electron beam. Varying the intensity of this beam by varying the voltage applied to the control grid of the tube changes the intensity of the spot of light on the screen. When signal voltage makes the control grid of the kinescope less negative, the beam current is increased, making the spot of light on the screen brighter. More negative grid voltage reduces the brightness. If the grid voltage is negative enough to cut off the electron-beam current of the kinescope, there will be no light. This corresponds to black.

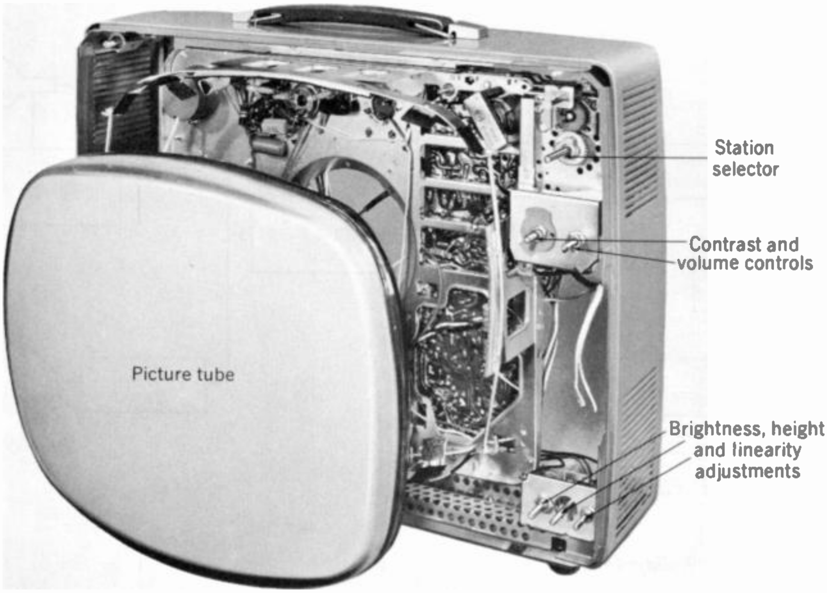


Fig. 1·7 Front view of portable television receiver, with cover off. (RCA.)

The video signal voltage corresponding to the desired picture information is coupled to the grid-cathode circuit of the picture tube so that the picture elements can be reproduced on the kinescope screen. Then the entire picture can be viewed through the clear glass faceplate. Figure 1·7 shows a picture tube mounted on the receiver chassis.

1·3 Scanning

In order to produce video signal for all the elements in the picture it is scanned by the electron beam, one element at a time, in sequential order. The scanning is done in the same way you read a written page to cover all the words in one line and all the lines on the page (see Fig. 1·8). Starting at the top left, all the picture elements are scanned in successive order, from left to right and from top to bottom, one line at a time. This method, called *horizontal linear scanning*, is used in the camera tube at the transmitter to dissect the image into picture elements and in the kinescope at the receiver to reassemble the reproduced image.

The sequence for scanning all the picture elements is as follows:

1. The electron beam sweeps across one horizontal line, covering all the picture elements in that line.
2. At the end of each line the beam is returned very quickly to the left side to begin scanning the next horizontal line. No picture information is scanned during this retrace time, as both camera tube and picture tube are blanked out for this period. The retraces must be very rapid, therefore, since they are wasted time in terms of picture information.

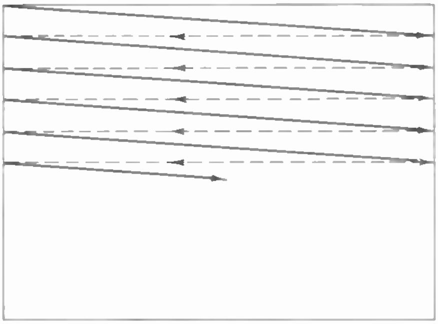


Fig. 1 · 8 Horizontal linear scanning. Solid lines for trace left to right; dashed lines indicate retrace from right to left.

3. When the beam is returned to the left side, its vertical position is lowered so that the beam will scan the next lower line and not repeat over the same line. This is accomplished by the vertical scanning motion of the beam, which is provided in addition to horizontal scanning.

The number of scanning lines for one complete picture should be large in order to include the highest possible number of picture elements and, therefore, more picture details. However, other factors limit the choice, and it has been standardized at a total of 525 scanning lines for one complete picture or frame. This is the optimum number of scanning lines per frame for the standard 6-Mc bandwidth of the television broadcast channels.

Note that the beam moves slowly downward as it scans horizontally. This vertical scanning motion is necessary so that the lines will not be scanned one over the other. The horizontal scanning produces the lines left to right while the vertical scanning spreads the lines to fill the frame top to bottom.

1 · 4 Motion pictures

With all the picture elements in the frame televised by means of the scanning process, it is also necessary to present the picture to the eye in such a way that any motion in the scene appears on the screen as a smooth and continuous change. In this respect the television system is very similar to motion-picture film practice.

Figure 1 · 9 shows a strip of motion-picture film. Note that it consists of a series of still pictures with each picture frame differing slightly from the preceding one. Each frame is projected individually as a still picture, but they are shown one after the other in rapid succession to produce the illusion of continuous motion of the hammer in this scene.

In standard commercial motion-picture practice, 24 frames are shown on the screen for every second during which the film is projected. A shutter in the projector rotates in front of the light source and allows the film to be projected on the screen when the film frame is still, but blanks out any light from the screen during the time when the next film frame is being

Fig. 1·9 Strip of motion-picture film. (Eastman-Kodak Co.)

moved into position. As a result, a rapid succession of still-film frames is seen on the screen. With all light removed during the change from one frame to the next, the eye sees a rapid sequence of still pictures that provides the illusion of continuous motion.

This illusion of motion is possible because of a fortunate property of the human eye. The impression made by any light seen by the eye persists for a small fraction of a second after the light source is removed. Therefore, if many views are presented to the eye during this interval of persistence of vision, the eye will integrate them and give the impression of seeing all at the same time. It is this effect of persistence of vision that makes possible the televising of one basic element of a picture at a time. With the elements scanned rapidly enough, they appear to the eye as a complete picture unit with none of the individual elements separately visible. To have the illusion of motion in the scene also, enough complete pictures must be shown during each second to satisfy this persistence-of-vision requirement of the eye. This can be done by having a picture repetition rate greater than 16 per second. The repetition rate of 24 pictures per second used in motion-picture practice is satisfactory and produces the illusion of motion on the screen.

However, the rate of 24 frames per second is not rapid enough to allow the brightness of one picture to blend smoothly into the next through the time when the screen is blank between frames. The result is a definite flicker of light that is very annoying to persons viewing the screen, which is made alternately bright and dark. The extent to which this flicker can be noticed depends on the brightness of the screen, the flicker effect being worse for higher illumination levels. In motion-picture practice the problem of flicker is solved by running the film through the projector at a rate of 24 frames per second but showing each frame twice so that 48 pictures are flashed on the screen during each second. A shutter is used to blank out light from the screen not only during the time when one frame is being changed for the next but once between. Thus each frame is projected twice on the screen. There are 48 views of the scene during each second and the screen is blanked out 48 times per second, although there are still the same 24 picture frames per second. As a result of the increased blanking rate, flicker is eliminated.

1·5 Frame and field frequencies

A similar process is carried out in the television system to reproduce motion in the picture. Not only is each picture broken down into its many individual picture elements, but the scene is scanned rapidly enough to provide sufficient complete pictures or frames per second to give the illusion of motion in the reproduced scene on the picture tube screen. Instead of the 24 in commercial motion-picture practice, however, the frame repetition rate is 30 per second in the television system. This repetition rate provides the required continuity of motion.

The picture repetition rate of 30 per second is still not rapid enough to overcome the problem of flicker at the light levels encountered on the picture tube screen. Again the solution is similar to motion-picture practice; each frame is divided into two parts, so that 60 views of the scene are presented to the eye during each second. However, the division of a frame into two parts cannot be accomplished by the simple method of the shutter used with motion-picture film, because the picture is reproduced one element at a time in the television system. Instead, the same effect is obtained by a method of interlaced horizontal linear scanning that divides the total number of lines in the picture frame into two groups of lines called *fields*. Each frame is divided into two fields, one field containing the odd-numbered, the other the even-numbered scanning lines. The repetition rate of the fields is 60 per second, since two fields are scanned during a single frame period and the frame frequency is 30 cps. In this way 60 views of the picture are presented to the eye during one second, providing a repetition rate fast enough to eliminate flicker.

The frame repetition rate of 30 is chosen in television, rather than the 24 of commercial motion pictures, because most homes in the United States are supplied with 60-cycle a-c power. Having the frame rate of 30 per second makes the field rate exactly equal to the power-line frequency of 60 cps. Then any effects of hum in the picture stay still, instead of drifting up or down the screen.

It may be of interest to note that in countries where the power-line frequency is 50 cps the frame rate is 25 cps, making the field frequency 50 cps. Television standards for the United States and other countries are compared in Appendix C at the back of the book.

1·6 Vertical and horizontal scanning frequencies

The field rate of 60 cps is the vertical scanning frequency. This is the rate at which the electron beam completes its cycles of vertical motion, from top to bottom and back to top again, ready to start the next vertical scan. Therefore, vertical deflection circuits for either the camera tube or picture tube operate at 60 cps. The time for each vertical scanning cycle, or field, is $\frac{1}{60}$ sec.

The number of horizontal scanning lines in a field is one-half the total 525 lines for a complete frame, since one field contains every other line. This results in 262½ horizontal lines for each vertical field. Since the time for a field is $\frac{1}{60}$ sec and it contains 262½ lines, the number of lines per second is $262\frac{1}{2} \times 60$, which equals 15,750. Or, considering 525 lines for a successive pair of fields, which is a frame, we can multiply the frame rate of 30 by 525, which equals the same 15,750 lines scanned in 1 sec. This is the rate at which the electron beam completes its cycles of horizontal motion, from left to right and back to left again, ready to start the next horizontal scan. Therefore, horizontal deflection circuits for either the camera tube or picture tube operate at 15,750 cps. The time for each horizontal scanning line is 1/15,750 sec. In terms of microseconds,

$$H \text{ time} = \frac{1{,}000{,}000}{15{,}750}\,\mu\text{sec} = 63.5\ \mu\text{sec, approx}$$

1 · 7 Synchronization

Time in scanning corresponds to distance in the image. As the electron beam in the camera tube scans the image, the beam covers different elements of the image and provides the corresponding picture information. Therefore, when the electron beam scans the screen of the picture tube the scanning must be exactly timed to assemble the picture information in the correct position. Otherwise, the electron beam in the picture tube can be scanning the part of the screen where a man's mouth should be while at that time the picture information being received corresponds to his nose. To keep the transmitter and receiver scanning in step with each other, special synchronizing signals must be transmitted with the picture information for the receiver. These timing signals are rectangular pulses used to control both transmitter and receiver scanning.

The synchronizing pulses are transmitted as a part of the complete picture signal for the receiver, but they occur during the blanking time when no picture information is transmitted. The picture is blanked out for this period while the electron beam retraces. A horizontal synchronizing pulse at the end of each horizontal line begins the horizontal retrace time, and a vertical synchronizing pulse at the end of each field begins the vertical retrace time, thus keeping the receiver and transmitter scanning synchronized. Without the vertical field synchronization, the reproduced picture at the receiver does not hold vertically, rolling up or down on the kinescope screen. If the scanning lines are not synchronized, the picture will not hold horizontally, as it slips to the left or right and then tears apart into diagonal segments.

In summary, then, the horizontal-line-scanning frequency is 15,750 cps, and the frequency of the horizontal synchronizing pulses is also 15,750 cps. The frame repetition rate is 30 per second, but the vertical field-scanning frequency is 60 cps and the frequency of the vertical synchronizing pulses is also 60 cps.

1 · 8 Picture qualities

Assuming it is synchronized to stay still, the reproduced picture should also have high brightness, strong contrast, sharp detail, and the correct proportions of height and width.

Brightness. This is the overall or average intensity of illumination, which determines the background level in the reproduced picture. The illumination of individual picture elements can then vary above and below this average level. The brightness must be enough to provide a picture that can easily be seen in daylight or in a room with normal lighting. This presents a problem because the fluorescent screen of the kinescope is illuminated on only one small spot at a time. Therefore, the brightness of the complete picture is much less than the actual spot illumination. The bigger the picture, the more light needed from the spot to produce adequate brightness. Brightness on the screen depends on the amount of high voltage for the kinescope and its d-c bias in the grid-cathode circuit. In television receivers the brightness or background control varies the kinescope bias.

Contrast. By contrast is meant the difference in intensity between black and white parts of the reproduced picture, as distinguished from brightness, which is average intensity. The contrast range should be great enough to produce a pleasing picture, with bright white and dark black for the extreme intensity values. The amount of a-c video signal determines the contrast of the reproduced picture, as the signal amplitude decides how intense the white will be, compared with black parts of the signal. Little contrast means a weak picture that is soft in appearance but dull and flat. Excessive contrast makes the picture appear hard, usually with distortion of the gray values. In television receivers, the contrast control varies the peak-to-peak amplitude of a-c video signal coupled to the kinescope grid-cathode circuit.

The contrast of the picture as you view it also depends on brightness, mainly because the background level determines how black the darkest parts of the picture will be. Furthermore, the room illumination affects how the black looks. Keep in mind the fact that black in the picture is the same light level you see on the kinescope screen when the set is shut off. With a picture, this level looks black in contrast to the white fluorescence. However, the black cannot appear any darker than the room lighting reflected from the kinescope screen. The surrounding illumination must be low enough, therefore, to make black look dark. At the opposite extreme, the picture appears washed out with little contrast when viewed in direct sunlight because so much reflected light from the screen makes it impossible to have dark black.

Detail. The quality of detail, which is also called *resolution* or *definition*, depends on the number of picture elements that can be reproduced. With many small picture elements, the fine detail of the image is evident. Therefore, as many picture elements as possible should be reproduced to have a picture with good definition. This makes the picture clearer. Small details can be seen and objects in the image are outlined sharply. Good definition also gives apparent depth to the picture by bringing in details of the background. The improved quality of a picture with more detail can be seen in Fig. 1 · 10, which shows how more picture elements increase the definition.

In our commercial television broadcasting system, the picture repro-

(a) (b)

Fig. 1 · 10 Picture quality improves with more detail. (a) Coarse structure having few details and poor definition. (b) Fine detail and good definition.

duced on the kinescope screen is limited to a maximum of 150,000 picture elements, approximately, counting all details horizontally and vertically. This definition allows about the same detail as in 16-mm film. The maximum applies to any size frame, since detail depends on the number of scanning lines and bandwidth of the transmission channel.

Aspect ratio. This is the ratio of width to height of the picture frame. Standardized at 4:3, this aspect ratio makes the picture wider than its height by the factor 1.33 (see Fig. 1 · 11). Approximately the same aspect ratio is used for the frames in motion-picture film. Making the frame wider than the height allows for motion in the scene, which is usually in the horizontal direction.

Note that only the proportions are set by the aspect ratio. The actual frame can be any size from a few square inches to 20 by 15 ft, as long as the correct aspect ratio of 4:3 is maintained. If the kinescope does not reproduce the picture with this proportion of width to height, people in the scene look too thin or too wide.

Viewing distance. Too close to the screen, we see all the details, but the speckled grain called *snow* in the picture is visible and individual scanning lines make the reproduction look coarse. Farther away, the fine picture detail may be wasted. The best viewing distance is a compromise, therefore, about four to eight times the picture height.

1 · 9 Television channels

It takes only 1/15,750 sec or 63.5 μsec to scan one line from left to right and retrace to start another line. Within each horizontal line there are many picture elements. Because so much picture information must be contained in an electric signal within so short a period of time, signal voltages of high frequency are produced. These video signal frequencies are as high as 4 million cycles per second. Since the frequency of the picture carrier wave that is used to transmit the signal must be above 4 Mc, television broadcast stations use transmitting frequencies much higher than the standard broadcast band (535 to 1,605 kc) for radio broadcast stations. Also, a much wider band of frequencies is necessary for transmitting a television program. The band of frequencies assigned to a broadcast station for transmission of their signals is called a *channel.*

Each television broadcast station is assigned by the Federal Communications Commission (FCC) a channel 6 Mc wide in one of the following television broadcast bands: 54 to 88 Mc, 174 to 216 Mc, and 470 to 890 Mc. The 54- to 88-Mc and 174- to 216-Mc bands are in the very-high-frequency (VHF) spectrum of 30 to 300 Mc originally used for television broadcasting. The 54- to 88-Mc band includes channels 2 to 6, inclusive, which are often called the low-band VHF television channels; the 174- to 216-Mc band includes channels 7 to 13, inclusive, which are the high-band VHF television channels.

Fig. 1 · 11 Picture with correct aspect ratio. (Zenith Radio Corp.)

4. The width of the channel assigned to a television broadcast station is 6 Mc. This bandwidth applies to VHF channels and UHF channels, either for monochrome or color.
5. The associated sound is transmitted as an FM signal. The sound carrier signal is included in the 6-Mc television channel.
6. The picture carrier is amplitude-modulated by both picture and synchronizing signals. The two signals have different amplitudes on the AM picture carrier.
7. The aspect ratio of the picture is 4 units horizontally to 3 units vertically, or 1.33.

SUMMARY

1. The smallest area of light or shade in the image is a picture element.
2. Picture elements are converted to electric signal by the camera tube at the studio. This becomes the video signal to be broadcast to receivers. The picture tube in the receiver converts the video signal back into visual information.
3. The electron beam scans all the picture elements from left to right in one horizontal line and all the lines in succession from top to bottom. There are 525 lines per picture frame.
4. The complete picture frame is scanned 30 times per second.
5. Blanking means going to black so that retraces cannot be seen.
6. For vertical scanning, the 525 lines in each frame are divided into two fields, each with 262½ lines. Odd fields contain odd lines; even fields have even lines. Each odd field is interlaced with an even field to provide a complete frame. This procedure is called *interlaced scanning*.
7. The vertical scanning frequency is the field rate of 60 cps.
8. The horizontal scanning frequency is 15,750 cps, equal to 525×30, or $262\frac{1}{2} \times 60$. The time to scan one horizontal line, including trace and retrace, equals 1/15,750 sec or 63.5 μsec.
9. Synchronization is necessary to time the scanning with respect to picture information. The horizontal and vertical synchronizing pulse frequencies are 15,750 cps and 60 cps, respectively, the same as horizontal and vertical scanning frequencies.
10. Brightness is the average or overall illumination. On the kinescope screen, brightness depends on high voltage and d-c grid bias for the picture tube.
11. Contrast is the difference in intensity between black and white parts of the picture. The peak-to-peak a-c video signal amplitude determines contrast.
12. Detail, resolution, or definition is a measure of how many picture elements can be reproduced. With many fine details, the picture looks sharp and clear.
13. The aspect ratio specifies 4:3 for the ratio of width to height of the frame.
14. A standard commercial television broadcast channel is 6 Mc wide to include the AM picture carrier signal and FM sound carrier signal. The television channel frequencies are in the VHF and UHF bands.
15. In color television broadcasting, red, green, and blue video signals corresponding to the color information are converted into luminance and chrominance signals for transmission in the standard 6-Mc broadcast channel. The luminance signal has black-and-white information for monochrome receivers; the chrominance signal provides the color information for color receivers.

SELF-EXAMINATION *(Answers at back of book.)*

Fill in the missing word or number in the following statements:

1. Picture frames are repeated at the rate of ______ per second.
2. The number of scanning lines is ______ per frame.

3. The number of scanning lines is ______ per field.
4. The number of fields is ______ per frame.
5. The number of scanning lines is ______ per second.
6. The time for one scanning line is ______ μsec.
7. The horizontal line-scanning frequency is ______ cps.
8. The vertical field-scanning frequency is ______ cps.
9. Video signal amplitude determines the picture quality called ______.
10. The number of picture elements determines the picture quality called ______.
11. Visual information is converted to electric signal by the ______ tube.
12. Video signal is converted to light by the ______ tube.
13. The bandwidth of a television channel is ______ Mc.
14. The type of modulation on the transmitted picture carrier signal is ______.
15. The type of modulation on the transmitted sound carrier signal is ______.
16. The assigned band of frequencies for channel 7 is ______ Mc.
17. Scanning in the receiver is timed correctly by ______ pulses from the transmitter.
18. Retraces are not visible because of ______ pulses.
19. Black on the kinescope screen results from ______ voltage on the kinescope grid.
20. The frequency of the horizontal synchronizing pulses is ______ cps.

ESSAY QUESTIONS

1. Why is the television system of transmitting and receiving the picture information called a sequential method?
2. Why is vertical scanning necessary in addition to the horizontal line scanning?
3. Define aspect ratio, contrast, brightness, and resolution.
4. Name the two signals transmitted in color television and give the function of each.
5. Why is the sawtooth waveform the required waveshape for linear scanning?
6. How is flicker eliminated by using interlaced scanning?
7. How would the reproduced picture look if it were transmitted with the correct aspect ratio of 4:3, but on the kinescope screen at the receiver the frame was square?

PROBLEMS *(Answers to odd-numbered problems at back of book.)*

1. Calculate the time of one horizontal line for the following examples: (*a*) frames repeated at 60 cps, with 525 lines per frame (for progressive scanning); (*b*) frames repeated at 25 cps with 625 lines per frame (for European standards).
2. Give the frequencies included in channels 2, 4, 9, 13, and 31.
3. A picture has 400 picture elements horizontally and 300 details vertically. What is the total number of details?
4. How does the reproduced picture look without vertical synchronization?
5. How long does it take to scan across 2 picture elements if 400 are scanned in 50 μsec?

Chapter 2 *Camera tubes*

The picture signal begins at the television camera. As illustrated in Fig. 2 · 1, light from the scene is focused by an optical lens onto the photosensitive surface of the camera tube. Note that the lens inverts the optical image so that top left in the scene becomes bottom right in the image. Figure 2 · 2 shows a television camera with several different lenses on a turret mount for close-up views or long shots of the televised scene. With the optical image of the scene focused on the photosensitive surface, the camera tube can then convert the picture elements into a succession of electrical variations corresponding to the visual information. There are two main requirements. First, the camera must have photoelectric properties to convert variations of light intensity into electrical variations. Second, the scanning process is necessary to produce signal variations for each of the picture elements, one at a time, in successive order, to televise all the visual information in the complete image.

2 · 1 Photoemission

Certain metals emit electrons when light strikes the surface. The emitted electrons are photoelectrons, and the emitting surface is a photocathode. Especially sensitive to light is the group of elements called *alkali metals,* which include cesium, sodium, potassium, and lithium. Cesium oxide is often used because its photoemission is most sensitive to incandescent light. The photoelectric action is practically instantaneous. Furthermore, the number of photoelectrons emitted can be made directly proportional to the amount of incident light. Therefore, the electric-signal variations correspond exactly to the variations in light intensity. This is the fundamental action by which the optical image is converted to an electrical image.

The photocathode can be enclosed in a vacuum with a positive anode to

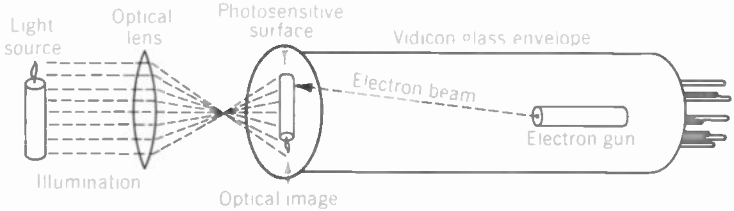
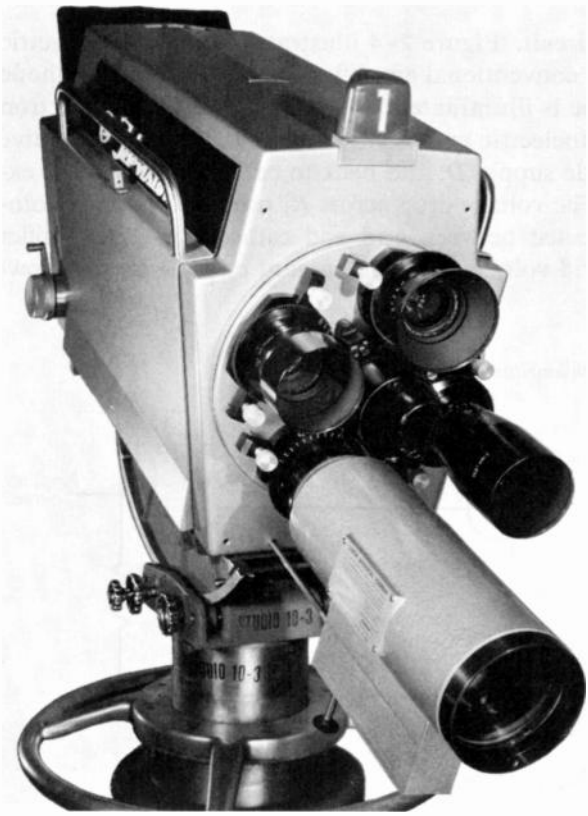


Fig. 2 · 1 Televising an image with vidicon type of camera tube. External coils for focusing and deflecting the electron beam are not shown.

collect the emitted electrons, forming a phototube such as the one in Fig. 2 · 3*a*. This photoelectric cell has a large curved cathode to intercept maximum light. Its inner surface is coated with a light-sensitive material that emits photoelectrons. The narrow wire down the center is the anode. With 100 to 400 volts positive on the anode, typical values of photoelectric current are 1 to 4 μa. Because the photoelectric current is so little, a large amount of amplification is necessary. The phototube in Fig. 2 · 3*b* includes an electron multiplier to amplify the photoelectric current. Here the photocathode is a coating on the inside of the glass faceplate at the top. This

Fig. 2 · 2 Television camera with four lenses of different focal lengths mounted on a revolving turret. The long barrel has a zoomar lens for extreme close-ups. An image orthicon camera tube is used in this camera head. (RCA.)



(a)

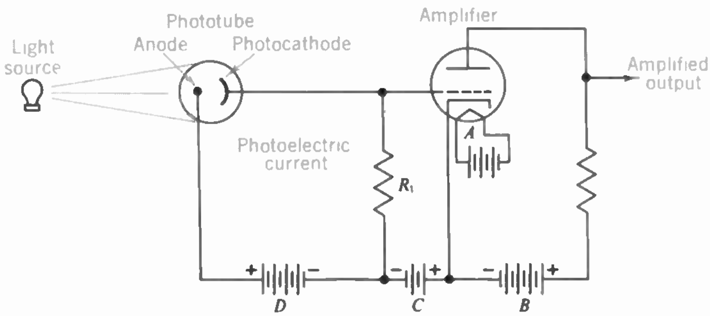
(b)

Fig. 2 · 3
Typical vacuum phototubes. (a) Diode type. Curved surface is photocathode and wire down center is anode. (b) Multiplier phototube. This is head-on type with semitransparent photocathode at top and electron multiplier below.

photocathode is semitransparent so that light can produce photoemission from the rear surface.

Phototube amplifier circuit. Figure 2 · 4 illustrates how a photoelectric tube can be used with a conventional amplifier circuit. When the cathode surface in the phototube is illuminated by the light source, the electron flow resulting from photoelectric emission is from cathode to the positive anode, through the anode supply D, and back to cathode through the external load resistor R_1. The voltage drop across R_1 produced by the photoelectric current is connected between grid and cathode of the amplifier tube. Any changes in grid voltage vary the amount of plate current flow-

Fig. 2 · 4 Phototube circuit with amplifier.



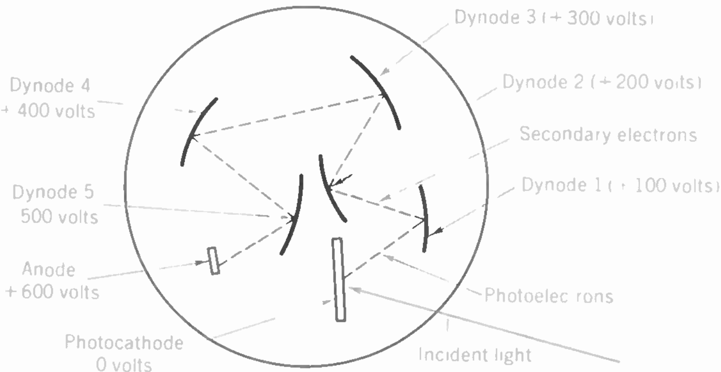
ing in the amplifier, thus providing an appreciable signal in the output circuit, which corresponds to the original photoelectric signal. Additional stages can be used after the first amplifier to produce the desired amount of amplification.

Secondary emission and multiplier phototubes. Metals have the property of emitting electrons when their surfaces are bombarded by incident electrons having a high velocity. This is called *secondary emission,* and the electrons emitted from the metallic surface by the incident primary electrons are called *secondary electrons.* These are not the same electrons that originally strike the surface but must be new electrons ejected from the metal, because there can be more secondary than primary electrons. Aluminum, as an example, can release seven secondary electrons for each incident primary electron. Almost all materials can produce secondary emission when the incident electrons have enough velocity.

Phototubes and camera tubes often include an electron-multiplier structure, making use of the secondary-emission effect to amplify the small amount of photoelectric current. The electron multiplier is a series of cold anode-cathode electrodes called *dynodes* mounted internally, with each at a progressively higher positive potential as illustrated in Fig. 2·5. The few electrons emitted by the photocathode are accelerated to a more positive dynode. Then the primary electrons can force the ejection of secondary-emission electrons when the velocity of the incident electrons is great enough.

The average number of electrons emitted from a dynode by each primary electron may be only three or four, depending on the potential and the type of surface. However, the number of electrons available is multiplied each time the secondary electrons strike the emitting surface of the next more positive dynode. The electron multiplier can be very useful, therefore, as a noise-free amplifier for very small photoelectric currents. Very little noise is produced while amplifying the current in an electron multiplier because it has no resistors. A relatively high voltage supply is

Fig. 2·5 Illustrating an electron multiplier for a phototube.



necessary for operation, however, because each stage must be at a progressively higher potential.

2·2 Flying-spot camera

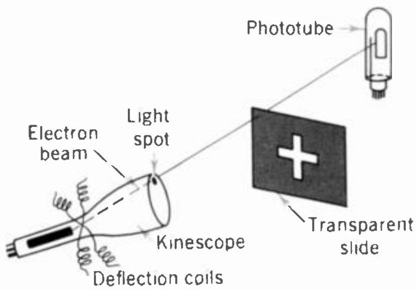
The arrangement shown in Fig. 2·6 is called a *flying-spot* camera because the spot of light moves over the image to scan the entire picture. This system combines the phototube and amplifier circuits with an electron scanning beam to convert an optical image to the desired camera signal. The screen of a cathode-ray tube is the source of light. Within the tube, the electron-gun structure produces a beam of electrons aimed at the fluorescent screen. Where the electron beam strikes, it produces a small spot of light. Scanning current in the deflection coils provides varying magnetic fields for horizontal and vertical scanning to move the light spot over the entire face of the tube. The brightness of the light spot is made as high as possible without burning the screen, to provide maximum light for the image. The entire assembly is enclosed in a lightproof case.

With the televised scene on a transparent film strip or slide, as shown in Fig. 2·6, the light from the scanning spot passes through the film to the phototube. The amount of light received by the phototube varies with the density of the film. Using a positive transparency, white parts of the picture are less dense and allow more light to go through. As the light spot scans the entire image, the amount of light to the phototube varies with the desired picture information. Then the output of the phototube contains signal variations for all the picture information, to provide the desired camera signal output. The electron-multiplier type of phototube is generally used for increased sensitivity, with the film very close to the kinescope screen for maximum light (see Fig. 2·7).

The flying-spot scanner is a convenient method of televising film slides. Excellent picture quality is obtained because of high-speed scanning by the electron beam, sensitive phototubes, and special flying-spot kinescopes that produce enough light.

Flying-spot kinescope. A common type is the 5ZP16 in Fig. 2·7. This cathode-ray tube has a 5-in.-diameter screen on a flat faceplate. It uses magnetic deflection but electrostatic focusing. The P16 phosphor for the screen produces purple florescence, near the ultraviolet-light spectrum, to be used with phototubes that are also sensitive to ultraviolet light.

Fig. 2·6 Flying-spot scanner arrangement.



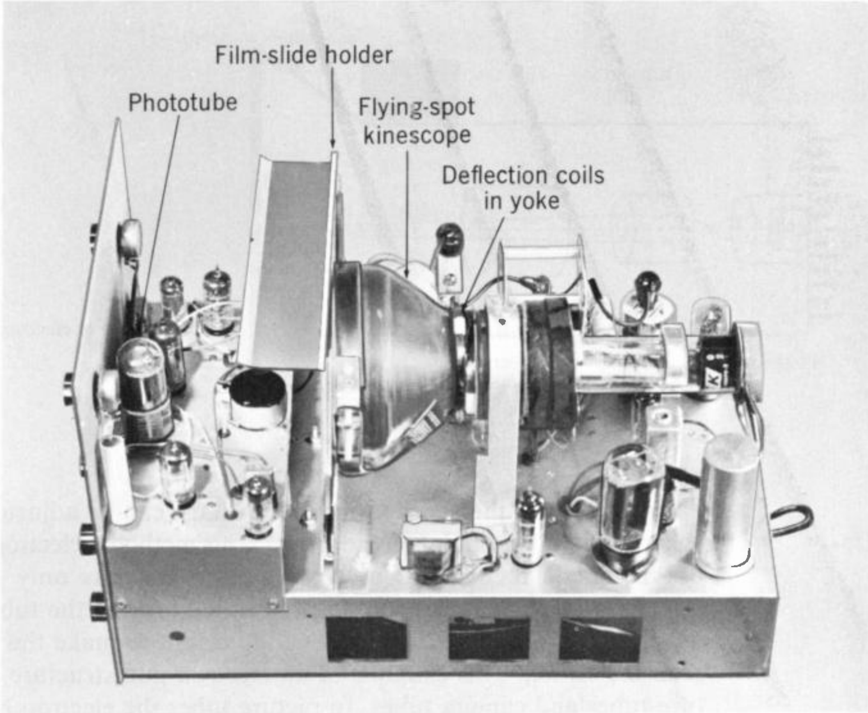


Fig. 2·7 Flying-spot scanner to generate signals for testing television receivers.
(B and K Manufacturing Co.)

Electron-gun structure. Enclosed in a vacuumed glass bulb, in the narrow part near the tube base, the gun produces a narrow beam of high-velocity electrons. Since the electrons emitted from the cathode must be concentrated into a beam and pass through the electrodes they are constructed in the form of a metal cylinder with a small pinhole or aperture. Referring to Fig. 2·8, at the left is the cathode. This is a thin metal sleeve enclosing the heater coil. Fitted on the cathode sleeve is a cap with a recess in the center to hold the oxide mixture that is heated to produce thermionic emission.

Directly after the cathode is the control-grid cylinder, with an aperture of 0.04 in. This sleeve actually encloses the cathode so that the negative grid bias can control the amount of beam current. The next two elements are accelerating electrodes, with positive voltages applied to attract electrons away from the cathode toward the screen. The positive accelerating voltages become progressively greater with the highest voltage on the final anode. Usually the anode is a conductive coating on the inside of the glass envelope. A separate anode connector is used for the high voltage, which is about 25,000 volts for flying-spot kinescopes.

Although the negative electrons repel each other, tending to form a

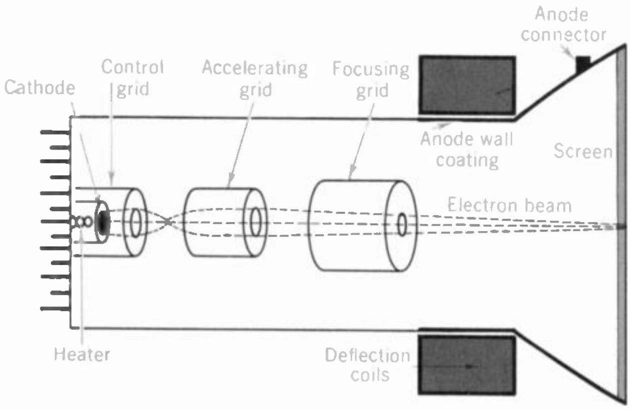


Fig. 2·8
Structure of electron gun in flying-spot kinescope.

diverging beam, the accelerating grid voltage can be adjusted to focus the electrons at a sharp point on the screen. This method is electrostatic focusing, but magnetic focusing is used where the gun has only one accelerating electrode. The deflection coils mounted around the tube are supplied with horizontal and vertical deflection current to make the electron beam scan the screen. This example of an electron-gun structure applies to picture tubes and camera tubes. In picture tubes the electron beam scans the screen; in camera tubes the beam scans an image plate or target plate.

2·3 Types of camera tubes

A camera tube includes in its vacuumed glass envelope (1) a photoelectric image plate that converts the visual information into electrical variations and (2) an electron gun for scanning the image to provide camera output signal for all the picture elements. The three main camera tubes, in their order of development, are the iconoscope (Fig. 1·3), image orthicon (Fig. 2·9), and vidicon (Fig. 2·13). The iconoscope, invented in 1928 by V. K. Zworykin, was commonly used for televising film but has been replaced by the vidicon. Not including film cameras, though, the image orthicon is the camera tube generally used in television broadcasting because of its high sensitivity. The vidicon is commonly used in small, portable television cameras because of its small size.

Camera sensitivity. This is the ratio of signal output voltage to the incident illumination. High sensitivity enables the camera tube to produce enough signal with moderate light. When the camera matches the sensitivity of the human eye, all types of scenes can be televised without restrictions on light level. This is especially important for remote pickups in the field, as the light level usually cannot be controlled outside the studio. In addition, high sensitivity provides benefits in the optical system for the camera. Smaller lens apertures can be used, improving depth of field while reducing the size and cost of the lenses needed to supply enough light for the camera.

Instantaneous pickups. The flying-spot camera is an instantaneous pickup, as it produces camera signal output for the light on each picture element only at the instant it is scanned. Most of the illumination is wasted. Because it cannot store the effect of light on an image plate, any instantaneous pickup has low sensitivity. Another example is the *image dissector*, an early type of camera tube invented by P. T. Farnsworth.

The storage principle. In storage-type camera tubes, the effect of illumination on every picture element is allowed to accumulate between the times it is scanned in successive frames. As an example, in the iconoscope the optical image is stored as a distribution of different electrical charges on the mosaic image plate. The charge image can be stored because each photoelectric globule on the mosaic plate is insulated from the others. As the plate is scanned by the electron gun, however, each point is discharged by the electron beam. Then signal current is produced in the circuit connected across the two output terminals on the glass bulb. In this case, the effect of light from the scene is stored in the electrical charge pattern on the mosaic plate. Each point stores charge from the time it is scanned in one frame until the same point in the next frame. With light storage, the amount of photoelectric signal can be increased 10,000 times, approximately, compared with an instantaneous system.

The image orthicon and vidicon also use the storage principle. These two camera tubes will now be described in detail.

2·4 Image orthicon[1]

Because of its high sensitivity, any scene visible to the eye can be televised with this camera tube. It is constructed in three main sections within the vacuumed glass envelope: the image section, scanning section, and electron multiplier (see Fig. 2·9). Light from the scene to be televised is focused onto the photocathode in the image section. This action produces a photoelectric image, which is then converted to an electrical charge image on the target plate. One side of the target plate receives the electrons emitted from the photocathode, while the opposite side of the target is scanned by the electron beam from the scanning section. As a result, signal current for the entire image is produced by the scanning beam. The signal current is then amplified in the electron-multiplier section, which provides the desired camera output signal.

Image section. The inside of the glass faceplate at the front is coated with a silver-antimony surface sensitized with cesium, to serve as the photocathode. This semitransparent photocathode receives the light image on the front side, while electrons are released from the opposite side facing the target.

The number of electrons emitted at any point in the photocathode is directly proportional to its illumination. Therefore, the photoelectrons have

[1] For more details on the image orthicon, see A. Rose, P. K. Weimer, and H. B. Law, *Proc. IRE,* July, 1946.

a distribution corresponding to the light variations in the optical image, forming an electron image of the picture. This electron image extends outward from the photocathode surface like bristles of a brush. However, the photocathode cannot store charge because it is a conductor. For this reason, the electron image is made to move to the glass target plate, which can store the charge image. Since the target plate is about 400 volts more positive than the photocathode, the photoelectrons emitted are attracted to the target. Although electrons tend to repel each other, this diverging effect is remedied by using a long-focus coil that extends over the image section. As a result, the electron image is focused at the target in order to produce a charge distribution on the target plate corresponding to the image.

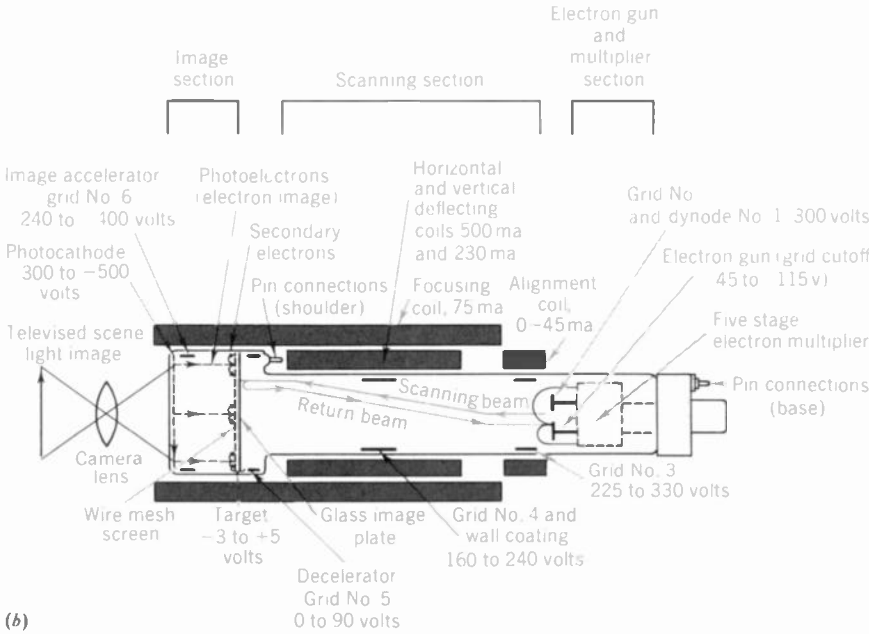
The glass target plate is 1½ in. wide and only 0.0001 in. thin. Because it is so thin, both sides have the required charge image. The charge is produced on the image side but is scanned on the opposite side.

Mounted close to the target on the photocathode side is a very fine wire-mesh screen. It has 500 to 1,000 meshes per inch, with an open area of 50 to 70 per cent so that the screen wires do not interfere with the electron image. The target assembly, including the glass plate and aluminum-mesh screen, is connected to a voltage source that can be adjusted for the best picture. Connections for all the image-section supply voltages are made through the pins mounted on the wide shoulder of the tube envelope, as can be seen in Fig. 2·9.

The scanning section. The electron-gun structure produces a beam of electrons that is accelerated toward the target. As indicated in Fig. 2·9, positive accelerating potentials of 160 to 330 volts are applied to grid 2, grid 3, and grid 4, which is connected internally to the metalized conductive coating on the inside wall of the tube. The electron beam is focused at the target by the magnetic field of the external focus coil and by the voltage supplied to grid 4. The alignment coil provides a magnetic field that can be varied to adjust the scanning beam's position exactly to correct for any mechanical misalignment of the electron gun. Deflection of the electron beam to scan the entire target plate is accomplished by the magnetic field of the vertical and horizontal deflecting coils mounted externally on the tube.

The target plate is close to zero potential. Therefore, electrons in the scanning beam can be made to stop their forward motion at the surface of the glass, and then return toward the gun structure. The grid 4 voltage is adjusted to produce uniform deceleration of electrons for the entire target area. As a result, electrons in the scanning beam are slowed down near the target. Depending upon the potential of individual areas in the target, some scanning-beam electrons land on the target while others stop at the glass surface and turn back to go toward the electron-multiplier structure. The electrons that return from the target provide the desired signal current. These return electrons are primary electrons, since the low-velocity scanning beam cannot produce secondary emission.

(*a*)



(*b*)

Fig. 2 · 9 Image orthicon camera tube. (a) Tube is 15½ in. long, 3 in. wide at front face, and 2 in. wide at narrow section. (b) Construction and operating voltages. All voltages indicated with respect to zero volts at cathode of electron gun. (RCA.)

The electron-multiplier section. The return beam from the target goes to the electron multiplier for amplification of the signal current. As illustrated in Fig. 2 · 10, the multiplier consists of several dynodes constructed as metal disks with cutouts, like a pinwheel (see Fig. 2 · 11). Each dynode is at a positive potential 200 to 300 volts greater than the preceding dynode.

Electrons returning from the target strike a disk on grid 2, which also serves as the first dynode of the multiplier section. The second dynode is a 32-blade pinwheel mounted behind dynode 1. Primary electrons from dynode 1 strike the blades of dynode 2 to produce more secondary electrons, which are attracted through the slots to the next stage. The same action occurs for each succeeding dynode. Five multiplier stages are used, each with a gain of approximately 4, providing a total gain of $4 \times 4 \times$

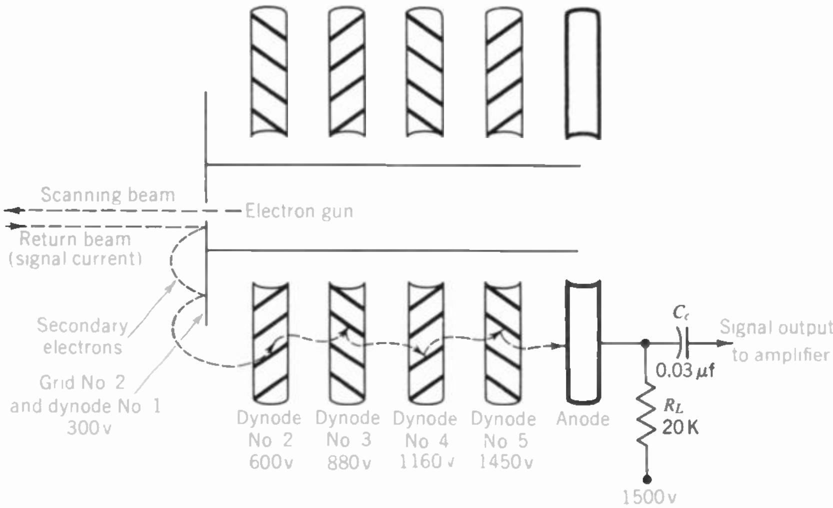
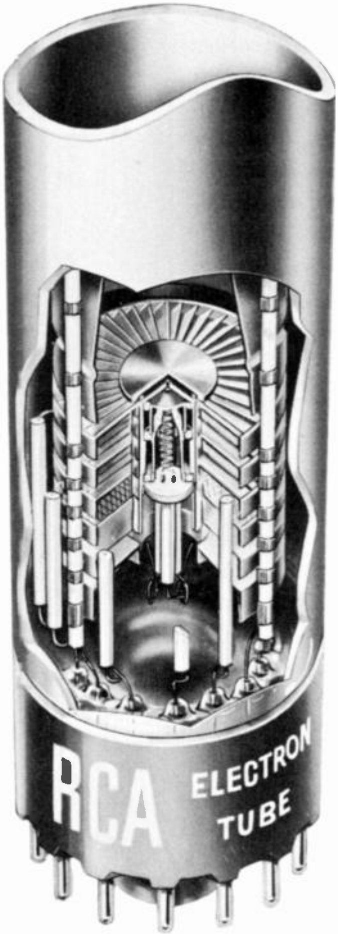


Fig. 2 · 10 Action of electron multiplier in image orthicon.

Fig. 2 · 11 Pinwheel dynodes in image orthicon. (RCA.)



$4 \times 4 \times 4$, or about 1,000. The secondary electrons are finally collected by the anode, which is connected to the highest supply voltage of +1,500 volts, in series with load resistor R_L. The anode current through R_L has the same variations that are present in the return beam from the target, amplified by the gain of the electron multiplier. Therefore, the voltage across R_L is the desired output, which is capacitively coupled to the camera signal amplifier.

Camera signal. The scene to be televised is focused by means of a suitable optical lens through the glass window of the tube directly onto the photocathode. Photoelectrons are emitted from the cathode surface in direct proportion to the light and shade in the scene, converting the optical image into an electron image. The electron image is accelerated toward the target, which is at a potential about 400 volts more positive than the negative photocathode.

When the photoelectrons in the electron image strike the target, secondary electrons are emitted from the screen side of the glass plate to produce a positive charge pattern on the plate. The charge is positive because the number of secondary electrons emitted is greater than the number of primary electrons. Brighter parts of the picture produce more photoelectrons and, therefore, more secondary emission. This makes the target more positive for the bright parts of the optical image, compared with dark areas in the picture. Secondary electrons ejected from the target are collected by the mesh screen close by so that they do not accumulate and cannot retard the secondary-emission process. As a result, a charge image is produced on the target plate corresponding to the picture elements in the optical image.

The light-storage principle is utilized effectively here as light in the image continuously provides photoelectrons that produce secondary emission, and the secondary electrons are removed by the wire screen to allow the charge to accumulate on the target plate. The charge distribution is preserved because the glass target plate has high electrical resistance along the surface. In this direction the glass plate is very narrow and has enough resistance to prevent the charge distribution from equalizing itself during the frame time of 1/30 sec. In the front-to-back direction, however, the glass target plate is a conductor of very large diameter and has a low resistance. Therefore, the charge pattern appears on both sides of the target. The brightest elements of the light image are most positive.

At the same time that the charge pattern is being formed on the target plate, it is scanned by the beam from the electron gun. The scanning-beam electrons have very little forward velocity at the target, but where the glass plate has a positive charge some of the scanning-beam electrons are attracted to land on the target. Enough electrons must be deposited on the glass plate from the electron stream to neutralize the positive charge. Therefore, more positive parts of the target require a greater number of electrons from the scanning beam than the less positive areas. The electrons in the beam in excess of the amount required to neutralize the charge on the spot being scanned can then turn back from the target and return to the electron gun.

As the electron beam scans the target, therefore, the charge distribution corresponding to the picture elements in the light image determines how many scanning electrons are returned toward the electron gun. Darker picture elements produce less positive areas on the target, and need fewer deposited scanning-beam electrons to neutralize the charge. A larger number of electrons from the scanning beam are returned to the electron gun for these elements. White picture elements produce more positive areas on the target, and fewer electrons turn back to the electron gun from these areas. In this way, an electron stream is started on its way back to the electron gun from the target plate, with variations in magnitude that correspond to the charge distribution on the target plate and the picture elements in the optical image.

The returning stream of electrons arrives at the gun close to the aperture from which the electron beam emerged. The aperture is part of a metal disk covering the gun electrode. When the returning electrons strike the disk, which is at a positive potential of about 300 volts with respect to the target, they produce secondary emission. The disk serves as the first stage of an electron multiplier. Succeeding stages of the electron multiplier are arranged symmetrically around and back of the first stage. Therefore, secondary electrons are attracted to the dynodes at progressively higher positive potentials. Five stages of multiplication are used. Since the amplified current from the final stage of the multiplier varies in magnitude with the picture information in the televised scene, the voltage output across load resistor R_L is the desired camera signal. With a signal current of 5 μa from the highlights in the scene and a 20,000-ohm R_L as typical values, the camera signal output from the image orthicon is 100,000 μv, or 0.1 volt.

The signal action in the image orthicon can be summarized briefly as follows:

1. Light from the televised scene is focused onto the photocathode, where the light image produces an electron image corresponding to the picture elements.
2. The electron image is accelerated to the target to produce secondary emission from the glass plate.
3. The secondary emission produces on the target a pattern of positive charges corresponding to the picture elements in the scene. White is most positive.
4. The low-velocity scanning beam from the electron gun provides electrons that land on the target to neutralize the positive charges. Scanning-beam electrons in excess of the amount needed to neutralize the positive charges turn back from the target and go toward the electron gun.
5. As the beam scans the target, therefore, the electrons turned back from the glass plate provide a signal current that varies in amplitude in accordance with the charge pattern and the picture information. The signal current is maximum for black.
6. The returning signal current enters the electron multiplier, where the current is amplified. The amplified current flowing through the load

Fig. 2 · 12 Light-transfer characteristics of image orthicon type 7513. (RCA.)

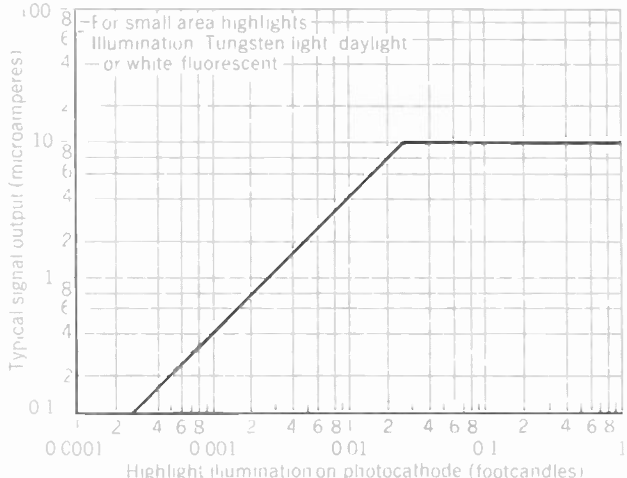


Plate I *Normal color picture.* (RCA.)

Plate II *Color fringing caused by incorrect convergence.* (RCA.)

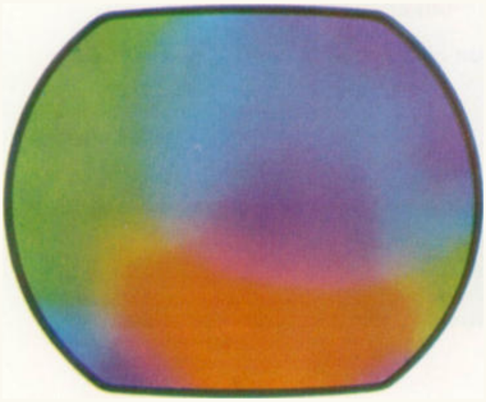


Plate III *Incorrect adjustment of purity magnet.* (RCA.)

Plate IV *No color sync.* (RCA.)

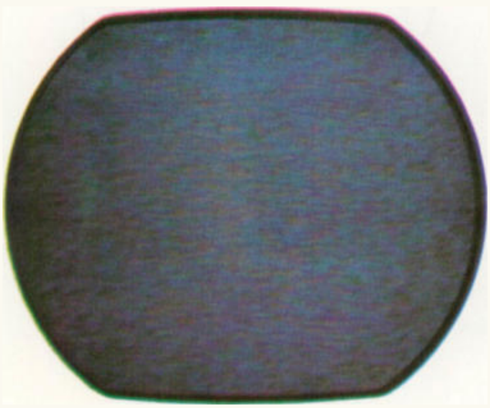
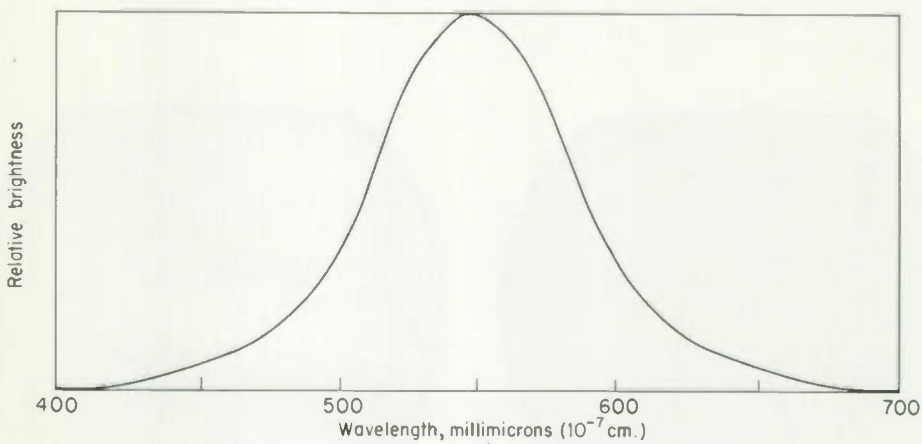


Plate V *Color noise or confetti.* (RCA.)

Plate VI *Green-stripe test signal.* (RCA.)

Plate VII Addition of colors.

a



b

Plate VIII Relative brightness response of the eye. (a) Hues of different wavelengths. (b) Brightness or signal-voltage response.

Plate IX NTSC flag illustrating resolution in color television. Largest areas are reproduced in full color as mixtures of red, green, and blue; smaller areas in orange or cyan; smallest details in black and white. (From *Proc. IRE*, Second Color Television Issue, January, 1954.)

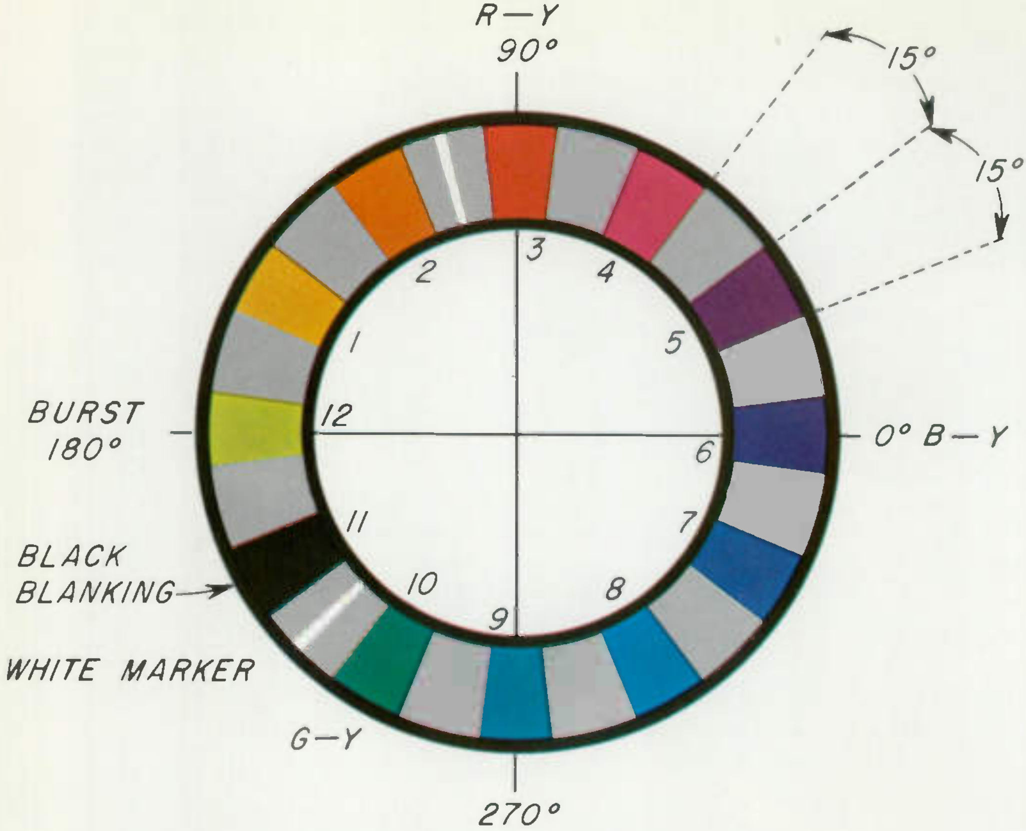


Plate X Color circle diagram, showing different hues for phase angles of chrominance signal.

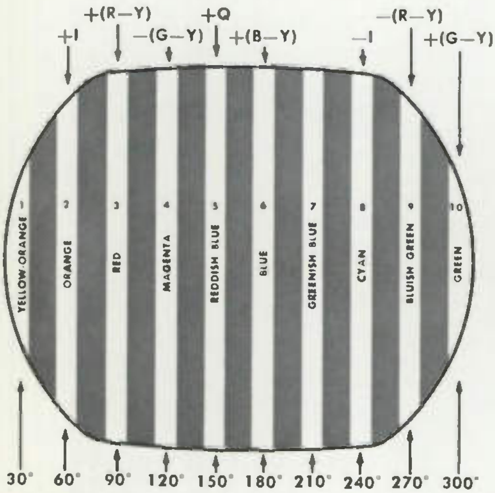


Plate XI Phase angles for color bar pattern in Plate XII.

Plate XII Color bar pattern corresponding to hue phase angles in Plate XI.

resistor in the multiplier's anode circuit produces the camera signal output voltage.

Sticking picture. This is an image of the televised scene with reversed black and white, which remains after the camera has been focused on a stationary bright image for several minutes, especially if the image orthicon is operated without sufficient warm-up. The sticking picture can usually be erased, however, by focusing on a clear white screen or wall. The time required to erase a sticking picture is usually a few minutes, but several hours may be necessary with older tubes. After the image orthicon is operated for 200 to 300 hr, it should be idle for 3 or 4 weeks to recover its original resolution and sensitivity. Steady use of the image orthicon causes cesium migration from the photocathode to the target, especially if the cesium surface is much hotter than the target plate.

Applications. The image orthicon is the camera tube generally used for live shows in the studio and in field cameras for televising sports and news events in black and white or in color. Matched sets of three are available for color cameras. One type of image orthicon has closer target-to-mesh spacing for very high sensitivity in remote camera pickups operating at very low light levels. However, this type has less resolution. A 4½-in. image orthicon is made with a larger photocathode and target area for increased detail in televising shows to be recorded on video tape.

The close-spaced image orthicon with extra sensitivity is equivalent to an ASA rating of 1,000 to 3,000 for film speed; the others correspond to ASA 400, equal to the fast Tri-X black-and-white film. The optical lenses are the same size used for 35-mm film, with focal lengths of 30 mm for long shots, 50 mm for medium view, and a telescopic longer lens for close-ups. Usually the long lens is the zoom type, which can maintain focus while being moved in and out manually or automatically.

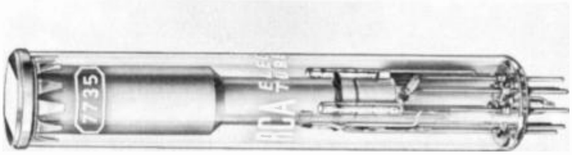
Light-transfer characteristic. Referring to Fig. 2 · 12, note the straight-line relation between light on the photocathode and camera signal, up to the knee at 10 μa output current. The linear characteristic is the reason for the name "orthicon," which means straight. For instance, increasing the light on the photocathode from 0.01 to 0.02 ft-candle doubles the output current from 4 to 8 ma.

One foot-candle unit is the illumination of a standard candle on a perpendicular surface 1 ft away. As examples, average indoor daylight is about 100 ft-candles; the required light on a library reading table is at least 10 ft-candles.

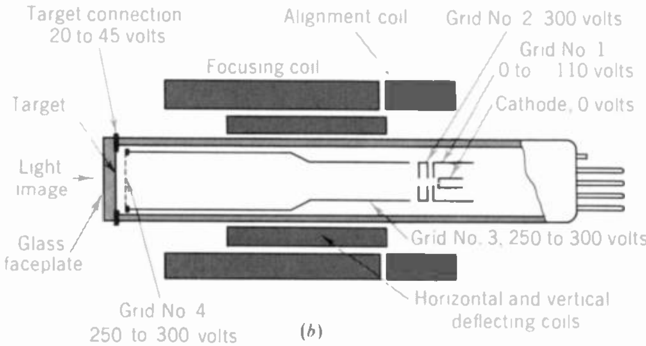
2 · 5 Vidicon

As shown in Fig. 2 · 13, the vidicon[2] is a very small camera tube of relatively simple construction. It has just a photosensitive target plate and electron gun. With the optical image focused on the target, it produces a

[2] For more details on the vidicon, see P. K. Weiner, S. V. Forgue, and R. R. Goodrich, *RCA Review*, September, 1951.



(a)



(b)

Fig. 2·13 Vidicon camera tube. (a) Tube is 6¼ in. long and 1 in. wide. (b) Construction and operating voltages. (RCA.)

charge image that is scanned by the electron beam from the gun. The image size on the target is only ½ by ⅜ in. The scanned raster is smaller than the faceplate to minimize distortion in the corners. Average illumination required is about 150 ft-candles in the scene or 1 to 10 ft-candles on the target plate. The vidicon has generally replaced the iconoscope for film cameras in television broadcasting. In addition, its small size allows many new applications in portable, economical vidicon cameras for educational and industrial uses in closed-circuit television.

Target. The target has two layers. One is a transparent film of conducting material, coated directly on the inside surface of the glass faceplate. This conductor is the signal-plate electrode for camera output signal. Light passes through to the second layer, which is an extremely thin coating of photoconductive material. Either selenium or antimony compounds are used. The photoconductive property means that its resistance decreases with the amount of incident light.

Charge image. The photolayer is an insulator with a resistance of approximately 20 megohms for the 0.00003-in. thickness, in the dark. Incident light can reduce the resistance to 2 megohms, as indicated in Fig. 2·14. Note that the image side of the photolayer contacts the signal plate at a potential of +40 volts, with respect to cathode. The opposite side returns to cathode through the electron beam, which has a resistance of 90 megohms approximately. With an image on the target, the potential of each point on the gun side of the photolayer depends on its resistance to the signal plate at +40 volts. As examples: a white area with low resistance can be close to the signal-plate voltage, approximately at +39.5 volts; the potential of a dark area with high resistance is lower at +35 volts. The result then is a pattern of positive potentials on the gun side of the photolayer, producing a charge image corresponding to the optical image. Maximum white in the picture is most positive in the charge pattern. The charge image on the target plate is scanned by the electron beam from the gun.

Electron gun. As shown in Fig. 2 · 14, the gun includes a heated cathode, control grid (No. 1), accelerating grid (No. 2), and focusing grid (No. 3). The electrostatic field of grid 3 and the magnetic field of an external-focus coil are both used to focus the electron beam on the target plate. Deflection of the beam for scanning is produced by horizontal and vertical deflection coils in an external deflection yoke.

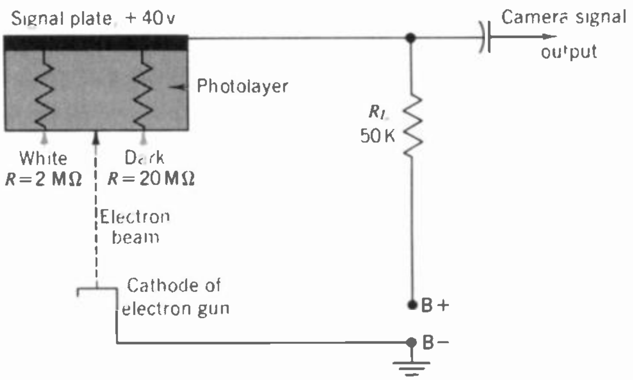
Note that grids 3 and 4 are connected internally. Grid 4 is a wire mesh to provide a uniform field near the target plate. Since the target is at a lower potential of 20 to 45 volts, compared with 250 to 300 volts on grid 4, electrons in the beam are decelerated just before they reach the target plate. The lower potential is used to slow down the electrons so that the low-velocity beam can deposit electrons on the charge image without producing secondary emission from the photolayer.

Signal current. Each point in the charge image has a different positive potential on the side of the photolayer toward the electron gun. Electrons in the beam are then deposited on the photolayer surface, reducing the positive potential toward the cathode voltage of zero. Excess electrons not deposited on the target are turned back, but this return beam is not used in the vidicon.

Consider a white picture element in the charge image on the photolayer. Its positive potential is close to the +40 volts on the signal plate just before the electron beam strikes. Then as electrons are deposited the potential drops toward zero. This change in potential causes signal current to flow in the signal-plate circuit, producing output voltage across R_L. For black in the picture, where the photolayer is less positive than white areas, the deposited electrons cause a smaller change in signal current.

The signal current results from the changes in potential difference between the two surfaces of the photolayer. The path for signal current can be considered a capacitive circuit provided by 5 $\mu\mu$f capacitance of the signal plate to the electron gun. Considering polarity, the output signal produces the least positive or most negative voltage output across R_L for

Fig. 2 · 14 Signal circuit for camera signal output of vidicon.



white highlights in the image. With 1 ft-candle illumination on the target, adjusted for a dark current of 0.02 μa, white highlights can produce 3 μa signal current for a voltage drop of 0.15 volt across the 0.05-megohm load resistor.

Applications. Four types of vidicon are manufactured for its many uses. One vidicon is for high-quality pictures in televising film for commercial broadcasting, either for monochrome film or in a set of three for color television. Another vidicon type has very high sensitivity for industrial uses in closed-circuit television. This vidicon has a sensitivity corresponding to the ASA rating of 600 for film speed. A vidicon camera for this application is shown in Fig. 2 · 16. The other two vidicon types are shorter with low heater-power requirements for transistorized cameras. One of these is for use under severe shock and vibration in military applications. Vidicon cameras are used in man-made satellites to televise pictures of the cloud cover around the earth to improve weather forecasting. The optical lenses for vidicon cameras are the same size used for 16-mm film equipment.

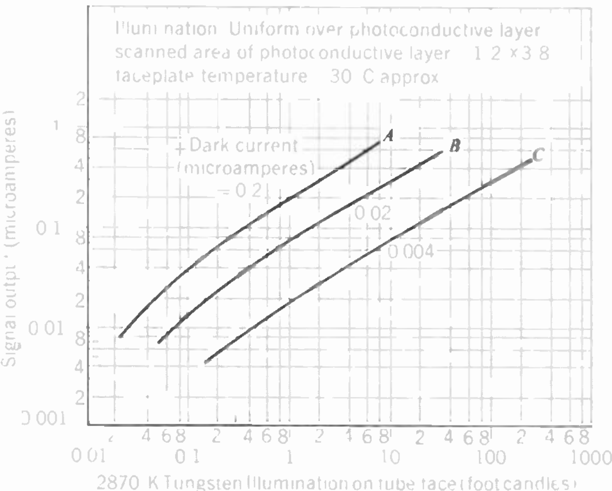
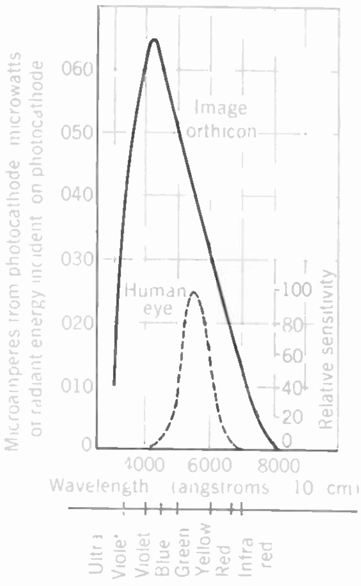


Fig. 2 · 15 Light-transfer characteristics of vidicon type 7263. (RCA.)

Fig. 2 · 16 Small vidicon camera with 16-mm lens mount for closed-circuit television. Entire unit is 6 in. wide and weighs 12 lb. (Thompson Ramo Woolridge, Inc.)

Fig. 2 · 17 Spectral sensitivity characteristic of image orthicon type 7037. (RCA.)



Light-transfer characteristics. Figure 2 · 15 shows three curves of vidicon output current for different sensitivity requirements according to the amount of illumination on the tube face. Each curve is for a specific value of dark current, which is the output with no light, corresponding to black in the picture. The dark current is set by adjusting target voltage. Sensitivity and dark current both increase as the target voltage is increased. The top curve *A* provides maximum sensitivity for live-scene pickup with highlight illumination as low as 2 ft-candles on the tube face. For this application, 60 to 100 volts on the target produces 0.2 μa dark current. White highlights at 2 ft-candles then produce output current of 0.3 μa. For average sensitivity with live scenes, setting the target at 30 to 50 volts produces 0.02 μa dark current. Then white highlights at 15 ft-candles produce output current of 0.4 μa. For film pickup with illumination at 100 ft-candles, the target is at 15 to 25 volts for 0.004 μa dark current, and the output current is 0.3 μa for white highlights.

2 · 6 Closed-circuit television

Also called *direct-wire* television, this system means the camera signal is not broadcast by radio but is connected by wire or cable directly to monitors at a remote position where the picture is reproduced on the kinescope screen. Generally, vidicon cameras are used because they are relatively inexpensive and compact. In many cases, the camera circuits are transistorized. There are any number of uses in education and industrial applications. Just a few examples are listed here.

Education. One teacher for multiple classrooms; close-up views of experiments.

Industrial. Inspection, watch nuclear reactions, night watchman.

Stores. Train personnel, observe customers and salespeople.

Home. Door monitor, baby sitter, observe someone sick in bed.

Medicine. Show operation in detail to students, observe patients in hospital.

Traffic Control. Observe both ends of tunnel or bridge, control freight traffic in railroad yards.

In addition, there are many applications in sports, advertising, marine or aviation traffic, and police or fire control.

2 · 7 Spectral response

This factor indicates sensitivity of the photosensitive surface in the camera tube for different colors in the visible light spectrum. The graph in Fig. 2 · 17 for a typical image orthicon also indicates approximate color response for the vidicon. Note the comparison with the spectral response of the human eye. Although the eye is most sensitive to light having the wavelengths for yellow-green, the camera tube responds best to blue. Furthermore, the relative color response indicates how a colorful scene is reproduced as different shadings in black and white. A blue blouse will be white but a red skirt will be dark gray, for the same illumination, while orange colors between yellow and red will appear medium gray. In color television, however, the entire visible spectrum is reproduced in the original colors.

SUMMARY

1. Any instantaneous camera pickup produces output signal for the light in each picture element only at the instant it is scanned. The flying-spot scanner is an instantaneous camera pickup. In storage-type camera tubes, the effect of light is allowed to accumulate between times the picture elements are scanned in successive frames. The image orthicon and vidicon utilize the storage principle to increase camera sensitivity.
2. The flying-spot camera pickup uses light from the screen of a kinescope as a flying-spot scanner. Light from each spot on the kinescope screen passes through a film slide to a photoelectric cell, which provides signal output.
3. The image orthicon has an image section, electron gun for scanning, and an electron multiplier to amplify the signal current. In the image section, light on the photocathode produces an electron image by photoemission. These electrons are attracted to the thin glass target plate, where secondary emission produces an image of positive charges corresponding to the electron image. The gun side of the target is scanned by a low-velocity beam to deposit enough electrons to neutralize the positive charges. Excess electrons are returned to the gun, where they enter the electron multiplier, for amplification by the dynode structure. The camera signal output is obtained as the varying voltage across the load resistor in the anode circuit of the multiplier.
4. The vidicon is a small camera tube with just the photoconductive target plate and electron gun. At the front of the target is a layer that is conductive for the output connection and transparent for the optical image to pass through to the photoconductive layer. The selenium photolayer is an insulator in the dark but allows slight conduction with light. As the electron beam scans the photolayer, light and dark elements allow more or less current to flow in the signal-plate circuit. This varying current through a load resistor produces camera signal output voltage.
5. In closed-circuit television, the camera signal is not broadcast. Vidicon cameras are generally used, connected by direct wire to monitors at a remote location where the picture is viewed on the kinescope screen. The many applications include industrial, military, and educational uses.

SELF-EXAMINATION *(Answers at back of book.)*

Answer true or false.

1. A photocathode emits photoelectrons.
2. Phototubes can use an electron multiplier.
3. A selenium photoconductive layer decreases its resistance with light.
4. A dynode emits secondary electrons.
5. An electron multiplier uses a series of dynodes.
6. Secondary emission from a point makes it positive.
7. Photoemission from a point makes it positive.
8. The flying-spot scanner is an instantaneous-type camera pickup.
9. Light is effectively stored by the target plate in the image orthicon.
10. In the image orthicon, maximum light on the photocathode causes maximum positive charge on the glass target plate.
11. Camera signal output of the image orthicon is taken from the anode of the electron multiplier.
12. The image orthicon has very high sensitivity because of the electron multiplier and charge storage by the target plate.

13. The photolayer in the vidicon emits photoelectrons.
14. The output signal current from the vidicon is approximately 3 μa, maximum for white.
15. The image orthicon spectral response is maximum for blue.
16. Closed-circuit television need not use FCC broadcast standards.
17. In the image orthicon, the target plate is scanned by the electron beam.
18. In the vidicon, the photolayer on the glass faceplate is scanned by the electron beam.
19. Camera signal output from any camera tube includes current variations with respect to time, corresponding to light variations of the picture elements.
20. Typical output current from the image orthicon is 6 ma.

ESSAY QUESTIONS

1. Define the following: photocathode, photoconductor, photoelectrons, secondary electrons, dynode.
2. Describe briefly the operation of a flying-spot camera pickup, with a drawing to show the physical arrangement.
3. In the image orthicon, describe briefly the function of: (*a*) photocathode; (*b*) glass target plate; (*c*) wire mesh of target; (*d*) decelerator grid 5; (*e*) dynode 1; (*f*) the anode in electron multiplier.
4. What causes sticking picture in the image orthicon? How is it erased?
5. In the vidicon, what are the two layers of the target plate? Give the function of each.
6. Make a drawing illustrating construction and operation of the vidicon. Explain briefly how camera signal output is obtained corresponding to the optical image.
7. Explain how the light-storage principle applies to: (*a*) the image orthicon; (*b*) the vidicon.
8. Give two practical applications of the image orthicon and two for the vidicon.
9. List five uses for closed-circuit television.

PROBLEMS *(Answers to odd-numbered problems at back of book.)*

1. With 10 μa peak-to-peak output from an image orthicon, how much is the peak-to-peak camera signal voltage across a 20,000-ohm R_L?
2. With 0.35 μa peak-to-peak output from a vidicon, how much is the peak-to-peak camera signal voltage across a 50,000-ohm R_L?
3. An electron multiplier has five dynodes. Each emits six secondary electrons for a primary electron. Assuming that two-thirds of the secondary electrons are attracted to the next dynode, how much is the current amplification of the electron multiplier?
4. Referring to Fig. 2·17, which color in the visible spectrum has the shortest wavelength? Longest wavelength? Give their wavelengths in meters, angstrom (10^{-10} m) units, and millimicrons (10^{-9} m).
5. Referring to the image orthicon light-transfer characteristics in Fig. 2·12, how much output current is produced by 0.005, 0.01, and 0.02 ft-candle on the photocathode?
6. Referring to the vidicon light-transfer characteristics in Fig. 2·15, with 0.02 μa dark current, how much output current is produced by 0.1, 1.0, and 10 ft-candles on the tube face?

Chapter 3 Scanning and synchronizing

The electron scanning beam is continuously deflected in a standard sequence of horizontal lines to scan all the elements in the picture. On the picture tube screen, the rectangular area of light produced by the electron beam as it scans horizontally and vertically is called the scanning *raster.* Figure 3 • 1 shows the raster on the kinescope screen, without any picture information. When video signal voltage is coupled to the kinescope control grid, the picture is reproduced on the raster. However, just scanning the raster is not enough because the deflection must be synchronized at the transmitter and receiver. To time the horizontal and vertical scanning correctly, synchronizing pulses are transmitted to the receiver. With synchronized scanning, the picture elements reproduced on the kinescope screen have the same relative position as on the image plate of the camera tube.

3 • 1 The sawtooth waveform for linear scanning

As an example of linear scanning consider the sawtooth waveshape in Fig. 3 • 2 as scanning current in the deflection coils for an electromagnetic tube. Let its peak value be 400 ma. If 100 ma is needed to produce a deflection of 5 in., then 400 ma will deflect the beam 20 in. Furthermore, the linear rise on the sawtooth wave provides equal increases of 100 ma for each of the four equal periods of time shown. Each additional 100 ma deflects the beam another 5 in.

If we consider horizontal scanning, this uniform rise of current in the horizontal deflection coils deflects the beam across the screen with a continuous, uniform motion that has constant velocity for the trace from left to right. At the peak of the rise the sawtooth wave reverses direction and decreases rapidly to its initial value. This fast reversal produces the retrace or flyback.

For linear vertical scanning the sawtooth deflection waveshape is provided in a similar manner but current in the vertical deflection coils moves the electron beam from top to bottom of the raster. While the electron beam is being deflected horizontally, the vertical sawtooth deflection waveshape moves the beam downward with uniform speed. Then the beam produces complete horizontal lines one under the other. The trace part of the sawtooth wave for vertical scanning deflects the beam to the bottom of the raster. Then the rapid vertical retrace returns the beam to the top.

Both trace and retrace are included in one cycle of the sawtooth wave. Since the number of complete horizontal lines scanned in 1 sec equals 15,750, for horizontal deflection the frequency of the sawtooth waves is 15,750 cps. For vertical deflection the frequency of the sawtooth waves equals the field-scanning rate of 60 cps. The vertical scanning motion at 60 cps is much slower than the horizontal sweep rate of 15,750, since many horizontal lines must be scanned during one cycle of vertical scanning.

During flyback time, both horizontal and vertical, all picture information is blanked out. Therefore, the retrace part of the sawtooth wave is made as short as possible, since retrace is wasted time in terms of picture information. For horizontal scanning, retrace time is approximately 10

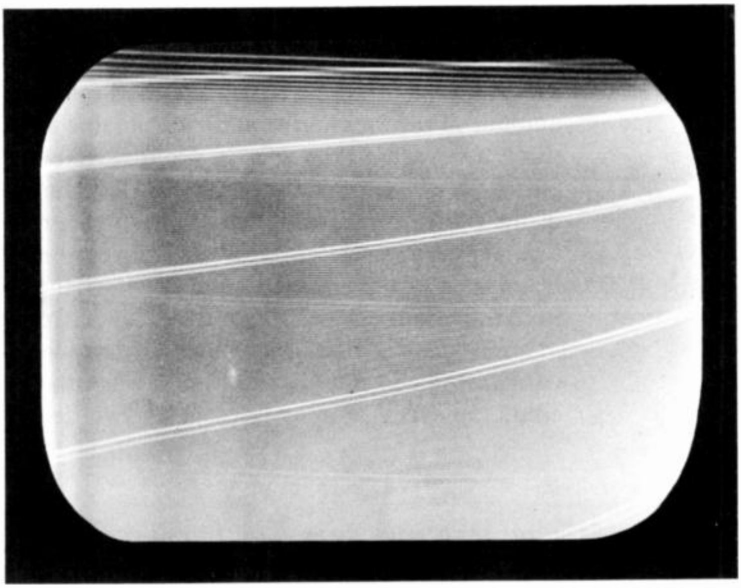


Fig. 3·1 Scanning raster on kinescope screen. Retrace lines are usually not visible. This raster is not interlaced because there is no vertical synchronization.

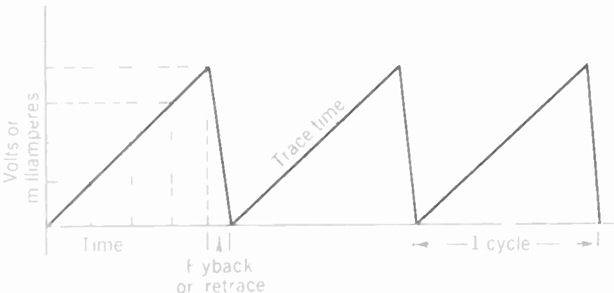


Fig. 3·2 Sawtooth scanning waveform.

per cent of the total line period. With 63.5 μsec for a complete line, 10 per cent equals 6.35 μsec for horizontal flyback time. Practical limitations in the circuits producing the sawtooth waveform make it difficult to produce a faster flyback. The lower frequency vertical sawtooth waves usually have a flyback time less than 5 per cent of one complete cycle. A vertical retrace 3 per cent of 1/60 sec, as an example, equals 0.0005 sec, or 500 μsec. Although vertical retrace is fast compared with vertical trace, note that 500 μsec is long enough to include almost eight horizontal lines scanned during vertical flyback, since one complete line is only 63.5 μsec.

3 · 2 Standard scanning pattern

The scanning procedure that has been universally adopted employs horizontal linear scanning in an odd-line interlaced pattern. The FCC scanning specifications for television broadcasting in the United States provide a standard scanning pattern that includes a total of 525 horizontal scanning lines in a rectangular frame having a 4:3 aspect ratio. The frames are repeated at a rate of 30 per second with two fields interlaced in each frame.

Interlacing procedure. Interlaced scanning can be compared with reading the interlaced lines written in Fig. 3 · 3. Here the information on the page is continuous if you read all the odd lines from top to bottom and then go back to the top to read all the even lines down to the bottom. If the whole page were written and read in this interlaced pattern the same amount of information would be available as though it were written in the usual way with all the lines in progressive order.

For interlaced scanning, therefore, all the odd lines from top to bottom of the frame are scanned first, skipping over the even lines. After this vertical scanning cycle, a rapid vertical retrace moves the electron scanning beam back to the top of the frame. Then all the even lines which were omitted in the previous scanning run are scanned from top to bottom.

Each frame is therefore divided into two fields. The first and all odd fields contain the odd lines in the frame, while the second and all even fields include the even scanning lines. With two fields per frame and 30 complete

The horizontal scanning lines are interlaced in
the odd lines are scanned, omitting the even lines.
the television system in order to provide two
Then the even lines are scanned to complete the
views of the image for each picture frame. All
whole frame without losing any picture information.

Fig. 3 · 3 Interlaced lines. Read the first and odd lines first and then the second and even lines.

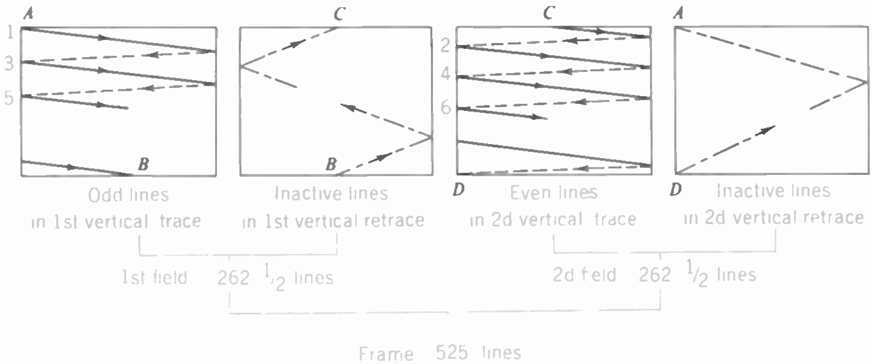


Fig. 3·4 Odd-line interlaced scanning procedure.

frames scanned per second the field repetition rate is 60 per second and the vertical scanning frequency is 60 cps. In fact, it is the doubling of the vertical scanning frequency from the 30-cps frame rate to the 60-cps field rate that makes the beam scan every other line in the frame.

Odd-line interlacing. The geometry of the standard odd-line interlaced scanning pattern is illustrated in Fig. 3·4. Actually, the electron gun aims the beam at the center, which is where the scanning starts from. For convenience, however, we can follow the motion starting at the upper left corner of the frame at point *A*. For this line 1, the beam sweeps across the frame with uniform velocity to cover all the picture elements in one horizontal line. At the end of this trace the beam then retraces rapidly to the left side of the frame, as shown by the dashed line in the illustration, to begin the next horizontal line.

Note that the horizontal lines slope downward in the direction of scanning because the vertical deflecting signal simultaneously produces a vertical scanning motion, which is very slow compared with horizontal scanning. Also note that the slope of the horizontal trace from left to right is greater than during retrace from right to left. The reason is that the faster retrace does not allow the beam so much time to be deflected vertically.

After line 1, the beam is at the left side ready to scan line 3, omitting the second line. This skipping of lines is accomplished by doubling the vertical scanning frequency from the frame repetition rate of 30 to the field frequency of 60 cps. Deflecting the beam vertically at twice the speed necessary to scan 525 lines produces a complete vertical scanning period for only 262 ½ lines, with alternate lines left blank. The electron beam scans all the odd lines, then, finally reaching a position such as *B* in the figure at the bottom of the frame.

At time *B* the vertical retrace begins because of flyback on the vertical sawtooth deflecting signal. Then the beam is brought back to the top of the frame to begin the second, or even, field. As shown in Fig. 3·4, the beam moves from point *B* up to *C*, traversing a whole number of horizontal lines.

This vertical retrace time is long enough for the beam to scan several horizontal lines. We can call these *vertical retrace lines*, meaning complete horizontal lines scanned during vertical flyback. Note that the vertical retrace lines slope upward, as the beam is moving up while it scans horizontally. The upward slope of vertical retrace lines is greater than the downward slope of lines scanned during vertical trace because the flyback upward is much faster than the trace downward. Any lines scanned during vertical retrace are not visible, though, because the electron beam is cut off by blanking voltage during vertical flyback time.

Horizontal scanning during the second field begins with the beam at point *C* in Fig. 3 · 4. This point is at the middle of a horizontal line because the first field contains 262 lines plus one-half a line. After scanning a half line from point *C* the beam scans line 2 in the second field. Then the beam scans between the odd lines to produce the even lines that were omitted during the scanning of the first field. The vertical scanning motion is exactly the same as in the previous field, giving all the horizontal lines the same slope downward in the direction of scanning. As a result, all the even lines in the second field are scanned down to point *D*. Points *D* and *B* are one-half line away from each other because the second field started with a half line.

The vertical retrace in the second field starts at point *D* in Fig. 3 · 4. From here, vertical flyback returns the beam to the top. With a whole number of vertical retrace lines, the beam finishes the second vertical retrace at *A*. The beam will always finish the second vertical retrace where the first trace started because the number of vertical retrace lines is the same in both fields. At point *A*, then, the scanning beam has just completed two fields or one frame and is ready to start the third field to repeat the scanning pattern.

All odd fields begin at point *A* and are the same. All even fields begin at point *C* and are the same. Since the beginning of the even-field scanning at *C* is on the same horizontal level as *A* with a separation of one-half line, and the slope of all the lines is the same, the even lines in the even fields fall exactly between the odd lines in the odd field. The essential requirement for this odd-line interlace is that the starting points at the top of the frame be separated by exactly one-half line between even and odd fields.

3 · 3 A sample scanning pattern

A complete scanning pattern is shown in Fig. 3 · 5 with the corresponding horizontal and vertical sawtooth waveforms to illustrate odd-line interlacing. A total of 21 lines in the frame is used for simplicity, instead of 525. The 21 lines are interlaced in two fields per frame. One-half the 21-line total or 10½ lines are in each field. Of the 10½ lines in a field, we can assume one line is scanned during vertical retrace for a convenient vertical flyback time. Then 9½ lines are scanned during vertical trace in each field.

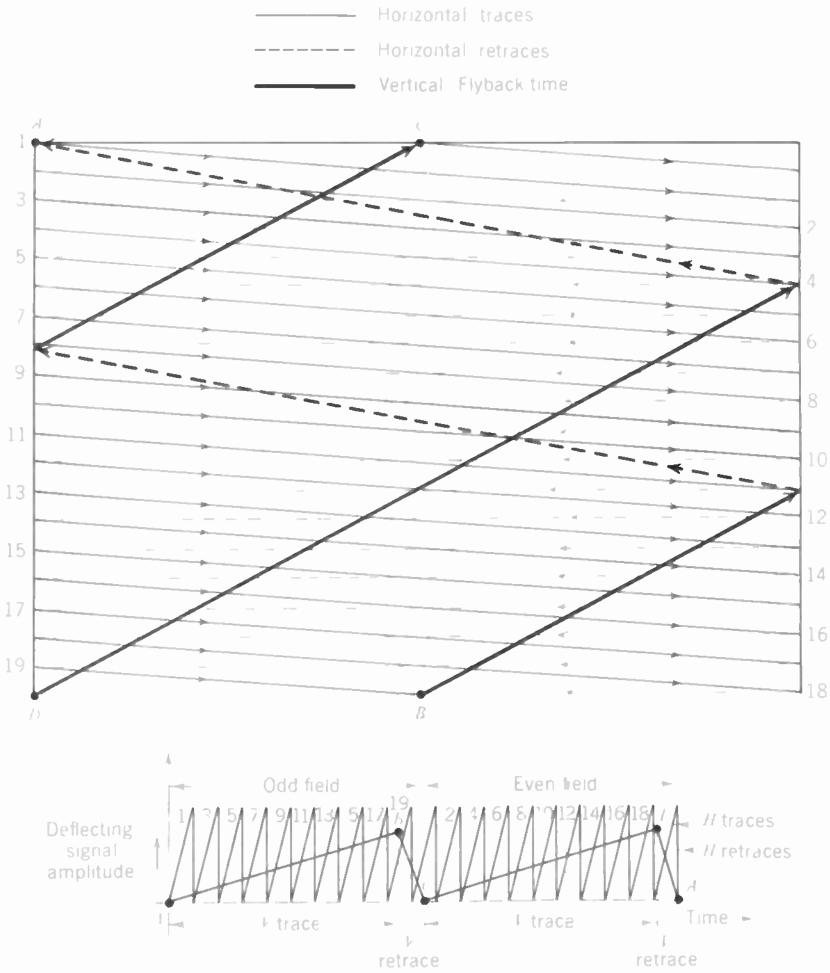


Fig. 3 · 5 A sample scanning pattern for 21 interlaced lines, with corresponding sawtooth deflection waveform. Beginning at point A, the scanning motion continues through B, C, D, and back to A again.

The entire frame has 2 × 9½ or 19 lines scanned during vertical trace, plus 2 vertical retrace lines.

Starting in the upper left corner at *A* in Fig. 3 · 5, the beam scans the first line from left to right and retraces to the left for the beginning of the third line in the frame. Then the beam scans the third and succeeding odd lines down to the bottom of the frame. After scanning 9½ lines the beam is at point *B* at the bottom when vertical flyback begins. Notice that this vertical retrace starts in the middle of a horizontal line. One line is scanned during vertical retrace, consisting of two half lines in this illustration, sloping upward in the direction of scanning. During this vertical retrace the scanning beam is brought up to point C, separated from point A by exactly one-half line, to start scanning the second field.

Because of this half-line separation between points A and C the lines scanned in the even field fall exactly between the odd lines in the previous field. The beam then scans 9½ even lines from point C down to D where the vertical retrace begins for the even field. This vertical retrace starts at the beginning of a horizontal line. Vertical retrace time is the same for both fields. Therefore, the one vertical retrace line in the second field returns the beam from D at the bottom to A at the top left corner of the frame where another odd field begins.

It should be noted that the points at which vertical retrace and the downward scan begin need not be exactly as shown in Fig. 3 · 5. These points could all be shifted by any fraction of a horizontal line without loss of interlace if the half-line difference were maintained. The half-line spacing between the starting points in alternate fields is automatically produced in the sawtooth deflecting signals and the scanning motion because there is an odd number of lines for an even number of fields. Proper interlacing is assured, therefore, when the required frequencies of the horizontal and vertical sawtooth scanning signals are maintained precisely and the flyback time on the vertical sawtooth wave is constant for all fields.

3 · 4 Flicker

Interlaced scanning is used because the flicker effect is negligible with 60 views of the picture presented each second. Although the frame repetition rate is still 30 per second, the picture is blanked out during each vertical retrace 60 times per second. Then the change from black between pictures to the white picture is too rapid to be noticeable. If progressive scanning were used instead of interlacing, with all the lines in the frame simply

Fig. 3 · 6 Trapezoidal raster. (a) Keystoned sides caused by unsymmetrical horizontal scanning. (b) Keystoned at top and bottom because of unsymmetrical vertical scanning.

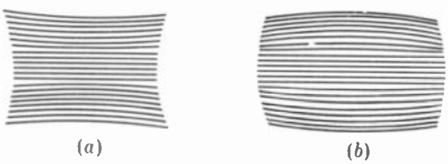


Fig. 3 · 7 Pincushion distortion of raster in (a) and barrel distortion in (b).

Fig. 3 · 8 Incorrect aspect ratio in raster. (a) Insufficient width. (b) Insufficient height.

scanned in progressive order from top to bottom, there would be only 30 blankouts per second and objectionable flicker would result. Scanning 60 complete frames per second in a progressive pattern would also eliminate flicker in the picture but the horizontal scanning speed would be doubled, which would double the video frequencies corresponding to the picture elements in a line.

Although the increased blanking rate with interlaced scanning largely eliminates the effect of flicker in the image as a whole, the fact that individual lines are interlaced can cause flicker in small areas of the picture. Any one line in the image is illuminated 30 times per second, reducing the flicker rate of a single line to one-half the flicker rate for the interlaced image as a whole. The lower flicker rate for individual lines may cause two effects in the picture called *interline flicker* and *line crawl.* The interline flicker is sometimes evident as a blinking of thin horizontal objects in the picture, such as the roof line of a house. Line crawl is an apparent movement of the scanning lines upward or downward through the picture, due to the successive illumination of adjacent lines. These effects may be noticed sometimes in bright parts of the picture because the eye perceives flicker more easily at high brightness levels.

3 · 5 Raster distortions

Since the picture information is reproduced on the scanning lines, distortions of the raster are evident in the picture. Two common problems are obtaining the correct aspect ratio and producing uniform deflection so that the edges are not distorted with respect to the center.

Keystone effect. In Fig. 3 · 6*a*, the scanning lines at the top are wider than those at the bottom, giving the raster the shape of a keystone. This effect is also called *trapezoidal distortion* because the geometrical form of a trapezoid has straight edges that are not parallel.

The keystone effect of Fig. 3 · 6*a* occurs in the iconoscope camera tube because the electron gun is not perpendicular to the image plate. In general, though, keystoning or trapezoidal distortion is a trouble in the raster caused by unsymmetrical deflection. Depending on whether horizontal or vertical deflection is unbalanced, the raster can be keystoned left to right as in *a*, or top to bottom as in *b*.

Pincushion and barrel distortion. If deflection is not uniform at the edges of the raster, compared with the center, the raster will not have straight edges. With the scanning lines bowed inward as in Fig. 3 · 7*a*, this effect is called *pincushion distortion.* Barrel distortion is shown in *b*.

Incorrect aspect ratio. Two cases are illustrated in Fig. 3 · 8. In *a* the raster on the kinescope screen is not wide enough for its height, compared with the 4:3 aspect ratio used in the camera tube. Then people in the picture look too tall and thin, with the same geometrical distortion as the raster. This raster needs more width. In *b*, the raster is too short for its width and people in the picture will look too short and wide. This raster needs more height.

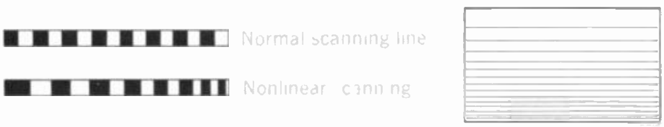
Nonlinear scanning. The sawtooth waveform with a linear rise for trace time produces linear scanning, as the beam is made to move with constant speed. With nonlinear scanning, however, the motion of the beam is too slow or too fast. If the scanning spot in the kinescope at the receiver moves too slowly, compared with scanning in the camera tube at the transmitter, picture information will be crowded together. Or, if scanning on the kinescope screen is too fast, the reproduced picture information is spread out. Usually the nonlinear scanning causes both effects at opposite ends of the raster. This is illustrated in Fig. 3·9*a* for a horizontal line with picture elements spread out at the left and crowded at the right. When the same effect occurs for all the horizontal lines in the raster, the entire picture is spread out at the left side and crowded at the right side. With people in the picture, a person at the left appears too wide while someone at the right looks too thin.

The vertical scanning motion must also be uniform. Otherwise the horizontal lines will be bunched at the top or bottom of the raster and spread out at the opposite end. This effect is illustrated in Fig. 3·9*b* for spreading at the top and crowding at the bottom. Then a person in the picture will appear distorted with long legs and flattened head.

Poor interlacing. In each field the vertical trace from the top must start exactly one-half line from the start of the previous field for odd-line interlacing. If the downward motion is slightly displaced from this correct position, the spot starts scanning too close to one of the lines in the previous field, instead of scanning exactly between lines. This incorrect start produces a vertical displacement between odd and even lines that is carried through the entire frame. As a result, pairs of lines are too close, with extra space between pairs. Then you can see too much black between the white scanning lines. This defect in the interlaced scanning is called *line pairing.* For the extreme case, lines in each successive field may be scanned exactly on the previous field lines. Then the raster contains only one-half the usual number of horizontal lines.

When the picture has diagonal lines as part of the image, poor interlacing makes them appear to be interwoven in the moire effect shown in Fig. 3·10. This effect is also called *fishtailing.* This is more evident in diagonal

Fig. 3·9 Effects of nonlinear scanning. (a) Picture elements spread out at left and crowded at right, caused by nonlinear horizontal scanning. (b) Scanning lines spread out at top and crowded at bottom, caused by nonlinear vertical scanning.



(*a*) (*b*)

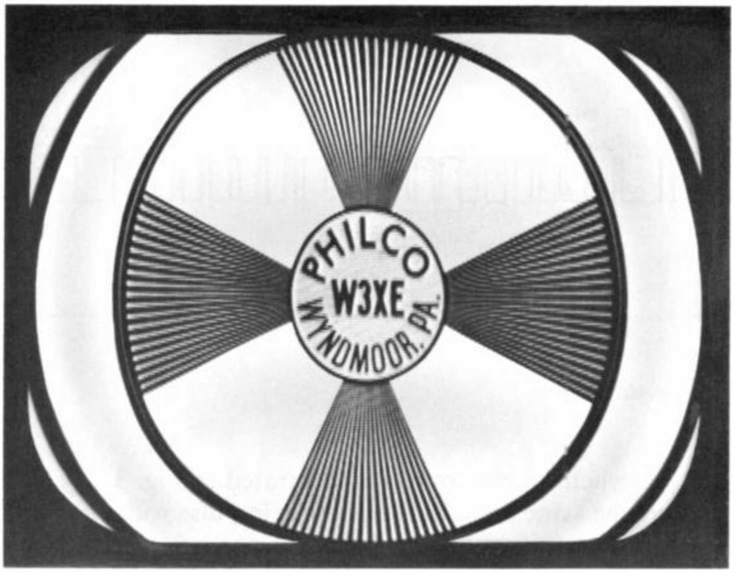


Fig. 3 · 10 Faulty interlacing. Note how the line divisions in horizontal wedges appear to be interwoven, in a moire effect. (Philco Corporation.)

picture information, when the interlacing varies in successive frames.

Poor interlacing is caused by inaccurate vertical synchronization. Although the period of a field is $\frac{1}{60}$ sec, which is a relatively long time, vertical scanning in every field must be timed much more accurately for good interlace. If the vertical timing is off by $\frac{1}{4}$ μsec in one field compared with the next, the interlaced fields are shifted the distance of one picture element.

3 · 6 The synchronizing pulses

At the receiver the kinescope scanning beam must reassemble picture elements on each horizontal line with the same left-right position as the image at the camera tube. Also, as the beam scans vertically the successive scanning lines on the kinescope screen must present the same picture elements as corresponding lines at the camera tube. Therefore, a horizontal synchronizing pulse is transmitted for each horizontal line to keep the horizontal scanning synchronized, and a vertical synchronizing pulse is transmitted for each field to synchronize the vertical scanning motion. The horizontal synchronizing pulses then have a frequency of 15,750 cps, and the frequency of the vertical synchronizing pulses is 60 cps.

The synchronizing pulses are transmitted as part of the picture signal but are sent during the blanking period when no picture information is transmitted. This is possible because the synchronizing pulse begins the retrace, either horizontal or vertical, and consequently occurs during retrace time. The synchronizing signals are combined with the picture signal in such a way that part of the modulated picture signal amplitude is used for the synchronizing pulses and the remainder for the camera signal. The term *sync* is often used for brevity to indicate the synchronizing pulses.

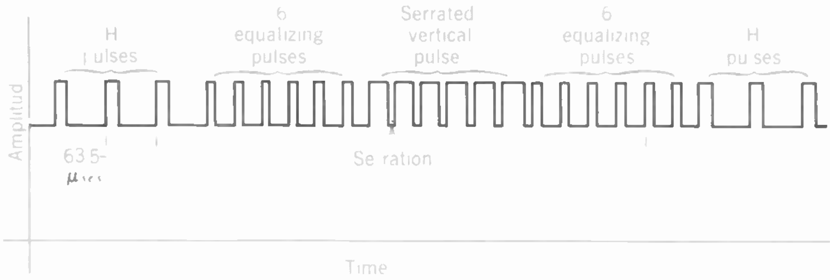


Fig. 3 · 11 The synchronizing pulses.

The form of the synchronizing pulses is illustrated in Fig. 3 · 11. Note that all pulses have the same amplitude but differ in pulse width or waveform. The synchronizing pulses shown include from left to right three horizontal pulses, a group of six equalizing pulses, a serrated vertical pulse, and six additional equalizing pulses which are followed by three more horizontal pulses. There are many additional horizontal pulses after the

(a)

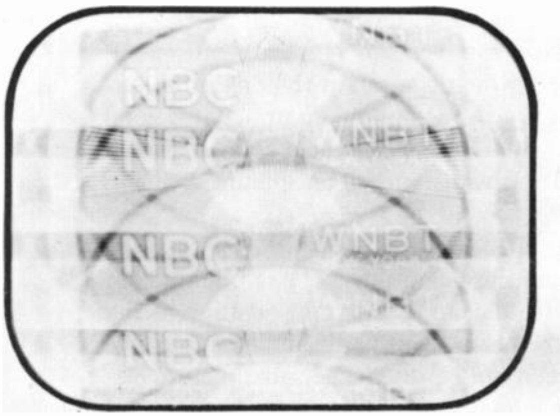
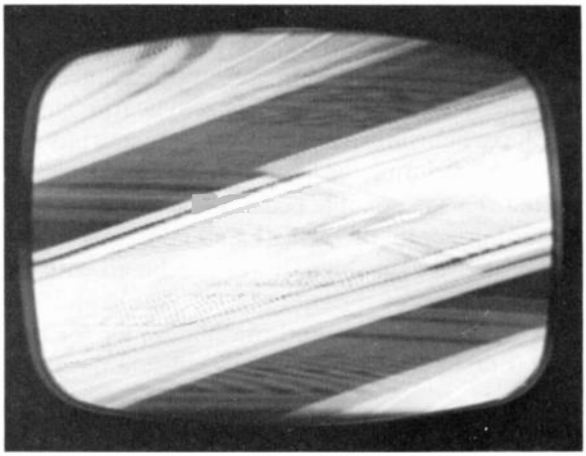


Fig. 3 · 12 Effects of no sync. (a) Picture rolling up or down without vertical sync. (b) Picture torn into diagonal segments without horizontal sync.

(b)



last one shown, following each other at the horizontal line frequency until the equalizing pulses occur again for the beginning of the next field. For every field there must be one wide vertical pulse, which is actually composed of six individual pulses separated by the five serrations.

Each vertical synchronizing pulse extends over a period equal to six half lines or three complete horizontal lines, making it much wider than a horizontal pulse. This is done to give the vertical pulses an entirely different form from the horizontal pulses. Then they can be completely separated from each other at the receiver, one furnishing horizontal synchronizing signals alone while the other provides only vertical synchronization.

The five serrations are inserted in the vertical pulse at half-line intervals. The equalizing pulses are also spaced at half-line intervals. These half-line pulses can serve for horizontal synchronization, alternate pulses being used for even and odd fields. The reason for using equalizing pulses, however, is related to vertical synchronization. Their effect is to provide identical waveshapes in the separated vertical synchronizing signal for even and odd fields so that constant timing can be obtained for good interlace.

Since the equalizing pulses are repeated at half-line intervals their repetition rate is twice 15,750, or 31,500 cps. Therefore, the horizontal pulse frequency is one-half the equalizing pulse rate. Also, the vertical pulse frequency is 60/31,500 or 1/525 of the equalizing pulse frequency. Since these are exact submultiples, the horizontal and vertical timing pulses can be obtained by frequency division of the equalizing pulses. In this way, all the timing pulses are derived from a common synchronizing signal generator at the transmitter and their frequencies are automatically interlocked in the correct ratios. It may be of interest to note that frequency-divider circuits are more practical for sync pulses, rather than the frequency multipliers used for sine waves.

The synchronizing signals do not produce scanning. Sawtooth generator circuits are needed to provide the deflection of the electron beam that produces the scanning raster. However, the sync enables the picture information reproduced on the raster to hold still in the correct position. Without vertical sync, the reproduced picture appears to roll up or down the raster; without horizontal sync the lines of picture elements are reproduced in diagonal segments. These effects are shown in Fig. 3 · 12.

SUMMARY

1. The sawtooth waveform for deflection provides linear scanning with uniform motion of the beam. The linear rise on the sawtooth is the trace part; the sharp drop in amplitude is for the retrace or flyback. Both trace and retrace are included in one cycle.
2. The frequency of the sawtooth waveform for horizontal deflection is the horizontal line rate of 15,750 cps.
3. The frequency of the sawtooth waveform for vertical deflection is the field rate of 60 cps. Vertical flyback time 5 per cent or less of 1/60 sec is long enough to include several complete lines. These horizontal lines scanned during vertical retrace are vertical flyback lines.
4. In odd-line interlacing, an odd number of lines (525) is used with an even number of fields (60), so that each field has a whole number of lines plus one-half. Then successive fields

start scanning one-half line away from the previous field, interlacing odd and even lines in the frame.

5. Interlaced scanning eliminates flicker because of the 60-cycle vertical blanking rate, while maintaining the 30-cycle rate for complete picture frames.
6. Distortions of the scanning raster include keystone, trapezoid, pincushion, and barrel effects (see Figs. 3·6 to 3·8).
7. Incorrect aspect ratio can make people in the picture look too tall or too short.
8. Nonlinear scanning spreads or crowds picture information at one end of the raster compared with the opposite end. This effect also distorts the shape of people in the picture.
9. The synchronizing pulses time the scanning with respect to the position of picture information on the raster. Horizontal sync pulses time every line at 15,750 cps; vertical sync pulses time every field at 60 cps. All the sync pulses have the same amplitude, but a much wider pulse is used for vertical sync. The equalizing pulses and the serrations in the vertical pulse occur at half-line intervals with the frequency of 31,500 cps.

SELF-EXAMINATION *(Answers at back of book.)*

Choose (*a*), (*b*), (*c*), or (*d*).

1. In the sawtooth waveform for linear scanning the: (*a*) linear rise is for flyback; (*b*) complete cycle includes trace and retrace; (*c*) sharp reversal in amplitude produces trace; (*d*) beam moves faster during trace than retrace.
2. With a vertical retrace time of 635 μsec, the number of complete horizontal lines scanned during vertical flyback is: (*a*) 10; (*b*) 20; (*c*) 30; (*d*) 63.
3. One-half line spacing between the start positions for scanning even and odd fields produces: (*a*) linear scanning; (*b*) line pairing; (*c*) fishtailing; (*d*) exact interlacing.
4. The number of lines scanned per frame in the raster on the kinescope screen is: (*a*) 525; (*b*) 262½; (*c*) 20; (*d*) 10.
5. In the interlaced frame, alternate lines are skipped during vertical scanning because the: (*a*) trace is slower than retrace; (*b*) vertical scanning frequency is doubled from the 30-cps frame rate to the 60-cps field rate; (*c*) horizontal scanning is slower than vertical scanning; (*d*) frame has the aspect ratio of 4:3.
6. With 10 per cent for horizontal flyback this time equals: (*a*) 10 μsec; (*b*) 56 μsec; (*c*) 6.4 μsec; (*d*) 83 μsec.
7. Which of the following is *not* true? (*a*) Line pairing indicates poor interlacing. (*b*) People will look too tall and thin on a square raster on the kinescope screen. (*c*) A person can appear to have one shoulder wider than the other because of nonlinear horizontal scanning. (*d*) The keystone effect produces a square raster.
8. The width of a vertical sync pulse with its serrations includes the time of: (*a*) six half lines or three lines; (*b*) five lines; (*c*) three half lines; (*d*) five half lines.
9. Sawtooth generator circuits produce the scanning raster but the sync pulses are needed for: (*a*) linearity; (*b*) timing; (*c*) keystoning; (*d*) line pairing.
10. Which of the following frequencies is wrong? (*a*) 15,750 cps for horizontal sync and scanning; (*b*) 60 cps for vertical sync and scanning; (*c*) 31,500 cps for equalizing pulses and serrations in the vertical sync pulse; (*d*) 31,500 for the vertical scanning frequency.

ESSAY QUESTIONS

1. Draw the interlaced scanning pattern for a total of 25 lines per frame, interlaced in two fields. Also show the corresponding sawtooth waveforms for horizontal and vertical scanning, as in Fig. 3·5. Assume one line scanned during each vertical flyback.
2. Define the following terms: (*a*) scanning raster; (*b*) pincushion effect; (*c*) line pairing; (*d*) interline flicker; (*e*) moire effect.

3. Why are the lines scanned during vertical trace much closer together than lines scanned during vertical flyback?
4. Suppose the sawtooth waveform for vertical scanning has a trace that rises too fast at the start and flattens at the top. Will the scanning lines be crowded at the top or bottom of the kinescope raster? How will people look in the picture?
5. Draw two cycles of the 15,750-cps sawtooth waveform, showing retrace equal to $0.08H$ to exact scale. Label trace, retrace, and time of one cycle in microseconds.
6. Draw two cycles of the 60-cps sawtooth waveform, showing retrace equal to $0.04V$ to exact scale. Label trace, retrace, and time of one cycle in microseconds.
7. Where is the electron scanning beam at the time of: (*a*) start of linear rise in *H* sawtooth; (*b*) start of *H* flyback; (*c*) start of linear rise in *V* sawtooth; (*d*) start of *V* flyback?

PROBLEMS *(Answers to odd-numbered problems at back of book.)*

1. If progressive scanning were used with a frame rate of 60 cps for the same 525 lines per frame, what would be the frequency of the vertical and horizontal sawtooth scanning waveforms?
2. How many flyback lines are produced during vertical retrace for each field and each frame for retrace time equal to (*a*) $0.02V$; (*b*) $0.08V$?
3. Compare the time in microseconds for horizontal flyback equal to $0.08H$ and vertical flyback of $0.04V$.
4. Referring to Fig. 4·4 calculate the width, in microseconds, of each horizontal sync pulse.
5. (*a*) How much time elapses between the start of one horizontal sync pulse and the next? (*b*) Between one vertical pulse in an odd field and the next in an even field?
6. What frequencies correspond to the following periods of time for one cycle: (*a*) 63.5 μsec; (*b*) 53.3 μsec; (*c*) 127 μsec?

Chapter 4 *Composite video signal*

Composite means that the video signal includes separate parts. These are (1) camera signal corresponding to the desired picture information, (2) synchronizing pulses to synchronize the transmitter and receiver scanning, and (3) blanking pulses to make the retraces invisible. How these three components are added to produce the composite video signal is illustrated in Fig. 4 · 1. The camera signal in *a* is combined with the blanking pulse in *b* and then the sync pulse is superimposed on the pedestal atop the blanking pulse to produce the composite video signal in *c*. The result shown here is composite video signal for one horizontal scanning line. With signal for all the lines, the composite video contains the information needed to reproduce the complete picture.

4 · 1 Construction of the composite video signal

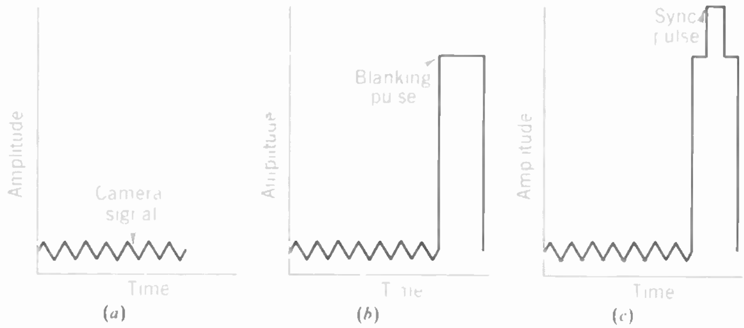
In Fig. 4 · 2, successive values of voltage or current amplitude are shown for the scanning of three horizontal lines in the image. Note that the amplitude of the video signal is divided into two sections, the lower 75 per cent being used for camera signal, with the upper 25 per cent for synchronizing pulses. In the camera signal, the lowest amplitudes correspond to the whitest parts of the picture while the darker parts of the picture have higher amplitudes. This is the way the signal is transmitted, using a standard negative polarity of transmission. Negative transmission means that white parts of the picture are represented by low amplitudes in the transmitted picture carrier signal. Higher amplitudes correspond to progressively darker picture information until the black level is reached, which is the fixed level at 75 per cent of maximum signal amplitude.

Black reference level. The black level is constant at 75 per cent amplitude and independent of picture information, in order to maintain a brightness reference in the television system. When the image is reproduced, the 75 per cent level of the video signal corresponds to the grid cutoff voltage of the picture tube and the absence of light, thus establishing a black level. The brightness values of various shades of white and gray are then defined in terms of their amplitude relative to the black level. The 75 per cent amplitude is also the pedestal level, or blanking level, because this represents the tops of the blanking pulses, providing pedestals on which the synchronizing pulses are placed. Blanking is accomplished by making the blanking level black.

Any signal amplitude greater than the black level is called *blacker than black,* because this voltage drives the picture-tube grid voltage more negative than cutoff. The synchronizing pulses are blacker than black.

The composite video signal and scanning. Referring again to Fig. 4·2, consider the amplitude variations shown as the desired video signal obtained in scanning three horizontal lines at the top of the image. Starting at the extreme left in the figure at zero time, the signal is at a white level and the scanning beam is at the left side of the image. As the first line is scanned from left to right, camera signal variations are obtained with various amplitudes that correspond to the required picture information. After horizontal trace produces the desired camera signal for one line, the scanning beam is at the right side of the image. The blanking pulse is then inserted to bring the video signal amplitude up to black level so that retrace can be blanked out.

Fig. 4·1 The three components of composite video signal. (a) Camera signal for one horizontal line. (b) Horizontal blanking pulse added to camera signal. (c) Sync pulse added to blanking pulse.



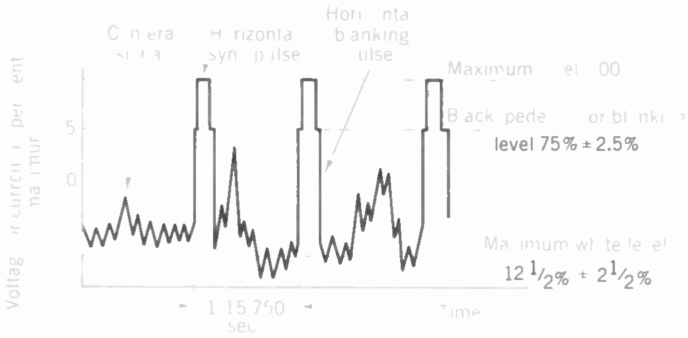


Fig. 4·2 Composite video signal for three consecutive horizontal lines.

After a blanking time long enough to include retrace, the blanking voltage is removed, since the scanning beam is then at the left side ready to scan the next line. Each horizontal line is scanned successively in this way. Notice that the second line shows dark picture information near the black level. The third line has gray values with medium amplitudes of 40 to 60 per cent.

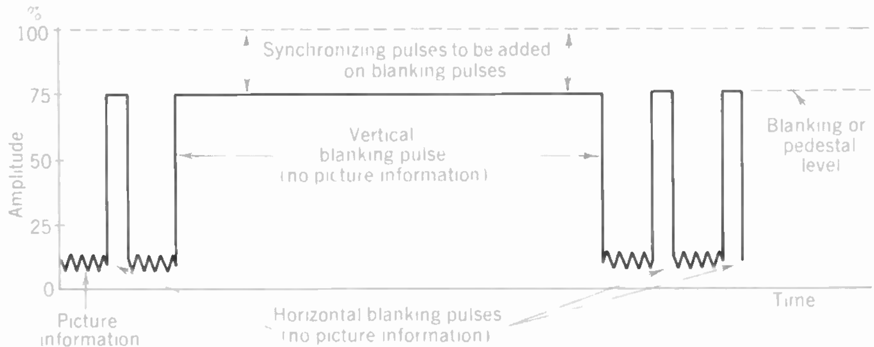
With respect to time the signal amplitudes just after blanking in Fig. 4·2 indicate information for the left side at the start of a scanning line. Just before blanking, the signal variations correspond to the right side. Information exactly in the center of a scanning line occurs at a time halfway between blanking pulses.

The blanking pulses. The composite video signal contains blanking pulses to make the retrace lines invisible by raising the signal amplitude to black level during the time the scanning circuits produce retraces. Retrace normally is produced during blanking time.

As illustrated in Fig. 4·3, there are horizontal and vertical blanking pulses in the composite video signal. The horizontal blanking pulses are included to blank out the retrace from right to left in each horizontal scanning line. The repetition rate of horizontal blanking pulses, therefore, is the line-scanning frequency of 15,750 cps. The vertical blanking pulses have the function of blanking out the scanning lines produced when the electron beam retraces vertically from bottom to top in each field. Therefore, the frequency of vertical blanking pulses is 60 cps.

Horizontal blanking time. Details in the horizontal blanking period are illustrated in Fig. 4·4. The interval between horizontal scanning lines is indicated by H. This time for scanning one complete line, including trace and retrace, equals 1/15,750 sec, or 63.5 μsec. However, the horizontal blanking pulse has a width only $0.14H$ to $0.18H$. We can consider the average of 16 per cent of the line period as a typical value. Then horizontal blanking time is 0.16×63.5 μsec for H, which equals 10.2 μsec. Subtracting from 63.5 μsec we have a remainder of 53.3 μsec as the time for visible

Fig. 4·3 Horizontal blanking pulses and vertical blanking pulses in video signal. Sync pulses not shown.



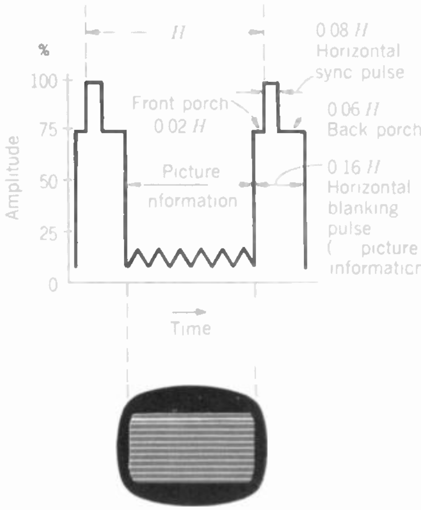


Fig. 4·4 Details of horizontal blanking and sync pulses. H equals 1/15,750 sec or 63.5 μsec.

scanning, without blanking, in each line. The 10.2 μsec for blanking allows time for retrace.

Superimposed on the pedestals provided by the tops of the blanking pulses at the black level are the narrower sync pulses. As noted in Fig. 4·4, each horizontal sync pulse is $0.08H$, or one-half the average width for the blanking pulse. This time equals 10.2/2 or 5.1 μsec.

For the remaining half of blanking time, which is also 5.1 μsec, the signal is at the pedestal level. The part of the pedestal just before the sync pulse is called the *front porch*, and the *back porch* follows the sync pulse. The front porch is $0.02H$ or 1.27 μsec and the back porch $0.06H$ or 3.81 μsec. Note that the back porch is three times longer than the front porch. In summary, then, with 10.2 μsec total blanking, 1.27 μsec is used for the front porch, 5.1 μsec for the sync pulse, and 3.81 μsec for the back porch.

The purpose of the blanking pulses is to make the retraces invisible. Furthermore, blanking time is slightly longer than typical values of retrace time, which depend on the horizontal deflection circuits in the receiver. As a result, a small part of the trace usually is blanked out at the start and end of every scanning line. This effect of horizontal blanking is illustrated by the black bars at the left and right sides of the raster in Fig. 4·4. The black at the right edge corresponds to the front porch of horizontal blanking, before retrace starts. Generally, horizontal flyback starts at the leading edge of the sync pulse. Just before retrace, when the scanning beam is completing its trace to the right, therefore, the blanking level of the front porch makes the right edge black. With a small part of every line blanked this way, a black bar is formed at the right edge. This black bar at the right can be considered as a reproduction of the front porch part of horizontal blanking.

After the front porch of blanking, horizontal retrace is produced when the sync pulse starts. The flyback is definitely blanked out because the sync level is blacker than black. Although retrace starts with the sync pulse, how much time is needed to complete the flyback depends on the scanning circuits. A typical horizontal flyback time is 7 μsec. Blanking time after the front porch is longer, however, equal to 9 μsec approximately. Therefore, 2 μsec of blanking remain after retrace is completed to the left edge. Although the blanking is still on, the sawtooth deflection waveform makes the scanning beam start its trace after flyback. As a result, the first part of trace at the left is blanked. After 2 μsec of blanked trace time at the left edge, the blanking pulse is removed. Then video signal reproduces picture information as the scanning beam continues its trace for 53.3 μsec of visible trace time. However, the small part of every line blanked at the start of trace forms the black bar at the left edge of the raster. This black edge at the left represents part of every back porch after horizontal sync.

The blanking bars at the sides have no effect on the picture other than decreasing its width slightly, compared with the unblanked raster. However, the amplitude of horizontal scanning can be increased to provide the desired width.

Vertical blanking time. The vertical blanking pulses raise the video signal amplitude to black level so that the scanning beam is blanked out during vertical retraces. The width of the vertical blanking pulse is 0.05 – 0.08 V, where V equals $\frac{1}{60}$ sec. If we take 6 per cent as an average, vertical blanking time is $0.06 \times \frac{1}{60}$ sec, which equals 0.001 sec or 1,000 μsec. Note that this time is long enough to include many complete horizontal scanning lines. Dividing 1,000 μsec vertical blanking time by the total line period of 63.5 μsec, the answer is 15.7 or 16. Approximately 16 lines are blanked out in each field, therefore, or 32 lines in the frame. This relatively long time blanks not only vertical retrace lines but also a small part of vertical trace at the bottom and top.

The sync pulses inserted in the composite video signal during the wide vertical blanking pulse are shown in Fig. 4·5. These include equalizing pulses, vertical sync pulses, and some horizontal sync pulses. The signals are shown for the time intervals between the end of one field and the next, to illustrate what happens during vertical blanking time. The two signals shown one above the other are the same except for the half-line displacement between successive fields necessary for odd-line interlacing.

Starting at the left in Fig. 4·5, the last four horizontal scanning lines at the bottom of the picture are shown with the required horizontal blanking and sync pulses. Immediately following the last visible line, the video signal is brought up to black level by the vertical blanking pulse in preparation for vertical retrace. The vertical blanking period begins with a group of six equalizing pulses, which are spaced at half-line intervals. Next is the serrated vertical sync pulse that actually produces vertical flyback in the scanning circuits. The serrations also occur at half-line intervals. Therefore, the complete vertical sync pulse is three lines wide. Following the vertical sync is another group of six equalizing pulses and a train of horizontal pulses. During this entire vertical blanking period no picture information is produced, as the signal level is black or blacker than black so that vertical retrace can be blanked out.

Notice the position of the first equalizing pulse at the start of vertical blanking in Fig. 4·5. In the signal at the top, the first pulse is a full line away from the previous horizontal sync pulse; in the signal below for the next field, the first pulse is one-half line away. This half-line difference in time between even and odd fields continues through all the following

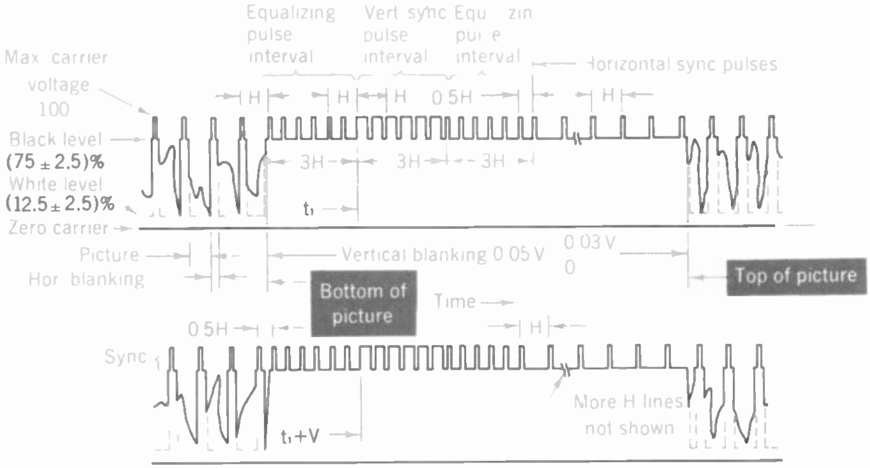


Fig. 4·5 Sync and blanking pulses for successive fields. V equals 1/60 sec.

pulses, so that the vertical sync pulses for successive fields have the timing required for odd-line interlacing.

The serrated vertical sync pulse forces the vertical deflection circuits to start the flyback. However, the flyback generally does not begin with the start of vertical sync because the sync pulse must build up charge in a capacitor to trigger the scanning circuits. If we assume vertical flyback starts with the leading edge of the third serration, the time of one line passes during vertical sync before vertical flyback starts. Also, six equalizing pulses equal to three lines occur before vertical sync. Then 3 + 1, or 4, lines are blanked at the bottom of the picture, just before vertical retrace starts.

How long the flyback is depends on the scanning circuits, but a typical vertical retrace time is five lines. As the scanning beam retraces from bottom to top of the raster, then, five complete horizontal lines are produced. This vertical retrace is easily fast enough to be completed within vertical blanking time.

With four lines blanked at the bottom before flyback and five lines during flyback, seven lines remain of the total 16 during vertical blanking. These seven blanked lines are at the top of the raster at the start of the vertical trace downward.

In summary, four lines are blanked at the bottom and seven lines at the top in each field. In the total frame of two fields, these numbers are doubled. The scanning lines that are produced during vertical trace but made black by vertical blanking form the black bars at top and bottom of the raster in Fig. 4·4. The resulting slight reduction in height of the picture with blanking, compared with the unblanked raster, is easily corrected by increasing the amplitude of the sawtooth waveform for vertical scanning.

4·2 Picture information and the video signal

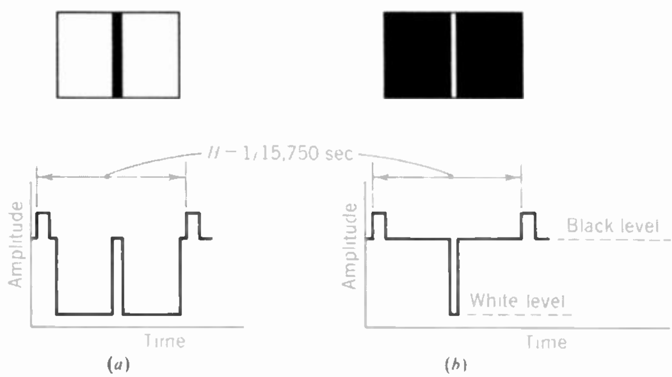
Two examples are shown in Fig. 4·6 to illustrate how the composite video signal corresponds to visual information. In *a*, the video signal corresponds to one horizontal line in scanning an image with a black vertical bar down the center of a white frame. In *b*, the black and white values in the picture are reversed from *a*.

Starting at the left in Fig. 4·6*a*, the camera signal obtained in active scanning of the image is initially at the white level corresponding to the white background. The scanning beam continues its forward motion across the white background of the frame and the signal continues at the same white level until the middle of the picture is reached. When the black bar is scanned the video signal rises to the black level and remains there while the entire width of the black bar is scanned. Then the signal amplitude drops to white level corresponding to the white background and continues at that level while the forward scanning motion is completed to the right side of the image.

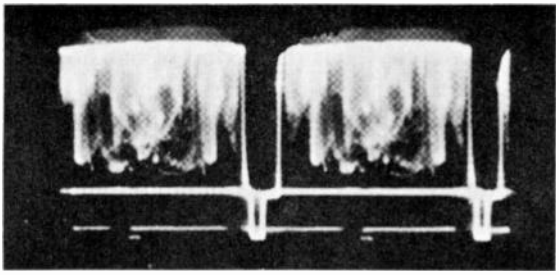
At the end of the visible trace the horizontal blanking pulse raises the video signal amplitude to black level in preparation for horizontal retrace. After retrace, the forward scanning motion begins again to scan the next horizontal line. Each successive horizontal line in the even and odd fields is scanned in this way. As a result, the corresponding composite video signal for the entire picture contains a succession of signals with a waveform identical with that shown in Fig. 4·6*a* for each active horizontal scanning line. For the image in *b* the idea is the same but the camera signal corresponds to a white vertical bar down the center of a black frame.

These are simple types of images, but the correlation can be carried over to an image with any distribution of light and shade. If the pattern contains five vertical black bars against a white background, the composite video signal for each horizontal line will include five rapid variations in

Fig. 4·6 Composite video signal and its corresponding picture information. (a) Image with black vertical line on white background. (b) White line on black background.



(a)



(b)

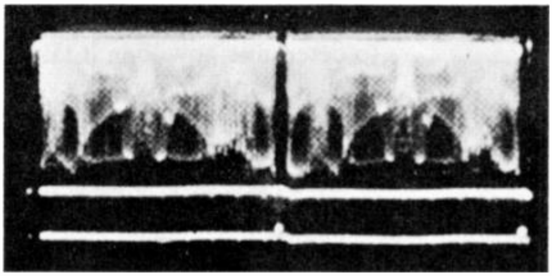


Fig. 4 · 7 Typical oscilloscope photographs of composite video signal shown with sync polarity downward. (a) Two lines of horizontal picture information between horizontal blanking and sync pulses. Oscilloscope sweep at 7,875 cps. (b) Two fields of vertical picture information between vertical blanking and sync pulses. Oscilloscope sweep at 30 cps.

amplitude from white to the black level. As another example, suppose the pattern consists of a horizontal black bar across the center of a white frame. Then most of the horizontal lines will contain white picture information for the entire trace period, with the camera signal amplitude remaining at white level except for the blanking intervals. However, those horizontal lines that scan across the black bar will produce camera signal that remains at black level for the complete active scanning time.

A typical image consists of picture elements having various degrees of light and shade with a nonuniform distribution. In this case, then, the corresponding video signal contains a succession of varying signals. Within each horizontal line there are variations in camera signal amplitude for different picture elements. Furthermore, the waveforms of camera signal for each horizontal line vary for different lines in the frame. See the oscilloscope photographs of typical video signal in Fig. 4 · 7.

4 · 3 Video frequencies and picture information

Referring to the checkerboard pattern in Fig. 4 · 8, the square-wave signal shown represents the camera signal variations of the composite video signal obtained in scanning one horizontal line at the top of the image. It is desired to find the frequency of this square wave. The frequency of the camera signal variations is very important in determining whether or not the television system can transmit and reproduce the corresponding picture information.

In determining the frequency of any signal variation, the time for one complete cycle must be known. A cycle includes the time from one point on the signal waveform to the next succeeding point with the same magnitude and direction. The frequency can then be found as the reciprocal of the period for one cycle. Thus the period of one horizontal scanning line is 1/15,750 sec and the line-scanning frequency is 15,750 cps. The camera signal variations within one horizontal line, however, necessarily have a shorter period and a higher frequency.

Note that one complete cycle of camera signal in Fig. 4 • 8 includes the information in two adjacent picture elements, one white and the other black. Only after scanning the second square does the camera signal have the same magnitude and direction as at the start of the first square. Therefore, to find the frequency of the camera signal variations it is necessary to determine how long it takes to scan across two adjacent squares. This time is the period for one cycle of the resultant camera signal.

Now the period of one complete cycle of the square-wave camera signal variations in Fig. 4 • 8 can be calculated. The horizontal line period is 1/15,750 sec, or 63.5 μsec, including trace and retrace. With a horizontal blanking time of 10.2 μsec, the time remaining for visible trace equals 53.3 μsec. This is the time to scan across all the picture elements in a line. For 12 squares across one line in 53.3 μsec, the beam scans two squares in $^2/_{12}$ or $^1/_6$ of 53.3 μsec. Then 53.3/6, or 8.9 μsec, is the time to scan two squares. This time is the period of one complete cycle of the square-wave signal. The reciprocal of 1/8.9 μsec is the frequency, therefore, which equals 0.11×10^6 cps or 0.11 Mc. This is the frequency of the square-wave camera signal variations in Fig. 4 • 8.

When a typical picture is scanned the scattered areas of light and shade do not produce symmetrical square-wave signal. However, the differences of light and shade correspond to changes of camera signal amplitude in the same way. The frequency of the resultant camera signal variations always depends on the time to scan adjacent areas with different light values. When large objects with a constant white, gray, or black level are scanned the corresponding camera signal variations have low frequencies because of the comparatively long time between changes in level. Smaller areas of light and shade in the image produce higher video frequencies. The highest signal frequencies correspond to variations between very small picture elements in a horizontal line.

The ability of the television system to transmit and reproduce the high video frequencies determines how well the horizontal detail in the image can be reproduced. High frequencies in the signal are associated with fine detail within the lines because the beam moves fast in horizontal scanning. However, the highest video frequency that can be transmitted is restricted to approximately 4 Mc with the use of 6-Mc channels for commercial television broadcasting.

At the opposite extreme, signal variations corresponding to picture elements adjacent in the vertical direction have low frequencies because the vertical scanning is comparatively slow. Variations between one line and the next correspond to a frequency of approximately 10 kc. Slower changes over larger distances in vertical scanning produce frequencies below 10 kc.

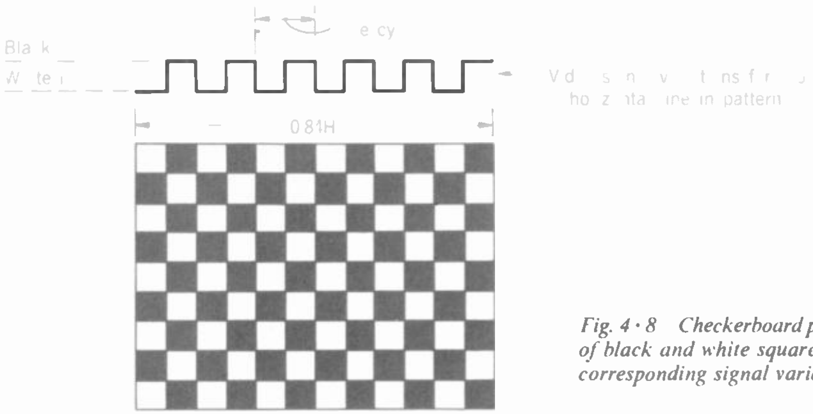


Fig. 4·8 Checkerboard pattern of black and white squares with corresponding signal variations.

The very low frequency of 30 cps corresponds to a variation in light level between two successive fields.

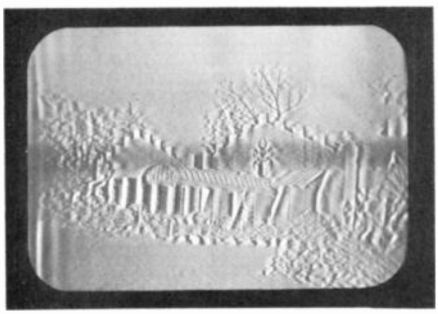
Figure 4·9 shows how the size of the picture information can be considered in terms of video frequencies. The main body of the image with the larger areas of black and white is reproduced in *b* with video frequencies up to 0.1 Mc or 100 kc. However, the detail with sharp edges and outlines is filled in by the high video frequencies from 0.1 to 4 Mc, as shown in *c*. Notice that the canopy of the building is reproduced in *b* but its stripes and the small lettering need the high-frequency reproduction in *c*.

(*a*)

Fig. 4·9 Effect of video frequencies on picture reproduction. (a) Normal picture reproduction. (b) Only large areas in picture reproduced by low video frequencies up to 0.1 Mc. (c) Only horizontal edges and outlines reproduced by high video frequencies between 0.1 and 4 Mc.

(*b*)

(*c*)



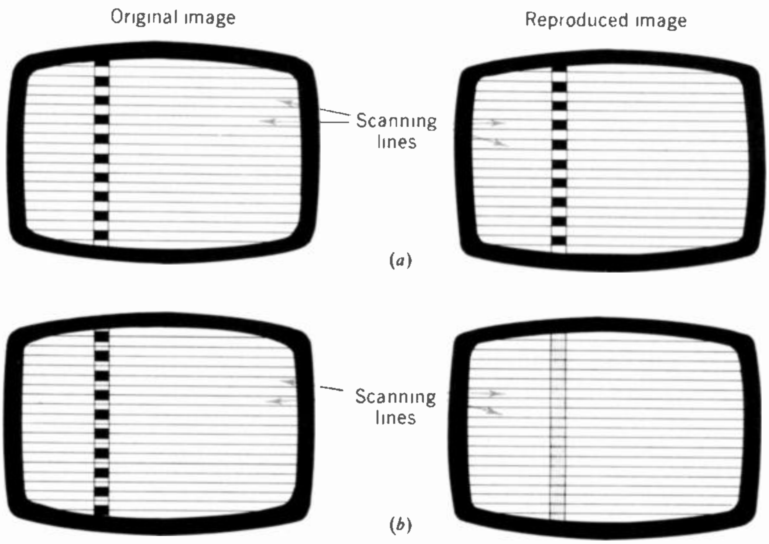
4 · 4 Maximum number of picture elements

If we consider a checkerboard pattern such as Fig. 4 · 8 with many more squares, the maximum possible number of picture elements can be calculated where each square is one element. The total elements in the area equal the maximum details in a line horizontally, multiplied by the details in a vertical row. However, horizontal detail and vertical detail must be considered separately in a television picture because of the scanning process. For horizontal detail, the problem is to determine how many elements correspond to the high-frequency limit of 4-Mc video signal. The vertical detail is a question of how many elements can be resolved by the scanning lines.

Maximum horizontal detail. Proceeding in the same manner as in the previous section, the number of elements corresponding to 4 Mc can be determined to show the maximum number of picture elements in a horizontal line and the size of the smallest possible horizontal detail. The period of one complete cycle for a 4-Mc signal variation is $1/(4 \times 10^6)$ sec, or 0.25 μsec. This is the time required to scan two adjacent picture elements. With two elements scanned in 0.25 or ¼ μsec, then eight elements are scanned in 1 μsec. Finally 8×53.3, or 426, picture elements can be scanned during the entire active line period of 53.3 μsec. If there were 426 squares in the horizontal direction in the checkerboard pattern in Fig. 4 · 8, therefore, the resultant camera signal variations would produce a 4-Mc signal.

In order to reproduce the squares of the checkerboard pattern as individual, discrete elements, a square-wave signal is needed. Response up to

Fig. 4 · 10 Vertical detail depends on how the scanning lines cover the picture elements. (a) Each line reproduces an individual black or white detail. (b) Scanning lines straddle the vertical details.



about the fifteenth harmonic of the fundamental frequency is required to reproduce a square wave. A 60-Mc sine-wave signal, equal to the fifteenth harmonic of 4-Mc, would be necessary, therefore, but this video frequency is beyond the capabilities of the television system. As a result, the maximum number of 426 horizontal details in a television picture can be reproduced only as continuous variations in shading for a 4-Mc sine-wave video signal, instead of individual, discrete elements corresponding to 4-Mc square-wave signal.

Utilization ratio and vertical detail. Each scanning line can represent at best only one detail in the vertical direction. However, a scanning line may represent no vertical detail at all. The two opposite cases are illustrated in Fig. 4 · 10 where the image to be scanned is a vertical bar containing a number of alternate black and white squares. The height of each square is considered to be equal to the width of a scanning line. When a square in the image has a position such that the scanning beam passes directly over it, as in *a*, the corresponding camera signal represents the vertical detail perfectly. This is the best possible case, and the reproduced pattern corresponds exactly to the original image. For the case illustrated in *b*, however, the details in the image are so placed that the scanning beam passes over the boundary between a black and a white square. Then the camera signal variation corresponds to a gray level intermediate between the black and white details, representing the average brightness of the two elements. When the scanning beam covers two picture elements in this way the vertical details are entirely lost. Instead, the reproduced image becomes the uniform gray bar in *b*.

Typical picture content has a nonuniform arrangement of elements. Some fall directly on a scanning line while others straddle the scanning lines. The problem in establishing the useful vertical detail, then, is determining how many picture elements can be reproduced along a vertical line by a given number of scanning lines. This factor depends on the average number of elements that can be expected to fall directly on a scanning line when there is a random distribution of light and dark picture elements. The ratio of the number of scanning lines useful in representing the vertical detail to the total number of visible scanning lines is called the *utilization ratio*. Theoretical calculations and experimental tests show that the utilization ratio ranges from 0.6 to 0.8 for different images with typical picture content. We can use 0.7 as an average.

Now the maximum possible number of vertical elements can be determined. The number of visible lines equals 525 minus those scanned during vertical blanking. With a vertical blanking time of 6 per cent the number of lines blanked out for the entire frame is 0.06×525, or approximately 32 lines. Some of these lines occur during vertical retrace, while others are scanned at the top and bottom of the frame but all are blanked out. Therefore, 493 visible lines remain. The number of lines useful in showing vertical detail is 483×0.7 since this is the utilization ratio, providing 338 effective lines. Therefore, the maximum number of vertical details that can

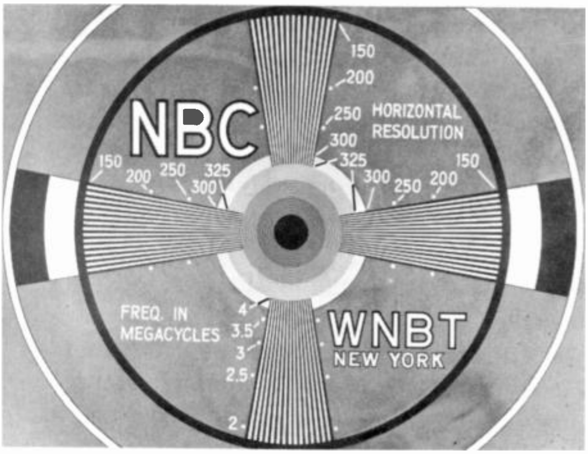


Fig. 4 · 11 NBC test pattern. (Copyright National Broadcasting Company.)

be reproduced with 525 total and 493 visible scanning lines is about 338, the exact value depending upon the utilization ratio.

Total number of picture elements. On the basis of the previous calculations, the maximum number of picture elements possible for the entire image is 426 × 338, or about 150,000. This number is independent of picture size. With different picture content, there may be 100,000 to 200,000 picture elements. Since the total number of picture elements can be regarded as the figure of merit it may be compared with motion-picture reproduction. A single frame of 35-mm motion-picture film contains about 500,000 picture elements. The smaller 16-mm frame contains one-fourth as many, or about 125,000. The televised reproduction, therefore, can have about the same amount of details as 16-mm motion pictures. The detail in a 16-mm film reproduction is superior to a television picture, however, because the picture elements in film are reproduced as discrete units.

4 · 5 Test patterns

In order to adjust a television system conveniently and compare performances, a standard picture is desirable. This is usually in the form of a *test pattern* (see Fig. 4 · 11). The test pattern is composed of black and white lines with a gray background halfway between black and white. The black and white areas at the outer ends of the horizontal wedges produce a signal that permits the voltage amplitude swing between white and black to be readily determined. The target circles range from black in the center to white on the outside in equal steps of gray.

The aspect ratio is 4:3. Proper aspect ratio is obtained when the height of the picture is equal to the diameter of the large black inner circle and the diameter of the large white outer circle equals the width of the picture frame. Linearity of the scanning motions may be judged by the circles after the proper width and height have been set. If the circles appear round, the scanning linearity is properly adjusted. Linearity can also be checked by

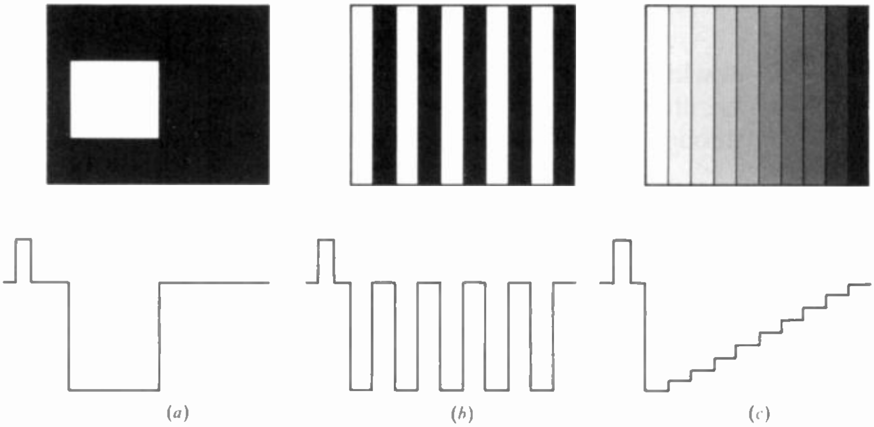


Fig. 4 · 12 Patterns for test purposes, with corresponding video signal for one horizontal line. (a) Window signal. (b) Bar pattern. (c) Staircase signal.

the wedges. The two vertical wedges are of equal length, as are the two horizontal wedges. When the top and bottom wedges are reproduced with the same length, the vertical scanning is linear; equal lengths for the two side wedges indicate linear horizontal scanning.

Resolution. The picture detail or resolution is measured on the test pattern in number of lines. If the vertical resolution is 150 lines in the reproduced picture, this means that it is possible to see 150 individual horizontal lines consisting of 75 black lines separated by 75 white lines. For equal resolution in the horizontal direction and an aspect ratio of 4:3, $150 \times 4/3$, or 200, vertical lines can be resolved in the picture, consisting of 100 black lines separated by 100 white lines. However, this is still considered 150-line resolution because the resolution is measured in terms of the picture height when indicating either horizontal or vertical detail, in order to provide a common basis for comparison.

The line divisions in the side wedges measure vertical resolution. They also indicate good interlacing when there is little moire effect in the diagonal lines.

Additional forms of test patterns for specific applications include the Electronic Industries Association[1] (EIA) resolution chart, which is like the NBC test pattern but more detailed. Resolution up to 600 lines and 6-Mc response can be checked with the EIA chart. Another test pattern features reproduction of an Indian head, which enables checking the quality of gray tones in an actual picture.

More specific test signals are illustrated in Fig. 4 · 12. The *window signal* in *a* provides maximum white and black in large areas with a sharp transition to check edge distortion, streaking, and smearing. In *b* the uniformly spaced black and white bars can be used to check scanning linearity. The bars may be vertical to check horizontal linearity, horizontal to check ver-

[1] Formerly Radio Manufacturers Association (RMA).

tical linearity, or both may be used in a *crosshatch* pattern. The *staircase signal* in *c* illustrates uniform changes in signal amplitude from white, progressing through gray values in equal steps to black.

4·6 D-c component of the video signal

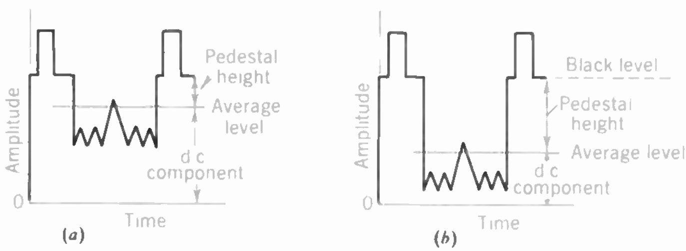
In addition to continuous amplitude variations for individual picture elements, the video signal must have an average value corresponding to the average brightness in the scene. Otherwise the receiver cannot follow changes in brightness. As an example of the importance of brightness level, the a-c camera signal for a gray picture element on a black background will be the same as signal for white on a gray background, if there is no average-brightness information to indicate the change in background.

Dark and light scenes. The average level of a signal is the arithmetic mean of all the instantaneous values measured from the zero axis. In Fig. 4·13*a* the average level is higher than in *b* because the camera signal variations have higher amplitudes. Now it is important to remember that for any signal variation, its average value for a complete cycle is its d-c component. Therefore, the d-c component in *a* is closer to the pedestal level than in *b*. Although illustrated here for one scanning line, for convenience, the required d-c component of the video signal is its average value for complete frames, since the background information of the frame indicates the brightness of the scene.

When the average value or d-c component of the video signal is close to the pedestal level, as in Fig. 4·13*a*, the average brightness is dark, since the axis is close to black reference level. The same a-c signal variations in *b* have a lighter background because the d-c axis is farther from the black pedestal level.

In the reproduced picture, its brightness depends on the kinescope d-c bias set by the brightness control. Furthermore, the d-c component of the video signal coupled to the kinescope can add or subtract from the bias to make the brightness lighter or darker. The average brightness, therefore, depends on how far the average-value axis is from the black level.

Fig. 4·13 Video signals with same a-c variations but different average brightness values. Only one line of frame illustrated. (a) Dark scene with average value close to black level. (b) Light scene with average value farther from black level.



Pedestal height. As noted in Fig. 4 · 13, the pedestal height is the distance between the pedestal level and the average-value axis of the video signal. This indicates average brightness since it measures how much the average value differs from the black level. Although not the d-c component measured from the zero axis, the pedestal height is a convenient measure of the average brightness because the distance between the pedestal and average-value levels stays the same if the signal loses its d-c component. With the pedestal level as the fixed black reference level, therefore, the pedestal height can always indicate the relative brightness for different video signals.

Setting the pedestal level. The method of using the pedestal level for black reference voltage can be followed by starting with the camera signal. The camera output voltage is amplified in several stages before being coupled to a control amplifier, where sync and blanking are added. At this point the camera signal has no d-c component, since the d-c level is blocked by capacitive coupling in either the camera tube or the amplifier stages.

To produce composite video signal, the sync pulses are superimposed on the pedestals provided by the blanking pulses. Before sync is added, though, the tops of blanking pulses are cut off by a clipper stage in the control amplifier (see Fig. 4 · 14). The level at which blanking pulses are clipped becomes the pedestal level that determines the black reference voltage for the entire system.

Note that setting the clipping level determines the pedestal height in the video signal. In Fig. 4 · 14, clipping lower on the blanking pulses reduces the pedestal height, making the brightness darker, as the average-value axis is then closer to the black reference. For the opposite case, a higher clipping level means a greater pedestal height to shift the black reference farther from the average axis and the signal will have a lighter background.

The bias and cutoff voltage of the clipper stage set the clipping level. At what level the pedestals will be clipped is decided in terms of average brightness in the scene being televised. The video control operator who observes the scene at the studio sets the level for the desired brightness in

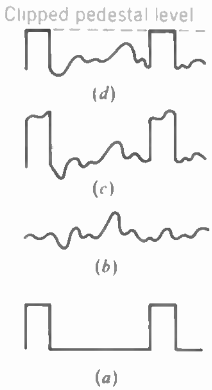


Fig. 4 · 14 Formation of pedestals by clipping top of blanking pulses. (a) Blanking pulses alone. (b) Camera signal without blanking. (c) Blanking pulses added to camera signal. (d) Pulses clipped to provide pedestal level for sync pulses.

the reproduced picture which he is viewing on a monitor kinescope.

D-c insertion. This means adding a d-c component to an a-c signal. Note that adjusting the clipping level of the video signal is equivalent to d-c insertion because its average value is shifted.

Once the d-c insertion has been accomplished, the pedestal level becomes the black reference and the pedestal height indicates correct relative brightness for the reproduced picture. However, the d-c level inserted in the control amplifier is usually lost in succeeding stages because of capacitive coupling. Still, the correct d-c component can be reinserted when necessary because the pedestal height remains the same.

D-c reinsertion. This means restoring the d-c component to a signal that has lost its d-c level. The d-c reinsertion is accomplished by a circuit that rectifies the a-c video signal to produce a d-c component proportional to the pedestal height. This procedure automatically reinserts the correct relative amounts of d-c component.

At the transmitter, d-c reinsertion is used for the video signal that modulates the transmitted picture carrier. With the correct d-c component in the video signal, all sync pulses are in line at the pedestal level so that pedestal voltage can produce a constant 75 per cent of peak amplitude in the transmitted picture carrier signal. In addition, d-c reinsertion may be needed in the control-grid circuit of the picture tube. Here the correct d-c component in the video signal keeps the black pedestal voltage at the grid-cutoff voltage of the kinescope. Then the blanking level is black and all light values are reproduced in their correct relation to black.

The signal could be transmitted without d-c reinsertion in the modulation, allowing the receiver to reinsert the d-c component. However, the constant pedestal level and sync pulse amplitudes provide greater efficiency at both the transmitter and receiver. The basic reason is that a constant peak amplitude allows maximum peak-to-peak signal voltage to be used in modulating an amplifier without producing overload distortion.

4 · 7 Gamma

This is a numerical factor used in television and film reproduction for indicating how light values are expanded or compressed. Referring to Fig. 4 · 15, the exponent of the equations for the curves shown is called *gamma* (γ). The numerical value of gamma is equal to the slope of the straight-line part of the curve where it rises most sharply. A curve with a gamma of less than one is bowed downward as in *a* of Fig. 4 · 15, with the greatest slope at the start and the relatively flat part at the end. When the gamma is more than one the curve is bowed upward as in *b*, making the start comparatively flat while the sharp slope is at the end. With a gamma of one the result is a straight line as in *c*, where the slope is constant.

A gamma value of one means a linear characteristic that does not exaggerate any light values. When gamma is greater than one for the white parts of the image, the reproduced picture looks "contrasty" because the increases in white level are expanded by the sharp slope, to emphasize the

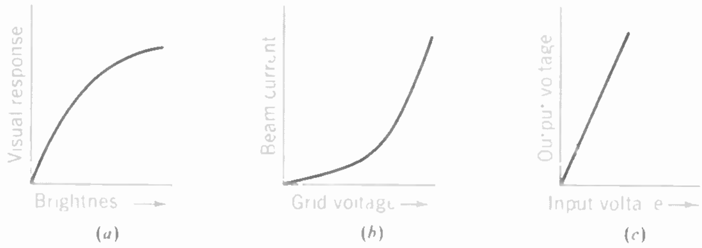


Fig. 4·15 Gamma characteristics. (a) Visual response of the eye; gamma less than one. (b) Control-grid characteristic of picture tube; gamma approximately three. (c) Linear characteristic of an amplifier; gamma equals one.

white parts of the picture. Commercial motion pictures shown in a darkened theater have this high-contrast appearance. Gamma values of less than one for the white parts of the image compress the changes in white levels to make the picture appear softer, with the gradations in gray level more evident.

Any component in the television system can be assigned a value of gamma to describe the shape of its response curve and contrast characteristics. As a typical example, picture tubes have the control-characteristic curve illustrated in *b* of Fig. 4·15. The video signal voltage is always impressed on the control grid of the picture tube with the polarity required to make the signal variations for the white parts of the picture fall on that part of the response curve with the steep slope. As a result, a variation in video signal amplitude at the white level produces a greater change in beam current and screen brightness than it would at a darker level. Therefore, picture tubes emphasize the white parts of the picture, with typical gamma values of 2.5 to 3.5. Commercial film also has a gamma greater than one, an average value being 1.5. In general, monochrome pictures are reproduced with high gamma to make up for the loss of color contrasts.

Amplifiers have a gamma characteristic that is very nearly unity, using linear operation (see Fig. 4·15*c*). The straight-line response shows that output signal voltage is proportional to input voltage without emphasizing any signal level. If desired, however, an amplifier can be made to operate over the curved portion of its transfer-characteristic curve by shifting the operating bias. The nonlinear amplifier can be used as a *gamma-control* stage, therefore, to expand or compress the white video signal amplitudes relative to the black level.

The image orthicon camera tube has an operating characteristic that is essentially linear with a gamma of one. However, the gamma-control stage may reduce gamma for the transmitted signal. The purpose of emphasizing dark parts of the signal is to minimize the effects of noise in dark parts of the picture. At the receiver, the high gamma of the kinescope provides the high contrast desired for the picture reproduction.

SUMMARY

1. The composite video signal includes camera signal with picture information, synchronizing pulses to time scanning, and blanking pulses to blank out retraces.
2. The signal is transmitted with the top 25 per cent in amplitude for sync, and the remaining 75 per cent for camera signal. Specifically, the peak amplitude is tip of sync; 75 per cent level is pedestal, blanking, or black reference level; maximum white is 12.5 ± 2.5 per cent amplitude. Gray values are between 12.5 per cent for white and the 75 per cent black level.
3. Horizontal blanking pulses at 15,750 cps blank out retrace for every line by raising the signal to 75 per cent black level. Average pulse width for horizontal blanking is 0.16H or 10.2 μsec. Horizontal sync pulses with a width of 5.1 μsec are on this pedestal provided by the top of each blanking pulse.
4. Vertical blanking pulses at 60-cps blank vertical retrace for every field by raising the signal to 75 per cent black level. Much longer than horizontal blanking time, a typical value is 0.06V for vertical blanking. This equals the time for 16 lines blanked in every field to make sure that the vertical flyback lines are blanked out.
5. The camera signal variations in the video signal correspond to the picture information in the image. Amplitude changes between the 12.5 and 75 per cent levels indicate variations in light level between white and black.
6. The high video frequencies in the camera signal correspond to horizontal detail. Approximately 4 Mc is the highest video frequency that can be broadcast in the 6-Mc transmission channel.
7. The average utilization ratio of 0.7 means 70 per cent of the visible scanning lines are useful in showing details in the vertical direction.
8. A test pattern usually includes black, white, and gray lines and areas to check picture reproduction. Vertical wedges indicate horizontal resolution as the beam scans across individual line divisions. The ability to resolve divisions in the side wedges indicates vertical resolution. The side wedges also show poor interlacing by moire effect in the diagonal lines. In addition, equal lengths for the horizontal wedges show linear horizontal scanning; equal lengths for the vertical wedges show linear vertical scanning.
9. The d-c component of any signal is its average-value axis. The pedestal height is measured from the average-value axis to the pedestal level.
10. Gamma is a numerical factor indicating how contrast is expanded or compressed. Picture tubes have a characteristic curve with gamma more than one, which emphasizes white signal voltages.

SELF-EXAMINATION *(Answers at back of book.)*

Answer true or false.

1. The three components of composite video signal are camera signal, blanking pulses, and sync pulses.
2. Sync pulses transmitted during vertical blanking time include equalizing pulses, the serrated vertical sync pulse, and horizontal sync pulses.
3. During the front porch time before a horizontal sync pulse the scanning beam is at the left edge of the raster.
4. The 10 per cent amplitude in composite video signal corresponds to maximum white picture information.
5. The video signal is at the 75 per cent black level during horizontal blanking but not during vertical blanking.
6. The visible trace time for one horizontal line is 10.2 μsec.
7. When the vertical blanking pulse starts, the scanning beam is at the top of the raster.
8. The equalizing pulses and serrations in the vertical sync pulse are spaced at half-line intervals.
9. The horizontal blanking pulses can produce vertical black bars at the sides of the raster.

10. The vertical sync pulse for one field starts a half line away from its timing in the previous field.
11. Camera signal variations between successive horizontal blanking pulses correspond to information from left to right in the picture.
12. The picture has the left half white and the right half black. The corresponding signal frequency in scanning across one line is approximately 19 kc.
13. The picture has the top half white and the bottom half black. The corresponding signal frequency in scanning vertically through one frame is slightly more than 60 cps.
14. The high video frequencies correspond to horizontal detail in the picture.
15. Ability to resolve individual lines in the top and bottom wedges of the test pattern indicates horizontal resolution.
16. Ability to resolve individual lines in the side wedges of the test pattern indicates vertical resolution.
17. The utilization ratio is the proportion of unblanked scanning lines to total lines.
18. Picture tubes have a gamma value greater than one, emphasizing white to increase contrast in the reproduced picture.
19. Average brightness of the reproduced picture depends on the d-c bias of the kinescope grid.
20. The pedestal height is a measure of brightness by indicating how far the average-value axis is from black level.

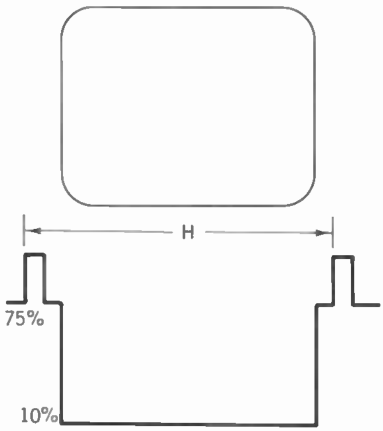
ESSAY QUESTIONS

1. Show the picture and draw the composite video signal of two consecutive lines in scanning across the following patterns: (*a*) all-white frame; (*b*) two vertical white bars and two black bars equally spaced; (*c*) 10 pairs of vertical bars. Why does this signal have a higher frequency than in *b*?
2. Why are the synchronizing pulses inserted during blanking time?
3. What is the function of the horizontal blanking pulses? The vertical blanking pulses?
4. Why are the horizontal blanking pulses wider than the horizontal sync pulses?
5. Trace the motion of the scanning beam from the beginning to end of vertical blanking.
6. Define: pedestal level, white level, pedestal height, gamma, utilization ratio, resolution, crosshatch pattern, and staircase signal.
7. Why do thin vertical lines produce higher video signal frequencies than wide vertical bars?

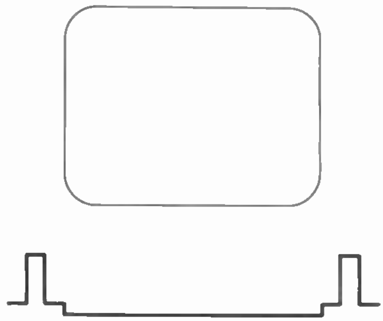
PROBLEMS *(Answers to odd-numbered problems at back of book.)*

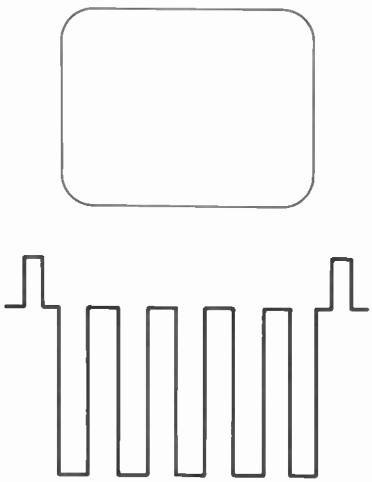
1. In the checkerboard pattern of Fig. 4 · 8, if there are 300 squares in a line, what is the frequency of the corresponding signal variations? Use 53.3 μsec for visible trace time.
2. With a utilization ratio of 0.7, what would be the maximum vertical detail for a vertical blanking time of 0.08 V?
3. In the test pattern of Fig. 4 · 11, calculate the vertical resolution at a point on the side wedge exactly midway between the 150 and 200 markers.
4. In the test pattern of Fig. 4 · 11, calculate the horizontal resolution at a point on the bottom wedge exactly between the markers for 2.5 and 3.5 Mc, measured in number of lines referred to picture height. With this resolution, how many horizontal details can be reproduced along a horizontal line?
5. Assume a facsimile reproduction with specifications of 200 lines per frame, progressive scanning, and five frames per second. Calculate the following: (*a*) time to scan one line, including trace and retrace; (*b*) visible trace time for one line with 4 per cent blanking; (*c*) Video frequency corresponding to 100 total black and white elements in a line.
6. Calculate the frequency of video signal produced in horizontal scanning of the window signal in Fig. 4 · 12. Assume 53.3 μsec visible trace time.

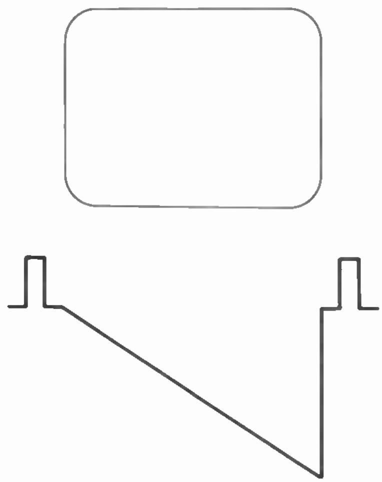
7. Referring to the Table of Television Standards for foreign countries in Appendix C, calculate the maximum vertical details with a utilization ratio of 0.72 for the systems in: (*a*) United States; (*b*) England; (*c*) Western Europe; (*d*) France.
8. For the same systems as in Question 7 calculate the time to scan one line, including trace and retrace.
9. Show the pictures corresponding to the video signals below. Assume that all the lines of video signal are the same as the one shown.



(*a*)


(*b*)


(*c*)


(*d*)

Chapter 5 Picture carrier signal

The method of transmitting the AM picture signal is similar to the more familiar system of sound transmission in the standard broadcast band for radio, where the amplitude of an r-f carrier wave is made to vary at the audio rate. In television broadcasting the composite video signal modulates a high-frequency carrier wave to produce the amplitude-modulated picture signal[1] illustrated in Fig 5 · 1. The amplitude of the transmitted carrier wave varies with the video modulating signal. In this way the desired camera signal, blanking pulses, and synchronizing pulses are transmitted to the receiver as the envelope of the modulated picture carrier signal. At the receiver the picture signal is detected to recover the composite video signal, which is then used to reproduce the picture.

5 · 1 Negative transmission

The transmitted picture carrier signal in Fig. 5 · 1 is shown with the negative polarity of modulation that is an FCC standard for all commercial television broadcast stations. Negative transmission means that changes toward white in the picture decrease the amplitude of the AM picture carrier signal.

In Fig. 5 · 1 the tips of sync voltage produce the peaks of r-f amplitude in the AM carrier wave. This peak carrier amplitude is its 100 per cent level. Pedestal level in the composite video signal is transmitted at the constant level of 75 per cent of peak carrier amplitude. Smaller amplitudes in the modulated r-f carrier correspond to picture information that varies between black and maximum white. The whitest parts of the picture produce a carrier amplitude 10 to 15 per cent of the peak value. All these relative amplitudes are the same for the top or bottom of the envelope be-

[1] The term *picture signal* is used here for the modulated r-f carrier wave, while *video* represents the signal that can be used directly to reproduce the desired visual information when applied to a picture tube, corresponding to *audio* in a sound system.

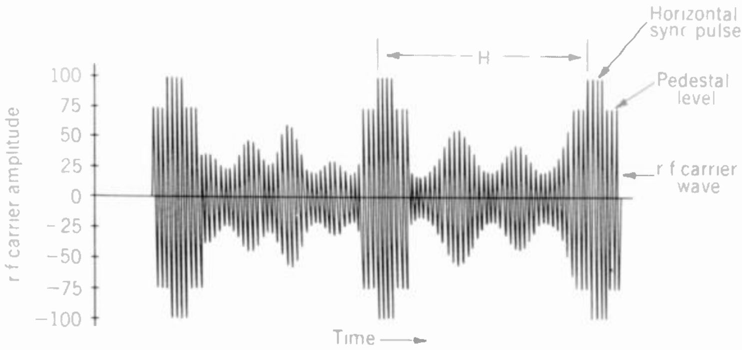


Fig. 5·1 Transmitted picture carrier wave, amplitude-modulated by composite video signal.

cause the modulated r-f carrier wave has a symmetrical envelope of amplitude variations.

The negative transmission refers to the polarity of video modulating voltage, not the individual cycles of r-f carrier voltage, which are both positive and negative. When the video modulation polarity is chosen to make the carrier amplitude decrease for white video voltage, the transmitted carrier has negative polarity of modulation. If the video modulating voltage had opposite polarity, the transmitted carrier would have positive polarity of modulation.

It may be of interest to note that, instead of negative transmission, positive transmission is used for television broadcasting in England and France.[2] With positive modulation, maximum white produces peak carrier amplitude and sync voltage reduces the carrier level toward zero. Each system has its own merits. However, negative transmission is standard practice for all broadcast stations in the United States so that receivers tuned to any station will produce a normal picture. The wrong polarity of video signal will reverse the light values in the reproduced picture with black instead of white and white for black, as in a film negative.

5·2 Vestigial-side-band transmission

The AM picture signal is not transmitted as a normal double-side-band signal. Instead, some of the side-band frequencies are filtered out before transmission in order to reduce the bandwidth of the channel needed for the modulated picture signal. To see how this vestigial-side-band transmission is accomplished, we can consider first the idea of how amplitude modulation produces side-band frequencies.

Amplitude modulation. In Fig. 5·2 an r-f carrier wave is amplitude-modulated by a sine-wave audio signal in a plate-modulation arrangement. For simplicity the r-f carrier frequency is taken as 100 kc and the audio as 5,000 cps. The B+ voltage for the r-f power amplifier is assumed to be

[2] See appendix C for television standards in countries other than the United States.

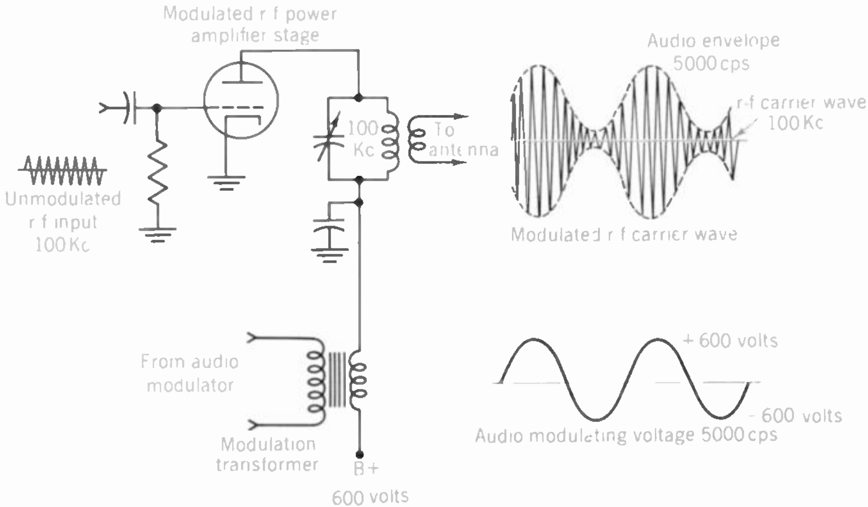


Fig. 5·2 Plate modulation circuit. The 100-kc r-f carrier is amplitude-modulated by 5,000 cps audio modulating voltage.

Table 5·1 Modulation values for Fig. 5·2

Audio voltage	*B+ voltage*	*R-f amplifier plate voltage*	*Modulated r-f signal amplitude*
0	600	600	Carrier level
+600	600	1,200	Double carrier level
0	600	600	Carrier level
−600	600	0	Zero
0	600	600	Carrier level

600 volts and the peak value of the audio sine-wave modulating voltage is also 600 volts, allowing 100 per cent modulation.

Note that the audio voltage across the secondary of the modulation transformer is in series with the B+ supply and the r-f amplifier plate circuit. Therefore, the audio modulating voltage varies the plate voltage of the r-f amplifier at the audio rate (see Table 5·1).

The varying amplitudes of the r-f carrier wave provide an envelope that corresponds to the audio modulating voltage. Both the positive and negative peaks of the r-f carrier wave are symmetrical above and below the center axis and have exactly the same amplitude variations. The envelope is symmetrical because the changes in amplitude of the negative and positive half cycles of the r-f signal are equal, as the carrier amplitude is varied at the audio rate, which is much slower than the r-f variations. Any point on the audio waveform includes many cycles of the r-f carrier. The result of the modulation in this case, then, is to produce an r-f carrier wave at a frequency of 100 kc with an amplitude that varies at the audio rate of

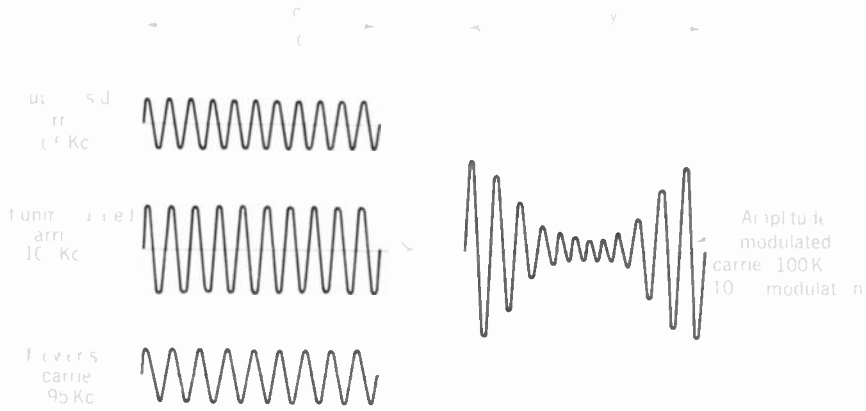


Fig. 5·3 Equivalence of amplitude-modulated wave to carrier plus two side carriers produced by modulation.

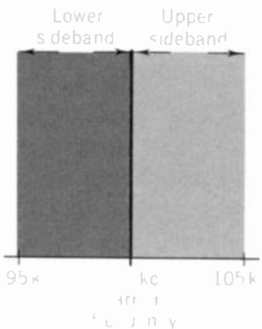


Fig. 5·4 Double side bands resulting from amplitude modulation of 100-kc carrier with all frequencies up to 5,000 cps.

5,000 cps. Either the top or bottom envelope of the amplitude-modulated carrier wave corresponds to the 5,000-cps audio modulating signal.

Side-carrier frequencies. Referring now to Fig. 5·3, it is shown that the AM wave is equal to the sum of the unmodulated r-f carrier and two side-carrier frequencies. Notice that the carrier and its equivalent side frequencies all have a constant level. Also, the amplitude of the side carriers equals one-half the unmodulated carrier level, for 100 per cent modulation. Each side frequency differs from the carrier by the audio modulating frequency. The upper side frequency is 105 kc and the lower side frequency 95 kc in this illustration.

The question as to whether the transmitted signal is a carrier with varying amplitudes or a constant-amplitude carrier with its two side carriers is without meaning, because the two concepts are the same. The constant-level side carriers plus the unmodulated carrier wave are equal to the AM carrier signal. Or, the AM carrier wave is equal to the unmodulated carrier plus two side carriers of proper amplitude, phase, and frequency. The equivalence of the two signals is due to the fact that the modulated r-f carrier wave is distorted slightly from true sine-wave form by the audio amplitude variations, producing new frequency components, which are the side frequencies.

The r-f side-carrier frequencies should not be confused with the audio envelope. The envelope is an audio frequency. The side carriers are radio frequencies close to the carrier frequency. For the case here, the envelope is the audio modulating signal of 5,000 cps while the r-f side frequencies are 105 and 95 kc. If the audio modulating frequency is 1,000 cps, the r-f side frequencies will be 101 and 99 kc. Then the modulated carrier will have an audio envelope of 1,000 cps.

When the carrier is modulated by a voltage that includes many frequency components, each audio modulating frequency produces a pair of r-f side frequencies. In each pair, one side frequency is higher than the carrier frequency and one is lower. All the higher side frequencies can be considered as the *upper side band* of the carrier with all the lower side frequencies the lower side band. This idea is illustrated in Fig. 5·4 for the case of modulation with a continuous band of audio frequencies from d-c to 5,000 cps. The graph here indicates the corresponding side frequencies for a 100-kc carrier. The upper side band includes all side frequencies from the carrier frequency of 100 kc up to 105 kc; the lower side band includes the side frequencies down to 95 kc. The bandwidth required for the two side bands in this case is ± 5 kc centered around the carrier frequency of 100 kc, or a total bandwidth of 10 kc. Note that the required bandwidth for the AM carrier with two side bands is double the highest modulating frequency.

The fact that different audio modulating frequencies produce different side frequencies in AM should not make it be confused with frequency modulation. In FM the r-f carrier frequency varies in accordance with the amount of audio modulating *voltage*, but in AM the side frequencies depend on the audio modulating *frequency*.

Vestigial side bands. Note that the information of the modulating signal is in the side bands of the amplitude-modulated r-f carrier. Frequency of the modulation is indicated by how much the side frequencies differ from the carrier frequency; modulating voltage is indicated by the amplitude of the two side carriers. For the case of 100 per cent modulation, each side carrier has one-half the unmodulated carrier amplitude. Furthermore, the upper and lower side frequencies have the same information, since they are of equal amplitude and each differs from the carrier frequency by the same amount. The desired modulating signal can be transmitted by one side band, therefore, and it does not matter whether the upper or lower side band is used. With only one side band, the transmitted signal has the advantage of only one-half the bandwidth of two side bands. The amplitude modulation normally produces double side bands but one side band can be filtered out if desired.

Figure 5·5 illustrates just one side frequency transmitted with the carrier. Notice that the resultant modulated wave has amplitude variations for only 50 per cent modulation, instead of the 100 per cent modulation produced with both side bands. In Fig. 5·5, the modulated wave varies in amplitude 50 per cent above and below the unmodulated carrier amplitude, but with 100 per cent modulation the peak carrier amplitude doubles the unmodulated level and decreases to zero. Therefore, a signal transmitted with one side band has effectively one-half the per cent modulation,

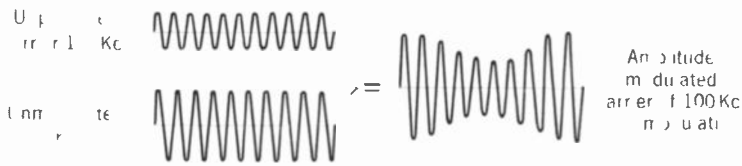


Fig. 5·5 Equivalence of carrier and one side carrier to amplitude-modulated wave.

compared with double-side-band transmission. The same reduction factor of one-half applies when the varying modulating voltage produces different amounts of modulation less than 100 per cent.

Except for the amount of amplitude swing, the envelope of the carrier plus one side band can be considered essentially the same as with double-side-band transmission, although there is slight distortion for high percentages of modulation. Notice that the envelope in Fig. 5·5 still corresponds to the audio modulating voltage. Furthermore, the envelope is not cut off at the top or bottom of the AM wave. Remember that one r-f side frequency is filtered out but not the audio envelope. In order to remove one part of the envelope it would be necessary to rectify the modulated carrier signal.

The method just described can be considered single-side-band transmission with the carrier. In many applications of single-side-band transmission, however, the carrier itself is not transmitted in order to save power by suppressing the carrier. Then only one side band is transmitted. In this case, the carrier must be reinserted at the receiver to detect the signal.

A compromise between double-side-band transmission and the single-side-band method is used for broadcasting the AM picture carrier signal. In this system, called *vestigial-side-band transmission*, all of one side band is transmitted but only a part, or vestige, of the other side band. The picture carrier itself is transmitted. More specifically, all of the upper side band of the AM picture signal is transmitted, to include all video modulating frequencies up to 4 Mc. The lower side band, however, includes only video modulating frequencies up to 0.75 Mc approximately, to conserve bandwidth in the broadcast channel.

5·3 The television channel

Each television broadcast station is assigned a channel 6 Mc wide for transmission of the AM picture signal and the FM sound signal. Vestigial-side-band transmission is used for the picture signal, so that video modulation frequencies up to 4 Mc can be broadcast in the 6-Mc channel.

Assigned channels. Since the picture carrier frequency must be much higher than the highest video modulating frequency of 4 Mc, the television channels are assigned in the VHF band of 30 to 300 Mc and the UHF band of 300 to 3,000 Mc. Table 5·2 lists the channels and frequencies assigned by the FCC for commercial television broadcast stations in the

Table 5·2 Television channel allocations

Channel number	*Frequency band, Mc*	*Channel number*	*Frequency band, Mc*
1*	....	42	638–644
2	54–60	43	644–650
3	60–66	44	650–656
4	66–72	45	656–662
5	76–82	46	662–668
6	82–88	47	668–674
7	174–180	48	674–680
8	180–186	49	680–686
9	186–192	50	686–692
10	192–198	51	692–698
11	198–204	52	698–704
12	204–210	53	704–710
13	210–216	54	710–716
14	470–476	55	716–722
15	476–482	56	722–728
16	482–488	57	728–734
17	488–494	58	734–740
18	494–500	59	740–746
19	500–506	60	746–752
20	506–512	61	752–758
21	512–518	62	758–764
22	518–524	63	764–770
23	524–530	64	770–776
24	530–536	65	776–782
25	536–542	66	782–788
26	542–548	67	788–794
27	548–554	68	794–800
28	554–560	69	800–806
29	560–566	70	806–812
30	566–572	71	812–818
31	572–578	72	818–824
32	578–584	73	824–830
33	584–590	74	830–836
34	590–596	75	836–842
35	596–602	76	842–848
36	602–608	77	848–854
37	608–614	78	854–860
38	614–620	79	860–866
39	620–626	80	866–872
40	626–632	81	872–878
41	632–638	82	878–884
		83	884–890

* The 44- to 50-Mc band was television channel 1 but is now assigned to other services. See Appendix D, Development of Television Broadcasting.

United States. The television channel frequencies can be considered in three groups: the five low-band channels 2 to 6 in the VHF range; seven high-band channels 7 to 13 also in the VHF range; and the 70 UHF channels 14 to 83.[3] Frequencies between these television broadcasting bands are used by other radio services.

The number of channels available for television broadcast stations in any one locality depends upon its population, varying from one channel in a smaller city to nine for New York City, including VHF and UHF channels. Most cities have one channel reserved for a noncommercial educational television broadcast station.

One channel can be used by many broadcast stations, but they must be far enough apart to minimize interference between them. Such stations using the same channel are *co-channel stations.* They must be separated by 170 to 220 miles for VHF channels or 155 to 205 miles for UHF channels. Stations that use channels adjacent in frequency, like channels 3 and 4, are *adjacent-channel stations.* To minimize interference between them, adjacent-channel stations are separated by 60 miles for VHF channels or 55 miles for UHF channels. However, channels consecutive in number but not adjacent in frequencies, such as channels 4 and 5, channels 6 and 7, or channels 13 and 14, can be assigned in one area because they are not adjacent-channel stations.

The standard channel. The structure of a standard television channel is illustrated in Fig. 5·6*a*. The width of the channel is 6 Mc, including the picture and sound carriers with their side-band frequencies. The picture carrier is spaced 1.25 Mc from the lower edge of the channel, and the sound carrier is 0.25 Mc below the upper edge of the channel. As a result there is always a fixed spacing of 4.5 Mc between the picture and sound carrier frequencies. The specific picture and sound carrier frequencies for all television broadcast channels are listed in Appendix A.

The standard channel characteristics shown in Fig. 5·6 should not be interpreted as an illustration of the picture signal. The graph merely defines the signal frequencies that can be transmitted in the television channel, with their relative amplitudes. The picture carrier is shown with twice the amplitude of the side-band frequencies, which are their relative amplitudes for 100 per cent modulation.

Since the sound signal is frequency-modulated its side-band frequencies do not have the same type of amplitude characteristic as in the picture signal, and these are not shown. The sound carrier signal is a conventional FM signal, with a bandwidth of approximately 50 kc for a frequency swing of $\pm$ 25 kc.

In the picture signal all upper side-band frequencies up to approximately 4-Mc video modulation are transmitted with their normal amplitude, as are all lower side-band frequencies that differ from the carrier frequency by 0.75 Mc or less. However, the lower side-carrier frequencies that differ from the picture carrier by more than 0.75 Mc but less than 1.25 Mc are

[3] After Apr. 30, 1964, all receivers sold in interstate commerce must be able to tune in UHF channels as well as VHF channels.

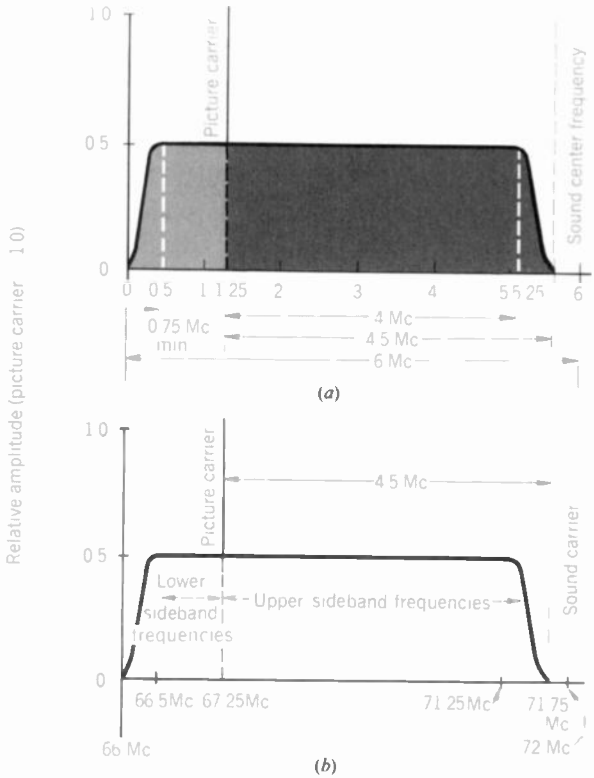


Fig. 5·6 The standard commercial television broadcast channel. (a) Frequency separations for any channel. (b) Frequencies for channel 4, 66 to 72 Mc.

gradually attenuated. Lower side-carrier frequencies below the picture carrier by 1.25 Mc or more are outside the channel. These frequencies must be completely filtered out at the transmitter so that they will not be radiated to interfere with the lower adjacent channel. Note that upper side-carrier frequencies more than 4 Mc above the picture carrier frequency are also attenuated to eliminate interference with the associated sound signal.

Numerical values for channel 4 as a typical television channel are shown in Fig. 5·6*b*. The channel has a bandwidth of 6 Mc from 66 to 72 Mc. The picture carrier is 1.25 Mc above the lower edge of the channel, which is 67.25 Mc for this channel. The sound carrier is 71.75 Mc, 4.5 Mc above the picture carrier frequency. With vestigial-side-band transmission, the upper side-band frequencies to 71.25 Mc and lower side-band frequencies to 66.5 Mc, approximately, are transmitted without attenuation. As an example, when the video modulating voltage has a frequency of 0.75 Mc, both the upper and lower side frequencies of 68 and 66.5 Mc are transmitted without attenuation. For this case, the picture carrier is a normal double-side-band signal. The same is true for any video modulating signal having a frequency less than 0.75 Mc.

However, for components of video modulating signal with a frequency higher than 0.75 Mc only the upper side carrier is transmitted with normal amplitude. For 2-Mc video modulation, as an example, the upper side frequency of 69.25 Mc is in the channel. The lower side frequency is 65.25 Mc, which is outside channel 4 and must be filtered out at the transmitter. In this case, then, only the upper side frequency is transmitted with the picture carrier, resulting in single-side-band transmission. The result is a vestigial-side-band transmission system because double-side-band transmission is used for video modulating frequencies lower than 0.75 Mc but single-side-band transmission is used for higher video modulating frequencies up to 4 Mc, approximately.

The advantage of using vestigial-side-band transmission can be seen from the fact that the picture carrier is 1.25 Mc from the end of the channel, allowing video modulating frequencies up to 4 Mc to be transmitted in the 6-Mc channel. A video-frequency limit of about 2.5 Mc would be necessary if double-side-band transmission were used with the picture carrier at the center of the channel. This would represent a serious loss in horizontal detail, since the high-frequency components of the video modulation determine the amount of horizontal detail in the picture.

It might seem desirable to place the picture carrier at the lower edge of the channel and use single-side-band transmission completely, allowing the use of video modulating frequencies higher than 5 Mc and increased horizontal detail, but this is not practicable. The elimination of undesired side-carrier frequencies is accomplished by a filter circuit at the transmitter, which cannot have ideal cutoff characteristics. Therefore, it would not be possible to remove side carriers that are too close to the carrier frequency without introducing objectionable phase distortion for the lower video signal frequencies, which causes smear in the picture.

The practical compromise of vestigial-side-band transmission that is used provides for complete removal of the lower side band only where the side-carrier frequencies are sufficiently removed from the picture carrier to avoid phase distortion. The picture carrier itself and all side frequencies close to the carrier are not attenuated. The net result is normal double-side-band transmission for the lower video frequencies corresponding to the

Fig. 5·7 Horizon distance r depends on antenna height h.

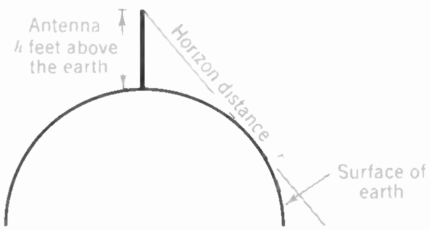
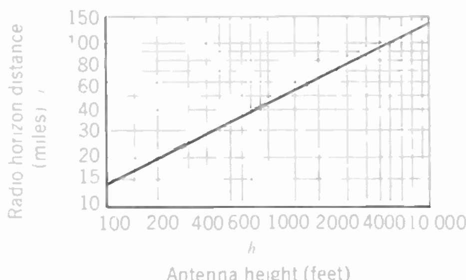


Fig. 5·8 Graph showing how the radio horizon distance increases with antenna height.



main body of picture information for large areas in the picture, while single-side-band transmission is used only for the higher video frequencies that represent details of edges or outlines in the picture.

It should be noted that the vestigial-side-band transmission distorts the picture signal in terms of relative amplitude for different frequencies. Remember that a signal transmitted with only a single side band and the carrier represents 50 per cent modulation in comparison with a normal double-side-band signal with 100 per cent modulation. Therefore, the higher video frequencies provide signals with one-half the effective carrier modulation produced by the lower video frequencies that are transmitted with both side bands. This is in effect a low-frequency boost in the video signal. However, it is corrected by deemphasizing the low video frequencies to the same extent in the i-f amplifier of the television receiver.

5 · 4 Line-of-sight transmission

Propagation of radio waves in the VHF and UHF bands is produced mainly by ground-wave effects, rather than sky waves from the ionized atmosphere. The ground wave is that part of the radiated signal affected by the presence of the earth and can be considered as being propagated along the surface of the earth from the transmitting antenna. Since the television broadcast channels are in the VHF and UHF bands, transmission of the picture and sound carrier signals is determined primarily by ground-wave propagation.

Horizon distance. The transmission distance that can be obtained for the ground-wave signal is limited by the distance along the earth's surface to the horizon, as viewed from the transmitting antenna. This is called *line-of-sight* transmission, and the line-of-sight distance to the horizon is the *horizon distance* (see Fig. 5 · 7). The horizon distance for the transmitted radio wave, however, is about 15 per cent longer than the optical horizon distance because the path of the ground wave curves slightly in the same direction as the earth's curvature. This bending of the radio waves by the earth's atmosphere is called *refraction.* The graph in Fig. 5 · 8 shows the radio horizon distance directly for any antenna height up to 10,000 ft.

Figure 5 · 9 shows several television transmitting antennas mounted at the top of the Empire State Building in New York City, in order to increase the antenna height and horizon distance. This antenna height is about 1,500 ft, providing a radio horizon distance of approximately 50 miles.

When considering the line-of-sight distance from the transmitting antenna to the receiving antenna, the horizon distance of each must be added. For an antenna height of 150 ft at the receiver, as an example, the radio horizon distance is approximately 17 miles, and line-of-sight communications could be obtained with a transmitting antenna atop the Empire State Building for a distance of 17 miles plus 50 miles, or 67 miles. The transmitting and receiving antennas should be mounted as high as possible, therefore, for line-of-sight transmission over appreciable distances.

Service area. Although the service area for reliable reception is within

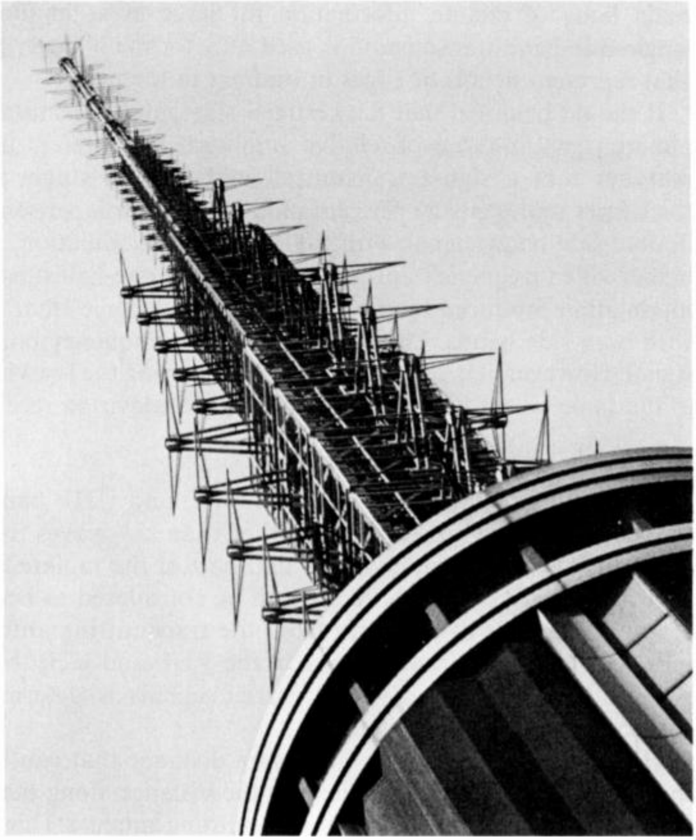


Fig. 5·9 VHF transmitter antennas atop Empire State Building in New York City.

the radio horizon distance, the strength of the ground-wave signal decreases rapidly with distance from the transmitter. In addition, higher frequencies have greater propagation losses. The service area is determined by either measuring or computing the boundary or contour within which the field strength of the transmitted signal is a minimum acceptable level.

Field strength is indicated by a receiving antenna at a height of 30 ft and is measured in microvolts per meter of antenna length. The minimum field strength for grade A service in cities and built-up areas is 2,510 μv per meter for channels 2 to 6 (54 to 88 Mc), 3,550 μv per meter for channels 7 to 13 (174 to 216 Mc), and 5,010 μv per meter for the UHF channels 14 to 83. Notice that more signal is needed for higher channels because of lower sensitivity at the receiver. For grade B service in rural areas, the minimum field strength required is lower because of less interference. The service area, therefore, may extend 25 to 75 miles from the station, depending on antenna height, radiated power, and channel frequencies for the transmitter. The smaller distances are for UHF channels. Greater distances can be obtained by increasing the height of the receiving antenna.

In unusual cases that depend on atmospheric conditions, television carrier frequencies may be returned from the ionosphere to provide reception over very long distances beyond the horizon. This condition probably results from *scatter propagation* in the earth's atmosphere but is not considered in the service area.

Reflections. As the ground wave travels along the surface of the earth, the radio signal encounters buildings, towers, bridges, hills, and other obstructions. When the intervening object is a good conductor and its size is an appreciable part of the radio signal's wavelength, the obstruction will reflect the radio wave, similar to the reflection of light from a mirror or other reflecting surface. What happens is that the conductor intercepts the radio wave, current flows as in an antenna, and the conductor reradiates the signal. Reflection of radio waves can occur at any frequency but more easily at higher frequencies because of the shorter wavelengths. For the television channel frequencies between 54 and 890 Mc the wavelengths are between 17 and 1 ft, depending on the frequency. Objects of comparable size, or bigger, can reflect the television carrier waves. When the reflected picture carrier signal arrives at the receiving antenna in addition to the direct wave or along with other reflections, the multipath signals produce multiple images called *ghosts* in the reproduced picture.

Shadow areas. Where an object in the path of the ground wave reflects the radio signal, the area behind the obstruction is shadowed and therefore has reduced signal strength. The shadowing effect is more definite at higher frequencies because of the shorter wavelengths, just like reflection of the radio waves. Reception of television signal in shadow areas behind an obstruction like a tall building is often accomplished by utilizing waves reflected from other buildings nearby.

Booster stations. Some areas are either shadowed by mountains or too far from the nearest transmitter for satisfactory television broadcast service. In this case, a booster station can be used. The station is in a suitable location for reception and rebroadcasts the program to receivers in the local area. Some booster stations convert the VHF channel frequencies for rebroadcasting on an unused UHF channel, to minimize interference problems. These are *translator* stations. Both booster and translator stations charge an annual fee.

Community television systems. This is another method of providing service to isolated areas. The signal from a distant station is received at the site of a master antenna. After being amplified, the signal is distributed by coaxial cable to subscribers who pay for this private service. The signal is usually converted to a low-band VHF channel for minimum cable losses.

Stratovision. Much greater line-of-sight transmission distance can be obtained by broadcasting from an airplane. The UHF channels 66 to 83 are available for experimentation with this airborne television system called *stratovision.* In one arrangement, an airplane circling at an altitude of 23,000 ft broadcasts educational television programs reaching most of six surrounding states.

Satellites for world-wide television broadcasting. One method is the

Telstar[4] project sponsored by the Bell System, American Telephone and Telegraph Company. A photograph is shown in Fig. 5·10. Its orbit is 575 to 3,500 miles high. Such an orbit has a period of 2½ hr for each pass around the earth. The satellite is visible in both the United States and Europe for ½ hr of each pass for testing purposes. However, additional satellites can be used for more available time. In the Syncom project, the orbits for three satellites are designed to be constant with respect to the earth's rotation. Then signals can be transmitted by line of sight from earth to one satellite, relayed between satellites and then back to earth.

The Telstar satellite contains one broad-band repeater amplifier for relaying either one television signal or the equivalent of about 600 telephone channels. Transmission from the ground station at Andover, Maine, is at the microwave carrier frequency of 6,390 Mc, with a power of 2 kw. The signal received by the satellite is amplified and shifted to the carrier frequency of 4,170 Mc, to be retransmitted with a power of 2 watts. For control functions, the VHF carrier frequencies of 120 and 136 Mc are used.

It should be noted that the satellite serves as a relay station, to provide signal for local stations broadcasting on their assigned channel frequencies. With satellites as relay stations, television programs can be broadcast between countries in any part of the world. Although different countries may use different scanning standards, as listed in Appendix C, the video signal can be converted from one set of standards to another. Television broadcasts from Europe and England have been on the United States standards.

5·5 *Television broadcasting*

A commercial television broadcast station includes equipment for producing camera signal, forming the composite video signal, and transmitting the picture signal. Since the associated sound must also be broadcast, audio facilities and an FM transmitter for the sound signal are included. The equipment used in broadcasting a televised scene can be considered in two parts: the studio and the transmitter. At the studio the camera pickup generates the camera signal to which blanking and synchronizing pulses are added to produce the composite video signal. The amount of video signal delivered to the transmitter is about 2 volts peak to peak across 75-ohm coaxial cable. At the transmitter, the video is amplified enough to modulate the picture carrier, and the modulated picture signal is radiated from the transmitting antenna.

The studio normally includes facilities for direct pickup of live-talent shows, motion-picture film, and video tape recordings. In addition, field equipment may be used for remote pickups in televising sport features and other special events outside the studio. Video signal from field equipment is relayed to the studio before being broadcast.

[4] For more details of the Telstar project, see H. E. Weppler, *IRE Trans. Broadcast Television Receivers,* April, 1962.

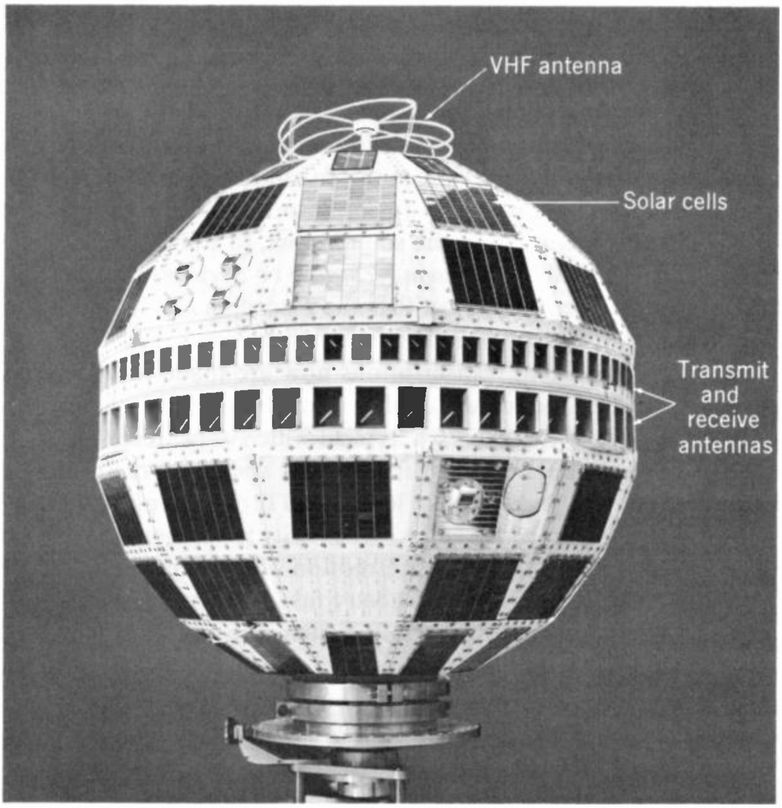


Fig. 5·10 Telstar satellite is 3 ft in diameter and weighs 170 lb. (A. T. & T. Long lines Department.)

Live-talent studio. Live-talent programs are staged in a television studio, as shown in Fig. 1·4, or in a theater for big musical productions. Two or more cameras provide close-ups or wide-angle views for long shots and allow switching from one scene to another. Each camera has a turret with three or four lenses of different focal length. One may be a *Zoomar lens,* which is a telephoto lens of variable focal length easily adjusted for quick close-up shots, while maintaining correct focus. Image orthicon cameras are generally used, with 35-mm optical equipment.

Televising motion-picture films. A separate studio is employed for televising film. Projectors are available for 35- and 16-mm motion-picture film and for slides, which can be used for titles, station identification, and commercial advertisements. The projector throws the light image directly onto the image plate of the television film camera. A mirror *triplexer* can be used to enable the image from any one of three projectors to be televised with one film camera. The film camera usually employs the iconoscope or the vidicon camera tube. When motion-picture film is televised, a special projector is used which allows the film to travel at the speed of 24 frames per second but projects 60 light images of the scene per second, instead of the normal 48. As a result, the time for 60 scanning fields or 30 television frames matches 24 film frames.

Television transcriptions. In order to distribute the same program material to many broadcast stations, or to transcribe a live-talent program to be shown some other time, the picture and sound can be recorded. The picture reproduced on the screen of a monitor kinescope at the studio is photographed on special high-quality 16- or 35-mm motion-picture film, and the audio signal is recorded on the sound track, to transcribe the program on film. This is called a *kinescope recording* or *teletranscription.* The film is run at 24 frames per second, so that the recording can be projected like commercial motion-picture film.

Video tape recording. Most transcriptions now are made by recording on magnetic tape because the picture quality is much better than kinescope film recordings. Figure 5·11 shows a typical video tape recorder. The magnetic tape is 2 in. wide and 1 mil thick. Tape speed for recording or playback is 15 in. per sec. A 14-in. reel with 7,200 ft of tape plays 96 min.

The main problem in recording video signal is its frequency range from 30 cps to 4 Mc. For high frequencies the motion of the tape with respect to the gap of the recording head must be very fast. This problem is solved by rotating the video heads transversely across the tape, as it moves

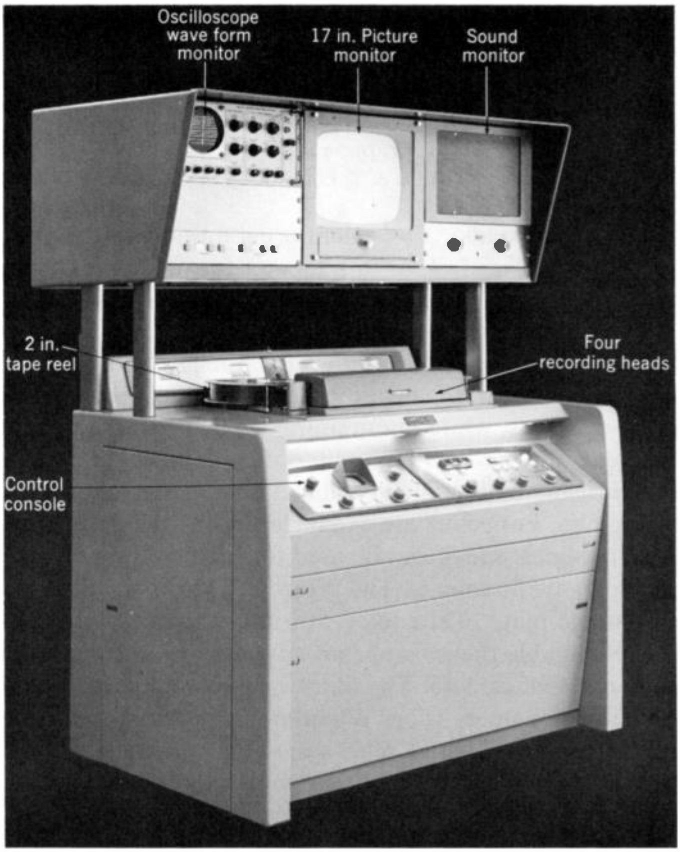


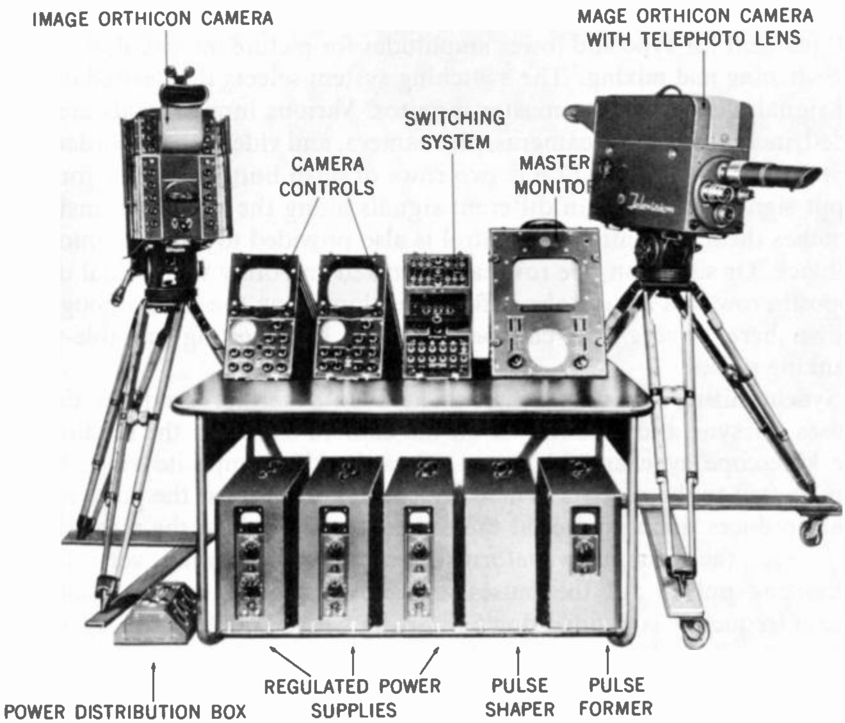
Fig. 5·11 Video tape recorder. (Ampex Corporation.)

lengthwise between reels. There are four record-or-playback heads mounted on a wheel rotating at 14,400 rpm. The heads are phased 90° apart so that the four gaps contact the tape in sequential order. As one head moves off the bottom of the tape, the next head comes on at the top. The video signal then is recorded as a series of transverse tracks across the tape. In addition to the video head assembly, provision is made for recording audio.

The problem of bandwidth is minimized by converting the video signal to a frequency-modulated signal for recording, so that all signal frequencies are in the megacycle range. A center carrier frequency of approximately 5.5 Mc is used for the FM signal with narrow-band deviation. The required bandwidth is still 4 Mc but the ratio of highest signal frequency to the lowest is much smaller in the modulated signal. Another advantage of using the FM signal is that its constant amplitude eliminates the need for bias current in recording. In playback, the FM signal from the tape is demodulated to recover the original video signal.

Camera chain. A single chain includes one camera with its control equipment. Figure 5·12 shows two image orthicon cameras, their two camera control units, and associated equipment needed for both camera chains. The portable suitcase-type units are convenient for remote pickups

Fig. 5·12 Portable field equipment for two camera chains. (RCA.)



in the field but can also be used for studio work. Within the camera head are the camera tube, deflection and blanking circuits for the camera tube, and a preamplifier to supply about 0.5 volt camera signal output for the control unit. In addition, an electronic view finder on the head displays the televised image on a small kinescope. The control unit provides remote control of gain, black level, beam current, target voltage, and electrical focus in the camera tube. Optical focus is varied by the cameraman.

Camera control and monitoring. As the video control operator views the televised scene, picture quality is monitored by adjusting gain and black level of the video signal, after the camera-tube controls have been set for best picture. The picture reproduction is observed on the kinescope monitor while the oscilloscope monitor shows video signal waveform. How the composite video signal looks on the monitor oscilloscope is shown in Fig. 5 · 13, for two horizontal lines. Note that sync is down on the oscilloscope screen, which is general practice for studio monitors. The blanking or pedestal level is set 5 to 10 per cent from black peaks in the camera signal. This difference in voltage between maximum black in the picture and the black blanking level is called *setup.* With this setup, black in the picture cannot extend past blanking into the sync amplitudes. The standard IRE scale of 140 divisions for video signal at the studio is used as follows: 40 divisions sync, 10 divisions black setup, and 90 divisions picture information between maximum white and black. These relative amplitudes are in the video signal sent to the transmitter. Then the amplitude-modulated picture carrier signal can have the standard relative amplitudes of 75 to 100 per cent for sync and lower amplitudes for picture information.

Switching and mixing. The switching system selects the desired *on-the-air* signal, viewed on the master monitor. Various input signals are provided, including all live cameras, film camera, and video tape recorder. The switcher unit in Fig. 5 · 12 has two rows of push buttons, a pair for each input signal. Punching in different signals along the same row instantly switches them. A continuous control is also provided to fade out smoothly to black. Or signal on one row can be mixed smoothly with signal on the opposite row in a *lap dissolve* before one alone is on the air. Although not shown here, *wipe effects* can be produced by inserting variable-width blanking pulses.

Synchronizing signal generator. The sync generator produces driving pulses for sync and blanking in all the camera chains at the studio, plus the kinescope sync and blanking pulses for the composite video signal transmitted to the receiver. There are two main sections: the pulse former that produces timing pulses at the correct frequency and the pulse shaper to provide the standard waveforms required for horizontal, vertical, and equalizing pulses. All the pulses are derived from a master oscillator, whose frequency is divided down. Starting with 31,500 cps for equalizing pulses, the oscillator output can be divided to $\frac{1}{2} \times 31{,}500$ for 15,750-cps horizontal pulses. The 60-cps vertical pulses can be divided in steps of $31{,}500 \times \frac{1}{7} \times \frac{1}{5} \times \frac{1}{5} \times \frac{1}{3}$, which equals 31,500 cps divided by 525. The

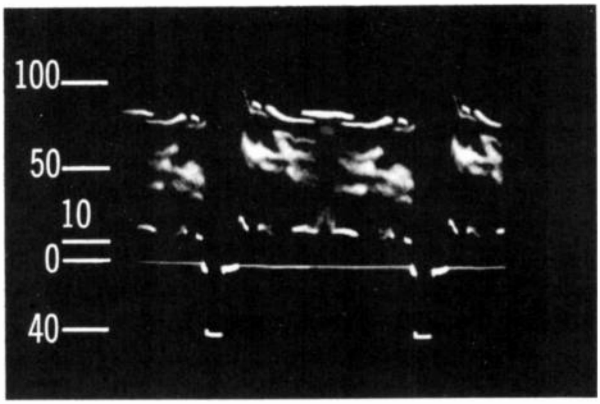


Fig. 5·13 Video signal on oscilloscope screen of studio monitor, with 10 per cent setup. Oscilloscope frequency at 7,875 cps to show two horizontal pulses. Sync polarity is down. Standard IRE scale of video amplitude at left. (American Telephone and Telegraph Co.)

master oscillator can be locked in phase with the 60-cycle power line or operate independently as a crystal-controlled oscillator. It should be noted, however, that for color television broadcasting the master oscillator is locked to the color subcarrier frequency of 3.579545 Mc.

Television relaying. To convey the signal for a television program from one location to another, relaying is used for studio-to-transmitter links, remote pickup-to-studio links, and intercity networks. When a program is broadcast over a network, each station in the network receives the program signal by means of intercity relay links. Then the station uses the relayed video signal as a program source to produce the standard AM picture signal broadcast in its assigned channel for the receivers in the area.

Television relaying is done by either cable or radio relays. Radio relays use microwave transmitters and receivers, operating in the range of 7,000 Mc. FM is generally used to transmit the picture signal by radio relay but this is converted to the standard AM picture signal broadcast by the station in its assigned channel.

Television transmitters. The peak r-f power output of a typical picture or sound signal VHF transmitter is 1 to 50 kw. However, the effective radiated power can be higher because it includes the gain of the transmitting antenna. The minimum effective radiated power specified by the FCC for a population of one million or more is 50 kw, with a transmitting-antenna height of 500 ft. For areas with a population under 50,000 the minimum effective radiated power is 1 kw with an antenna height of 300 ft.

The radiated power of the sound carrier signal is not less than 50 per cent or more than 150 per cent of the radiated power for the picture carrier signal, for monochrome transmission. In color transmission, the sound power is limited to 50 to 70 per cent of the picture power for minimum sound interference in the picture.

The frequency tolerance for the picture or sound carrier is ±0.002 per cent. However, the exact carrier frequencies for different stations on the same channel are offset from each other by plus 10 kc or minus 10 kc, in order to reduce interference between co-channel stations. This system is called *offset carrier operation.*

SUMMARY

1. The amplitude-modulated picture carrier signal has a symmetrical envelope, which is the composite video signal used to modulate the carrier wave. The two main features of the transmitted picture carrier signal are negative polarity of modulation and vestigial-side-band transmission.
2. Negative transmission means that the video modulating signal has the polarity required to reduce the carrier amplitude for white camera signal. Darker picture information raises the carrier amplitude. Tip of sync produces maximum carrier amplitude for the 100 per cent level.
3. Vestigial-side-band transmission means that all the upper side frequencies but only some of the lower side frequencies are transmitted in the 6-Mc channel for the modulated picture carrier signal. The band of upper side frequencies approximately 4 Mc above the carrier frequency is transmitted with full amplitude. Similarly, lower side frequencies separated by 0.75 Mc, or less, from the carrier frequency are also transmitted. However, the lower side frequencies below the picture carrier enough to be outside the assigned channel are not transmitted.
4. Channels 2 to 6 are low-band VHF channels between 54 and 88 Mc; channels 7 to 13 are high-band VHF channels from 174 to 216 Mc; channels 14 to 83 are UHF channels from 470 to 890 Mc. In all bands the station broadcasts in a standard channel 6 Mc wide.
5. The standard 6-Mc channel includes the AM picture carrier 1.25 Mc above the low end, and the FM sound carrier 0.25 Mc below the high end, with 4.5 Mc between the picture and sound carrier frequencies.
6. The equipment for broadcasting picture carrier signal includes studio facilities to produce composite video signal and transmitter equipment to generate the r-f carrier which is modulated by the video signal. At the studio, two or more camera chains are used for live-talent shows, plus camera chains for film and video tape. The synchronizing signal generator provides timing pulses for all camera chains, plus blanking and sync for the composite video signal transmitted to receivers.

SELF-EXAMINATION *(Answers at back of book.)*

Choose (*a*), (*b*), (*c*), or (*d*).

1. The modulated picture carrier wave includes the composite video signal as the: (*a*) average carrier level; (*b*) symmetrical envelope of amplitude variations; (*c*) lower side band without the upper side band; (*d*) upper envelope without the lower envelope.
2. Which of the following statements is true? (*a*) Negative transmission means the carrier amplitude decreases for black. (*b*) Negative transmission means the carrier amplitude decreases for white. (*c*) Vestigial-side-band transmission means both upper and lower side bands are transmitted for all modulating frequencies. (*d*) Vestigial-side-band transmission means the modulated picture carrier signal has only the upper envelope.
3. With 2-Mc video signal modulating the picture carrier for channel 4 (66 to 72 Mc), which of the following frequencies are transmitted? (*a*) 66-Mc carrier and 68-Mc upper side frequency; (*b*) 71.75-Mc carrier, with 69- and 73-Mc side frequencies; (*c*) 67.25-Mc carrier, with 65.25- and 69.25-Mc side frequencies; (*d*) 67.25 Mc carrier and 69.25 upper side frequency.
4. With 0.5-Mc video signal modulating the picture carrier; (*a*) both upper and lower side frequencies are transmitted; (*b*) only the upper side frequency is transmitted; (*c*) only the lower side frequency is transmitted; (*d*) no side frequencies are transmitted.
5. In all standard television broadcast channels the difference between picture and sound carrier frequencies is: (*a*) 0.25 Mc; (*b*) 1.25 Mc; (*c*) 4.5 Mc; (*d*) 6 Mc.

6. The difference between the sound carrier frequencies in two adjacent channels equals: (*a*) 0.25 Mc; (*b*) 1.25 Mc; (*c*) 4.5 Mc; (*d*) 6 Mc.
7. The sync for the receiver is produced in the: (*a*) deflection circuits in the receiver; (*b*) carrier generator at the transmitter; (*c*) camera control unit at the studio; (*d*) synchronizing signal generator at the studio.
8. Which of the following has the function of camera pickup? (*a*) synchronizing signal generator; (*b*) video tape recorder; (*c*) master monitor; (*d*) vestigial-side-band filter.
9. With 5 per cent black setup, maximum black in the picture corresponds to what per cent amplitude in the modulated picture carrier signal? (*a*) 5; (*b*) 70; (*c*) 75; (*d*) 95.
10. Line-of-sight transmission is a characteristic of propagation for the: (*a*) VHF and UHF bands of frequencies; (*b*) VHF band but not the UHF band; (*c*) low radio frequencies below 1 Mc; (*d*) AM picture signal but not the FM sound signal.

ESSAY QUESTIONS

1. Define negative transmission.
2. Give one advantage and one disadvantage of vestigial-side-band transmission.
3. For each of the following channels, list the picture carrier and sound carrier frequencies with their frequency separation: channels 2, 5, 7, 13, 14, 83.
4. Define the following terms: (*a*) black setup; (*b*) offset carrier operation; (*c*) camera chain; (*d*) booster station; (*e*) shadow area (*f*) intercity network; (*g*) carrier generator.
5. Which television channel numbers are in the 30- to 300-Mc VHF band? Which are in the 300- to 3000-Mc UHF band?
6. Draw a graph similar to Fig. 5·6 showing frequencies transmitted in channel 8, indicating picture and sound carriers with their frequency separation.
7. Why is reflection of the transmitted carrier wave a common problem with the picture signal in television, but not in the radio broadcast band of 535 to 1,605 kc? What is the effect in the reproduced picture of multipath signals caused by reflections?
8. Describe briefly three technical jobs in producing television programs.

PROBLEMS *(Answers to odd-numbered problems at back of book.)*

1. List the r-f side frequencies transmitted in addition to the picture carrier for the following modulation examples: (*a*) 0.25-Mc video modulation of channel 2 carrier; (*b*) 3-Mc video modulation of channel 5 carrier; (*c*) 0.5-Mc video modulation of channel 14 carrier; (*d*) 4-Mc video modulation of channel 14 carrier.
2. In negative transmission, what is the relative amplitude of the modulated picture carrier wave for: (*a*) maximum white; (*b*) tip of sync; (*c*) blanking level; (*d*) black in picture; (*e*) medium gray?
3. Give the exact picture carrier frequencies for: (*a*) channel 2 offset -10 kc; (*b*) channel 5 offset $+10$ kc; (*c*) channel 9 offset $+10$ kc; (*d*) channel 10 offset -10 kc.
4. Give the frequency separation for the following combinations: (*a*) picture and associated sound carriers; (*b*) picture carrier and lower adjacent-channel sound carrier; (*c*) picture carriers in two adjacent channels; (*d*) sound carriers in two adjacent channels.
5. Give the picture and sound carrier frequencies for channel 3.
6. List the picture and sound carrier frequencies for all channels from 2 to 14, inclusive.

Chapter 6 Television receivers

In effect, two receivers are included in the television chassis: an AM receiver for the transmitted picture carrier signal and an FM receiver for the associated sound signal. The sound section of the receiver provides audio signal for the loudspeaker, while the picture section provides video signal for the grid-cathode circuit of the picture tube. In addition, the television receiver has horizontal and vertical deflection circuits for scanning the raster on the kinescope screen. Finally, with video signal to reproduce the picture on the raster, synchronizing circuits are necessary to time the horizontal and vertical scanning. All these functions are provided by 15 to 20 tubes in the television receiver, many being dual-purpose tubes. Figure 6·1 shows a typical television receiver chassis.

6·1 Receiver circuits

Figure 6·2 illustrates the typical arrangement in television receivers. The superheterodyne circuit is used, as both the r-f picture signal and r-f sound signal beat with the local oscillator to produce the lower intermediate frequencies. The fact that the sound is an FM signal does not affect the heterodyning action.

Notice that both the sound and picture signals are amplified in the r-f section and the common i-f section. These circuits have enough bandwidth to amplify both signals, even though the sound and picture carrier frequencies are 4.5 Mc apart. Then the sound signal can beat with the picture carrier in the video detector stage, to produce an i-f signal centered at 4.5 Mc for the FM sound. This arrangement is called *intercarrier sound,* because the associated sound signal is obtained as the 4.5-Mc beat between the picture and sound carrier frequencies.

R-f section. Starting at the antenna, the picture and sound r-f carrier signals are intercepted by a common receiving antenna for both signals. A transmission line connects the antenna to the receiver input terminals, coupling the r-f picture and sound signals to the r-f amplifier stage. The amplified r-f output is then coupled into the mixer stage. Also coupled into the mixer is the output of the local oscillator to heterodyne with the incoming r-f picture and sound carrier signals. The oscillator usually beats above the r-f signal frequencies. When the oscillator frequency is set for the channel to be tuned in, the carrier signals of the selected station are heterodyned to the lower intermediate frequencies of the receiver.

The oscillator beating with the two r-f carrier signals produces two i-f carrier signals. One is the picture i-f signal corresponding to the r-f picture signal and the other is the sound i-f signal corresponding to the sound r-f signal. The original modulating information of the r-f carrier signals is present in the i-f signals out of the mixer, for both the AM picture signal and the FM sound signal. Furthermore, the 4.5-Mc separation between the r-f carrier frequencies is maintained in the i-f carrier frequencies. For most television receivers the intermediate frequencies in the output of the mixer stage are 45.75 Mc for the picture carrier and 41.25 Mc for the sound carrier, or 25.75 and 21.25 Mc in older receivers. In either case, note the 4.5-Mc difference between picture and sound carrier frequencies.

The r-f amplifier, mixer, and local oscillator stages are usually on an individual subchassis, which is called the *front end, head end, r-f unit,* or the *tuner.* Most often the local oscillator and mixer functions are combined in one tube, as indicated by the dotted lines around the *frequency-converter* stage in Fig. 6·2. With its station selector and fine tuning controls, the tuner selects the channel to be received by converting its picture and sound r-f carrier frequencies to the intermediate frequencies of the receiver, so that the selected signals can be amplified in the i-f stages.

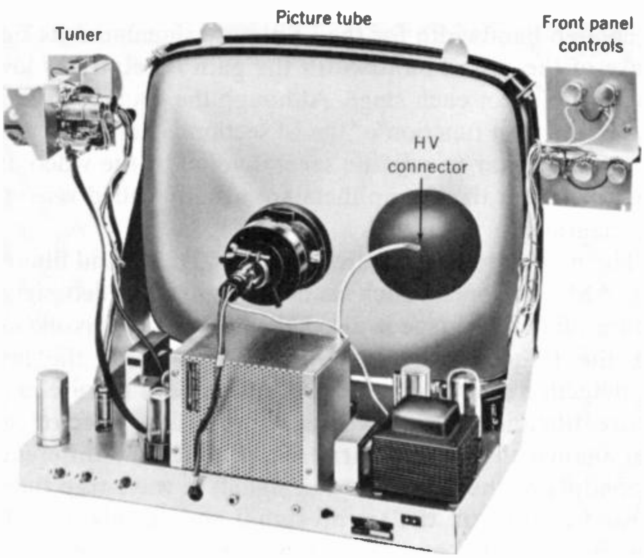


Fig. 6·1 Television receiver chassis. (Andrea Radio Corp.)

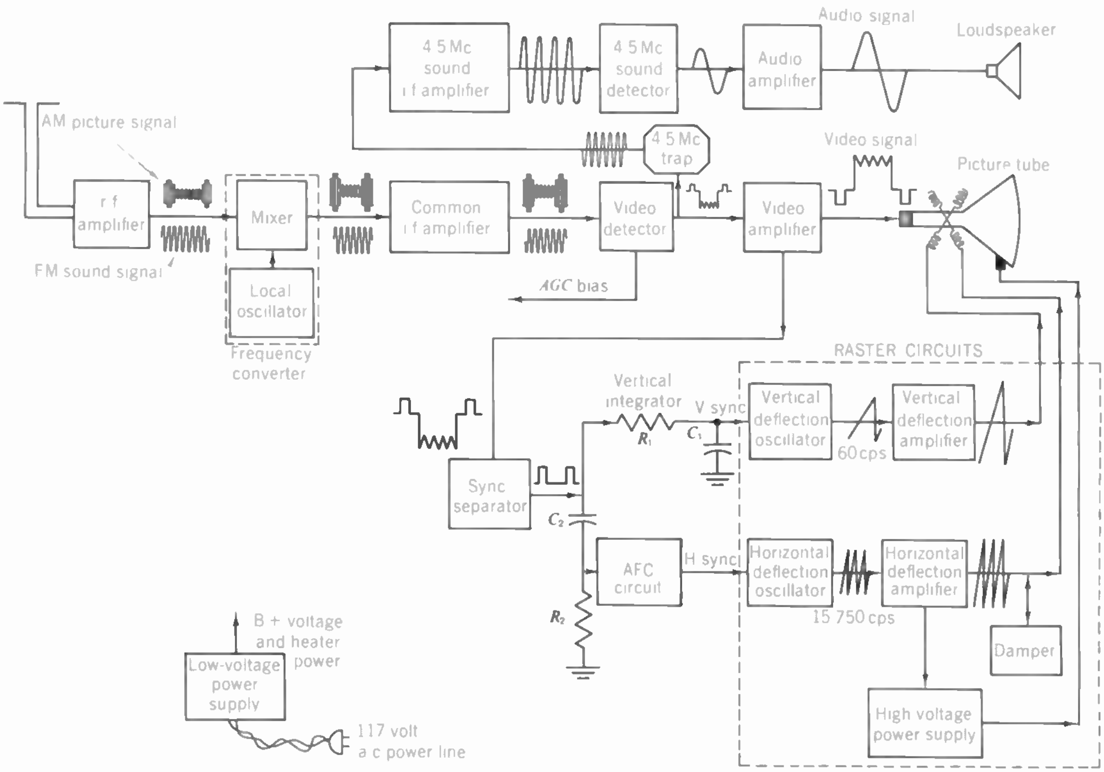


Fig. 6·2 Block diagram of a typical television receiver. Amplitudes of signal waveshapes not shown to scale.

Picture I-f signal. The common i-f amplifier usually includes three tuned stages, with enough bandwidth for the i-f picture signal and its side frequencies. Because of the broad bandwidth the gain is relatively low, typical values being 20 to 30 for each stage. Although the sound i-f signal is also amplified here, the main function of the i-f section is amplifying the picture i-f signal from the mixer to provide several volts for the video detector. It should be noted that these amplifiers are usually called *video i-f stages* in schematic diagrams.

Video detector. The modulated i-f picture signal is rectified and filtered here to recover its AM envelope, which is the composite video signal needed for the picture tube. This stage is also called the *picture second detector,* considering the frequency converter in the tuner as the first detector. Actually, detection and frequency conversion both require non-linear operation or rectification. For this reason, the video detector can beat the FM signal against the picture carrier to produce the difference frequencies corresponding to the 4.5-Mc sound signal. A wave trap tuned to 4.5 Mc is then used to take off the sound signal and couple it to the sound section of the receiver. However, the main purpose of the video detector is to provide for the video amplifier the composite video with its camera signal, sync, and blanking information needed for reproduction of the picture.

Video amplifier. Consisting of one or two stages, this section amplifies the composite video signal enough to drive the grid-cathode circuit of the picture tube. The camera signal variations change the instantaneous grid-cathode voltage, modulating the intensity of beam current. Then the variations of light intensity as the spot scans the screen enable the picture to be reproduced on the raster.

The blanking pulses in the composite video signal drive the kinescope grid-cathode voltage negative to cutoff, blanking out retraces. Although the sync is included in the video signal to the kinescope, the only effect of sync voltage here is to drive the grid more negative than cutoff. Note that the composite video signal is also coupled to the sync circuits, where the synchronizing pulses are separated for use in timing the receiver scanning.

The amount of composite video signal required is about 80 volts peak to peak to drive the kinescope grid for strong contrast. With 4 volts output from the video detector, one video amplifier stage with a gain of 20 can supply the required signal for the picture tube. When the composite video signal is coupled to the kinescope control grid, this method is called *grid drive;* coupling the signal to the cathode is *cathode drive.*

Automatic gain control. In Fig. 6·2 the video detector is shown providing a bias voltage for automatic control of the gain of the preceding i-f and r-f stages. This automatic-gain-control (AGC) circuit is similar to the automatic-volume-control (AVC) system in conventional sound receivers. The stronger the picture carrier signal is, the greater the negative AGC bias voltage produced and the less the gain of the receiver. The result is relatively constant video signal amplitude for different carrier-signal strengths. Therefore, the automatic gain control in the picture amplifier chain is useful as an automatic control of contrast in the reproduced picture. AVC is not generally used in the FM sound i-f amplifier. However, the AGC circuit affects both the picture and sound, since it controls the gain of the r-f and common i-f stages, which amplify the picture and sound signals.

Synchronizing circuits. The video detector output includes the synchronizing pulses as part of the composite video signal. Therefore, the composite video is also coupled to the sync circuits to provide the timing pulses needed for controlling the frequency of the vertical and horizontal deflection oscillators in the receiver.

The synchronizing circuits include one or more amplifier and separator stages. A sync separator is a clipper stage that can separate the synchronizing pulse amplitude from the camera signal in the composite video. Remember that the top 25 per cent of the video signal amplitude is used for sync. When the tips of the video signal are clipped off and amplified, the output consists only of sync pulses.

Since there are synchronizing pulses for both horizontal and vertical scanning, Fig. 6·2 shows the output of the sync separator divided into two parts. The integrator is a low-pass *RC* filter circuit that filters out all but the vertical pulses from the total separated sync voltage. Then the vertical synchronizing signals can lock in the vertical deflection oscillator at 60 cps. For horizontal synchronization, an automatic-frequency-control (AFC) circuit is used to lock in the horizontal deflection oscillator at 15,750 cps.

Overall receiver gain. The picture amplifier circuits produce a total voltage amplification of one million, approximately, from the antenna input terminals to the kinescope grid-cathode circuit. This is the product of the voltage amplification for individual stages. Although the r-f and i-f gain depends on AGC bias, typical values are about 10 in the r-f section, 10,000 for picture i-f signal, ½ for the diode video detector, and 20 in the video amplifier. The product then is

$$10 \times 10{,}000 \times \tfrac{1}{2} \times 20 = 1{,}000{,}000$$

For 80 volts of video signal at the kinescope, therefore, 80 μv signal is needed at the antenna terminals. However, more antenna signal may be necessary for a picture with good sync and no snow.

Deflection circuits. As shown in Fig. 6·2, the deflection circuits include the vertical oscillator and amplifier for vertical scanning at 60 cps, plus the horizontal oscillator and amplifier for scanning at 15,750 cps. For either vertical or horizontal scanning, the oscillator stage generates deflection voltage to drive the amplifier at the required frequency. The deflection amplifiers are power output stages to provide enough scanning current in the deflection yoke for a full-sized raster. The horizontal output circuit also includes the damper stage, which has the function of reducing sine-wave oscillations in the horizontal sawtooth scanning current. In addition, the damper provides boosted B+ voltage for the horizontal amplifier. Therefore, the horizontal output circuit cannot operate without the damper. The horizontal output is also coupled to the high-voltage rectifier to supply anode voltage for the picture tube.

The horizontal and vertical deflection circuits produce the illuminated scanning pattern forming the raster on the kinescope screen. The spot intensity can then be varied by the video signal voltage coupled to the kinescope grid-cathode circuit, to reproduce the picture on the raster.

The deflection circuits produce the required scanning current and the resultant raster with or without the synchronizing signals. Since the deflection oscillators are free-running, they do not require any external signal for operation. However, the sync is needed to hold the deflection oscillators at exactly the right frequency so that the picture information is reproduced in the correct position on the raster. Without sync, the deflection circuits can scan the raster but the picture will not hold still.

Power supplies. Two power supplies are needed in the television receiver. One is the usual B+ supply with about 300 volts output to supply d-c operating potentials for all amplifier stages. This is called the *low-*

voltage supply in the television receiver because its output voltage is low compared with the *high-voltage supply*, which provides anode voltage for the picture tube. For sufficient brightness, the anode voltage for direct-view picture tubes is generally 9 to 18 kv, while projection kinescopes require 20 to 80 kv.

The high-voltage rectifier obtains its a-c input from the horizontal amplifier. This arrangement is called a *flyback supply* because the high voltage is generated as an induced voltage during the fast horizontal retrace. The resultant voltage is stepped up by the horizontal output transformer for the required amount of high voltage. The rectified output is the d-c anode voltage needed by the kinescope to produce brightness on the phosphor screen. Because the kinescope anode voltage depends on the horizontal output, there cannot be any brightness on the kinescope screen if the horizontal scanning circuits are not operating.

6 • 2 Sound take-off circuit

Since the bandwidth of the sound signal is narrow compared with the picture signal, a resonant circuit can be used as a wave trap to filter out the sound signal and couple it to a separate sound i-f amplifier. The sound take-off block in Fig. 6 • 2 is an *LC* tuned circuit resonant at 4.5 Mc for intercarrier sound receivers. Connected to the video detector output circuit, where the sound signal is converted to 4.5 Mc, the sound take-off circuit rejects 4.5 Mc from the video circuits and couples the sound signal to the sound i-f amplifier. In some circuits the 4.5-Mc sound take-off is in the output of the video amplifier. Then its frequency response must extend up to 4.5 Mc to amplify the sound signal in addition to the video signal.

Regardless of where the sound take-off tuned circuit is in the receiver, this point marks the separation of sound and picture signals. All stages before the sound take-off circuit amplify both the sound and picture signals. After the sound take-off the signal amplifiers can be considered in two groups, one for sound and the other for picture. In Fig. 6 • 2, only the video amplifier is for picture alone. The 4.5-Mc i-f amplifier and FM detector with the audio amplifier are for sound alone.

Intercarrier sound receivers. Note the sequence of stages for sound in Fig. 6 • 2. After being amplified in the r-f tuner with the picture signal, the sound i-f signal passes through all the i-f amplifiers for picture signal so that both i-f signals are coupled into the video detector. This is accomplished by having a little more bandwidth in the common i-f amplifier to provide some gain for the sound i-f signal. However, there is much more gain for the picture signal. As a result, the video detector can produce the 4.5-Mc sound signal as the difference frequencies between the strong i-f picture carrier beating with the weaker i-f sound signal. This action corresponds to the operation of the converter in the tuner, where the relatively strong output of the local oscillator stage beats with the weak r-f signal input.

We can consider the picture carrier as a second local oscillator to heterodyne with the i-f sound signal. The receiver then is a double super-

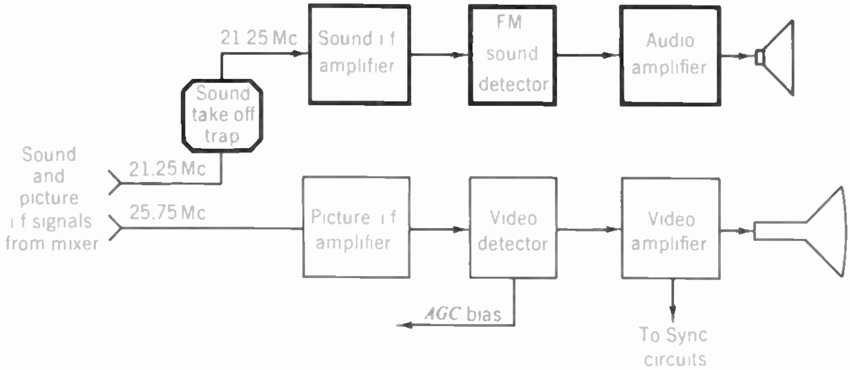


Fig. 6·3 Signal circuit in a split-sound receiver, showing sound take-off in mixer output.

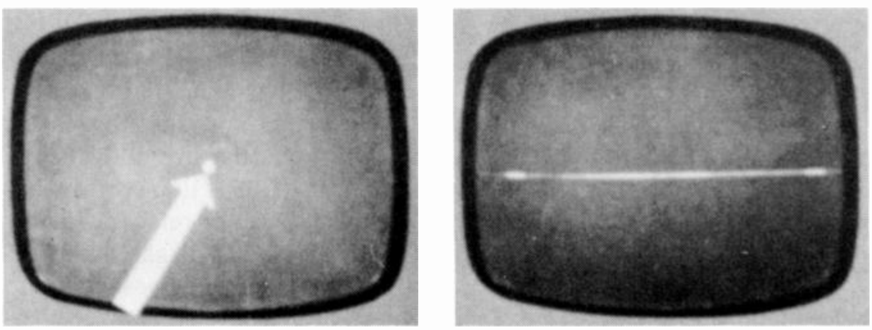
heterodyne with two frequency conversions for the associated sound. First, the r-f sound carrier of any selected station is converted by the tuner to 41.25 Mc in the mixer output, assuming this value as the sound i-f carrier frequency for the receiver. Then in the video detector the picture carrier at 45.75 Mc heterodynes with the 41.25-Mc sound i-f signal to beat it down to 4.5 Mc. This second sound i-f value is always exactly 4.5 Mc because this is the exact difference between picture and sound carrier frequencies, as determined by the transmitter, for all stations.

The sound take-off circuit filters out the 4.5-Mc sound signal from the video frequencies and couples it to the sound i-f amplifier. It is important to note that the take-off circuit, the sound i-f amplifier, and the FM detector are always tuned to 4.5 Mc. This 4.5-Mc combination is the mark of all intercarrier sound receivers.

The sound i-f amplifier has the function of amplifying the 4.5-Mc signal enough to drive the detector. The FM detector is necessary because the 4.5-Mc sound signal is not an audio signal. It is just the FM sound i-f signal at the lower center frequency of 4.5 Mc. The FM signal must still be detected to recover the desired audio signal. After detection, the audio signal is amplified in a conventional audio amplifier to operate the loudspeaker.

Split-sound receivers. In older receivers, the sound signal is separated from the picture signal in the i-f circuits, before the video detector. In this

(*a*) (*b*)



case, the sound take-off is usually in the mixer output circuit, as illustrated in Fig. 6 • 3. This arrangement is considered a split-sound receiver because there is a separate i-f amplifier for picture signal alone and a separate sound i-f amplifier. Figure 6 • 3 does not show sync and deflection circuits or power supplies because these circuits can be the same for either intercarrier sound or split sound.

In a split-sound receiver, the local oscillator in the r-f tuner must be tuned exactly to make the sound i-f signal from the mixer be the same as the frequency to which the sound i-f circuits are tuned. In Fig. 6 • 3, this sound i-f value required from the mixer plate circuit is 21.25 Mc. If you detune the oscillator slightly, the sound signal changes from its 21.25-Mc center frequency. Then the sound may be weak or distorted. Furthermore, the oscillator can easily be detuned enough to cause no sound at all, when the i-f signal is too far off the frequency to which the sound i-f circuits are aligned. However, this problem of tuning in the sound is practically eliminated in intercarrier-sound receivers. Even when the oscillator tuning changes the sound i-f signal out of the mixer, the picture i-f carrier changes by the same amount and the difference frequency is still 4.5 Mc. As a result, the 4.5-Mc sound circuits are always tuned to the same frequency as the intercarrier-sound signal. This ability to provide good sound automatically when the picture is tuned in is the main reason why the intercarrier-sound arrangement is used now in all receivers.

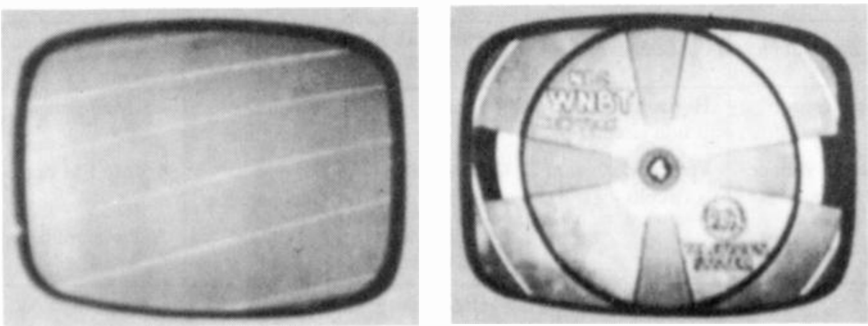
6 • 3 Functions of the receiver circuits

The television receiver can be considered in sections, each with a specific function, as follows:

Illumination. The picture tube with its auxiliary components and the high-voltage supply, which provides the kinescope anode voltage, have the function of producing brightness on the kinescope screen. Just the spot of light on the screen illustrated in Fig. 6 • 4*a* shows that the kinescope and high-voltage supply are operating. It should be noted that the illuminated

Fig. 6 • 4 Illustrating how the receiver puts the picture on the raster. (a) Illumination on kinescope screen. (b) Illumination plus horizontal scanning. (c) Horizontal and vertical scanning to produce the illuminated raster. (d) The picture on the raster.

(*c*) (*d*)



spot for this illustration was produced with an external high-voltage supply, as the flyback supply in the receiver cannot produce high voltage without horizontal output.

Horizontal scanning. The horizontal deflection circuits produce horizontal scanning lines. The single horizontal line on the kinescope screen in Fig. 6·4*b* shows illumination and horizontal scanning.

Vertical scanning. The vertical deflection circuits produce vertical scanning to spread the horizontal scanning lines over the entire screen area, to form the scanning raster. The illuminated raster on the kinescope screen in Fig. 6·4*c* shows that the vertical and horizontal deflection circuits, kinescope, and high-voltage supply are operating.

Picture. Figure 6·4*d* shows a picture reproduced on the raster. The circuits for picture signal from the antenna input to the kinescope grid provide video signal corresponding to the desired picture information. The video signal-voltage variations on the kinescope grid vary the intensity of the electron beam, while the deflection circuits produce scanning, to reproduce the picture as shades of white, gray, and black on the raster.

Synchronization. The sync circuits in the receiver hold the line structure of the picture together and make it stay still by timing the horizontal and vertical scanning correctly with respect to the reproduced picture information. Horizontal synchronization prevents the line structure of the picture from tearing apart into diagonal segments. Vertical synchronization allows successive frames to be superimposed over each other so that the picture will not roll up or down on the screen.

Sound. The signal circuits for the sound provide audio signal for the loudspeaker to reproduce the sound.

Table 6·1 illustrates how the receiver can be divided into circuits that produce the illuminated raster and circuits for the picture and sound signals. The signal circuits are subdivided between picture and sound for the typical case of an intercarrier-sound receiver that has the 4.5-Mc sound take-off circuit in the video detector output. Note that the low-voltage

Table 6·1 Receiver circuit functions

Circuits for the raster		*Circuits for the signal*		
Illumination	*Scanning*	*Picture and sound*	*Picture only*	*Sound only*
Kinescope	Horizontal deflection circuits	R-f tuner	Video amplifier	4.5-Mc i-f stages
High-voltage supply	Vertical deflection circuits	Common i-f stages	Sync circuits	4.5-Mc FM detector
		Second detector AGC circuit		Audio amplifier
Low-voltage power supply				

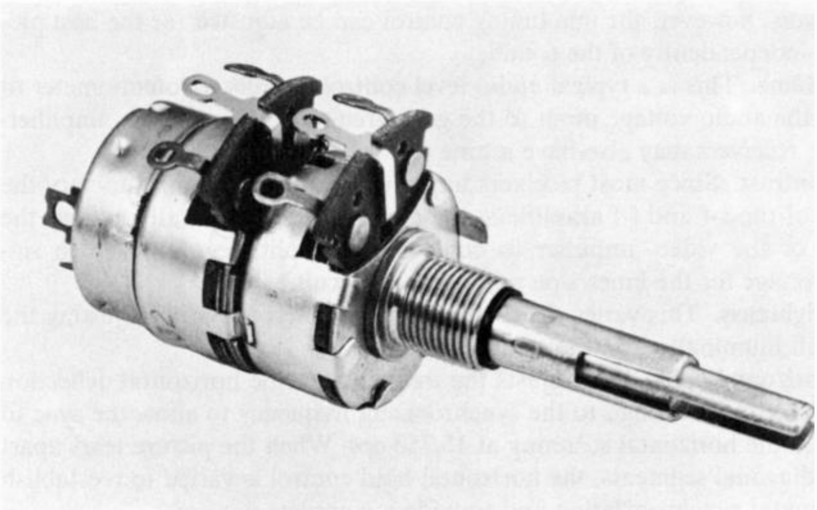


Fig. 6 · 5 Dual potentiometer with concentric shafts and on-off switch at back. (Clarostat Mfg. Co.)

power supply is common to the raster and signal circuits, since all the amplifiers need B+ voltage for operation.

The sync section is part of the circuits for picture only. Without sync, the receiver can still produce the raster. Signal is needed to provide synchronization, however, so that picture information is reproduced in its correct position on the raster. Otherwise, the picture will roll vertically or tear apart horizontally.

6 · 4 Receiver operating controls

The controls for adjusting the operation of the receiver circuits can be considered in two groups: the setup adjustments for the raster circuits, and the operating controls in the signal circuits. The setup adjustments for scanning are generally mounted on the rear apron of the chassis. These are installation or servicing adjustments to provide suitable height, width, and linearity of the scanning. The operating controls are on the front panel of the receiver, where they can be varied for different stations. Dual controls, as shown in Fig. 6 · 5, are generally used to save space. Also, the on-off switch on the volume control is often the push-pull type so that the volume setting does not change when the set is turned off. The controls have the following functions:

Station selector. This adjusts the resonant frequency of the r-f circuits in the front end to the desired channel frequencies and sets the local oscillator at the frequency necessary to tune in the station.

Fine tuning. This provides more exact setting of the local oscillator frequency. In split-sound receivers, the fine tuning control is adjusted for the best sound, which normally results in the best picture also. With intercarrier receivers, however, the fine tuning control can be adjusted for the best picture, independently of the sound.

Volume. This is a typical audio level control, usually a potentiometer to vary the audio voltage input to the grid circuit of the first audio amplifier. Some receivers may also have a tone control.

Contrast. Since most receivers have automatic gain control to vary the gain of the r-f and i-f amplifiers, the contrast control usually adjusts the gain of the video amplifier to control the amplitude of the video signal voltage for the kinescope grid-cathode circuit.

Brightness. This varies the d-c bias for the kinescope grid, adjusting the overall illumination on the screen.

Horizontal hold. This adjusts the frequency of the horizontal deflection oscillator close enough to the synchronizing frequency to allow the sync to lock in the horizontal scanning at 15,750 cps. When the picture tears apart into diagonal segments, the horizontal hold control is varied to reestablish horizontal synchronization and provide a complete picture.

Vertical hold. This adjusts the frequency of the vertical deflection oscillator close enough to the vertical synchronizing frequency to allow the sync to lock in the vertical scanning at 60 cps. When the picture rolls up or down, the vertical hold control is varied to reestablish vertical synchronization and make the picture stay still.

6 · 5 Projection television

The practical limit for a direct-view picture tube is a screen diameter of about 30 in. For larger pictures, it is necessary to use an optical projection system to enlarge the image. In one method, a projection lens throws light from the kinescope screen image onto a larger viewing screen. This optical arrangement is essentially the same as a film projector for slides or motion pictures. More common for television projection is the method shown in Fig. 6 · 6, which uses a special mirror to enlarge the optical image, instead of a projection lens. This reflective system is an adaptation of the Schmidt astronomical camera and therefore is called a *Schmidt-type* optical projector.

The main problem in television projection is producing enough brightness, for two reasons. First, the brightness of the enlarged picture is much less than the original image on the kinescope screen because this light is distributed over a much larger area on the viewing screen. The light intensity in the image is reduced in inverse proportion to its increase in area. Second, the optical projection system is inefficient. Only part of the light from the kinescope screen is collected by the optical projection system and delivered to the viewing screen. In the Schmidt system about 20 to 30 per cent of the available light is transmitted to the screen, while the projection-lens method has an efficiency of only 5 to 10 per cent.

Theater television. Theaters can provide television programs by means of large-screen projection, presenting news and sports events as they occur. The Schmidt-type reflective optical system is used for the greatest light efficiency. With 80 kv on the anode of the 7WP4, this projection kinescope provides enough brightness for a picture 15 by 20 ft, at a projection throw distance of about 80 ft.

The picture and sound signals from the source of the program can be transmitted to the theaters by special telephone lines or by microwave radio-relay links. In either case, the transmission for theater television is outside the commercial television broadcast channels.

6·6 *Pay television*

Several systems of coded or private television transmission have been proposed, with the purpose of charging a fee for special program material. In effect, a box office is provided for watching first-run motion pictures, popular sports events, and similar programs that would normally not be available to broadcasters and advertisers because of excessive costs. Conventional television receivers can reproduce the program but only after the signal has been decoded. The transmitted signal can be modified so that the reproduced picture is blurred without sync or reversed in black-and-white values. However, when the viewer pays the fee for the decoding information, a normal picture is reproduced. The decoder, which is attached to the receiver, may be operated by a coin box, a key, a punched card with coded holes, or a telephone connection.

In one method of scrambling black and white are reversed, while the picture is cut into horizontal strips and alternate strips shifted in synchronization. See Fig. 6·7. Furthermore, the strips are changed from frame to frame, which makes the picture information roll vertically. The associated

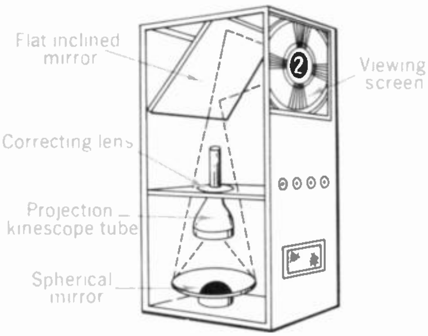
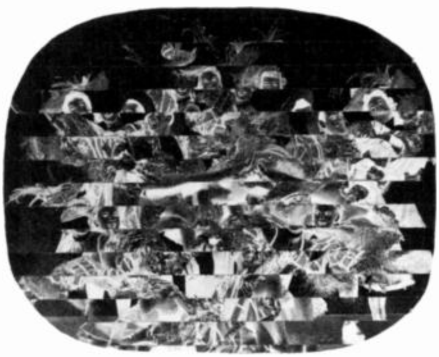


Fig. 6·6 Folded Schmidt-type projection system.

Fig. 6·7 Picture reproduction in Phonevision system of pay television. (a) Blurred picture without coding signal at receiver. (b) Clear picture with coding signal. (Zenith Radio Corp.)

(*a*)



(*b*)

Table 6·2 Television receiver tubes

Circuit	*Tube type*	*Notes*
R-f amplifier in tuner	6BQ7-A, 6BK7, 12BZ7	Dual triodes for low-noise cascode circuit
	2CY5, 6CY5, 6ER5	Tetrodes
	6AG5, 6CB6	Pentodes
Oscillator-mixer in tuner	6CG8, 6CQ8	Triode-tetrode
	6J6	Dual triode
	6AF4, 6DZ7	Triode UHF oscillator
I-f amplifiers, picture or sound	3BZ6, 6BZ6, 6DK6, 6AU6, 6CB6, 6EW6	Pentodes
	6AN8, 6K8, 6U8, 6AU8	Pentode-triode
Video detector	6AL5 (one section)	7-pin miniature dual diode
	1N64, 1N60, 1N87, 1N295	Germanium crystal diodes
Video amplifier	6AW8, 6EB8	Pentode-triode
	6CL6, 6AC7, 6AU6, 12BY7	Pentodes
FM sound detector	6AL5 dual diode or crystal diodes	Ratio detector or discriminator circuit
	6T8, 6FM8, 6BN8	Dual diode and triode
	3DT6, 6DT6, 3BN6, 6BN6, 6HZ6	Pentagrid tube with two control grids for gated-detector circuit
Audio amplifiers	6AV6, 6AT6, 6T8, 6FM8	Triode section for first audio
	6AQ5, 6BQ5, 6CU5, 6V6	Audio power output stage
Sync circuits	6CG7, 6SN7, 12AU7	Dual triodes
	6CS6, 6BE6, 6BY6	Pentagrid types
	3BU8, 6BU8, 6HS8	Dual pentodes for sync and AGC
	2EN5 twin diode or crystal diodes	Sync discriminator for horizontal AFC
Deflection oscillators, horizontal or vertical	6CG7, 6SN7, 6FQ7	Dual triodes for multivibrator or sync separator
	6DN7, 6EM7, 6DE7, 6CM7, 6DR7, 6GF7	Dual triodes for vertical oscillator and amplifier
Horizontal deflection amplifier, power output tubes	6BG6, 6CD6, 6JB6, 6BQ6, 6DQ6, 6JE6	Top cap is plate
	6GJ5, 6GT5	No top cap
Horizontal damper	5V4, 6AX4, 6AU4, 6BH3, 6BS3, 6W4, 6DE4, 6DA4, 6DW4, 12AX4	Diode power rectifiers
High-voltage rectifier	1B3, 1G3, 1K3, 1J3, 2B3, 3A2, 3A3	Diode rectifiers; top cap is plate
	1V2, 1X2	Miniature type
Low-voltage rectifier	3DG4, 5CB3, 5U4, 5V3, 5AU4, 5Y3	Full-wave rectifier tubes
	Silicon diodes or selenium diodes	Need series *R* to limit peak current

sound signal is also scrambled to make it sound like a high-pitched monkey chatter. When the private keying signal is received, however, a decoder unit installed at the receiver reverses the encoding process. Then the decoded signals provide picture and sound.

The private program may be sent to the receiver either by broadcasting in standard transmission channels or by distributing video signal to subscribers in a closed-circuit system. For the case of broadcasting coded signals, however, FCC approval is needed. Several systems are now being tested on an experimental basis, as an auxiliary service to regular television broadcasting sponsored by advertisers.

6·7 Receiver tubes

Table 6·2 lists typical tube types often used in television receivers. Familiarity with these often means you can know the function of the stage by the tube used. Amplifiers for the picture and sound signals are generally the miniature glass types shown in Fig. 6·8. The heater pins are usually 3 and 4 for the 7-pin base in *a*, or pins 4 and 5 for the 9-pin base in *b*. Note that the wider spacing between end pins serves as a guide for inserting the tube. The 6-volt types such as 6BZ6 are for parallel heater

(*a*) (*b*)

Fig. 6·8 Miniature glass tubes. (a) 7-pin base. IS is internal shield. (b) 9-pin base. (RCA.)

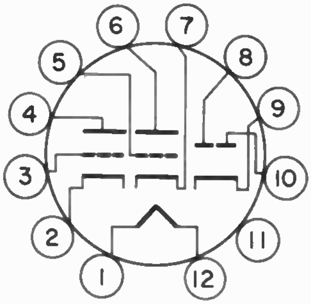


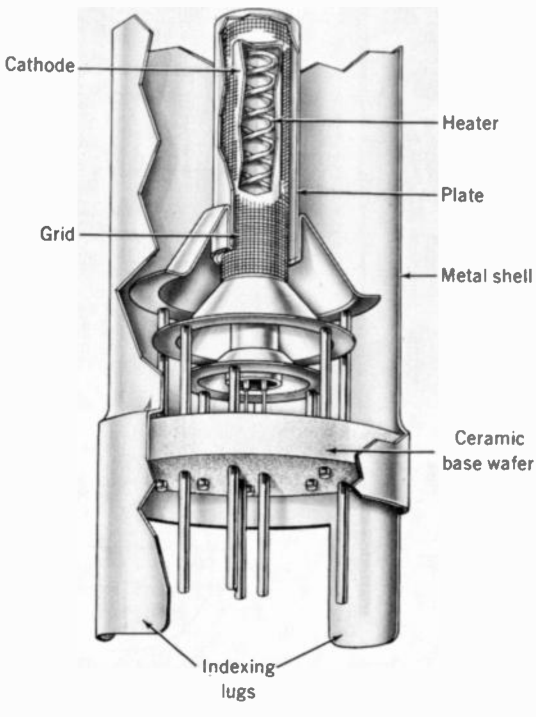
Fig. 6·9 Compactron multifunction tube 6B10, combining two diodes and two triodes. Height 1½ in. (GE.)

circuits in receivers using a power transformer. Heater ratings such as 3BZ6, 5CG8, 10DE6, and 17DQ6 are for series heater circuits in transformerless (a-c/d-c) receivers. Many 12-volt types, such as 12BZ7, have a center-tapped heater for either 6-volt operation in a parallel circuit or 12-volt operation in a series circuit.

Recently developed tube types are the *compactron* in Fig. 6·9 and the *nuvistor* in Fig. 6·10. The compactron saves space by combining several tubes in one relatively small envelope with a 12-pin base. The nuvistor is a miniaturized tube, not much larger than a transistor. Note the cylindrical electrode in Fig. 6·10*b*, which improves high-frequency operation and provides a rugged construction.

Fig. 6·10 Nuvistor ultraminiature tube. (a) 6CWA triode r-f amplifier. Height is ½ in. (b) Internal construction. (RCA.)

(*a*)



(*b*)

Fig. 6 · 11 *Horizontal power output tube with octal base. Top cap is the plate.* (RCA.)

Fig. 6 · 12 *Half-wave power rectifier tube 6BH3 for damper diode in horizontal output circuit. This is a novar tube with dark heater.* (RCA.)

Fig. 6 · 13 *High-voltage rectifier tube 1K3, with octal base. Top cap is the plate.* (RCA.)

Power tubes are needed for the deflection circuits, especially in the horizontal output stage, which requires the most power from the B+ supply. The 6DQ6 in Fig. 6 · 11 has maximum ratings of 15 watts plate dissipation, 140 ma average cathode current, with a peak of 440 ma for full width in the raster. Note the plate connection to the top cap, instead of the base, where the high plate voltage could arc to other pins or to chassis. Similarly, the damper stage in the horizontal output circuit is a diode power rectifier, as it rectifies the deflection voltage to produce boosted B+.

The 6BH3 damper diode in Fig. 6 · 12 is one of a new line of *novar* tubes. These are power tubes in a 9-pin miniature glass envelope. An important feature is the dark heater, which operates 350° below the temperature of conventional heaters that are red-hot. The advantages are less heat and longer tube life.

The high-voltage diode rectifier in Fig. 6 · 13 features a peak inverse voltage rating of 26,000 volts, in a flyback supply. The heater power at 1.25 volts and 0.2 amp is obtained from the horizontal output circuit, as is the high voltage of 15 to 20 kv. Little heater power is necessary because average direct load current is less than 1 ma for beam current in the picture tube.

Tube-location guide. See Fig. 6 · 14. This chart is usually pasted on the inside of the receiver cabinet or on the back cover. Not all receivers have the same tube layout but certain patterns follow in logical order. The

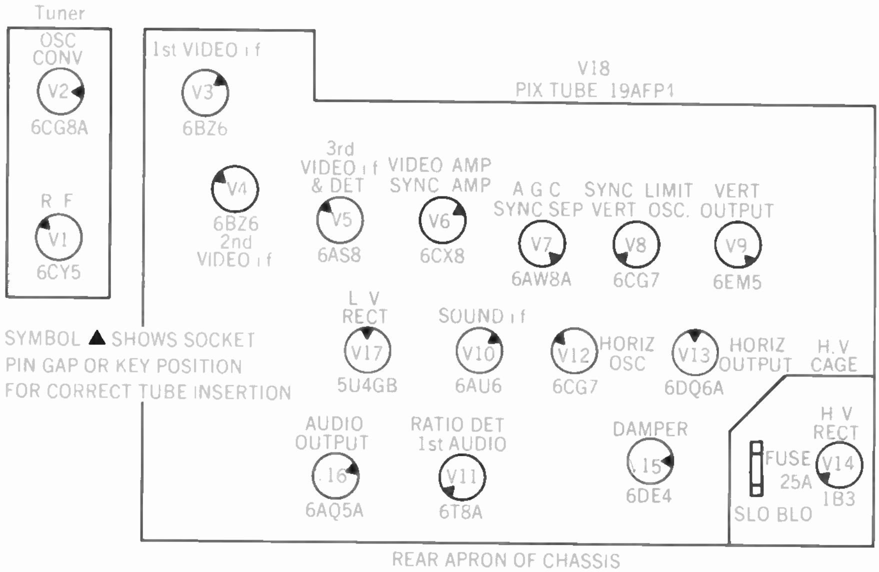


Fig. 6·14 Tube-layout chart for typical television receiver chassis. (Andrea Radio Corp.)

tuner tubes are on a separate subchassis that has the station selector. The i-f tubes are in sequential order. The vertical deflection stages are together. The horizontal deflection stages are also together, usually near the high-voltage cage because of the flyback supply. A fuse in or near the cage usually protects the horizontal output circuit alone.

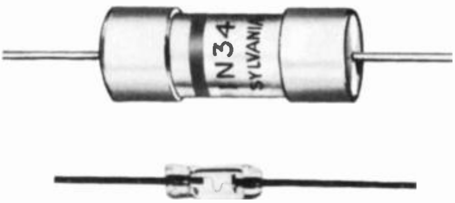
6·8 Semiconductors

This category includes diode rectifiers and triode transistor[1] amplifiers. Two advantages are smaller size and no heater power for the semiconductor. Common applications are the video detector, where a germanium diode is generally used, and silicon diodes for the low-voltage power supply (see Fig. 6·15*a*). Also, two semiconductor diodes are often used as the phase detector for the horizontal AFC circuit.

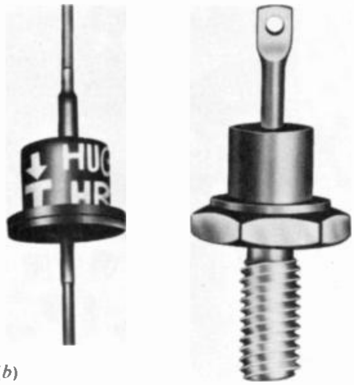
The back resistance in the inverse direction for semiconductors is much lower than a vacuum-tube diode. Typical values for a germanium diode are 50 ohms forward resistance and 0.5 megohm back resistance, approximately. When checking its resistance, the main requirement is that the forward and back resistances be very much different when you reverse the ohmmeter leads.

For power rectifiers in the B+ supply, silicon diodes are generally used (see Fig. 6·15*b*). These have ratings of 350 to 500 ma or more for average direct load current. Their extremely low forward resistance means less *IR*

[1] For references on transistor theory, see Bibliography at back of book.



(a)



(b)

Fig. 6·15 Semiconductor diodes. (a) Germanium detectors 1N34 and 1N277. Length about ½ in. (b) Silicon diodes for B+ supply. Note symbol on left diode. Arrow opposite to direction of electron flow. Height about ½ in.

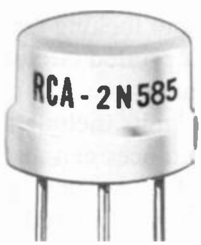
drop across the rectifier, allowing about 20 volts more B+ output for the same a-c input, compared with the 5U4.

In the type numbers for semiconductors, the first digit is one less than the number of electrodes. Diodes begin with 1, therefore, as in 1N277. Triode transistor types begin with 2, as in 2N285 (see Fig. 6·16). The N indicates a semiconductor.

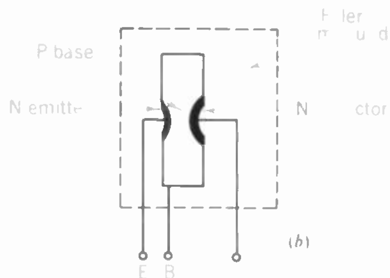
6·9 Printed circuits

Whether tubes or transistors are used, most television receivers have printed wiring. Generally, the chassis includes several printed-circuit boards for individual sections, such as i-f amplifier, sync and deflection circuits, and audio amplifier. A typical board is shown in Fig. 6·17. The wiring consists of paths of copper foil bonded to the plastic board.

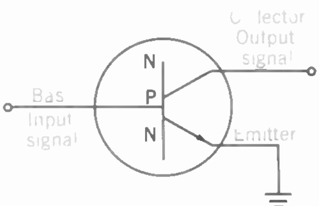
Fig. 6·16 Triode NPN transistor. Polarities are opposite for PNP type. (b) Physical construction. (c) Schematic symbol. Arrow is opposite to direction of electron flow. (d) Comparison to triode tube.



(a)



(b)



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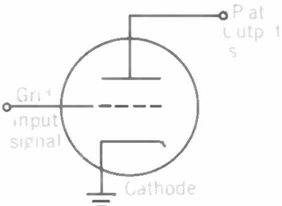

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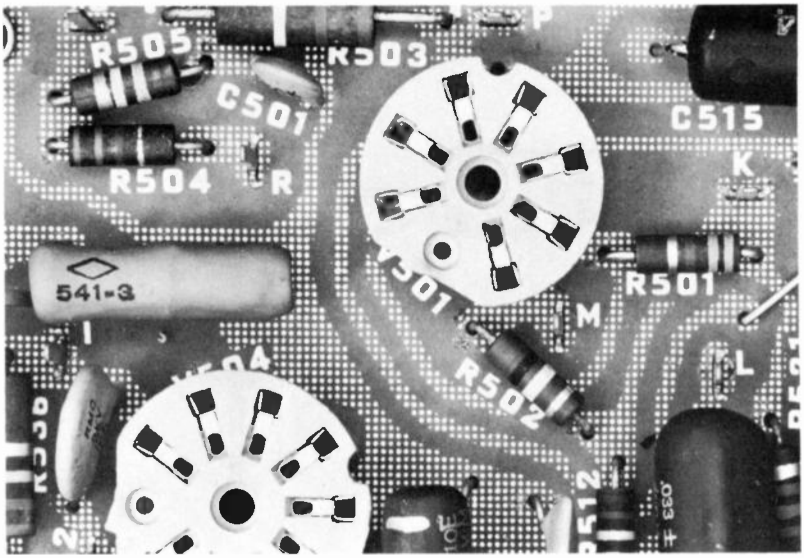


Fig. 6 · 17 Printed-circuit board for i-f amplifier. Shaded areas indicate printed wiring on opposite side of board. (RCA.)

When working on a printed board, the following techniques may be helpful:

1. Use a small iron (25 to 50 watts), to prevent excessive heat from lifting the printed conductors off the board.
2. Conventional resistors and capacitors on the board can usually be replaced without disturbing the printed wiring. Just break the old component in the middle, to save its pigtail connecting leads, and then solder the new component to the old leads.
3. A short break in the printed wiring can be repaired by soldering a piece of bare wire over the open. Apply only enough heat to melt the solder.
4. If a larger section of printed wiring is damaged, substitute the required length of hookup wire, soldered at two convenient end terminals.
5. You can see the printed wiring more clearly by placing a bright light at one side of the board and looking through the opposite side.

Another application of printed circuits is for complete units, as shown in Fig. 6 · 18. This particular unit is a three-section *RC* integrating filter for vertical sync. Similar units include an *RC* coupling network with resistance plate load. Inductances can also be in printed form by means of copper loops. The printed-circuit unit is much smaller than the combination of individual components replaced.

6 · 10 Localizing troubles to a receiver section

In order to localize troubles, we can use three indicators in the television receiver: the illuminated raster, the picture, and the sound. Several examples are given here, based on the typical receiver in Fig. 6 · 2.

No illumination, with normal sound. In this case, the kinescope screen is completely blank, without any light output or scanning lines and, therefore, no picture. Even with video signal at the kinescope grid, the picture cannot be reproduced unless the illuminated raster is present. The first step necessary to provide a picture, therefore, is to find out why there is no brightness. Since the sound circuits are operating, the receiver has a-c power input and the low-voltage supply is providing B+ output.

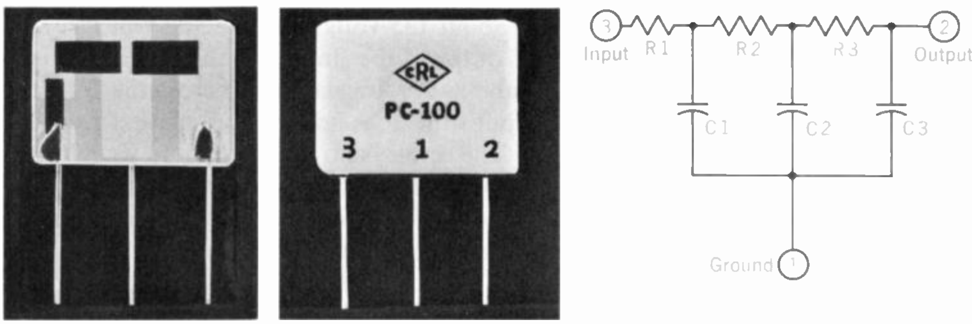
In order to have brightness, the kinescope itself must be operating normally and the high-voltage supply must provide kinescope anode voltage. It is important to remember that the horizontal deflection circuits produce the a-c input for the high-voltage power supply. Therefore, the trouble causing no illumination can be in the kinescope and its associated circuits, the high-voltage supply, or the horizontal deflection circuits. If there is d-c high voltage for the kinescope anode, the trouble is in the kinescope or its associated circuits. If there is no d-c high-voltage output, the trouble is in the high-voltage supply or the horizontal deflection circuits. The trouble can be localized further by noting whether the horizontal deflection circuits produce a-c high-voltage input at the plate cap of the high-voltage rectifier tube.

No picture and no sound, with normal raster. The normal raster indicates that the kinescope, deflection circuits, and power supplies are operating. This trouble is in the signal circuits, before the sound take-off point, because both the picture and sound are affected. The circuits common to the picture and sound signals are the r-f section, i-f amplifier, second detector, and AGC circuit. If the trouble occurs only on some channels but not on others the defect is probably in the r-f section, including the antenna and transmission line, since this is the only part of the receiver operating on the r-f signal frequencies for each individual channel.

No picture, with normal raster and normal sound. Figure 6·4*c* shows the kinescope screen with just the raster. The fact that the receiver produces the raster means the kinescope and the raster circuits are functioning, which includes the vertical and horizontal deflection circuits and the power supplies. In the signal circuits all the stages operating on the sound signal must be normal. The one section operating only on signal for the kinescope grid is the video amplifier. Therefore, the trouble must be in this stage.

Fig. 6·18 Printed couplate used as filter for integrating vertical sync. (a) Internal construction. Black areas show carbon resistance; gray areas are capacitor electrodes. (b) Complete unit. (c) Schematic diagram.

(*a*) (*b*) (*c*)



No sound, with normal raster and normal picture. The normal picture on the raster means the kinescope, deflection circuits, picture signal circuits, and power supplies are operating. The trouble must be in the sound circuits, after the sound take-off point, because only the sound is affected. This includes the 4.5-Mc sound take-off circuit in the video detector output, the 4.5-Mc sound i-f amplifier, the 4.5-Mc FM detector, the audio amplifier, and the loudspeaker. The trouble can be localized to a specific stage in the sound circuits by signal tracing to find out which circuit does not pass signal.

Only a horizontal line on the screen. Figure 6·4*b* shows how the kinescope screen looks with horizontal deflection only. Any illumination on the screen means the kinescope and its auxiliary circuits for low-voltage and high-voltage d-c operating potentials are functioning. The horizontal deflection circuits are producing the horizontal line on the kinescope screen and high voltage for brightness. Only vertical deflection is missing. Therefore, the trouble must be in the vertical deflection section of the raster circuits, which includes the vertical deflection oscillator and the vertical output stage.

No raster and no sound. The screen is completely blank, without illumination, and there is no sound. This trouble means the raster circuits and signal circuits are not operating. The defect is likely to be in the low-voltage power supply, since this section affects both the raster and the signal. Either there is no a-c input power, or the low-voltage supply is not producing B+ voltage output. If the tubes are lit, the receiver has a-c power input and the trouble must be in the B+ supply.

6·11 Multiple troubles

Usually only one defect occurs at a time, but the circuit arrangement can cause multiple effects in different sections of the receiver. The most common examples are listed here.

Series heaters. When the heaters of all tubes are connected in a series string across the power line, an open in any one heater means none of the tubes can light. Then the receiver will be completely dead, with no raster, no picture, and no sound, as all the tubes are cold, including the picture tube.

Stacked B+ voltage. In many receivers the picture i-f amplifier stages obtain plate supply voltage of about 135 volts from the 300-volt B+ supply, through the audio power output tube. In effect, the i-f section is in series with the audio output tube as a voltage divider across the B+ supply. Then when the audio output tube does not conduct plate current the i-f section has no B+ voltage. The result is no picture and no sound if the heater is open in the audio output tube. Furthermore, a weak audio output tube not only causes weak sound but picture contrast is reduced also because of lower B+ voltage for the picture i-f amplifiers.

Multiple tubes. Where a dual triode such as 6CG7 or triode-pentode such as 6AN8 is used, an open heater causes multiple troubles when the

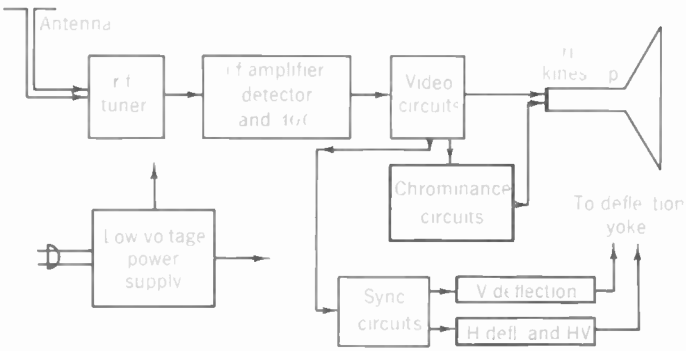


Fig. 6·19 Essentials of color television receiver.

one tube is used for two functions in different sections of the receiver. As an example, the pentode section of the 6AN8 may be a common i-f amplifier, while the triode is used for the vertical deflection oscillator. Then an open heater in this 6AN8 stage results in no vertical deflection and no picture or sound.

AGC troubles. The AGC circuit affects both picture and sound by controlling the bias on the r-f and common i-f amplifiers. Especially when a separate AGC amplifier stage is used, it can produce enough negative bias voltage to cut off the i-f amplifier, resulting in no picture and no sound. At the opposite extreme, zero AGC bias produces a reversed picture out of sync, with buzz in the sound, caused by overload distortion (see Fig. 13·13). The AGC circuit is probably the most common cause of troubles that affect both picture and sound, when the raster is normal.

6·12 Color receivers

Referring to Fig. 6·19, it can be seen that a color television receiver has all the circuits of a monochrome receiver, plus the chrominance circuits for color. This section provides color video signals for the three electron guns in the tricolor kinescope. Then the red, green, and blue phosphor dots on the kinescope screen can reproduce the picture in natural colors. Note that the conventional monochrome video signal is also necessary. Furthermore, the raster circuits, sync section, r-f tuner, i-f circuits, and power supplies are basically the same as in a monochrome receiver. More details of color television are explained in Chaps. 24 and 25.

6·13 Monitor receivers

A receiver without the r-f and i-f tuned circuits can be considered as a monitor or a "slave" receiver (see Fig. 6·20). This receiver has video cir-

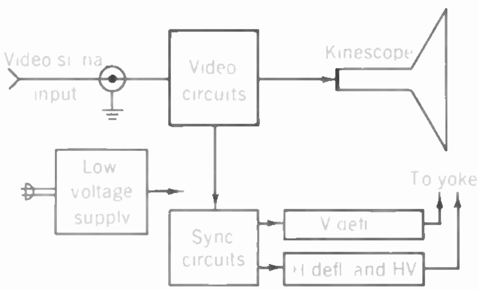


Fig. 6·20 Essentials of monitor receiver.

cuits, sync, deflection, and power supplies to reproduce a picture on the kinescope screen, but only for the video signal input supplied by coaxial cable. Generally, 1.4 volts peak-to-peak amplitude is required. The applications for such a receiver without r-f and i-f tuning include monitors in the broadcast studio and receivers for closed-circuit television where the signal is not broadcast by radio transmission.

SUMMARY

1. The functions for all stages in the block diagram of Fig. 6·2 are summarized in Table 6·3. The signal stages are listed first, followed by the raster circuits and power supplies.

Table 6·3 Functions of stages in receiver

Stage	*Input*	*Output*	*Notes*	*Common tubes*
R-f amplifier	Picture and sound r-f carrier signals from antenna	Amplified r-f picture and sound signals for mixer	Low tube noise for minimum snow in picture	6BQ7A, 6CY5, 6ER5
R-f local oscillator	None	Unmodulated r-f wave to beat with r-f signal in mixer	Usually combined with mixer stage	6J6, 6CG8
Mixer or converter	R-f picture and sound signals, plus local oscillator output	I-f picture and sound signals	Also called first detector. Usually combined with oscillator	6J6, 6CG8
Common i-f or video i-f section	I-f picture and sound signals from mixer	Amplified i-f picture and sound signals	Includes two or three stages	6BZ6, 6AU6, 6AN8
Video detector	Amplified i-f picture and sound signals	Video signal for video amplifier and 4.5-Mc FM sound signal for sound i-f amplifier	Also called second detector, or picture detector	6AL5 or germanium diode
Video amplifier	Composite video signal	Amplified composite video signal	Drives kinescope grid-cathode circuit	6AW8, 12BY7
Sound i-f amplifier	4.5-Mc sound i-f signal from video detector	Amplified 4.5-Mc sound signal for FM detector	Always tuned to 4.5 Mc for intercarrier sound	6AU6
FM sound detector	Amplified 4.5-Mc sound i-f signal	Audio signal	Common circuits are discriminator, ratio detector, and gated-beam detector	6AL5, 6T8, 6DT6, germanium diodes

Stage	*Input*	*Output*	*Notes*	*Common tubes*
Audio section	Audio signal from FM sound detector	Audio power output for loudspeaker	May have one or two stages	6AV6, 6BQ5
Sync separator	Composite video signal from video amplifier	Sync pulses for horizontal and vertical synchronization	Also amplifies the separated sync	6BU8, 6SN7, 12AU7, 12BH7
Vertical oscillator	60-cps vertical sync pulses	60-cps deflection voltage to drive vertical amplifier	Can produce output with or without sync input	6DE7, 6EM7
Vertical amplifier	Deflection voltage at 60 cps from vertical oscillator	60-cps sawtooth current for vertical scanning coils in deflection yoke	Output has same frequency as oscillator	6DE7, 6EM7
Horizontal AFC	15,750-cps horizontal sync pulses	D-c control voltage for horizontal oscillator	Makes horizontal synchronization immune to noise pulses	6SN7, 6CG7, 6AL5, germanium diodes
Horizontal oscillator	D-c control voltage from horizontal AFC circuit	15,750-cps deflection voltage to drive horizontal amplifier	Can operate with or without d-c control voltage	6SN7, 6CG7
Horizontal amplifier	Deflection voltage at 15,750 cps from horizontal oscillator	15,750-cps deflection current for horizontal scanning coils in deflection yoke	Also supplies power input to high-voltage rectifier and the damper stage	6DQ6, 6BG6
Damper	A-c oscillations after flyback in horizontal output	Rectified horizontal deflection voltage	Supplies B+ boost voltage	6AU4, 6AX4, 6W4, 6BH3
High-voltage rectifier	Horizontal flyback pulses at 15,750 cps	D-c high voltage of about 15 kv for kinescope anode	Needs horizontal output for operation	1B3, 1J3, 1K3
Low-voltage power supply	A-c power from 120-volt main line	B+ supply voltage and heater power for all tubes	May be a-c supply or a-c/d-c transformerless type	5U4, silicon diodes

2. Intercarrier-sound receivers obtain the associated FM sound signal as the 4.5-Mc beat between picture and sound carrier frequencies in the output of the video detector. In split-sound receivers, the sound take-off is before the video detector, usually in the output of the mixer stage. The advantage of intercarrier sound is that the 4.5-Mc sound i-f signal is automatically present at its correct frequency when the picture is tuned in.
3. In projection television, a very bright picture is enlarged by an optical lens, as light from the small kinescope screen is projected onto the large viewing screen.
4. In pay television, the picture is scrambled unless the viewer pays a fee for the decoding signal that produces a normal picture.
5. Troubles can be localized to a specific stage by noting its function in producing picture, sound, and raster. However, circuits that produce multiple troubles include series heaters, stacked B+ supply, AGC amplifier, and multiple-tube arrangements.

SELF-EXAMINATION *(Answers at back of book.)*

Part A

Match the circuits listed at the left with the troubles at the right, for the intercarrier-sound receiver in Fig. 6·2. Assume parallel heater connections.

TROUBLE SYMPTOM	DEFECTIVE STAGE
1. No picture, no sound, no raster	*a.* Vertical oscillator
2. Horizontal line across screen	*b.* R-f amplifier
3. Snowy picture	*c.* 4.5-Mc i-f amplifier
4. No sound, picture normal	*d.* Horizontal AFC
5. No brightness, normal sound	*e.* Low-voltage rectifier
6. Picture torn apart in diagonal bars	*f.* Video amplifier
7. No picture, normal raster and sound	*g.* Audio output (stacked B+)
8. No picture, no sound, normal raster	*h.* Horizontal output

Part B

Match the tube type numbers at the left with stages at the right.

1. 5U4	*a.* Horizontal damper
2. 1B3	*b.* R-f amplifier in tuner
3. 6AX4	*c.* Converter in tuner
4. 6BQ7	*d.* High-voltage rectifier
5. 6J6	*e.* Horizontal output
6. 6BQ6	*f.* Audio output
7. 6BQ5	*g.* Low-voltage rectifier
8. 6DT6	*h.* I-f amplifier
9. 6CG7	*i.* Horizontal oscillator
10. 6BZ6	*j.* FM sound detector

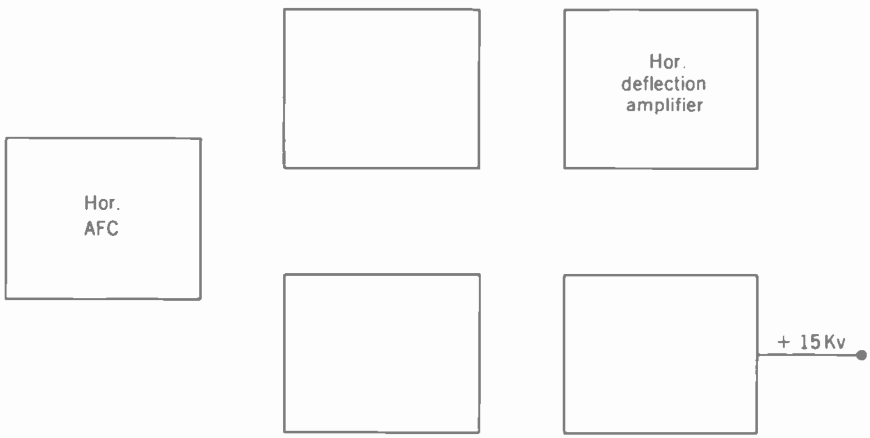
Part C

Match the controls at the left with the functions at the right.

1. Fine tuning	*a.* Tunes r-f, oscillator, and mixer stages
2. Volume	*b.* Varies frequency of local oscillator
3. Tone	*c.* Gain or level for audio signal
4. Contrast	*d.* Frequency response of audio amplifier
5. Brightness	*e.* Gain or level for video signal
6. Station selector	*f.* D-c bias for kinescope

ESSAY QUESTIONS

1. Using all the stages listed in Table 6·2, classify them under the following headings: picture and sound; picture alone; sound alone; illuminated raster.
2. Give the function of: (*a*) antenna and transmission line; (*b*) sound take-off circuit.
3. Give two functions of the composite video signal.
4. Give two reasons why sufficient brightness is a problem in projection television.
5. Give two locations for the sound take-off circuit in intercarrier-sound receivers.
6. Give two locations for the sound take-off circuit in split-sound receivers.
7. For pay television give two methods for each of the following parts of the system: (*a*) sending out the signal to subscribers; (*b*) making the signal private; (*c*) collecting the fee from subscribers.
8. Give two advantages of a crystal diode detector compared with a diode section of the 6AL5 vacuum tube.
9. Give one advantage and one disadvantage of a silicon rectifier for the low-voltage power supply, compared with a vacuum tube.
10. Give two advantages of transistors for vertical and horizontal deflection circuits.
11. Give two features of printed wiring.
12. In Fig. 6·2, why is the section for sync circuits included in the signal circuits, not in the raster circuits?
13. For the block diagram below, indicate the name of each stage, with typical tubes and arrows for the signal paths.



Chapter 7 *Picture tubes*

The picture tube or kinescope is a cathode-ray tube consisting of an electron gun and phosphor screen inside the vacuumed glass envelope. The narrow glass neck contains the electron gun, which produces a beam of electrons accelerated to the screen by the high voltage on the anode. To form the screen, the inner surface of the wide glass face at the front of the tube is coated with a luminescent material that produces light when excited by electrons in the beam. At the end of the narrow neck, either an electric or magnetic field is provided to make the electron beam scan the entire surface of the screen to produce the raster. With video signal applied to the grid-cathode circuit of the electron gun to vary the beam intensity, picture information is shown as light and dark variations on the screen.

7 • 1 Deflection, focusing, and centering

Either electrostatic or electromagnetic deflection can be used for a cathode-ray tube. For electrostatic deflection, two pairs of metal deflection plates are attached to the gun structure within the tube. As the electron beam moves between the plates, sawtooth voltage applied across the plates provides an electrostatic field to deflect the beam. Electrostatic deflection is generally used in oscilloscope tubes, with anode voltage of 5 kv or less. In electromagnetic deflection, however, two pairs of deflection coils are mounted in the yoke housing on the kinescope neck. The associated magnetic field of sawtooth current in the coils deflects the electron beam. Practically all picture tubes use magnetic deflection.

Deflection yoke adjustment. In Fig. 7 • 1, note that the deflection yoke is placed as far forward as possible, against the wide bell of the envelope. If the yoke is too far back, the beam will be deflected off the screen for large deflection angles and the corners of the screen will be dark, or shadowed. The entire yoke can be turned in its housing, by loosening the wing nut at the top. Moving the yoke left or right tilts the raster. Adjust for a raster parallel with the screen, or the picture will be tilted left or right.

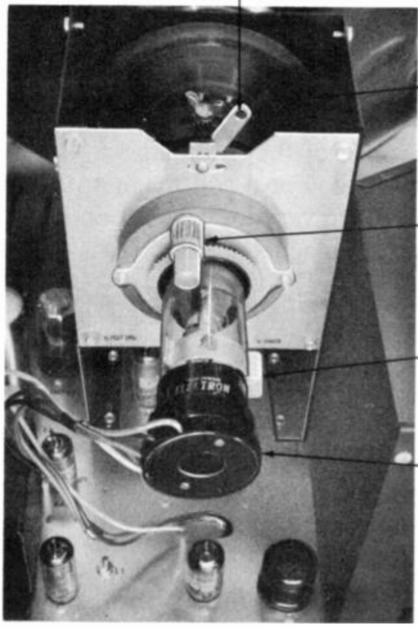


Fig. 7·1 Kinescope using magnetic deflection and electrostatic focus. (RCA Institutes Home Study School.)

Focus adjustment. The electron beam must be focused to a small spot of light on the screen for sharp scanning lines and fine details in the reproduced picture. Usually focus is sharpest in the center area. As with deflection, focusing of the electron beam can be either electrostatic or magnetic. With electrostatic focusing, voltage applied to the focusing electrode of the electron gun controls beam focus.

For magnetic focusing, a magnet is mounted externally on the kinescope neck, as in Fig. 7·2. This may be either a permanent magnet or a coil magnet, with direct current for a steady magnetic field. In either case, the focusing magnet is mounted approximately ½ in. behind the yoke. For sharp focus,

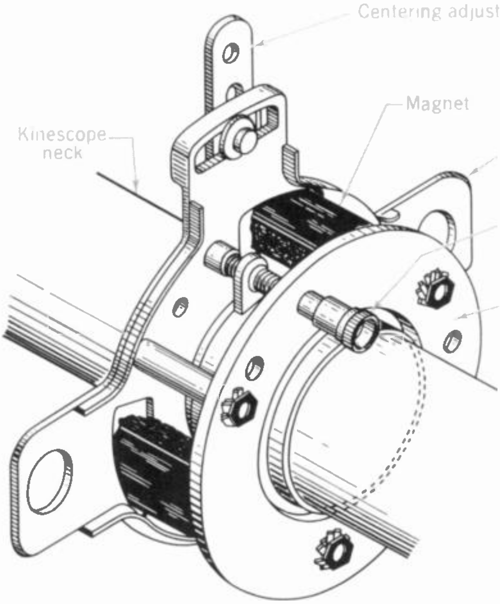


Fig. 7·2 Focusing magnet, with magnetic shunt and centering ring.

move the magnet back and forth and shift it slightly around the kinescope neck. Best focus is obtained with the tube neck approximately centered in the hole of the ring magnet. Usually, a fine focusing adjustment is also provided. For a coil magnet, the fine focus control is a rheostat on the chassis to vary the amount of direct current in the focus coil. With a permanent magnet, the fine focus control is a movable iron ring, as shown in Fig. 7·2. The thumbscrew focusing adjustment moves the iron magnetic shunt to control the strength of the magnetic focusing field.

Centering adjustment. Provision is generally made for a steady displacement of the electron beam for centering. When you center the beam while it is scanning the screen, the entire raster is centered.

For electrical centering, direct current can be supplied through the horizontal and vertical deflection coils. The direct current adds a steady displacement component to the deflection field of the a-c sawtooth scanning current. Two rheostats on the chassis adjust the direct current to serve as horizontal and vertical centering controls. This method is not used often, except for color kinescopes, because of the added current drain on the low-voltage power supply.

Mechanical arrangements with a permanent magnet can provide a steady displacement of the magnetic field lines through which the electron beam passes on its way to the screen. In one method, a small permanent-magnet ring is clamped onto the back of the yoke housing. In some cases, the yoke housing includes adjustable permanent magnets for centering. Or, the centering ring magnet may be in the focus-magnet assembly, as in Fig.

Fig. 7·3 Placement of ion-trap magnet, centering rings, and pincushion magnets used for some kinescopes.

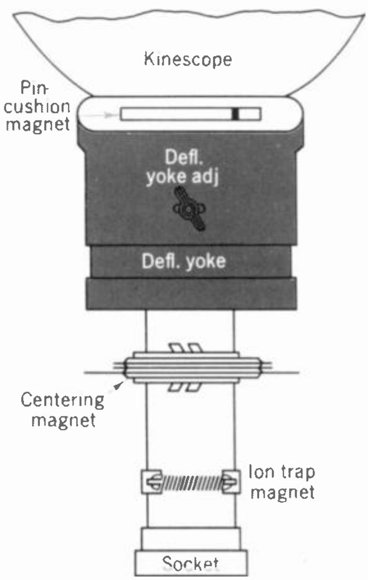
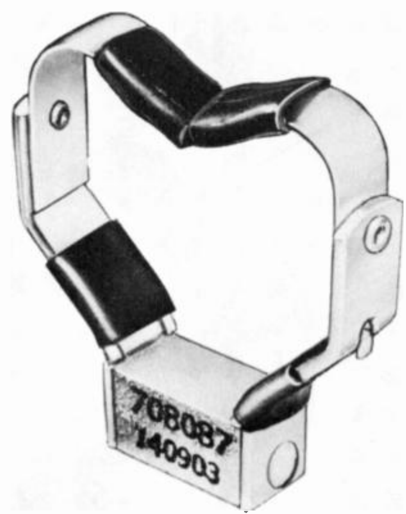


Fig. 7·4 Ion-trap magnet. (Clarostat Mfg. Co., Inc.)



7·2. The centering ring is eccentric with the center hole in the focus magnet. When you move the lever on the eccentric ring up or down, the raster moves left or right; moving the magnet horizontally shifts the centering vertically. If there is no centering ring, the focus magnet itself can be adjusted. Kinescopes using electrostatic focusing generally have centering ring magnets as shown in Fig. 7·3, because there is no focus magnet.

Pincushion magnets. Yokes for wide-angle deflection may have pincushion magnets, as in Fig. 7·3, to correct bowing at the top and bottom edges of the raster. There are two bar magnets in removable plastic holders at the top and bottom of the yoke housing. The magnets, with opposite polarities, are placed to the front or back for straight raster edges. See the yoke in Fig. 7·16. There can also be two magnets for the sides.

Ion-trap magnet. When magnetic deflection is used for a cathode-ray tube, a brown spot can form at the center of the screen because of bombardment by ions in the beam current. This circular area about 1 in. diameter on a 20-in. screen is an ion burn or ion spot. It is produced by negative ions emitted from the cathode of the electron gun and attracted to the screen by the anode voltage. Any residual gas ionized by the electron beam is a source of ions in the tube. When positive ionized gas molecules bombard the cathode of the electron gun its oxide coating emits negative ions, which become part of the beam current and produce the ion spot.

Remember that ions have much more mass and therefore less velocity than electrons in the beam. The slow-moving ions correspond to a small current and a weak magnetic field. In magnetic deflection then, there is little reaction between the ions and the magnetic field of the scanning current. As a result, the ions are not deflected with the electron beam. The steady bombardment by ions at the center of the screen causes the burnt spot. Note that ion spot is no problem in electrostatic deflection, which depends on charge rather than current. In this case, the deflecting voltage moves the ion charges with the electrons in the scanning beam.

In one method of eliminating ion spot in picture tubes using magnetic deflection the total cathode beam current, including electrons and ions, is made to leave the cathode at such an angle that the beam will strike the side of the tube instead of the screen. Either a bent gun can be used, or a straight gun with a diagonal cut. A bent gun is shown in Fig. 7·12. Then a magnetic field to deflect the electron beam back to the screen is provided by a small external magnet, called the *ion-trap magnet* (Fig. 7·4). Such a magnet is also called a *beam bender.* The ions are not deflected by the magnet because of the weak magnetic field of the ion current. The ions strike the electrodes in the tube and are collected as waste current.

The ion-trap magnet is placed over the electron gun, near the tube base, as in Fig. 7·3. By moving the magnet slowly back and forth, and rotating it, you can find the position where the magnet bends the electron beam back to center to produce illumination on the screen. The ion-trap magnet must be adjusted exactly for maximum brightness.

Picture tubes with an aluminized screen do not require an ion-trap magnet because the metalized backing prevents ions from going through to the screen. These tubes have a straight gun that aims the electron beam along the center line of the tube. In this case no ion-trap magnet should be used, as it would deflect the beam off the screen. Any tube with a bent gun needs an ion trap magnet, however, even if it has an aluminized screen.

7·2 The luminescent screen

The inside surface of the glass faceplate is coated with a chemical phosphor to form the screen of the picture tube. The screen emits light when bombarded by electrons in the scanning beam, thereby converting electric signal into visible light. Radiation of light from the screen as it is excited by the electron beam is *luminescence.* This term applies to light that is not produced by incandescent heating. When the radiated light is extinguished practically instantaneously after excitation by the electron beam has ceased, the screen is *fluorescent.* Continued emission of light after excitation is *phosphorescence.*

Table 7·1 lists several screen phosphors; the missing P numbers have similar uses but some types are obsolete. Most common are the P1 green phosphor for oscilloscope tubes and the P4 white phosphor for picture tubes. Note the persistence characteristics. This can be defined as the time for light emitted from the screen to decay to one per cent of its maximum value. Some persistence is desirable to increase brightness and reduce flicker. However, the persistence time must be less than 1/30 sec for picture tubes so that one frame cannot persist into the next and cause blurring of moving objects. The decay time for most black-and-white kinescopes is approximately 0.005 sec, a medium-short persistence. For the P1 phosphor, with medium persistence, the decay time is 0.05 sec.

Table 7·1 Typical screen phosphors for cathode-ray tubes

Phosphor number	*Color*	*Persistence*	*Uses*
P1	Green	Medium	Oscilloscopes
P2	Yellow-green	Medium	
P4	White	Medium-short	Picture tubes. Sulfide type for direct view; silicate type for projection tubes
P7	White, yellow	Short, long	Two-layer screen
P11	Blue	Medium-short	
P12	Orange	Long	
P14	Blue, orange	Short, long	Two-layer screen
P15	Green-ultraviolet	Very short	Flying-spot scanner
P16	Purple-ultraviolet	Very short	Flying-spot scanner
P22	Red, green, and blue	Medium	Tricolor kinescope. Phosphor dots for each color

The phosphors are generally light metals such as zinc and cadmium in the form of sulfide, sulfate, and phosphate compounds. The screen material is processed to produce very fine particles which are then applied to the glass faceplate in a thin uniform layer. In terms of molecular structure, the phosphors are imperfect crystals where an activator element such as manganese or silver can be added to distort the crystal lattice. Then high-velocity electrons can excite the phosphor to emit light. For the green P1 phosphor, a form of zinc silicate called *willemite* is generally used. The P4 phosphor usually combines compounds of zinc sulfide, cadmium sulfide, or zinc silicate. For color television, the P22 phosphor includes zinc sulfide for blue, zinc silicate for green, and zinc phosphate for red.

Sticking potential. The phosphor materials should be able to produce secondary emission of electrons because this is how the screen becomes charged to the anode voltage. In cathode-ray tubes, the return path for the electron-beam current that strikes the screen is provided by secondary emission of electrons from the phosphor. These secondary electrons are collected by the anode. Because of the secondary electrons emitted from the screen, it can charge to a positive potential approximately equal to the anode voltage. Then electrons in the incident beam are accelerated to the screen with high velocity to produce high brightness. The electron beam is not attracted to the anode wall because its positive potential is symmetrical around the tube envelope.

The amount of secondary emission from the screen is limited, however. With an anode voltage higher than 10 kv, approximately, the screen may not be able to charge to the anode potential. Then the screen remains at the highest positive potential to which it can charge by secondary emission. This value is the *sticking potential* of the screen. It is not possible to have a higher effective accelerating voltage than the sticking potential. As an example, if the anode voltage is 10 kv but the sticking potential is 8 kv, only 8 kv is effective in accelerating the electron beam because this is the potential difference between the phosphor screen and the electron-gun cathode.

Aluminized screen. To overcome the problem of sticking potential, many picture tubes have a coating of aluminum on the back surface of the screen phosphor. The metal film is very thin, being vaporized aluminum condensed upon the screen. With anode voltages of 10 kv or more, the electrons in the beam have enough velocity to pass through the aluminum coating and excite the phosphor. Because of its metalized backing the screen potential is essentially the same as the anode voltage.

There are several other advantages provided by the aluminized screen. The ion spot is eliminated without the need for an ion trap because slowly moving ions cannot pass through the metal film. Also, the conductive coating makes the screen insensitive to stray capacitance effects. Finally, the metal-backed screen provides an important optical advantage because light from the phosphor that would radiate to the inside of the tube is reflected from the smooth metal film. As a result, the amount of light available at the front of the screen is increased. The combination of a high effective screen voltage and efficient use of the emitted light enables the aluminized screen to have about twice the light output of a nonaluminized screen, when the anode voltage is 10 to 20 kv.

7·3 Types of picture tubes

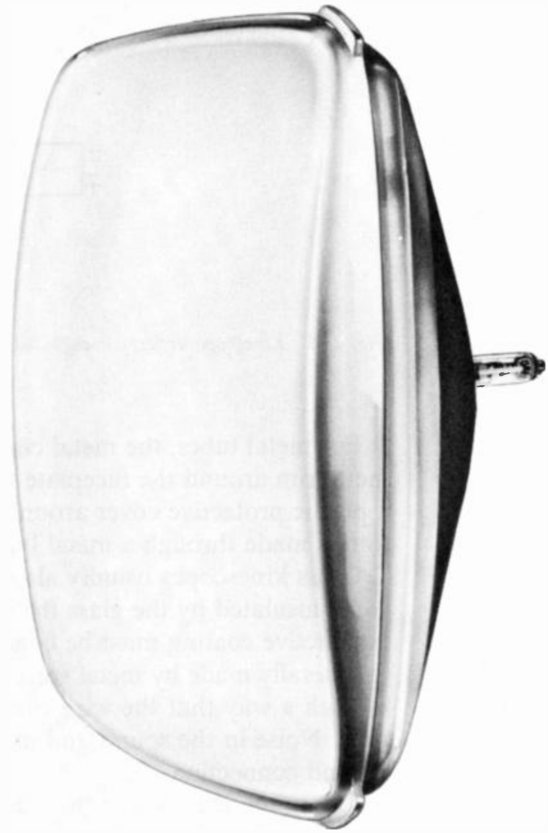
If we take the 23CP4 as an example, 23 in. is the diagonal of its rectangular screen, within ½ in. For a round screen, its diameter is specified. The P number at the end specifies the type of phosphor screen, which is P4 for all black-and-white kinescopes. The letter or letters in the middle are assigned alphabetically for different types with the same screen size. Note that tubes of different sizes with the same letter, such as 23CP4 and 20CP4, do not necessarily have the same features. These two kinescopes are definitely not interchangeable because the 20CP4 uses magnetic focusing while the 23CP4 uses electrostatic focusing; the 20CP4 requires an ion-trap magnet but the 23CP4 does not; and each requires a different type of deflection yoke for different deflection angles. Table 7·2 lists typical tubes.

A suffix letter such as the A in 20CP4A may be added to indicate a slight change that is not enough to require a new type number. The modified type can be substituted for any previous model. However, an earlier model may not be a direct substitute for a modified type. This system applies to tubes in general. For picture tubes, the suffix letter A or B often indicates the addition of an external conductive coating on the glass en-

Table 7·2 Types of picture tubes

Description	*Type numbers*	*Remarks*
Magnetic focus and deflection	10BP4, 12LP4, 16LP4, 16WP4	Round, glass
	14BP4, 16RP4, 17AP4, 20CP4, 21EP4, 24CP4, 27EP4	Rectangular, glass
	16AP4, 16GP4, 19AP4	Metal, round
Low-voltage focus and magnetic deflection	14XP4, 16AEP4, 17ATP4, 17LP4, 17TP4, 20HP4, 21AQP4, 21MP4, 21YP4, 21ZP4, 24BP4, 27SP4	Deflection angle 65° to 90°. Ion trap for these and types listed above.
	17DKP4, 19YP4, 21CEP4	Deflection angle 110°. No ion trap for these and types listed below
	23CP4, 23EP4	Bipanel faceplate
Projection kinescopes	5TP4, 7NP4, 7WP4	30 to 80 kv anode voltage
Flying-spot scanner	5ZP16	For camera pickup
Test kinescopes	5AXP4, 8YP4	For checking receivers
Monitor viewfinder	3KP4, 5FP4, 8DP4, 8HP4	For compact equipment
Tricolor kinescope	21AXP22, 21FBP22, 21FJP22	Three guns for red, green, and blue phosphor dots

Fig. 7·5 *Bipanel picture tube 23CP4. Note glass mounting lugs at corners of protective panel.* (RCA.)



velope, or an aluminized screen that can be operated with higher anode voltage, or a different faceplate. Kinescopes differing only in the type of glass faceplate are interchangeable.

Faceplate. Picture tubes are made with a thick glass faceplate that is almost flat. Approximately ½-in. thickness provides the strength required to withstand the air pressure on the vacuumed envelope. The thick glass also makes it possible to construct the envelope with a rectangular faceplate. Practically all picture tubes now are made with a rectangular screen to eliminate the wasted area of a rectangular raster on a circular screen. The result is a more compact installation in the receiver cabinet.

The faceplate is manufactured with a neutral light-absorbing material in the glass, making it gray. This feature makes black areas darker to increase optical contrast when the picture is viewed in a lighted room. In addition, the faceplate is etched or frosted, like a light bulb, to diffuse the effect of bright light in the room reflected from dark areas of the screen. This combination of a gray filterglass faceplate transmits approximately 78 per cent of the light from the screen to the viewer.

A later type of faceplate construction is illustrated by the 23CP4 in Fig. 7·5, showing a filter-glass protective panel sealed to the faceplate of the tube. As a result, the tube can be mounted in the receiver cabinet without a safety-glass window and the rubber dust seal around the tube is not necessary. Without any air separation between faceplate and protective window, reflections are reduced and contrast is improved. The total light transmission of the faceplate and protective panel is approximately 40 per cent.

The anode. In glass kinescopes the anode is a conductive coating on the inside wall of the tube. Because of the high voltage, a separate anode connector on the side of the cone is used, instead of connecting to the tube base. This recessed-cavity connector is a hole of about ¼ in. diameter through the glass wall, into which a metal ball spring or a spring clip fits to contact the anode coating on the inside wall.

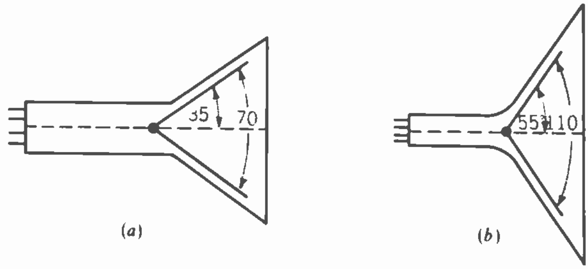


Fig. 7·6 Kinescope deflection angle. (a) 70°. (b) Shorter tube with 110° angle.

For metal tubes, the metal cone is the anode. A metal strip contacts the metal rim around the faceplate for the high-voltage connection. If there is a plastic protective cover around the metal cone, the high-voltage connection is made through a metal button on the cover.

Glass kinescopes usually also have an external graphite coating on the cone, insulated by the glass from the anode wall coating inside. The outer conductive coating must be connected to chassis ground. This connection is generally made by metal spring wire riveted to the yoke mounting frame in such a way that the wire can brush against the coating on the picture tube. Noise in the sound and flashes in the picture can result from a poor ground connection.

The grounded coating provides a filter capacitance for the high voltage of the anode supply. The external coating serves as one plate; the anode coating is the other plate, with the glass bulb the insulator between the two plates. The capacitance is 2,000 $\mu\mu$f, approximately.

For some tubes where one or more accelerating grids in the electron gun are connected internally to the anode wall coating these electrodes are considered collectively as the *ultor,* since all have the ultimate high voltage. In any case, the ultor or anode connection provides the high voltage needed to accelerate the electron beam to the screen to produce brightness.

Deflection angle. This specifies the maximum angle through which the beam can be deflected to fill the screen without striking the sides of the bulb. The deflection angle for typical kinescopes is 50° for older tubes to 114°, approximately, for later tubes that are not so long. As shown in Fig. 7·6, the total angle is specified. A deflection angle of 110° means the electron beam can be deflected 55° from the center line of the tube. For rectangular kinescopes, the maximum deflection angle is along the diagonal between opposite corners of the screen. The advantage of a larger deflection angle is the fact that the shorter bulb can be installed in a smaller receiver cabinet.

The angle specified for a picture tube only means it can accommodate so much deflection but the power needed to deflect the electron beam is supplied by the electromagnetic field of current in the deflection yoke. This scanning current is generated in the deflection circuits of the receiver.

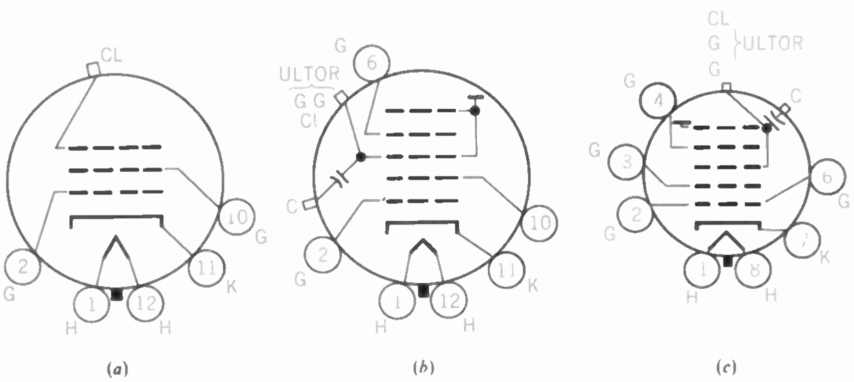
Therefore, deflection yokes also are specified in term of deflection angle. As an example, a 70° yoke can fill the screen of a kinescope with a 70° deflection angle, but on a 90° tube, the 70° yoke will produce a raster that is too small. However it can be useful to note that different screen sizes can be filled if the deflection angle is the same as the yoke. For instance, a 70° yoke can fill a screen of 14-, 17-, or 21-in. kinescopes if all these tubes have a deflection angle of 70°, since bigger tubes with the same angle will be longer.

Larger deflection angles require a stronger electromagnetic field for scanning. For this reason, 110° tubes are made with a narrow neck to allow greater field intensity at the point where the electron beam is deflected. A 110° yoke has a hole diameter of 1⅛ in. compared with 1 7/16 in. neck diameter for tubes with a smaller deflection angle.

Base and socket connections. The 12-pin (duodecal) base in Fig. 7·7*a* is typical for picture tubes using magnetic focus. Then the triode electron gun includes heater, cathode, control grid G1, and an accelerating grid G2. The anode is labelled CL. Sometimes G2 is called an accelerating anode but the standard name is grid because G2 is not intended to collect electrons. The basing diagram in *b* indicates a pentode gun that provides for electrostatic focus. G6 is the focusing grid, which requires about 300 volts. The same type of gun for low-voltage electrostatic focusing is shown in *c* but this is a narrower 8-pin button base for 110° tubes. In this type, the pins go right through the glass base, as in miniature glass tubes, but there is a plastic indexing plug.

Input capacitance. This is the capacitance for input video signal applied to either grid or cathode. For practically all picture tubes, C_{in} from control grid to all other electrodes is approximately 6 $\mu\mu f$; from cathode to all other electrodes C_{in} is 5 $\mu\mu f$.

Fig 7·7 Typical base and socket connections, bottom view. CL is anode collector and C is external coating. (a) Duodecal 5-pin for tubes with magnetic deflection and focus. (b) Duodecal 6-pin for tubes with low-voltage focus. (c) Miniature base for 110° tubes.



7·4 The electron beam

The electron gun in Fig. 7·8 includes a heated cathode to emit electrons, a control grid G1, and the accelerating grid G2. Not counting the anode, there are three electrodes in this triode electron gun. Tetrode and pentode electron guns have additional grids. Each grid structure is a cylinder with a small aperture or hole in the center. Figure 7·9 shows a control-grid cylinder and its disk cover with an aperture of 0.04 in. The electrodes are generally made of nickel or a nickel alloy, mounted on ceramic insulator supports.

The control-grid voltage is a negative bias to control the space charge at the cathode. The accelerating grids have progressively higher positive potentials, with the anode at the highest voltage to accelerate the electron beam to the screen. Most of the electrons go through the apertures and are not collected by the electrodes because their circular structure provides a symmetrical accelerating field around all sides of the beam. The electron beam has a complete circuit because of secondary emission from the screen, as secondary electrons collected by the anode return to cathode through the high-voltage supply. A typical value of beam current is 0.6 ma with 15 kv anode voltage.

Electron lenses. Electrons emitted from the cathode tend to diverge because they repel each other. Consequently, special attention must be given to forming the electrons into a narrow beam. This action is analogous to focusing a beam of light by optical lenses. Therefore, the term *focusing* is used for the action of obtaining a narrow beam, while a focusing system for the electrons is an *electron lens.* Two electron lenses are generally used. The first lens is the electrostatic field between cathode and control grid produced by their difference in potential. The second lens may be either an electrostatic field or magnetic field, provided for focusing the beam just

Fig. 7·8 Elements of a triode electron gun. Focusing and deflection not shown.

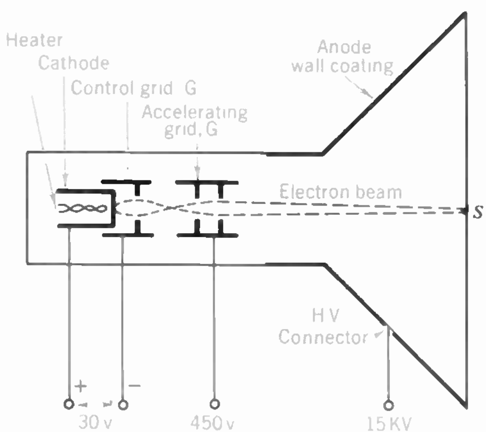
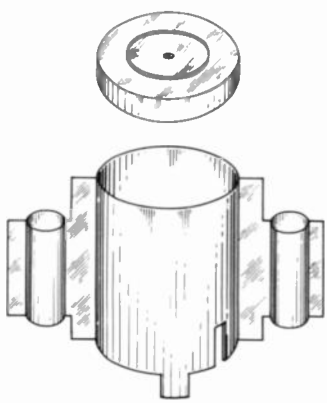


Fig. 7·9 Control-grid cylinder with aperture disk. (Sylvania Electric Products, Inc.)



before deflection. As a result of the two electron lenses the beam is focused to produce a small, sharp spot of light on the screen.

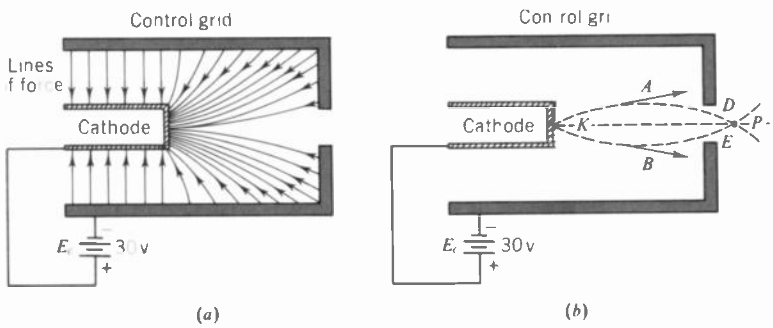
Crossover point. Referring to Fig. 7·10*a*, electrostatic lines of force are shown repelling electrons back to cathode because the control-grid bias is negative. The lines are straight where the cathode and grid structures are parallel. Such straight lines indicate a uniform change of potential in the space between grid and cathode. However, where the grid does not have uniform distances to cathode, the lines of force curve. Notice that the curved lines of force have the direction that repels electrons toward the center axis. Electrons diverging the most have the greatest force toward the center.

Remember now that the G2 voltage and anode voltage provide a forward accelerating force. The net result is that diverging paths are bent so that the electrons go through the grid aperture (see Fig. 7·10*b*). The diverging beam then is focused at point *P* just beyond the control grid. Note that electrons emitted in the direction *KA* are made to follow the curved path *KDP*. Similarly, electrons from path *KB* are forced into path *KEP*. Electrons in a straight path along the center axis stay in this line.

The focal point *P* is the crossover point produced by the first electron lens. *P* serves as a point source of electrons to be imaged onto the screen by the second electron lens for a sharp spot. Fine focus can be produced this way because the crossover point is much smaller than the cathode area supplying electrons for the beam.

Bias control. The d-c voltage between control grid and cathode is the bias determining the amount of beam current that strikes the screen. Less negative bias allows more electrons to be accelerated through the grid aperture to form the beam. More negative bias reduces the beam current. The amount of negative bias voltage needed to reduce the beam current to the point of visual cutoff, when no light is produced by the screen, is about 50 volts, depending on the gun structure and accelerating voltages.

Fig. 7·10 Action of grid-cathode electrostatic field as first electron lens to focus beam at crossover point P. (a) Electrostatic lines of force without beam. (b) Diverging beam from K focused at P (lines of force omitted).



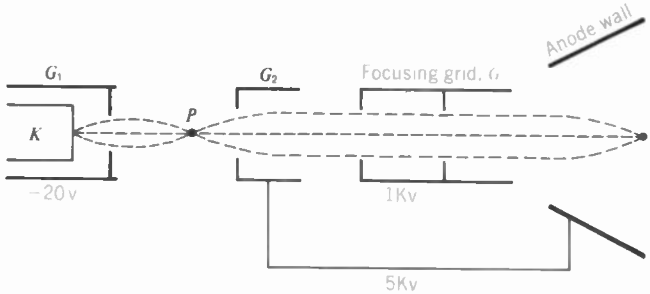


Fig. 7·11 Tetrode electron gun using high-voltage electrostatic focusing for second electron lens.

For oscilloscopes, the *intensity* control varies the bias on the cathode-ray tube to adjust screen brightness; in television receivers the bias control for the picture is called *brightness.* Although the bias controls the amount of beam current, focus is also affected to some extent, as the crossover point moves farther from the cathode with less negative control-grid voltage.

Intensity modulation. In addition to its d-c bias the grid-cathode circuit can have a-c signal voltage applied. Then the signal varies the beam current above and below its average value set by the bias. The variation by control-grid voltage is called *intensity modulation* or *Z-axis modulation* of the beam current. Most oscilloscopes have a terminal connection for intensity modulation, when desired, which is coupled to the control grid of the cathode-ray tube. For picture tubes, the a-c video signal is coupled to the grid-cathode circuit to vary the beam intensity and illumination to reproduce the visual information for all the picture elements.

7·5 *Focusing the electron beam*

For the second electron lens, three methods of converging electrons in the beam to a sharp point on the screen are described here. Either an electric field or magnetic field can be used.

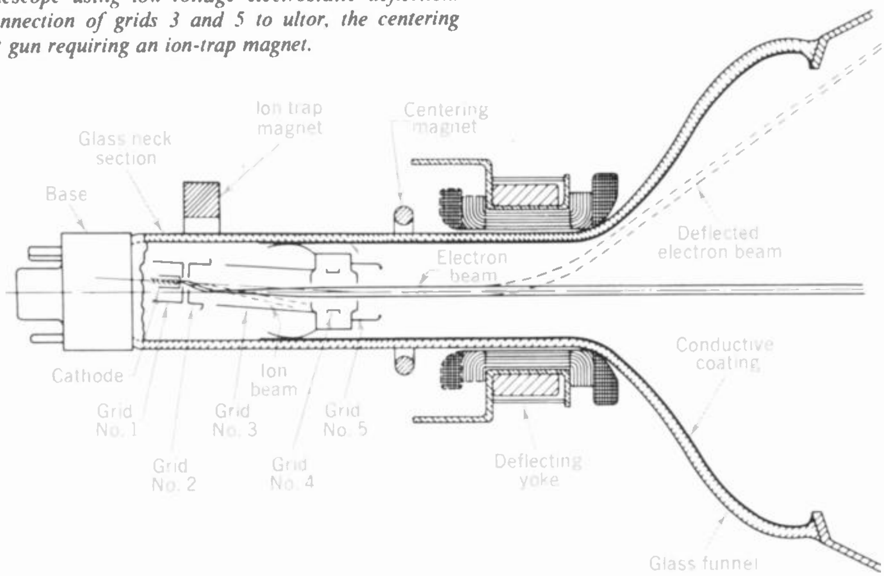
High-voltage electrostatic focusing. The second electron lens can be provided by the electrostatic field of the focusing grid in a tetrode gun as illustrated in Fig. 7·11. Voltages indicated are for an oscilloscope tube, which generally uses this type of gun. G2 is connected internally to the anode wall coating and has its potential of 5 kv. The focusing grid has a potential about one-fifth the anode voltage. This method is generally used for color kinescopes, with a separate high-voltage supply for the focusing voltage.

In Fig. 7·11, note that the focusing grid voltage of 1 kv provides a decelerating field for electrons coming from G2 at 5 kv. This decelerating field repels diverging electrons toward the center line of the beam, corresponding to the action of the first electron lens between cathode and grid. By adjusting the amount of voltage applied to the focusing grid, the electrostatic focusing field can be varied to produce electron paths with the curvature required to image the crossover point *P* onto the screen at point *S*.

Low-voltage electrostatic focusing. Most pictures tubes with electrostatic focus use the electron gun illustrated in Fig. 7 · 12. Grid 1 is the control grid, grid 2 is an accelerating grid at about 450 volts, grids 3 and 5 are connected internally to the anode wall coating at about 15 kv, and grid 4 is the focusing grid. As a result, the electron beam can be decelerated by grid 4 between grid 3 and grid 5. With this type of electron lens, the B+ supply can provide the low focusing voltage required, which may be between −30 volts and +300 volts. Usually there is no focusing control but a wire jumper on the tube base connects G4 to one of the other pins, as the required focusing voltage is not critical.

Magnetic focusing. Referring to Fig. 7 · 13, the focus magnet provides lines of force that are essentially in the direction of the electron beam. However, electrons in the beam also have an associated magnetic field. These lines of force are circular around the beam axis in a plane perpendicular to both the electron beam and the focusing field. When the magnetic field of the electron beam is at right angles to the focusing field, these two fields do not react with each other. Remember that the field strength is not changed with two perpendicular fields. Therefore, these electrons can proceed along

Fig. 7 · 12 Kinescope using low-voltage electrostatic deflection. Note internal connection of grids 3 and 5 to ultor, the centering magnet, and bent gun requiring an ion-trap magnet.



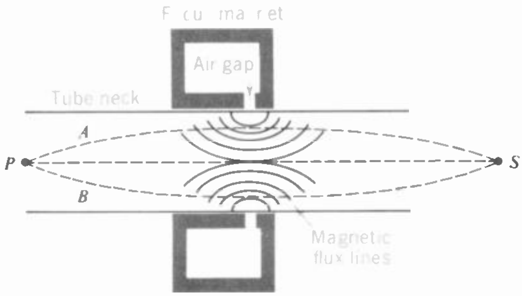


Fig. 7 · 13 Focus magnet as second electron lens.

the center axis toward the screen, accelerated by the anode voltage. However, an electron that travels along the line *PA* is moving at an angle and has a component of motion perpendicular to the focusing field. Therefore, part of the magnetic field associated with this electron can be considered in parallel with the field produced by the focus magnet. Where the lines of force are in the same direction they add to produce a stronger field, while opposing lines of force cancel to produce a weaker field. The reaction between the two fields produces a force that moves the electrons toward the weaker field.

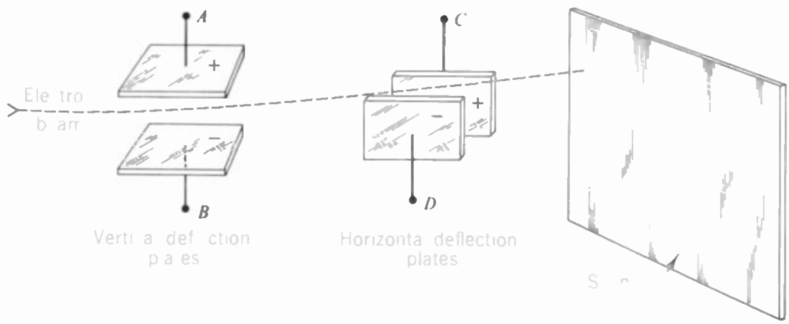
As a result, a force is applied to those electrons that travel in paths such as *PA* and *PB* in Fig. 7·13 at an angle with respect to the beam axis. The electrons are forced to move toward the center axis of the electron gun. Electrons in the outer edges of the diverging beam have paths that make the largest angle with the horizontal axis. They are subject to a greater force toward the center than electrons nearer the axis. The result is a converging electron beam. By adjusting the position of the focus magnet and its strength the electrons can be given a component of motion toward the axis of the gun which persists after the focusing field is passed, so that the electron beam converges at point *S* on the screen.

The motion of the electron resulting from the action of the magnetic field must be perpendicular both to the direction of beam current and to the focusing field. Therefore, the electrons actually follow a circular spiral motion toward the center axis as they are accelerated to the screen by the anode voltage. However, the spiral motion is not shown in Fig. 7·13.

7·6 Electrostatic deflection

In this method two pairs of deflection plates are attached to the electron gun, internally at the neck of the tube just before it begins to flare out. When a positive potential is applied to a deflection plate the negatively charged electrons in the beam are attracted toward the positive plate. If one plate is made negative with respect to the other, the electron beam will be repelled away from the negative plate. Therefore, a deflecting

Fig. 7·14 Electrostatic deflection.



voltage can be applied to the vertical pair of deflection plates to produce horizontal deflection of the electron beam, to either left or right. Deflecting voltage can also be applied to the horizontal plates to deflect the electron beam up or down toward the more positive plate. As a result, deflection voltages applied to the vertical and horizontal plates simultaneously deflect the beam both horizontally and vertically (see Fig. 7 · 14). While plate *A* is made more positive than *B* to deflect the beam upward, plate *C* can be made more positive than *D* for deflecting the beam to the left, moving the electron beam to the upper left corner in the illustration.

Electrostatic deflection is generally used for oscilloscope tubes with anode voltage of 2 to 5 kv, but not for picture tubes. Because of the high anode voltage too much deflection voltage would be necessary to fill the raster. As a typical value of *deflection factor*, 30 volts between the plates moves the beam 1 in. on the screen, with 1 kv anode potential. For 15 kv anode voltage and 20 in. deflection, the required sawtooth deflection amplitude would be $30 \times 15 \times 20$, which equals 9,000 volts peak to peak. The required deflection is obtained much more easily, by using magnetic deflection, with a shorter tube having a larger deflection angle.

7 · 7 *Magnetic deflection*

Two pairs of deflection coils are used as illustrated in Fig. 7 · 15, mounted externally around the neck of the tube just before the bell. The two coils in each set are usually in series with each other. The pair of coils above and below the beam axis produce horizontal deflection; the coils left and right of the beam deflect it vertically. This perpendicular displacement results because current in each coil has a magnetic field that reacts with the magnetic field of the electron beam to produce a force that deflects the electrons at right angles to both the beam axis and the deflection field.

To analyze the deflecting action, remember that the reaction between two parallel fields always exerts a force toward the weaker field. Consider the horizontal deflection coils first in Fig. 7 · 15. The windings are in a horizontal plane above and below the beam axis. Using the left-hand rule, the thumb points in the direction of the field inside a coil when the fingers curve in the direction of the electron flow around the coil. Therefore, the deflection field for the horizontal windings is downward. When the direction of the electron beam is into the paper, as indicated by the cross in the center, its magnetic field has lines of force counterclockwise around the beam in the plane of the paper. To the left of the beam axis, the magnetic field of the electron beam is down in the same direction as the deflecting field, while the fields are opposing on the right. The electron beam is deflected to the right, therefore, as the resultant force moves the beam toward the weaker field. In a similar manner, the vertical deflection coils deflect the electron beam downward. Deflecting current for both sets of coils can be provided simultaneously, deflecting the beam to the lower right corner of the screen.

Each of the deflecting coils is a flat rectangular winding which is then

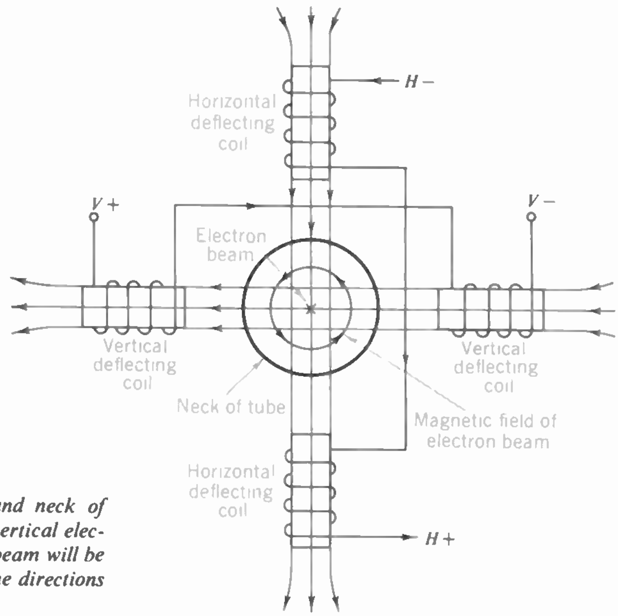


Fig. 7·15 Two pairs of coils around neck of cathode-ray tube for horizontal and vertical electromagnetic deflection. The electron beam will be deflected down and to the right for the directions of current shown here.

curved to fit the yoke form. This pancake type of winding is illustrated in Fig. 7·16. The entire yoke is in a housing that allows its magnetic field to penetrate the tube. In a typical yoke, the vertical deflection coils may have a total inductance of 40 mh and require 300 ma peak-to-peak sawtooth deflection current, while the horizontal deflection coils may have a total inductance of 13 mh and require 1,000 ma peak-to-peak current.

7·8 Characteristic curves of picture tubes

As shown in Fig. 7·17, the transfer characteristic curves of picture tubes are similar to the grid-plate characteristic curves for amplifier tubes. The kinescope control grid has a fixed negative bias and the video signal voltage varies the instantaneous grid voltage, varying the amount of beam current and the light from the screen. At cutoff, the grid voltage is negative enough to reduce the beam current to a value low enough to extinguish the beam, which corresponds to the black level. The parts of the screen without any luminescence look black in comparison with the adjacent white areas. The cutoff bias for kinescopes is about 50 volts, the exact value depending upon control grid–cathode spacing and G2 voltage.

As the a-c video signal swings in the positive direction to make the control grid less negative, the beam current increases. The slope of the steep part of the kinescope characteristic curve is approximately 2.2, providing a gamma of 2.2. It should be noted that the values of the grid voltages shown in Fig. 7·17 are positive signal volts applied in addition to the cutoff bias voltage. For a value of grid-driving voltage 30 volts positive

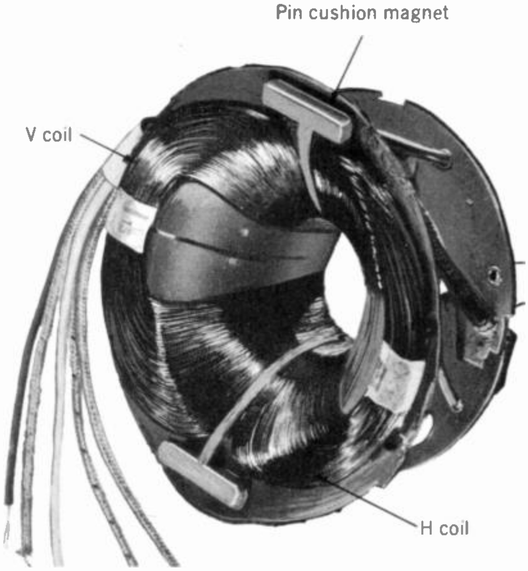
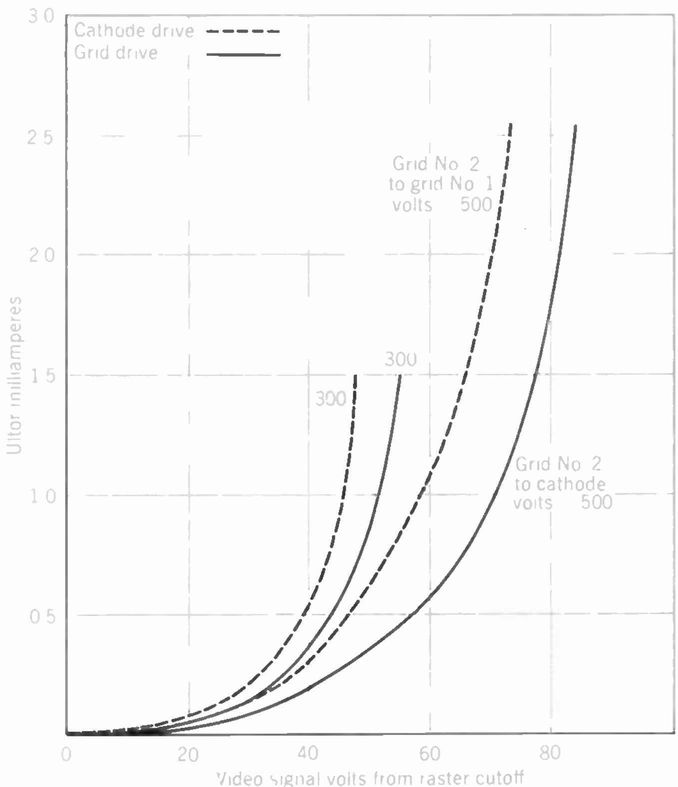


Fig. 7·16 Construction of deflection yoke for 110° deflection angle. (Stancor Electronics Corporation.)

with respect to a typical cutoff bias of −50 volts, the actual grid voltage becomes −20 volts. The corresponding beam current 30 volts from cutoff is 100 to 200 μa depending on accelerating grid G2 voltage. When the video signal voltage of varying amplitudes is applied to the control grid of the kinescope, the beam current and screen illumination change with the signal variations to reproduce the desired picture information. With a cutoff voltage of −50 volts, in order to use the operating range of the kine-

Fig. 7·17 Average grid-drive characteristics of 23CP4 kinescope. (RCA.)



scope for full contrast the amount of composite video signal required is approximately 70 volts peak-to-peak, including the sync-pulse amplitude.

Grid drive and cathode drive. The video signal can be applied either to the control grid of the kinescope or to the cathode. In either case, however, white in the video signal must make the instantaneous kinescope grid voltage less negative than the bias, producing more beam current and increased light for white parts of the picture. Cathode drive is used more often. This method provides greater variations of beam current and light output because the G2-cathode voltage is also video-modulated, increasing the modulation sensitivity of the electron gun. The difference in drive characteristics can be seen in Fig. 7 · 17, where the dashed curves are for grid drive. The higher curves are for higher G2 voltage.

Heater power. Most kinescopes require a heater voltage of 6.3 volts at 0.6 amp, for parallel heater circuits. For a series string, 8.4 volts at 0.45 amp is a typical heater rating for kinescopes; other heater voltages are 2.34 and 2.68 volts. If the heater does not light, there will be no beam current and no screen brightness. You can look through the glass neck to see if the heater is operating.

7 · 9 Picture tube precautions

The anode voltage for picture tubes is usually 12,000 volts or more. Therefore, the safety precautions important in working with high-voltage equipment should be observed. Before any part of the kinescope is touched, the power should be turned off and the high-voltage filter capacitor discharged, including the anode capacitance.

X rays are invisible radiation with wavelengths much shorter than visible light. Lead and leaded glass are used for shielding against X-ray penetration. Since they are produced by bombarding a metal target with high-velocity electrons, X rays can be produced in cathode-ray tubes. However, picture tubes may be operated with anode voltages up to 16 kv without producing harmful X-ray radiation and without danger of personal injury on prolonged exposure at close range. Most receivers meet the requirement for maximum radiation with a high safety factor. Above 16 kv, special X-ray shielding precautions may be necessary, especially for projection kinescopes.

Picture tubes should be handled more carefully than smaller vacuum tubes. The kinescope bulb encloses a high vacuum, and because of the large surface area, the envelope has a strong force on it produced by air pressure. For a typical 20-in. rectangular picture tube, the surface area is about 1,000 sq in. With an air pressure of approximately 15 lb per sq in. the total force on the bulb is 15,000 lb. The envelope is strong enough to withstand this force if there are no flaws and the tube is handled gently.

Dust on the glass faceplate of the kinescope can reduce the brightness appreciably. Dust particles are attracted by electrostatic induction, resulting from the high voltage. To form an airtight dust seal between the tube and the safety-glass window, a rubber strip is placed around the face of the kinescope where it fits into the mask on the front panel of the receiver cabinet. However, tubes with a bipanel faceplate do not need a dust seal.

7·10 Picture tube troubles

Typical troubles are low emission from the cathode, an open heater, or an internal short circuit from cathode grid or heater. An open heater means no emission and no brightness. However, the most common cause of no brightness is no high voltage from the flyback high-voltage supply. In old kinescopes, low emission combined with gas in the tube usually produces a distinctive weak picture that has shimmering silver gray instead of white in the highlights.

An internal short at the grid or cathode usually causes no control of brightness. The actual brightness may be too high or too low, but the picture contrast is weak and the brightness cannot be varied.

When a kinescope must be replaced either a new or rebuilt tube can be used. Rebuilt picture tubes cost less because they use the old envelope, but the electron gun and phosphor screen are new.

Weak emission in old kinescopes can often be improved temporarily by using a *tube brightener*, or booster. This is just a small filament transformer to step up the kinescope heater voltage, usually from 6.3 to 7.8 volts. Different types are available for parallel or series heater circuits, or the universal type in Fig. 7·18 can be used. The connections are simple, as the kinescope plug goes on the brightener and its plug on the tube. Some brighteners are autotransformers while others have an isolated secondary. It is useful to note that an isolated heater winding of the brightener can eliminate the effects of a heater-cathode short in the kinescope, as the ground connection of the heater is removed.

Continuing spot. This is a luminous spot that remains at the center of the kinescope screen just after the receiver is turned off. The afterglow results because anode voltage remains on the high-voltage filter capacitance while the cathode is still hot, after the deflection circuits stop operating. The anode filter capacitance can discharge only through the equivalent

Fig. 7·18 Kinescope tube brightener. Universal type for series or parallel heater circuits. (Perma-Power Company.)

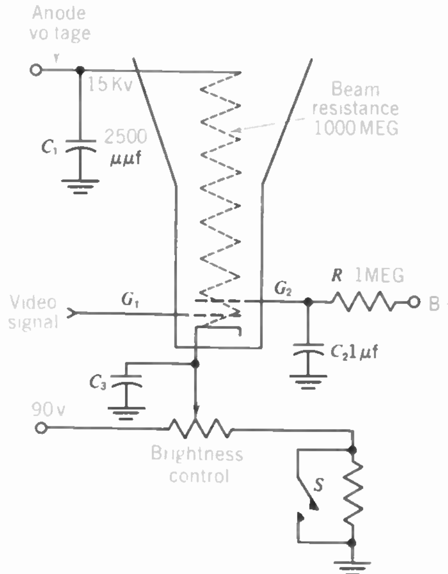


Fig. 7·19 Methods of discharging anode capacitance after receiver is turned off to eliminate continuing spot. Note switch S to short-circuit part of kinescope bias and RC filter to maintain G2 voltage.

resistance of beam current from cathode to anode in the kinescope. With 15 kv on the anode and 15 μa beam current, as examples, the equivalent beam resistance is 1,000 megohms as illustrated in Fig. 7·19. The time constant with 2,500 $\mu\mu$f anode capacitance equals 2.5 sec for the high-voltage filter.

The remedy for continuing spot is to allow faster discharge so that the anode voltage will not be enough to produce illumination when the receiver is off. More beam current means less beam resistance and a shorter time constant to allow faster discharge. This is why turning up the brightness control to reduce the kinescope bias, just before or after switching the receiver off, is able to eliminate the spot. The same result can be obtained automatically by using an extra pair of contacts on the on-off switch to reduce kinescope bias when the receiver is turned off. Another remedy is to provide a long-time-constant filter for the G2 supply voltage, as the accelerating voltage helps to produce enough beam current to discharge the anode capacitance. Both these methods are illustrated in Fig. 7·19.

Screen burn. If a spot or line is on the screen instead of the complete scanning raster, the brightness should be turned down to avoid burning the screen. Since the energy of the electron beam is concentrated in one small area without deflection, instead of being distributed over the entire screen, the excessive excitation can destroy the fluorescent properties of the phosphor. The result is a dark brown area on the burned part of the screen that cannot produce any light. The higher the anode voltage, the greater is the danger of screen burn.

Blooming. This term describes a picture that increases in size, with defocusing in white areas, as the brightness is increased. The blooming results from a sharp decrease in the amount of high voltage for the kinescope anode when the beam current increases. Usually, the trouble is a weak high-voltage rectifier having excessive internal resistance.

SUMMARY

Summary of picture tube setup adjustments

ADJUSTMENT	DESCRIPTION
Deflection yoke	Forward against bell for no shadowed corners. Set straight to eliminate tilted raster. If pincushion magnets attached, adjust for straight edges on raster. Wire spring on housing must contact outside coating on kinescope bell
Focus magnet	Move back and forth for sharp scanning lines in raster. Have kinescope neck centered and do not tilt magnet to prevent shadowed corner. Fine focus varied by rheostat on chassis to vary current in coil magnet, or thumbscrew to vary magnetic shunt on permanent magnet
Focus voltage	Potentiometer varies focus-grid voltage in tetrode gun; adjust for fine focus. With low-voltage focusing, connect wire from focus grid to one of the other pins for different focus voltages; this adjustment is not critical
Ion-trap magnet	Position over bend or diagonal cut in gun. Move back and forth and rotate slowly for maximum brightness. This adjustment is critical.
Electrical centering	Adjust rheostats on chassis to vary direct current in deflection coils of yoke. With electrostatic deflection, potentiometers vary d-c voltage of deflection plates
Mechanical centering	Move tab on centering ring magnet vertically and horizontally. The ring magnet may be alone, on focus magnet, or clamped to back cover of yoke. Tilting focus magnet also shifts centering. Yoke housing may include permanent magnets for centering
Pincushion magnets	Adjust for straight edges at top and bottom of raster

SELF-EXAMINATION *(Answers at back of book.)*

Choose (*a*), (*b*), (*c*), or (*d*).

1. The ion-trap magnet is adjusted for: (*a*) minimum brightness; (*b*) maximum brightness; (*c*) maximum deflection; (*d*) proper centering.
2. Which of the following kinescope types will not require an ion-trap magnet? (*a*) bent-gun; (*b*) slashed-field; (*c*) straight gun with aluminized screen; (*d*) magnetic deflection with electrostatic focus.
3. The raster is tilted with respect to the screen. To straighten the raster you can adjust the: (*a*) ion-trap magnet; (*b*) picture tube position; (*c*) focus magnet; (*d*) deflection yoke.
4. Current in the deflection coils for horizontal scanning provides magnetic field lines that are: (*a*) circular around the beam; (*b*) horizontal above and below the beam; (*c*) vertical left and right of the beam; (*d*) parallel to the beam.
5. The 23CP4 picture tube has the following characteristics: (*a*) 23-volt heater and 110° deflection angle; (*b*) 4-volt heater and white screen; (*c*) 23-in. screen, green phosphor; (*d*) 23-in. screen, white phosphor.
6. Which of the following is true? (*a*) The beam is attracted to the anode wall because of its

capacitance to the outside coating. (*b*) The screen can charge to the anode voltage by secondary emission from the phosphor. (*c*) The screen phosphor is a low-resistance conductor. (*d*) A medium-persistence screen has a decay time of 1 sec.

7. Which of the following is false? (*a*) A 90° tube is shorter than a 70° tube for the same screen size. (*b*) The heater pins are usually 1 and 12 on a duodecal base. (*c*) A bipanel faceplate does not require a dust shield. (*d*) An 8-pin kinescope base is wider than a duodecal base.
8. Refer to the graph in Fig. 7·17. With 300 volts on G2, cathode drive, and video signal 40 volts less negative than cutoff, the ultor current equals: (*a*) 0.35 ma; (*b*) 0.55 ma; (*c*) 1.0 ma; (*d*) 2.5 ma.
9. A kinescope brightener: (*a*) boosts anode voltage about 10 per cent; (*b*) boosts heater voltage about 15 per cent; (*c*) improves secondary emission from G2; (*d*) increases the anode capacitance.
10. Black in the reproduced picture corresponds to: (*a*) grid-cathode voltage equal to zero; (*b*) grid-cathode voltage negative at cutoff; (*c*) minimum anode voltage; (*d*) average beam current equal to 1 ma.

ESSAY QUESTIONS

1. Give the function of the electron gun, control grid, accelerating grid, anode wall coating, phosphor screen, external conductive coating.
2. Define: ultor, deflection factor, crossover point, electrostatic lens, magnetic lens, bent gun, pincushion magnet.
3. Describe briefly how to mount and adjust the deflection yoke and a PM focus magnet.
4. What is ion spot? How is it eliminated by an ion trap? By an aluminized screen?
5. Why is the P4 phosphor generally used for black-and-white picture tubes? What phosphor is used for tricolor kinescopes? For oscilloscope tubes?
6. What is meant by sticking potential of the phosphor screen?
7. Give two advantages of an aluminized screen.
8. Draw the base connections, bottom view, for a kinescope using magnetic deflection and focus with a duodecal base. Also for a tube using low-voltage electrostatic focus with an 8-pin base.
9. Give two precautions to keep in mind when installing a picture tube.
10. List the kinescope requirements for producing screen brightness.
11. What causes a continuing spot on the kinescope screen?
12. What is a brightener or booster for the kinescope?
13. How is the high voltage usually connected to the anode?
14. Why does the anode retain high voltage after the receiver is turned off?
15. Referring to the 23CP4 characteristics in Fig. 7·17, assume that 60-cycle sine-wave alternating voltage with a peak-to-peak value of 160 volts is applied with a bias of 80 volts. Show by a drawing the bar pattern that appears on the raster.
16. Which of the following can produce magnetic field lines parallel to the beam axis: (*a*) ion-trap magnet; (*b*) focus magnet; (*c*) vertical deflection coils; (*d*) horizontal deflection coils?

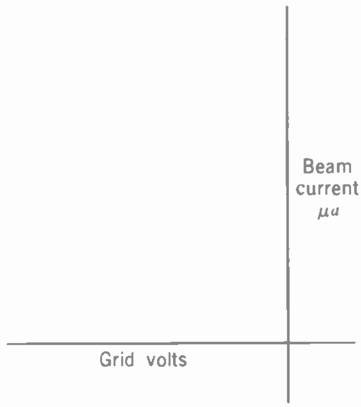
PROBLEMS *(Answers to odd-numbered problems at back of book.)*

1. The raster has all four corners shadowed. How would you correct this?
2. The raster has one corner shadowed. How would you correct this?
3. The raster is tilted. How would you straighten it?
4. The raster is off to one side, leaving a wide black bar at the left edge of the screen. What would you adjust?
5. The raster has top and bottom edges bowed in. What would you adjust?
6. A kinescope has a deflection angle of 90°. How many degrees off center should the beam

be deflected by maximum scanning current? What is the advantage of this tube compared with a 53° tube?

7. Give the approximate d-c voltage with respect to cathode for each of the following for a typical picture tube: (*a*) control grid; (*b*) accelerating grid; (*c*) anode; (*d*) external coating; (*e*) focus grid in tetrode gun; (*f*) focus grid for low-voltage focusing.
8. What is the probable cause of the following troubles? (*a*) Kinescope 4 years old has little brightness and weak contrast. (*b*) Screen suddenly becomes black with no brightness. Kinescope heater lights. (*c*) Screen bright but cannot be made darker. Bias voltage varies as brightness control is turned, with kinescope socket disconnected. (*d*) Picture blooms out of focus and screen becomes black when brightness control is turned up for maximum brightness.
9. Referring to the 23CP4 characteristics in Fig. 7·20: (*a*) What are the values of ultor current with grid-cathode voltage 40 volts from cutoff for each of the four curves shown? (*b*) Which of the types of operation shown allows the most video signal driving voltage for maximum contrast?
10. For the values of grid voltage and beam current listed below, draw the corresponding kinescope grid-drive characteristic curve.

Grid volts	Beam current, μa
−50	0
−40	50
−30	100
−20	250
−10	500
0	800

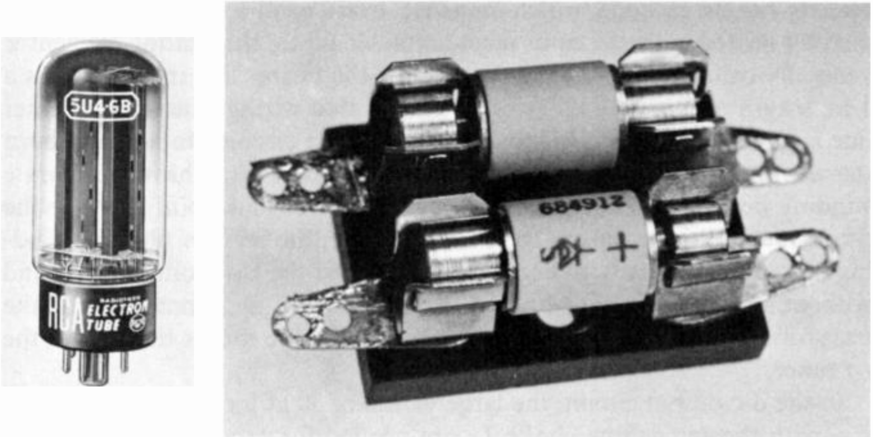


Chapter 8 Power supplies

The television receiver has a low-voltage supply for B+ voltage and a high-voltage supply for the kinescope anode voltage. Typical B+ output is 250 to 350 volts at approximately 200 ma load current for the plate and screen circuits of the amplifier tubes. Generally, either a vacuum tube such as the 5U4 is used as a full-wave rectifier, or two semiconductor diodes are connected in a voltage doubler circuit. Silicon, selenium, or germanium diodes can be used. Figure 8 · 1 shows typical rectifiers. The high-voltage supply uses a half-wave rectifier to produce 15 to 20 kv anode potential for direct-view picture tubes, with a current drain less than 1 ma. Because of the different power requirements, a common supply is not practicable.

8 · 1 Full-wave rectifier

The B+ supply in Fig. 8 · 2 can use a full-wave rectifier because of the center-tapped secondary L_2 on the a-c power transformer T_1. This circuit is generally called an a-c power supply since T_1 operates only with a-c input. The 120 volts across L_1 is stepped up to 600 volts across L_2, but stepped down for heater voltages across L_3 and L_4. The a-c input voltage to each 5U4 diode is 300 volts, from either side of L_2 to the grounded center tap. When each plate is driven positive on alternate half cycles the diodes conduct to produce d-c output. The input filter capacitor C_{2A} charges to 320 volts. With 200 ma load current, this d-c voltage output is slightly more than the rms value of 300 volts but less than the peak a-c input. The filtered output across C_{2B} is 20 volts less because of the IR drop with 0.2 amp across the 100-ohm resistance of the filter choke L_5. This E_1 supply of 300 volts is B+ for the power amplifiers, such as the audio, vertical, and horizontal output stages. The lower supply voltage of 140 volts (E_2) is for miniature glass tubes used in the r-f and i-f amplifiers. Note that R_2 has a voltage drop of 160 volts, equal to 3,200 ohms $\times$



(*a*) (*b*)

Fig. 8·1 Typical rectifiers for low-voltage supply. (a) 5U4 vacuum-tube full-wave rectifier. (b) Silicon diodes; width about ½ in. Metal case is cathode connection. (Sarkes-Tarzian, Inc.)

0.05 amp, leaving 140 volts for E_2. Both E_1 and E_2 are d-c output voltages measured with respect to chassis ground.

In the a-c input circuit, C_1 is a bypass across the primary to filter out r-f interference from the power line. R_1 provides a high-resistance path from chassis to earth ground, through the power line, to discharge any electrostatic charge that may accumulate on the chassis. The safety interlock plug S_1 on the back panel of the receiver must be closed to supply a-c power input when the on-off switch S_2 is closed. Usually S_2 is on the volume control. Instead of a rotary switch, S_2 is often the push-pull type so that the volume setting need not change when turning power on or off. F_1 is a slow-blow type of fuse to protect the primary of the power transformer against excessive current if there is a short circuit in the power supply.

Included in the power transformer are step-down secondary windings for the filaments and heaters of all tubes in the receiver. A filament is a

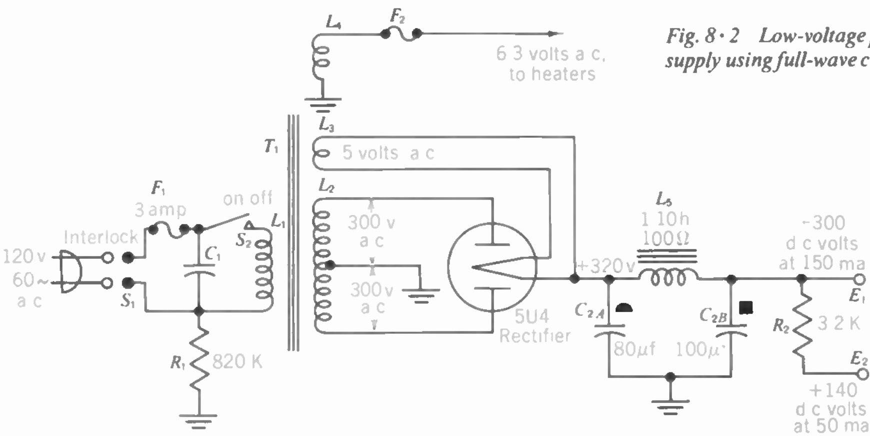


Fig. 8·2 Low-voltage power supply using full-wave circuit.

directly heated cathode, which emits electrons, as in the 5U4. With an indirectly heated cathode, as in most amplifier tubes, the heating element is generally called a *heater.* The fuse F_2 for the heater line in Fig. 8·2 is a 1-in. length of No. 24 gage wire. When printed wiring is used, the heater line is generally fused this way because a short circuit can actually burn the wiring board. Note that the rectifier filament must have a separate winding because the B+ voltage is present here. This point is where the d-c output voltage is taken from. Furthermore, the rectifier filament winding must be insulated well enough to withstand the B+ voltage to ground without arcing. It should be noted that in receivers for remote control, the transformer may include a 24-volt winding for the motor that drives the r-f tuner.

In the d-c output circuit, the large values of 80 μf for C_{2A} and 100 μf for C_{2B} with the smoothing choke L_5 are needed for excellent filtering. The square and half-circle markings for C_{2A} and C_{2B} in Fig. 8·2 indicate the corresponding sections on the electrolytic filter (see Fig. 8·3). C_{2A} is the input filter capacitor for high d-c output voltage and good regulation. However, the input filter cannot have too much capacitance, or peak charging current will damage the rectifier when it conducts. The output filter capacitor C_{2B} provides additional filtering in bypassing a-c ripple around the d-c load circuit. This is a capacitor input filter because rectifier current that charges C_{2A} does not flow through the choke.

The a-c ripple is only 1 volt rms, or less, which is less than ½ per cent of the d-c output. Good filtering is necessary because excessive hum voltage in the raster and signal circuits can produce dark and light bars or bend in the reproduced picture. Also, low video frequencies from 30 to 100 cps must be amplified without mutual coupling between stages through the B supply filter as a common impedance. Finally, the relatively large d-c load current, compared with a radio receiver, corresponds to a low value of load resistance that requires more filter capacitance.

Fig. 8·3 Electrolytic filter capacitor with multiple sections. (Sprague Electric Co.)

8·2 D-C voltage polarities

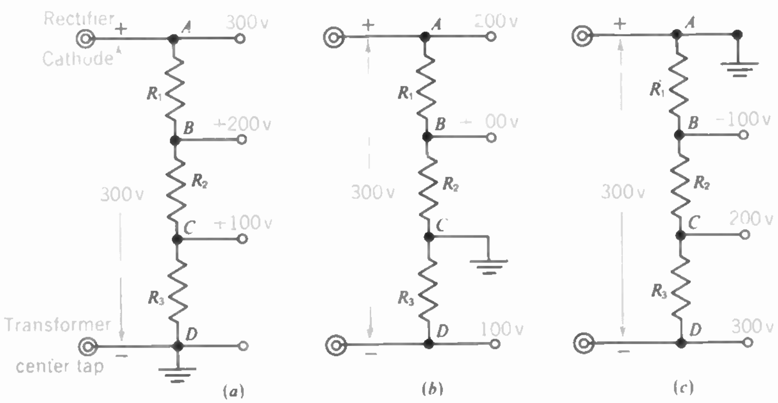
Although chassis ground is generally used as a voltage reference, this is not necessarily the most negative point of the power supply. Fundamentally, the most negative potential is at the center tap on the transformer secondary for a full-wave rectifier, or any other point in the a-c input circuit that returns to the rectifier cathode. This idea can be illustrated by assuming a voltage divider across the d-c output, as in Fig. 8·4. In *a* the transformer center tap is connected to chassis ground at *D*. Then all other points on the divider are positive to ground because *D* is connected to the transformer center tap. Note that in *b* point *C* is grounded. However, *D* is still the most negative point. Now *D* is 100 volts negative to the chassis ground reference. This arrangement can be used to obtain negative voltage for grid bias and other uses. Finally, in *c* the top of the divider is grounded at *A*. Then all other points on the divider are negative with respect to ground, because the most positive point is connected to chassis ground. This can be done as long as no other point is grounded. Actually, this arrangement is an *inverted power supply,* for negative d-c output voltage to ground. Here the d-c load is in the plate-to-ground circuit, instead of the cathode circuit. The rectifier conducts when a-c input voltage drives the cathode negative, with respect to chassis ground.

Note the arbitrary nature of assigning polarity. The power supply will operate in exactly the same way if there are no ground connections. Grounding one point in the supply only provides output voltage with the desired polarity to chassis ground, which is a good reference because it is a convenient return connection for the amplifier stages on the chassis. For the different arrangements in Fig. 8·4, point *A* will always be more positive than *B*, *C*, or *D*; also *D* will always be the most negative point in the supply, regardless of where the ground connection is made. Furthermore, the total d-c output voltage between *A* and *D* equals 300 volts in all three cases.

8·3 Heater circuits

With a transformer to step down the a-c line voltage from 120 to 6.3 volts, all the heaters are connected in parallel across the heater winding. Then all the parallel heaters have the same voltage rating, although the in-

Fig. 8·4 Effect of moving ground point on voltage divider in d-c output of power supply. See text for explanation.



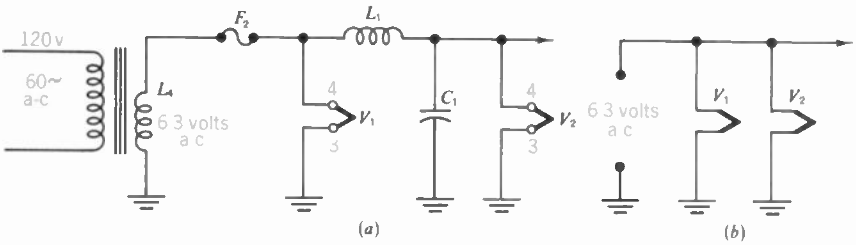


Fig. 8·5 Parallel connections for 6.3-volt heaters. (a) With r-f decoupling chokes. (b) Equivalent circuit for 60 cps.

dividual branch currents can be different. In an a-c/d-c type of supply, without a power transformer, all the heaters are connected in series to the 120-volt a-c power line. Then the current must be the same in all parts of the series string, while the different voltage drops add to equal 120 volts. If one heater is open there is no current in the entire series string.

Parallel heaters. In Fig. 8·5, one end of L_4 is tied to chassis ground to save wiring. The opposite end is the high side connecting to all heaters in parallel. Each heater returns to L_4 by an individual ground connection. The coil L_1 is a few turns of heavy wire to serve as a filament choke for r-f signals. With its r-f bypass C_1, the choke L_1 provides a decoupling filter to prevent feedback of r-f or i-f signal through the common heater line. The choke has high impedance for r-f signal but practically none for the 60-cps filament current. Therefore, the heaters are effectively in parallel, as in *b*. Such r-f decoupling filters are also commonly used in series heater circuits.

Series heaters. In Fig. 8·6 all the heaters, including the picture tube, are in series with each other across the 120-volt a-c power line. The current is 600 ma in all these series heaters. The sum of the individual heater voltages equals 99.5 volts. R_1 and R_2 in series provide an IR drop of 20.4 volts, including 3.6 volts across the 6 ohms of R_1 and 16.8 volts across R_2. Therefore, all the series IR drops total 120 volts, equal to the line voltage. The series heater string can operate with either a-c input or d-c input.

The thermistor R_1 is a compensating resistor with a negative temperature coefficient. Its resistance decreases at higher temperatures. This is opposite from the characteristic of the tungsten heaters, which increase their resistance with temperature. When the receiver is first turned on R_1 is 200 ohms to limit the current while the heaters are cold and have low resistance. After warmup, R_1 decreases to 6 ohms, allowing normal operation of the heaters. The thermistor prevents excessive heater current until the tubes reach operating temperature, therefore serving as a "tube saver."

Tubes for a series string should have heaters with a controlled warmup time so that all heaters reach operating temperature at the same time. Otherwise the cold heaters have too little resistance, allowing excessive IR voltage drop across the hot tubes when the set is turned on. The uneven warmup time can cause repeated troubles of a burned-out heater in the series string. To eliminate this problem, tubes for series strings using 600 or 450 ma heater current are designed to reach 80 per cent of rated heater voltage in 11 sec, with specified test conditions. In actual operation, typical warmup time for a series string without a thermistor is about 50 sec, for a full raster. This is appreciably longer than the warmup time for parallel heaters. With a thermistor in the series string, warmup time is about 2 min.

Note the order of connections in Fig. 8 · 6. The series heaters have the same current but a point closer to the grounded end of the string has a smaller potential difference to chassis ground for 60-cycle hum voltage. At the opposite extreme, you can measure 99.5 volts of the a-c line input from one side of the 12AX4 heater to ground, since this tube is first in the series string. A higher 60-cycle voltage to ground allows more 60-cps hum to be coupled into the signal circuits by heater-cathode leakage in the tube. Therefore, the tubes most susceptible to hum are connected closer to the grounded end of the series string.

Some tubes have a center-tapped heater with two sections that can be wired either in series for 12.6-volt operation or in parallel for 6.3 volts (see Fig. 8 · 7). The 12AU7, 12AT7, and 12BY7 are examples of such tubes with three heater pins.

8 · 4 Voltage doublers

A voltage doubler uses two diodes to charge two capacitors on alternate half cycles in a circuit arranged to produce d-c output double the a-c in-

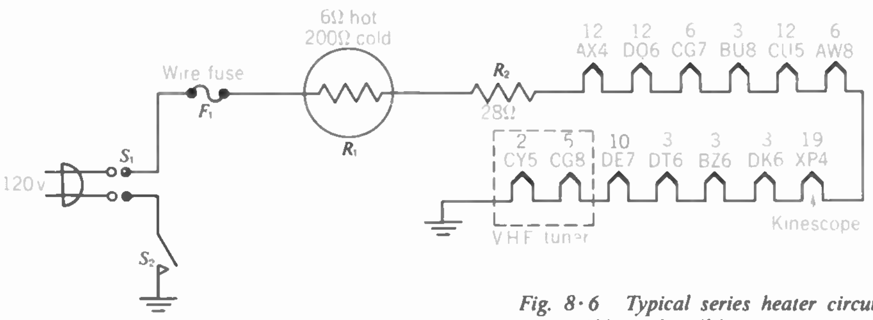


Fig. 8 · 6 Typical series heater circuit. Current rating 600 ma for all heaters in string.

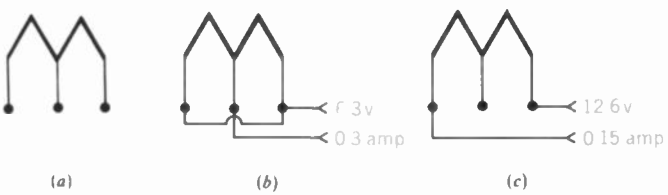


Fig. 8 · 7 (a) Heater with two sections. (b) Parallel connection. (c) Series connection.

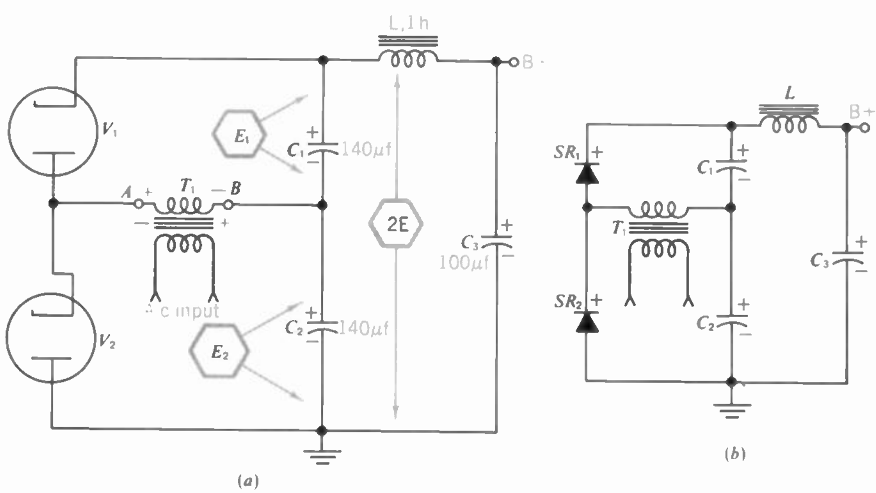
put voltage. Since each capacitor charges to peak value, without a load the d-c output can be twice the peak input. As an example, for a-c input of 120 volts rms, or 168 volts peak, the d-c output voltage can be 336 volts. With typical load current in a B+ supply, however, the d-c output voltage is about 270 volts. The doubler circuit is often used for a low-voltage power supply with selenium or silicon diodes, because their peak inverse voltage rating of 400 to 500 volts is low compared with vacuum tubes. The peak inverse voltage is across the diode when it does not conduct. This voltage equals the d-c output at the cathode plus the peak a-c input.

These circuits are often used without a power transformer, but a-c voltage input is necessary for doubler operation. A doubler circuit is seldom needed in the high-voltage power supply, as the horizontal output current for wide-angle picture tubes can generate the required amount of flyback high voltage.

Full-wave doubler. The circuit in Fig. 8·8 is a full-wave doubler with a ripple frequency of 120 cps for 60-cycle input. C_1 and C_2 are charged separately on alternate half cycles of the a-c input but they discharge in series through the d-c load so that the two capacitor voltages are added in the output. Operation of the circuit in *a* can be analyzed as follows:

1. When the a-c input makes point A and the V_1 plate positive this diode conducts to charge C_1, with the cathode side plus. The charging path is through V_1 and the secondary winding of the power transformer T_1.
2. On the next half cycle, the a-c input makes point A and the V_2 cathode negative. Since negative cathode voltage corresponds to positive plate voltage, V_2 now conducts to charge C_2. The terminal of C_2 closer to the cathode of V_2 is the plus side.

Fig. 8·8 Full-wave voltage doubler circuit. (a) With tubes. (b) With selenium diodes.



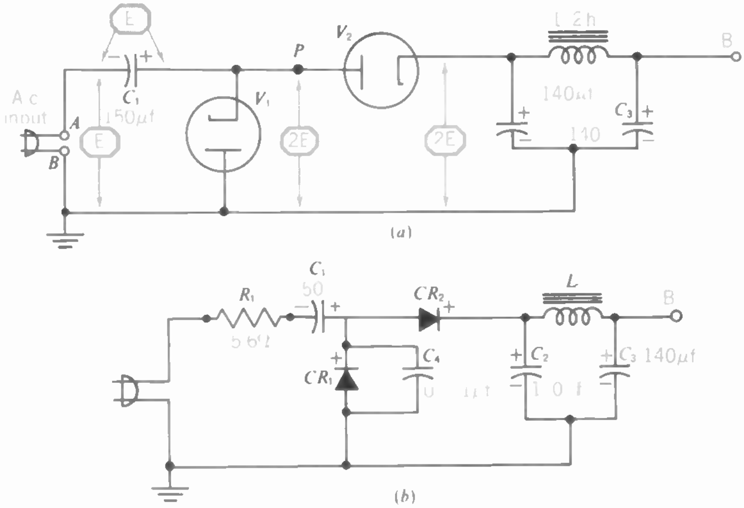


Fig. 8·9 Half-wave or cascade voltage doubler circuit. (a) With tubes. (b) With silicon diodes.

3. The d-c output voltage from the plus side of C_1 to chassis ground equals the sum of E_{C_1} and E_{C_2}, as they are series-aiding, similar to two batteries in series. The result is d-c output voltage for B+ approximately double the a-c input voltage.

This circuit is a full-wave rectifier because on both half cycles power is supplied to the output by each rectifier and its filter capacitor. The smoothing choke L and output filter capacitor C_3 provide additional filtering of the d-c output. A disadvantage of this circuit is the fact that the a-c input and d-c output cannot have a common ground, as a ground at A or B would short-circuit the voltage across C_2.

The circuit in Fig. 8·8*b* operates the same as in *a* with selenium rectifiers SR_1 and SR_2 instead of V_1 and V_2. Silicon or germanium rectifiers can be used also. In any case, the power transformer T_1 is convenient for isolating the a-c input from ground. Its turns ratio may be 1.5:1 to step up the secondary voltage for more B+ without exceeding the peak inverse ratings.

Note the rectifier symbol in Fig. 8·8*b*. The arrowhead shows the direction of forward or easy current through the NP semiconductor junction, which is opposite to the direction of electron flow through a vacuum tube. The terminal marked + or coded red is where B+ is taken from, corresponding to the cathode of a vacuum tube.

Half-wave doubler. This circuit shown in Fig. 8·9 is also called a *cascade* doubler. As a half-wave rectifier, its ripple frequency is 60 cps for 60-cycle input. This doubler has the advantage of a common ground for one side of the a-c input and d-c output, which is convenient for operation without a power transformer. Referring to the circuit in *a* with vacuum tubes:

1. On the half cycle of a-c input that makes point A and the V_1 cathode negative it conducts to charge C_1. Without any load current we can assume C_1 charges to approximately 150 volts (E_{C_1}), close to the peak

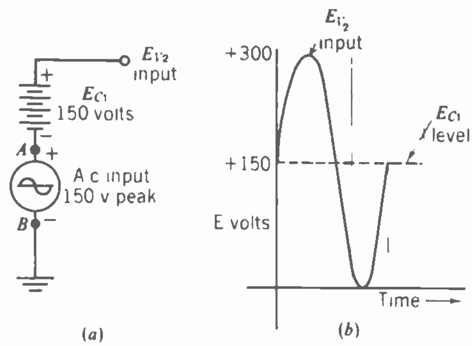


Fig. 8·10 Doubled voltage input to second diode V_2 or CR_2 in Fig. 8·9. (a) D-C voltage across C_1 in series-aiding polarity on positive alternation of a-c input. (b) Graph of a-c input shifted to positive d-c level of E_{c2}.

input. E_{C_1} is a d-c voltage because charging current flows in just one direction only when V_1 conducts. Although the V_1 diode is inverted, the C_1 terminal at the V_1 cathode is still the positive side of E_{C_1}. At this time V_2 cannot conduct because its plate is negative.

2. On the next half cycle the a-c input voltage makes point A and the V_2 plate positive and V_2 conducts.
3. The input voltage for the V_2 plate is approximately double the a-c line voltage. At P, the voltage to ground consists of E_{C_1} in series with the a-c input. This is illustrated in Fig. 8·10a. When the input is plus, this polarity makes the two voltages series-aiding. The result of adding these d-c and a-c voltages is shown in b. Notice the a-c input voltage is shifted to the axis of E_{C_1}. The effect of inserting this d-c level is that the varying voltage now has a positive peak of 300 volts, equal to 150 volts peak above the 150-volt axis, instead of 150 volts above the zero axis of the a-c input.
4. With doubled voltage input ($2E$), when the diode V_2 conducts it charges C_2 to produce d-c output voltage equal to $2E$, approximately.

This circuit is a half-wave rectifier because power is supplied to the output only on the half cycle that makes V_2 conduct. At this time C_1 can discharge through V_2, the a-c input circuit, and the load, which reduces E_{C_1} slightly. However, C_1 can charge much faster through the low resistance of V_1 when it conducts, compared with discharge through V_2 and the load resistance, to maintain E_{C_1} at a voltage slightly less than the peak of the a-c input. Then E_{C_1} can shift the peak of E_{V_2} to twice the input voltage to double the d-c output voltage across C_2.

The circuit in Fig. 8·9b operates the same as in a with silicon diodes CR_1 and CR_2 instead of V_1 and V_2. Selenium rectifiers can be used also. Note the series resistor R_1 to limit peak current through the semiconductor rectifiers, which have very low resistance in the forward direction. R_1 is often a fusible resistor, which opens with excessive current to act as a fuse, with snap-in contacts for easy replacement. Such a protective resistor is especially important without the resistance of a power transformer to limit the current. The small bypass capacitor C_4 across CR_1 protects the silicon rectifier against high-voltage transients produced when the power is turned off or on. In addition, C_4 filters out harmonics of the rectified current, which can cause interference in the signal circuits.

8·5 Transformerless low-voltage power supply

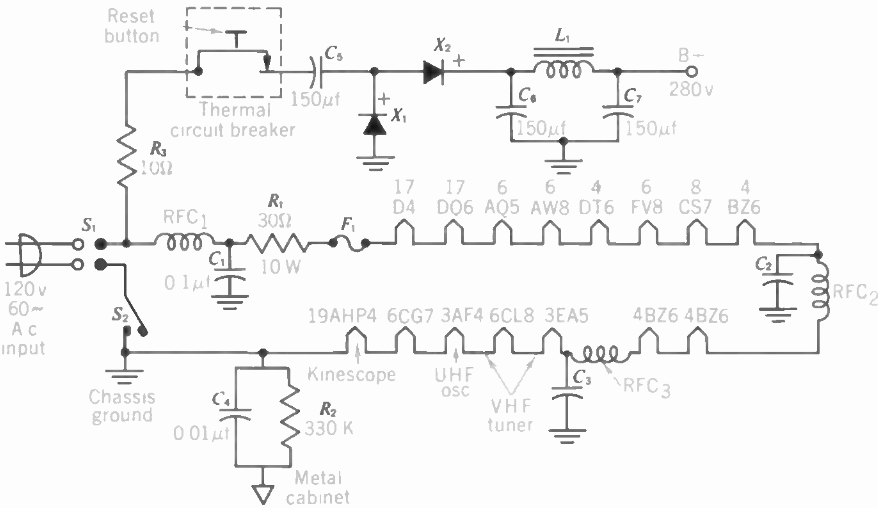
The circuit in Fig. 8·11 is generally called an a-c/d-c power supply because it has no power transformer. The series heater string, similar to Fig. 8·6, will operate with either d-c power or a-c input. However, a-c input is necessary for the doubler to produce 280 volts output. This is a cascade doubler with silicon diodes X_1 and X_2, similar to Fig. 8·9*b*. If d-c power were applied then only X_2 would conduct on one polarity of the d-c input. Note the thermal cutout, or circuit breaker, used in the B+ supply so that heat caused by excessive current will open the circuit.

Shock hazard. In a transformerless supply the B− line must return to one side of the a-c power line, as the source of a-c input. Therefore, when the chassis is used as the B− line, the chassis is connected to one side of the power line. By touching the chassis and the opposite side of the line, you are connected across the 120-volt a-c power input. Or, if two chassis plugged into opposite sides of the power line should touch, this produces a short circuit across the line. To eliminate these dangers when working on a transformerless chassis, an isolating transformer should be used. This is a power transformer with a 1:1 turns ratio for the isolated secondary. The isolation transformer is plugged into the line instead of the receiver, which connects to the transformer secondary.

8·6 Stacked B+ circuits

Ordinarily the plate-to-cathode circuit for each amplifier tube connected to B+ is a separate branch in parallel across the B+ supply. This method

Fig. 8·11 Transformerless power supply, with series heater string and cascade doubler for B^+ voltage.



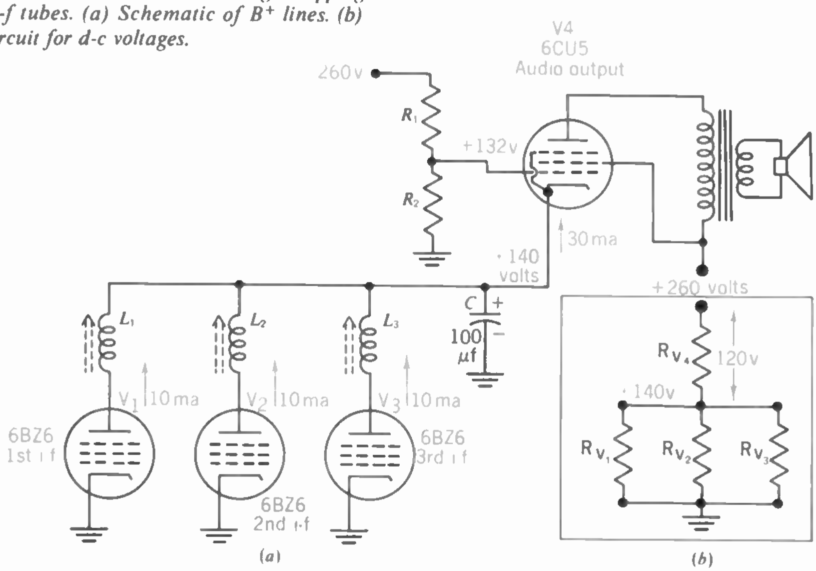
is satisfactory when all the tubes use approximately the same supply voltage. In a television receiver the i-f and r-f tubes usually need about +135 volts supply, however, while the power stages require full B+ voltage, which we can take as 260 volts. In this case, it is convenient to connect some tubes in series for B+ voltage, in a stacked B+ arrangement. Figure 8 · 12*a* shows a common circuit, with the equivalent tube resistances in *b*.

The three i-f tubes in Fig. 8 · 12 are in parallel and this group is in series with the audio output tube. Therefore, the B+ voltage is divided in proportion to the series resistances. Then V_4 has 120 volts from plate to cathode while the parallel bank has 140 volts, as proportional parts of the 260 volts of B+ supply. We can now consider the cathode of V_4 as a source for +140 volts. The bypass C keeps audio signal out of the 140-volt supply line for the i-f tubes. Note that the control grid of V_4 must be at +132 volts with respect to chassis ground, or 8 volts less than the cathode, for a typical grid bias of −8 volts.

Since the current must be equal in series components, this combination is good because the three parallel branch currents total 30 ma, equal to the required current through V_4. All these values are for direct current. The alternating signal current flows in a separate branch for each stage by means of appropriate bypass capacitors. This stacked B+ circuit is also called a *shelf-type supply.*

With stacked B+ as in Fig. 8 · 12, it is important to remember that the audio output tube determines the plate voltage of the i-f amplifiers for picture signal. If V_4 is open, there will be no B+ voltage for the i-f stages. When the audio output tube is weak, its high resistance causes lower plate voltage for the i-f amplifiers, resulting in weak picture and poor sync.

Fig. 8 · 12 Stacked B+ arrangement with audio output tube plate-cathode circuit used as voltage-dropping resistor for i-f tubes. (a) Schematic of B+ lines. (b) Equivalent circuit for d-c voltages.



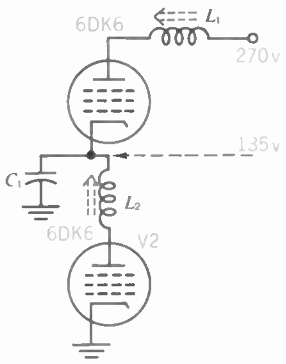


Fig. 8 · 13 Two i-f tubes in series with each other across the B^+ supply.

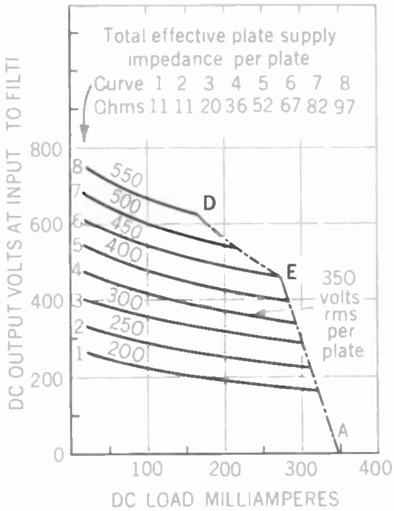


Fig. 8 · 14 Operation characteristics of 5U4 in 60-cps full-wave rectifier circuit. Input filter capacitor 40 μf. (From RCA Tube Manual.)

Another method of stacking tubes in series for B+ voltage is shown in Fig. 8 · 13. Here two i-f tubes with the same plate current are in series with each other across the B+ supply of 270 volts. Each tube then has the potential difference of +135 volts from plate to cathode.

Stacked B+ circuits can be used with either an a-c power supply or the a-c/d-c type, since the stacking affects only the plate-supply branches connected to B+. The advantage is less load current in the B+ supply, with fewer parallel branches.

8 · 7 Rectifier ratings

The chart in Fig. 8 · 14 shows 5U4 operation characteristics in a full-wave rectifier circuit with capacitor input filter as in Fig. 8 · 2. For any one curve the graph shows d-c voltage available at the output for different values of d-c load. Output voltage decreases with more load current.

Each one of the family of curves is for a specified amount of a-c input voltage. If we use curve 3 as an example, for 300 volts a-c input across each half of the high-voltage secondary in the power transformer, the d-c output is approximately 320 volts with a load current of 200 ma. This value assumes 20 ohms source impedance of the a-c input circuit, normally provided by the transformer windings, to limit peak plate current when each diode charges the input filter capacitor. Note that the boundary points *DEA* for all the curves indicate limits not to exceed maximum ratings of the tube. If a choke input filter is used the peak current ratings will be slightly higher because the choke inductance in series with the input filter capacitor limits its charging current.

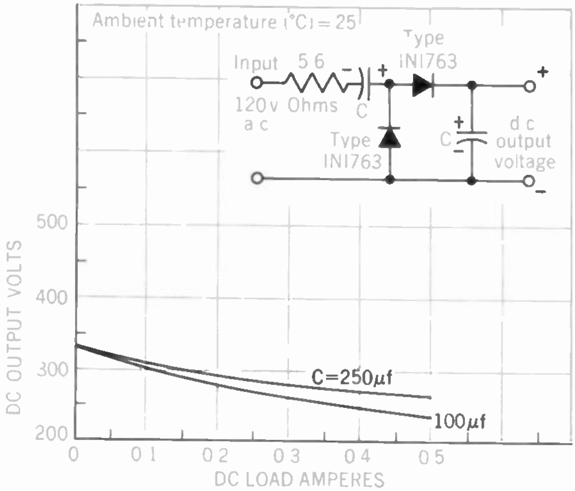


Fig. 8·15 Operation characteristics of type 1N1763 silicon diode in 60-cps half-wave voltage doubler circuit. (From RCA Technical Data.)

For silicon diode rectifiers, Fig. 8·15 shows operation characteristics in the typical application of a doubler circuit. This is a half-wave doubler operating directly from the 120-volt a-c line without a power transformer. The 5.6-ohm surge resistor provides the source impedance to limit peak charging current. If we use 100 μf for C, the d-c output voltage is about 280 volts with a load current of 200 ma. For a full-wave doubler, d-c output voltage is approximately 15 volts higher.

The comparison of a typical silicon diode with the 5U4 rectifier in Table 8·1 shows that in general the semiconductor can conduct more d-c load current because of its lower internal resistance. This means a very small internal voltage drop. As a result, the silicon diode can provide about 40 volts more d-c output with a load current of 200 ma.

The vacuum-tube rectifier can have more a-c input voltage because of its higher *peak inverse voltage rating.* This is the maximum voltage that can be applied in reverse polarity (rectifier not conducting) without internal arcing. The peak inverse voltage actually applied equals the sum of the

Table 8·1 Comparison of vacuum-tube and silicon diode rectifiers

Maximum rating for capacitor input filter	*Full-wave vacuum rectifier* 5U4GB *(each plate)*	*Silicon diode type* 1N1763
Rms a-c input volts	450–550	400
D-c output, ma	160–270	500
Peak plate current, ma	1,000	5,000
A-c source resistance, ohms	11–97	5.6–6.8
Peak inverse volts	1,550	400*
Internal forward resistance, ohms	196	0.08
Internal reverse resistance, ohms	Infinite	400,000

* At ambient temperature up to 100°C.

peak a-c input plus the d-c voltage across the input filter capacitor. At this time these two voltages are series-aiding from plate to cathode of the rectifier. As an example, in Fig. 8 · 2 the peak inverse voltage applied to the 5U4 diodes when each plate is negative equals 424 plus 320, or 744 volts.

8 · 8 High-voltage power supplies

The requirements include a source of high-voltage a-c input, usually at about 15 kv, the high-voltage rectifier, and a filter for the rectified d-c output voltage for the kinescope anode. Instead of using stepped-up 60-cycle voltage, the receiver generates its own a-c source of high voltage at a frequency much higher than 60 cps. With an r-f type of supply, a power oscillator operating at about 200 kc supplies a-c input to the rectifier. More common is the flyback type of supply, where high-voltage a-c input is obtained from the horizontal deflection circuits. During horizontal retrace, high-voltage flyback pulses are produced by the output transformer at the repetition rate of 15,750 cps.

These circuits have been developed because high-voltage power supplies operating directly from the 60-cps power line have definite disadvantages. Most important is the danger of serious electrical shock since the 60-cps line can supply a large amount of power. In addition, a 60-cps high-voltage supply is bulky and costly because of the large transformers and filter capacitors needed. These disadvantages are eliminated to a great extent when smaller components are used and the amount of power output is limited.

Flyback high-voltage supply. Practically all television receivers use this circuit because the required high-voltage a-c input for the rectifier is produced by the horizontal output transformer. Details of the horizontal output circuit are explained in Chap. 18 but Fig. 8 · 16 shows the basic requirements for flyback high voltage. Note the plate current of the 6DQ6 increasing from zero at *A* to 300 ma at *B* during horizontal trace time. Then the cur-

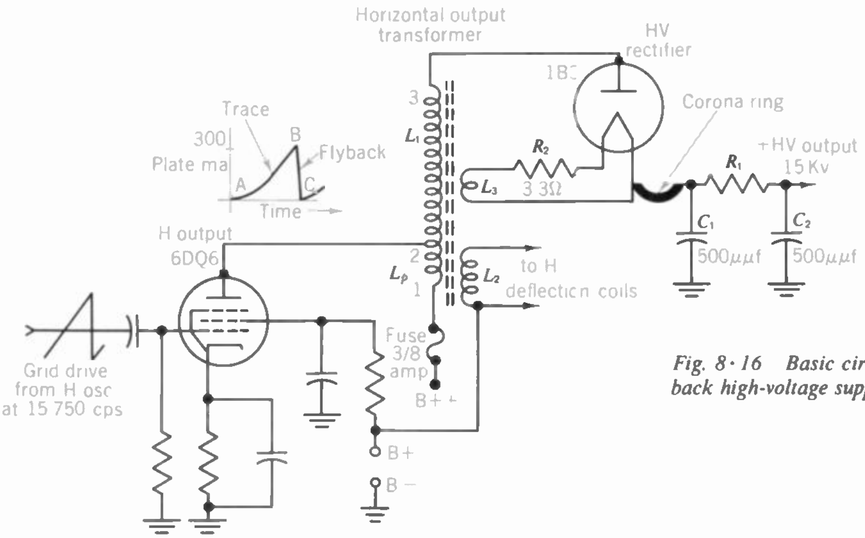


Fig. 8 · 16 Basic circuit of flyback high-voltage supply.

rent decreases sharply to zero for a fast flyback from B to C. The same cycle is repeated for every horizontal line with grid drive from the horizontal scanning oscillator at 15,750 cps.

The plate current flows through the primary winding L_p between terminals 2 and 1 on the horizontal output transformer. While the current increases, its associated magnetic field is expanding. Then plate current is suddenly cut off by flyback on the sawtooth grid voltage, driving the grid far negative. Now the magnetic field collapses very fast, inducing a high self-induced voltage across L_p. The polarity of this self-induced voltage is positive to oppose the decrease in plate current.

As a typical value, the peak amplitude of this flyback voltage at the plate of the horizontal output tube is 6,000 volts. This voltage is stepped up by the high-voltage winding between terminals 2 and 3 on the transformer. Assuming a 3:1 voltage step-up, 18 kv is available for the high-voltage rectifier. This a-c voltage is applied between plate and ground of the 1B3. When the rectifier plate is made positive it conducts to charge the input filter capacitor C_1 to the peak value of the input voltage. The result is approximately 18 kv d-c output voltage for the kinescope anode. The 500-$\mu\mu$f filter has enough capacitance because of the relatively high ripple frequency. Figure 8 · 17 shows a typical high-voltage filter capacitor, with a voltage breakdown rating of 20 kv. The output filter capacitance C_2 is usually the anode capacitance in the picture tube. The filter resistor R_1 can be used instead of a choke because the load current is less than 1 ma.

Figure 8 · 18 shows the main components of a flyback supply and the high-voltage cage. In addition to protection, the metal cage is a dust cover and shield to minimize radiation to the receiver signal circuits. The fuse for the horizontal output transformer and high-voltage supply, with a rating of ¼ or ⅜ amp, is generally in or near the high-voltage cage.

High-voltage rectifier. The tubes commonly used are 1B3, 1G3, and 1K3. These have essentially the same ratings of 30,000 volts maximum peak inverse voltage up to 300 kc, 1.25 volts at 0.2 amp for the heater, and direct load current of 0.5 ma or less as a pulsed rectifier. Because of the small power required, heater voltage for the 1B3 can be taken from the horizontal output transformer by the filament winding L_3 in Fig. 8 · 16. This is only one turn of heavy wire on the transformer core. This convenience eliminates the need for a separate filament transformer, which would require enough insulation to withstand the high d-c voltage at the rectifier filament without arcing to ground. Note that R_2 drops the heater voltage to 1.25 volts. This may be a carbon resistor, or resistance wire can be used for the filament leads.

Corona and arcing. A point at high potential can ionize the surrounding air to produce a visible corona effect, which is light blue in color. The corona causes loss of power and eventual insulation breakdown with arcing. In addition, precipitation of ionized dust particles is a result of the corona. It is important to eliminate sharp edges in the wiring; also all soldered joints should be smooth and round to minimize corona. The

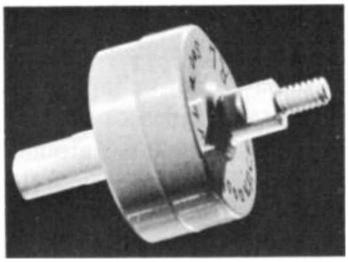


Fig. 8 · 17 500-μμf high-voltage filter capacitor, often called a skatewheel. Voltage breakdown rating 20 kv. Diameter 1 in. (Centralab.)

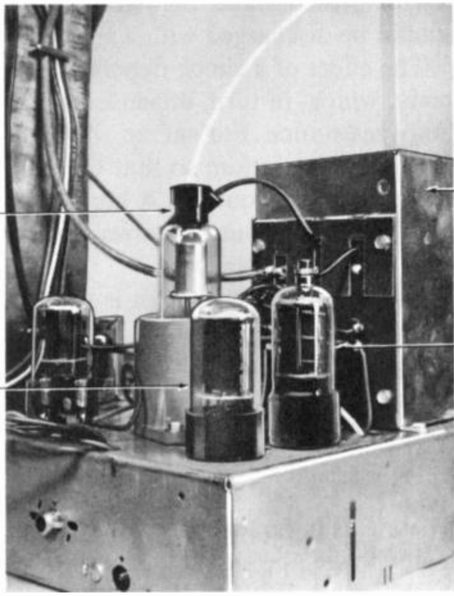


Fig. 8 · 18 Flyback power supply in high-voltage cage on receiver chassis. (RCA.)

solid metal ring under the 1B3 socket in Fig. 8 · 16 is connected to the filament in order to distribute the potential here over as large an area as possible to minimize corona. Thick wires also have less corona, as the voltage gradient of the conductor surface to the surrounding air is reduced.

Corona and arcing in the high-voltage supply can produce streaks in the picture on the kinescope screen. To recognize these effects, arcing can usually be heard as a snapping noise, while corona produces a sizzling sound. Often you can see the corona or arcing by looking into the high-voltage cage. All high-voltage points must be well separated from the chassis to eliminate arcing. If we take the voltage breakdown rating of air at about 20 kv per in., high voltage at 15 kv can jump a gap of ¾ in.

8 · 9 High-voltage safety precautions

If receiver operation is checked with the back cover off, a cheater cord for the a-c line can be used to bypass the safety interlock. With power on, it is a good idea to use only one hand when touching any part of the receiver. This precaution against a severe shock applies to both the anode voltage and B+ voltage. In the high-voltage supply, the rectifier plate cap is dangerous because the current here is limited only by the resistance of the input circuit. The a-c high voltage here can cause a skin burn. The highest d-c voltage is at the rectifier filament with its lead to the kinescope anode and the high side of the filter capacitance.

If the receiver can be checked without high voltage, disable the horizontal amplifier or damper stages to remove the flyback high voltage. Taking the rectifier out removes d-c high voltage but the a-c high-voltage input is still present. The high-voltage filter capacitor and kinescope anode should be discharged with a lead connected at one end to chassis ground.

The effect of a shock depends on how much current flows through the body, which, in turn, depends on the amount of applied voltage and the body resistance. For safety, the high-voltage supply has high resistance and poor regulation, so that the voltage can drop sharply with a partial short across the output. A high-resistance filter instead of a choke helps limit maximum output current. Also, the high-voltage rectifier itself has high internal resistance.

Another important factor is the effect of the charged filter capacitor discharging through the body. Low values of filter capacitance are necessary as a safety measure. The amount of energy stored in the charged capacitor equals

$$\mathcal{E} = \tfrac{1}{2}(CV^2) \qquad (8 \cdot 1)$$

where C is in farads, V in volts, and the energy $\mathcal{E}$ in joules. Because of the high voltage, the stored energy should be no more than 1 joule, which is 1 watt-second. With 15 kv across 500 $\mu\mu$f, $\mathcal{E}$ equals 0.056 joule. It is interesting to note that, in the B+ supply, a 40-μf capacitor charged to 300 volts stores 1.8 joules, which is more energy than in the high-voltage filter capacitor.

8·10 High-voltage troubles

The kinescope anode voltage affects screen brightness and focus. Also, the raster size can vary, as deflection sensitivity increases with less high voltage. Without any high voltage, there is no brightness. Insufficient high voltage reduces brightness and makes the raster bigger but usually with poor focus. When the high voltage is increased the raster becomes smaller, since more deflection power is needed with a greater accelerating force on the electron beam.

If the trouble is no anode voltage for the kinescope there is no illumination at all on the screen—no raster, no line, or no spot. In a flyback supply, either the trouble is in the high-voltage rectifier and its d-c output circuit or there are no horizontal flyback pulses to produce the a-c high voltage. Therefore, no high-voltage a-c input to the rectifier can be the result of trouble in the horizontal oscillator, amplifier, or damper stages.

The output voltage of flyback high-voltage supply normally drops slightly with more beam current for more brightness. However, the picture may become too large and defocused as the raster blooms when brightness is increased. This trouble indicates the internal resistance of the high-voltage supply is too high, causing insufficient voltage regulation. A weak high-voltage rectifier tube is often the cause of this trouble of a *blooming raster.*

8·11 Troubles in the low-voltage supply

Since it is common to all sections of the receiver, a trouble here can produce multiple effects. Two common indications are (1) no brightness and no sound at the same time; (2) a small raster with both height and width reduced. Table 8·2 illustrates how the raster size shrinks and sound volume decreases with B+ voltage less than normal, because of either low a-c line voltage or trouble in the B+ supply. A common cause of low B+ voltage is weak rectifiers. With a stacked B+ circuit, a weak audio output tube can cause low B+ voltage for the r-f and i-f amplifiers.

If there is no B+ voltage at all the receiver will be dead with no brightness and no sound, although the heaters can light. For this case keep in mind the fact that in many receivers the B+ voltage is normally disconnected when the yoke plug is removed or when a TV-Phono switch is in the Phono position.

With series heaters, an open in one means no current in the entire string. The full a-c line voltage is across the open heater because its resistance is infinitely high. To find the open heater in a long string the following technique can be helpful. Assuming the schematic diagram is available, check for continuity with an ohmmeter from a heater in the middle of the string to chassis ground and then to the a-c input. A reading of infinite ohms indicates which half of the string has the open. Then check from the center of this half to either end, and continue until the open heater is localized. This way, groups of heaters with continuity are eliminated as the source of trouble, instead of checking one at a time.

8·12 Hum

Excessive hum voltage at 60 or 120 cps can cause one or two pairs of dark and light horizontal bars or vertical bend in the picture, in addition to hum in the sound. These effects of hum in the picture result from hum in the video signal. When the hum frequency is 120 cps, this is excessive ripple in the B+ voltage from a full-wave rectifier with insufficient filtering. The 60-cps hum may be caused by heater-cathode leakage in a tube or ripple in the B+ voltage from a half-wave rectifier.

Table 8·2 Effects of changes in power-supply voltage

A-c input, volts	*B+, volts*	*Effect*
110	385	Normal raster, 10 × 13¼ in.
100	340	Height reduced ½ in., width reduced ½ in., slight defocusing
90	300	Height reduced 1 in., width reduced 1 in., out of focus, reduced brightness
80	240	No brightness, sound volume low
65	120	No brightness, no sound volume

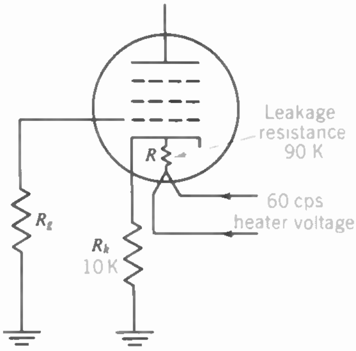


Fig. 8·19 Heater-cathode leakage resistance coupling 60-cycle hum voltage into cathode circuit.

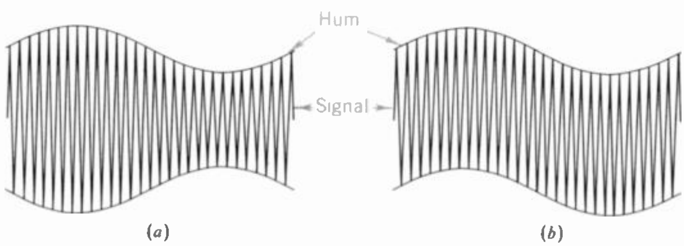
Heater-cathode leakage. Referring to Fig. 8·19, the leakage resistance R between heater and cathode inside the tube forms a voltage divider with the cathode resistor R_k. In this case R_k is 10 per cent of the total resistance in the voltage divider. Therefore, 10 per cent of the heater voltage is developed across R_k from cathode to ground. Note that the control grid returns to the grounded side of R_k through R_g. Then the 60-cycle hum voltage across R_k varies the grid-cathode voltage and the plate current. It should be noted that transistors and semiconductor diodes cannot have hum injected this way because there is no heater.

Higher values of R_k allow more hum voltage to be introduced by heater-cathode leakage. Two examples of high cathode resistance are in the discriminator and ratio detector FM detector circuits. These stages may have reduced heater voltage to minimize hum injection by heater-cathode leakage.

Modulation hum. Hum can be present in high-frequency signal circuits, even though they are not able to amplify the hum frequency, if the hum voltage modulates the high-frequency signal. In a picture i-f amplifier tuned to 45.75 Mc, as an example, leakage between heater and cathode can allow the filament voltage to modulate the grid signal voltage. The result is i-f output amplitude-modulated by the 60-cps hum. Succeeding stages amplify the modulated i-f signal and when it is rectified the detected output includes the 60-cycle hum in the video signal.

Hum modulation usually occurs in the grid and cathode circuits of low-level stages. Modulation hum is often called *tunable hum,* since the hum

Fig. 8·20 Waveforms of hum combining with higher frequency signal. (a) Modulation hum with opposite envelopes. (b) Additive hum with parallel envelopes.



is evident only when a signal is tuned in. The effects of modulation hum can be seen in the picture, therefore, but not in the raster alone. The local oscillator, r-f amplifier, converter, picture i-f, and sound i-f stages can have modulation hum.

Additive hum. In a circuit that has a plate load impedance for 60-cps or 120-cps a-c voltages, the hum can be present in addition to the desired signal. Modulation is not necessary, since the circuit can amplify the low-frequency hum voltage itself. The audio amplifiers, video amplifiers, sync amplifiers, vertical deflection circuits, and horizontal deflection circuits can amplify the hum voltage. The effects of additive hum are evident in the raster and in the picture. Figure 8 · 20 illustrates the difference in waveforms for modulation hum and additive hum.

SUMMARY

1. Using a 5U4 full-wave rectifier for B+, the d-c output voltage is approximately equal to the rms a-c input per plate, with a total direct load current of about 200 ma. Ripple frequency is 120 cps.
2. In a full-wave voltage doubler circuit, two diodes are charged on alternate half cycles and discharged in series to provide d-c output voltage double the a-c input voltage. Ripple frequency is 120 cps.
3. In a half-wave or cascade doubler one diode inserts a d-c level to raise the peak input voltage for the other diode. The d-c output voltage is double the a-c input voltage with a ripple frequency of 60 cps.
4. Compared with vacuum-tube rectifiers, the semiconductor diodes are smaller and more efficient, with less internal resistance. This requires a series-limiting resistor for protection against current surges. However, they have much lower ratings for peak inverse voltage.
5. In a transformerless or a-c/d-c type of power supply, all the heaters are in a series string and a voltage doubler is generally used for B+ voltage. The shock hazard with B− connected to one side of the a-c input can be eliminated by using an isolation transformer. An open in one heater opens the entire series string.
6. A common arrangement for stacked B+ uses the audio output tube plate-cathode circuit as a dropping resistor in series with the plate supply for the common i-f stages. In this case, the audio output tube determines B+ voltage for the picture i-f section.
7. In the flyback high-voltage supply, the fast drop of current in the horizontal output transformer produces high-amplitude positive pulses of self-induced voltage, stepped up for the plate of the high-voltage rectifier, which conducts to charge its filter capacitor to about 15 kv d-c high voltage for the kinescope anode.
8. No d-c high voltage for the kinescope anode means no brightness. With a flyback supply, the horizontal oscillator, output, and damper stages must be operating to have high voltage.
9. No B+ voltage results in no brightness and no sound. Low B+ voltage can reduce both height and width of the raster.
10. Protective devices often used include: 3-amp fuse in the primary of the power transformer, wire fuse for heater current, thermistor in heater line to reduce current until warmup, thermal overload with reset button for B+ circuit, 5-ohm surge resistor in series with semiconductor diodes, and ⅛-amp fuse for high-voltage circuit.
11. Hum at 120 cps is ripple from a full-wave rectifier; 60-cycle hum is caused by heater-cathode leakage or ripple from a half-wave rectifier. For either frequency, modulation hum is in the picture but not in the raster, as signal input is necessary; additive hum is in the raster and picture, with or without signal.

SELF-EXAMINATION *(Answers at back of book.)*

Choose (*a*), (*b*), (*c*), or (*d*).

1. The ripple frequency for a cascade doubler is: (*a*) 60 cps; (*b*) 120 cps; (*c*) 240 cps; (*d*) 800 cps.
2. An advantage of a silicon diode compared with the 5U4 tube is: (*a*) higher peak inverse voltage rating; (*b*) higher temperature rating; (*c*) high forward resistance; (*d*) smaller size.
3. Referring to Fig. 8·2 if the entire secondary L_2 supplied 800 volts rms the d-c output voltage would be approximately: (*a*) 80; (*b*) 320; (*c*) 425; (*d*) 800.
4. For the same case as Question 3 the peak inverse voltage at each plate of the 5U4 would be approximately: (*a*) 425; (*b*) 800; (*c*) 1,020; (*d*) 1,120.
5. In a full-wave doubler circuit: (*a*) both diodes conduct on the same half cycle; (*b*) each diode conducts on alternate half cycles; (*c*) both capacitors charge on the same half cycle; (*d*) the ripple frequency is 60 cps.
6. Referring to Fig. 8·6, if an ohmmeter from kinescope heater to ground reads infinite ohms, which tube can have the open heater? (*a*) 6AW8; (*b*) 12AX4; (*c*) 6CG7; (*d*) 5CG8.
7. A transformerless power supply has: (*a*) a voltage doubler and series heaters; (*b*) a voltage doubler and parallel heaters; (*c*) a full-wave rectifier and parallel heaters; (*d*) one side of the power line connected to B+.
8. In the flyback high-voltage supply: (*a*) ripple frequency is 60 cps; (*b*) a full-wave rectifier is used; (*c*) the 1B3 filament can be in a series string with all other heaters; (*d*) the sharp drop in horizontal output current produces the a-c high voltage.
9. With an open ⅜-amp fuse in the high-voltage cage the result will be: (*a*) no brightness, normal sound; (*b*) no brightness, no sound; (*c*) normal raster, no heaters lit; (*d*) small raster, weak sound.
10. Heater-cathode leakage in an i-f amplifier tube can cause: (*a*) additive hum at 120 cps; (*b*) modulation hum at 60 cps; (*c*) modulation hum at 120 cps; (*d*) 400-cycle hum.

ESSAY QUESTIONS

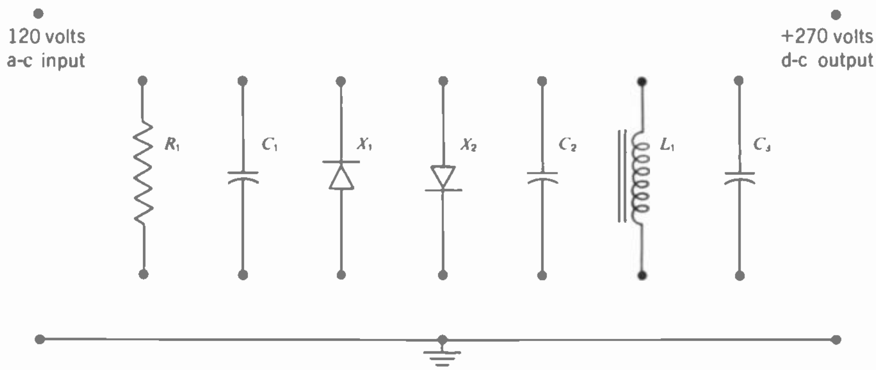
1. Draw the schematic diagram of a power supply with a full-wave rectifier for B+, parallel heaters, and π-type filter, for 400 volts d-c output voltage at 200 ma. Give the function of each component.
2. Draw the schematic diagram of a transformerless power supply with cascade doubler for B+ and series heaters. How much is the d-c output voltage with 200 ma load current using 100-μf capacitors? Give the function of each component.
3. Draw the schematic diagram of a power supply using a full-wave doubler with an isolation transformer and parallel heaters.
4. Give one advantage and one disadvantage of series heaters compared with parallel heaters.
5. What is corona? Describe briefly two insulation problems in a high-voltage supply.
6. Give two safety measures to keep in mind when working with high-voltage equipment.
7. Referring to Fig. 8·2, give the function of R_1, C_1, S_1, S_2, F_1, F_2, C_{2A}, C_{2B}, L_5, and R_2.
8. Referring to Fig. 8·11 give the function of R_1, R_3, RFC_2, C_2, C_5, X_1, and X_2.
9. If R_3 opened in Fig. 8·11, what would be the effect on picture and sound?
10. By referring to a tube manual, list the stage and function for each tube type in Fig. 8·6.
11. Why does no output from the high-voltage supply result in no brightness on the kinescope screen? Give three other possible causes of no brightness with normal sound.
12. What is a possible source of hum causing one pair of dark and light bars across the picture but not in the raster, with hum in the sound? This is an intercarrier receiver with the sound take-off in the video detector output circuit.

PROBLEMS *(Answers to odd-numbered problems at back of book.)*

1. Refer to Fig. 8·2. (*a*) If L_5 had a resistance of 200 ohms, how much would E_1 be? (*b*) If R_2 had a resistance of 2,000 ohms, how much would E_2 be?(*c*) Assume E_2 is used for

five identical stages. How much is the combined plate and screen current for each stage?

2. Refer to Fig. 8·14. With 400 volts a-c input per plate, list d-c output voltages for load of 50, 100, 150, 200, and 250 ma. Plot a graph of output voltage on the *Y* axis vs. load current on the *X* axis. Why does output voltage decrease with more load current?
3. Refer to Fig. 8·11. (*a*) How much is the total hot resistance of the heater string across the 120-volt line with 450-ma heater current? (*b*) How much is the resistance of the 6CL8 heater?
4. A series string totals 90 volts with 600 ma in all the heaters. What size voltage-dropping resistor is needed for a 120-volt supply? How much power is dissipated in this series resistor?
5. A 6.3-volt heater rated for 450 ma is in series with other heaters rated at 600 ma. What size resistor should be across the 6.3-volt heater to bypass 150 ma with 6.3 volts across both? How much power is dissipated in this parallel resistance?
6. The current through a 2-henry coil drops from 10 to 9 amp in 1 sec. How much is the self-induced voltage across the coil produced by this current change? Ignore the resistance of the coil.
7. Refer to Fig. 8·6. Give a-c voltmeter readings for the following: (*a*) across F_1; (*b*) F_1 to ground; (*c*) across 12DQ6; (*d*) high side of 12DQ6 to ground.
8. Show how to connect the components below in a cascade voltage doubler circuit, with π-type filter. Give functions and typical values.



Chapter 9 Video amplification

The video amplifier in the television receiver amplifies the output from the video detector, so that there will be enough video signal voltage coupled to the kinescope (see Fig. 9 • 1). With 3 volts peak-to-peak video signal from the detector and a voltage gain of 25, as typical examples, the video amplifier output for the kinescope is 75 volts for good contrast in the picture. This function corresponds to an audio amplifier in a sound system, amplifying the audio voltage from the second detector to provide enough signal to drive the loudspeaker.

When you watch the picture on the kinescope screen, the reproduction depends on the video signal. Without video, there is no picture on a blank raster. Insufficient video signal causes weak contrast. Distortion of the signal in the video amplifier causes distortion in the picture.

9 • 1 The video signal and picture reproduction

The function of the video signal is to vary the intensity of kinescope beam current. This intensity modulation reproduces the degree of light or shade of the picture elements on the kinescope screen. Figure 9 • 2 illustrates how video signal voltage for one horizontal line, impressed between grid and cathode of the picture tube, results in reproduction of the picture elements in the line.

Intensity modulation of kinescope. The operating characteristic illustrated in Fig. 9 • 2 is for the kinescope, not the video amplifier. Still, the effect of a-c video signal on the kinescope grid is the same as any a-c signal coupled to the control grid of a vacuum tube having a steady negative d-c bias voltage. The bias sets the operating point around which the instantaneous grid voltage varies, as the a-c signal makes the grid more or less negative with respect to cathode.

The kinescope beam current and spot illumination both vary with the grid voltage. As the grid goes more negative there is less beam current and

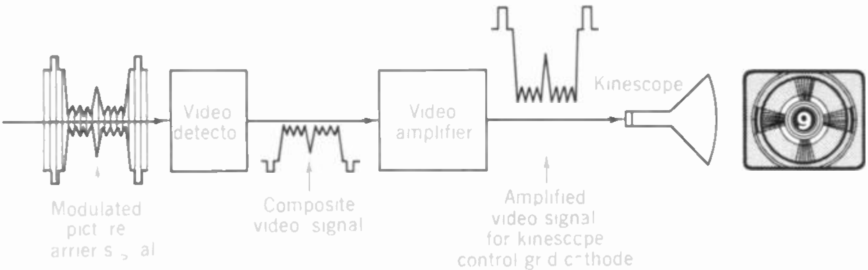


Fig. 9·1 The video signal coupled to kinescope grid-cathode circuit reproduces the picture information on the raster.

the intensity of the spot on the screen is reduced. This produces a darker picture element. When the grid voltage goes as far negative as cutoff, the beam current is cut off completely and there is no spot of light on the screen. This corresponds to black. As the grid voltage is made less negative or more positive than the d-c bias value, the beam current in the tube increases to produce a brighter spot on the screen. This corresponds to the white parts of the video signal, with maximum white driving the grid voltage the most positive. Although the effect of the white parts of the video signal is to drive the kinescope in the positive direction, the control-grid voltage remains negative because of the d-c bias.

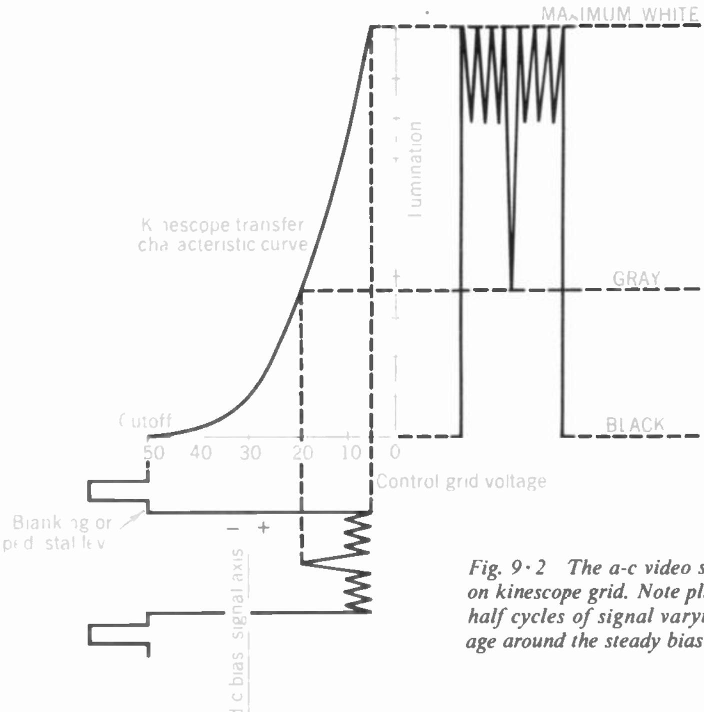
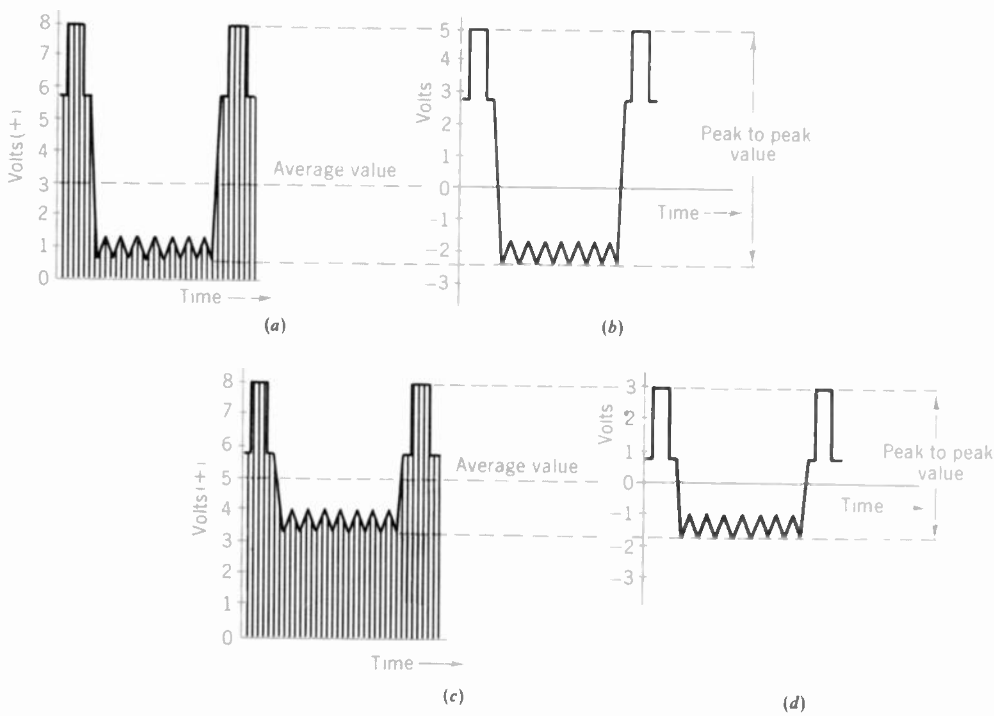


Fig. 9·2 The a-c video signal voltage on kinescope grid. Note plus and minus half cycles of signal varying grid voltage around the steady bias value.

Video signal amplitude determines contrast. The extent to which the video signal voltage swings away from its average-value axis determines the contrast of the reproduced picture. Suppose that the video voltage has only one-half the amplitude shown in Fig. 9·2. The maximum white of this weaker signal will not be so bright because the kinescope grid voltage will not be driven so far positive. This means that there will be less difference between the maximum white parts of the picture and the black level. The reproduced picture will not have so much contrast, resulting in a picture that appears weak and flat with no highlights. The amount of composite video signal required for most picture tubes is about 75 volts peak to peak which includes the sync pulses. With an output of 3 to 5 volts from the video detector and a gain of approximately 20 in a video amplifier, one stage usually provides enough video signal for the kinescope.

The video signal amplitude is given in peak-to-peak voltage instead of rms, average, or peak value. This is necessary because the video signal is not symmetrical, and its waveshape changes with picture content. Figure 9·3 shows that the peak-to-peak value of the composite video signal includes the total voltage swing between the tips of the synchronizing

Fig. 9·3 Peak-to-peak value of video signal. (a) D-c form for white video signal. (b) A-c form for same white video signal. (c) D-c form for a gray video signal. (d) A-c form for same gray video signal.



pulses and the maximum white level. Note that the peak-to-peak voltage for the white information in *a* and *b* is greater than for the gray information in *c* and *d*. The common a-c relations—that the average value is 0.636 the peak value and the effective value is 0.707 the peak value—apply only to sine waves. Since the positive and negative half cycles of the video signal are not necessarily the same, any notation other than peak-to-peak value is useless.

Polarity. If the polarity is opposite from that shown for the video signal in Fig. 9 • 2, a negative image in the same sense as a photographic negative will be produced. The dark parts of the original image will be white on the kinescope screen while the white parts of the picture will be reproduced as black. The correct polarity of video signal makes the kinescope grid voltage less negative than the d-c bias to produce more beam current for white picture information, and the synchronizing pulses drive the kinescope grid voltage more negative beyond cutoff. Video signal polarity is usually indicated by the polarity of sync voltage. Therefore, the signal shown in Fig. 9 • 2 has negative sync polarity.

The steady d-c bias determines average brightness. Since the a-c video signal varies the grid voltage above and below the d-c bias voltage, the picture elements reproduced on the screen vary in intensity around the brightness value corresponding to the bias voltage. Therefore, the kinescope grid bias determines the average amount of beam current and the average brightness of the reproduced picture.

Black level. The grid-cutoff voltage for the picture tube corresponds to black. When the negative grid voltage is equal to the cutoff value there is no beam current and no illumination of the screen. The part of the video signal that corresponds to black, therefore, should drive the kinescope grid voltage to cutoff. Any grid voltage more negative than cutoff is blacker than black. The pedestal or blanking level of the composite video signal should drive the kinescope grid voltage to cutoff.

9 • 2 Polarity of the video signal

At the kinescope, the correct polarity is the result of three factors: (1) whether the video signal is coupled to the kinescope control grid or to the cathode, (2) the number of video amplifier stages, and (3) polarity of video detector output. When the output from the video amplifier is coupled to the kinescope control grid, for grid drive, the video signal must have negative sync phase. Then white in the signal drives the grid voltage in the positive direction for more beam current. However, when the video signal is coupled to the kinescope cathode, for cathode drive, it must have the opposite polarity. Remember that driving the cathode more negative, for white information, corresponds to making the grid less negative, as both changes increase the beam current. Therefore, video signal of positive sync polarity is required at the kinescope cathode.

One video amplifier stage inverts the polarity of its input signal, because the circuit is essentially an *RC*-coupled amplifier. Then the output from an

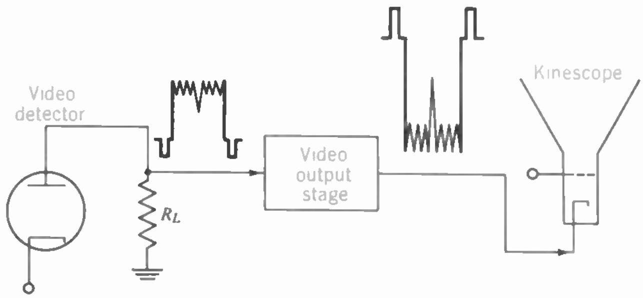


Fig. 9·4 Polarity of video signal for kinescope. (a) Positive sync for cathode drive.

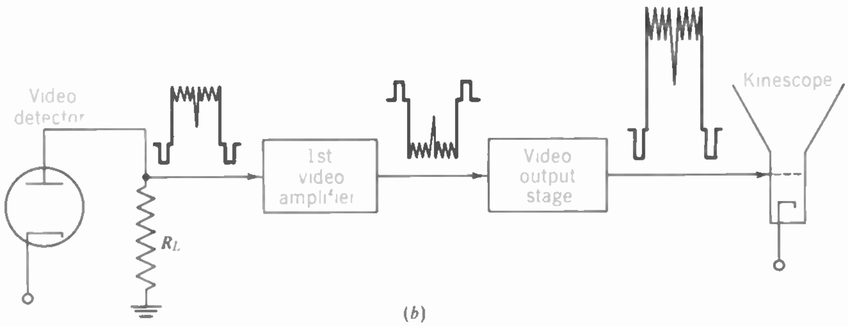
odd number of video amplifiers has opposite polarity from the input because each stage inverts the signal voltage. An even number of stages, however, has output signal of the same polarity as the input, since two phase inversions produce the original input polarity.

For the video detector circuit alone, when its diode load resistor is in the plate circuit the video signal output has negative sync polarity, with the carrier signal modulated for negative transmission. If the diode load resistor is in the cathode circuit the video signal output from the detector will have positive sync polarity.

Many combinations of these factors are possible, but receivers generally have the arrangement in either *a* or *b* of Fig. 9·4. In *a*, negative sync polarity from the video detector is inverted by one video amplifier. The output video signal of positive sync polarity is coupled to the kinescope cathode. In *b*, the same negative sync polarity from the video detector is inverted twice by two video amplifier stages, providing video signal with negative sync polarity for the kinescope grid.

9·3 Amplification of the video signal

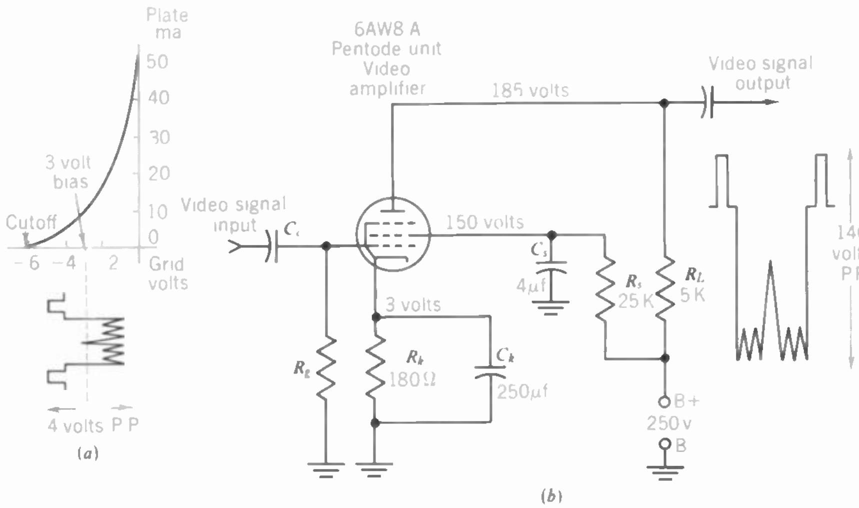
As shown in Fig. 9·5, the circuit of a video amplifier is basically an *RC*-coupled stage operating essentially the same as a class A audio amplifier. Composite video signal input swings the instantaneous grid voltage around the bias axis. The 3-volt bias is less than cutoff grid voltage and the ±2-volt grid drive allows plate current for the full cycle. The resulting variations in plate current produce amplified video signal output voltage across the plate load resistor R_L. With signal output of 140 volts peak to peak produced by 4 volts input, the voltage gain of this stage equals 140/4 or 35. Note the polarity inversion. The output signal with positive sync polarity is then coupled to the kinescope cathode for reproducing the picture information on the raster. Its amplitude can be reduced if necessary by the contrast control, not shown in the circuit. Also omitted are peaking coils generally used to boost the high-frequency response but they do not affect the peak-to-peak amplitudes shown.



(b) Negative sync for grid drive.

Although the grid signal in Fig. 9·5 has negative sync voltage, either polarity can be used. For one stage, however, the polarity shown is preferable. Notice that white parts of the video signal vary the grid voltage within the linear portion of the tube's operating characteristic. Signal amplitudes near black level may be compressed but this is not so noticeable in the picture. Compression of the white amplitudes can cause a flat pasty appearance in large white areas, such as the faces of people in the scene. Another advantage of operating the video amplifier with input signal of positive picture phase is that noise voltage having higher amplitudes than the negative sync pulses can be clipped when the noise voltage drives the grid more negative than cutoff.

Fig. 9·5 Video amplifier circuit and its operating characteristic. (a) Video signal with negative sync and positive white for control grid. (b) Basic RC-coupled amplifier.



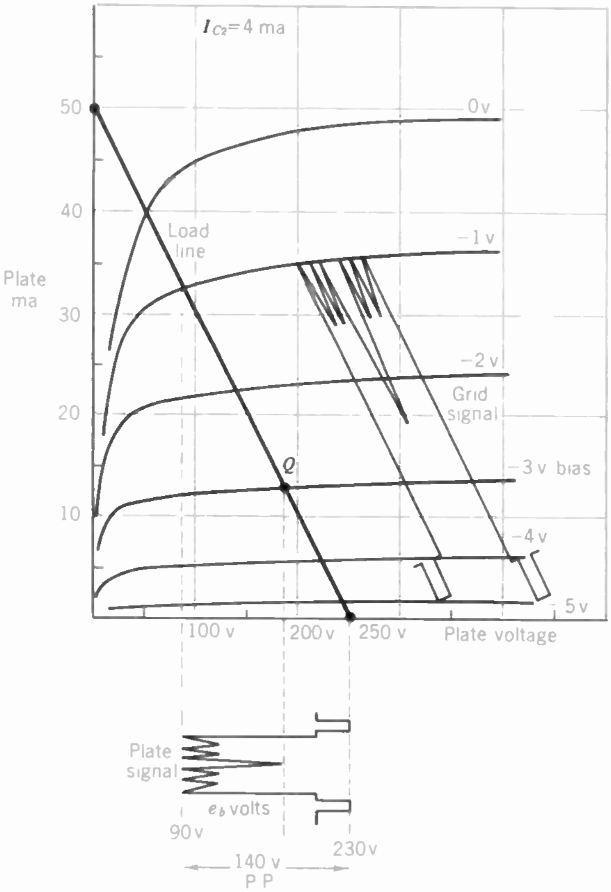


Fig. 9·6 Load line calculations for video amplifier circuit in Fig. 9·5. Plate characteristic curves for 6AW8 pentode section.

D-C voltages. If you measure with a d-c voltmeter from plate to chassis ground at B−, the meter will read 185 volts for the average d-c plate voltage, as indicated in Fig. 9·5*b*. This value equals the B+ voltage minus the average i_bR_L drop. Similarly the screen-grid voltage equals the B+ voltage minus the *IR* drop across R_s. The cathode bias of +3 volts makes the control grid 3 volts negative with respect to cathode.

A-C voltages. Checking with an oscilloscope at the grid of the amplifier, you can see the composite video signal waveform. The oscilloscope can be calibrated to read the 4-volt peak-to-peak amplitude of the a-c signal input. With a gain of 35 for the stage, the output video signal equals 140 volts peak-to-peak.

Amplitude distortion. If the d-c voltages are not correct in the video amplifier the a-c video signal can be distorted, just as in an audio amplifier. Typical problems are limiting and clipping of the video signal amplitudes or weak signal output. When the white signal amplitudes are compressed by limiting or clipping, the white parts of the picture appear too flat. If the sync voltage is compressed, synchronization can be lost because the video amplifier usually provides composite video signal for the sync circuits.

Load line. The amplitude variations of the video signal can be analyzed by a load line for this amplifier, as shown in Fig. 9·6. One end of the load line is at E_{bb} of 250 volts, equal to B+, as this is the plate voltage when plate current is zero. The other end is at 50 ma, which is the plate current equal to 250 volts/5,000 ohms when the plate voltage is zero because the B+ voltage is dropped across R_L.

All instantaneous values of grid voltage e_c, plate current i_b, and plate voltage e_b are on the load line. The quiescent point marked *Q* is set by the 3-volt bias. As grid signal swings e_c 2 volts less negative to −1 volt and 2 volts more negative to −5 volts, i_b varies between 32.5 and 1.5 ma. The corresponding peak values of e_b are 90 and 230 volts. Therefore, the amplified a-c signal voltage at the plate is 140 volts peak to peak, equal to the difference between these two peak voltages (see Fig. 9·7).

By means of the values on the load line, we can calculate the d-c voltages. The Q point at 3 volts bias sets the average plate current I_b at 13 ma. The I_bR_L voltage drop equals $5{,}000 \times 0.013$, or 65 volts. Subtracting this 65-volt drop from the E_{bb} supply of 250 volts results in 185 volts for the average d-c plate voltage.

With 13 ma I_b and 4 ma screen-grid current the total average cathode current equals 17 ma. Therefore, R_k has the value of 3/0.017, or approximately 180 ohms for the 3-volt bias.

Similarly, R_s is 25,000 ohms to produce a voltage drop of 100 volts with 4 ma screen current. This 100-volt IR drop across R_s subtracted from the supply of 250 volts provides 150 volts for the screen-grid voltage.

9·4 Manual contrast control

Corresponding to an audio volume control for sound, the *contrast control* or *picture control* on the receiver front panel can be varied for the desired amount of contrast in the reproduced picture. As the control is turned clockwise, more a-c video signal is provided for the kinescope. The result is a greater variation between black and the whitest parts of the picture to increase the contrast.

Any control that varies the amount of a-c video signal for the kinescope will operate as a contrast adjustment. Therefore, the contrast varies when the gain is varied in either the picture i-f section or the video amplifier. The contrast control is generally in the video amplifier, however, as practically all receivers have automatic gain control to adjust the bias on the picture i-f amplifiers automatically according to the signal level. Also, with intercarrier sound a variation of i-f bias affects the sound volume.

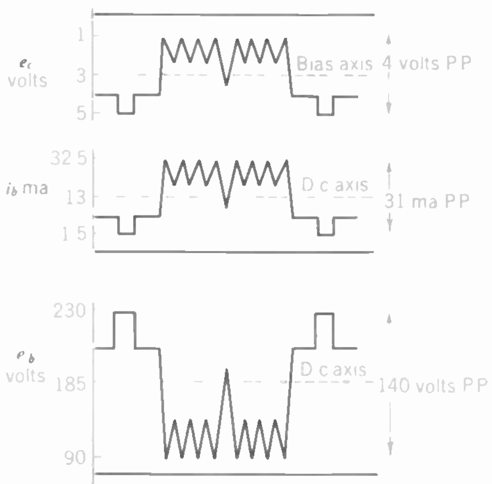


Fig. 9·7 Video signal waveshapes for load-line operation in Fig. 9·6.

The two most common methods of contrast control in the video amplifier are shown in Fig. 9·8. In *a* the variable cathode resistor R_1 varies the bias for the video amplifier. The control is unbypassed in order to provide degeneration of the a-c signal. Then the amount of degeneration also varies as the bias resistor is varied. The degeneration is important for varying the gain because with linear amplification little change in gain results from changing the bias. Moving the variable arm of R_1 closer to the cathode end reduces the bias and the degeneration. This allows more gain to increase the video signal amplitude and the contrast.

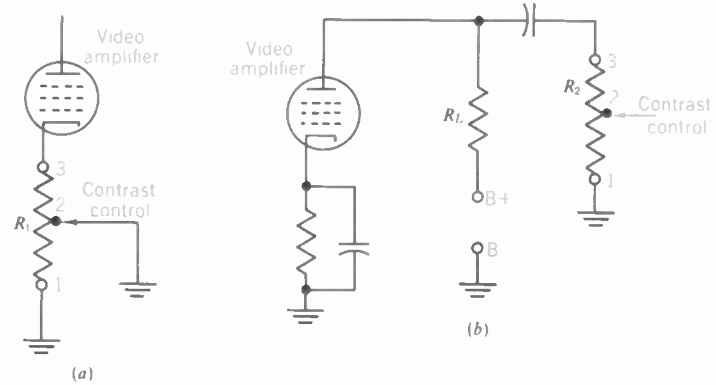
The variable bias method of contrast control has the disadvantage of changing the amplifier operating characteristics, which can introduce amplitude distortion. To overcome this problem, the arrangement in *b* taps off the desired amount of a-c signal output without changing the bias. Now the contrast control is in the video signal circuits, however, where the stray capacitance of the potentiometer and its connecting leads can reduce the high-frequency response of the amplifier. To minimize shunt capacitance, the control is usually mounted near the video amplifier with the shaft mechanically linked to the front panel of the receiver.

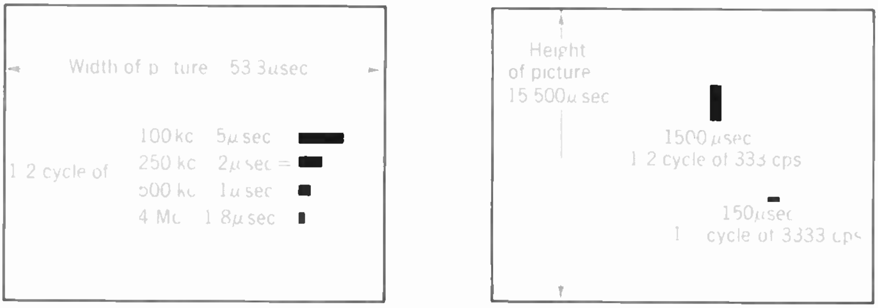
9·5 Video frequencies

The video stage amplifies signal voltage that may have frequency components from 30 cycles to 4 Mc per sec. High frequencies are produced because the video signal contains, within a line, rapid changes in voltage that occur during very much less time than the active line-scanning time of 53.3 μsec. These fast variations in signal voltage can correspond to frequencies infinitely high but are limited to approximately 4 Mc by the restriction of a 6-Mc transmission channel. There are frequencies lower than 4 Mc, but this sets an upper limit on the signal frequencies that the video amplifier in the receiver has to amplify.

The relationship between the frequency of a video signal variation and

Fig. 9·8 Methods of contrast control in a video amplifier. (a) Variable cathode bias. (b) Potentiometer to tap off desired amount of a-c video signal voltage.





(*a*) Horizontal information

(*b*) Vertical information

Fig. 9·9 Relation of video signal frequencies to size of picture information scanned horizontally in (a) or vertically in (b). Time for width of picture in (a) is 63.5 μsec minus 16 per cent H for horizontal blanking. Time for height of picture in (b) is 1/60 sec minus 7 per cent V for vertical blanking.

its associated picture information is illustrated in Fig. 9·9. To convert the size of an element of picture information into frequency, it is first necessary to calculate the time required to scan the element. This time can be considered as the period of a half cycle of the video signal required to reproduce the information. Then you can:

1. Multiply the time by 2 to obtain the period of a full cycle.
2. Take the reciprocal of the period. The answer is the desired frequency. With time in microseconds, the frequency is in megacycles.

As an example, in Fig. 9·9*a* the top black bar has a width slightly less than one-tenth of the picture width. Therefore, the bar is scanned horizontally in 5 μsec. This time is a little less than one-tenth of 53.3 μsec for the entire picture width. As a result, the video signal for this black bar preceded and followed by white information corresponds to one-half cycle of a signal variation with a period of 10 μsec for the full cycle. The corresponding frequency is 0.1 Mc or 100 kc. The lowest frequency for picture information in the horizontal direction can be considered as 9.4 kc, corresponding to a line all white or all black producing video signal with a half period of 53.3 μsec. The narrowest variations produce the highest frequencies up to 4 Mc.

The signal frequencies corresponding to picture information scanned in the vertical direction can be calculated in a similar manner (see Fig. 9·9*b*). Here the height of the black bars is converted to frequency in terms of active vertical scanning time equal to 0.0155 sec. This is relatively low frequency information compared with details reproduced within a line. If the video voltage is taken from top to bottom, through all the horizontal lines in a field, this variation will correspond to a half cycle of a signal with a frequency of approximately 30 cps. When the brightness of the picture varies from frame to frame, the resultant signal frequency is lower than 30 cps, but this is considered as a change in d-c level to be reproduced by means of a d-c reinsertion circuit.

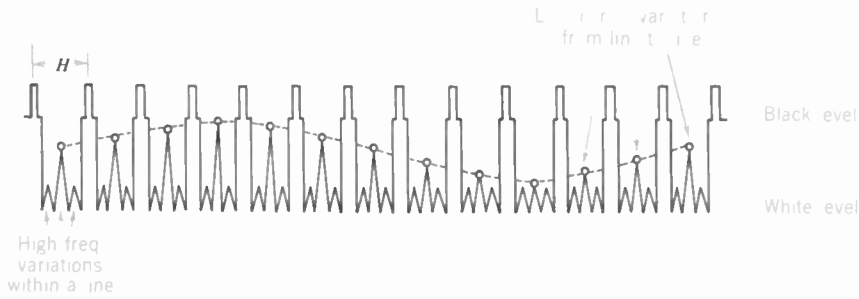


Fig. 9·10 Video signal voltage waveform, illustrating high-frequency components as variations during horizontal scanning within a line. Lower frequency variations during vertical scanning are from line to line within a field.

The a-c video signal, therefore, can be regarded as a complex waveform, not being a sine wave, containing signal voltages that range in frequency from 30 cycles to 4 Mc, approximately. Figure 9·10 illustrates typical video signal waveshapes. Note that the high-frequency variations correspond to white, gray, and black shadings across every line. The lower frequency variations correspond to vertical shadings of gray, black, and white.

9·6 *Frequency distortion*

The amplifier has more gain for some frequencies than for others. Excessive frequency distortion cannot be tolerated because it changes the picture information. As shown in the amplifier response curve of Fig. 9·11 the amplifier response should be flat within a tolerance of about ± 10 per cent. Note that the frequency units are marked off on the horizontal axis in powers of 10, making the spacing logarithmic. This is necessary for a graph of reasonable size that will still show the ends of the response curve. To illustrate the extremely wide range, the highest video frequency at 4 Mc is almost a million times higher than the lowest frequency at 30 cps.

When the amplifier has a flat response curve the relative gain of the amplifier is the same for all signal frequencies. Then the amplifier introduces no frequency distortion. As an example, the 100 per cent response in Fig. 9·11 may correspond to a voltage gain of 35 at the middle frequency of 10 kc. Then the flat response means the gain is 35 for all frequencies from 30 cps to 4 Mc. With an actual video signal containing typical picture information coupled to the amplifier, the different frequency components do not all have the same amplitude. When the amplifier response is flat, however, all frequencies are amplified equally well without any frequency distortion. Then each frequency component in the amplified output signal has the same relative amplitude as in the input signal.

The video amplifier response will not ordinarily be flat over the required frequency range unless precautions are taken in building the amplifier and special compensating circuits are added. Usually the response of the uncompensated video amplifier is down for the high video frequencies (about 0.5 Mc and above). High-frequency compensation is necessary, therefore, for a flat frequency response up to 4 Mc. At the low-frequency end the video amplifier response is often inadequate at about 100 cps and below. Then low-frequency correction of the video amplifier may also be necessary. The response of the video amplifier over the middle range of frequencies is normally flat and requires no compensation.

Loss of high video frequencies. Insufficient response for the high video frequencies reduces the amount of horizontal detail in the picture because the high-frequency components of the video signal correspond to the smallest picture elements in a horizontal line. If these signal frequency variations are lost, the rapid changes between black and white for small adjacent picture elements in the horizontal lines cannot be reproduced on the kinescope screen, with the resultant loss of horizontal detail. Figure 9·12 shows the effect of loss of the high video frequencies on the reproduced test pattern. Notice the lack of separation between the black and white divisions in the top and bottom wedges. The extent to which the divisions in either of these wedges can be resolved indicates the high-frequency response. Normally, the divisions can be seen all the way in to the center

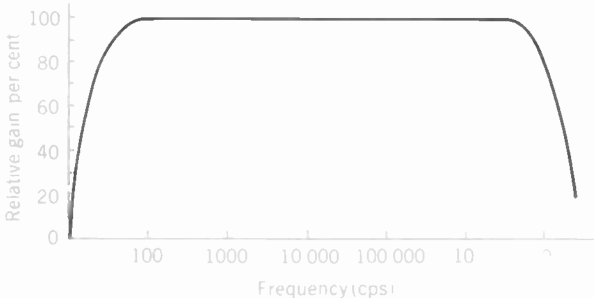


Fig. 9·11 Desired flat frequency response of video amplifier.

Fig. 9·12 Test pattern illustrating loss of high video frequencies. Note the lack of separation of the divisions in top and bottom wedges. Also they have weaker intensity than the side wedges. (RCA Pict-O-Guide.)

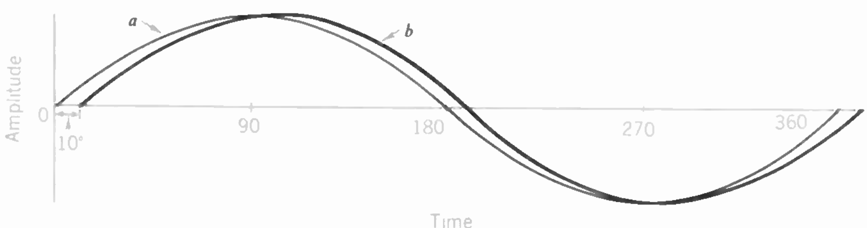
Fig. 9 · 13 Test pattern illustrating loss of low-video signal frequencies. Background is weak, lettering is not solid, and the side wedges have weaker intensity than the wedges at top and bottom. (RCA Pict-O-Guide.)

circle, indicating video-frequency response up to 4 Mc. The divisions in the widest part of the vertical wedges correspond to a video frequency of about 2 Mc.

In a televised scene, loss of the high-video frequency information is evident as reduced detail. The picture does not appear sharp and clear. Small details of picture information, such as individual hairs in a person's eyebrows and details of the eye, are not reproduced. In addition, the edges between light and dark areas, as in the outline of lettering or the outline of a person's face, are not reproduced sharply but trail off gradually instead. The effects of reduced high-frequency response causing insufficient detail in the picture are not so noticeable when the camera at the studio presents a close-up view. Then the subject occupies a large area of the picture and the highest video signal frequencies are not necessary for adequate reproduction.

Loss of low video frequencies. The video frequencies from about 100 kc down to 30 cps represent the main parts of the picture information, such as background shading, lettering, and any other large areas of black and white. This follows from the fact that it takes a longer period of time for the scanning beam to change from black to white over large areas. Frequencies from 100 kc down to about 10 kc correspond to black-and-white

Fig. 9 · 14 Phase delay. Wave b lags behind wave a by the amount of time equal to 10° of the cycle.



information in the horizontal direction having a width one-tenth or more of a horizontal line. Frequencies from 10 kc down to 30 cps can represent changes of shading in the vertical direction. If a solid white frame is scanned, the signal is a 30-cps square wave. If this low-frequency square wave is not amplified with its waveshape preserved, the reproduction will show a white screen having a gradual change of intensity from top to bottom.

Figure 9·13 shows a test pattern reproduced with insufficient low-frequency response. The background is dull gray, instead of white, and the lettering is not solid. Actually, the picture as a whole is weak with poor contrast because low video frequencies represent the main areas of picture information. Notice that the side wedges, which represent low video frequencies, are weaker than the top and bottom wedges corresponding to the high video signal frequencies. The changes from black to white in the vertical direction between the divisions in the side wedges represent frequencies of 4 to 8 kc, approximately.

9·7 Phase distortion

Before the harmful effects of phase distortion at low frequencies can be fully appreciated, the effect of time delay in the picture reproduction must be analyzed. The relative time delay of some parts of the signal with respect to the others is important because one element of the picture is being reproduced at a time as the scanning beam traces out the frame. As a result, a great enough time-delay distortion in the video signal can have the effect of displacing the picture information on the kinescope screen. The result is smear in the picture.

The time it takes the beam to complete its visible scanning from left to right for one horizontal line is 53.3 μsec. Consider now the scanning speed in a picture 20 in. wide. Since it takes 53.3 μsec to cross the screen from left to right and the picture is 20 in. wide, it takes the beam 53.3 μsec to move 20 in. across the screen or 5.33 μsec for 2 in. If some low-frequency video signal suffers a time delay of 5.33 μsec because of phase distortion, the variation of the spot intensity on the screen corresponding to that signal will be displaced 2 in. to the right from its proper position.

Time delay. Phase delay is equivalent to time delay. If one signal voltage is 10° out of phase with another and lagging, as shown in Fig. 9·14, it reaches its maximum and minimum values at a later time. The delay is the amount of time that corresponds to 10° of the cycle. This time varies with the frequency. For a signal voltage having a frequency of 100 cps, it takes $\frac{1}{100}$ sec for one complete cycle of 360°. The amount of time equivalent to 10° in this cycle is $\frac{10}{360} \times \frac{1}{100}$ sec, which is approximately 0.000278 sec, or 278 μsec. In this time the scanning beam can be displaced in a vertical direction by more than four lines.

Phase distortion is very important at low video frequencies, therefore, because even a small phase delay is equivalent to a relatively large time delay. For the extremely high video frequencies, the effects of phase distortion are not as evident on the screen because the time delay at these high frequencies is relatively small. Normally a video amplifier stage that has flat frequency response up to the highest useful video frequency has negligible time-delay distortion for the high frequencies.

Square-wave response. Consider the waveshape shown at the right in Fig. 9 · 15*a*. This wave is actually composed of two sine waves. One is the fundamental, with the same frequency as the combined waveform. The other sine wave is the third harmonic, having a frequency three times the fundamental frequency and an amplitude one-third of the fundamental. Addition of the two sine waves, fundamental and third harmonic, produces the resultant nonsinusoidal wave. This has the same frequency as the fundamental but tends to become more of a square wave than a sine wave. If enough odd harmonics are added to the fundamental, the result will be a square wave.

The waveshape resulting from combining the fundamental and its harmonics is critically dependent on the phase relation between the waves. Figure 9 · 15*b* shows the combination of the same fundamental and third harmonic as in *a*, the amplitude and frequency of each being preserved, but with a different phase angle between the two waves. Assume that in a stage amplifying such a signal the fundamental and third harmonic are originally in phase, as shown in *a*. Then because of phase distortion, the fundamental is made to lag by 60° of its cycle and the third harmonic is made to lag by 120° of its cycle, with the result shown in *b*. The resultant waveshape is distorted from the original wave because the original phase relations between the fundamental and harmonic are not maintained. Phase distortion always produces an unsymmetrical distortion of the waveshape, as in *b*. Notice that the peak amplitudes of the wave in *b* have different values than the original wave and they occur at different times than the

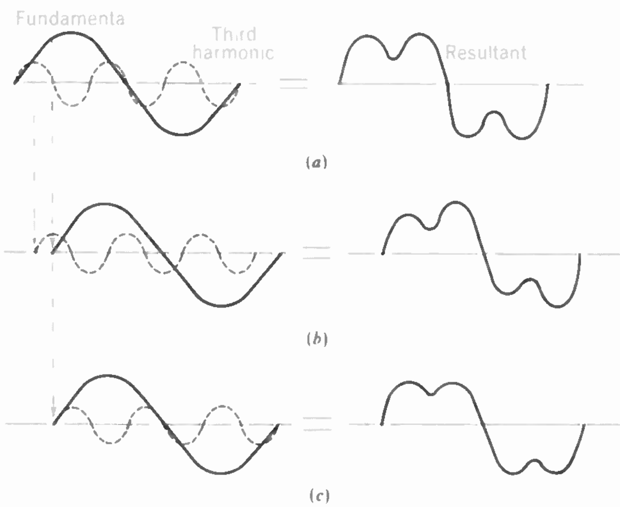


Fig. 9 · 15 How phase distortion changes waveshape of nonsinusoidal signal. (a) Fundamental and third harmonic in phase. (b) Fundamental delayed by 60° and third harmonic by 120°. Phase delay not proportional to frequency. (c) Fundamental delayed by 60° and third harmonic by 180°, proportional to frequency.

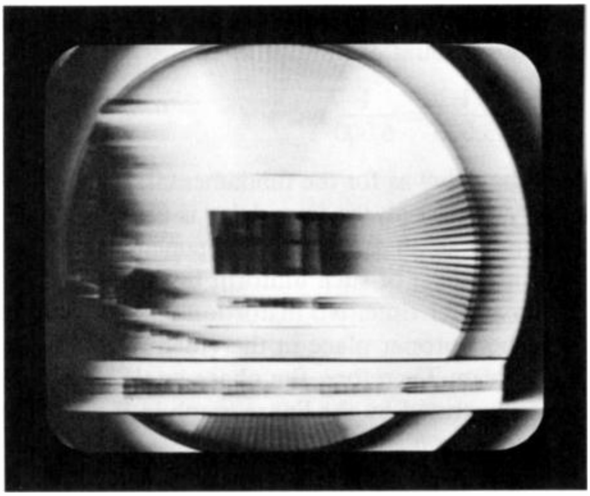


Fig. 9·16 Severe smear caused by excessive low-frequency response with phase distortion. (RCA Pict-O-Guide.)

peaks for the undistorted wave. When the signal amplitudes are distorted this way, the corresponding picture information has incorrect light values and is displaced in time, producing smear in the reproduced image (see Fig. 9·16).

Phase shift should be proportional to frequency. It is important to note that the phase distortion is introduced because the amount of phase shift is not proportional to frequency. The distortion is not caused by the phase shift in itself. When the second harmonic is delayed twice as much as the fundamental, the third harmonic three times as much, and so on, then the phase shift is proportional to frequency and there is no phase distortion. As an example, in Fig. 9·15*c* the third harmonic is delayed by 180° instead of 120°. Then the phase shift of the fundamental and third harmonic are proportional to frequency. As a result, there is no phase distortion and the wave maintains its original shape.

To see why the phase shift should be proportional to frequency, the phase delay must be translated into time delay. Consider a 1,000-cps signal, corresponding to the fundamental frequency in Fig. 9·15, delayed by 60°. This is a delay of 60/360 of the complete cycle that takes 1/1,000 sec. The amount of time delay can be calculated from the formula

$$T_D = \frac{\theta}{360} \times \frac{1}{f} \qquad (9\cdot 1)$$

In this example,

$$T_D = \frac{60}{360} \times \frac{1}{1{,}000} = \frac{1}{6{,}000} \text{ sec}$$

When a 3,000-cps signal corresponding to the third harmonic of the fundamental is delayed by 180° the amount of time delay is

$$\frac{180}{360} \times \frac{1}{3{,}000} = \frac{1}{6{,}000} \text{ sec}$$

This is the same amount of time delay as for the fundamental. With phase shift proportional to frequency, therefore, the time delay is uniform.

The time delay is not harmful if all frequency components have the same amount of delay. The only effect of such uniform time delay would be to shift the entire signal to a later time. No distortion results because all components would be in their proper place in the video signal waveshape and on the kinescope screen. Therefore, the phase angle should be proportional to frequency, as illustrated by the linear graph in Fig. 9·17*a*, to provide the uniform time delay in *b*.

Sources of phase distortion. Nonuniform time delay can be introduced in the video amplifier because of reactance in the plate and grid circuits. Unless the reactance of the grid coupling capacitor is negligible, the signal voltage across the grid resistor will have a different phase angle for different frequencies. Also, with stray capacitance in shunt with the plate load for the video amplifier, the amplified output will be out of phase with the input by more or less than 180°; and this phase angle will vary with frequency.

It should be noted that an inversion of the complete signal of exactly 180° by the video amplifier does not contribute to phase distortion. There is no actual time delay resulting from transit time in the tube at video frequencies, and the phase inversion of 180° caused by the amplifier only reverses the polarity of the video signal with no delay or advance in time.

9·8 Frequency response of the video amplifier

As shown in Fig. 9·18, the resistance plate load makes the response flat over a wide range of middle frequencies. Therefore, an *RC*-coupled stage is best for amplifying the wide band of video frequencies with minimum frequency and phase distortion. Audio-transformer coupling would not be suitable for the high video frequencies that correspond to radio frequencies. A video frequency of only 2 Mc is higher than any r-f signal in the standard radio broadcast band. This entire band from 535 to 1,605 kc would occupy only a small part of the video spectrum from 30 cps to 4 Mc. Coupling with r-f transformers would not be suitable for the lowest video frequencies that correspond to audio frequencies.

Total shunt capacitance. The video amplifier gain is down at the high-frequency end because of the shunting effect of capacitances in the plate circuit to chassis ground, in parallel with R_L. All these capacitances add in parallel to provide the total value labeled C_t in Fig. 9·18. Typical values for C_t are 15 to 30 $\mu\mu$f. This may seem a small value but remember that capacitive reactance is inversely proportional to frequency. For the high video frequencies the capacitive reactance of C_t is low enough to reduce

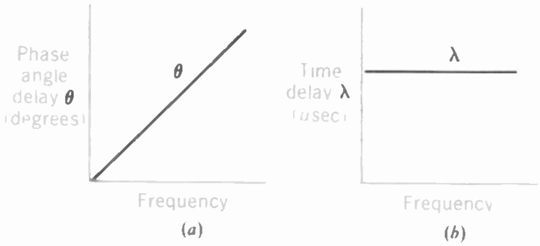


Fig. 9·17 Graphs of linear phase response in (a) and corresponding uniform time delay in (b).

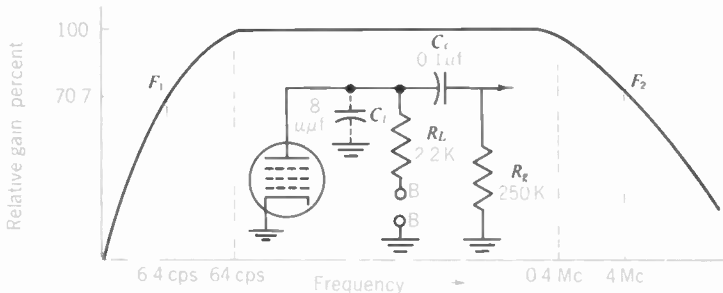


Fig. 9·18 Frequency response curve for values shown in RC-coupled amplifier.

the plate load impedance, consisting of R_L in parallel with the reactance of C_t. The gain of the amplifier decreases in the same proportion as the decrease in plate load impedance. Specifically, the gain is down to 70.7 per cent of maximum response at the frequency $F_2 = 1/(2\pi R_L C_t)$. C_t should be as small as possible, therefore, so that its reactance will be high. Tubes with little interelectrode capacitance are used and the wiring is arranged for minimum stray capacitance.

Plate load resistor. Video amplifiers use a relatively low value of R_L, typically 2,000 to 7,000 ohms. Although the gain is reduced, lowering R_L extends the high-frequency response. The effect of C_t does not become important until its reactance is low enough to be comparable with the resistance of R_L. Then the shunt reactance X_{c_t} lowers the plate load impedance. As examples, suppose R_L is 2,000 ohms and X_{c_t} is 1 megohm. The impedance of the two in parallel equals practically 2,000 ohms, the same as R_L. However, if R_L were also 1 megohm the combined impedance would be reduced to 70.7 per cent of either one.

Because R_L is small in the video amplifier, the frequency necessary to make X_{c_t} low enough to have a shunting effect becomes higher. The lower the value of R_L and the smaller the capacitance of C_t, the better is the high-frequency response.

High-frequency compensation. For typical values of R_L and C_t, video amplifiers use peaking coils such as L_o in Fig. 9·19*a* to boost the high-frequency response. A typical peaking coil is shown in Fig. 9·20. This circuit is called *shunt peaking* because L_o in the branch with R_L is in parallel with C_t. The peaking coil L_o resonates with C_t to boost the high-frequency gain, where the response of the uncompensated RC-coupled amplifier would drop off. Because its inductance is small, typically 20 to 200 μh, the

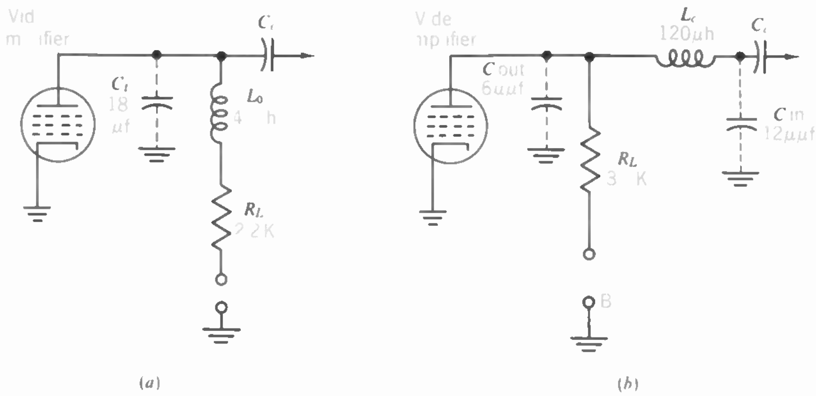


Fig. 9·19 Methods of using peaking coils in plate circuit to boost high-frequency response. (a) Shunt peaking. (b) Series peaking.

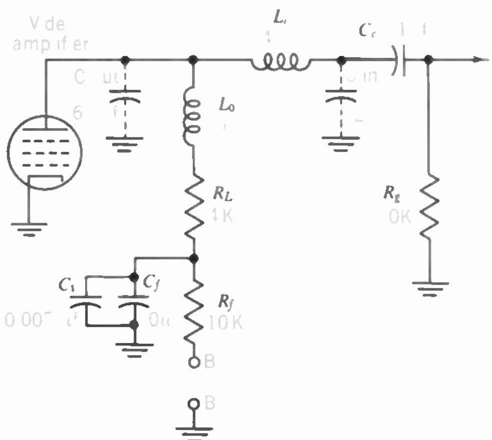
peaking coil has no effect for d-c voltage and the middle a-c frequencies. The peaking coil is generally effective for frequencies from 400 kc to 4 Mc.

In the series peaking circuit in Fig. 9·19*b*, L_c is in series with the two main components of C_t. The shunt capacitance at the plate side of L_c is the output capacitance of the amplifier. At the opposite end of L_c is the input capacitance of the next stage. The stray capacitance is included with C_{in} and C_{out} and their sum equals C_t. The series peaking circuit provides more gain than shunt peaking because there is less shunt capacitance across R_L while L_c provides a resonant rise of voltage across C_{in}.

The combination peaking circuit in Fig. 9·21 combines both the shunt and series peaking methods. Of these three methods, combination peaking allows the most gain. Shunt peaking has the least gain. Any of the

Fig. 9·20 Typical video peaking coil. Width ½ in. Inductance 180 μh.

Fig. 9·21 Video amplifier with series-shunt combination peaking for high-frequency compensation and R_fC_f decoupling filter to boost low-frequency response.



three circuits can give uniform response up to the desired frequency. The method of calculating values for these types of compensation is explained in the next chapter.

Low-frequency compensation. For the low frequencies, phase and frequency distortion result because of insufficient bypassing in the cathode and screen-grid circuits and, most important, by the increasing reactance of the grid coupling capacitor. Remember that as the frequency becomes lower the capacitive reactance increases. This is a problem for bypass and coupling capacitors at low frequencies because they should have very little reactance. With the R_gC_c-coupling circuit, specifically, the signal voltage developed across R_g for the next stage is reduced to 70.7 per cent of maximum at $F_1 = 1/(2\pi R_gC_c)$. This is shown in Fig. 9·18.

The amplifier low-frequency response can be corrected by using the largest possible values for the bypass capacitors, coupling capacitor C_c, and grid resistor R_g. In addition, the decoupling filter R_fC_f in the plate circuit of Fig. 9·21 can boost the gain and reduce phase distortion for very low frequencies. The filter has no effect on the amplifier response for middle frequencies or at the high end because then C_f serves as bypass capacitor. Note the 0.005-μf capacitor in parallel with the large 10-μf bypass. This is done in video amplifiers for effective bypassing at high frequencies where the inductance of the large tubular capacitor may be enough to result in appreciable inductive reactance.

For the very low video frequencies C_f has enough capacitive reactance to provide appreciable impedance for the R_fC_f combination. This impedance in series with R_L boosts the amplifier gain for low frequencies. The rise in gain can compensate for the reduced signal output caused by the increasing reactance of C_c. Furthermore, the phase angle shift caused by C_f is opposite from the phase angle introduced by C_c, correcting phase distortion of the signal. How to calculate values for the R_fC_f compensating filter is explained in the next chapter.

Gain of the stage. Voltage gain is the ratio of output to input signal voltage. Over the frequency range for which the response is flat, the gain of a pentode video amplifier equals g_mR_L. The g_m is grid-plate transconductance of the tube. As an example, with a g_m of 7,000 μmhos or 0.007 mho and R_L of 5,000 ohms, the gain is 35. There are no units for gain because the ohms and mhos cancel.

To increase gain, R_L should be as large as possible. However, to extend the high-frequency response, R_L must be low to minimize the shunt effect of C_t. Therefore, C_t should be as small as possible to allow a higher value of R_L with a specified high-frequency response. Higher g_m for the tube increases gain independently of the frequency response. Tubes used for video amplifiers, therefore, are generally pentodes with low interelectrode capacitances and high g_m. In this application g_m is more important than amplification factor for gain. Here the tube must produce large changes in plate current, when the grid signal varies, to have a large signal voltage across a small load resistance.

9·9 Typical video amplifier circuit

About 3 volts peak-to-peak input signal with negative sync polarity is supplied by the video detector. The video section must amplify this signal with a gain of about 25 for 75 volts video signal output. Usually, the video amplifier output has positive sync polarity for the cathode of the kinescope. Cathode drive is generally used because it requires less video signal for the kinescope, compared with grid drive. Figure 9·22 shows an oscillogram of composite video signal output of the video amplifier. The bandwidth extends up to 3.2 Mc, usually, in monochrome receivers. This limit on high-frequency response allows rejection of 3.58 Mc, which is the color subcarrier frequency transmitted for programs televised in color.

Referring to the circuit in Fig. 9·23, C_{168} and R_{171} couple the video detector output signal to the control grid of the 6CX8 pentode unit. Cathode bias of 2.4 volts is produced by R_{172}. Its bypass C_{401D} is a low-voltage electrolytic capacitor of 100 μf for low reactance at low frequencies. C_{167} in parallel is a bypass for high video frequencies. The screen grid also has an r-f bypass C_{170} in shunt with the large electrolytic C_{312}. Note that the

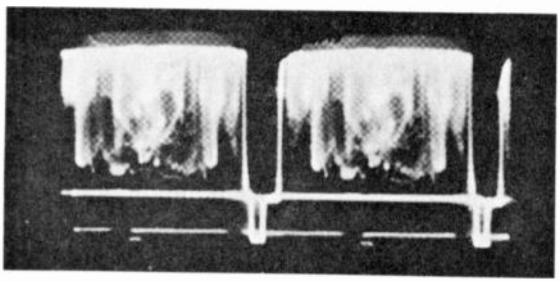
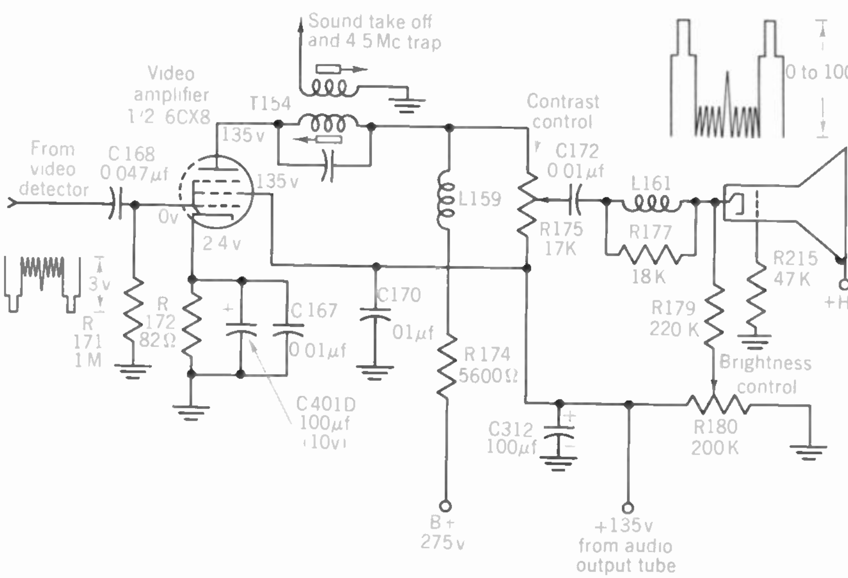


Fig. 9·22 Oscillogram of composite video signal. Oscilloscope internal sweep at 15,750/2 cps to show video signal for two horizontal scanning lines.

Fig 9·23 Typical single-stage video amplifier. D-c voltages shown at tube pins. A-c video signal oscilloscope waveforms in peak-to-peak values. (From GE Chassis M569.)



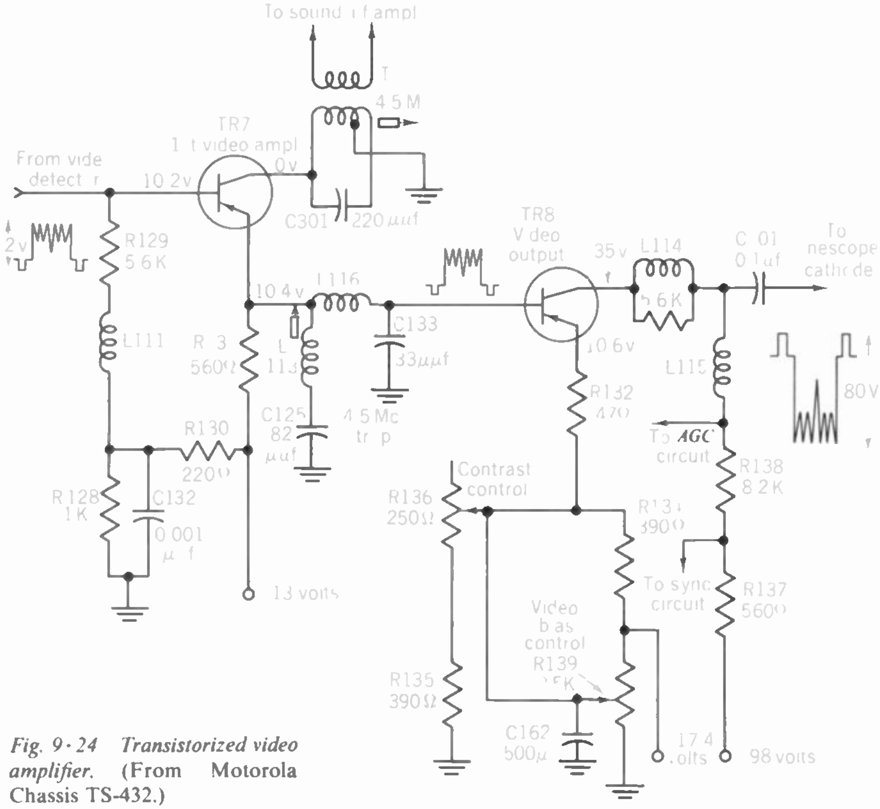


Fig. 9·24 Transistorized video amplifier. (From Motorola Chassis TS-432.)

+135-volt supply is obtained from the audio output tube cathode, in a stacked B+ arrangement.

In the plate circuit the 4.5-Mc transformer T_{154} has two functions. First, it couples 4.5-Mc sound signal to the sound i-f section in this intercarrier sound receiver. The 4.5-Mc sound signal is produced in the video detector and amplified by the pentode with the video signal. In addition, the tuned primary of T_{154} is a parallel-resonant wave trap in series with the plate load to prevent 4.5-Mc sound signal from being coupled with the video signal to the kinescope.

The 5,600-ohm R_{174} is the video amplifier plate load resistor. L_{159} and L_{161} are peaking coils to boost the high-frequency response. R_1 is a damping resistor across L_{161}. Note that the contrast control R_{175} is a potentiometer to tap off the desired proportion of video signal for the kinescope. Moving the variable arm toward the top increases the amount of plate output signal coupled by C_{172} to the kinescope cathode. R_{180} functions as the brightness control by varying the d-c bias on the kinescope.

9·10 Transistorized video amplifier

Figure 9·24 shows a circuit using two PNP transistor stages. With 2 volts video signal from the detector, the output of 80 volts is enough to drive the cathode of a conventional 19-in. kinescope. Note the d-c coupling without blocking capacitors between the detector and the transistors. The d-c voltages for emitter, base, and collector are indicated on the diagram.

The TR7 stage has two functions. It is a common-emitter amplifier for the 4.5-Mc sound signal but serves as a common-collector stage for the video signal with frequencies up to 3.2 Mc. In either case, forward bias between base and emitter is 0.2 volt. At 4.5 Mc the emitter is grounded for a-c signal by the series resonant trap with L_{113} and C_{125}. In addition, T_{301} in the collector circuit is tuned to 4.5 Mc so that parallel resonance can provide a high impedance for the sound signal output from the collector. For video frequencies below 4.5 Mc, however, the collector is effectively grounded for a-c signal by the primary of T_{301} while the emitter has the impedance of 560-ohm R_{131} and the series peaking provided by L_{116}. Now the stage functions as a common collector circuit, with input signal to the base and output signal from the emitter. This circuit is similar to a cathode follower.[1] Its characteristics are a good match from high input impedance to low output impedance. However, the gain is less than one. Also there is no phase inversion of the signal.

The TR8 video output stage is a common-emitter amplifier. R_{138} and R_{137} provide the load resistance in the collector output circuit. L_{114} with its damping resistor and L_{115} are peaking coils to boost the high-frequency response. The PNP common-emitter amplifier needs negative collector voltage for reverse bias and negative base-emitter voltage for forward bias. In the emitter circuit, the bias network supplies 10.6 volts. With respect to 10.4 volts on the base, then, the emitter has -0.2 volt forward bias. To allow degeneration of the input signal, R_{132} is not bypassed, similar to an unbypassed cathode resistance in a vacuum-tube amplifier. R_{136} controls contrast by varying the amount of degeneration. Decreasing its resistance in series with R_{139} decreases the degeneration, increasing the gain for more contrast. In the output circuit, the high collector supply of -98 volts is needed for the desired signal amplitude. Although the collector is at -35 volts, because of IR drop in its external circuit, the actual collector-emitter voltage is -45.6 volts.

Since the amplified video signal output is inverted, its positive sync polarity is correct for cathode drive of the kinescope. The composite video signal output is also coupled to the AGC circuit where the signal is rectified to provide bias for automatic control of i-f gain, and to the sync circuits where the sync pulses are clipped to provide synchronization of the deflection circuits.

9·11 Hum in the video signal

An excessive amount of 60 or 120-cps hum voltage in the video signal coupled to the kinescope produces *hum bars* on the screen (see Fig. 9·25).

Consider the case of a 60-cps sine-wave hum voltage varying the kinescope control-grid voltage in synchronism with the vertical scanning motion, as in *a*. The positive half cycle of hum voltage makes the grid more positive, increasing the beam current and screen illumination; the

[1] The cathode follower circuit is explained in Sec. 10·13.

(a) (b)

Fig. 9·25 Hum voltage at kinescope grid and horizontal bars reproduced on screen. (a) 60-cps hum. (b) 120-cps hum.

negative half cycle reduces beam current and screen illumination. Since it takes 1/120 sec for a half cycle of the 60-cps voltage, the scanning beam moves approximately halfway down the screen during this time. Therefore, if the positive half cycle of the hum voltage begins at the same time as the vertical scan, the top half of the screen will be lighter than the bottom half. The screen then has two horizontal bars, one light and the other dark.

If the frequency is 120 cps, as in *b*, two pairs of bars are produced. Then one complete cycle of the hum voltage occurs during 1/120 sec to produce a pair of black-and-white bars during one-half the vertical scan. Therefore, two cycles of the 120-cps voltage produce two pairs of bars during the field interval of 1/60 sec.

When the hum has the opposite polarity from the voltage shown in Fig. 9·25 the black-and-white bars will be reversed but there will be the same number. Often one of the hum bars appears in two sections, a part at the bottom of the screen and the remainder of the bar at the top, because the phase of the hum voltage does not coincide with the start of vertical scanning. Figure 9·26 shows how hum bars look in the picture. Since the hum in the video signal is coupled to the sync circuits, the picture also has bend caused by hum in the sync.

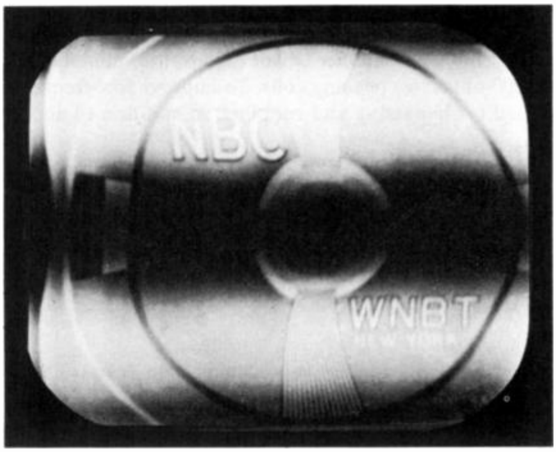


Fig. 9·26 Hum bars in picture. Bend also present because of hum in the horizontal sync. (RCA.)

Any 60- or 120-cps hum voltage in the video section of the receiver is amplified by the video stages because these frequencies are in the video-frequency range of 30 cps to 4 Mc. The hum voltage may be present in the video amplifier as the result of (1) modulation hum produced in either the r-f or i-f stages and rectified in the video detector, or (2) hum that can be introduced in the video amplifier section itself.

Hum can be produced in the video section by heater-to-cathode leakage in a tube, or because of insufficient filtering of the B supply voltage. Note that the filament hum always has a frequency of 60 cps, while hum from the B supply is 120 cps in a full-wave power supply. Hum voltage introduced in the video section itself is the additive type of hum, which does not require any modulation and is present with or without a signal. Hum bars that are in the raster without any picture, therefore, are caused by hum introduced in the video section.

SUMMARY

1. Composite video signal is coupled to the kinescope grid-cathode circuit to vary beam-current intensity. This intensity modulation reproduces the picture information. Cutoff-grid voltage means black in the picture; maximum beam current produces maximum white. About 75 volts peak-to-peak a-c video signal is required for good contrast. Average brightness is set by the kinescope d-c bias.
2. Usually, the video detector provides signal with negative sync polarity, which is inverted by one video amplifier for cathode drive at the kinescope.
3. The video stage operates as a class A, *RC*-coupled amplifier. A weak tube or incorrect d-c voltages, especially bias, can distort the signal amplitudes. Reduced a-c signal causes weak contrast.
4. The two most common methods of manual contrast control are: (*a*) variable cathode bias to control gain of the video amplifier; (*b*) variable potentiometer to tap off desired amount of plate signal voltage.
5. In the range of video frequencies from 30 cps to 4 Mc, smaller areas of picture information correspond to higher frequencies.
6. Frequency distortion means unequal gain for different frequencies. The video amplifier gain is down for very high frequencies because the shunt capacitance has decreasing reactance that bypasses the plate load resistance. The response is down for very low frequencies mainly because the voltage divider effect between R_g and the increasing reactance of C_c results in less signal coupled to the next stage.
7. Phase distortion and its resultant time-delay distortion is most important for low video frequencies, causing smear in large areas of the picture.
8. For good high-frequency response, the video amplifier circuit requires: (*a*) minimum C_t; (*b*) relatively low R_L of about 5,000 ohms; (*c*) peaking coils. To improve low-frequency response, large capacitors are used for bypassing and coupling, in addition to an *RC*-decoupling filter in series with R_L.
9. The g_m of the tube should be high since the voltage gain of a pentode with a small R_L is equal to $g_m R_L$.
10. Hum at 60 cps in the video signal coupled to the kinescope produces one pair of dark and light hum bars across the screen. With 120-cps hum, two pairs of hum bars result.

SELF-EXAMINATION *(Answers at back of book.)*

Choose (*a*), (*b*), (*c*), or (*d*).

1. The composite video signal for the video amplifier is obtained from the: (*a*) video detector; (*b*) sound i-f section; (*c*) kinescope cathode; (*d*) r-f tuner.
2. The video amplifier gain determines: (*a*) brightness; (*b*) contrast; (*c*) resolution; (*d*) detail.

3. Which of the following is true for video signal at the kinescope? (*a*) Amplitude is about 300 volts peak to peak. (*b*) Negative sync phase is required for cathode drive. (*c*) Cutoff-grid voltage corresponds to black. (*d*) Less signal allows more contrast.
4. If the video amplifier frequency response drops after 1 Mc: (*a*) Contrast will be weak. (*b*) Horizontal edge details will not be clear. (*c*) Large lettering will be weak. (*d*) Top and bottom wedges in test pattern will be too strong.
5. A video amplifier tube with a g_m of 8,000 μmhos and R_L of 4,000 ohms has a gain of: (*a*) 8; (*b*) 32; (*c*) 40; (*d*) 400.
6. Referring to the load line in Fig. 9·6, the peak-to-peak current swing is: (*a*) 2 ma; (*b*) 13 ma; (*c*) 31 ma; (*d*) 140 ma.
7. Referring to Fig. 9·23, which of the following is true? (*a*) L_{159} is a series peaking coil. (*b*) R_{171} is the cathode bias resistor. (*c*) R_{177} is the plate load resistor. (*d*) R_{174} is the plate load resistor.
8. The most important factor for good high-frequency response is: (*a*) low g_m; (*b*) large C_c; (*c*) low C_t; (*d*) low cathode bias.
9. Which of the following combinations would be best for a video amplifier? (*a*) R_L of 1 megohm, C_t 40 $\mu\mu$f; (*b*) R_L 100 ohms, C_t 200 $\mu\mu$f; (*c*) R_L 5,000 ohms, C_t 40 $\mu\mu$f; (*d*) R_L 6,000 ohms, C_t 20 $\mu\mu$f.
10. One pair of hum bars on the raster without a picture can be caused by: (*a*) 120-cps modulation hum in the r-f tuner; (*b*) 60-cps modulation hum in the i-f amplifier; (*c*) heater-cathode leakage in the video amplifier; (*d*) ripple in the B+ voltage from a full-wave rectifier.

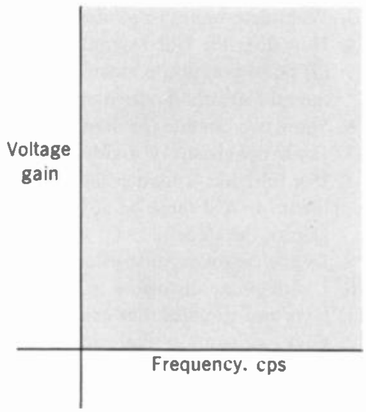
ESSAY QUESTIONS

1. What is the effect of composite video signal on: (*a*) kinescope beam current; (*b*) screen illumination?
2. What determines contrast of the picture?
3. What determines brightness of the raster?
4. How does the video signal reproduce black in the picture?
5. (*a*) Draw composite video signal of 75 volts amplitude peak to peak with the polarity required for cathode drive at the kinescope. (*b*) Do the same for grid drive.
6. Show two circuits for manual contrast control.
7. Draw the circuit of a video amplifier with series peaking and an R_fC_f-decoupling filter.
8. If a tube has a load resistance of zero ohms in the plate circuit, and there is no other load: (*a*) Will there be any polarity inversion of the signal? (*b*) How much is the voltage gain of the stage?
9. Define frequency distortion in an amplifier.
10. Define phase distortion in an amplifier.
11. Give two troubles that can cause low gain in a video amplifier.
12. Give two troubles that can cause amplitude distortion in a video amplifier.
13. What is the effect on the picture of loss of the high video frequencies?
14. Why can we say that insufficient gain for low video frequencies causes weak contrast?
15. Draw the frequency response curve for an *RC*-coupled amplifier. What are the important factors for the high-frequency end? What are the important factors for the low-frequency end?
16. If the total shunt capacitance in the video amplifier were zero, would high-frequency compensation be necessary? Why?
17. Why is phase distortion more important for low video frequencies than high frequencies?
18. Referring to the schematic in Fig. 9·23, give the function of all the components.
19. Referring to the schematic in Fig. 9·24, give the function of the following components: T_{301}, R_{132}, L_{114}, R_{138}, R_{137}, and C_{201}.
20. Referring to Fig. 9·10, draw the picture information corresponding to the 13 lines of signal shown.
21. Is the hum in Fig. 9·26 at 60 cps or 120 cps? Explain.

PROBLEMS *(Answers to odd-numbered problems at back of book.)*

1. Refer to the kinescope characteristic curve of Fig. 7·17. Assume 34 volts bias, cathode drive, and 500 volts on grid 2. List the following values: (*a*) peak-to-peak video signal for maximum contrast; (*b*) average beam current; (*c*) beam current for maximum white.
2. Referring to Fig. 9·6, draw a graph showing to correct scale the fluctuating d-c plate voltage. State the value of average d-c plate voltage and calculate the plate power dissipation in watts. How much is the peak-to-peak value of the a-c signal component? Compare this with the grid signal voltage input and calculate the gain.
3. An audio amplifier has an R_L of 250 kΩ and C_t of 40 $\mu\mu$f. At what frequency will the gain be down to 70.7 per cent of the mid-frequency response?
4. An amplifier has R_g of 1 megohm and C_c of 0.1 μf. At what frequency will the reactance of C_c equal R_g?
5. How much is the time delay in microseconds corresponding to the phase angle of 36° for the frequencies of: (*a*) 4 Mc; (*b*) 40 cps?
6. Redraw the circuit in Fig. 9·5*b* but use 2 volts cathode bias instead of 3 volts. From the load line in Fig. 9·6, then determine the following: (*a*) voltage gain; (*b*) d-c voltages at cathode, plate, and screen grid; (*c*) size of R_k and of R_s; (*d*) capacitance for C_k and for C_s to bypass 30 cps.
7. Referring to the waveshapes in Fig. 9·10, calculate the signal frequency for: (*a*) the high-frequency variations across every line; (*b*) the one cycle of low-frequency variation from line to line in each field.
8. Do the same as in question 7 but for a scanning system with 819 lines per frame and 25 frames per second.
9. Draw the frequency response curve for an *RC*-coupled amplifier having the voltage amplification values listed below.

Frequency, cps	Voltage gain
0	0
1×10^2	8
1×10^3	16
1×10^4	16
1×10^5	16
1×10^6	8
10×10^6	2



Chapter 10 Practical design of video amplifiers

In addition to voltage gain for enough contrast in the picture, the video amplifier must preserve the complex waveform of the video signal for proper reproduction of the visual information. In order to do this, there should be no appreciable frequency or phase distortion for the range of video frequencies. At the low-frequency end, starting at 30 cps, zero time delay is the main problem to prevent smear in large areas of picture information. The frequency limit at the high end is generally 2.5 to 4 Mc, the same as the bandwidth of the i-f section of the receiver. Sufficient gain for the high video frequencies is necessary to show horizontal detail in the picture. Phase distortion is less important for the high video frequencies because of the small time delay. Also, the high-frequency compensation for uniform gain reduces time-delay distortion.

10 · 1 Wide-band amplifiers

The video amplifier is a good example of a wide-band amplifier, which can be defined as a stage that amplifies both audio frequencies and radio frequencies. To see the difference in frequency response between a wide-band amplifier and a tuned amplifier with wide bandwidth, refer to Fig. 10 · 1. In *a*, the response curve is for an i-f amplifier tuned to 43 Mc. Its bandwidth is 4 Mc, meaning it can amplify the band of frequencies from 41 to 45 Mc with 70.7 per cent response or more. However, the video amplifier frequency response in *b* is entirely different, although it also has a bandwidth of 4 Mc. Here the bandwidth includes audio frequencies from 30 cps and radio frequencies up to 4 Mc. This response is really an example of amplifying a wide range of frequencies because the ratio of the highest to the lowest frequency is very great.

For wide-band amplifiers it is convenient to consider the bandwidth in octaves. An *octave* is a range of 2 to 1 in frequency. As examples: from 100 to 200 cps is one octave; 400 cps is 1 octave above 200 cps. Note that

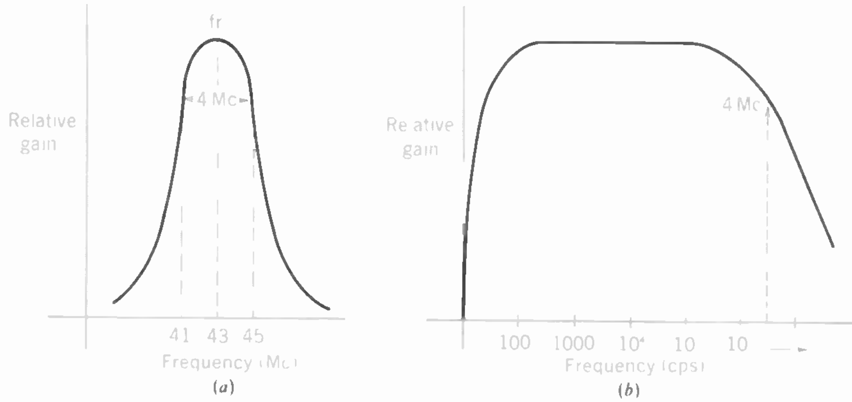


Fig. 10·1 Comparison of response curves with bandwidth of 4 Mc. (a) Tuned response centered at 43 Mc. (b) Wide-band response from low audio frequencies up to 4 Mc.

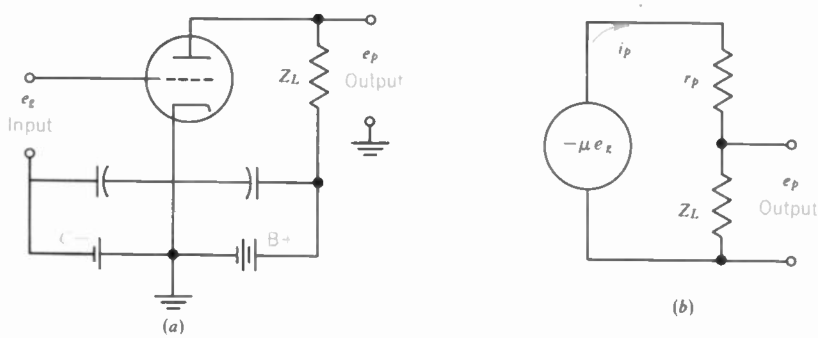
100 to 400 cps is a range of 2 octaves, as each octave increase means doubling the frequency. The general formula for the number of octaves is $N = 3.32 \times \log$ of the frequency ratio, which is the ratio of the highest frequency to the lowest frequency. For the video-frequency ratio of 3 Mc to 30 cps, equal to 1×10^5, this is a range of 3.32×5 or 16.6 octaves.

The applications of wide-band amplifiers include circuits that must amplify sine waves of audio and radio frequencies, or amplification of complex waveforms having a wide range of sine-wave frequency components. Common examples are amplifiers for the input signal in an oscilloscope, pulse amplifiers, and the video amplifier circuits in television receivers. When we analyze the requirements of video amplifiers in this chapter, the results can be applied to wide-band amplifiers in general.

10·2 Voltage amplification or gain

This factor equals the output a-c signal voltage divided by the input a-c signal voltage, usually in peak-to-peak values. As shown in Fig. 10·2, the

Fig. 10·2 Vacuum-tube amplifier circuit. (a) Typical operation. (b) Equivalent plate circuit.



output is e_p and the input is e_g. The equivalent circuit shows the tube as a constant-voltage generator supplying amplified voltage equal to $\mu \times e_g$. The μ is the amplification factor of the tube. The input e_g is multiplied by μ to increase the signal voltage in the plate circuit. This equivalent circuit applies for any signal waveshape, as long as the tube is operating as a class A amplifier over the linear portion of its characteristic curve.

However, the μe_g voltage is not the output signal. Actually, the μe_g voltage is divided between the internal r_p of the tube and the external plate load impedance Z_L. Only the voltage across Z_L is available as signal output. Therefore, the voltage amplification, equal to e_p/e_g, can be stated as

$$A_v = \mu \frac{Z_L}{r_p + Z_L} \tag{10·1}$$

This formula shows that the voltage gain A approaches the value of μ but is actually less because of the voltage divider effect between r_p and Z_L. The gain is μ times the proportion of Z_L to the total of Z_L plus r_p. The higher Z_L is, the greater is the proportion of μe_g voltage that can be developed across Z_L for output signal, and the higher is the gain.

This formula can be used to calculate gain for triode circuits with a known value of Z_L, where μ and r_p are given in the tube manual. As an example, for a μ of 50 and both r_p and Z_L equal to 40,000 ohms, A equals 25. The gain is one-half the μ because r_p and Z_L are equal, providing a voltage divider that splits the μe_g voltage into two equal parts.

For pentode tubes, r_p is usually very high, commonly about 1 megohm. Z_L usually cannot be made that large because the d-c voltage drop in the plate circuit would reduce the plate voltage too much. For the case where Z_L is very small with respect to r_p in the denominator of Eq. (10·1) it becomes

$$A_v = \mu \frac{Z_L}{r_p} = \frac{\mu}{r_p} \times Z_L$$

However, μ/r_p is the transconductance (g_m). Therefore,

$$A_v = g_m Z_L \tag{10·2}$$

As an example, if g_m is 7,000 μmhos, or 0.007 mho, and Z_L is 5,000 ohms, the gain equals 35. This formula also applies for any signal waveshape as long as the tube is operating as a linear class A amplifier. However, Z_L must be low compared with r_p. Otherwise Eq. (10·1) is used to calculate the gain.

Equation (10·2) fits the case of video amplifier circuits because a pentode is generally used, with a Z_L that is very low compared with r_p. Pentodes are used for their low grid-plate capacitance, which is important at the high video frequencies. A low value of Z_L is required for good high-frequency response. In practical terms, the $g_m Z_L$ gain formula indicates that the g_m of the tube is more important than μ for gain when the external plate load impedance is relatively low. Then what is needed is high g_m so that e_g can produce large variations in plate current across the small Z_L to produce a large signal output voltage e_p.

In any case, the voltage amplification increases or decreases with the value of Z_L. In order to have uniform gain, or flat frequency response in the video amplifier, the product of $g_m \times Z_L$ must be constant through the desired frequency range. The g_m value is a constant for the tube, depending only on plate voltage and grid bias. The Z_L value can vary with frequency, however. Although a resistance plate load R_L is used for wide frequency response, there is also shunt capacitance C_t in the plate circuit. At high frequencies the reactance of C_t combines with R_L to provide a complex plate load impedance Z_L that varies with frequency. As the value of Z_L changes for different frequencies, the gain of the amplifier will vary in exactly the same way.

10·3 Shunt capacitance

C_t is the only reason why the video amplifier response is down for high frequencies, as the shunt capacitive reactance reduces Z_L. It is very important, therefore, to keep C_t as small as possible. Remember that a small C_t means high capacitive reactance, as C and X_c are inversely proportional. Typical values of C_t are 15 to 40 $\mu\mu$f or pf. Both $\mu\mu$f for micromicrofarad and pf for picofarad are equal to 10^{-12} farad.

Components of C_t. As shown in Fig. 10·3, the following parallel capacitances are added to form the total C_t:

$$C_t = C_{\text{stray}} + C_{\text{out}} + C_{\text{in}} + C_m \qquad (10\cdot3)$$

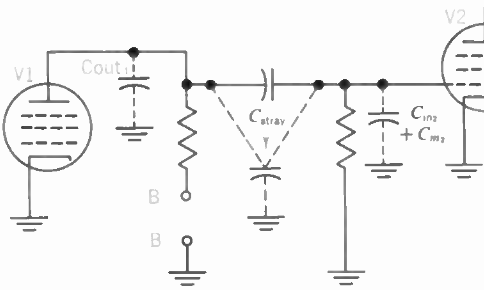


Fig. 10·3 The shunt capacitances in an amplifier.

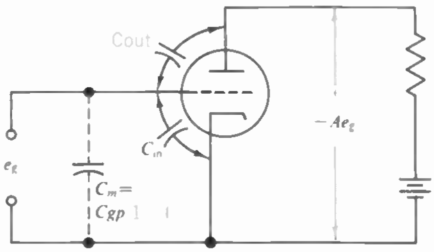


Fig. 10·4 The Miller effect capacitance C_m adds to C_{in}.

C_{stray} includes capacitance to chassis ground of the wiring, component parts such as the coupling capacitor, plate load resistor, and peaking coils, in addition to capacitances between the tube elements through the tube socket. These are not included in the tube manual listing of the capacitances but can add appreciably to C_t. The stray capacitances can be held to a minimum, greatly improving the amplifier's high-frequency response, by short wiring, use of low-capacitance sockets, and proper placement of parts. The coupling elements should be mounted away from and perpendicular to the chassis to reduce C_{stray}. A rough estimate of the stray capacitance in a video amplifier stage is about 5 $\mu\mu$f.

C_{out} is the capacitance from plate to cathode of the tube itself. As an example, the 6AC7 has an output capacitance of 5 $\mu\mu$f. Similarly, C_{in} is the static input capacitance between grid and cathode. This is taken for the next tube, since the input circuit of V_2 is part of the plate load for V_1. The 6AC7 has an input capacitance of 11 $\mu\mu$f. In the case of a video stage driving the kinescope, its input capacitance must be included. C_{in} for kinescopes is 6 $\mu\mu$f with grid drive, or 5 $\mu\mu$f with cathode drive. These values apply for practically all picture tubes.

C_m is a dynamic input capacitance, which is added to the static value of C_{in} for an amplifier. This increase of input capacitance in an amplifier stage, called *Miller effect*, can add appreciable capacitance to C_{in} for a triode, or a pentode having high gain.

Miller effect. The amount of input capacitance added by the Miller effect can be derived as

$$C_m = C_{gp}(1 + A) \tag{10·4}$$

where A is the gain and C_{gp} is grid-plate capacitance (see Fig. 10·4). Note that C_m increases with more C_{gp} and high gain. The dynamic input capacitance of an amplifier, therefore, is the sum of the static C_{in} plus the C_m added to the grid circuit when the tube is amplifying signal.

Example 1. Find the total input capacitance of a 6CL6 amplifier with a gain of 14. From the tube manual, C_{in} is 11 $\mu\mu$f and C_{gp} is 0.12 $\mu\mu$f.

$$\begin{aligned} C &= C_{in} + C_{gp}(1 + A) \\ &= 11 + 0.12(1 + 14) \\ &= 11 + 1.8 \\ &= 12.8\ \mu\mu\text{f} \end{aligned}$$

Example 2. Find the total input capacitance of the triode amplifier 6J5 with a gain of 14. From the tube manual C_{in} is 3.4 $\mu\mu$f and C_{gp} is 3.4 $\mu\mu$f.

$$\begin{aligned} C &= C_{in} + C_{gp}(1 + A) \\ &= 3.4 + 3.4(1 + 14) \\ &= 3.4 + 51.0 \\ &= 54.4\ \mu\mu\text{f} \end{aligned}$$

From these examples it is seen that with a pentode tube the additional input capacitance resulting from the Miller effect is small because of the small C_{gp}. In a triode, however, the high value of grid-to-plate capacitance reflects back into the grid circuit when the tube is used as an amplifier stage, greatly increasing the dynamic input capacitance. When the gain of a stage is very high, though, even a small C_{gp} can increase the input capacitance appreciably. Kinescopes do not have any Miller effect capacitance, because there is no amplification of signal.

Calculating C_t. The first step in designing the video amplifier is to find the value of the shunt capacitance, since this is a direct measure of the high-frequency response of the amplifier and how much compensation is needed. C_t can be calculated, if desired, as the sum of the individual capacitances. The values of C_{out}, C_{in}, and C_{gp} are available from the tube manual, and the gain of the next stage can be either measured or calculated as $g_m \times R_L$.

Example. How much is C_t for a 6AU6 video amplifier driving a 6K6GT video output tube whose gain is 10? The stray capacitance is 5 $\mu\mu$f.

$$
\begin{aligned}
C_t &= C_{out} + C_{stray} + C_{in} + C_{gp}(1 + A) \\
&= 5 + 5 + 5.5 + 0.5(1 + 10) \\
&= 5 + 5 + 5.5 + 5.5 \\
&= 21\ \mu\mu\text{f}
\end{aligned}
$$

Measuring C_t. Better than calculating C_t is finding its value experimentally by resonating with a known value of inductance in the plate circuit of the video amplifier. Sine-wave signal is coupled to the video amplifier input and the amplified output is measured with an a-c voltmeter capable of measurements at the high video frequencies used. Varying the frequency of the applied signal, the output voltage will show a marked increase at resonance when the reactance of the inductance equals the reactance of C_t. Normally, no output reading is obtained until resonance is reached because the coil is inserted in the plate circuit of the amplifier in place of the

Fig. 10·5 Plate load impedance Z_L of uncompensated RC-coupled amplifier. (a) Actual circuit. (b) Equivalent circuit.

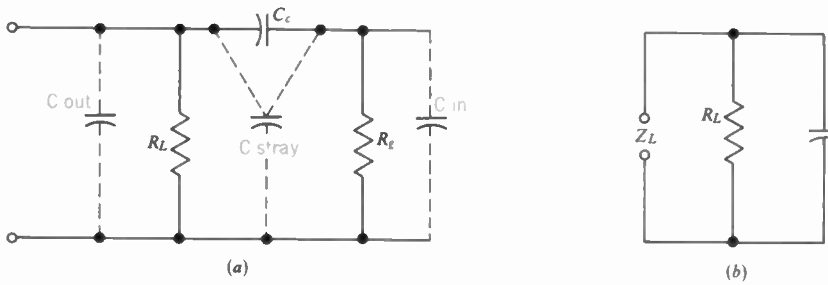


plate load resistor. R_L is omitted when making this measurement in order to obtain sharper resonance. The coil should be similar to the actual peaking coil that will be used for the high-frequency compensation. The circuit for this measurement is the same as in Fig. 10·18.

With the value of inductance and the resonant frequency known, C_t can be computed from the resonant frequency formula

$$f_r = \frac{1}{2\pi\sqrt{LC}}$$

Squaring both sides,
$$f_r^2 = \frac{1}{4\pi^2 LC}$$

Transposing,
$$C = \frac{1}{4\pi^2 f_r^2 L} \tag{10·5}$$

As an example, resonance at 1.6 Mc with a 350-μh coil means C_t equals 28.6 $\mu\mu$f. With C_t known, the high-frequency response of the amplifier can be analyzed.

10·4 High-frequency response of the uncompensated amplifier

At the high-frequency end of the video band, the plate load impedance Z_L is equal to the plate resistor R_L in parallel with the total shunt capacitance C_t (see Fig. 10·5). That this is so can be seen from the following:

1. The reactance of the coupling capacitor C_c is negligibly small for the high video frequencies, since C_c is generally about 0.1 μf.
2. Then all the individual input and output capacitances are in parallel with the plate resistor R_L and grid resistor R_g.
3. However, R_L is usually quite small, while R_g is at least 100,000 ohms. The parallel combination of R_L and R_g, therefore, is practically equal to the resistance of R_L, the smaller one. As a result, Z_L is equal to R_L in parallel with X_{c_t}, as shown in Fig. 10·5*b*.

How Z_L decreases at higher frequencies. We can use numerical values to show how the decreasing reactance of C_t reduces Z_L for high video frequencies. Assume 18 $\mu\mu$f for the shunt capacitance C_t, with R_L 2,200 ohms. To find Z_L, first calculate the capacitive reactance X_{c_t}. The formula $X_c = 1/(2\pi fC)$ gives the reactance of C_t at any one frequency. Then combine this value of X_{c_t} with the fixed value of R_L, in parallel. They must be combined vectorially, however, because X_{c_t} and R_L have branch currents that are 90° out of phase. The results are shown in Table 10·1, with X_{c_t} and Z_L calculated for five different frequencies, including the phase angle θ for Z_L.

The negative angle θ for Z_L in the plate circuit means that i_p leads e_p. However, the output signal voltage is e_p, which lags by the angle θ. When θ is converted to time, therefore, this value is the amount of time delay for signal voltage at a particular frequency.

At low frequencies and mid-frequencies, the reactance X_{C_t} is so large it has practically no shunting effect. Then Z_L has the same value as R_L. If we take the frequency of 10 kc as an example, X_{C_t} is 880,000 ohms, as listed in the top row of Table 10·1. The resultant Z_L of 2,200 ohms R_L in parallel with 880,000 ohms shunt reactance is practically equal to 2,200 ohms. For 400 kc in the next row, Z_L is still 2,200 ohms as X_{C_t} is ten times R_L.

At the frequency of 2 Mc, however, X_{C_t} is 4,400 ohms. Since this reactance is not much greater than R_L, now X_{C_t} has a shunting effect that reduces Z_L to a value less than R_L. For the higher frequencies listed at 4, 8, and 16 Mc, X_{C_t} continues to decrease and Z_L is reduced even further. Remember that the gain of the amplifier decreases in exactly the same way Z_L is reduced because the gain equals $g_m Z_L$.

Definition of F_2. This is the frequency at which $X_{C_t} = R_L$. In Table 10·1, the frequency of 4 Mc in the middle row is F_2 because the reactance of the shunt capacitance is down to 2,200 ohms, exactly equal to the 2,200-ohm R_L. This frequency is important because it is a convenient point for marking the frequency response.

For any parallel combination of resistance and reactance, when they are equal their total impedance equals 0.707 or 70.7 per cent of either one. In Table 10·1, Z_L is 0.707 × 2,200 ohms or 1,556 ohms. Therefore, Z_L drops to 70.7 per cent of R_L at the frequency labeled F_2. Or, we can say F_2 is the frequency at which Z_L is 0.707 R_L.

The gain of the amplifier is $g_m Z_L$. At mid-frequencies Z_L has the same values as R_L. Therefore, the mid-frequency gain is $g_m R_L$. This value is taken as 100 per cent response. In Table 10·1 the $g_m R_L$ value is equal to a gain of 22, for the flat response at mid-frequencies.

At higher frequencies the gain drops because Z_L is reduced by the shunting effect of X_c. At F_2, the plate impedance Z_L is 0.707 R_L. Then the gain at F_2 equals 70.7 per cent of the mid-frequency gain. In Table 10·1, the $g_m R_L$ value is equal to a gain of 15.56 at F_2, which is 70.7 per cent of 22.

The reduction of gain from 100 to 70.7 per cent is a loss of 29.3 per cent.

Table 10·1 How Z_L decreases with frequency in uncompensated amplifier

R_L = 2,200 ohms and C_t = 18 μμf in Fig. 10·5b

	Frequency	R_L, ohms	X_{C_t}, ohms	Z_L, ohms	θ	Gain = $g_m{}^* Z_L$
	10 kc	2,200	880,000	2,200	0°	22
	400 kc	2,200	22,000	2,200	−5.7°	22
	2 Mc	2,200	4,400	1,970	−26.6°	19.7
F_2 →	4 Mc	2,200	2,200	1,556	−45°	15.56
	8 Mc	2,200	1,100	990	−63.4°	9.8
	16 Mc	2,200	550	530	−76°	4.4

* Gain values for g_m of 10,000 μmhos.

Stated in decibels, this voltage ratio corresponds to a loss of 3 db. Therefore, F_2 can be considered the frequency at which the high-frequency response of the uncompensated resistance-loaded amplifier is down 3 db, compared with the gain at mid-frequencies.

The gain of the uncompensated video amplifier is down 3 db at F_2 because X_{c_t} equals R_L at that frequency. The reactance of C_t at F_2 must equal

$$X_{ct} = \frac{1}{2\pi F_2 C_t}$$

This value of X_{c_t} is equal to R_L by the definition of F_2. Therefore,

$$R_L = \frac{1}{2\pi F_2 C_t} \qquad (10\cdot 6)$$

or transposing,

$$F_2 = \frac{1}{2\pi R_L C_t} \qquad (10\cdot 7)$$

or transposing,

$$C_t = \frac{1}{2\pi R_L F_2} \qquad (10\cdot 8)$$

R is in ohms, C in farads, and F_2 in cps in all three formulas.

The value of F_2 is inversely proportional to R_L and C_t. In order to have a high value of F_2 and good high-frequency response, both R_L and C_t must be small.

Example 1. An uncompensated video amplifier has a plate load resistor of 4,000 ohms and a total shunt capacitance of 18 $\mu\mu$f. At what frequency is the gain down 3 db? What is the mid-frequency gain, using a tube with a g_m of 6,000 μmhos?

$$\begin{aligned} F_2 &= \frac{1}{2\pi R_L C_t} \\ &= \frac{1}{2\pi \times 4{,}000 \times 18 \times 10^{-12}} \\ &= 2.2 \times 10^6 \text{ cps (approx)} \\ &= 2.2 \text{ Mc} \\ \text{Gain} &= g_m \times R_L \\ &= 0.006 \times 4{,}000 \\ &= 24 \end{aligned}$$

Example 2. What value of R_L is needed to make F_2 equal 4 Mc for the amplifier of Example 1? What is the mid-frequency gain using a tube with a g_m of 10,000 μmhos?

$$\begin{aligned} R_L &= \frac{1}{2\pi F_2 C_t} \\ R_L &= \frac{1}{2\pi \times 4 \times 10^6 \times 18 \times 10^{-12}} \\ &= 2{,}200 \text{ ohms (approx)} \\ \text{Gain} &= g_m \times R_L \\ &= 0.01 \times 2{,}200 \\ &= 22 \end{aligned}$$

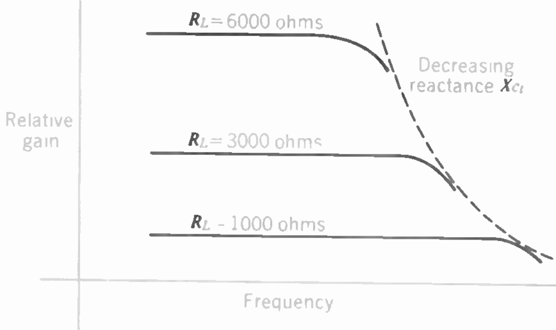


Fig. 10·6 Relative gain vs. frequency for different values of plate load resistance R_L.

Note that these are the same values used for the calculations in Table 10·1. In general, C_t is made as small as possible with special attention to the strays, and R_L is given the value necessary for a desired value of F_2.

While it would seem that any value of F_2 can be obtained by suitable choice of R_L, the smaller the R_L the less the gain. It is much better to obtain a high value of F_2 by keeping C_t down to a minimum. In this way, R_L can have a higher value for a given F_2 and the gain of the stage will be greater. The effect of different values of R_L on the amplifier response curve, with a fixed value of C_t, is shown in Fig. 10·6.

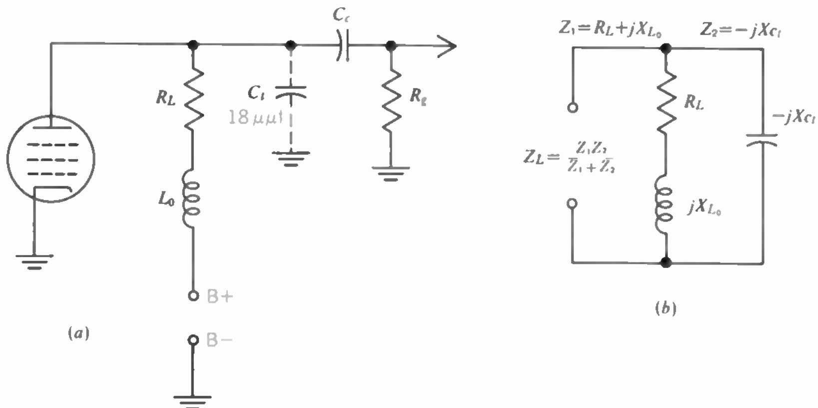
It can be useful to note that the uncompensated amplifier has flat frequency response only up to $0.1F_2$. This response is shown by curve 1 in Fig. 10·12. At this frequency, X_{c_t} is ten times R_L and the parallel reactance has practically no shunting effect. As an example, if F_2 is 4 Mc, the amplifier is flat to 0.4 Mc or 400 kc without compensation.

The higher the value of F_2 the better is the high-frequency response. However, the video amplifier cannot be used up to F_2 without compensation. At F_2 the gain is down 29.3 per cent, or 3 db. A variation of more than 10 per cent, or 1 db, is generally not considered flat response. Even so, F_2 is an important frequency. When the video amplifier is compensated with the procedures described here, the design values are calculated for flat response up to F_2, instead of being 3 db down at this frequency without compensation.

10·5 Shunt peaking

Once C_t is known, R_L can be chosen for the desired value of F_2, using Eq. (10·6). This frequency is chosen as the limit up to which the video

Fig. 10·7 Shunt peaking. (a) Circuit with L_o in shunt with C_t to boost high-frequency response. (b) A-c equivalent plate load impedance Z_L.



amplifier response is desired to be flat, because the high-frequency compensation is designed to make the amplifier response uniform up to F_2. The simplest method of doing this is with the shunt peaking circuit shown in Fig. 10·7. The peaking coil L_o is in the branch with R_L shunting C_t. Although R_L and L_o can be reversed, the connection shown reduces the stray capacitance of L_o. The peaking coil resonates with C_t, canceling its shunt reactance, to allow flat response up to F_2, instead of the 3-db loss at this frequency without compensation.

The values required for shunt peaking are

$$R_L = X_{c_t} = \frac{1}{2\pi F_2 C_t} \tag{10·9}$$

$$X_{L_o} = 0.5 R_L \tag{10·10}$$

Note that X_{L_o} is the inductive reactance, in ohms. To find the inductance of L_o in henrys, Eq. (10·10) can be converted to

$$L_o = 0.5\, C_t R_L^2 \tag{10·11}$$

In all three formulas the units for R and X are in ohms, C in farads, L in henrys, and F in cps. These values for shunt peaking make R_L equal to the capacitive reactance of the shunt C_t, at the desired F_2, while the inductive reactance of L_o is one-half either R_L or X_{c_t}. Then, the effect of X_{L_o} is to maintain the total plate load impedance Z_L essentially equal to R_L up to F_2 because of parallel resonance. With Z_L at the same value as R_L, the gain is uniform up to F_2.

Plate load impedance. With the peaking coil inserted in series with R_L, the equivalent plate load illustrated in Fig. 10·7*b* takes a form different from the case of the uncompensated amplifier. The impedance Z_1 of the R_L branch is the vector sum of $R_L + jX_{L_o}$. The parallel Z_2 branch is $-jX_c$. When Z_1 and Z_2 are combined in parallel, as $Z_1Z_2/(Z_1 + Z_2)$, their resultant impedance Z_L can be calculated as in Table 10·2.

Table 10·2 Z_L values for shunt peaking circuit

R_L = 2,200 ohms, C_t = 18μμf, and L_o = 43.6 μh in Fig. 10·7*b*

	Frequency f	f/F_2	R_L, *ohms*	X_{c_t}, *ohms*	X_{L_o} *ohms*	Z_L, *ohms*	θ	*Gain* = g_m* Z_L
	10 kc	0.0025	2,200	880,000	2.75	2,200	0°	22
	400 kc	0.1	2,200	22,000	110	2,200	−5.5°	22
	2 Mc	0.5	2,200	4,400	550	2,250	−15.6°	22.5
F_2 →	4 Mc	1.0	2,200	2,200	1,100	2,200	−36.8°	22
	8 Mc	2.0	2,200	1,100	2,200	1,400	−71.6°	14
	16 Mc	4.0	2,200	550	4,400	610	−86.7°	6.1

* Gain values for g_m of 10,000 μmhos.

A graph of impedance or relative voltage amplification against frequency for the shunt-peaked amplifier is shown by curve 2 in Fig. 10·12. Note that frequencies on the horizontal axis are in terms of the ratio f/F_2, instead of actual frequency. This method allows the curves to be used universally for amplifiers having different values of F_2. As an example, if F_2 of the amplifier is 4 Mc and it is desired to know the response at 2 Mc it can be read from the curve at $f/F_2 = 0.5$. At this point the relative gain for curve 2 is 1.0, or 100 per cent response, compared with 89 per cent response in the uncompensated amplifier. Another shunt-peaked amplifier with F_2 at 6 Mc would have this same relative response at 3 Mc, which is also $0.5F_2$.

The graph in Fig. 10·12 uses logarithmic spacing on the horizontal axis, with values increasing in multiples of 10. This method is generally used for frequency-response curves because the logarithmic spacing compresses the range of frequencies. Then the lowest and highest values can be included without extending the graph too far.

Although the high-frequency boost does not take much space on the graph, numerically the compensation extends the high-frequency response by the factor of 10, from $0.1F_2$ up to F_2. This range includes almost all the high frequencies. The reason for the increase in impedance for high frequencies and rise in gain over the uncompensated circuit is that the peaking coil resonates with C_t to form a parallel resonant circuit broadly tuned to the high-frequency end of the video band. The resonant circuit has very low Q because of the damping effect of R_L in series with L_o in the tuned circuit. The resonant frequency of the peaking circuit, when $X_L = X_c$, is not at $0.6F_2$ where the maximum rise in impedance occurs but is at some frequency higher than F_2. With such a high series resistance, and low Q, the tuned circuit has a resonance curve that is neither symmetrical nor sharply peaked, and the point of maximum impedance is not at the frequency where the reactances are equal.

While these design values are not the only possibilities, they are suitable for obtaining uniform response in the shunt-peaked amplifier. If L_o is too large, the Q of the compensated circuit is too high and the response rises too sharply just below F_2. Also, the phase distortion increases. If L_o is too small the plate impedance will not increase enough and the gain drops below F_2. Reducing L_o to the extreme case of zero inductance gives the same results as the uncompensated amplifier. The values given, then, represent a good compromise for flat response up to F_2.

Design procedure for shunt peaking. In practice, the procedure is this:

1. The highest frequency up to which the amplifier is desired flat is chosen as F_2. This may be from 2.5 to 4 Mc or higher, depending on the use of the video amplifier.
2. C_t is calculated or preferably measured in the chassis.
3. With C_t known, R_L is chosen as equal to the reactance of C_t at the top correction frequency F_2 as decided in 1 above: $R_L = 1/(2\pi F_2 C_t)$.
4. The peaking coil L_o has an inductance $L_o = 0.5C_t R_L^2$.

5. The gain of the stage for mid-frequencies and very high frequencies up to F_2 is now equal to $g_m R_L$, approximately, since the plate impedance is essentially equal to R_L up to the frequency F_2 with the added compensation.

Example. An amplifier is desired to be flat to 4 Mc. C_t is 18 $\mu\mu$f. What are the sizes of R_L and L_o required? What is the gain of the stage when a tube with a g_m of 10,000 μmhos is used?

1. F_2 is 4 Mc, the top correction frequency.
2. The reactance of C_t at F_2 is

$$\begin{aligned} X_{c_t} &= \frac{1}{2\pi F_2 C_t} \\ &= \frac{1}{2\pi \times 4 \times 10^6 \times 18 \times 10^{-12}} \\ &= 2{,}200 \text{ ohms (approx)} \end{aligned}$$

3. R_L is 2,200 ohms, therefore, since it equals the reactance of C_t at F_2. The value for R_L is taken to the nearest 100 ohms.

4.
$$\begin{aligned} L_o &= \tfrac{1}{2} C_t R_L^2 \\ &= \tfrac{1}{2} \times 18 \times 10^{-12} \times (2{,}200)^2 \\ &= 43.6 \times 10^{-6} \text{ henry (approx)} \\ &= 43.6\ \mu\text{h} \end{aligned}$$

5.
$$\begin{aligned} \text{Gain} &= g_m \times R_L \\ &= 0.01 \times 2{,}200 \\ &= 22 \end{aligned}$$

If the 2,200-ohm plate load resistor is used with the 43.6-μh inductance in the shunt peaking circuit, having a C_t of 18 $\mu\mu$f, the amplifier response will be essentially flat to 4 Mc with a gain of 22. Note that these are the values used for the shunt peaking in Table 10·2.

Experimental procedure for determining L_o. The reactance of L_o is one-half the reactance of C_t at the frequency F_2. Therefore, the two reactances will be equal at some frequency higher than F_2, as X_L increases and X_c decreases with increasing frequency. The reactances are equal at the resonant frequency $f_r = 1.4F_2$.

For an F_2 of 4 Mc, as an example, f_r is $1.4 \times 4 = 5.6$ Mc. This can be checked by referring back to the previous example calculated for a 4-Mc amplifier with C_t 18 $\mu\mu$f and L_o 43.6 μh for shunt peaking. If these values are substituted for L_o and C_t in the resonant frequency formula $f_r = 1/(2\pi\sqrt{L_o C_t})$, the answer is 5.6 Mc for f_r. This shows that resonating L_o at 1.4 F_2 provides the inductance equal to $0.5C_t R_L^2$.

To resonate L_o with C_t connect the coil in place of R_L, temporarily, in order to obtain sharp resonance at the correct frequency (see Fig. 10·8). Use a larger coil to start, and take off turns until the inductance resonates with C_t at 1.4 F_2. After the correct inductance has been obtained, connect the coil in series with R_L. Then the shunt-peaked amplifier will have flat frequency response up to F_2, with a gain equal to $g_m R_L$.

10·6 Series peaking

As shown in Fig. 10·9*a*, the series peaking coil, labeled L_c, is connected in series with the signal path to the next stage. This connection separates C_t into two smaller parts. C_{out} is the shunt capacitance at the plate side of L_c and C_{in} is at the opposite end. The sum of C_{out} plus C_{in} equals C_t. However, the capacitance division allows higher gain while still extending the flat frequency response up to F_2. Note that L_c forms a π-type filter with C_{out} and C_{in}.

The reason for the higher gain is a higher R_L, while L_c resonates with C_{in} at F_2. This is an example of series resonance because L_o and C_{in} are in the same path for signal current. Remember that, for series resonance, the voltage across either X_L or X_c increases because of the resonant rise of current. This voltage rise across C_{in} compensates for the reduced $g_m Z_L$ value at high frequencies near F_2, where the reactance of C_{out} shunts R_L. The result is uniform signal voltage to the next grid circuit, up to F_2. See curve 3 in Fig. 10·12.

Since C_t is usually easier to determine than C_{in}, we can specify the required values for series peaking as follows:

$$R_L = \frac{1.5}{2\pi F_2 C_t} \qquad (10\cdot 12)$$

$$L_c = \tfrac{2}{3}\, C_t R_L^2 = 0.67\, C_t R_L^2 \qquad (10\cdot 13)$$

R is in ohms, C in farads, L in henrys, and F in cps. Note that these values are in terms of C_t, not C_{in}. This conversion is based on a ratio of C_{in} twice C_{out}. This ratio makes C_t equal to $\tfrac{3}{2}$ C_{in} or 1.5 C_{in}. Usually, this ratio ap-

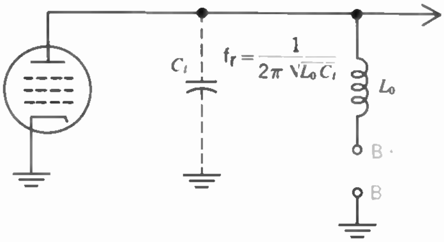


Fig. 10·8 Resonating L_o with C_t.

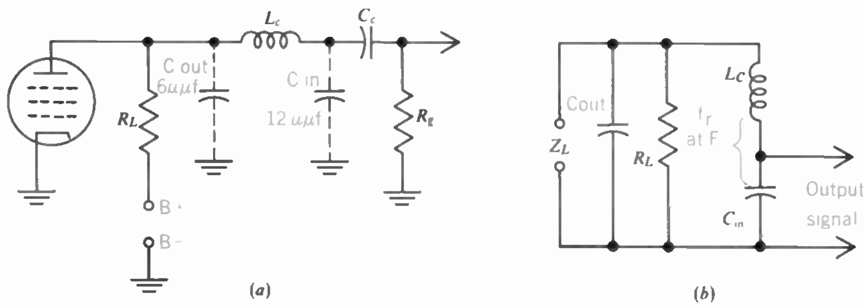


Fig. 10·9 Series peaking. (a) Circuit for $C_{in} = 2\ C_{out}$. (b) A-c equivalent plate load.

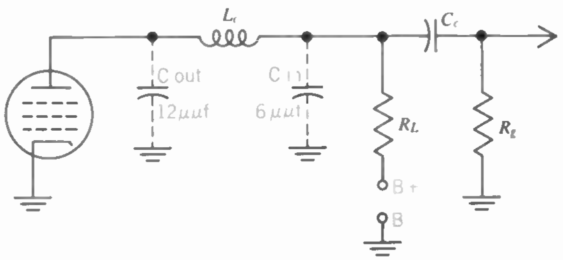


Fig. 10·10 Series peaking circuit with R_L across C_{in} because it is ½ C_{out}.

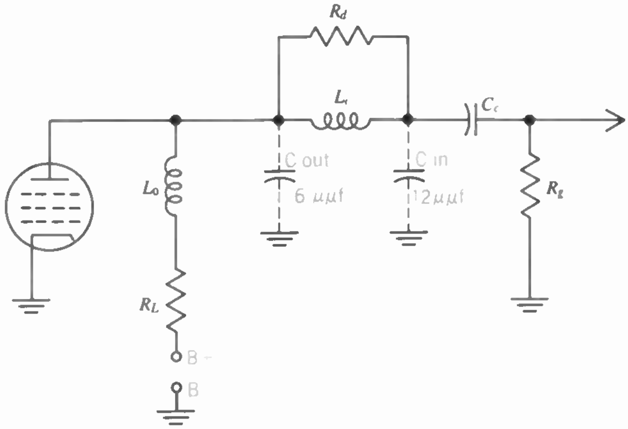


Fig. 10·11 Combination peaking circuit combining the shunt and series methods.

plies because of the higher input capacitance of tubes. Small changes in the ratio can be made, if necessary, by adding the stray capacitance of C_c to either the plate or grid side of L_c.

If the capacitances are reversed, with C_{out} twice C_{in}, the circuit looks like Fig. 10·10. The rule is to keep R_L on the low capacitance side of the filter. If the capacitances do not have the 2:1 ratio, L_c is given the value that makes X_{L_c} equal $X_{c_{in}}$ at F_2. This reactance, in ohms, is also the required value for R_L.

The plate load R_L for series peaking is 50 per cent more than R_L for shunt peaking with the same F_2. Therefore, the gain for the entire video-frequency range up to F_2 is one and one-half times higher than in shunt peaking. For the same example calculated for shunt peaking in Sec. 10·5, if series peaking were used R_L would be 3,300 ohms, L_c 120 μh, and the gain 33 with flat frequency response up to 4 Mc.

When the Q of the series resonant circuit is too high, usually with a small value of C_{in}, an excessive peak in the response may result at F_2. Then a damping resistor of about 20,000 ohms is connected across L_c, as shown in Fig. 10·11. This is not necessary for L_o because it is damped by the series resistance of R_L.

10·7 Series-shunt combination peaking

When the filter coupling of L_c is combined with L_o in the R_L branch, as shown in Fig. 10·11, the plate load resistor can be increased over either of

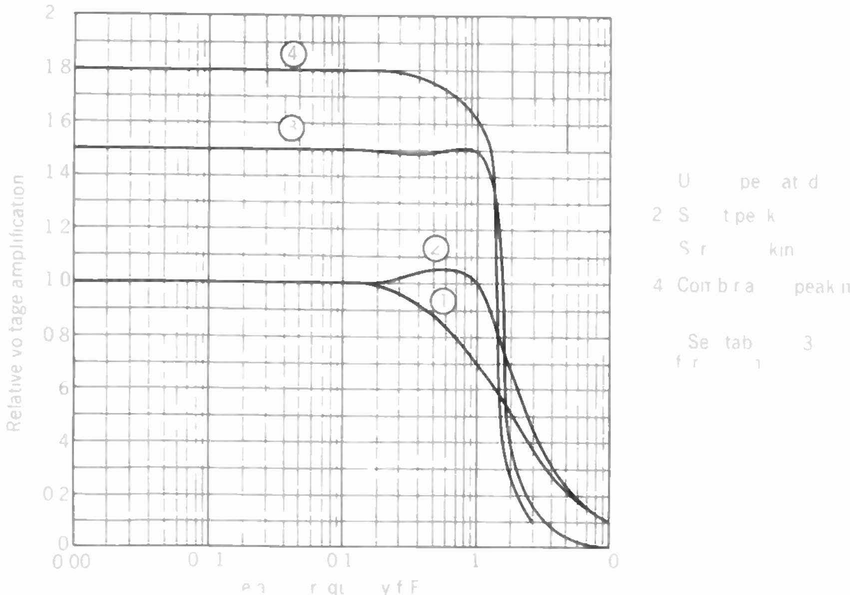


Fig. 10·12 Universal frequency response curves for the video amplifier.

the previous methods for still more gain. Design values for the combination peaking are

$$R_L = 1.8X_{c_t} \text{ at } F_2, \text{ or } R_L = \frac{1.8}{2\pi F_2 C_t} \qquad (10\cdot 14)$$

$$L_c = 0.52C_t R_L{}^2 \qquad (10\cdot 15)$$

$$L_o = 0.12C_t R_L{}^2 \qquad (10\cdot 16)$$

With the higher value of R_L, the gain of the amplifier is 80 per cent higher than in shunt peaking, and the response is still uniform within 10 per cent up to F_2. See curve 4 in Fig. 10·12. For the same example calculated for shunt peaking in Sec. 10·6, the plate load resistor would be 3,960 ohms, L_c 145 μh, L_o 35 μh, approximately, and the gain of the stage 39.6, if combination peaking were used.

Table 10·3 Comparison of high-frequency compensation methods

Type	R_L	L_o	L_c	*Relative gain at* F_2
Uncompensated	$1/2\pi F_2 C_t$	. . . .	. . . .	0.707
Shunt	$1/2\pi F_2 C_t$	$0.5C_t R_L{}^2$	. . . .	1.0
Series ($C_{in}/C_{out} = 2$)	$1.5/2\pi F_2 C_t$	. . . .	$0.67C_t R_L{}^2$	1.5
Combination ($C_{in}/C_{out} = 2$)	$1.8/2\pi F_2 C_t$	$0.12C_t R_L{}^2$	$0.52C_t R_L{}^2$	1.8

10·8 Summary of high-frequency compensation

The essential design data for the shunt, series, and combination peaking methods are given in Table 10·3. In all three cases F_2 is the same top correction frequency. This is $1/(2\pi R_L C_t)$ for shunt peaking, $1.5/(2\pi R_L C_t)$ for series peaking, or $1.8/(2\pi R_L C_t)$ for combination peaking. These three formulas have equal values of F_2 for the same C_t because the different R_L values in the denominator cancel the different factors in the numerator.

Figure 10·12 shows the response curves. The uncompensated amplifier is also included for comparison. Combination peaking provides the most gain but the sharpest cutoff, which makes the amplifier more susceptible to oscillations or ringing. Edges between black-and-white picture information then are reproduced with multiple outlines. However, for one or two stages combination peaking is used most often because of its high gain.

For shunt, series, or combination peaking, the design procedure is essentially the same as given previously for shunt compensation. The top correction frequency F_2 is decided, C_t measured, and R_L can then be calculated. With R_L and C_t known, the inductance of the peaking coils can be determined. If necessary, the inductance can be measured on a Q meter.

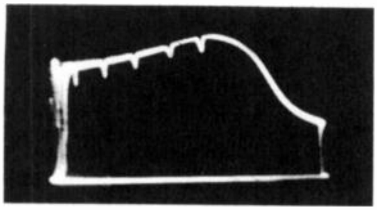
With the compensation added, the high-frequency response of the amplifier can be checked by measuring its gain from about 500 kc to 8 Mc. Use a signal generator that has a good sine waveform, and keep the signal voltage low to avoid amplitude distortion. Output is measured with a voltmeter having an r-f probe. At each frequency, the voltage amplification can be calculated as output divided by input. When the input voltage is monitored for a constant level, the output voltages indicate relative gain.

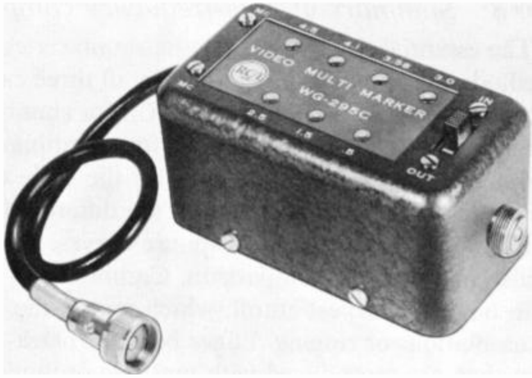
The video response curve can be observed visually on an oscilloscope similar to the way that an i-f response curve is obtained. However, the sweep generator for input signal to the amplifier must sweep through the video range of frequencies, from mid-frequencies up to about 8 Mc. Also, the video amplifier output must be connected to the oscilloscope vertical input through a detector probe. A typical video response curve on the oscilloscope screen is shown in Fig. 10·13.

10·9 Low-frequency response

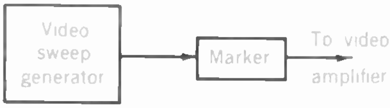
The video amplifier not only must have uniform response for the very high video frequencies but also must amplify signal frequencies down to the frame frequency of 30 cps. At low frequencies, C_t has such high reactance that its shunting effect on the plate load resistor is negligible. Now, however, the increasing reactance of the coupling and bypass capacitors at the low frequencies introduces distortion. Phase distortion is especially troublesome because of the large time delay at very low frequencies. Referring to Fig. 10·14, low-frequency distortion can be introduced in any of the following parts of the amplifier circuit: the screen resistor and its bypass $R_s C_s$, the power-supply impedance Z_b, the cathode resistor and bypass $R_k C_k$, and the coupling network $R_g C_c$.

(a)





(c)



(b)

Fig. 10·13 (a) Video-frequency response curve. Note suckouts at frequencies of absorption markers. (b) Marker unit is connected in series with signal from sweep generator. (c) Video multimarker. Size is 4 in. wide. When you touch each terminal, detuning the absorption trap makes that suckout disappear. (RCA.)

Screen-grid impedance. The screen-grid resistor R_s must be bypassed to cathode or chassis ground by C_s to maintain the screen at a steady d-c potential. Otherwise the screen-grid voltage can vary with the variations in current produced by the control-grid signal voltage. If the screen voltage varies at the signal frequency, the gain of the stage will drop because of degeneration.

Sufficient bypassing results when the reactance of C_s at the lowest operating frequency is no more than one-tenth of R_s or of the internal screen-cathode resistance, whichever is less. Usually, the screen-grid circuit can be properly bypassed with a 10-μf electrolytic capacitor, which is a common size. This normally eliminates the screen circuit as a source of low-frequency distortion.

Power-supply impedance. Unless a regulated power supply is used, the impedance of the power supply Z_b is essentially equal to the reactance of the output filter capacitor. This reactance increases at lower frequencies. If appreciable, it may introduce frequency and phase distortion at low frequencies because of the corresponding variations in plate load impedance. This problem is minimized by using a large output filter capacitor of about 80 to 150 μf. In addition, the decoupling filter R_fC_f in series with the plate circuit, as shown in Fig. 10·14, isolates R_L from the power supply. Furthermore, with correct design values for the R_fC_f filter, it can correct for low-frequency distortion introduced in other parts of the circuit. Section 10·12 describes how to use the R_fC_f filter to boost low-frequency gain and compensate for phase distortion.

Distortion caused by cathode impedance. Attenuation of the low fre-

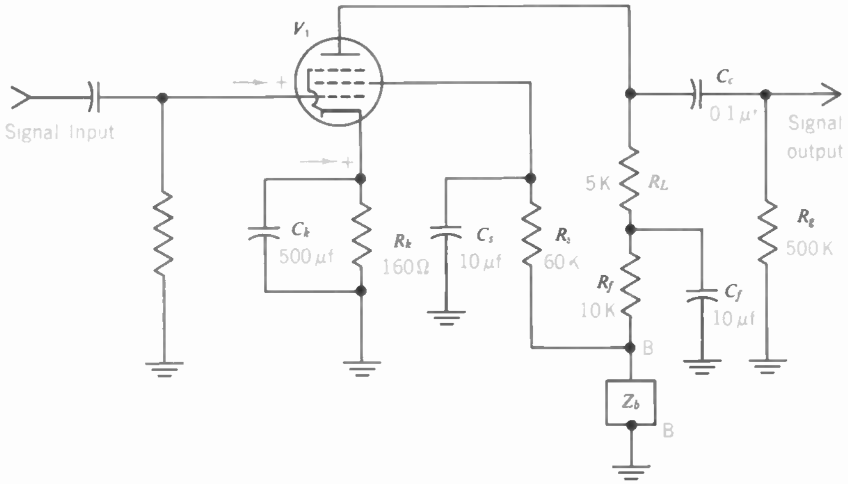


Fig 10·14 Low-frequency distortion is caused by increasing reactance of cathode bypass C_k, screen-grid bypass C_s, power-supply impedance Z_b, and coupling capacitor C_c.

quencies can result from insufficient bypassing of the cathode bias resistor. If the resistance is not at least ten times the reactance of its bypass capacitance at the lowest frequency, the cathode voltage will vary at the signal frequency instead of remaining fixed as a d-c bias voltage. This causes degeneration, with reduced output, as some of the amplified signal in the cathode is fed back to the grid circuit.

With degeneration, the cathode signal voltage opposes the input to cancel part of the grid drive (see Fig. 10·22). If the input signal drives the control grid in a positive direction, the cathode simultaneously will be driven more positive with respect to chassis ground because of the increased plate current. Making the cathode more positive is the same as driving the control grid more negative. Then the feedback to the control grid has opposite phase from the input to reduce both the input and output signal voltages.

When C_k is not large enough, its bypassing is not effective for low frequencies. The result is more degeneration and less gain for lower video signal frequencies.

The effects of insufficient cathode bypassing can be corrected by eliminating the cathode bypass capacitor. Then the cathode resistor degenerates the input signal the same amount for all frequencies. The reduction in gain is constant for all frequencies, and no low-frequency distortion is produced. However, the overall gain is reduced because of the cathode degeneration. The reduced gain can be calculated as

$$A_v = \frac{g_m R_L}{1 + g_m R_k} \qquad (10\cdot 17)$$

where R_L is the plate load resistor and R_k is the unbypassed cathode resistor. As an example, consider a stage with R_L 2,200 ohms and g_m 10,000 μmhos. The normal gain without degeneration is 22, equal to g_mR_L. However, with a 100-ohm unbypassed R_k, the gain would be reduced to 11 because of the cathode degeneration. The calculations are

$$A_v = \frac{0.01 \times 2{,}200}{1 + (0.01 \times 100)} = \frac{22}{1+1} = \frac{22}{2} = 11$$

This formula is based on μ of the tube being much greater than one, which is usually the case.

10·10 Distortion caused by the coupling circuit

In Fig. 10·14, C_c has increasing reactance at lower frequencies. For the extremely low video frequencies its reactance can be comparable with the resistance of the grid resistor R_g. Then C_c and R_g act as a voltage divider for the amplified output of the preceding stage. Only that part of the signal voltage developed across R_g is impressed across grid and cathode for amplification, however. Any voltage developed across C_c is lost as far as output signal is concerned because the capacitor voltage is not connected between grid and cathode.

Definition of F_1. This is the frequency at the low end of the response curve where the reactance of C_c equals the resistance of R_g. Then the a-c signal voltage across C_c equals the voltage across R_g, since they are equal arms in series as a voltage divider across the a-c input voltage. The voltage across C_c is reactive, however. Adding these equal voltages vectorially, each is $1/\sqrt{2}$ of the total applied voltage. Therefore, at F_1 where $X_{C_c} = R_g$ the frequency response is down to $1/\sqrt{2}$ or 0.707 of the mid-frequency response. This is a loss of 29.3 per cent in voltage or 3 db. We can define F_1 as follows:

$$F_1 = \frac{1}{2\pi R_g C_c} \qquad (10\cdot 18)$$

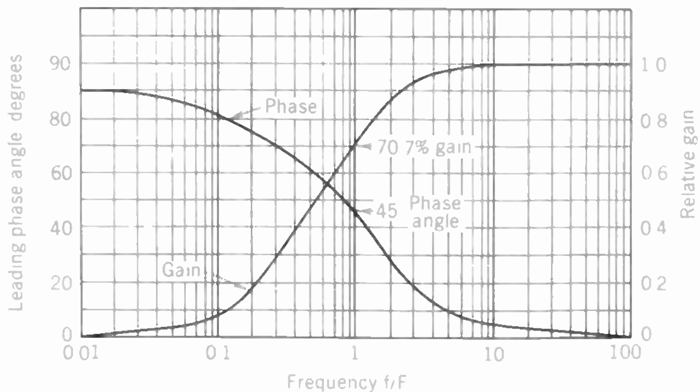


Fig. 10·15 How the gain and phase-angle distortion vary at low frequencies, as determined by the R_gC_c coupling circuit. At F_1 gain is down 3 db and phase angle is 45°.

Table 10·4 Low-frequency distortion produced by R_gC_c coupling circuit

$C_c = 0.1\ \mu f$, $R_g = 300$ kilohms, $e_p = 100$ volts

	f, cps	f/F_1	R_g, ohms	X_{C_c}, ohms	e_{C_c}, volts	e_{R_g}, volts	θ	T_θ, sec
	2.65	0.5	300,000	600,000	89.4	44.8	$-63.4°$	0.048
$F_1 \rightarrow$	5.3	1	300,000	300,000	70.7	70.7	$-45°$	0.024
	10.6	2	300,000	150,000	44.8	89.4	$-26.6°$	0.012
	53	10	300,000	30,000	9.9	99.5	$-5.70°$	0.0024

For any values of R_g and C_c, in ohms and farads, F_1 can be calculated in cps to find the frequency at which the coupling circuit reduces the response by 3 db. As an example, with 0.1 μf for C_c and 0.3 megohm for R_g,

$$F_1 = \frac{1}{2\pi R_g C_c} = \frac{0.159}{0.3 \times 10^6 \times 0.1 \times 10^{-6}}$$

$$F_1 = \frac{0.159}{0.03} = 5.3 \text{ cps}$$

The video amplifier is definitely not usable down to F_1 because the response drops 3 db, as shown by the gain curve in Fig. 10·15. Most important, the phase angle is $-45°$ at F_1 because X_{C_c} equals R_g. Even at 10 F_1, where the response is flat, the phase angle is still $-5.7°$. These values are shown in Table 10·4. Note that the small phase angle of 5.7° can result in excessive distortion because of the large time delay at the low frequency of 53 cps.

Values for R_gC_c. To obtain good performance at low frequencies F_1 should be as low as possible. This means R_g and C_c should be large. However, R_g cannot be too large because of reverse grid current resulting from residual gas in the vacuum tube, which produces a positive grid bias. The maximum permissible value of R_g varies with different tubes and depends on whether fixed bias or self-bias is used. When C_c is too large, its physical size adds shunt capacitance to the circuit, which reduces the high-frequency response. Also, the larger the capacitance of C_c, the greater its leakage current. The resulting positive bias may cause trouble with a high value of grid resistance. It is advisable, therefore, to use values of R_gC_c small enough to avoid these troubles and, if necessary, to use design values for the decoupling filter R_fC_f that compensate for the coupling circuit. Values used in the grid coupling circuit usually have an R_gC_c product with a time constant of 0.05 to 0.5 sec.[1]

Phase distortion caused by R_gC_c. More important than the loss in response is the phase-angle distortion because even a small angle corresponds

[1] The details of RC time constants are explained in Appendix E.

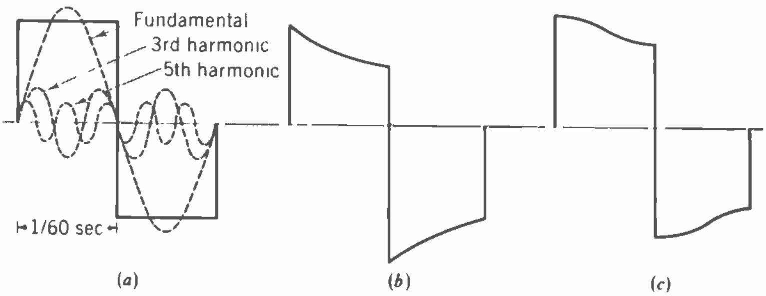


Fig. 10·16 Analysis of low-frequency distortion. (a) Square wave with its odd harmonics. Fundamental frequency is 30 cps, with 1/60 sec for half cycle. (b) Low-frequency droop because R_gC_c time constant of 0.05 sec is not long enough. (c) Square-wave tilt due to phase distortion of 6° for fundamental frequency. Flat frequency response assumed.

to a relatively long time at low frequencies. The effect of the reactance of C_c is to give the signal voltage across R_g a leading phase angle that increases for lower frequencies as X_{c_c} increases. At F_1 when $X_{c_c} = R_g$, the phase angle is 45°, or ⅛ cycle. The corresponding time for a frequency of 5.3 cps equals 0.024 sec. This time is longer than the field time of 1/60 sec. Even for 53 cps, at $10 \times F_1$ where the response is flat, the phase angle of 5.7° corresponds to 334 μsec, which is more than the time for five horizontal lines. Such large time-delay distortion causes smear in large areas of picture information.

In order to minimize this source of distortion, the phase angle should be reduced as close to zero as possible at the lowest operating frequency. Furthermore, for cascaded stages the total phase-angle distortion equals the sum of the individual phase angles. The best way to check phase distortion in the amplifier is by observing its square-wave response.

10·11 Square-wave analysis of low-frequency distortion

As shown in Fig. 10·16*a*, a square wave is the sum of a sine wave at the fundamental frequency with all its odd harmonics. The amplitude of each harmonic component is inversely proportional to its harmonic number. Thus the third harmonic has one-third the amplitude of the fundamental. All the harmonics start at zero phase with the fundamental.

The square-wave response is useful in checking an amplifier because the fundamental and all its harmonics must be amplified with flat frequency response and uniform time delay to preserve the waveshape. In general, the flat top and bottom of the square wave represent the lowest frequencies down to the fundamental. Especially evident is the effect of low-frequency droop, as in *b*, caused by insufficient response for the fundamental with phase distortion also present. Furthermore, even a fraction of a degree of phase distortion produces noticeable tilt of the square wave. In *c*, the effect of phase distortion alone in tilting the square wave is shown for a 6° leading angle at the fundamental frequency. As little as 1° phase distortion will tilt the square wave 2 per cent of the peak-to-peak amplitude.

To test the amplifier, apply a low-frequency square wave to the input, at 30 to 60 cps, and observe the output on an oscilloscope. The input square-wave signal should be coupled directly to the oscilloscope first in order to check the original waveshape and the oscilloscope response. Then the waveshape of the amplifier output can be interpreted in terms of the amplifier characteristics at low frequencies.

The inability of the square wave to maintain its flat top and bottom is generally called *droop*. This results from an RC coupling circuit with a time constant that is not long enough. For practically zero droop, the time constant should be at least five times longer than one-half cycle of the square wave. In this example, the RC product of 0.3 megohm for R_g and 0.1 μf for C_c equals 0.03 sec. Then the required half cycle is 0.03/5 or 0.006 sec. The full period is 0.012 sec, and the lowest frequency is 83.3 cps for zero droop.

10·12 Low-frequency compensation

As illustrated in Fig. 10·17, R_fC_f provides an impedance Z_f at very low frequencies to boost the amplifier gain. Furthermore, Z_f provides a lagging phase angle for the signal, to correct for leading-phase-angle distortion.

Compensating for the coupling circuit. The required values are

$$R_LC_f = R_gC_c \tag{10·19}$$

This equation says make the time constant of the R_LC_f combination equal to the time constant of the R_gC_c coupling circuit. In Fig. 10·17, R_LC_f is $3{,}000 \times 10 \times 10^{-6}$, which equals 0.03 sec. Equal to it is the R_gC_c time constant of $300{,}000 \times 0.1 \times 10^{-6}$, which also is 0.03 sec. Note that the plate load resistor R_L is used in Eq. (10·19), not R_f. Although R_L is determined by the high-frequency response, C_f provides the required time constant.

The same time constant for R_LC_f and R_gC_c makes X_{c_f} equal to R_L at F_1. Then Z_F provides the increased impedance and $-45°$ phase angle required to compensate for the voltage across R_g. See Table 10·5 for calculated values. In this example, 10 μf for C_f provides reactance equal to the 3,000-ohm R_L at 5.3 cps, for the required 3-db increase in impedance and

Table 10·5 Low-frequency compensation by C_f

$C_f = 10$ μf and $R_L = 3$ kilohms

	f, cps	f/F_1	X_{c_f}, ohms	R_L, ohms	Z_L,* ohms	θ_L
	2.65	0.5	6,000	3,000	6,700	$-63.4°$
$F_1 \rightarrow$	5.3	1	3,000	3,000	4,242	$-45°$
	10.6	2	1,500	3,000	3,350	$-26.6°$
	53	10	300	3,000	3,000	$-5.7°$

* $Z_L = R_L$ with X_{c_f}, assuming R_F very high so that $Z_F = X_{c_f}$. Also, $\tan \theta_L = X_{c_f}/R_L$.

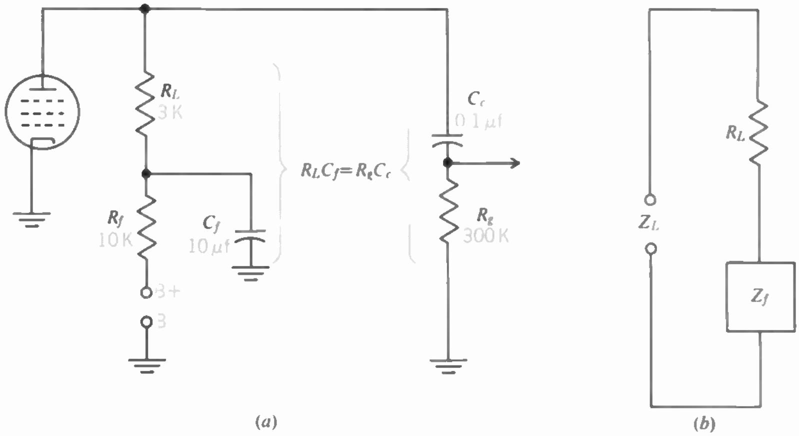


Fig. 10·17 Low-frequency compensation to correct for reactance of C_c. (a) Circuit with R_fC_f time constant equal to R_gC_c. Peaking coils omitted. (b) Equivalent impedance of R_fC_f as Z_f in series with R_L to boost plate load impedance Z_L.

$-45°$ phase angle at F_1. Furthermore, the impedance and phase angle decrease for less compensation at higher frequencies, corresponding to less distortion by the coupling circuit.

Figure 10·17 shows how Z_f can increase the plate load impedance Z_L. Note that R_f and X_{c_f} in the filter are in parallel. However, Z_f and R_L are in series to calculate the total plate load impedance Z_L.

The only function of R_f in the compensation is to be very large compared with X_{c_f}, so that Z_f is essentially the same as X_{c_f}. Then the reactance of C_f can compensate exactly for the reactance of C_c. Note that the negative phase angle θ for Z_f means it is leading i_p but the signal voltage across C_f lags i_p. This is why C_f compensates for the leading phase angle across R_g.

At mid-frequencies C_f is a perfect bypass and R_L alone is the plate load. Then Z_f is too small to be effective because of the low reactance of C_f. However, at low frequencies when C_f has appreciable reactance, its reactance in series with R_L provides a higher value of plate load impedance.

To allow the reactance of C_f to be effective, the resistance of R_f in parallel must be at least ten times X_{c_f} at F_1. This means ten times R_L. However, R_f cannot be too large because its voltage drop reduces the plate voltage. A typical value is 10,000 ohms for R_f. Note that C_f can be considered in parallel with R_f because the power-supply impedance should be very small.

Suitable values for the low-frequency compensation are shown in Fig. 10·17*a*. The compensation is only approximate, however, because R_f is not ten times R_L. Still, practical values of R_f can be three to five times R_L. Then, the compensation is effective down to the frequency where X_{c_f} is one-tenth R_f. It may be of interest to note that this method of low-frequency boost is not so easily applied to audio amplifiers because their higher values of R_L make it difficult to have R_f big enough without an excessive drop in plate voltage.

Compensating for the cathode circuit. The R_fC_f filter can also boost the gain to compensate for degeneration in the cathode circuit when the

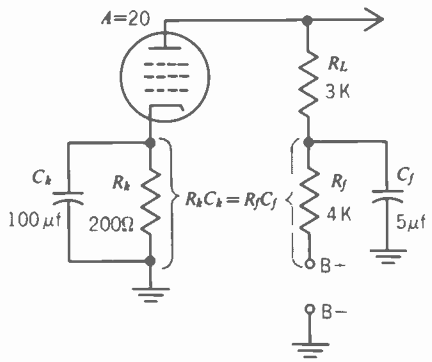


Fig. 10·18 Low-frequency compensation for cathode impedance with R_fC_f time constant equal to R_kC_k. Peaking coils omitted.

cathode bypass does not have enough capacitance. As shown in Fig. 10·18, compensation for the cathode bias impedance can be obtained by using the following values for R_fC_f:

$$R_f = R_k \times g_m R_L \tag{10·20}$$

and

$$C_f = C_k \times \frac{1}{g_m R_L} \tag{10·21}$$

These relations state that the time constant of R_fC_f equals R_kC_k, while the resistors and capacitors have the same proportion as the gain of the stage. The required cathode bias determines the value of R_k, which is 200 ohms in Fig. 10·18. C_k is given a convenient value such as 100 μf. For a gain of 20, as shown, R_f of 4,000 ohms is 20 times greater than R_k and C_f of 5 μf is 1/20 C_k. Then the two time constants match and R_LC_f produces perfect compensation for the cathode impedance R_kC_k.

Square-wave test. This method provides the best overall check of low-frequency compensation because the square-wave response shows phase distortion and frequency distortion. The square-wave test can be used to correct for the cathode impedance, the coupling circuit, or both. With a single test, the best values for low-frequency compensation are quickly obtained. The procedure can be as follows:

1. Use as large a value of R_f as possible, without dropping the plate voltage too much.
2. Use a large capacitor of 2 to 10 μf for C_f. A capacitance decade box will be convenient in experimenting for the best value of C_f. The actual C_f is an electrolytic capacitor.
3. Use the value of plate load resistor that is required for the high-frequency compensation.
4. Use a coupling capacitor of 0.05 to 0.1 μf for C_c.
5. Make the grid resistor a variable resistance to find the best value of R_g.

To make the test, couple square-wave signal at 30 to 60 cps to the input of the amplifier and observe the output coupled to an oscilloscope. Vary

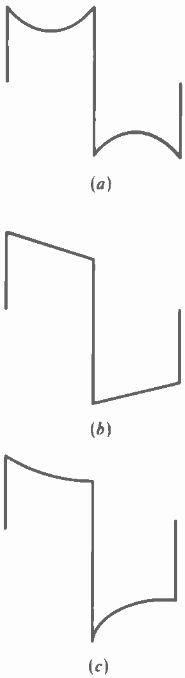


Fig. 10·19

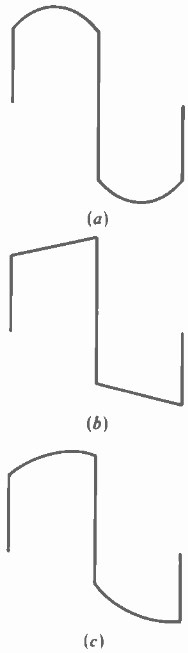


Fig. 10·20

Fig. 10·19 Typical low-frequency droop caused by R_gC_c time constant too small. (a) Attenuation of low frequencies without phase distortion. (b) Leading phase shift at low frequencies without frequency distortion. (c) Sum of (a) and (b).

Fig. 10·20 Excessive compensation caused by C_f too small. (a) Rise in gain for low frequencies without phase distortion. (b) Lagging phase shift at low frequencies without frequency distortion. (c) Sum of (a) and (b).

the grid resistance R_g until a perfect square wave is obtained in the output. This determines the value of R_gC_c that allows the compensating filter R_fC_f to provide the best low-frequency correction. Then measure the experimentally determined value of R_g and insert a fixed grid resistance of this size. If the top and bottom of the square wave are not the same in the amplifier output, reduce the input voltage to eliminate overload distortion.

Typical distorted waveshapes are shown in Figs. 10·19 and 10·20. If the output waveform has low-frequency droop as in Fig. 10·19*c*, this shows the R_gC_c time constant is too small and there is no compensation. Either R_gC_c must be increased or the low-frequency compensating filter R_fC_f must provide more boost. More reactance for C_f will result when its capacitance is made smaller, boosting Z_f. The waveform in Fig. 10·20*c* shows too much low-frequency boost caused by excessive compensation. In this case, C_f should be increased in capacitance for less reactance. When the output is a perfect square wave, the low-frequency compensation is exactly correct.

10·13 Cathode follower

A stage where the output load is in the cathode circuit, instead of the plate circuit, is a cathode follower or cathode-coupled stage (see Fig. 10·21). The input signal is coupled to the control grid, as usual, but the output signal is taken across the cathode resistor R_k. Note that R_k is not bypassed, in order to provide signal voltage output. Because of the degeneration in the cathode-grid circuit, the gain of the cathode follower is always less than 1.

Triodes or pentodes can be used. A series resistor may be used to drop the plate voltage, as in the pentode cathode follower in Fig. 10·21*b*, but the dropping resistor is bypassed so that it cannot serve as the plate load for signal voltage. The screen grid in a pentode tube is bypassed to cathode, rather than chassis ground, because the cathode is not grounded.

Although there is no cathode bypass, the average value of cathode

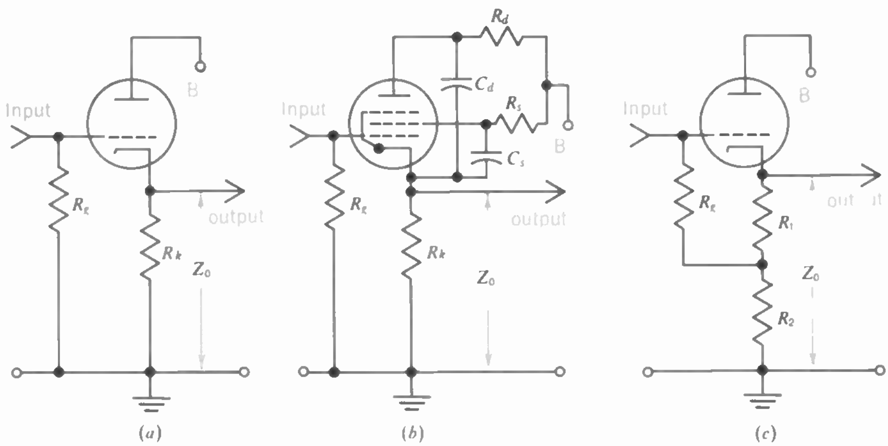


Fig. 10·21 Cathode follower stage. Load is in cathode instead of plate circuit. (a) Triode circuit. (b) Pentode circuit, with plate-dropping resistor. Both plate and screen are bypassed to cathode. (c) Triode with grid returned to tap on R_k for less bias.

current produces an IR drop across R_k to set the d-c bias. If the desired output impedance of the cathode follower requires a value of R_k that is too large for the required bias, R_k can be divided in two parts as in Fig. 10·21*c*. Note that only the voltage across R_1 serves as the bias because it is connected between cathode and grid.

The main purpose of a cathode follower is to serve as an impedance match between a high-impedance source coupled to the grid and a low-impedance load connected to the cathode. This idea is the same as an impedance-matching transformer, such as an audio output transformer, but without its limitations on frequency response. The cathode follower has especially good high-frequency response because of its low internal resistance. Also, the response for low frequencies is excellent, as one side of the cathode load is grounded directly without any bypass capacitor.

In video work, the cathode follower is used at the transmitter to couple video signal with its wide frequency range from the high-impedance plate circuit of a video amplifier to a transmission-line cable of much lower impedance, usually 50 to 72 ohms. In addition, cathode followers are often used as isolation stages. Similarly, the cathode follower is used for coupling video signal to a long line to a slave receiver. It should be noted, however, that a cathode follower stage is practically never used for video signal in the receiver chassis. Here the video output stage can drive the high-impedance kinescope grid-cathode circuit without any impedance match and no cathode follower is necessary. Still, the cathode follower is an important circuit often used for isolation and impedance-matching.

The output impedance Z_o across the cathode resistance R_k is equal to $1/g_m$ in parallel with R_k. The formula is

$$Z_o = \frac{1/g_m \times R_k}{1/g_m + R_k} \tag{10·22}$$

As an example, with R_k 100 ohms and g_m 10,000 μmhos, Z_o equals 50 ohms. This formula assumes the internal r_p is much greater than R_k. Tubes having high g_m are good for low output impedance with minimum loss of gain.

The gain of the cathode follower stage is

$$A_v = \frac{g_m R_k}{1 + g_m R_k} \tag{10·23}$$

For the same example with R_k 100 ohms and g_m 0.01 mho, the voltage gain is ½ or 0.5. The voltage gain is always less than 1, as shown by Eq. (10·23), since all the a-c output signal provides inverse feedback to degenerate the input signal. However, the higher the value of R_k the closer to unity the gain of the stage becomes. Output impedances of 50 to 100 ohms can be obtained with a voltage gain close to 1.

Although its voltage gain is less than 1, the cathode follower has power gain. Actually the stage has a useful function in taking input signal voltage across a high impedance and transferring it to a very low impedance with little loss in signal amplitude. Bear in mind that in any case an amplifier with a high-resistance plate load cannot be coupled to a very low impedance without a drastic drop in voltage amplification.

When it is necessary to find the value of R_k needed for a definite output impedance, Eq. (10·22) can be rearranged as follows:

$$R_k = \frac{1/g_m \times Z_o}{1/g_m - Z_o} \tag{10·24}$$

Example. What size R_k is needed for an output impedance of 72 ohms with a tube having g_m of 10,000 μmhos? How much is the gain?

Answer

$$R_k = \frac{100 \times 72}{100 - 72} = \frac{7{,}200}{28}$$

$$R_k = 260 \text{ ohms (approx)}$$

for the gain,

$$A_v = \frac{0.01 \times 260}{1 + (0.01 \times 260)} = \frac{2.6}{3.6} = 0.72$$

There is no phase inversion of the output signal, as the cathode signal voltage follows the same polarity as the grid input signal, which is the reason for the name of cathode follower. When the grid voltage is driven more negative by input signal, the decreased plate current makes the cathode voltage less positive; less negative input signal results in more plate current to make the cathode voltage more positive. These polarities for input and output signals are shown in Fig. 10·22. Notice that the cathode signal following the grid-signal polarity is actually opposing the input voltage from grid to cathode, degenerating the input signal. The

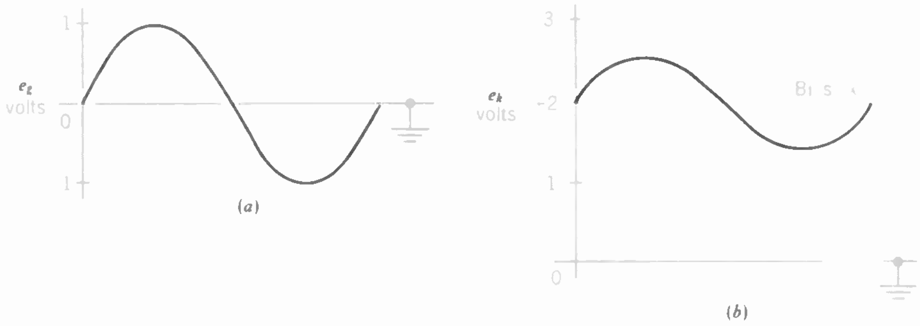


Fig. 10·22 How the average cathode voltage serves as bias. (a) Grid signal. (b) Cathode voltage.

reason why can be illustrated by considering both e_g and e_k with respect to the common ground reference. In Fig. 10·22, when e_g drives the grid +1 volt to ground, e_k goes ½ volt more positive. The potential difference of the a-c signal between grid and cathode then is only ½ volt, instead of the 1-volt grid drive.

Also note that the average-value axis of e_k is a d-c voltage that serves as the bias axis, around which the signal can vary. In the example here the cathode bias is 2 volts.

SUMMARY

1. The video amplifier is an example of a wide-band amplifier, which amplifies audio frequencies and radio frequencies.
2. The voltage gain of a pentode video amplifier equals $g_m R_L$ up to the highest frequency for which the gain is uniform. Higher values for R_L and g_m produce higher gain.
3. The high-frequency response of the *RC*-coupled amplifier drops because of the decreasing reactance of C_t shunting the plate load R_L, which reduces the plate load impedance.
4. C_t can be measured by resonating with a known inductance L. Then $C_t = 1/(4\pi^2 f_r^2 L)$. Typical values of C_t are 15 to 40 $\mu\mu$f.
5. For high frequencies, the gain of the uncompensated *RC*-coupled amplifier is down 3 db, equal to 70.7 per cent response, at the frequency F_2 where the shunt reactance of C_t is equal to R_L. Then for a given value of R_L the frequency $F_2 = 1/(2\pi R_L C_t)$. Also, for a desired F_2 the value of $R_L = 1/(2\pi R_L C_t)$.
6. For shunt peaking, L_o in shunt with C_t boosts the gain at high frequencies to provide flat response up to F_2, instead of 70.7 per cent response. See Table 10·3 for design values.
7. For series peaking, L_c divides C_t into two parts, C_{in} and C_{out}. L_c resonates with C_{in} at F_2. R_L is 50 per cent more than the value for shunt peaking. Therefore, the series peaking circuit has 50 per cent more gain and the frequency response is still flat up to F_2. See Table 10·3 for design values in terms of C_t.
8. Combination peaking uses both L_o and L_c, with a higher R_L that is 80 per cent more than for shunt peaking. Therefore, the gain is increased 80 per cent over shunt peaking, with uniform response up to F_2. See Table 10·3 for the design values required.
9. Low-frequency distortion results when the bypass and coupling capacitors are not large enough to have negligibly small reactance at the lowest signal frequencies.
10. The frequency at which the reactance of the coupling capacitor C_c equals the grid

resistance R_g is designated $F_1 = 1/(2\pi R_g C_c)$. At F_1 the response of the coupling circuit is down 3 db and the signal voltage across R_g has a leading phase angle of $-45°$.

11. A cathode bypass capacitor produces distortion when it allows more degeneration for lower frequencies because of increasing reactance.
12. The plate decoupling filter $R_f C_f$ can be used to compensate for low-frequency distortion because the reactance of C_f has two desirable effects. X_{c_f} in series with R_L increases the plate load impedance Z_L at lower frequencies to boost the gain; also X_{c_f} introduces a lagging phase angle to cancel the leading-phase-angle distortion produced in other parts of the circuit.
13. To compensate for the coupling circuit, $R_L C_f = R_g C_c$. Also, R_f is greater than 10 X_{c_f} at the lowest correction frequency.
14. To compensate for the cathode bypass, the design values are $R_f = R_k \times g_m R_L$ while $C_f = C_k / g_m R_L$.
15. The best test of low-frequency response is how a square wave is amplified. The most obvious distortion is low-frequency droop caused by too small a time constant for $R_g C_c$.
16. The cathode follower is a stage that has no plate load so that all the output signal can be taken from the unbypassed cathode load resistance. Because of degeneration, the gain is always less than 1. However, the cathode follower is useful for matching from a high-impedance source in the grid input circuit, to a low-impedance load in the cathode circuit. There is no polarity inversion of the signal in a cathode follower.

SELF-EXAMINATION *(Answers at back of book.)*

Fill in the numerical answer.

1. A pentode video amplifier with R_L 5,000 ohms and g_m 7,000 μmhos has a gain of ______.
2. For an audio amplifier with R_L 250 K and C_t 40 $\mu\mu$f, the gain is down 3 db at ______ kc.
3. With 1,000-ohm R in parallel with 1,000-ohm X_c, the combined Z_t equals ______ ohms.
4. With 1,000-ohm R in series with 1,000-ohm X_c, the total Z_t equals ______ ohms.
5. Three octaves above 5 kc is ______ kc.
6. If F_2 is 2 Mc with a 3,200-ohm R_L, C_t equals ______ $\mu\mu$f.
7. (*a*) To make F_2 4 Mc, with 25 $\mu\mu$f C_t, R_L should be ______ ohms. (*b*) For shunt peaking, L_o is ______ μh. (*c*) L_o resonates with C_t at ______ Mc.
8. (*a*) For series peaking the above amplifier, R_L is ______ ohms. (*b*) Then L_c is ______ μh.
9. For combination peaking the above amplifier, R_L is ______ ohms.
10. With C_{gp} of 0.002 $\mu\mu$f and a gain of 49, C_m is ______ $\mu\mu$f.
11. If C_t resonates with 350 μh L at 1.4 Mc, C_t equals ______ $\mu\mu$f.
12. (*a*) If C_t is 30 $\mu\mu$f and C_{in} is 20 $\mu\mu$f then C_{out} equals ______ $\mu\mu$f. (*b*) If X_{c_t} is 2,000 ohms, $X_{c_{in}}$ at the same frequency is ______ ohms.
13. With 0.5 megohm for R_g and 0.05 μf for C_c, F_2 is ______ cps.
14. With 5,000-ohm R_L, C_c of 0.05 μf, and R_g of 0.5 meg ohm, C_f is ______.
15. To make F_2 3.2 cps with C_c of 0.5 μf, R_g must be ______ megohm.
16. (*a*) A cathode follower with g_m 12,000 μmhos and R_k 200 ohms has a gain of ______. (*b*) Its output impedance is ______ ohms. (*c*) To make the output impedance 50 ohms, R_k should be ______ ohms. (*d*) Then the gain equals ______.
17. With 12 ma average current through a 200-ohm R_k, the cathode bias equals ______ volts.
18. With 8 ma average plate current the IR drop across a 10,000-ohm R_f is ______ volts.

ESSAY QUESTIONS

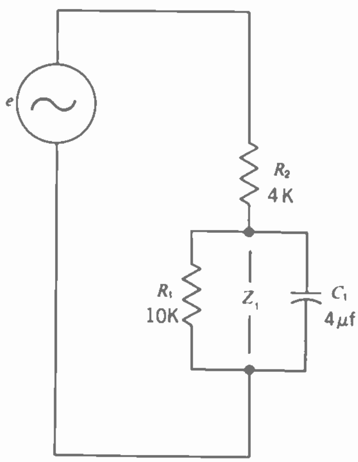
1. An amplifier has 3 Mc bandwidth from 30 to 33 Mc. Another has bandwidth from 100 cps to 3 Mc. Which would you call a wide-band amplifier? Explain why.
2. Explain briefly why the gain of a cathode follower must be less than 1.

3. Draw the schematic diagram of a triode cathode follower. Show waveshapes of input and output signals to illustrate there is no phase inversion.
4. Explain briefly why the shunt capacitance is the main reason why the response of the RC-coupled amplifier is down for the high frequencies.
5. Why does reducing R_L extend the high-frequency response? What is the disadvantage of reducing R_L?
6. Make a graph showing gain on the vertical axis, for different values of R_L on the horizontal axis equal to 2,000, 4,000, 6,000, 8,000, and 10,000 ohms. Assume a g_m value of 10,000 μmhos, with gain $= g_m R_L$.
7. Explain briefly why C_c is the main reason for low-frequency distortion. Give the effect on amplitude and phase for the voltage across R_g.
8. What two factors determine the amount of dynamic input capacitance added by the Miller effect?
9. Draw the schematic diagram of a video amplifier with combination peaking and an $R_f C_f$ plate decoupling filter.
10. Describe briefly how to test the amplifier square-wave response. Show the distorted output for the case of: (*a*) $R_g C_c$ time constant too small; (*b*) excessive low-frequency compensation.
11. In the formula $R_L C_f = R_g C_c$, what is the circuit relation of R_L and C_f that makes this time constant important for low-frequency compensation? What determines R_L?
12. For video amplifier tubes: (*a*) Why is g_m more important than μ for high gain with a low R_L? (*b*) Why are low input and output capacitances an advantage? (*c*) How does C_{gp} affect the total shunt capacitance? (*d*) From the tube manual, list the g_m and interelectrode capacitance values of two receiver tubes used as video amplifiers.
13. Draw the schematic diagram of a video amplifier using series peaking and low-frequency compensation. C_{in} is 20 $\mu\mu$f and C_{out} is 10 $\mu\mu$f. The g_m is 8,000 μmhos. Give values of all components for flat response from 60 cps to 2.5 Mc. Also show the frequency response curve.
14. Explain briefly why the degeneration with a cathode bypass capacitor too small causes low-frequency distortion but there is no distortion without any bypass capacitor.
15. If less low-frequency boost is desired, should C_f be made a smaller or larger capacitance? Explain why.
16. Derive the statement $f_r = 1.4\ F_2$ for shunt peaking, from the formula $f_r = 1/(2\pi\sqrt{L_o C_t})$. Remember to substitute $0.5\ C_t R_L^2$ for L_o and $1/(2\pi f_2 C_t)$ for R_L.
17. Prove that $f_r = 0.8\ F_2$ for L_c resonating with C_t in series peaking, where C_t is 1.5 C_{in}. Remember that L_c resonates with C_{in} at F_2.

PROBLEMS *(Answers to odd-numbered problems at back of book.)*

1. (*a*) A 12BY7 tube drives the kinescope cathode. If C_{stray} is 8 $\mu\mu$f, how much is C_t? (*b*) At what frequency will this C_t resonate with a 350-μh L?
2. At what frequency will C_t of 18 $\mu\mu$f resonate with 120-μh L?
3. An uncompensated amplifier has R_L of 5,000 ohms and C_t 40 $\mu\mu$f. Calculate F_2.
4. (*a*) With C_t of 40 $\mu\mu$f, what size R_L is needed for F_2 of 3 Mc? (*b*) With C_t of 20 $\mu\mu$f, what size R_L is needed for F_2 of 3 Mc?
5. Calculate the combined impedance of 3,000 ohms R in parallel with 4,000 ohms X_c.
6. Show the calculations to obtain the five values of Z_L in the last column of Table 10·1.
7. Calculate the total impedance of 3,000 ohms R in series with 4,000 ohms X_c.
8. Referring to Fig. 10·14: (*a*) calculate F_1; (*b*) at this frequency determine X_{c_f}, Z_L, and θ.
9. Referring to Fig. 10·14, assume B+ is 320 volts, I_b is 8 ma, and I_s is 3 ma. How much are the plate voltage, screen-grid voltage, and cathode bias?
10. It is desired to compensate an amplifier for flat response up to 3.2 Mc. C_t is 20 $\mu\mu$f. The g_m is 9,000 μmhos. Make a table listing values of R_L, peaking coils and gain for the three cases of shunt, series, and combination peaking.

11. If F_2 is 3.2 Mc at what frequency will: (*a*) $X_{L_o} = X_{C_t}$ for shunt peaking; (*b*) $X_{L_c} = X_{C_t}$ for series peaking; (*c*) $X_{L_c} = X_{C_{in}}$ for series peaking?
12. Referring to Table 10·5 calculate the exact values of Z_f and θ for the five frequencies listed, when R_f is 10,000 ohms.
13. In a cathode follower with g_m of 14,000 μmhos: (*a*) How much is Z_o with R_k of 200 ohms? (*b*) How much is the gain?
14. Draw the schematic diagram of a pentode cathode follower. Calculate R_k for an output impedance of 50 ohms with g_m of 12,000 mhos. How much is the gain? With 4 volts peak-to-peak video signal input of negative sync polarity, show the output signal.
15. Referring to Fig. 10·18, calculate the reduced gain of the amplifier at mid-frequencies if C_k were eliminated.
16. For the series-parallel a-c circuit below, calculate the values required for each frequency listed. Z_1 consists of R_1 and X_c in parallel. Z_2 consists of R_2 and Z_1 in series.



F, cps	X_c	Z_1	Z_2
5			
10			
20			
40			
80			
200			

Chapter 11 Brightness control and d-c clamping

The pedestal or blanking level of the composite video signal corresponds to black. This black level is transmitted at a constant 75 per cent of peak carrier amplitude so that the pedestal voltage can be a definite brightness reference. In the receiver, the kinescope grid cutoff voltage corresponds to black because then there is no beam current and no screen illumination. Therefore, the pedestal voltage of the composite video signal must drive the kinescope grid-cathode voltage to cutoff. Then black at the receiver corresponds to black in the transmitted picture. Also, the variations in light and shade are from a common black level, allowing the kinescope to reproduce light and dark parts of the picture with their correct relative values. Furthermore, the d-c component of the video signal should keep the pedestal voltage clamped at kinescope cutoff even when the brightness changes in the scene.

11 · 1 Brightness control

Brightness is the average illumination or background level, as distinguished from contrast, which is the difference between black and maximum white. The d-c bias on the kinescope determines brightness because it sets the operating point about which the a-c video signal swings the instantaneous grid voltage (see Fig. 11 · 1).

Remember that the average value of any a-c signal is zero. When the a-c video signal is coupled to the kinescope grid, therefore, the bias voltage becomes the signal axis. The average grid voltage then is the d-c bias. As a result, the average beam current or the average screen illumination is the amount corresponding to the bias voltage. Making the bias more negative toward cutoff decreases the average beam current, which reduces the brightness. Less negative bias toward zero grid voltage increases the beam current and brightness.

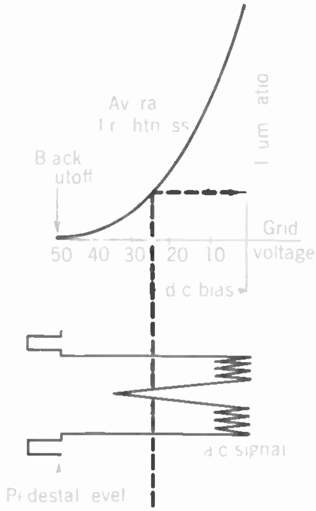


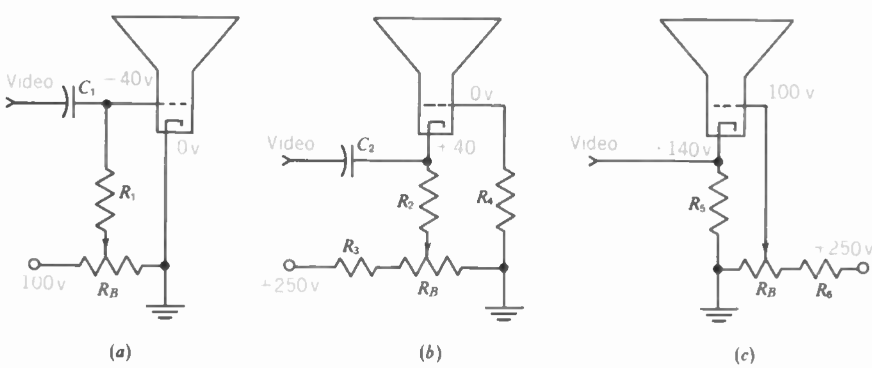
Fig. 11·1 The d-c bias on kinescope grid determines average brightness of screen illumination.

Manual bias control. This is generally called the *brightness* or *background* control, usually located on the receiver front panel. Its function is to vary the kinescope d-c bias voltage to set average brightness in the reproduced picture. Figure 11·2 shows three circuit variations for the manual brightness potentiometer R_B. In *a*, the negative bias voltage tapped off R_B is connected in series with the kinescope grid resistor R_1. Then the grid has negative bias voltage with respect to the grounded cathode.

The circuit in *b* uses the B+ voltage to supply bias for the kinescope cathode. Making the cathode positive with respect to the control grid corresponds to negative grid bias. Note that R_2 is needed to develop the a-c signal voltage coupled by C_2 to the kinescope cathode. The kinescope grid is at ground potential for d-c voltage because there is no grid current. The purpose of R_3 is to drop the B+ for R_B, allowing smaller voltage changes when the brightness control varies the kinescope bias.

In *c* the kinescope cathode has relatively high positive d-c voltage because of direct coupling from the plate of the video output stage. Then R_B must supply grid voltage less positive than the cathode. Here the grid at +100 volts is 40 volts less positive, or negative, with respect to the cathode at +140 volts.

Fig. 11·2 Manual brightness control circuits to set d-c bias for kinescope. R_B is brightness control. (a) Negative bias in control-grid circuit. (b) Positive bias in cathode circuit. (c) Kinescope grid made less positive than cathode d-c coupled to video output stage.



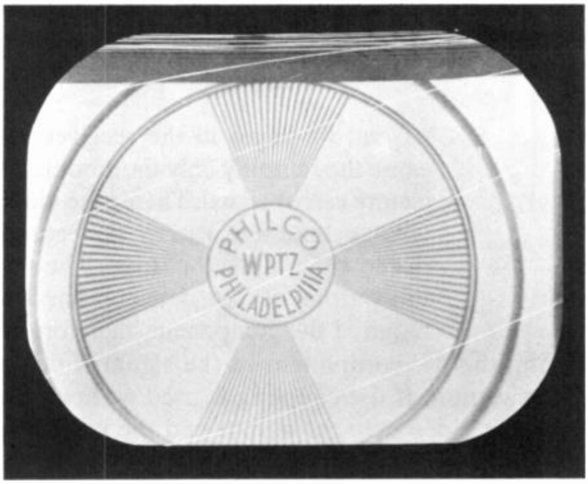


Fig. 11 · 3 Brightness too high. Black values not dark enough. Vertical retraces are seen here but they are not visible in receivers with internal vertical blanking circuit.

Incorrect brightness. Figure 11 · 3 shows a picture with the brightness too high, because of insufficient negative grid bias on the kinescope. Then black parts of the picture are too light. Also, with the pedestal voltage less negative than cutoff, the blanking level is not black. At the opposite extreme, when the bias is too negative dark parts of the signal drive the kinescope to cutoff. Then shadow detail in the picture is lost, as dark-gray values are reproduced as black. The correct bias, therefore, allows the pedestal voltage of the composite video signal to drive the kinescope grid voltage to cutoff, for the correct black reference level. Then the brightness of the reproduced picture corresponds to the actual scene brightness.

11 · 2 D-C component of the video signal

As an example of the need for the d-c component, assume that −25 volts is the grid bias needed for the brightness of a light scene. Then the picture changes to a very dark scene and back to the light background. The dark scene may require a grid bias of −35 volts. Although the manual brightness control set the bias correctly for the light scene, 10 volts more negative bias is needed for the darker scene. The manual bias cannot be varied fast enough. However, the d-c component of the video signal can shift the kinescope bias by the required amount to follow brightness changes in the picture.

The video signal is transmitted with a d-c component that indicates changes in background level for different scenes. This is done to make all the pedestals line up at the 75 per cent level, for different values of average brightness. Actually, the inserted d-c component shifts the signal axis the amount required to line up the pedestals. Consequently, the d-c component of the video signal can be recognized by either (1) pedestals in line or (2) average d-c level that varies with scene brightness. The alignment of all the pedestals at the common blanking level is maintained through the

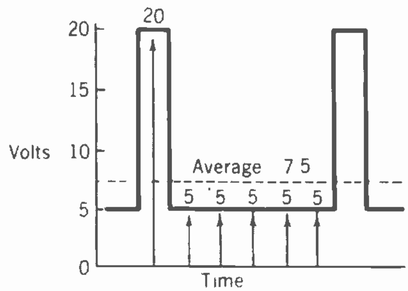


Fig. 11 · 4 Average value of a fluctuating d-c voltage.

r-f and i-f stages in the receiver because they amplify only the modulated picture carrier signal. There is no video voltage until the carrier is rectified in the video detector stage. The signal output of the detector has the required d-c component, therefore.

The problem of preserving the d-c component in the signal is determined in the video amplifier section. If d-c coupling is used without any blocking capacitors between the video detector output and the kinescope input, the d-c component will not be lost. With capacitive coupling in the video amplifier, however, the d-c component in the signal is blocked. What happens is that the coupling capacitor becomes charged to the average value of the signal voltage, which is its d-c level. The variations above and below the average axis develop a-c signal around a zero axis in the next stage. Still, the desired d-c component can be recovered by a d-c restorer circuit that rectifies the a-c signal.

Actually, many receivers use capacitive coupling in the video amplifier, without a d-c restorer. The most obvious effect with no d-c component is unblanked vertical retrace lines when the scene shifts to a darker background. Although this problem is solved by using a vertical retrace-suppressor circuit,[1] it does not change the fact that changes in brightness are not reproduced exactly right without the d-c component. In color television, the d-c level of the video signal is essential for correct reproduction of color values.

11 · 3 Average value of the video signal

For any signal, its average value is the arithmetical mean of all the values taken over a complete cycle. With fluctuating d-c voltage, the average is some d-c level. In Fig. 11 · 4, as an example, if we add the five values of 5 volts each and the 20-volt value, the sum is 45 volts. This sum divided by the six values included for the average equals 45 ÷ 6 or 7.5 volts for the average value. If we included more instantaneous values, the average could be calculated more exactly. Notice that the average axis of 7.5 volts is close to the 5-volt level, not in the middle, because this signal voltage is 5 volts for most of the cycle. The d-c voltages are measured from the zero base line. However, the average-value axis divides the signal into two equal areas above and below the axis.

When a fluctuating d-c voltage is coupled by an *RC* circuit, the average d-c axis becomes the zero axis in the a-c signal. This idea is illustrated for

[1] The vertical retrace-suppressor circuit is explained in Sec. 17 · 4.

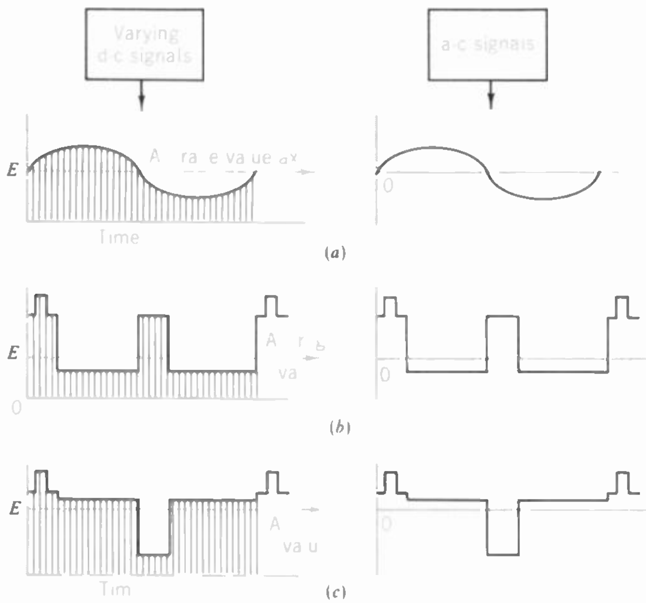


Fig. 11·5 Average-value level for different types of signal waveforms, shown in d-c form at left and a-c form at right. (a) Symmetrical sine wave with average-value axis in center of waveform. (b) Unsymmetrical video signal, mostly white. (c) Unsymmetrical video signal, mostly black.

three types of signals in Fig. 11·5. The sine wave in *a* has an average-value axis exactly in the center because it is a symmetrical waveform. In *b* the video signal corresponds to a black bar down the center of a white frame. Now the average-value axis is closer to white level because the signal is white for most of the cycle. The opposite case in *c* is for a white bar down the center of a black frame. Then the average-value axis is close to black level, because the signal is mostly black.

In all cases the average d-c axis becomes the zero a-c axis, with equal areas of positive and negative signal above and below the axis. However, notice that for the dark signal in *c* its axis is much closer to the black pedestal level than in *b*. The distance between the pedestal level and the axis of the signal variations is the *pedestal height.* This voltage is a convenient measure of how far the average signal level is from the black reference signal. The light signal in *b* has a large pedestal height, indicating the signal varies around an average far from black.

If we put the dark and light video signals on a common axis, the pedestals are out of line, as shown in Fig. 11·6. The common axis for the signals in a-c form is the zero level. In the kinescope grid circuit, however, the zero axis of the a-c signal corresponds to the d-c bias. When the

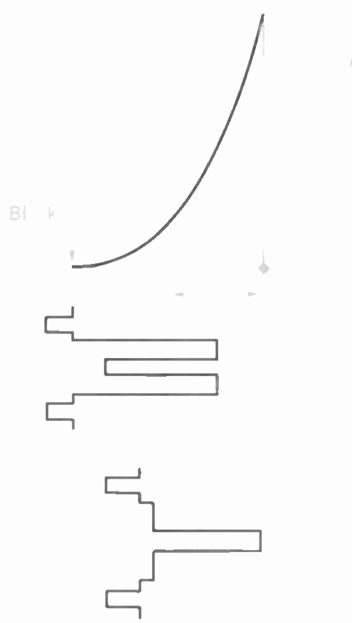


Fig. 11·6 Pedestals out of line on kinescope grid, shown for the two lines of video signal in Fig. 11·5a and b. A single value of bias cannot be correct for both signals.

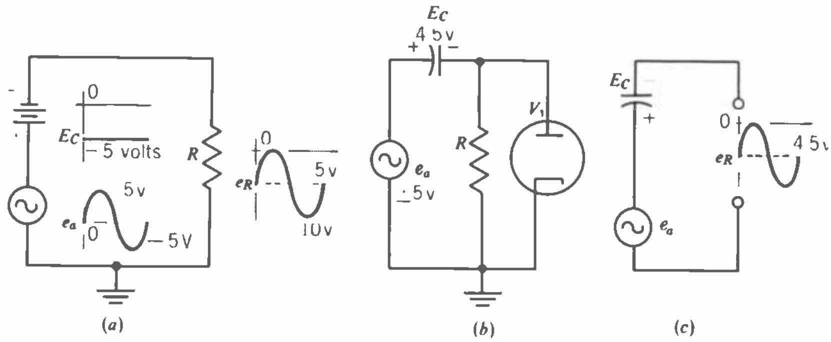


Fig. 11·7 Illustrating negative d-c insertion. (a) Battery providing d-c component in series with a-c signal e_a. (b) Diode rectifies signal to produce E_c for d-c component. (c) Equivalent circuit of E_c in series with e_a.

pedestals have different voltage variations from the common axis, one setting of the brightness control cannot be correct for both dark and light scenes. The bias illustrated in Fig. 11·6 is right for the light video signal at the top. Then the pedestals drive the kinescope grid voltage to cutoff. However, for the dark video signal below, the bias indicated is not negative enough. In this case blanking level is not black and the dark scene is reproduced too light. It should be noted that the video signals shown here are not for two consecutive horizontal lines, but rather as typical lines from two different frames for scenes that do not have the same brightness. In order for the pedestals to be in line at the blanking level for frames of different average brightness, the video signal must have its original d-c component.

11·4 D-C insertion

The basic principle of d-c insertion is illustrated in Fig. 11·7*a*. Note that the output voltage e_R is the sum of the a-c generator voltage e_a and the battery voltage E_c because they are in series with each other across R. The effect of E_c is to insert 5 d-c volts and shift the entire a-c signal to a new axis of -5 volts. Although the variations in the a-c signal are still the same, they now vary above and below the -5-volt axis, instead of the original zero axis. The output e_R is a fluctuating d-c voltage of negative polarity. This is negative d-c insertion or restoration. If the polarity of the battery is reversed, it will produce positive d-c insertion. Then the variations will be around a new axis of $+5$ volts.

In the waveshape for e_R in Fig. 11·7*a*, notice that the positive peak of the a-c signal is kept at the zero level, because of the -5 volts added by E_c. This idea of shifting the signal axis to keep one point at a fixed voltage level is called *clamping*. In this case, the positive signal peak is clamped at zero volts. Actually the signal can be clamped at any desired level, depending on the amount of d-c voltage inserted.

It is more useful to accomplish the d-c insertion for clamping by using a diode to rectify the a-c signal, as in *b* of Fig. 11·7. The diode V_1 is a rectifier to produce d-c voltage proportional to the a-c signal. When its plate is driven positive by input signal, the diode conducts to charge C, with the plate side negative. Between signal peaks when V_1 does not conduct, C can discharge through the high resistance of R. With an RC time constant very long compared with the time between positive peaks of the signal, C discharges very slowly, compared with the fast charge with diode current. As a result, C accumulates a negative charge, equal to approximately 90 per cent of the positive peak of the a-c signal. The resulting d-c voltage E_c can be considered a d-c component in series with the signal, as shown in *c*.

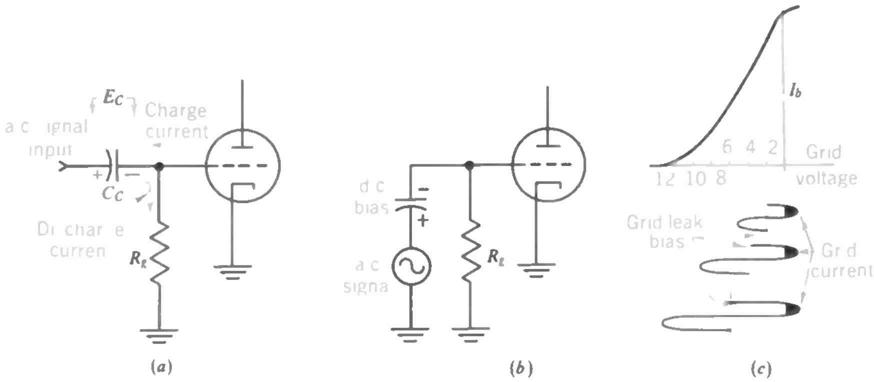
Such a circuit for restoring a desired d-c level is called a *d-c restorer* or *clamping circuit*. In *c*, the restored voltage e_R will be clamped at approximately zero voltage for any a-c signal level. If the signal changes its amplitude, the diode restorer V_1 will provide the d-c component E_c needed to clamp the positive signal peak at zero voltage.

For the application in television receivers, the composite video signal is rectified to provide the required d-c component. Once the d-c component has been inserted, direct coupling must be used to keep the d-c axis of the signal.

11·5 Clamping action of grid-leak bias

A more general example of clamping signal voltage is illustrated by the ordinary grid-leak bias circuit, in any stage where the signal drives the control grid positive to produce grid current (see Fig. 11·8). This type of bias is obtained by rectifying the a-c signal input. Effectively, the grid-cathode circuit serves as a diode rectifier that charges C_c to a d-c voltage almost equal to the peak positive amplitude of the a-c grid drive. Note that the capacitor voltage E_c is shown in *b* as a negative d-c voltage

Fig. 11·8 Clamping action of grid-leak bias. (a) C_c is charged by grid current. (b) E_c is a d-c component in series with the a-c signal input. (c) The grid-leak bias adjusts itself for different signal amplitudes to keep the positive peak clamped at approximately zero grid voltage.



in series with the a-c input. The clamping action is shown in *c*. Grid-leak bias, therefore, is an example of negative d-c insertion.

The bias produced is the amount of negative voltage needed to clamp the positive peak of the a-c signal at approximately zero grid voltage. When the grid signal has an unsymmetrical waveform it is important to remember that the peak to be clamped must have positive polarity so that it can produce grid current. It is the grid current charging C_c faster than the discharge through R_g that causes the clamping action.

How C_c produces d-c bias. When a-c signal drives the grid positive in Fig. 11·8*a*, grid current flows and electrons accumulate on the grid side of C_c, repelling electrons from the opposite side. During the time when the grid is not positive C_c can discharge through R_g. Note that C_c charges fast with grid current but discharges very slowly through the high resistance of R_g. After a few cycles of signal, C_c has a net charge with its grid side negative. The resultant voltage E_c is a d-c bias because it has one polarity. It can be considered a steady voltage compared with the signal variations, since the long time constant of the R_gC_c coupling does not allow much change in E_c during one cycle of the a-c input signal. Therefore, the d-c bias action due to grid current and the a-c coupling can be considered superimposed, as in *b*. The net result is a-c signal varying the instantaneous grid voltage around the d-c bias voltage as an axis, as shown in *c*.

The question of whether the grid is negative because of the bias or positive because of the signal is answered by the following facts. The grid becomes both positive and negative but not at the same time. After the bias has been established, the positive peak of grid signal drives the grid from its negative values toward zero grid voltage. At any instant, the grid voltage is the algebraic sum of the a-c signal voltage and the negative d-c bias.

The grid-leak bias is proportional to signal voltage. The positive drive of the signal produces enough grid current to charge C_c to a negative d-c voltage about 90 per cent of the peak positive amplitude. This feature of grid-leak bias is illustrated in Fig. 11·8*c* for different amounts of signal. With increasing signal, more grid current increases the bias to make the instantaneous grid voltage back off, so that only the peak amplitude can drive the grid positive. If the signal decreases, C_c will discharge through R_g, until the bias is low enough to allow grid current. The bias voltage settles at a value where the grid current that flows at the positive peak of signal is enough to make up for the charge lost during the remainder of the cycle when C_c discharges through R_g.

Once the bias has been established, grid current flows for only a small part of the a-c cycle, at the positive peak. This area of grid current is indicated by the shaded area at the tip of each cycle. It should be noted, however, that plate current can flow for part or all of the cycle, as determined by the grid-cutoff voltage of the tube. Therefore, a stage can operate class A, B, or C with grid-leak bias. For all three signals in Fig. 11·8*c*, plate current flows 360° of the cycle for class A operation.

D-C insertion. The grid-leak bias voltage can be considered a d-c component inserted in series with the a-c signal input. This is an example of negative d-c insertion. In Fig. 11·8c, the d-c voltage inserted is −2, −4, or −6 volts for each of the three signal amplitudes shown. The amount of negative d-c voltage inserted by grid-leak bias is proportional to the peak positive signal amplitude.

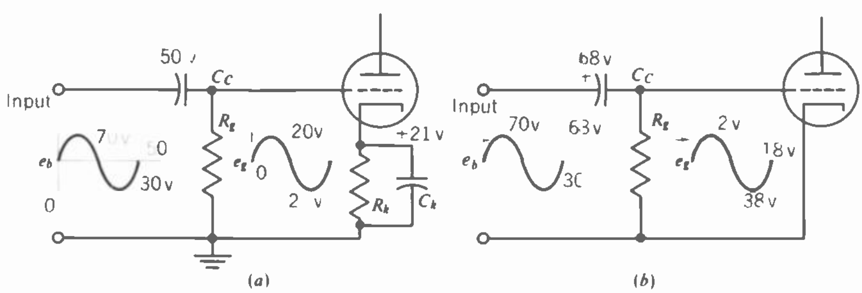
For all three examples in Fig. 11·8c, the positive peak of signal voltage is clamped at approximately zero grid voltage. When the peak positive drive increases, the bias will increase; when the drive decreases the bias decreases. The bias adjusts itself to the signal level for the amount of negative d-c voltage needed to clamp the positive peak at zero grid voltage.

Action with plate signal voltage. When R_gC_c couples signal from the preceding plate circuit where e_b is a fluctuating d-c voltage, the grid-leak bias action is the same as for a-c signal input. However, in this case we can consider the effect of grid-leak bias as a shift in the axis of the coupled signal. See Fig. 11·9. In *a*, there is no grid-leak bias. The cathode bias is enough to keep the grid signal negative so that no grid current flows. In this case, C_c charges to the average-value axis of e_b at 50 volts. This level corresponds to the zero axis of the a-c signal e_g.

The reason C_c charges to the average of 50 volts, however, is the fact that its time constant is the same for charge and discharge through R_g. With grid current flowing for the positive signal peak, conditions are different, as shown in *b*. Now C_c has a short time constant for charging through the low-resistance grid-cathode circuit. The discharge is through the high resistance of R_g. With a shorter time constant for charge, C_c will charge to a value higher than the average e_b. The example here shows C_c charged to 68 volts, which is 18 volts more than the average e_b at 50 volts.

Now the plate voltage e_b will cause C_c to produce charge or discharge current according to the changes above or below 68 volts. This axis corresponds to the zero level in e_g. However, the e_g signal variations are around the −18-volt axis, corresponding to the e_b axis of 50 volts. The result is a d-c component of −18 volts for e_g, which is the amount of grid-leak bias produced. A d-c voltmeter across R_g will read −18 volts.

Fig. 11·9 Action of R_gC_c coupling circuit for fluctuating d-c plate voltage e_b from previous stage. (a) Without grid current, C_c charges to average value of e_b. The cathode bias prevents the grid from going positive. (b) With grid current C_c charges to a voltage higher than the average e_b.



11·6 Diode clamping circuit

When d-c restoration is used for the video signal, a typical circuit arrangement has an inverted diode restorer inserting positive d-c voltage in the kinescope grid circuit to clamp the pedestal voltage at cutoff. The effect on kinescope bias is illustrated in Fig. 11·10. The manual brightness control is adjusted at cutoff bias for zero a-c video signal. Then, with video signal input for contrast, the diode restorer rectifies the signal to provide the required d-c component. The inserted d-c voltage must have positive polarity in the kinescope grid circuit in order to reduce the negative manual bias. Then, as a light scene with greater pedestal height produces a larger d-c component than a dark scene, the bias moves away from cutoff to reproduce a brighter picture. Furthermore, the bias backs off the amount needed to clamp the pedestal voltage at cutoff.

Figure 11·11 shows a diode restorer in the kinescope grid circuit. The video signal is coupled to the kinescope in two paths. C_c provides video signal with its high-frequency components. Also, video signal voltage without high-frequency compensation is taken from R_L and coupled by C_1 to the inverted diode restorer. This way the capacitances in the restorer circuit will not be in shunt with the video output. The high-frequency components are not needed for the d-c restorer. The diode is inverted so that negative voltage applied to the cathode makes plate current flow. Then C_1 in the diode cathode circuit can reinsert d-c voltage of positive polarity. R_3 provides d-c coupling for the restorer output to the kinescope grid, while isolating the diode from the video amplifier. Additional isolation is provided by R_2 in series with C_1. Only 10,000 ohms is used for R_2 because it limits the current that flows when the diode conducts.

Negative sync polarity is required for the video signal at the kinescope grid. Since the sync pulses are to be clamped, their negative polarity must make the diode conduct. Therefore, the diode is inverted so that negative drive at the cathode causes conduction. It should be noted that the inverted diode conducts when C_1 discharges, because this action provides negative signal voltage for the cathode. As a result, C_1 can discharge fast through the diode but it charges slowly through R_1.

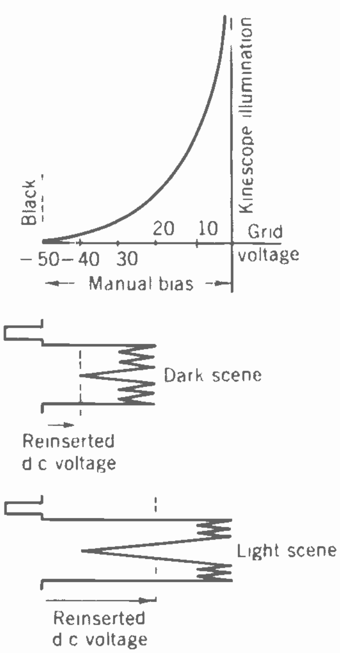


Fig. 11·10 Video signals at kinescope grid with pedestals clamped at cutoff because of positive d-c insertion. Manual brightness control set for cutoff bias with no video signal.

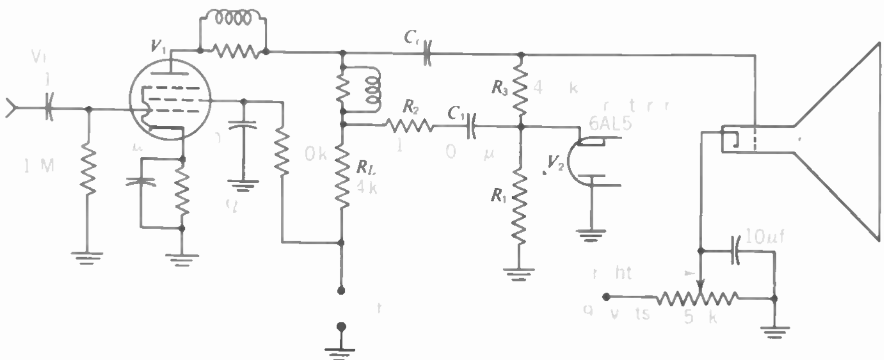


Fig. 11 · 11 Inverted diode restorer for positive d-c restoration in kinescope grid circuit.

The path for charging C_1 is from chassis ground through R_1, making this side of C_1 more negative and repelling electrons from the opposite side through R_2 and R_L back to B+. When C_1 discharges, however, the current path is through the diode and the plate-cathode circuit of the video output tube V_1. It is only because plate current increased in V_1 that its plate voltage dropped to provide negative drive for the V_2 cathode. Since discharge through V_2 is faster than charge through R_1, the d-c voltage across C_1 is reduced. The polarity of this voltage is positive on the side connected to B+ and negative on the diode side. However, the reduced voltage caused by excess discharge makes the diode side of C_1 less negative, or more positive. This action reinserts the required d-c component.

The amount of d-c voltage restored equals approximately 90 per cent of the amount of peak signal swing causing diode conduction. Since the circuit is arranged to conduct on the peak of negative sync pulses in the video signal, the d-c voltage inserted is proportional to the pedestal height of the composite video signal. Lighter scenes correspond to video signal with a larger pedestal height. This produces a larger positive d-c reinsertion voltage, allowing the kinescope grid voltage to back off more from the cutoff bias set by the brightness control.

D-C restorer time constant. Differences in brightness level correspond to a shift from a light scene to a dark scene, or vice versa. Such variations occur at a rate lower than the frame frequency of 30 cps. Furthermore, if the brightness of the scene is constant, there is no change at all in the d-c level. Then the manual brightness control could set the correct kinescope bias without the need for any d-c reinsertion. As an example, when a test pattern is transmitted, the d-c level of the video signal is constant.

The reinserted d-c voltage should vary from frame to frame when the average brightness changes, but not from line to line. Therefore, the time constant of the d-c restorer is very long compared with the horizontal-line period but not with respect to the frame time. A suitable value is 0.03 to 0.1 sec. The *RC* time constant includes the coupling capacitor for the diode rectifier and the diode shunt resistor. For the case of grid-leak bias clamping, R_gC_c determine the time constant. Both circuits operate essentially the same way as peak-signal rectifiers.

D-C component in direct-coupled video amplifiers. In a video amplifier having d-c coupling from the video detector output to the kinescope grid-cathode circuit, the d-c component is amplified with the a-c video signal. Then no restorer is needed for the correct d-c level. However, the d-c component of the video signal is negative at the kinescope grid. In this case the negative d-c component shifts the kinescope bias toward cutoff. Without video signal, the manual bias is near zero and the raster is visible. For a light scene, the bias shifts slightly more negative; with a dark scene the bias moves far negative toward cutoff. This negative polarity of the d-c component in video signal d-c coupled from the video detector to the kinescope grid is opposite to the positive d-c reinsertion of an inverted diode restorer in the kinescope grid circuit.

SUMMARY

1. The brightness or average background level of the reproduced picture is determined by the d-c bias in the kinescope grid-cathode circuit. The bias includes the d-c voltage set by the manual brightness control and the d-c component of the signal.
2. The manual bias of the brightness control can be a variable voltage from a negative source for the kinescope grid, or from a positive source for the cathode. Making the cathode more positive is equivalent to more negative grid bias, to reduce the beam current.
3. The transmitted signal has the d-c component needed to clamp the pedestals at 75 per cent carrier level, for black reference voltage. This d-c component is maintained in the r-f and i-f amplifiers of the receiver and is present in the video detector output.
4. The d-c component of composite video signal is indicated by its pedestal height. This voltage is the difference between the average-value axis of the signal and the pedestal level. A dark scene has a small pedestal height as signal variations are around an axis close to black. A light scene has a large pedestal height, with the average of signal variations farther from black level.
5. With a d-c coupled video amplifier, the video signal has a negative d-c component at the kinescope grid. No d-c restorer is necessary for the correct d-c level.
6. An inverted diode restorer, at the kinescope grid, can insert a positive d-c component proportional to pedestal height, to follow changes in brightness level.

SELF-EXAMINATION *(Answers at back of book.)*

Choose (*a*), (*b*), (*c*), or (*d*).

1. The average d-c voltage at the kinescope grid determines: (*a*) contrast; (*b*) detail; (*c*) brightness; (*d*) resolution.
2. When the kinescope bias is too negative: (*a*) contrast is too great; (*b*) brightness is too low; (*c*) pedestal level is too light; (*d*) blanking level is too light.

3. If the kinescope d-c cathode voltage is +120 volts, for −30 volts grid bias, the grid voltage is: (*a*) −30 volts; (*b*) +30 volts; (*c*) +90 volts; (*d*) +150 volts.
4. Referring to Fig. 11·9*a*, if the cathode bias is changed to 10 volts: (*a*) the stage cannot amplify signal; (*b*) operation must be class C; (*c*) an inverted diode restorer will be necessary; (*d*) grid current flows to produce about 9 volts grid-leak bias.
5. Which of the following is true for the d-c component? (*a*) It is always present in signal output from the video detector. (*b*) It is removed by d-c coupling in the video amplifier. (*c*) It is not blocked by transformer coupling. (*d*) It is not blocked by capacitive coupling.
6. If you measure the voltage shown in Fig. 11·4 with a d-c voltmeter, it will read: (*a*) 5 volts; (*b*) 7.5 volts; (*c*) 15 volts; (*d*) 20 volts.
7. Symmetrical sine-wave signal of 40 volts peak-to-peak can produce grid-leak bias of: (*a*) 18 volts; (*b*) 36 volts; (*c*) 40 volts; (*d*) 103 volts.
8. Positive d-c voltage can be inserted by: (*a*) grid-leak bias; (*b*) inverted diode restorer; (*c*) conventional diode restorer; (*d*) a transformer.
9. With an inverted diode rectifier, plate current is produced by: (*a*) negative drive at the plate; (*b*) negative drive at the cathode; (*c*) positive drive at the cathode; (*d*) positive peaks of the a-c signal input.
10. For correct brightness with changes in background level, the pedestal level should be clamped at: (*a*) zero voltage at kinescope grid; (*b*) cutoff voltage at grid of video output tube; (*c*) cutoff voltage at kinescope grid; (*d*) halfway between zero and cutoff voltage at kinescope grid.

ESSAY QUESTIONS

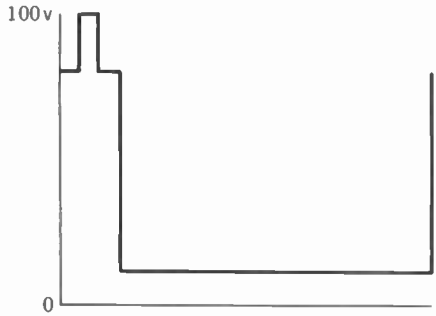
1. What kinescope voltage is varied by the manual brightness control?
2. What is the effect on the picture when the brightness control is set for too much negative bias and too little bias?
3. Why does a direct-coupled video amplifier provide the d-c component with the a-c video signal at the kinescope?
4. A video detector is directly coupled to the video output stage, which is capacitively coupled to the kinescope cathode. Is the d-c component of the signal present at the kinescope? Why?
5. One video amplifier of two stages has two capacitive couplings, while another amplifier has only one capacitive coupling. Is there any difference in the amount of d-c insertion required? Explain.
6. Referring to the diagram in Fig. 11·11, give the function of R_1, R_2, R_L, C_1, and C_c.
7. In Fig. 11·11, what would be the effect on the reproduced picture if R_L opened? If C_1 opened?
8. Redraw the waveshape in Fig. 11·4 with different values and show the new average value.
9. Redraw the circuit in Fig. 11·7*a* to show positive d-c insertion. Include waveforms.
10. In a receiver having a d-c restorer in the kinescope grid, reproducing a light scene, the kinescope d-c grid voltage is −25 volts with video signal and −40 volts without it. How much is the inserted d-c component? Why do these measurements show the d-c restorer is operating?
11. A diode restorer can follow a change in brightness from dark to light in less time than a change from light to dark. Explain this statement in terms of diode conduction and *RC* time constant in the restorer circuit.

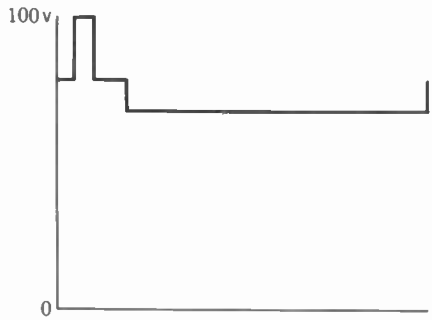
PROBLEMS *(Answers to odd-numbered problems at back of book.)*

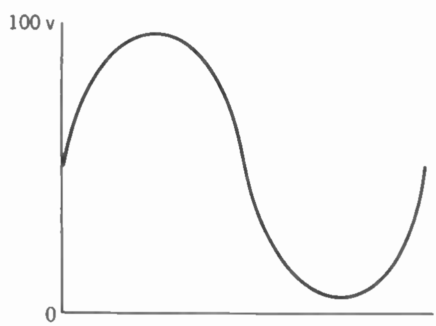
1. If the waveshape in Fig. 11·4 has a peak value of 30 volts and minimum of 6 volts, how much will the average value be?
2. Draw an a-c sine wave, having a peak value of 100 volts. Label its values at 0°, 30°, 60°,

90°, 120°, 150°, and 180°. How much is the average of these values? How much is the average value taken over the complete cycle of 360°?

3. A symmetrical square wave of fluctuating d-c voltage varies between +180 volts and +120 volts. (*a*) How much is the average d-c level? (*b*) If this fluctuating d-c voltage is applied to an R_gC_c coupling circuit, how much will E_{c_c} be? (*c*) If the next stage has grid-leak bias, how much bias will the signal produce? (*d*) In the case of *c*, how much will E_{c_c} be? (*e*) If an inverted diode restorer is connected across R_g, without grid-leak bias, what will the inserted d-c voltage be? (*f*) In the case of *e*, how much will E_{c_c} be?
4. Draw the composite video signal waveshapes for one line all white and another all black. Calculate the average value for each, using 100 volts for peak of sync and 10 volts for white. Also calculate the voltage corresponding to the pedestal height for each signal.
5. Grid-leak bias clamps a symmetrical signal with the following peak-to-peak values: (*a*) 2 volts; (*b*) 8 volts; (*c*) 60 volts. How much is the grid-leak bias voltage in each case?
6. For each of the three waveforms below: (*a*) indicate the average d-c level, and (*b*) show the a-c form with peak-to-peak amplitude.







Chapter 12 Video detector

The video detector is the second detector for the modulated picture carrier signal, following the last i-f stage in the superheterodyne television receiver. Since the AM picture signal is rectified and filtered in the video detector stage, its output corresponds to the modulation envelope of the i-f input signal (see Fig. 12·1). This envelope is the original composite video signal containing all the information needed for reproduction of the picture. The video signal output from the detector is coupled to the video amplifier to be amplified enough for the kinescope grid-cathode circuit.

12·1 Detection

The video detector action is the same as any diode detector for an AM signal. In Fig. 12·1, the signal input provides a-c voltage to drive the diode plate positive with respect to cathode. Then plate current i_p flows through the diode, the tuned input circuit, and returns through chassis ground and R_L to cathode. R_L is the diode load resistance.

The plate current through R_L produces an IR voltage drop, which is the rectified d-c output voltage. Its polarity is plus at the cathode side, with respect to the grounded end of R_L. With more a-c signal input voltage, the plate current increases. Less a-c input voltage reduces i_p. As a result, i_p consists of half cycles of direct current, corresponding to the i-f carrier sine waves. The amplitude of each half cycle of current depends on the amplitude of carrier-signal input. Therefore, the amplitude variations of i_p correspond to the modulation envelope.

Since the voltage across R_L is equal to i_pR_L, the d-c voltage e_{R_L} has the same signal variations as the current. However, the bypass capacitor across R_L eliminates variations at the i-f rate, so that the output voltage has only the lower frequency variations of the envelope. In an actual circuit, the stray capacitance bypasses R_L. As a result, the detector output across R_L is a fluctuating d-c voltage with the video signal variations corresponding to the modulation envelope of the i-f input signal.

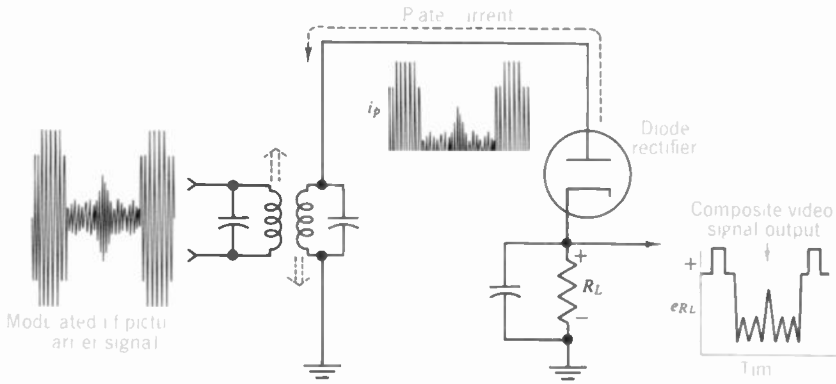


Fig. 12·1 Function of video detector. The AM picture carrier signal from last i-f stage is rectified and filtered to provide composite video signal for the video amplifier.

It is necessary to rectify the modulated carrier signal because its amplitude variations corresponding to the desired intelligence have an average value of zero, as a result of the symmetrical envelope. In terms of the desired video signal information, the carrier is just as much positive as negative at any one instant of time. However, after rectification the amplitude variations of the modulation can be obtained as variations in average value or peak value of the rectified carrier signal. With adequate filtering, the detector output voltage cannot follow the rapid variations in the individual cycles of i-f input signal, but only the relatively slow variations in amplitude corresponding to the envelope. The detector output voltage across R_L, therefore, is the desired video signal.

Either polarity of i-f input signal can be rectified by the detector, since both sides of the modulation envelope have the same amplitude variations, even though vestigial-side-band transmission is used. Remember that the side bands indicate frequencies, not voltage amplitudes.

12·2 Detector polarity

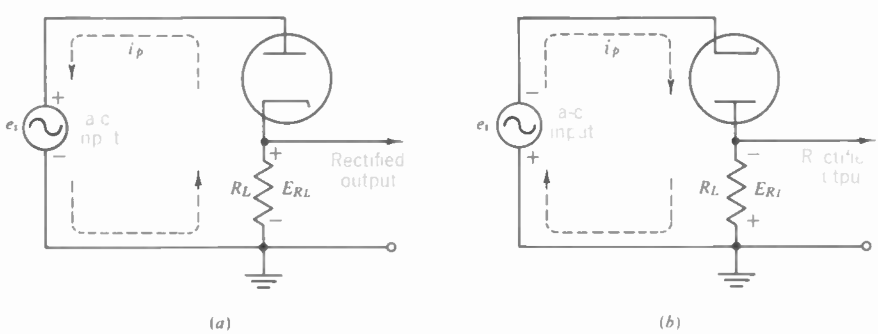
Two polarities are possible for the output voltage of a diode detector, depending on whether the diode load resistor is in the plate or cathode circuit (see Fig. 12·2). It should be noted that the polarity is not important in an audio system because the phase of a-c audio signal for the loudspeaker does not matter in reproduction of the sound. For the kinescope, though, polarity inversion of the video signal driving the grid circuit produces a negative picture.

The question of polarity in a rectifier depends on the reference point. In Fig. 12·2, equivalent diagrams are shown for a diode rectifier with positive d-c voltage output and the other having negative d-c voltage output, both with respect to chassis ground. The input signal e_i of the a-c generator reverses in polarity for each half cycle, making the top terminal alternately positive and negative, compared with the grounded side. Current flows through the load resistor R_L only when the diode conducts. To produce plate current, the generator can either drive the diode plate positive, as in *a*, or the cathode negative for the inverted diode in *b*. Notice that in *b* both the diode and generator polarity are reversed from *a*. As a result, for either circuit i_p flows from cathode to plate in the diode, to complete the path through the source and R_L in the external circuit. The only difference is that R_L is in the cathode-to-ground return in *a*, while e_i applies positive drive to the plate; R_L is in the plate-to-ground return in *b*, while e_i applies negative drive to the cathode.

Positive d-c output voltage. In Fig. 12·2*a* assume e_i supplies a positive peak of 10 volts for the diode plate. Then diode current flows, producing d-c output voltage across R_L. The cathode side of R_L is the positive terminal for E_{RL} to ground because the cathode end is closer to the positive terminal of the generator voltage, through the conducting diode. The electron flow in the external circuit is from minus to plus through R_L. Note that the diode plate is still more positive than its cathode. Otherwise diode current cannot flow. For 10 volts peak a-c input at the plate of a 6AL5 diode with 5,000-ohm R_L, the d-c output is 7 volts, positive with respect to chassis ground. The plate is still 3 volts more positive than the cathode, at the positive peak of the a-c input voltage. For the negative half cycles of a-c input voltage, there is no current and the diode is effectively an open circuit.

Negative d-c output voltage. With the diode inverted as in Fig. 12·2*b*, current flows through R_L when the generator voltage drives the cathode negative. Now the rectified d-c output voltage is obtained from the plate side of R_L. This end is the negative terminal of E_{R_L} to ground because the plate end is closer to the negative terminal of the generator voltage, through the conducting diode. The electron flow is still in the same direction, from cathode to plate in the diode and from minus to plus through R_L. For negative d-c output across R_L of 7 volts with 10 volts negative peak a-c input, the cathode is still 3 volts more negative than the plate to allow conduction by the diode.

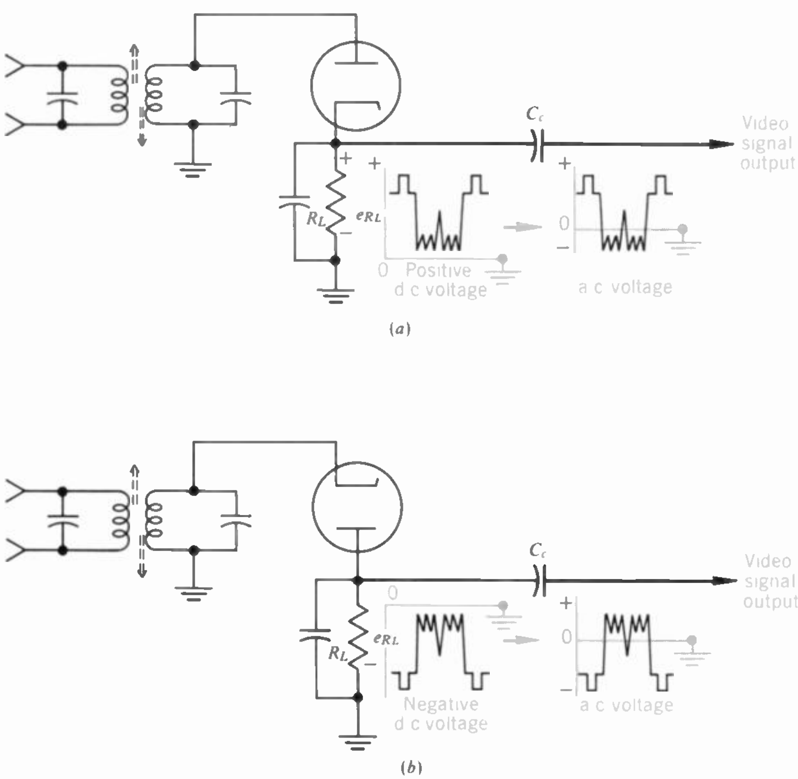
Fig. 12·2 Polarity of rectifier output voltage with respect to chassis ground. (a) R_L in cathode circuit produces positive d-c voltage at cathode. (b) R_L in plate circuit produces negative d-c voltage at plate.



Positive sync polarity. Now we can examine the polarity of the video detector output, keeping in mind that the tips of the synchronizing pulses correspond to the positive and negative peaks of the modulated carrier. Therefore, the sync voltage provides maximum amplitude of a-c input voltage to the video detector. White in the i-f signal corresponds to the lowest signal amplitudes.

In Fig. 12·3*a*, the video detector load resistance is in the cathode circuit to produce d-c output voltage positive with respect to chassis ground. The rectified output voltage is maximum positive for the peaks of sync. Output voltage is minimum for white. As shown, the rectified output varies with the amount of a-c input signal to provide the desired composite video signal as the output voltage e_{R_L}. The video signal in *a* has positive sync polarity because the sync pulses have the most positive amplitudes. With capacitive coupling, the video signal voltage coupled to the next stage is in its a-c form. Then the sync pulse voltage produces maximum positive swing from the a-c zero axis, and the white voltage swings negative. Note that the fluctuating d-c voltage in the rectifier output has an average posi-

Fig. 12·3 Polarity of video detector output signal. (a) R_L in cathode circuit provides video signal with positive sync polarity. (b) R_L in plate circuit provides video signal with negative sync polarity.



tive value corresponding to a steady d-c voltage that can be measured by a d-c voltmeter across R_L. This average d-c voltage is the d-c component blocked by the coupling capacitor C_c.

Negative sync polarity. In Fig. 12 · 3*b*, the video detector with R_L in the plate-to-ground circuit produces video signal output having negative sync polarity. The rectified output voltage is most negative for sync pulses, which correspond to maximum i-f signal input. With a-c coupling to the next stage, the a-c form of the video signal has maximum negative swing for the sync pulses. Also, white camera signal voltage produces the most positive variations from the a-c zero axis. The average value of E_{R_L} is a steady d-c voltage of negative polarity. In summary, the detector in *a* produces positive d-c output and video signal with positive sync polarity, while the one in *b* produces negative d-c output and video signal with negative sync polarity.

The video detector circuit with negative d-c output is generally used. Its output video signal with negative sync is the polarity required for the kinescope grid. When two video amplifier stages follow this detector, the video output is coupled to the kinescope grid; with one stage the video output is coupled to the kinescope cathode.

Composite video signal of negative sync polarity in the video detector output is preferred for several reasons. First, the rectified d-c voltage output is negative, which means it can be used for AGC bias in a simple AGC circuit. Also, no hum can be injected into the cathode circuit from the heater because the i-f coil in the cathode is effectively at ground potential for 60 cps. Finally, for just one video stage, amplification of the white amplitudes with positive polarity is more linear.

12 · 3 Video detector load resistance

Although not an amplifier, the video detector must have uniform output up to the highest video frequency, just as a uniform gain is required for the video amplifier. Then the video signal supplied by the detector contains the original picture information without frequency distortion.

The problem of high-frequency response in the video detector is the same as in the video amplifier. The output capacitance of the rectifier and stray capacitances are in shunt with the detector load resistance. Their decreasing reactance at high video frequencies reduces the detector load impedance, causing less output signal voltage at these frequencies. This is why the detector load resistance is about the same size as the video amplifier load resistance, typically 5,000 ohms. Note that this value is very low compared with 0.25 to 2.0 megohms commonly used in detectors for audio signal. The low value of R_L in the video detector is necessary for good high-frequency response.

12 · 4 Video detector filter

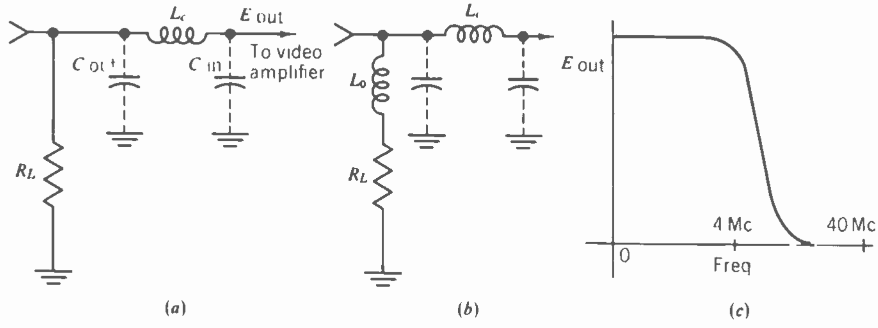
The filtering required corresponds to the function of the usual r-f bypass across the detector load resistance to eliminate variations at the intermediate frequency. The bypass capacitor cannot discharge through R_L fast enough to follow the variations of individual cycles in the i-f carrier, but the voltage output can change with the relatively slow variations corresponding to the envelope. Then the detector output voltage is the desired video signal, without the i-f carrier. However, the simple bypass capacitor cannot provide adequate filtering because the high video frequencies are not separated enough from the intermediate frequency. A capacitor large enough to bypass i-f variations would have a reactance too low for the high video frequencies; a capacitor small enough for appreciable reactance at the high video frequencies would not bypass the i-f variations effectively.

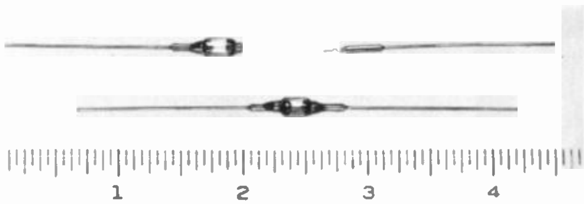
The video detector output circuit usually includes filter coupling into the video amplifier, as shown in Fig. 12·4. This method allows bypassing of the intermediate frequencies without attenuating the high video frequencies. The filter coupling has a much better band-pass characteristic than a single bypass capacitor. The circuit in Fig. 12·4*a* is actually the same as the series peaking circuit for boosting the high-frequency response of the video amplifier; the one in *b* is a combination peaking circuit. For both circuits, the values of R_L and the peaking coils can be calculated by using the design formulas in Table 10·3. Then the response will be flat up to the highest correction frequency F_2, which is 4 Mc in Fig. 12·4*c*. The sharp attenuation above 4 Mc means there will be practically no output at the intermediate frequency, which is generally 45.75 Mc. Note that the output is uniform down to 0 cps, as the detector output includes the d-c component of the signal. As a result of the filter coupling, therefore, the intermediate frequencies are filtered out and the video detector output is compensated for uniform high-frequency response. It should be noted that for intercarrier sound receivers, the response must include 4.5 Mc.

12·5 Detector diodes

The video detector uses one-half the 6AL5 dual diode, or more often, a germanium crystal diode (see Fig. 12·5). Note the point-contact construc-

Fig. 12·4 Peaking circuits for video detector output. (a) π-type filter coupling equivalent to series peaking. (b) Combination peaking circuit. (c) Typical frequency response.





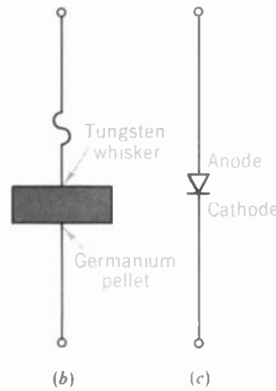


Fig. 12·5 Germanium crystal diode rectifier, type number 1N55A. (a) Photograph. Length about ¾ in. (b) Construction. (c) Symbol. (Sylvania Electric Products, Inc.)

tion, which has better high-frequency response but allows less power dissipation, compared with junction diodes used for power rectifiers. Features of the crystal diode detector are small size, no heater voltage, zero contact potential, good linearity at low signal levels, and low capacitance of 0.3 to 1 $\mu\mu$f. However, they are temperature-sensitive, with a typical maximum rating of 75°C. In the type number, such as 1N55, the N indicates a semiconductor, the 1 means a diode with two terminals, while the last digits indicate the EIA registration number.

In the diode symbol, which is generally marked on the unit, the arrowhead indicates direction of flow for positive hole charges. This direction of current is opposite from electron flow. Actually, either electrons or hole charges can provide the forward current, depending on the materials used for the semiconductor pellet and the wire whisker. In any case, though, the bar terminal is the cathode, corresponding to the cathode in a tube, while the arrowhead terminal is the anode. The desired polarity of d-c output can be obtained as follows: Apply a-c input to the anode (arrowhead) for positive d-c output at the bar (cathode). For opposite polarity, apply a-c input to the bar for negative d-c output at the arrowhead.

The graph in Fig. 12·6 shows forward current of 40 ma with 3 volts ap-

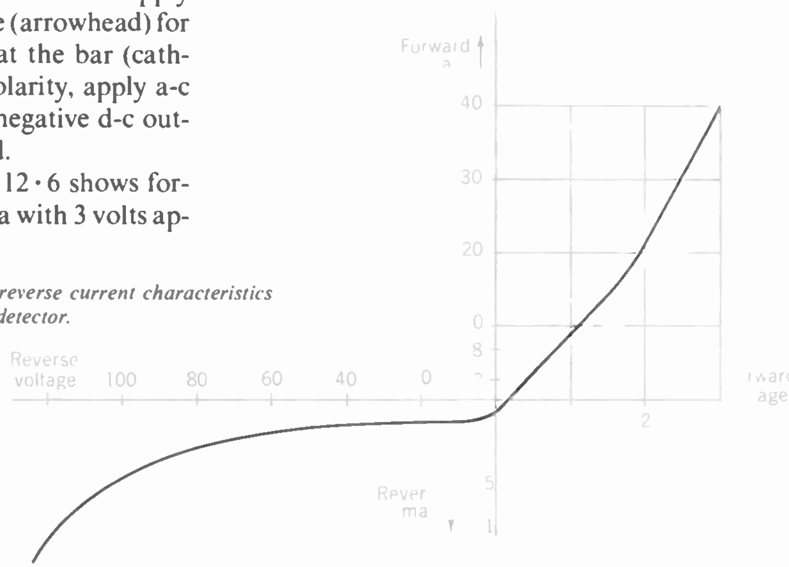


Fig. 12·6 Forward and reverse current characteristics for a typical crystal diode detector.

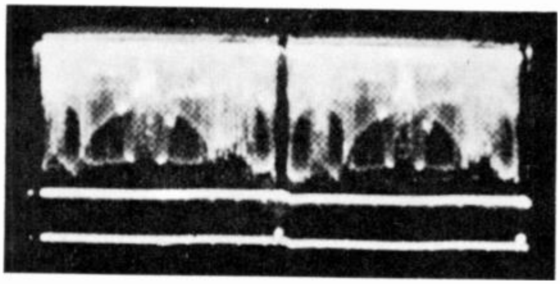


Fig. 12·7 Oscillogram of composite video signal from video detector. Amplitude 10 volts peak to peak. Sync at bottom of waveform with negative polarity. Oscilloscope internal sweep at 30 cps to show signal for two scanning fields.

plied, equal to a forward resistance of 75 ohms, for a typical crystal diode detector. The back resistance is about 30,000 ohms. Individual values may vary but most important, a good crystal diode should have a ratio of at least 100:1 for back resistance compared with forward resistance.

12·6 Video detector circuits

The picture signal from the last i-f stage has a peak amplitude of about 5 volts, allowing use of a diode detector for good linearity. In Fig. 12·8, the video detector has the diode load resistor in the plate circuit to provide composite video signal output with negative sync polarity. Figure 12·7 shows an oscillogram of video signal with this polarity. You can see this waveshape by connecting the oscilloscope input across the diode load resistance.

In Fig. 12·8*a*, the last i-f transformer couples amplitude-modulated i-f picture signal to the cathode of the video detector, which uses one-half the 6AL5 twin diode as a half-wave rectifier. The i-f transformer resonates with stray capacitance to provide at the detector cathode i-f signal voltage with respect to chassis ground. This circuit is an inverted diode rectifier, as current flows when the signal drives the cathode negative. The diode load resistance is R_1 in the diode plate circuit. Good high-frequency response is obtained by using the relatively low value of 3,900 ohms for R_1, with the peaking coils L_1 and L_2. This low-pass filter circuit also removes the intermediate frequencies from the detector output voltage. R_2 is a damping resistor for L_2. C_1 couples the variations in detector output voltage as a-c signal to the grid of the first video amplifier, while blocking the d-c level. The video signal here has negative sync polarity. Two video amplifier stages follow this detector circuit and the video output is coupled to the kinescope grid.

In Fig. 12·8*b*, the i-f signal is coupled to the cathode of a germanium crystal diode rectifier. Corresponding to an inverted diode, the crystal conducts forward current when the signal drives the cathode negative. Inside the crystal diode, positive charges flow in the direction of the arrowhead. In the external circuit, we can consider the corresponding electron flow in the opposite direction down through the diode load resistance R_1. Therefore, R_1 develops rectified output voltage with negative sync phase, as in *a*. Similarly L_1 and L_2 are peaking coils to boost the high-frequency

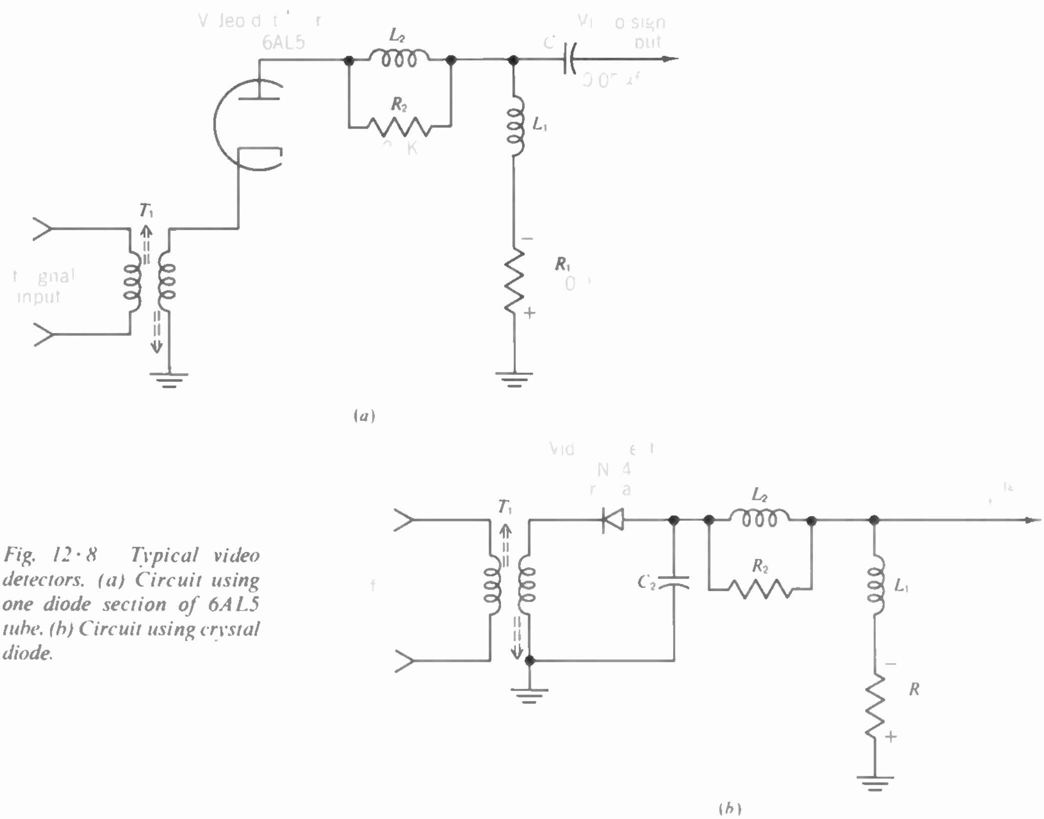


Fig. 12·8 Typical video detectors. (a) Circuit using one diode section of 6AL5 tube. (b) Circuit using crystal diode.

response. The 5-$\mu\mu$f capacitance of C_1 is added to obtain the desired ratio for C_{out} and C_{in}. In the receiver using this circuit, the detector output is directly coupled to the one video amplifier stage, which is also d-c coupled to the kinescope cathode. The average level of detector output voltage is a steady d-c voltage of negative polarity that serves as grid bias for the video amplifier. It should be noted that, because of the small size of the crystal diode, it can be mounted inside the shield can of the last i-f transformer.

A printed-circuit board for the video detector is shown in Fig. 12·9. You can trace this circuit and check against the video detector in the receiver schematic diagram of Fig. 23·16.

12·7 Functions of the composite video signal

Figure 12·10 illustrates three paths for the video signal output of the detector. We can consider that the signal is coupled to three parallel branches for the different functions. Therefore, each circuit can function independently of the others. For instance, clipping the synchronizing pulses in the sync separator stage does not interfere with normal operation of the video amplifier, which still provides the complete composite video signal for the picture tube.

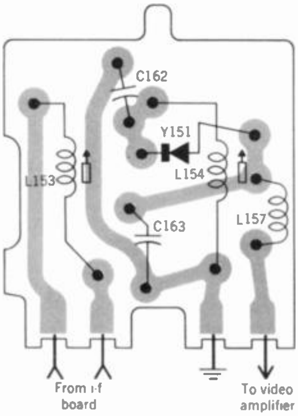


Fig. 12·9 Printed circuit for video detector, viewed from conductor side of board. (General Electric Co.)

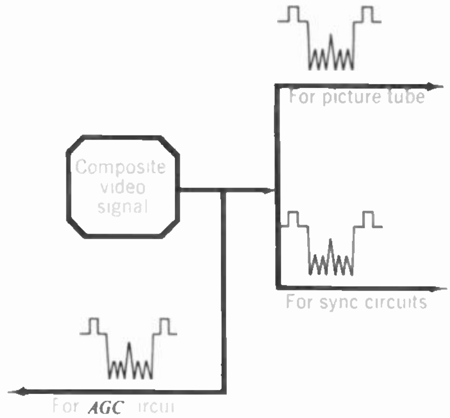


Fig. 12·10 Three functions of the composite video signal.

12·8 Detecting the 4.5-Mc intercarrier beat

In addition to the fact that the video detector diode rectifies the modulated i-f picture carrier to recover the composite video signal, this stage serves as a frequency converter to produce the 4.5-Mc second sound i-f signal (see Fig. 12·11). In terms of intercarrier sound, the 45.75-Mc picture carrier in the detector stage is equivalent to a local oscillator at this frequency because its amplitude is much higher than the 41.25-Mc sound signal. Therefore, the heterodyning action of the 45.75-Mc picture carrier beating with the 41.25-Mc center frequency of the sound signal results in the lower center frequency of 4.5 Mc. The frequency response of the video detector output circuit is made wide enough to include 4.5 Mc for the intercarrier sound signal.

As the instantaneous frequency of the FM sound signal deviates from center, it still beats with the 45.75-Mc picture carrier. The result is the same amount of frequency swing around the lower center frequency at 4.5 Mc. This FM signal is then coupled to the 4.5-Mc sound i-f section for amplification and detection to recover the audio modulation.

12·9 Measuring the detector output voltage

As illustrated in Fig. 12·12, the voltage across the diode load resistance can be checked with a d-c voltmeter to see if the detector is producing rectified signal voltage in the output circuit. This is a convenient measurement because a d-c meter is used to check i-f signal. Most receivers have a test point available for a d-c voltmeter connection to measure detector output voltage from the top of the chassis.

The video output signal of the diode detector is a varying d-c voltage of fixed polarity because the rectifier conducts in only one direction. There is no B+ voltage applied to the diode for plate voltage. Instead, the i-f signal operates the detector by supplying a-c input voltage that makes diode current flow. The d-c output voltage of the detector, therefore, is rectified signal. With typical i-f signal input, the d-c output voltage is about 2 to 5 volts. When the receiver is switched off channel to remove signal input, the detector output drops to a much lower value. The small output voltage without signal is rectified receiver noise. Measuring the d-c voltage output of the video detector, therefore, is a convenient way of checking the signal output of the i-f amplifier section.

If the d-c voltage measurement indicates no signal output from the detector, either there is no signal input or the rectifier is defective. When the detector does not function, the result is no picture and no sound, assuming an intercarrier receiver, while the raster is normal. A weak rectifier can cause poor contrast. Usually the picture is affected more than the sound.

To test a crystal diode, disconnect one end from the circuit and measure its resistance with an ohmmeter. Reverse the leads to read forward resistance and back resistance. Keep the same range for both measurements to have the same voltage applied to the diode. The $R \times 1{,}000$ ohms range should be satisfactory. If its back resistance is not more than 100

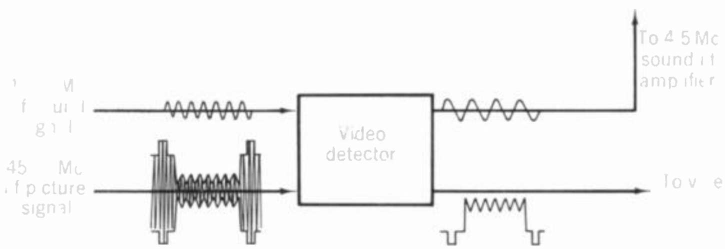


Fig. 12·11 The video detector produces the 4.5-Mc intercarrier-sound i-f signal, in addition to the composite video signal.

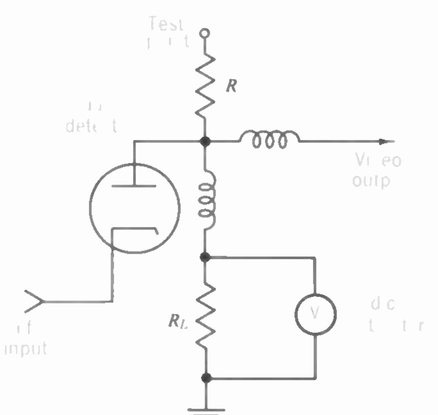


Fig. 12·12 Measuring rectified i-f signal with d-c voltmeter across diode detector load resistor.

times greater than the forward resistance, the diode is defective. When replacing a crystal diode it is important to note the correct polarity. Also, do not apply excessive heat in soldering, as high temperatures can destroy the rectifying characteristic of a semiconductor diode.

SUMMARY

1. The video detector is a diode that rectifies the amplitude-modulated picture carrier signal to recover the composite video signal envelope. Either a vacuum tube or semiconductor diode can be used. An inverted diode rectifier circuit is generally used, providing video signal output with negative sync polarity, varying around a negative d-c voltage axis.
2. A low value of load resistance, typically 5,000 ohms, is used with peaking coils in the video detector output circuit to boost the high-frequency response. The compensation procedure is the same as high-frequency peaking for video amplifiers. The wide-band response includes the 4.5-Mc sound signal but cuts off at a frequency low enough to filter out the intermediate frequencies.
3. The rectified i-f signal output of the detector can be checked with a d-c voltmeter by measuring the d-c voltage across the diode load resistor. Typical readings are 2 to 5 volts of rectified signal. The detector can also be checked by connecting an oscilloscope across the diode load resistor to see the composite video signal output.
4. The composite video signal output of the video detector has the following three functions: (*a*) signal for the video amplifier and kinescope to reproduce the picture on the raster; (*b*) signal for the sync circuits, where the sync pulses are clipped and used as timing pulses for the deflection oscillators; (*c*) signal for the AGC circuit to provide automatic bias on the r-f and i-f stages according to the r-f signal level.
5. The video detector output includes the 4.5-Mc intercarrier-sound signal.

SELF-EXAMINATION *(Answers at back of book.)*

Answer true or false.

1. The video detector is a half-wave diode rectifier.
2. The i-f signal has enough amplitude to make current flow in the detector.
3. An inverted diode rectifier has R_L in the cathode circuit.
4. Current flows in an inverted diode when the cathode is driven negative.
5. A video detector with R_L in the plate circuit produces video signal output having negative sync polarity.
6. The average d-c level of the voltage across R_L equals zero, for typical i-f signal input to the detector.
7. A typical value for R_L in the video detector is about 0.5 megohm.
8. The video detector does not require high-frequency compensation.
9. White parts of the picture correspond to maximum i-f signal driving the video detector.
10. With direct coupling, the negative d-c level of the detector output can serve as bias for the first video amplifier stage.
11. A good crystal diode has a back resistance not more than ten times the forward resistance.
12. The video detector compensation usually filters out the 4.5-Mc sound i-f signal along with the picture i-f carrier frequency in intercarrier sound receivers.
13. A crystal diode generally has less output capacitance than a vacuum-tube diode.
14. A crystal diode detector cannot provide negative sync polarity for its video signal output.
15. To check the d-c output of the detector, connect a d-c voltmeter across R_L.

ESSAY QUESTIONS

1. Explain briefly how the video detector produces composite video signal output.

2. What two input signals are required for the video detector to produce 4.5-Mc sound signal in the output?
3. Explain briefly why any modulated carrier signal must be rectified in order to recover the envelope signal.
4. Why is R_L so low in the video detector, compared with a detector for audio signal?
5. (*a*) Draw the diagram of a video detector circuit that has video signal output with negative sync polarity. (*b*) Draw a similar circuit for positive sync polarity.
6. For the circuit in question 5*a*, draw the graph of rectified output for an all-white signal. Label polarity and indicate average d-c level.
7. Referring to the oscillogram of composite video signal in Fig. 12 • 7, how can you know if the sync pulses shown are at the horizontal scanning frequency or at the vertical scanning frequency?
8. Give three functions of the composite video signal output of the video detector.
9. Referring to Fig. 12 • 8*a*, give the functions for T_1, R_1, C_1, and L_1. What is the function of the diode?
10. Explain briefly why the output of the video detector can be checked with a d-c voltmeter. Referring to Fig. 12 • 8, where would the d-c voltmeter be connected in *a* and *b*?

PROBLEMS *(Answers to odd-numbered problems at back of book.)*

1. Calculate the values required for R_L and L_c in a video detector circuit with series peaking for flat response up to 4.6 Mc. C_{out} is 4 $\mu\mu$f and C_{in} for the succeeding video amplifier is 8 $\mu\mu$f.
2. For the same specifications as in Question 1, calculate R_L, L_o, and L_c using combination peaking.
3. Referring to Fig. 12 • 6, how much is the forward current of the crystal diode for forward voltages of 0.5 volt, 1 volt, 2 volts, and 3 volts?
4. Tabulate the values of forward resistance for the crystal diode characteristics in Fig. 12 • 6, at forward voltages of 0.5 volt, 1 volt, 2 volts, and 3 volts. Draw a graph plotting forward resistance on the vertical axis against forward voltage on the horizontal axis.

Chapter 13 Automatic gain control

The automatic gain control (AGC) circuit varies receiver gain according to the strength of the carrier signal received. Less gain is needed for strong signals than for weak signals. Therefore, the AGC reduces the gain for strong signals by increasing the negative grid bias on the r-f and i-f tubes. The manual contrast control then can be in the video amplifier section for easy control of picture contrast, while the AGC circuit automatically sets the i-f and r-f bias at a value that is best for the signal level at the antenna. Practically all television receivers use AGC to prevent overload distortion on strong signals. Note that the AGC bias on the common r-f and i-f amplifiers affects both the picture and sound signals in inter-carrier-sound receivers.

13 · 1 Requirements of the AGC circuit

The circuit in Fig. 13 · 1 is called simple AGC because it does not have extra details of delay bias, amplification, or keying, which can be added to improve the AGC action. The first requirement is producing the negative AGC bias voltage. This is done by the diode V_6, which rectifies the i-f signal. R_L is the diode load resistor in the plate circuit of the inverted diode rectifier to produce negative d-c voltage output proportional to signal strength.

The voltage across R_L must be filtered to remove the signal variations, since a steady d-c voltage is required for bias. This is the function of the AGC filter R_1C_1. The filtered output across C_1 is the negative d-c voltage that serves as the AGC bias. Each i-f and r-f stage controlled by the AGC bias has its control-grid circuit returned to C_1 instead of chassis ground. Then the negative AGC bias voltage across C_1 is in series with the signal input to the grid. As a result, the AGC voltage determines the grid bias. The controlled stages may also have cathode bias of 1 to 3 volts for minimum bias when the AGC voltage is zero with no signal input.

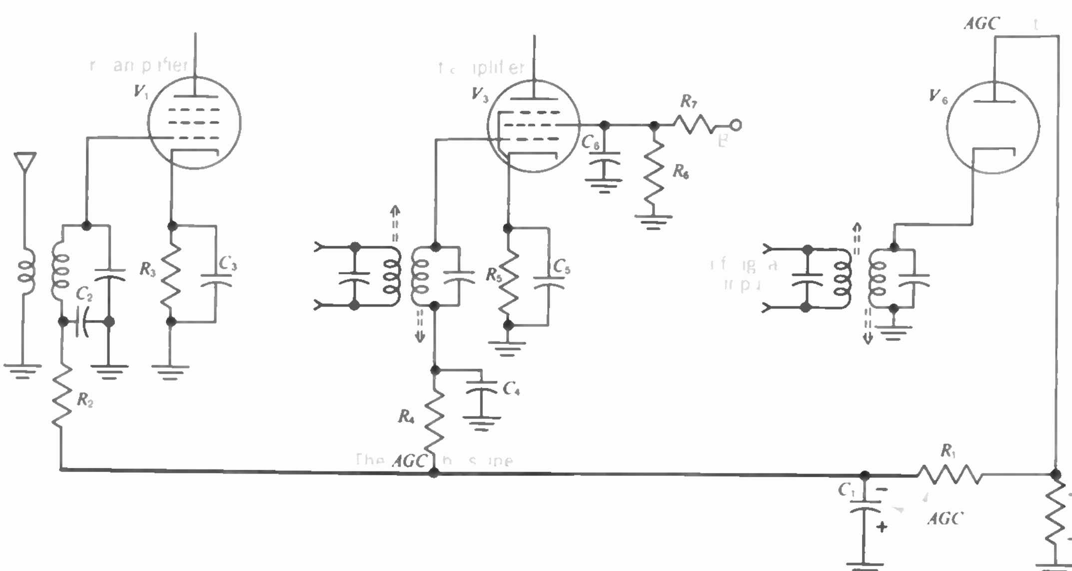
The AGC bias in Fig. 13·1 can vary between −2 and −10 volts, approximately, depending on signal level. An increase in signal strength produces more negative AGC bias, which reduces the gain of the controlled stages. With less signal input to the receiver, for the opposite case, the AGC circuit develops less negative bias, allowing more gain than for strong signals. Notice, though, that in both cases the gain is reduced in comparison with zero AGC bias.

The constant detector output results because the AGC bias reduces the receiver gain more for strong signals than for weak signals. As an example, suppose that the r-f and i-f stages have an overall gain of 10,000 for antenna signal of 0.2 mv, allowing 2 volts detector output. Now assume a much stronger antenna signal of 2 mv produces enough negative AGC bias to reduce the overall gain to 2,000. The resulting detector output is 4 volts. Notice that, for a 10:1 increase in antenna signal, the detector output only doubled because of the AGC action in reducing the receiver gain. Although the receiver has less gain, there is more detector output because of the much stronger input signal.

The function of each part of the AGC circuit in Fig. 13·1 can be summarized as follows:

1. The AGC rectifier V_6 produces the negative d-c bias voltage. More signal produces more AGC bias; less signal results in less AGC bias. The rectified d-c output voltage across the diode load resistor R_L is negative because R_L is in the plate-to-ground circuit. It should be noted that the negative d-c output voltage of a diode detector can also be used for AGC bias, instead of having a separate AGC rectifier. In either case

Fig. 13·1 A simple AGC circuit.



the d-c voltage must be filtered to provide a steady bias for AGC.

2. The R_1C_1 filter removes the a-c variations from the rectified signal voltage to provide a steady AGC bias voltage. The filter time constant is generally about 0.3 sec. A shorter time constant will not filter out the lowest frequency variations in the rectified signal. Too long a time constant will not allow the AGC bias to change fast enough when the receiver is tuned to stations having different signal strengths.
3. The AGC line is connected to the grid circuit in each controlled stage to supply the negative AGC voltage as grid bias. The filtered AGC voltage across C_1 can be considered the source of AGC bias to be distributed by the AGC line. This idea is the same as automatic volume control (AVC) in radio receivers. In television receivers the AGC line usually controls the bias on the r-f amplifier, plus the first and second i-f amplifier stages.
4. The decoupling filters R_2C_2 and R_4C_4 isolate each grid circuit from the common AGC line. The isolation is provided by the resistors to minimize feedback between stages through the AGC filter capacitor as a common impedance. Note that the bypass capacitor C_4 provides a low-impedance path for alternating signal current returning to cathode in the i-f amplifier. Otherwise the signal circuit would include the high resistance of the d-c path with R_3. C_2 is also a bypass capacitor to connect the coil to the grounded end of the tuning capacitor in the r-f amplifier grid circuit, without shorting the d-c voltage on the AGC bias line to chassis ground.
5. Remote-cutoff tubes, also called *variable-mu tubes,* should be used in i-f stages controlled by AGC bias. The remote-cutoff characteristic helps reduce amplitude distortion for the case of high signal level and a large negative bias.
6. Stages controlled by AGC bias may have a voltage divider across the B+ supply to stabilize the screen-grid voltage, as shown for V_3 in Fig. 13·1. With just a series resistor, the screen voltage rises when the screen current decreases because of more negative bias. This effect opposes the AGC action. Therefore, the voltage divider to stabilize the screen-grid voltage is preferable in a stage controlled by AGC bias.

Delayed AGC. The AGC circuit reduces the receiver gain more for strong signals than for weak signals. However, there is still some AGC bias and loss of gain for very weak signals, produced mostly by noise, just when maximum sensitivity is needed. For this reason some receivers have provision for shorting the AGC line to chassis ground when necessary to allow maximum receiver gain.

The problem of preventing AGC action on weak signals can be solved by using a delayed AGC circuit. In this arrangement, a reverse-bias voltage is connected to the AGC rectifier to prevent conduction untii the signal is strong enough to overcome the delay bias. Note that the delay is in voltage, not in time, as the AGC rectifier is biased out of conduction for weak signals. As an example, suppose that a 3-volt battery is connected to the AGC diode, negative at the plate. The diode cannot conduct, therefore, until the signal has more than 3 volts amplitude to make the plate positive and allow diode conduction. However, for all signal levels higher than 3 volts, the diode conducts to produce AGC bias voltage.

The delayed AGC circuit requires a separate AGC rectifier because it is not desired to bias the detector, which should be allowed to operate for the weakest signals. In addition, the delay bias for the separate AGC rectifier must come from a source other than the signal itself, since the delay bias must be present without any signal input.

Amplified AGC. With this arrangement, a small change in signal level can produce a relatively large change in AGC bias. The result is better control of receiver gain because of the wider range of bias voltage, compared with simple AGC. The larger bias voltage with amplified AGC is especially helpful in eliminating overload distortion for very strong signals. Either of two methods can be used. In one arrangement, the signal is amplified before being applied to the AGC rectifier. Then more AGC bias can be produced because of the larger rectified signal, compared with the level at the second detector. The other method uses a d-c amplifier stage to amplify the AGC bias voltage itself. Receivers with an amplified AGC circuit generally require an adjustment for AGC level, because enough negative AGC bias can be produced to cut off the controlled stages. The AGC level is adjusted just below the point of overload distortion on the strongest station.

Keyed AGC. In this circuit, voltage pulses are used to key on the AGC rectifier to conduct only during a small part of each cycle of the signal. The advantage is that noise pulses in the signal have little effect on the AGC bias voltage. This circuit is explained in Sec. 13 · 5.

13 · 2 How the AGC bias controls gain

As the AGC voltage makes the bias on the controlled stages more negative toward cutoff, the grid has less control on plate current because fewer electrons pass through to the plate. Therefore, the grid-plate transconductance g_m and amplification factor (mu) of the tube both decrease when the AGC bias becomes more negative.

Remote-cutoff tubes. These are also called *variable-mu* or *supercontrol* tubes. Because the control grid is constructed with variable spacing between turns in the grid winding, the tube has a gradual cutoff characteristic, compared with a sharp-cutoff tube (see Fig. 13 · 2). Note that the remote-cutoff characteristic includes a wider range of grid-bias values. Then the bias can be made more negative without having the signal drive the grid beyond cutoff. Also, the remote-cutoff characteristic has a continuous variation in slope. The change in plate current for a change in grid voltage, which equals g_m, decreases toward the grid cutoff voltage. With less g_m, the tube provides less gain for more negative AGC bias voltage.

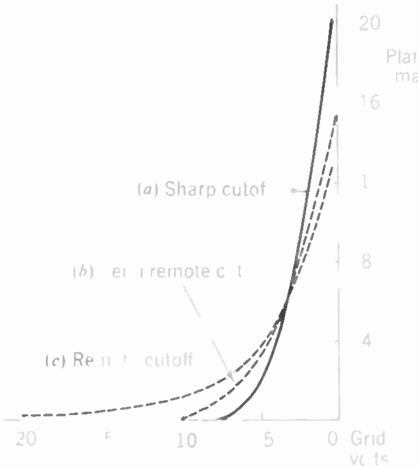


Fig. 13·2 Comparison of grid-plate transfer characteristic curves for pentodes with sharp cutoff, remote cutoff, and semiremote cutoff.

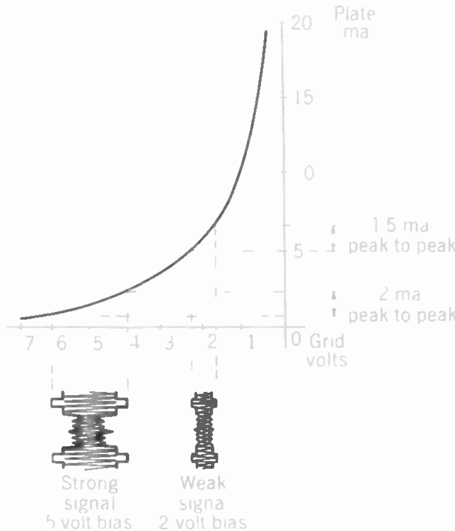


Fig. 13·3 Illustrating the effect of more negative AGC bias in reducing gain for strong i-f signal.

It should be noted, though, that a sharp-cutoff tube has higher g_m, allowing more gain than the equivalent remote-cutoff tube. Also, the sharp-cutoff characteristic is more linear, which results in less amplitude distortion of the signal. Because of these factors, tubes have been developed with a semiremote-cutoff characteristic, as illustrated by the middle curve in Fig. 13·2. This compromise between opposing factors allows control of gain for a range of about 10 volts grid bias, without excessive nonlinearity and with little loss of g_m.

Control of i-f gain. The effect of a change in d-c grid bias in changing the gain and a-c signal output is illustrated in Fig. 13·3. Examples are given for bias values of -2 and -5 volts. If we considered equal amounts of signal for the two examples, the more negative bias would provide less gain and less signal output. However, the input signals taken here have different amplitudes to illustrate the case of stronger signal producing more negative AGC bias. Then the plate-current variations of the two output signals are almost the same in amplitude because the AGC voltage reduces the gain for the stronger signal.

With more AGC bias, the gain of the i-f amplifier is reduced because its g_m is reduced. The gain is numerically equal to $g_m \times Z_L$, where Z_L is the impedance of about 5,000 ohms for the i-f tuned circuit at resonance. Figure 13·4 shows g_m values for different values of negative grid bias for a typical semiremote cutoff pentode. In the curve, g_m drops from 10,000 μmhos with 0 volts e_c, for maximum gain, down to practically zero for 16 volts e_c. Notice that negative d-c bias voltage greater than -16 volts can cut off the amplifier. In this case, there is no i-f signal output.

Because of the reduced i-f gain for stronger signals, the AGC action provides relatively constant detector output. This effect is illustrated by the AGC characteristic curve in Fig. 13·5 for a receiver with several stages controlled by AGC bias. The curve shows detector output voltage for antenna signal input. Without AGC action, the output just increases with signal input until overload distortion occurs. For curve *c* with AGC bias but no delay, the output increases much more slowly as the input signal increases. Furthermore, a much stronger input signal can be received without overload distortion. Curve *b* shows AGC action like curve *c* but there is no control until the delay bias is overcome at the knee of the curve.

Decibel calculations. In a typical receiver, the AGC circuit can maintain the output signal voltage of the detector within a range of 2:1 (6 db), while the antenna input voltage varies over a range of 100 to 1 (40 db). These db calculations are shown below.

For a voltage ratio of 2,

$$\text{db} = 20 \times \log 2 = 20 \times 0.3 = 6$$

For a voltage ratio of 100,

$$\text{db} = 20 \times \log 100 = 20 \times 2 = 40$$

Subtracting 6 db from 40 db, the answer of 36 db indicates the amount of AGC action. This means that if the maximum gain of the receiver is assumed as 10,000 without AGC, the largest negative AGC bias can reduce this gain by 36 db below 10,000. To convert to an actual ratio for voltage gain, we can use the antilog of the db formula, as follows:

$$\text{antilog (db/20)} = \text{voltage ratio}$$

or,

$$\text{antilog (36/20)} = 10{,}000/\text{gain}$$

The larger gain of 10,000 is in the numerator to have a fraction greater than 1 for positive logarithms. To find the antilog,

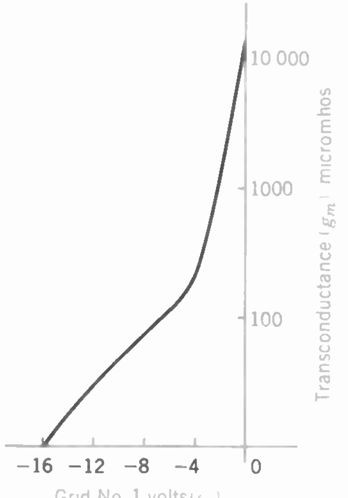


Fig. 13·4 Characteristic variation of g_m with grid bias for typical semiremote-cutoff pentode.

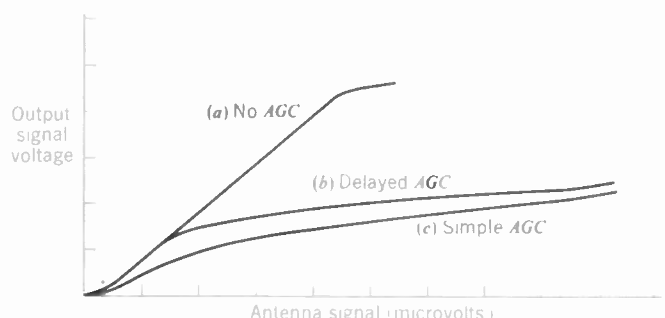


Fig. 13·5 Typical AGC characteristic curves for receiver.

$$\text{antilog }(36/20) = \text{antilog } 1.8 = 63$$

Therefore,

$$63 = 10{,}000/\text{gain} \qquad \text{or} \qquad \text{gain} = 10{,}000/63 = 159$$

This value of 159 is the lowest gain, 36 db below the maximum gain of 10,000, as a result of the most negative AGC bias for the strongest signal input.

13 · 3 Advantages of AGC for picture signal

The main advantage is relatively constant contrast in the reproduced picture. With the manual control set for the desired contrast, the AGC circuit maintains this contrast level by providing constant video signal amplitude from the detector. Although the strength of antenna signal is usually different when switching from one station to another, the picture AGC circuit can automatically keep the video signal amplitude at the same level so that the manual contrast control need not be readjusted. Since the AGC controls the i-f gain, the manual contrast control is in the video amplifier section. The manual control has the function of setting the contrast over a relatively narrow range, while the picture AGC circuit controls the bias on the r-f and i-f amplifiers automatically to adjust for the wide range of input signal levels. Adjusting the manual control for the desired contrast is easier, therefore, because of the picture AGC circuit.

The picture AGC circuit minimizes the problem of overloaded picture due to excessive signal. See Fig. 13 · 13 for an example of a picture out of sync with reversed black and white values caused by overload distortion. Another advantage is that separation of the sync pulses is easier in the sync separator section when the composite video signal has constant amplitude.

Airplane flutter. This is a rise and fall of picture intensity, making the picture fade in and out, when airplanes are flying nearby. The cause is fading of the picture carrier signal as its amplitude increases and decreases because of the moving airplane. For the carrier frequencies used in television broadcasting, the airplane is several wavelengths long and can act as an antenna. Since it is in the field of the radiated wave, the airplane body intercepts some signal, current flows as on an antenna, and signal is reradiated. The reradiated signal may aid or oppose the original transmitted signal picked up by the receiving antenna. Furthermore, the phase relations change continuously as the airplane moves. The resultant fading occurs at a rate of about 10 to 25 fluctuations per second. In addition, the reradiated signal can produce a ghost in the picture. The sound may also be distorted and pulsate in volume. If the airplane is transmitting its own radio signal, this r-f interference can cause diagonal bars in the picture. The effects of fading in the antenna signal caused by airplane flutter can be practically eliminated by the picture AGC circuit.

13·4 AGC circuits for picture signal

Either the modulated picture i-f signal or the detected video signal can be rectified to provide the AGC bias voltage. When the video signal is used, however, it is d-c coupled to the AGC stage to preserve the d-c component. Then the pedestals of the video signal are in line so that a peak rectifier can indicate sync voltage level to measure the carrier strength correctly.

The stages controlled by AGC bias are generally the r-f amplifier and the first two i-f stages, assuming three i-f stages. The last i-f stage usually has cathode bias without AGC control, for two reasons. The effectiveness of the AGC control is proportional to the gain for signal from the grid of the controlled stage to the AGC rectifier. Therefore, AGC bias on the earlier stages can provide more effective control. In addition, the last i-f stage has a relatively high signal amplitude, which can easily be distorted by a change in bias.

Another important factor is the time constant of the AGC filter. The question is how to produce AGC bias voltage that measures peak carrier level for a correct indication of carrier strength, without making the filter time constant too long. This time constant determines how fast the AGC bias can change to a different value when necessary. With a long time constant of 0.3 sec or more, the AGC circuit cannot control the receiver gain fast enough to compensate for airplane flutter. In addition, noise voltages can set up an AGC bias voltage too negative for the amount of signal. The filter capacitor charges fast through the AGC rectifier, but if discharge through the filter resistor is too slow, the excessive AGC bias reduces the receiver gain more than necessary. The result can be a low video signal level, with temporary loss of synchronization because of weak video.

At the opposite extreme, too short a time constant does not allow the peak signal amplitude to be measured for correct indication of carrier strength. One particular problem is that a fast AGC circuit can interpret the vertical blanking pulses as an increase in signal because of their greater pulse width. Then the AGC bias rises slightly during this time. The reduced receiver gain at 60 cps makes the vertical blanking pulse droop in the i-f signal voltage input to the detector. Most important, if the vertical sync pulses have reduced amplitude, the picture may roll out of sync easily because of poor vertical synchronization.

In spite of these problems, all television receivers use picture AGC for its convenient control of contrast with minimum overload distortion in the picture, and the reduction of airplane flutter. We can consider three types of circuits. A separate diode used as a peak rectifier for AGC bias is shown in Fig. 13·6*a*. The more economical arrangement in *b* uses the average d-c output of the video detector as the source of AGC bias. The best solution to the problem of a fast AGC circuit capable of indicating carrier strength correctly is the keyed AGC circuit in Fig. 13·10.

Separate AGC diode as peak rectifier. Referring to Fig. 13·6*a*, T_1

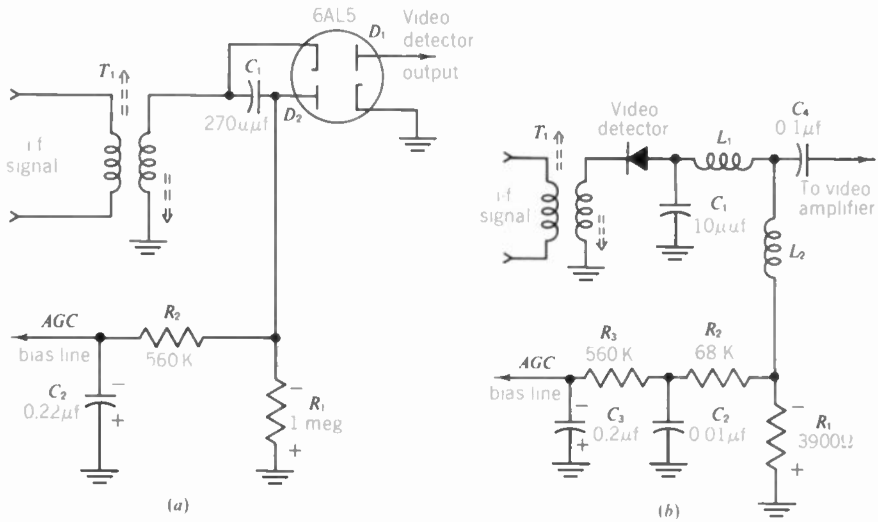


Fig. 13·6 · Diode AGC circuits. (a) Separate AGC diode as peak rectifier. (b) AGC bias from video detector diode.

supplies i-f signal input for the video detector D_1 and the AGC rectifier D_2 in the 6AL5 twin diode. The i-f signal voltage makes diode current flow. The coupling capacitor C_1 with its discharge resistor R_1 allow the AGC diode to function as a peak rectifier. When i-f signal drives the AGC diode plate positive, C_1 charges fast through the diode. However, the R_1C_1 time constant of 270 μsec is longer than 63.5 μsec so that C_1 cannot discharge appreciably between horizontal sync pulses. Therefore, the amount of rectified voltage is close to the peak of the input signal. The polarity is negative at the ungrounded end of R_1 because it is in the diode plate circuit.

The R_1C_1 time constant is not made too long in order to minimize AGC setup on noise pulses. The additional filter R_2C_2 has a longer time constant of 0.12 sec to filter out 60-cycle ripple on the AGC bias line, caused by response to the vertical blanking pulses. Finally, the filtered AGC bias controls the gain of the r-f amplifier, plus the first and second i-f amplifiers. In each of the controlled stages the control grid returns to the AGC line through decoupling filters.

AGC bias from video detector. The circuit in Fig. 13·6*b* is a typical video detector using a crystal diode. R_1 is the diode load resistor producing negative rectified output. The average rectified voltage indicates carrier-signal strength, approximately. Therefore, the negative d-c voltage across R_1 is filtered to provide AGC bias. The double-section filter has a shorter time constant for R_2C_2 in order to minimize noise setup across C_2.

The video detector cannot be a peak rectifier, as the output voltage must follow the envelope variations of the i-f input signal. The average d-c output voltage used for AGC bias, therefore, can vary when the d-c component of the signal changes for different brightness levels. Since the AGC circuit reduces receiver gain, this circuit reduces the d-c component in the video signal. Other than this, though, the AGC take-off does not affect the operation of the video detector in supplying composite video signal for the video amplifier. Note that R_2 and R_3 serve as isolating resistors that separate the AGC filter capacitors from the detector load R_1.

D-C amplifier for AGC voltage. A larger bias voltage allows more control of receiver gain, which is especially needed for very strong signals. For this reason amplified AGC is better than simple AGC. A separate amplifier may be used for the AGC bias voltage before it is applied to the controlled stages, as illustrated in Fig. 13·7. A d-c amplifier is necessary, with direct coupling to the source of AGC bias voltage in the input and to the AGC line in the output. Note that positive AGC bias is applied, as the amplifier inverts the polarity. The source of bias can be an AGC diode, with positive output. However, in some receivers the cathode-grid circuit of the sync separator stage provides the positive rectified signal voltage for input to the AGC amplifier.

Bias-clamp diode. For any type of AGC circuit, the bias line to the controlled stages can have a diode to clamp the bias voltage for the r-f amplifier at zero. The basic idea is to keep the bias on the r-f amplifier lower than the i-f stages. Remember that the r-f amplifier is the first stage with the lowest signal level. In this stage, maximum gain is most important for a good signal-to-noise ratio.

In order to reduce the r-f bias, a positive voltage is inserted in the AGC line going to the r-f amplifier (see Fig. 13·8). The value of R_1 is determined in proportion to the resistance of the AGC circuit so that the B+ supply of 250 volts is dropped to a small value at point X. We can assume +3 volts. R_2 isolates point X and the r-f bias line from the i-f bias line, which has the normal AGC voltage. Suppose the AGC bias is −5 volts. At point X, though, the bias for the r-f stage is only −2 volts because of the +3 volts added. With any AGC voltage more than −3 volts, the r-f bias is 3 volts less than the i-f bias. For these values, the diode clamp does not conduct as its plate is negative.

Fig. 13·8 Bias-clamp diode on AGC line for r-f amplifier.

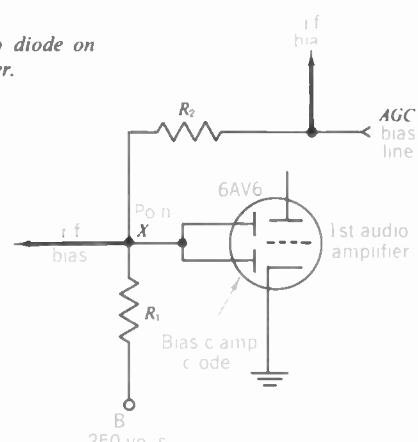
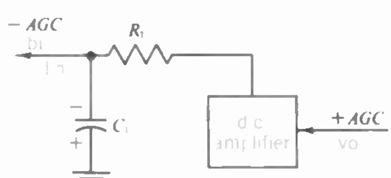


Fig. 13·7 D-C amplifier for AGC bias voltage.



The need for the clamp diode arises when the AGC bias is less than -3 volts. Then, if not for the diode clamp the r-f bias would be positive, with severe amplitude distortion. For any positive voltage at point X, however, the diode plate is positive and it conducts. Effectively, the low resistance of the conducting diode shorts point X to chassis ground, clamping the r-f bias at zero.

In summary, then, a small positive voltage is added to reduce the negative bias on the r-f stage. Also, the clamp diode makes the r-f bias zero to prevent positive voltage on the line when the AGC bias is too small. Note that the diode clamp in Fig. 13·8 uses part of a duodiode-triode. However, the bias-clamp circuit is for the AGC line and has nothing to do with the audio amplifier.

13·5 Keyed AGC circuit

In this circuit, the AGC tube is driven into conduction by positive keying pulses applied to the plate. The keying is done by flyback pulses from the horizontal output circuit. Then the AGC tube produces plate current only during horizontal sync pulse time. As a result, its output is free from noise because the circuit does not operate between keying pulses. In addition, fast AGC action can be obtained without any error from the vertical sync pulses because their greater width is not measured by the keyed AGC tube.

With normal phasing of horizontal scanning when the picture is in sync, flyback occurs during blanking time. This coincides with the time of the horizontal sync pulses in the composite video signal (see Fig. 13·9). Therefore, the AGC tube is keyed on to conduct only during horizontal sync pulse time.

A typical keyed AGC circuit is shown in Fig. 13·10. Composite video signal with its d-c component is directly coupled from the video output circuit to the grid of the AGC tube. At the same time, horizontal flyback pulses drive the plate positive during horizontal sync pulse time. With no video signal input, the AGC tube is biased beyond cutoff by the negative voltage between control grid and cathode. In this circuit the grid bias is -25 volts, equal to the difference between $+100$ volts at the grid and $+125$

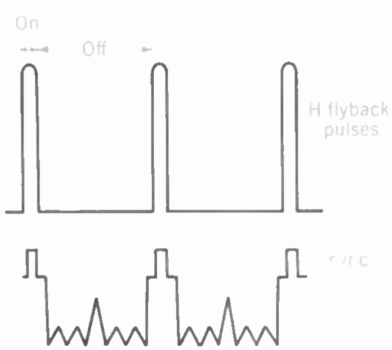


Fig. 13·9 Keying pulses at horizontal sync rate for keyed AGC tube.

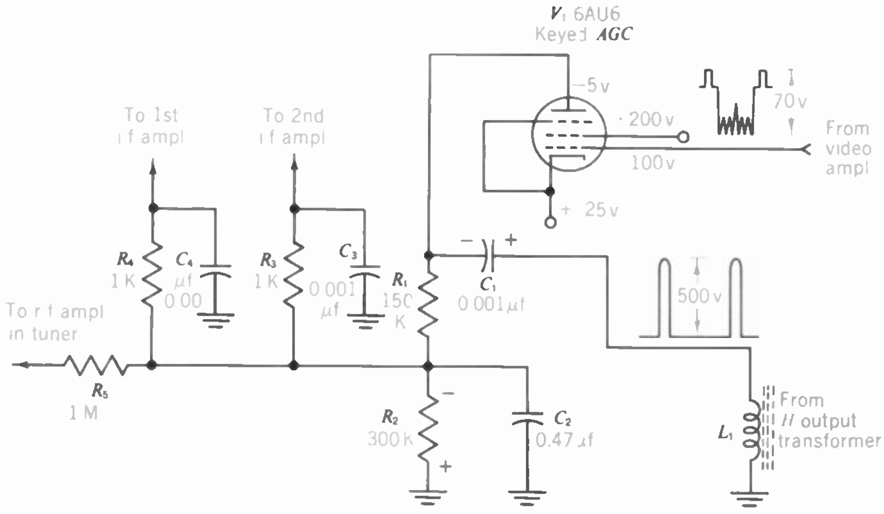


Fig. 13·10 Typical keyed AGC circuit.

volts at the cathode. However, the sync pulses at the positive peak of the video input signal drive the grid voltage to zero so that the tube can conduct as the plate voltage is pulsed positive during this time. The grid bias is negative enough to allow only the sync pulses to produce plate current. Therefore, the rectified output in the plate circuit is proportional to the peak amplitude of the composite video signal with its correct d-c component, which indicates the picture carrier-signal strength.

The plate circuit of the AGC tube must produce negative voltage for the AGC bias line. This required negative plate voltage is produced by the coupling circuit for flyback pulses when plate current flows. The idea is the same as negative grid-leak bias produced by an *RC* coupling circuit when positive signal makes grid current flow. In Fig. 13·10, C_1 couples the positive pulses to the plate of the AGC tube. When the tube conducts, plate current charges C_1 with the plate side negative, as shown. The path for charging current includes the cathode-to-plate circuit in the tube and L_1, which is the source of the applied voltage pulses. Between pulses, when the tube does not conduct, C_1 has a discharge path through R_1, R_2C_2, and L_1. However, the resistance is high enough to make the discharge time constant much longer than the 63.5 μsec between pulses. Therefore, the voltage across C_1 is a relatively steady negative d-c voltage proportional to the peak value of the video signal input.

Since the d-c voltage across C_1 is in parallel with the R_1R_2 divider, R_2 has two-thirds of the negative plate voltage. R_1 is used for isolation, to separate the AGC line from L_1 supplying the keying pulses. C_2 with R_2 form the AGC filter. Its time constant of 0.14 sec is fast enough to follow airplane flutter. The vertical blanking pulses have no effect on the AGC because of the keying pulses and the circuit is immune to noise because the AGC tube is cut off most of the cycle. Finally, the AGC line connects the filtered AGC bias to the control-grid return for the first and second i-f amplifiers, of three i-f stages, and the r-f amplifier in the tuner, which is a separate subchassis. The standard color code for wiring in the AGC line is white.

The keyed AGC circuit is often combined with the sync separator stage, using one tube including twin sharp-cutoff pentodes (see Fig. 13·11). This combination is convenient because both stages need the same composite video signal input, which is applied to grid 1. The video signal is d-c coupled, in order to preserve the correct amplitude of peak carrier signal. Grid 2 is also common, serving as the screen grid for both sections. Grid 3 is separate in each pentode, to be used as either a suppressor grid or an additional control grid for noise gating. The sync separator function in this combination is explained in Sec. 14·6.

13·6 AGC level adjustment

The bias supplied by the AGC circuit is usually made adjustable to provide the amount of gain control needed for different signal strengths. In strong-signal areas, more AGC bias is needed to prevent excessive signal from overdriving an amplifier stage. The overload results in a reversed, dark picture, out of sync (Fig. 13·13). At the opposite extreme, for very

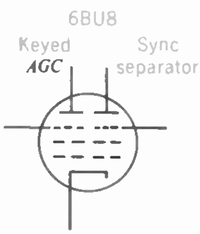


Fig. 13·11 Schematic of 6BU8 twin sharp-cutoff pentode.

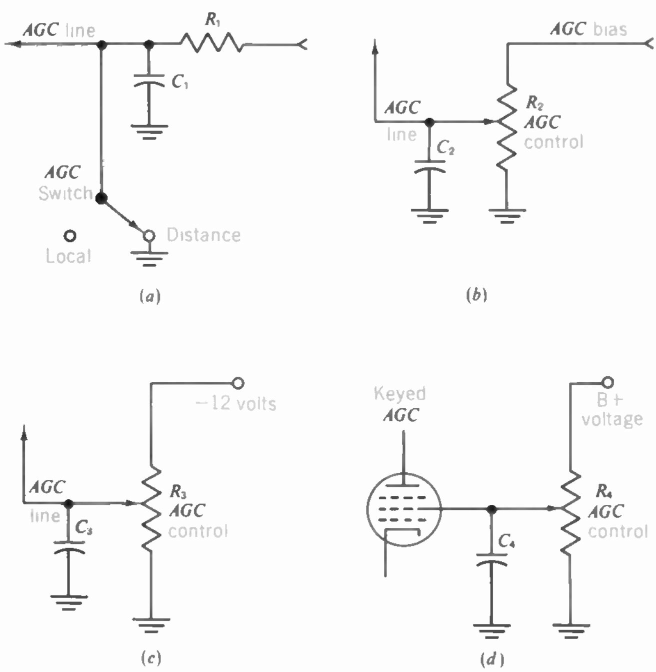


Fig. 13·12 Circuit arrangements for adjusting AGC level. (a) Switch to short AGC line for zero AGC bias. (b) Varying the AGC bias voltage. (c) Varying the delay bias voltage on AGC line. (d) Varying screen-grid voltage on AGC amplifier.

weak signal in fringe areas less bias or no AGC voltage is desirable to allow more receiver gain.

Figure 13 · 12 illustrates typical arrangements for adjusting AGC level. In *a* the local-distance switch can short the AGC line to chassis ground for zero AGC bias, on the distance position for very weak signals. Then the r-f and i-f stages operate with their minimum self-bias. The circuit in *b* allows a more gradual control, between zero and maximum AGC bias. In *c* an adjustable negative voltage is inserted in the AGC line. The source of −12 volts in this circuit is the grid-leak bias voltage on the horizontal amplifier, but any negative supply can be used. This arrangement improves the AGC action for a diode AGC circuit, as the negative voltage added to the AGC line provides enough bias for very strong signals. The circuit in *d* adjusts the screen-grid voltage on the keyed AGC tube. This control determines the amount of plate current and the resultant AGC bias. More screen-grid voltage allows more plate current to increase the AGC bias for strong signals. The same results can be obtained with an AGC adjustment for cathode voltage or d-c voltage at the control grid of the keyed AGC tube.

The AGC control is usually mounted on the rear apron of the chassis as a setup adjustment. It can generally be set by adjusting for less AGC bias until the top of the picture begins to bend, indicating compression of the sync pulses. Then back off a little. This adjustment should be done for the strongest station, with the manual contrast control at maximum. Otherwise the picture may bend on strong stations, or when the contrast control is turned up.

More exact adjustment can be made by connecting an oscilloscope to the video detector output to see the composite video signal waveform. Adjust the AGC level until the sync pulses start to compress and then back off the control just enough to remove the compression. The compression of the sync means the pulses have less than 25 per cent relative amplitude compared with the peak-to-peak signal voltage. If the signal is not strong enough to produce bend in the picture, the AGC control can be adjusted by watching the snow in the picture. Reduce the AGC bias to the point where the snow becomes a little more obvious and then back off the control a little. In weak-signal areas, it is generally necessary to reduce the AGC bias to minimum, or short the AGC line completely, for maximum gain in the receiver.

13 · 7 AGC troubles

The AGC circuit is a common cause of troubles with contrast in the picture. Since the d-c bias on the AGC line controls gain for the a-c signal in the r-f and i-f amplifiers, the AGC bias is the most important factor in determining contrast. Too much AGC bias reduces the receiver gain, causing a weak picture. When the bias is negative enough to cut off an r-f or i-f stage the result is no picture, with a blank raster. The sound is also cut off in intercarrier receivers.

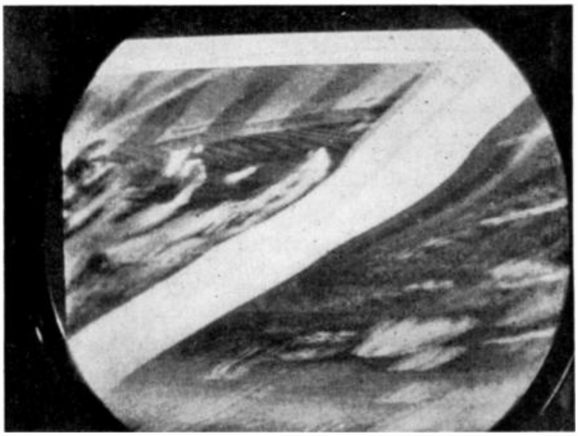


Fig. 13·13 Overloaded picture. Black and white are reversed and picture is out of sync. Diagonal white bar is horizontal blanking. (RCA Institutes, Inc.)

For the opposite case, insufficient AGC bias allows too much gain. Then an i-f stage or the video amplifier can easily have too much signal input for the bias, causing overload distortion. Figure 13·13 shows an overloaded picture. It has reversed black-and-white values and is out of sync. The white diagonal bar is horizontal blanking which should be black and at the side. The picture is out of synchronization vertically and horizontally because the sync pulses are either compressed or lost with the severe amplitude distortion. The overload also can reverse the polarity of modulation for the composite video signal, causing the reversed picture. Usually, there is also 60-cycle buzz in the sound, produced by vertical blanking pulses modulating the sound signal in an overloaded stage. These AGC troubles are summarized in Table 13·1.

Localizing to the AGC circuit. Since the symptoms produced by AGC troubles are similar to troubles in the signal amplifiers, it is helpful to localize the trouble. Taking an example of no picture and no sound, the trouble can be localized by reducing the AGC bias to zero. This can be done by shorting the AGC line directly to chassis ground. If the sound is heard and an overloaded picture results with zero bias, but there is no picture and no sound when the AGC bias is on the amplifiers, the trouble must be excessive bias produced by the AGC circuit. This trouble usually occurs in amplified AGC circuits. With only an AGC rectifier, it cannot produce enough AGC bias to cut off the stages amplifying the signal because without any signal input there will be no AGC bias.

Table 13·1 AGC troubles

TROUBLE	EFFECT
Zero bias on AGC line	Too little bias on r-f and i-f stages causes overload distortion, resulting in overloaded picture and buzz in the sound
Excessive bias on AGC line	Too much bias cuts off r-f and i-f stages, resulting in no picture and no sound

A more general method of localizing troubles in the AGC circuit requires the use of a *battery bias box*. As shown in Fig. 13·14, the bias box provides an adjustable negative voltage to take the place of the AGC bias. Connect the negative output lead of the bias box directly to the AGC line and the positive lead to chassis ground. The low resistance of the bias box effectively shorts the high-resistance AGC circuit, so that it need not be disconnected. If the picture and sound are normal when the bias box supplies the amplifier bias, but not when the AGC bias is functioning, the trouble must be in the AGC circuit. The bias box is also useful for supplying the fixed bias that should be used in i-f alignment.

13·8 AGC bias for transistor amplifiers

The function of AGC bias is the same in controlling the gain of transistorized r-f and i-f amplifiers, but there are important differences in the methods used compared with vacuum tubes. First, it is important to note that transistor collector current is cut off with zero forward bias. Remember that forward bias is the voltage between emitter and base with the polarity required to increase current through the transistor. Therefore, reducing the forward bias toward zero reduces the collector output current. This case is opposite from tubes, which need a high negative grid voltage for cutoff. The conditions for saturation of output current are also different. In a transistor, increasing its forward-bias voltage away from zero increases the collector output current toward saturation.

In one method of applying AGC bias to a transistorized amplifier its forward bias is decreased. Then the circuit is arranged to insert AGC bias of opposite polarity from the forward-bias voltage. This method of reducing gain by less forward bias is called *reverse AGC* because the collector output current decreases with more signal and more AGC bias. Reverse AGC for transistors corresponds to conventional AGC bias for tubes, as output current is reduced toward cutoff with more AGC bias.

However, the cutoff characteristic of a transistor near the zero forward-bias value is very nonlinear, compared with remote-cutoff tubes. The result is severe amplitude distortion for wide variations in reverse AGC bias with strong signals. More common is the method called *forward AGC* to increase collector current toward saturation for reduced gain. Then the AGC bias is inserted with the polarity required to increase the forward bias with more signal.

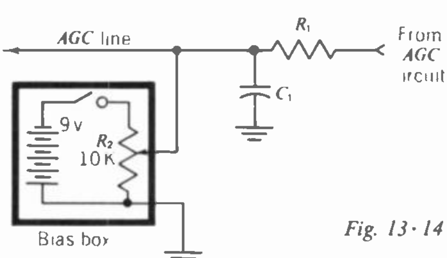


Fig. 13·14 Bias box connected to AGC line.

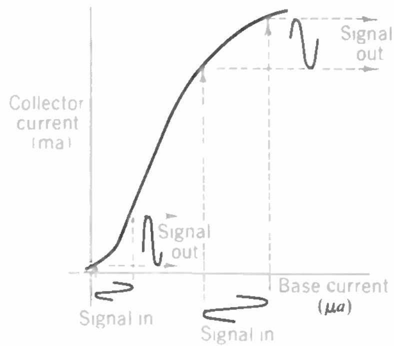


Fig. 13·15 Collector characteristic curve for grounded-emitter circuit.

A comparison of the two methods is illustrated by the graph in Fig. 13·15. This curve shows collector output current for different values of base current in the input circuit, depending on the amount of forward bias in a grounded-emitter amplifier. Compared with bias at the middle linear part of the characteristic curve, note that: (1) at the bottom of the curve collector current has smaller variations, producing less gain, when less forward bias reduces the current toward zero; (2) near the top of the curve, collector-current variations are also reduced at saturation but with better symmetry and less amplitude distortion.

The best linearity and maximum gain result with bias values in the middle of the curve because of its sharp linear rise. This means large proportional changes in collector current in the output for small changes of base current in the input. Moving the bias toward saturation collector current or zero current reduces the gain in either case. However, the curve has less nonlinearity for operation near saturation collector current. Forward AGC, therefore, allows less amplitude distortion for wide variations of signal level and AGC bias, as the operating bias is shifted toward saturation to reduce the gain.

SUMMARY

1. The AGC circuit rectifies the signal to produce d-c bias proportional to signal strength so that the AGC bias can control receiver gain. More carrier signal at the antenna produces more AGC bias to reduce the gain enough to maintain relatively constant output at the video detector.
2. The AGC filter removes video variations so that the AGC bias does not vary with the signal information. A typical filter time constant is about 0.3 sec.
3. The filtered AGC bias is distributed by the AGC line to the controlled stages, usually the r-f amplifier and the first and second i-f amplifiers. The r-f amplifier usually has less AGC bias for a better signal-to-noise ratio.
4. The AGC line is usually connected to each controlled stage through an *RC* decoupling filter, to isolate the signal in each stage from the common impedance in the AGC line.
5. The controlled stages generally use variable-mu tubes. Their remote-cutoff characteristic minimizes amplitude distortion as the AGC voltage varies its gain by shifting the bias.
6. In delayed AGC, no AGC bias is produced until the signal is strong enough to overcome the delay bias. The result is a better signal-to-noise ratio for weak signals.
7. In amplified AGC, more AGC bias is available for better control of gain with strong signals.

8. In keyed or gated AGC (Fig. 13 · 10), the circuit is pulsed on by horizontal flyback pulses. The result is AGC bias relatively free from noise because the AGC rectifier conducts only during horizontal sync pulse time.
9. When AGC bias is obtained from the video detector (Fig. 13 · 6*b*), its plate load resistor must be in the plate circuit to produce negative d-c voltage proportional to average carrier amplitude. This rectified signal voltage is filtered for AGC bias. However, the detector output is also coupled by a parallel path to provide video signal for the video amplifier.
10. The AGC level control sets the least amount of AGC bias needed to prevent overload on the strongest signal.
11. Too little AGC bias causes an overloaded picture and 60-cycle buzz in the sound. The overloaded picture (Fig. 13 · 13) has reversed black and white values and is out of sync.
12. Too much AGC bias causes no picture and no sound, caused by cutoff of one or more stages common to both signals.
13. The gain of transistor amplifiers can be controlled by either reverse AGC bias or forward AGC bias.

SELF-EXAMINATION *(Answers at back of book.)*

Choose (*a*), (*b*), (*c*), or (*d*).

1. The signal must be rectified for AGC bias so that: (*a*) AGC bias can decrease with more signal; (*b*) AGC bias can increase with more signal; (*c*) average white level can be measured; (*d*) good high-frequency response can be obtained.
2. The AGC bias is a d-c voltage that can be measured: (*a*) at the r-f amplifier plate; (*b*) at the kinescope grid; (*c*) at the output of the low-voltage power supply; (*d*) across the AGC filter capacitor.
3. A keyed AGC tube conducts: (*a*) between *H* sync pulses; (*b*) during *H* sync pulses; (*c*) only during vertical blanking pulses; (*d*) only during vertical sync pulses.
4. For the keyed AGC tube in Fig. 13 · 10, the negative AGC bias is taken from: (*a*) cathode; (*b*) control grid; (*c*) suppressor grid; (*d*) plate.
5. Which of the following can cause overloaded picture and buzz in the sound? (*a*) Excessive AGC bias; (*b*) AGC line shorted to chassis; (*c*) brightness control set too low; (*d*) i-f amplifier cut off.
6. When more negative AGC bias is applied to the grid of a remote-cutoff tube: (*a*) plate current increases; (*b*) plate current is saturated; (*c*) gain increases; (*d*) gain decreases.
7. When more forward bias is applied to a transistor: (*a*) collector current decreases; (*b*) collector current increases; (*c*) emitter current decreases; (*d*) base current decreases.
8. Which of the following is false? (*a*) A fast AGC circuit can be affected by the vertical sync pulses. (*b*) A fast AGC circuit helps in minimizing airplane flutter. (*c*) A slow AGC circuit can produce excessive bias because of noise. (*d*) The average value of rectified picture signal is a correct measure of carrier level.
9. Which of the following is true? (*a*) The r-f amplifier generally has more AGC bias than the i-f stages. (*b*) Remote-cutoff tubes generally have higher g_m than comparable sharp-cutoff tubes. (*c*) A transistor is cut off with zero forward bias. (*d*) With forward AGC on a transistor, stronger signal results in less collector current.
10. A voltage gain of 1,000 equals: (*a*) 20 db; (*b*) 60 db; (*c*) 100 db; (*d*) 1,000 db.

ESSAY QUESTIONS

1. Give two advantages of picture AGC and one disadvantage.
2. In a simple AGC circuit, give the function of: (*a*) AGC rectifier; (*b*) AGC filter; (*c*) AGC line; (*d*) AGC decoupling filters.
3. What is the purpose of delayed AGC?
4. Give two methods of having amplified AGC.
5. When the carrier level increases, what is the effect on the amount of negative AGC bias voltage and the gain of the receiver?

6. When the carrier level increases, what is the effect on the amount of forward bias of the controlled transistor amplifiers with: (*a*) forward AGC; (*b*) reverse AGC?
7. When the AGC bias line is shorted to ground what is the effect on the amount of bias and gain of the controlled stages?
8. Describe how to connect a bias box to the receiver, noting the polarities. Give two uses for a bias box.
9. Draw the circuit for a bias box using a step-down transformer and half-wave rectifier for a-c power input, instead of batteries.
10. Explain briefly what a keyed AGC circuit is. What is its advantage?
11. In a keyed AGC circuit: (*a*) Where are the keying pulses taken from and where are they applied? (*b*) What signal is rectified? (*c*) Where is the AGC bias voltage taken from?
12. Referring to the keyed AGC circuit in Fig. 13·10: (*a*) How much would a d-c voltmeter read from each electrode to chassis ground? (*b*) How much is the grid-cathode bias? (*c*) If the control-grid voltage was +110 volts, how much should the cathode voltage be for the same bias?
13. Describe briefly how to adjust the AGC level control by watching the picture, for the typical case of medium signal strength on all stations.
14. Describe briefly why excessive AGC bias can result in no picture and no sound. Give two troubles in components that can cause this.
15. Describe briefly why zero AGC bias can result in overloaded picture and buzz in the sound. Give two troubles in components that can cause this.
16. What are two characteristics of an overloaded picture?
17. Where would a d-c voltmeter be connected to read the bias on the AGC line in the circuits of Figs. 13·1, 13·6*a* and *b*, and 13·10?
18. In Fig. 13·14, if R_1 opened, what would be the effect on the amount of bias on the AGC line?
19. Give the function of each component in Fig. 13·6*b*, with two functions for R_1.
20. Give the function of each component in Fig. 13·10.
21. In Fig. 13·1, what would you assume are the omitted stages corresponding to V_2, V_4, and V_5?

PROBLEMS *(Answers to odd-numbered problems at back of book.)*

1. Calculate the time constant in seconds for the following *RC* combinations: (*a*) 2 megohms and 0.05 μf; (*b*) 150,000 ohms and 0.001 μf; (*c*) 300,000 ohms and 0.001 μf; (*d*) 300,000 ohms and 0.47 μf.
2. Draw *RC* filters for the following: (*a*) *RC* is 0.2 sec with *R* of 0.5 megohm; (*b*) *RC* of 1,270 μsec with *C* of 0.002 μf.
3. Calculate the decibels corresponding to the following voltage ratios: (*a*) 10:1, (*b*) 10,000:1; (*c*) 2:1; (*d*) 1.4:1, (*e*) 5:1; (*f*) 20:1; (*g*) 50:2.
4. Calculate the voltage ratios corresponding to: (*a*) 6 db; (*b*) 10 db; (*c*) −6 db; (*d*) 50 db; (*e*) 38 db.
5. A signal voltage is amplified by the factor of 10,000 and then attenuated by −10 db. (*a*) Calculate the db corresponding to the gain of 10,000. (*b*) How much is the overall db gain? (*c*) How much is the overall voltage gain?
6. Refer to the tube manual for the family of plate characteristic curves of any typical miniature glass remote-cutoff pentode tube. Estimate the g_m for five different values of control-grid voltage e_c and tabulate these values. Then draw a graph plotting g_m against e_c. Also, calculate the gain for the five values of g_m, assuming an i-f plate load impedance of 5,000 ohms.
7. Antenna signal of 100 μv produces 1 volt at the video detector. Calculate the overall voltage amplification and db gain.
8. Do the same for 4,000 μv antenna signal producing 2 volts at the detector. Why is there less gain now?

Chapter 14 Sync separation

The synchronizing pulses are included in the composite video signal transmitted to the receiver in order to time the scanning with respect to the camera signal variations. At the broadcast station, the synchronizing signal generator produces pulses to time the scanning in the camera tube; the same generator supplies the synchronizing pulses that are added to the video signal transmitted for the receiver. These synchronizing pulses, generally called *sync*, are separated by the sync circuits in the receiver and coupled to the deflection circuits in order to control the timing of the scanning for the picture tube. As a result, the picture information reproduced on the screen of the picture tube is in the same relative position as on the image plate of the camera tube, since the scanning for both tubes is synchronized by one common source—the sync generator at the broadcast station. The amount of time for the transmitted signal to travel to the receiver has no effect on synchronization because the sync pulses must be present at the same time as the camera signal variations in the composite video signal at the receiver. In short, the sync is necessary to time the scanning in the raster with respect to picture information in the video signal.

The sync separation circuits in the receiver provide vertical sync to time every scanning field correctly at 60 cps, and horizontal sync to time the scanning lines at 15,750 cps. The sync voltage does this by controlling the frequency of the vertical and horizontal deflection oscillators. It is important to remember that the deflection circuits in the receiver can produce vertical and horizontal scanning to form the raster with or without sync. However, the position where the picture information is reproduced on the raster depends upon the vertical and horizontal synchronization. The effects of no sync are shown in Figs. 14·1 and 14·2.

14·1 Vertical synchronization of the picture

When every scanning field is produced at the correct time, the frames

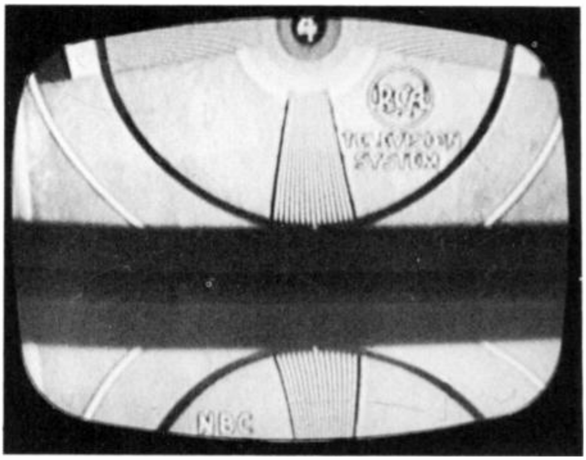
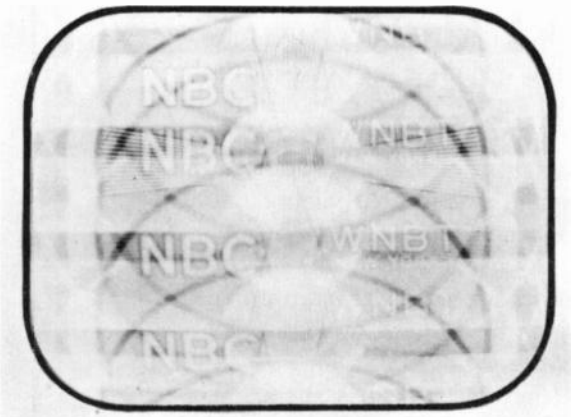


Fig. 14 · 1 Vertical scanning not synchronized, resulting in no vertical hold. (a) Picture slowly slips frames vertically. (b) Picture rolls fast vertically.

(a)



(b)

are superimposed on each other. The result is a steady picture locked in frame on the kinescope screen. If the vertical scanning is not locked in at the 60-cps sync frequency, successive frames cannot overlap. Instead, the frames are displaced either above or below the first frame. Without vertical synchronization, therefore, the picture on the kinescope screen appears to roll up or down.

The faster the picture rolls the farther the vertical scanning frequency is from 60 cps. If the vertical scanning frequency is just slightly off the 60-cps synchronizing rate the picture will be recognizable, as shown in Fig. 14 · 1*a*, but it slips out of frame slowly and continuously. In *b* the picture is rolling fast. The wide black bar across the picture in Fig. 14 · 1*a* is produced by vertical blanking, which now occurs during vertical trace time because the scanning is out of sync. In some receivers white retrace lines appear while the picture is rolling. These are horizontal scanning lines produced during the vertical flyback. They are visible because vertical retrace is not occurring during vertical blanking time, without synchronization. When the picture is in vertical sync, though, it stays still, locked in frame vertically, with the black vertical blanking at the top and bottom edges, and the retrace lines are blanked out.

14 · 2 Horizontal synchronization of the picture

When every scanning line is produced at the correct time, the line structure of the reproduced picture holds together to provide a complete image that stays still horizontally. If the horizontal scanning is just slightly

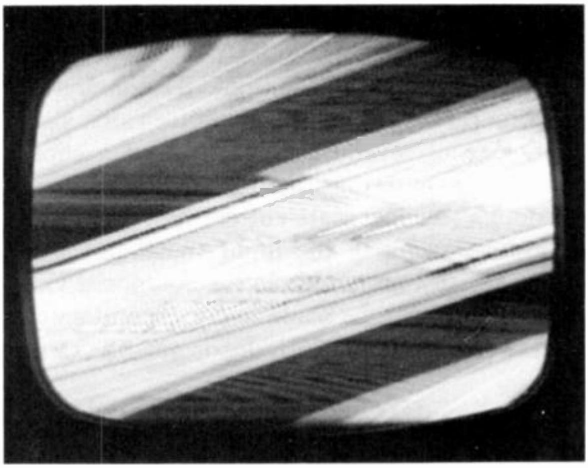


Fig. 14·2 Horizontal scanning not synchronized, resulting in no horizontal hold. The picture can have more or less than the two diagonal black bars.

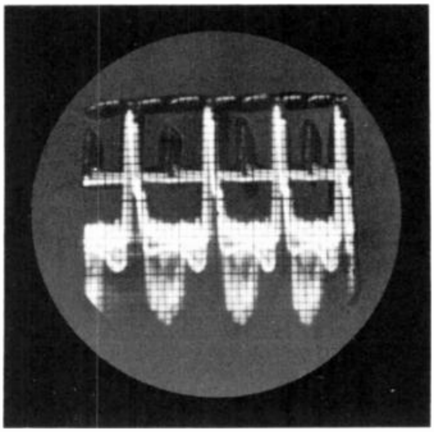
off the 15,750-cps sync frequency, the line structure is complete but the picture slips horizontally, as the picture information on the lines is displaced horizontally in successive frames. The faster the picture slides horizontally the farther the scanning frequency is from 15,750 cps. When the horizontal scanning frequency departs from the 15,750-cps synchronizing frequency by 60 cycles or more, the picture tears apart into diagonal segments, as shown in Fig. 14·2. The black diagonal bars represent parts of horizontal blanking, which is normally at the sides of the picture. The more bars, the farther the horizontal oscillator is from the horizontal sync frequency of 15,750 cps.

14·3 Separating the sync from the video signal

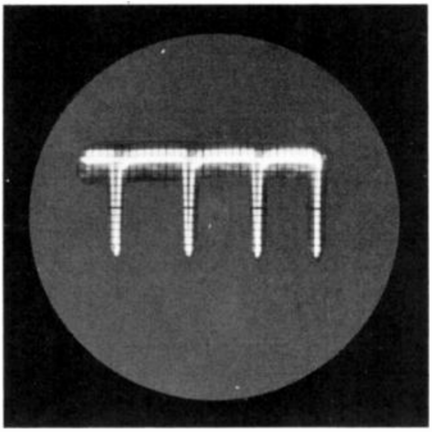
Figure 14·3 shows oscillograms of the composite video signal input to

Fig. 14·3 Oscillograms of composite video signal input to sync separator and clipped sync output voltage. Only horizontal sync can be seen with oscilloscope internal sweep frequency at 15,750/4 cps but vertical sync and equalizing pulses are also in the signal. (a) Video signal with positive sync polarity. (b) Separated sync with negative polarity. (RCA.)

(*a*)



(*b*)



a sync separator stage and its sync voltage output. This amplitude separation of the synchronizing pulses from the composite video signal can be accomplished by any one of several arrangements of a clipper stage. All are fundamentally the same in that the tube is biased beyond cutoff by an amount great enough to allow plate-current flow and output signal only for the most positive swing of the input signal. Since the synchronizing pulses have the greatest amplitude in the composite video signal, the output of the clipper stage can be made to correspond only to the synchronizing pulses of the input. A diode, triode, or pentode can be used.

Grid-leak bias sync separator. Figure 14·4 illustrates the operation of a typical triode clipper using grid-leak bias. The grid input voltage is composite video signal of positive sync polarity. Then the sync pulses can drive the instantaneous grid voltage positive to produce grid current for grid-leak bias and maximum plate current for sync output. The negative grid-leak bias voltage automatically adjusts itself to the value that allows just the peaks of the input signal to drive the grid voltage slightly positive. As a result, the tips of the sync pulses are clamped at approximately zero grid voltage.

With the sync pulses in line at a constant voltage level in the grid circuit, it is now necessary only to have the grid-cutoff voltage of the amplifier correspond to the sync voltage of the grid signal. There are two factors to consider, the amount of signal and the grid-cutoff voltage. With a large signal voltage, taken from the kinescope grid circuit, the composite video signal can easily overdrive the amplifier. As shown in Fig. 14·4, only the sync voltage drives the clipper grid voltage less negative than cutoff. Then the pulses in the grid signal produce plate-current output just for the sync. If a smaller grid-cutoff voltage is necessary with less grid signal, the plate voltage can be reduced for a triode, or screen-grid voltage

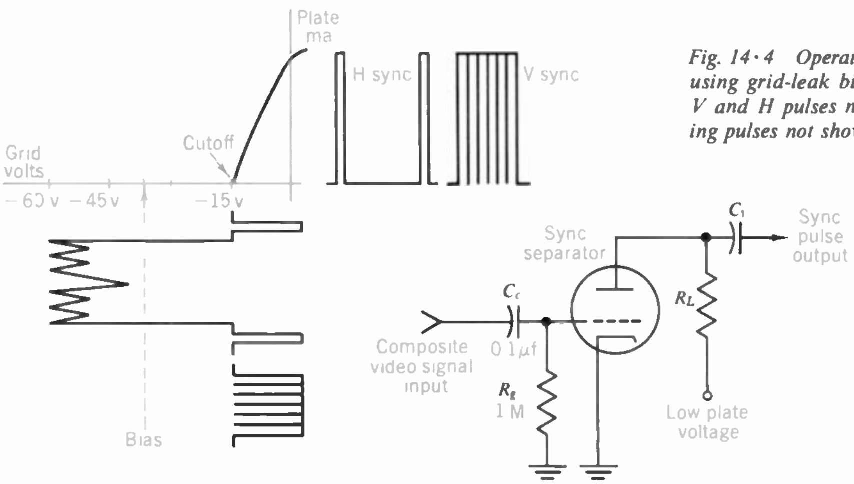


Fig. 14·4 Operation of sync separator using grid-leak bias clipper amplifier. V and H pulses not to scale. Equalizing pulses not shown.

reduced for a pentode. It should be noted that the composite video signal can be clipped at any amplitude between pedestal level and tip of sync and still provide separated sync output because plate current would flow only during sync pulse time.

Only the sync pulses are amplified by the clipper stage. The remainder of the grid signal voltage is cut off as it does not produce plate current. If there is a question of what happens to this part of the signal, the answer is—nothing. An oscilloscope connected to the clipper grid circuit will show the entire composite video signal input. In the plate circuit, though, the oscilloscope shows only the separated sync output. These input and output signals are shown in Fig. 14·3. Furthermore, the fact that the separator stage separates the sync voltage does not prevent the kinescope grid circuit from having its normal composite video signal input.

In Fig. 14·4, the amount of grid-leak bias shown is -35 volts, approximately equal to the peak positive swing of the input voltage. Since the grid-cutoff voltage here is -15 volts, the video signal must drive the instantaneous grid voltage at least 20 volts more positive than the bias to produce plate current. Only the sync pulses of the input signal are positive enough to drive the grid voltage inside of cutoff. As a result, pulses of plate current flow just for the sync pulses. This plate current through R_L produces signal voltage output consisting of only the separated sync pulses. The camera signal variations are missing from the output because this part of the input signal is more negative than cutoff and cannot produce plate current.

The negative-polarity sync output voltage is amplified and inverted, compared with the positive sync input. Both the vertical and horizontal sync pulses are clipped, as they have the same voltage amplitude in the composite video signal. Figure 14·3 shows horizontal sync but if the oscilloscope internal sweep frequency is set at 60/4 cps you will see four vertical sync pulses. The equalizing pulses are also clipped. In summary, the voltage output of the clipper is total separated sync including horizontal, vertical, and equalizing pulses, but without the camera signal.

Amplitude separation. Clipping the sync pulses from the composite video signal is amplitude separation of the sync, and the stage having this function is a *sync separator.* This separation is possible because the sync pulses have higher amplitude than the camera signal variations. The separated sync can be clipped again by an additional stage, generally called a sync *clipper,* to limit the effect of noise in the sync.

Waveform separation. The total separated sync voltage includes horizontal, vertical, and equalizing pulses, as shown in Fig. 14·5. All have the same amplitude but they differ in frequency and pulse width. Most important, the horizontal and vertical sync pulses have different waveforms so that they can be separated from each other. The horizontal sync with a narrow pulse width of 5.1 μsec, repeated at 15,750 cps, represents a high-frequency signal, compared with the 60-cps rate of the wide vertical pulses. Therefore, they can be separated by *RC* filters. A low-pass

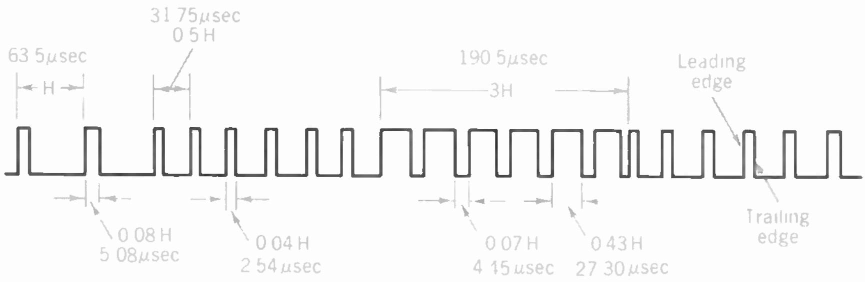


Fig. 14·5 Waveform of the synchronizing pulses. H is the horizontal line time of 63.5 μsec.

filter with a time constant long enough to bypass the horizontal sync allows the vertical sync alone to be coupled to the vertical deflection oscillator. This is the function of the *RC* integrator[1] circuit in Fig. 14·6. In addition, an *RC* coupling circuit with a time constant too short for the vertical sync can couple the horizontal sync to the horizontal AFC circuit.

Sequence of sync separation. To summarize the amplitude and waveform separation, Fig. 14·6 shows how the desired synchronizing signals are obtained to hold the vertical and horizontal scanning at the correct frequency. First the sync in the composite video signal is separated by one or more clipper amplifiers. The total separated sync is then coupled to two parallel circuits for waveform separation. The R_1C_1 integrator receives all the sync pulses, but C_1 builds up charge only during the relatively wide vertical sync pulse. The equalizing pulses help in making the integrated voltage across C_1 the same for even and odd fields. As a result, the vertical deflection oscillator is triggered at 60 cps to hold the picture locked in vertically. For horizontal hold, all receivers use automatic frequency control for the horizontal deflection oscillator. As explained in Chap. 16, the horizontal AFC circuit is better than triggered sync for holding the horizontal oscillator at the correct frequency in the presence of noise pulses.

It is important to remember that the synchronizing signals do not scan. The vertical and horizontal deflection oscillators generate the deflection voltage needed to drive the deflection amplifiers to produce scanning current in the deflection yoke. The only function of the sync is to time the scanning. The vertical sync pulses trigger the vertical oscillator at 60 cps to lock in the vertical scanning. The horizontal sync provides the timing information needed by the horizontal AFC circuit to hold the horizontal oscillator frequency at 15,750 cps.

14·4 Integration of the vertical sync

An integrating circuit, having a time constant long compared with the duration of the horizontal pulses but not with respect to the vertical pulse

[1] See Appendix E for explanation of integrator and differentiator circuits.

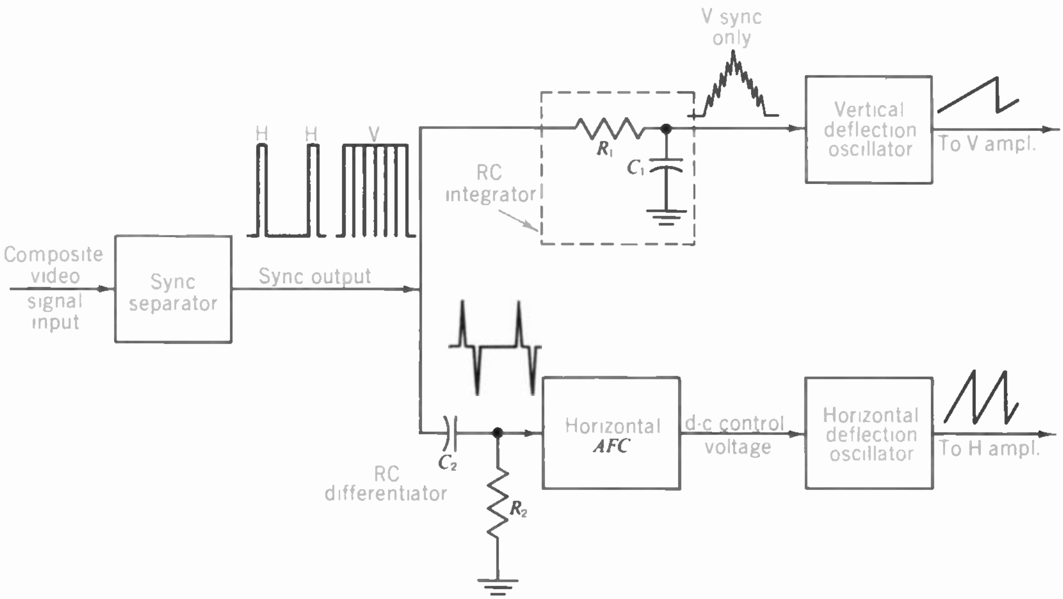


Fig. 14·6 Sequence of sync separation including clipper stage for amplitude separation followed by waveform separation of vertical and horizontal sync.

width, is used to provide the waveform separation needed for vertical synchronization. As shown in Fig. 14·7, the total sync voltage is coupled to the *RC* integrating circuit. However, the output voltage across the C_1 provides vertical synchronizing voltage alone.

The time constant of the *RC* circuit in Fig. 14·7 is 100 μsec. The horizontal pulse width is 5.1 μsec. Therefore, C_1 can charge to only a small percentage of the applied voltage for the short period during which the horizontal pulse is applied. The period between horizontal pulses, when no voltage is applied to the *RC* circuit, is so much longer than the horizontal pulse width that the capacitor has time to discharge almost down to zero during this time. The equalizing pulses apply voltage at half-line

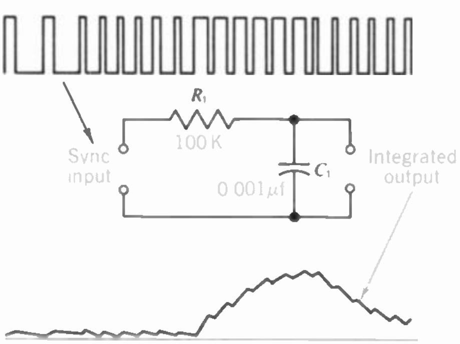


Fig. 14·7 Integration of vertical sync voltage. RC time constant here is 100 μsec.

intervals, but their duration is only one-half the horizontal pulse width, and they cannot charge C_1 to any appreciable voltage.

When the vertical pulse is applied, however, the voltage across C_1 can build up to the value required for triggering the vertical deflection oscillator. The serrated vertical pulse consists of six individual pulses, each of approximately 27 μsec. In this time, C_1 charges to 27 per cent of the applied voltage, as obtained from the universal charge curve in Appendix E. During the serration, applied voltage is removed and the capacitor discharges. This is only for 4.4 μsec, however. Therefore, C_1 loses little of its charge before the next 27-μsec pulse provides sync voltage to recharge the capacitor. Thus the integrated voltage across C_1 builds up toward maximum amplitude at the end of the vertical pulse, followed by a decline practically to zero for the equalizing pulses and horizontal pulses that follow. The result is a pulse of the triangular waveshape shown for the complete vertical synchronizing pulse.

You can see the 60-cps integrated vertical sync pulses with an oscilloscope at the grid of the vertical oscillator (Fig. 14·11). However, the oscillator must be off, as its grid voltage is much larger than the sync voltage. Or, with the oscillator on but slightly out of sync, you can see the vertical sync pulse moving across the grid-voltage waveform.

Effect of the equalizing pulses. Their function is improving the accuracy of vertical synchronization in even and odd fields for good interlace (see Fig. 14·8). Note that, with the equalizing pulses, the voltage across C_1 can adjust itself to practically equal values for even and odd fields, although there is a half-line difference in time before the vertical pulse begins. Then the integrated output for vertical sync is the same for even and odd fields.

Cascaded integrator sections. A very long time constant for the integrating circuit removes the horizontal sync pulses but reduces the vertical sync amplitude across the integrating capacitor. Also, the rising edge on

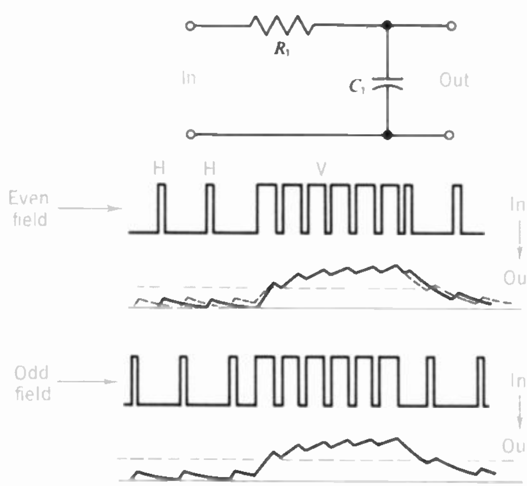


Fig. 14·8 Integrated output across C_1 for even and odd fields without equalizing pulses. Dotted waveform in upper figure is the lower waveform superimposed. Horizontal dashed lines across both waveforms are the level to which C_1 voltage must rise to trigger vertical oscillator.

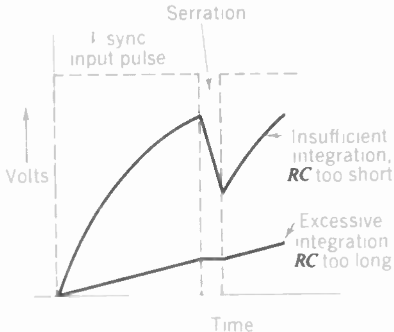
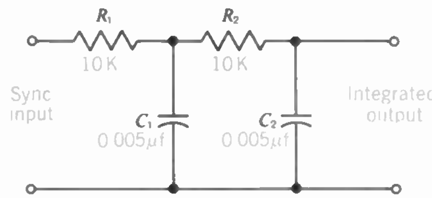


Fig. 14·9 Integrated vertical sync voltage for time constant too long and too short.

Fig. 14·10 Two-section integrating circuit.



the integrated vertical pulse is not sharp enough. With a time constant that is not long enough, the horizontal sync pulses cannot be filtered out and the serrations in the vertical pulse produce notches in the integrated output. The notches should be filtered out because they give the integrated vertical synchronizing signal the same amplitude value at different times. Figure 14·9 illustrates too much integration and not enough integration.

For good vertical synchronization, the integrated vertical pulse should have sufficient amplitude and rise quickly with a smooth increase up to the amount of voltage required to trigger the vertical deflection oscillator. In order to provide a compromise between good filtering of the horizontal sync with a long time constant and the sharper leading edge and higher amplitude produced by a shorter time constant, the integrating circuit generally consists of two or three sections in cascade, as shown in Fig. 14·10. This is a two-section integrating circuit with each RC section having a time constant of 50 μsec. The operation of the circuit can be considered as though the R_1C_1 section provided integrated voltage across C_1 that is applied to the next integrating section R_2C_2. The overall time constant for both sections is long enough to filter out the horizontal sync, while the shorter time constant of each section allows the integrated voltage to rise more sharply because each integration is performed with a time constant of 50 μsec. If one section is open, there will be less filtering but actually more vertical sync voltage because of a shorter time constant.

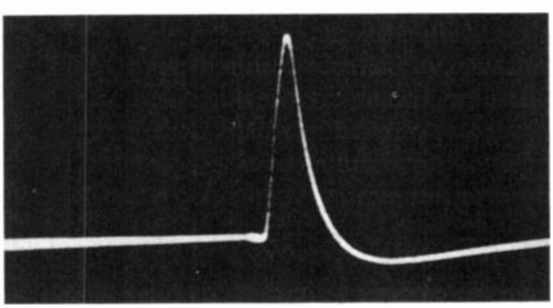


Fig. 14·11 Oscillogram of vertical sync voltage output from cascaded integrator. Peak-to-peak amplitude 10 volts. Oscilloscope sweep at 60 cps.

14 · 5 Noise in the sync

Noise pulses are produced by ignition interference from automobiles, arcing in neon signs or arcing brushes in motors, and atmospheric static. The man-made noise is either radiated to be added to the received signal or is coupled to the receiver through the power line. Especially with weak video signal, the noise can act as false synchronizing pulses. In the vertical integrator, the noise pulses charge the capacitor and allow it to reach the voltage required to trigger the vertical oscillator too soon. Then the noise makes the picture roll temporarily, until the sync locks in the oscillator again. For horizontal synchronization, the noise can be mistaken for horizontal sync pulses. Then the picture tears into diagonal segments during the loss of horizontal sync. Furthermore, when the noise pulses have much higher amplitude than the sync voltage, the increased grid-leak bias caused by *noise setup* makes the bias on the sync separator stage too negative for the signal level. The result is weak sync or no output at all, temporarily, from the sync separator. Then the picture does not hold still until the synchronization is established again.

In order to reduce the effect of noise on the synchronization, the sync circuits generally include one or more of the following:

1. Sync clipper stage for the separated sync. The additional clipping prevents noise pulses from having higher amplitude than the sync.
2. Double time constant for the grid of the sync separator. A long time constant is necessary for coupling the vertical sync but a shorter time constant reduces noise setup.
3. Noise-inverter or noise-canceler stage. Such a stage can cancel most of the high-amplitude noise pulses in the sync circuits.

It should be noted that reducing the effect of noise in the sync circuits does not prevent noise pulses in the video signal from producing horizontal streaks in the picture. The purpose of the noise-reduction circuits in the sync section is to enable the picture to hold still in the presence of noise.

Fortunately, the noise pulses do not have to be reduced to zero amplitude. Good sync can be obtained when the noise pulses are limited to the same amplitude as sync voltage. The reason is that an interfering noise pulse occurring between sync pulses must have much more amplitude than the sync to trigger the oscillator in the middle of a cycle, when the oscillator is not ready for triggering.

Sync clipper. The sync output of the separator can be clipped and amplified again in the next stage. The purpose is to provide sharp sync pulses with high amplitude, free from noise and without any camera signal. Clipping in successive stages allows the top and bottom of the sync pulses to be clipped by cutoff, which is sharper than limiting by plate saturation. The operation of a sync separator followed by a sync clipper is illustrated in Fig. 14 · 12.

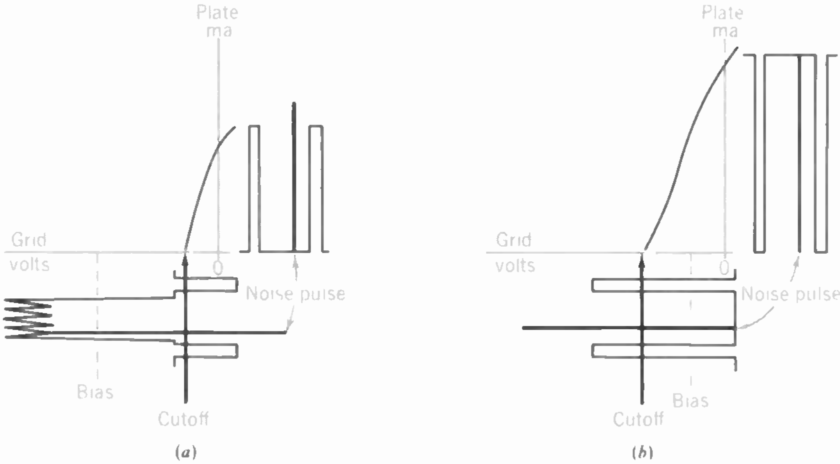


Fig. 14·12 Signal with noise pulse in successive sync stages. (a) Composite video with positive sync and noise pulse at grid of sync separator. (b) Separated sync of negative polarity with noise pulse at grid of sync clipper.

Sync separator time constant. The time constant of the grid-leak bias circuit in the input to the separator must be long enough to maintain the bias from line to line and through the time of the vertical synchronizing pulse, in order to maintain a constant clipping level. Typical values are 0.1 μf for C_c and 1 megohm for R_g, providing a time constant of 0.1 sec. These values allow the bias to vary from frame to frame for different brightness values, keeping the tip of sync clamped at zero grid voltage. However, a time constant of 0.1 sec is too long for the bias to follow amplitude variations produced by noise pulses occurring between lines, with a frequency higher than the horizontal sync pulses.

Noise pulses in the input signal to the grid of the sync separator can increase the amount of grid-leak bias produced. Then the bias is more negative than required for clipping the sync from the composite video signal. As a result, the gain is reduced for the separated sync. In addition, the noise is amplified with the sync. If the time constant of the grid-leak bias circuit in the sync separator is made shorter for noise pulses, it will not be long enough to maintain the bias between sync pulses, especially during the vertical sync pulse time. The result may be inadequate sync separation during and immediately after the vertical pulse. Some receivers have separate horizontal and vertical sync separator circuits with a much shorter time constant for horizontal sync separation.

In order to reduce the effect of high-frequency noise pulses on the grid-leak bias for the sync separator, an *RC* circuit with a short time constant can be added to the grid circuit of the sync separator, as shown in Fig. 14·13. The grid coupling circuit R_gC_c provides normal grid-leak bias, with a time constant of 0.1 sec, for the sync signal. The small 150-$\mu\mu$f C_1 and the 270,000-ohm R_1 provide a short-time-constant circuit. Then the grid

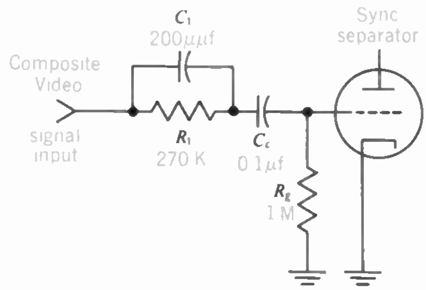


Fig. 14·13 Short-time-constant network R_1C_1 to reduce impulse-noise amplitude at grid of sync separator.

leak bias can change quickly to reduce the effect of noise pulses in the input to the sync separator. C_1 can charge fast when noise pulses produce grid current, increasing the bias for noise. The change in voltage across C_c and C_1 is inversely proportional to their capacitance values. Therefore, a noise pulse will charge C_1 to a voltage 500 times more than across C_c. Since the R_1C_1 time constant is 54 μsec, C_1 can discharge through R_1 between sync pulses. The bias then remains at the voltage produced by R_gC_c for the sync.

14·6 Sync separator circuits

The sync stages use the *RC*-coupled amplifier circuit. Values of 5,000 to 50,000 ohms for the plate load resistor minimize the effect of shunt capacitance. Good high-frequency response up to about 1 Mc is important for amplifying harmonics of the horizontal sync pulses for sharp edges. Grid-leak bias is generally used to reinsert the d-c component needed to line up the sync pulses for a constant clipping level.

Sync separator, clipper, and phase splitter. In Fig. 14·14, composite video signal from the output of the video amplifier, with positive sync polarity, is coupled to the grid of the first sync stage, which uses one triode section of the 12AU7 twin triode V_{304}. This stage is a grid-leak bias sync separator to provide separated sync voltage of negative polarity in the plate circuit, without camera signal. The negative sync output from the plate is coupled through C_{409} to the grid of the next sync stage, which amplifies the separated sync and clips the negative side of the sync pulses. The amplified output in the plate of this clipper is positive sync that is coupled through C_{410} to the grid of the sync inverter.

The inverter operates as a phase splitter to provide push-pull sync output voltage, because the horizontal AFC circuit in this chassis requires sync voltages of equal amplitude and opposite polarity. Negative sync voltage from the plate of the phase inverter, across the 2,200-ohm plate load resistor R_{419}, is coupled by C_{412} to one side of the AFC circuit, while positive sync voltage of equal amplitude across the 2,200-ohm cathode resistor R_{424} is coupled by C_{413} to the opposite side of the AFC circuit. The bypass C_{414} returns R_{424} to chassis ground for the horizontal sync signal, so that no horizontal sync voltage will be developed across R_{425} to unbalance the push-pull circuit. The total sync voltage of positive polarity

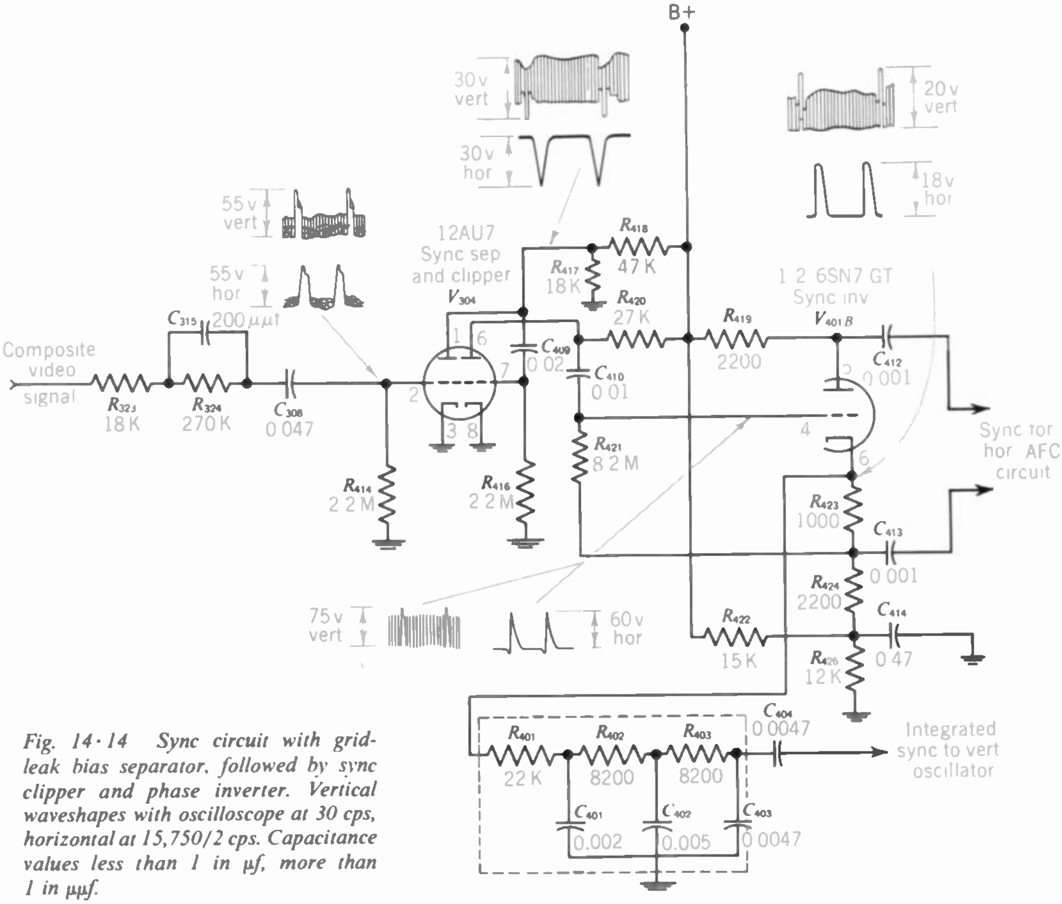


Fig. 14·14 Sync circuit with grid-leak bias separator, followed by sync clipper and phase inverter. Vertical waveshapes with oscilloscope at 30 cps, horizontal at 15,750/2 cps. Capacitance values less than 1 in μf, more than 1 in μμf.

at the cathode, across R_{423}, R_{424}, and R_{425}, is connected to the three-section integrating circuit for vertical synchronization. The integrated voltage across C_{403} is vertical sync only, for the grid of the vertical deflection oscillator, to synchronize the vertical scanning. The dotted lines around the integrator indicate it is a printed circuit in one unit.

In the cathode circuit of the sync inverter, R_{425} forms a voltage divider with R_{422}, across the B supply voltage, applying positive voltage across R_{425} to the cathode. This reduces the plate-to-cathode voltage for a lower grid-cutoff voltage, allowing additional clipping of the sync. Notice that the grid circuit of the inverter returns to the junction of R_{423} and R_{424} in the cathode, making only the cathode voltage across R_{423} effective as bias for the grid signal.

In summary, then, video signal with positive sync polarity is coupled into V_{304}, which combines sync separator and sync clipper stages. Its positive sync output is coupled into the triode phase inverter. At the cathode, the sync voltage is also positive, as in a cathode follower. The sync is integrated to provide positive triggering pulses for the grid of the vertical blocking oscillator. Push-pull sync voltage is taken from plate and cathode for the horizontal AFC circuit.

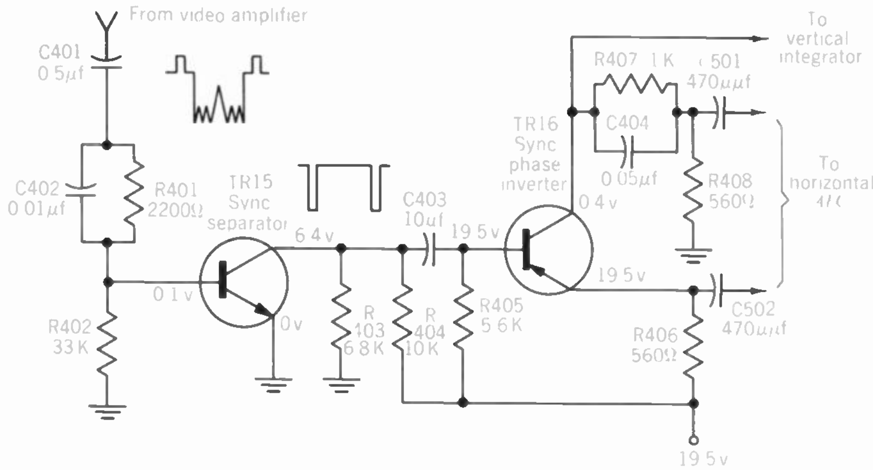


Fig. 14·15 Transistorized sync circuits with separator and phase splitter. D-C voltages at transistor sockets with no signal input. (Motorola chassis TS-432.)

Transistorized sync separator and phase splitter. The transistorized sync circuit in Fig. 14·15 is similar to Fig. 14·14, but there is just the one transistor sync separator TR15. Composite video signal with positive sync polarity is coupled to the base of this NPN transistor connected in a grounded-emitter circuit. Positive voltage at the base is the polarity needed for forward bias. Without signal, the transistor is cut off, with -0.1 volt at the base. The positive sync signal provides enough forward voltage to allow conduction. Therefore, the output is separated sync voltage. Its polarity is inverted, resulting in negative sync voltage to drive the phase inverter.

TR16 is also in a common-emitter circuit, but the PNP transistor requires negative forward bias at the base. The negative sync input voltage is amplified and inverted to produce positive sync output at the collector. However, the signal voltage across the emitter load resistor R_{406} has the original negative input polarity. All the positive sync voltage at the collector is used for input to the vertical integrator. However, this voltage is divided down to produce the same amount of sync across the 560-ohm R_{408} and the 560-ohm R_{406} for push-pull sync into the horizontal AFC circuit.

Noise inverter for the sync signal. In Fig. 14·16, signal is coupled to the sync separator V_2 in two paths, one for composite video and the other for inverted noise pulses. One branch into V_2 supplies normal signal from the video amplifier. In addition, the clipped output of the noise-inverter stage V_1 goes to the sync separator. Notice that the V_1 output includes only noise pulse voltage, with opposite polarity from the video signal input of V_2. As a result, noise pulses are canceled to a large extent in the sync signal output.

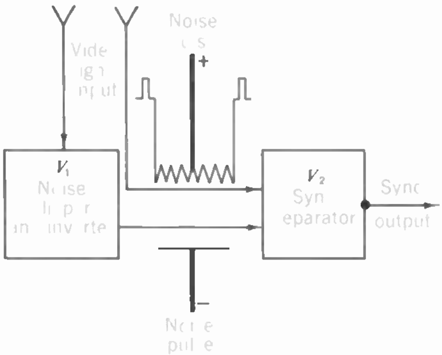
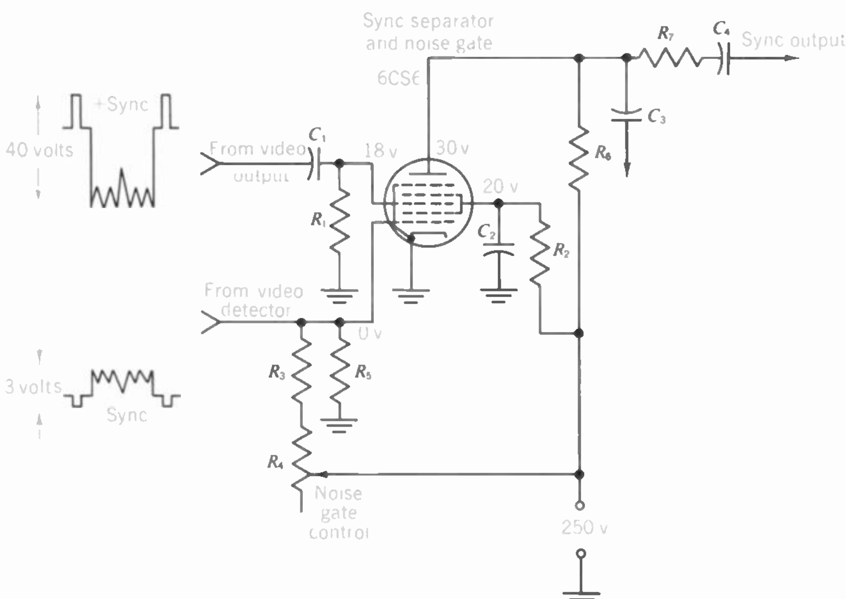


Fig. 14·16 Noise-inverter circuit with waveshapes showing noise pulses canceling in sync separator.

The noise inverter clips only noise pulses because its fixed bias is negative enough to cut off this stage, except for amplitudes greater than the tip of sync. Composite video signal for the noise inverter can be obtained from the video detector and video amplifier.

Pentagrid sync separator and noise gate. The circuit in Fig. 14·17 uses a tube having two control grids with sharp cutoff. The first grid next to the cathode is one control grid and the other is grid 3. Grids 2 and 4 are connected internally to serve as a screen grid and grid 5 is the suppressor, providing pentode characteristics for the two control grids. Cutoff for each control grid is approximately -2 volts. It is important to note that either control grid can cut off plate current. If grid 1 is more negative than cut-

Fig. 14·17 Pentagrid tube used as sync separator and noise gate.



off while grid 3 is not, plate current cannot flow. Both grids must be less negative than cutoff at the same time to produce plate current. We can consider this a gated circuit, therefore. Both control-grid gates must be on to produce output in the plate circuit.

Video signal is applied to each control grid. Grid 3 has enough composite video signal to allow clipping the sync voltage. Grid 1 has video signal of opposite polarity, to serve as a noise gate. Negative noise pulses at this grid can cut off plate current, reducing noise voltages in the sync output.

Composite video signal with positive sync polarity from the plate of the video amplifier is coupled to grid 3 by C_1 and R_1. The negative grid-leak bias allows only the peak positive sync voltage to produce plate current. The result is separated sync output across the plate load resistor R_6, as grid voltages more negative than -2 volts are beyond cutoff. Assuming grid 1 does not have any high-amplitude noise voltage so that plate current can flow, the part of the circuit using grid 3 functions as a conventional grid-leak bias sync separator. The sync amplitude in the output is about 25 volts peak to peak.

Grid 1 has composite video signal of low amplitude, with negative sync polarity, from the video detector. The bias on grid 1 is set by R_4 to allow sync voltage for normal signal to remain less negative than cutoff. This adjustment allows plate current to flow for producing separated sync output from the signal at grid 3. However, high-amplitude noise pulses can drive the grid 1 voltage more negative than cutoff. For the noise, then, plate current is cut off. As a result, noise pulses in the composite video signal input are not present in the separated sync output. Furthermore, the grid-leak bias on grid 3 cannot be increased by noise setup, since grid current does not flow while current in the tube is cut off by grid 1.

The noise-gate control R_4 may be labeled a *fringe, range,* or *local-distant* adjustment, on the rear apron of the chassis, as it is set for the signal level in different areas. The idea is to set R_4 for the least positive bias that still allows normal sync output. Less positive bias provides better noise rejection with weak signals, allowing noise pulses to drive grid 1 negative enough to cut off plate current. However, too little positive bias may reduce the separated sync output or eliminate the sync completely on strong signals. The control can be set approximately by adjusting to the point where the picture is out of sync for the strongest signal, and then back off the control just enough for normal sync.

Sync separator and gated AGC. The noise-gating function can also be helpful in reducing the effects of noise in the automatic gain control circuit of the receiver. Therefore, special tubes are manufactured to combine the keyed AGC stage and sync separator stage with a noise-gate circuit common to both. These tubes are twin sharp-cutoff pentodes, as shown in Fig. 14·18. The cathode is common to both sections; grid 1 serves as a noise gate for the sync separator and AGC; grid 2 is a common screen grid. Grid 3 in each section is an individual control grid, and there are two plates

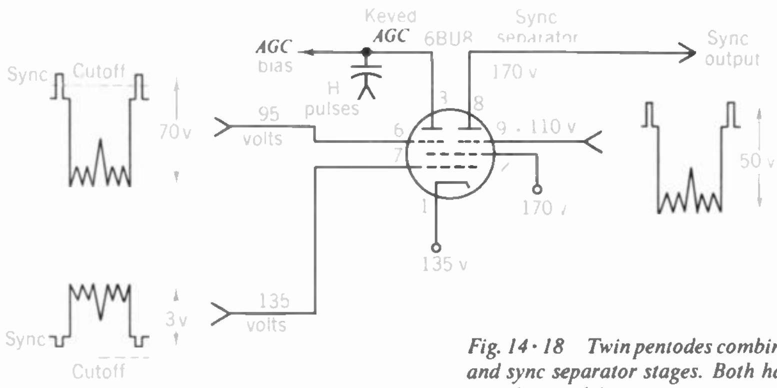


Fig. 14·18 Twin pentodes combining keyed AGC and sync separator stages. Both have noise-gating signal at grid 1.

for separated sync output from one pentode unit and AGC voltage from the other. The sync separator circuit is the same as in Fig. 14·17; the keyed AGC circuit is the same as explained previously for Fig. 13·10.

Note the d-c voltages in Fig. 14·18. Pin 6 is at +95 volts because of d-c coupling to the video amplifier plate. The cathode is at +135 volts to make the grid bias at pin 6 equal to −40 volts, for the AGC section. Pin 7 is also at +135 volts, making this grid-cathode bias zero for the noise-gate video signal voltage. However, this grid voltage can be varied slightly by the noise-gate adjustment. The plate of the AGC section is keyed into conduction by horizontal flyback pulses. In the separator, its plate is at +170 volts to provide 35 volts from plate to cathode. The screen grid has the same voltage. Pin 9 at +110 volts has grid-cathode bias equal to −25 volts for the sync separator.

14·7 Sync and blanking bars on the kinescope screen

Although the video signal is coupled to the sync separator to provide the desired sync pulses, the entire composite video signal with sync and blanking pulses is also coupled to the kinescope grid-cathode circuit. The amplitude variations of the sync and blanking pulses can be seen, therefore, as relative intensities on the kinescope screen. Figure 14·19 shows the details of the vertical sync and blanking bars on the kinescope screen. This can be seen by varying the vertical hold control to allow the picture to roll slowly, out of sync, so that the vertical blanking bar is in the middle of the picture instead of at the top and bottom. Brightness is turned up higher than normal to make the blanking level gray instead of black. The sync amplitude, which is 25 per cent above the blanking level, then becomes black. The black bar within the vertical blanking bar, often described as the "hammerhead" pattern, represents the equalizing and vertical sync pulses occurring during vertical blanking time.

The appearance of the horizontal sync within the horizontal blanking bar on the kinescope screen is shown in Fig. 14·20. This can be seen by adjusting the phasing of the horizontal deflection oscillator to put the

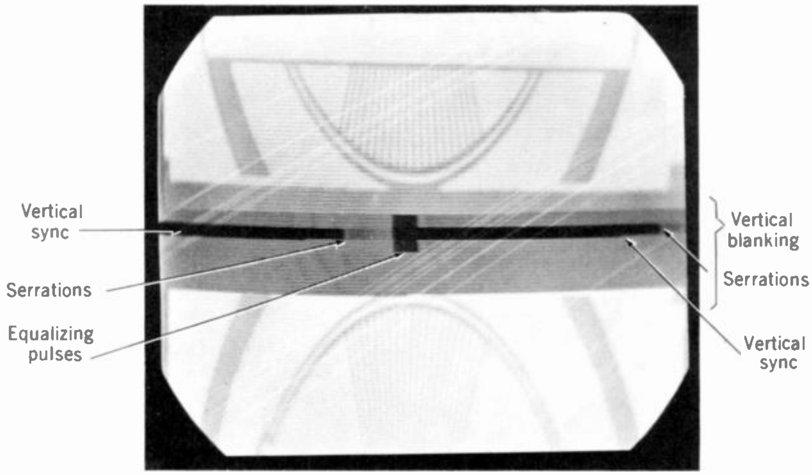


Fig. 14·19 Hammerhead pattern on kinescope screen. Vertical sync and equalizing pulses are within the vertical blanking bar. (RCA.)

horizontal blanking bar in the picture, with the brightness higher than normal. Normally the sync and blanking bars are at the edges of the picture behind the mask of the screen and are not visible.

The bars can be examined, to check the sync voltage at the kinescope grid. The sync should be blacker than blanking and the darkest parts of the picture. Normal bars on the kinescope screen show normal sync voltage in the composite video signal input to the sync separator. If the sync is not blacker than blanking, this indicates compression of the sync pulses in the video or i-f section.

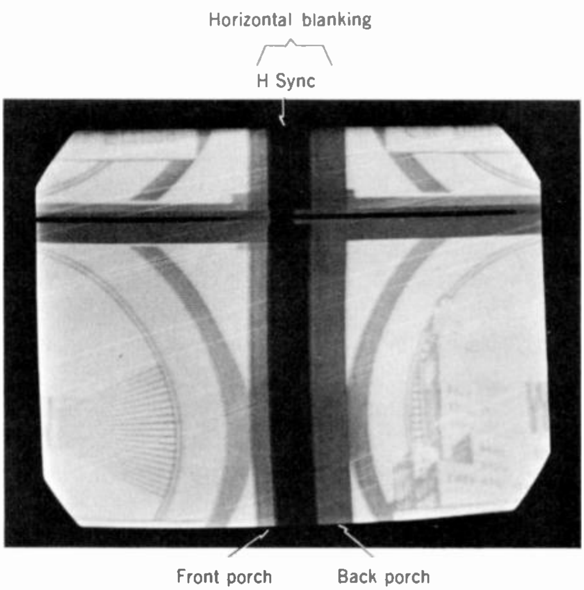


Fig. 14·20 Horizontal sync and blanking bars on kinescope screen, with hammerhead pattern. The intersection is called a pulse-cross display. (RCA.)

14·8 Sync troubles

Vertical rolling and horizontal tearing or bending in the picture mean faulty synchronization of the scanning raster with respect to the reproduced picture information. It should be noted, though, that the separated sync is part of the received picture signal and therefore the receiver must have enough signal to provide good synchronization. In most receivers a weak picture with hardly enough contrast to be visible cannot be synchronized. Also, interfering noise pulses can easily interrupt the synchronization with a weak signal. When the picture has good quality and contrast, however, poor synchronization indicates sync trouble. All channels are affected by a trouble in the sync circuits.

A trouble in the sync circuits where only vertical sync is used, such as the vertical integrating circuit, can cause loss of vertical synchronization while horizontal sync is normal. Or trouble in a horizontal sync circuit can cause loss of just the horizontal synchronization. The AFC circuit for the horizontal deflection oscillator provides horizontal synchronization only. When there is trouble in a stage where both horizontal and vertical sync are present, as in the sync separator, the picture rolls vertically and tears horizontally at the same time.

It is important to remember that the sync pulses are part of the r-f and i-f picture signal and the video signal for the sync separator. Distortion of the signal by limiting or clipping in the signal circuits can reduce the amplitude of the sync. Poor response for the low video frequencies can distort the vertical sync. Checking the sync and blanking on the kinescope screen will show whether the sync is normal in the composite video signal for the sync separator. In cases of distorted sync in the signal circuits, the picture contrast will be distorted, usually, in addition to the faulty synchronization.

No vertical hold. This is illustrated in Fig. 14·1. Vertical rolling of the picture means no vertical synchronization. The trouble may be either no vertical sync to lock in the vertical deflection oscillator, or the oscillator is too far off the correct frequency of 60 cps. In order to hold an oscillator synchronized, not only is sync voltage necessary but the oscillator frequency must be close enough to the synchronizing frequency to enable the sync voltage to lock in the oscillator. To check whether the trouble is no vertical sync or incorrect oscillator frequency, vary the vertical hold control to see if one complete picture can be stopped, in frame, for just an instant. If varying the vertical hold control cannot stop one complete picture, the trouble is incorrect frequency of the vertical oscillator. When the hold control stops the picture but it slips vertically out of hold, the trouble is no vertical sync input to the vertical deflection oscillator. Check the sync signal input and output of the vertical integrator.

No horizontal hold. The picture tears apart in diagonal segments, as shown in Fig. 14·2. Again, the trouble can be in either the oscillator or the sync. Vary the horizontal oscillator frequency control to see if a whole picture can be produced. If not, the trouble is incorrect frequency of the horizontal oscillator. When varying the frequency control can produce a whole picture but it does not stay still horizontally, the trouble is either no horizontal sync input to the AFC circuit for the horizontal deflection oscillator, or the AFC circuit is not synchronizing the oscillator.

Poor interlace. Inaccurate timing of the vertical scanning in even and odd fields causes poor interlacing of the scanning lines, resulting in partial pairing or complete pairing, which reduces the detail in the picture. Stray pickup of pulses generated by the receiver's horizontal deflection circuits and coupled into the vertical sync and deflection circuits can cause interlace troubles. The pulses generated locally by the horizontal deflection circuits for scanning are high in amplitude and do not have exact half-line difference in timing for even and odd fields. They can override the vertical sync, therefore, and produce inaccurate timing of the vertical scanning, with poor interlace.

Horizontal pulling in the picture. Weak horizontal sync allows the picture to bend or pull horizontally, as successive scanning lines are slightly displaced with respect to the picture information, but not enough to make the picture tear apart. Horizontal pulling often appears only at the top of the picture, as illustrated in Fig. 14·21. The edges of the raster are straight, showing that the weak horizontal sync causes bend in the picture but not in the raster. The bend at the top of the picture indicates weak horizontal sync just after the vertical sync pulse, since the horizontal scanning at the top immediately follows the vertical flyback. Weak horizontal sync all the time generally makes the picture tear apart, but when the horizontal sync is weak for only part of the picture it pulls or bends. The bend in the picture is caused by a small continuous change in the frequency of the horizontal oscillator. The scanning amplitude can remain the same to produce constant width with straight edges on the raster.

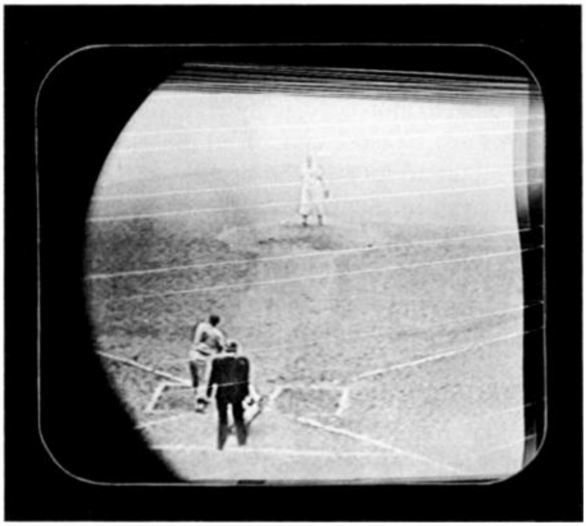


Fig. 14·21 Bend only at top of picture, with straight edge at side of raster. (RCA.)

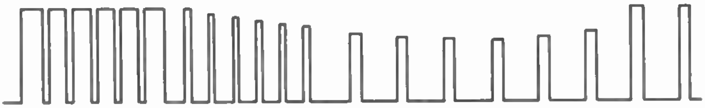


Fig. 14·22 Weak horizontal sync following the vertical pulse. Caused by short time constant in grid of sync separator.

Figure 14·22 illustrates how the horizontal sync voltage can be reduced, starting with the vertical sync pulse, to cause weak horizontal sync for the top of the picture. This can happen because the average value of the sync voltage increases during the vertical pulse. Therefore, an *RC* bias circuit with too short a time constant for the vertical sync will temporarily charge to a value too high for the average value of the total sync. As an example, if the vertical sync pulse makes the grid-leak bias too negative in the sync separator stage, the amplification will be reduced for the horizontal sync. Then the horizontal sync is weak following the vertical pulse, resulting in bend at the top of the picture.

Hum in the sync. Excessive hum in the horizontal sync produces bend in the picture as shown in Fig. 14·23. The hum in the sync bends the picture but the edge of the raster is straight. This shows that the hum is in the horizontal sync but not in the raster circuits. With excessive hum in the vertical sync, the picture tends to lock in halfway out of phase, with the vertical blanking bar across the middle of the picture, staying still in the position shown in Fig. 14·1*a*. Especially with weak signal, the oscillator can easily be locked in by 60-cycle hum voltage instead of vertical sync.

Fig. 14·23 60-cycle hum bend in picture, but not in the raster, without hum bars. Caused by heater-cathode leakage in horizontal AFC tube. (RCA.)

SUMMARY

1. The synchronizing pulses time the receiver scanning to make the reproduced picture hold still on the raster. Without vertical hold the picture rolls up or down. Without horizontal hold, the picture is torn apart with diagonal black bars. The bars are parts of horizontal blanking, which should be at the edges of the picture when it is in horizontal sync.
2. In the sync circuits, the sync separator clips the sync voltage from the composite video signal. Then the integrator filters out all but the vertical sync voltage, which triggers the vertical deflection oscillator at 60 cps. Also, the horizontal sync is coupled to the AFC circuit to control the frequency of the horizontal deflection oscillator.
3. A grid-leak bias sync separator circuit has composite video signal input with positive sync polarity. Then the sync voltage draws grid current to clamp the tip of sync close to zero grid voltage. With a large signal voltage compared with the grid-cutoff voltage, only the sync pulses allow plate current. Then the output voltage in the plate circuit is separated sync.
4. The *RC* integrator for vertical sync has a time constant very long for the horizontal pulse width but not for the vertical pulses.
5. Noise in the sync can cause loss of synchronization, especially with weak signal.
6. A noise-inverter stage cancels noise pulses in the composite video signal input to the sync separator (see Fig. 14·16).
7. A common arrangement for sync separation uses a double triode, with one section a grid-leak bias separator and the other triode a sync clipper (see Fig. 14·14).
8. Another common circuit is the gated sync separator. Either a pentagrid tube can be used as in Fig. 14·17, or a twin pentode combining the sync separator and AGC stages, as in Fig. 14·18.
9. When the picture rolls vertically but it can be stopped temporarily with the hold control, this means no vertical sync. Check the input and output of the vertical integrator.
10. If varying the horizontal oscillator frequency control can produce a picture but it drifts into diagonal bars, this means no horizontal hold. Check the horizontal AFC circuit.
11. When the picture has normal black and white amplitudes but it rolls vertically and tears horizontally on all channels, the trouble is no sync. Check the sync separator and clipper amplifier stages.
12. Hum in the horizontal sync can cause sine-wave bend from top to bottom of the picture, as in Fig. 14·23. Hum in the vertical sync can make the picture lock in with the blanking bar in the middle as in Fig. 14·1*a*.

SELF-EXAMINATION *(Answers at back of book.)*

Answer True or False.

1. Without vertical sync, the bar produced by vertical blanking rolls up or down the screen.
2. The diagonal black bars resulting from no horizontal hold are produced by horizontal blanking.
3. The integrated vertical sync voltage triggers the vertical deflection oscillator.
4. The horizontal deflection oscillator frequency is controlled by an AFC circuit.
5. A grid-leak bias sync separator requires composite video signal with positive sync polarity at the grid.
6. Grid-leak bias in a sync separator or sync clipper stage is proportional to grid signal voltage.
7. A time constant of 1/60 sec is typical for the vertical integrator.
8. Sharp grid cutoff and low plate voltage are two features of a sync separator stage.
9. When a noise pulse triggers the vertical oscillator, the picture can roll out of sync.
10. The 6BU8 twin pentode is often used to combine the gated sync separator and AGC stages.
11. Video signal voltage at the kinescope grid is about 3 volts peak to peak.
12. The hammerhead pattern in Fig. 14·19 is a reproduction of the serrated vertical sync and equalizing pulses.

13. If the sync separator stage does not function, the picture will roll vertically and have diagonal bars at the same time.
14. The top of the picture in the transmitted signal is always immediately after vertical blanking.
15. The left edge of a line of picture information in the transmitted signal is always immediately after horizontal blanking.
16. Grid-leak bias in a sync separator clamps the tip of sync close to zero grid voltage.
17. In Fig. 14·18 if the cathode voltage rises to 165 volts, the plate voltage on the sync separator will be +5 volts with respect to cathode.
18. Weak picture signal can cause poor vertical and horizontal synchronization.
19. The 60-cycle bend in Fig. 14·23 shows heater-cathode leakage in the video amplifier.
20. The TR16 transistor stage in Fig. 14·15 is a PNP common-emitter circuit.

ESSAY QUESTIONS

1. Give the specific function of the horizontal sync pulses, vertical pulses, and the equalizing pulses.
2. Describe briefly the sequence of separating the sync from composite video signal to time the frequency of the vertical and horizontal deflection oscillators.
3. Draw the schematic diagram of a pentode sync separator. Why is grid-leak bias used? Label typical values of components. Show input and output waveshapes with polarities.
4. Describe briefly the operation of a noise-inverter stage, as in Fig. 14·16.
5. With the oscilloscope connected to the sync separator output, how would you set the internal sweep control to see two vertical sync pulses, or two horizontal sync pulses?
6. Give the functions of the three sync stages in Fig. 14·14. Show input and output waveshapes, with sync polarity, for the three stages.
7. Describe briefly the functions for the two transistor stages in Fig. 14·15.
8. Give two reasons for vertical rolling.
9. Give two reasons for diagonal bars.
10. Give two causes of horizontal pulling or bend in the picture.
11. In Fig. 14·14, give the function of C_{315}, R_{425}, R_{422}, C_{410}, R_{417}, R_{418}, R_{419}, R_{420}, C_{403}, C_{412}, C_{413}, and C_{414}.
12. In Fig. 14·15, give the function of $C_{402}R_{401}$, $C_{401}R_{402}$, R_{403}, R_{404}, R_{405}, R_{406}, R_{408}, C_{501}, and C_{502}.
13. What is the function of the $R_3R_4R_5$ divider in Fig. 14·17?
14. For the gated sync separator in Fig. 14·17, what input signals are applied to the two control grids? Where is sync output taken from? Give the function of R_1C_1.
15. Give two names for the R_4 adjustment in Fig. 14·17. Explain briefly how to set this control.
16. Find the load resistor for output signal in each amplifier stage in Figs. 14·4, 14·14, 14·15, and 14·17.

PROBLEMS *(Answers to odd-numbered problems at back of book.)*

1. A grid-leak bias circuit has 200 $\mu\mu$f C_c and 200,000-ohm R_g. (*a*) Calculate the time constant. (*b*) After the capacitor has been charged how long will it take for C_c to discharge to one-half its initial voltage? (*c*) To 37 per cent of its initial voltage? (*d*) To zero voltage?
2. Calculate the time constant for each of the following *RC* circuits: (*a*) R_1C_1 in Fig. 14·4; (*b*) R_1C_1 in Fig. 14·6, with R_1 7,500 ohms and C_1 0.01 μf; (*c*) R_2C_2 in Fig. 14·6, with R_1 500 ohms and C_1 100 $\mu\mu$f; (*d*) $C_{401}R_{402}$ in Fig. 14·15 and $R_{408}C_{501}$ in Fig. 14·15. Which of these time constants are short compared with the vertical sync pulse width?
3. For the integrator in Fig. 14·14, calculate the time constant for each of the three sections.
4. Draw the circuit of an *RC* integrator with 6,350-ohm *R* and 63.5-μsec time constant. Show total sync input with one serrated vertical pulse, equalizing pulses, and two horizontal pulses before and after vertical sync. Draw to scale the integrated sync output voltage.

Chapter 15 Deflection oscillators

The scanning of the electron beam in the picture tube is made possible by the deflection oscillator stages in the receiver. As illustrated in Fig. 15 · 1, the deflection voltage generated by the vertical oscillator is coupled to the vertical deflection amplifier, which supplies the amount of current needed for the vertical scanning coils in the yoke mounted on the picture tube. Similarly, the horizontal oscillator produces the deflection voltage required for horizontal scanning. The deflection oscillator stage is often called a *deflection generator, sweep oscillator, sawtooth oscillator,* or *sawtooth generator.*

15 · 1 The sawtooth deflection waveform

The sawtooth waveform is required for scanning because it has a linear rise in amplitude to deflect the electron beam at uniform speed for the linear trace, with a sharp drop in amplitude for the fast retrace (see Fig. 15 · 2). Note that the sawtooth scanning current is an a-c wave. The electron beam is centered by positioning controls, and the alternating sawtooth current in the deflection coils deflects the beam away from center. Zero amplitude on the sawtooth average axis is the time when the beam is undeflected, at the center. The electron beam is at the extreme positions at the left and right sides, or top and bottom, of the raster when the sawtooth deflection wave has its peak negative and positive amplitudes.

Figure 15 · 2 shows a sawtooth waveform of current for magnetic deflection. However, the deflection oscillator generates sawtooth voltage. The sequence of waveforms is as follows. The deflection oscillator generates a fluctuating d-c voltage. This is capacitively coupled for a-c sawtooth voltage drive at the input to the deflection amplifier. The grid-voltage waveform produces sawtooth plate current. Finally, the output stage is transformer-coupled to the deflection coils to supply the required amount of a-c sawtooth current in the yoke. The peak-to-peak amplitudes of the

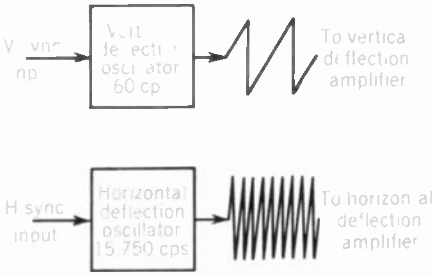


Fig. 15·1 The horizontal and vertical deflection oscillators generate sawtooth voltage for the scanning circuits.

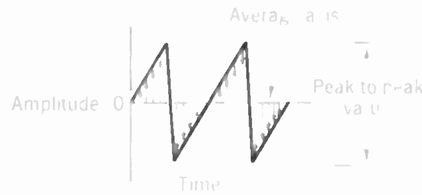


Fig. 15·2 The sawtooth deflection waveform.

a-c sawtooth current in the horizontal and vertical deflection coils determine the width and height of the raster.

Notice that the average axis is through the center of the sawtooth wave. With a fast retrace, the waveform is essentially a right triangle. Then the axis through the center divides the sawtooth waveform into equal areas above and below.

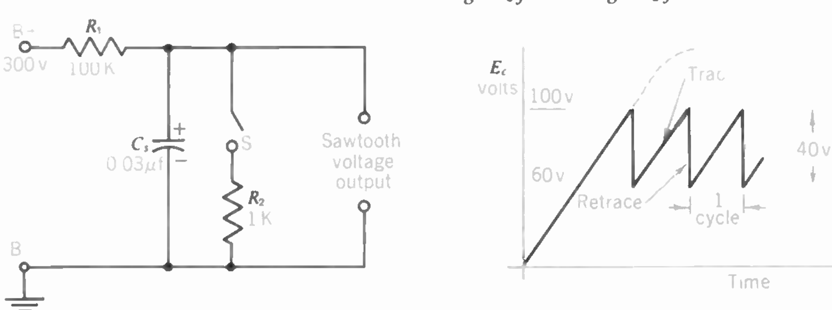
15·2 *Producing sawtooth voltage*

Usually, sawtooth voltage is obtained as the voltage output across a capacitor that is alternately charged slowly and discharged fast (see Fig. 15·3). There are two requirements:

1. Charge the capacitor through a high resistance in series with C_s for a long time constant. The charging produces a linear rise of voltage across C_s for the trace part of the sawtooth wave.
2. Discharge the same capacitor through a much smaller resistance for a relatively fast time constant. The sharp drop in capacitor voltage as it discharges is the retrace part of the sawtooth wave.

This sawtooth waveform is also called a *ramp voltage, sweep voltage,* or *time base.*

Fig. 15·3 Fundamental method of generating sawtooth voltage. C_s charges slowly through high resistance of R_1 for linear rise. Then switch is closed to discharge C_s fast through R_2 for retrace.



In Fig. 15·3, when d-c voltage is applied the capacitor charges toward the B+ voltage applied with an exponential charge curve, at a rate determined by the RC time constant.[1] The path for charging current is from B− to accumulate negative charge at the grounded side of C_s. Electrons are then repelled from the opposite plate to produce current through R_1, returning to the B+ terminal of the voltage source. As charge builds up in the capacitor, the voltage across C_s increases. The rate of increase in voltage depends on the RC time constant, which equals 0.003 sec for charge in this example.

Let us assume that, when the voltage across C_s builds up to 100 volts, we close the switch. Then C_s discharges through the low-resistance path in parallel. The path for discharge current is from the negative side of C_s through R_2 and the closed switch back to the positive side of C_s. Practically no current flows through the R_1 path because of its high resistance compared with R_2. Since electrons lost from the negative side are added to the positive side, the charge in the capacitor is neutralized. Then the voltage across C_s decreases toward zero. The R_2C_s time constant of 0.00003 sec for discharge is only one-hundredth of the time constant for charge. Therefore, the decrease in capacitor voltage, with a short time constant on discharge, is much faster than the voltage rise on charge.

If we open the discharge switch when the voltage across C_s has discharged down to 60 volts, the capacitor will start charging again. From now on, the capacitor starts charging from a 60-volt level. However, we can still discharge the capacitor by closing the switch every time E_c reaches 100 volts. Also, we open the switch to allow recharging when E_c is down to 60 volts. The result is the sawtooth waveform shown for the output voltage across C_s. One cycle of the sawtooth wave includes a linear rise for trace and the sharp drop in amplitude for the retrace. The frequency of the sawtooth wave here depends on the rate of switching. Note that the peak-to-peak amplitude of this sawtooth voltage waveform is 40 volts, equal to the difference between the maximum at 100 volts and the minimum at 60 volts.

Actually, the switch S in Fig. 15·3 represents a gas tube, vacuum tube, or transistor that can be switched on or off at almost any speed. When the tube or transistor is not conducting, this condition is the off position. Maximum conduction is the on position. Connected in parallel with C_s, the tube or transistor conducting maximum current provides a low-resistance discharge path to allow a fast decay of capacitor voltage for the retrace. In this application, we can call the tube a *discharge tube.* C_s is often called the *sawtooth capacitor* or *saw maker.*

The voltage rise on the sawtooth wave is kept linear by using only a small part of the exponential RC charge curve. As shown in Appendix E, the first 40 per cent of the curve is linear within 1 per cent. To limit the voltage rise to this linear part of the curve, the charging time should be

[1] See Appendix E for details of RC time constant.

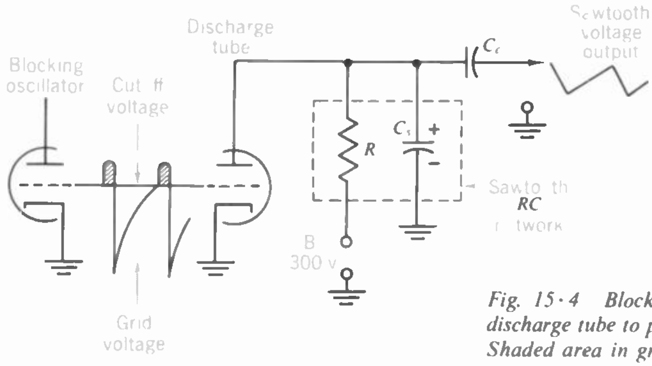


Fig. 15·4 Blocking oscillator grid voltage driving discharge tube to produce sawtooth voltage across C_s. Shaded area in grid voltage indicates time of plate-current conduction.

no more than one-half the RC time constant. Furthermore, a higher B+ voltage improves linearity, as a given amplitude of E_c is a smaller percentage of the applied voltage. Then a smaller part of the RC charge curve is used, where it is most linear.

15·3 Blocking oscillator and discharge tube

Figure 15·4 shows how a vacuum tube can be used as a discharge tube in parallel with the sawtooth capacitor C_s. While its grid voltage is more negative than cutoff, the discharge tube cannot conduct plate current. Then it is an open circuit. While the discharge tube is cut off, therefore, the sawtooth capacitor C_s in the plate circuit charges toward the B+ voltage, through the series resistance R, to produce the linear rise on the sawtooth voltage wave. When the grid voltage drives the discharge tube into conduction, its plate-to-cathode circuit becomes a low resistance equal to several hundred ohms. Then C_s discharges quickly from cathode to plate through the discharge tube, producing the rapid fall in voltage for the flyback on the sawtooth voltage wave. Therefore, by applying narrow positive pulses to the grid of the discharge tube and keeping it cut off between pulses, a sawtooth wave of voltage is produced in the output.

Figure 15·5 illustrates how the linear rise of voltage in the output cor-

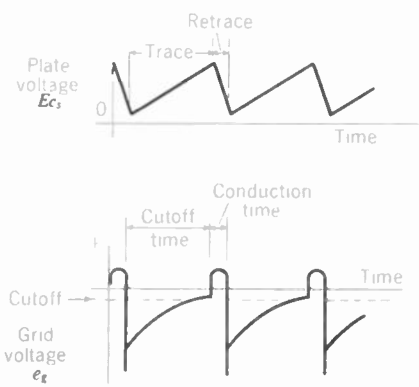


Fig. 15·5 How sawtooth voltage across C_s (above) corresponds to grid drive (below).

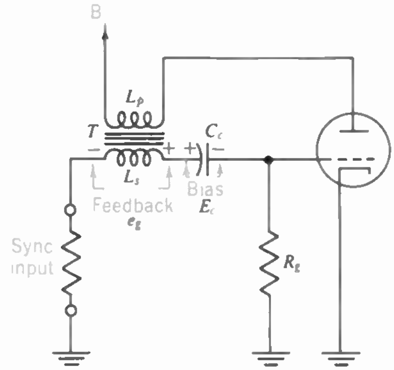


Fig. 15·6 Blocking oscillator circuit.

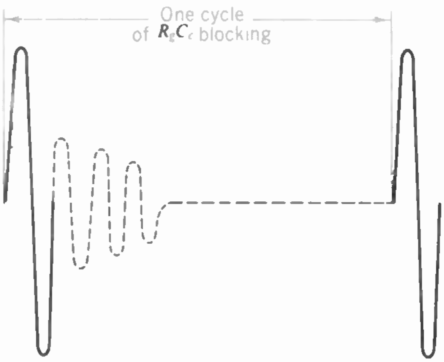


Fig. 15·7 Ringing of blocking oscillator transformer at resonant frequency of its inductance and stray capacitance. Resistance of transformer prevents dotted cycles from having enough amplitude to start plate current.

responds to the time when the grid is more negative than cutoff, while the flyback time coincides with the positive grid pulse. The result is a sawtooth wave of voltage output from the plate of the discharge tube with the same frequency as the grid-voltage pulses from the blocking oscillator.

Blocking oscillator circuit. Figure 15·6 is essentially a transformer-coupled oscillator with grid-leak bias. The transformer provides grid feedback voltage with the polarity required to reinforce grid signal and start the oscillations. When the oscillator feedback drives the grid positive, grid current flows to develop grid-leak bias. This regenerative circuit could oscillate with continuous sine-wave output at the natural resonant frequency of the transformer, depending on its inductance and stray capacitance. However, several factors enable the oscillator to cut itself off with high negative grid-leak bias. A large amount of feedback is used. Also, the R_gC_c time constant is made long enough to allow the grid-leak bias to keep the tube cut off for a relatively long time. Finally, the transformer has high internal resistance for low Q so that after the first cycle the sine-wave oscillations do not have enough amplitude to overcome the negative bias.

The tube remains cut off until C_c can discharge through R_g to the point where the grid-leak bias voltage is less than cutoff. Then plate current can flow again to provide feedback signal for the grid and the cycle of operation repeats itself at the blocking rate. Therefore, the circuit operates as an intermittent or blocking oscillator. The tube conducts a large pulse of plate current for a short time and is cut off for a long time between pulses.

Pulse repetition frequency. As shown in Fig. 15·7, one sine-wave cycle of high amplitude is produced at the blocking rate. The succeeding sine waves, in dotted lines, do not have enough amplitude to overcome the negative grid-leak bias blocking the oscillator. The number of times per second that the oscillator produces the pulse and then blocks itself is the pulse repetition rate, or frequency. This blocking frequency is much lower than the frequency of the sine waves, which is the ringing frequency or resonant frequency of the transformer.

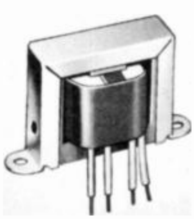
For a deflection oscillator, the pulse repetition rate is what we consider to be the oscillator frequency. This is the rate at which the tube oscillates between conduction and cutoff. The blocking frequency is determined mainly by the R_gC_c time constant. In a horizontal deflection oscillator, its R_gC_c time constant allows the blocking oscillator to operate at a pulse repetition frequency of 15,750 cps; a vertical blocking oscillator operates at 60 cps.

The total grid voltage of the blocking oscillator, with the pulses and grid-leak bias, is shown at the bottom of Fig. 15·9. Note that this grid voltage is exactly the waveform needed to operate the discharge tube. The narrow positive pulses drive the discharge tube into conduction. Then the sawtooth capacitor voltage in the plate of the discharge tube can drop sharply for a fast retrace. The negative grid voltage beyond cutoff keeps the plate current of the discharge tube cut off for a relatively long time between pulses. During this time the sawtooth capacitor voltage rises toward B+ voltage for the linear rise on the sawtooth wave.

The frequency of the sawtooth voltage output from the plate circuit of the discharge tube is the same as the blocking oscillator frequency. Also, the retrace time on the sawtooth wave corresponds to the width of the positive grid-voltage pulse from the blocking oscillator. This follows from the fact that the sawtooth capacitor can discharge only during the time the blocking oscillator pulse makes the discharge tube conduct. The width of the positive grid pulse is approximately one-half cycle of a sine wave at the natural resonant frequency of the transformer.

Blocking oscillator transformers. Figure 15·8 shows a transformer for the vertical deflection oscillator, operating at 60 cps. Such a transformer has a ratio 4:1 for secondary to primary turns, with a primary inductance approximately 1 henry. For a horizontal deflection oscillator, the trans-

Fig. 15·8 Vertical blocking oscillator transformer. Height is 2 in. (Stancor Electronics Corporation.)



former is smaller with less inductance. A typical unit has a ratio 2:1 for secondary to primary turns, with a primary inductance approximately 16 mh. In either case, the leads are color-coded red for B+, blue to plate, green to grid, and yellow for the grid return.

15·4 Analysis of blocking oscillator circuit

The waveform of instantaneous grid voltage e_c can be considered in two parts. One is the a-c feedback voltage e_g for the grid, induced across the transformer secondary by a change in primary current. The other is d-c bias voltage E_c developed when the feedback signal drives the grid positive to produce grid current. E_c can increase fast when grid current charges the coupling capacitor to produce the grid-leak bias. The bias voltage must decrease slowly, however, as C_c discharges through the high resistance of R_g when the feedback signal voltage decreases. At any instant, the net grid voltage e_c is equal to the algebraic sum of the bias E_c and the signal drive e_g. The grid signal voltage caused by feedback can drop to zero instantaneously when the feedback ceases, but the bias cannot. This is why the oscillator can cut itself off with the grid-leak bias produced by its own feedback.

It is worth reviewing the fundamentals of induced voltage with a transformer to see how the feedback voltage can increase, decrease, or reverse polarity. First, the primary current must change to induce voltage in the secondary. The amount of induced voltage in the secondary winding increases with a sharper rate of change in primary current. A slower rate of current change means less induced voltage. When the current stops changing, there is zero induced voltage for the steady primary current. Furthermore, we must consider the polarity of induced voltage. In the blocking oscillator transformer, plate current in the primary can flow in only the one

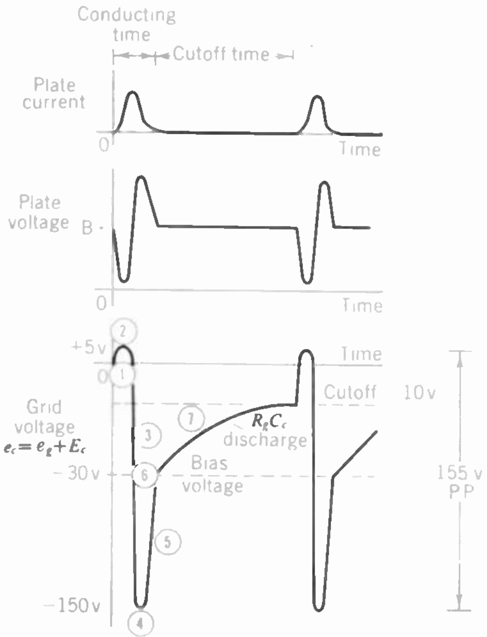


Fig. 15·9 Plate and grid waveforms in blocking oscillator circuit. See text for explanation of numbered steps in grid-voltage waveform. PP indicates peak-to-peak amplitude.

direction, from plate to the B+ voltage source. However, the current can either increase or decrease. Whatever polarity the induced voltage has for an increase of current, the polarity reverses for a decrease in current. The polarity reversal results from the fact that increasing current has an expanding magnetic field cutting across the winding, but the field collapses into the winding with decreasing current. The polarity of induced voltage reverses across both the primary and secondary when the current change reverses from increasing to decreasing values, or vice versa.

The cycle of operations can be followed from the time power is applied to produce plate current. Note the plate and grid waveforms in Fig. 15·9. The numbers in the e_c waveform correspond to the following steps in the circuit analysis.

1. **Increasing plate current.** Plate current flows immediately because the grid-leak bias is zero at the start. Now the plate current through L_p is increasing from zero. The windings are poled to make the resultant induced voltage positive at the grid side of L_s. Therefore, increasing plate current induces feedback voltage that drives the grid positive. The positive feedback increases the plate current still more. As a result of the regeneration, the plate current increases very quickly to its maximum value.

During this time when the feedback is driving the grid positive, grid current flows to develop negative bias a little less than the peak positive drive. As an example, for +35 volts at the peak, the grid-leak bias may be about −30 volts. Then the net grid voltage is +5 volts. The positive grid voltage is maintained while e_g is positive and more than E_c. This part of the cycle occurs while the plate current is increasing to produce positive grid feedback voltage.

2. **Saturation plate current.** Because of saturation, the plate current cannot increase indefinitely. When the plate current increases at a slower rate, the amount of positive feedback decreases. However, the negative bias remains. With less positive grid drive, now the plate current starts to decrease. At this turning point, the grid feedback voltage changes in polarity from positive to negative. Plate current is still flowing, but it is decreasing instead of increasing.

3. **Negative grid feedback.** As the grid becomes less positive, the negative grid feedback decreases the plate current still further. The amplification of the tube then makes the plate current drop sharply toward zero. This fast drop in current produces a very large negative voltage at the grid.

4. **Maximum negative grid feedback.** The fastest drop in plate current produces the maximum negative voltage at the grid, which can be as much as −150 volts. The large peak voltage in the negative direction is the result of two factors. First, there is no grid current now. Without the load of grid current in the secondary, the primary current can drop faster than it rises. Also, the negative grid feedback is series-aiding with the negative grid-leak bias so that now e_g is added to E_c.

5. **Decreasing negative grid feedback.** Because inductance opposes a change in current, the plate current in L_p decreases at a slower rate as it

drops toward zero. The current decay in an inductance corresponds to a capacitor discharge curve, with the sharpest slope at the start. Therefore, the amount of negative feedback voltage decreases as the drop in plate current approaches zero. Plate current can flow even though the grid voltage is very negative, as the self-induced voltage across L_p makes the plate voltage rise to a positive value much higher than B+.

6. Plate current drops to zero. Now there is no feedback voltage at all. However, the grid-leak bias produced while the grid was positive still remains after the feedback stops. This bias voltage keeps the tube cut off.

7. C_c discharges through R_g. As C_c loses charge, its negative bias voltage decreases. If we take the example of -30 volts maximum grid-leak bias, and a grid-cutoff voltage of -10 volts, the tube is cut off for the time it takes C_c to discharge 20 volts, from -30 to -10 volts. Therefore, the cutoff time depends on the R_gC_c time constant.

When the grid-leak bias becomes less than the grid-cutoff voltage, plate current can flow again to produce the next pulse and the cycle is repeated. Therefore, the blocking oscillator continuously generates sharp positive grid pulses followed by relatively long periods of cutoff. The blocking rate is faster, for a higher frequency, with a shorter time constant for R_gC_c. The blocking oscillator frequency also depends on the tube characteristics, especially its grid-cutoff voltage, the plate voltage, and the transformer. However, the oscillator frequency is usually varied by adjusting either the R_gC_c time constant or the bias voltage in the grid circuit.

15·5 Deflection generators with blocking oscillator and discharge tube

Figure 15·10 shows a sawtooth voltage generator for horizontal deflection. However, the same circuit can be used with different values for a vertical deflection generator operating at 60 cps. For horizontal scanning, the 180-$\mu\mu$f grid capacitor C_1 and the total grid resistance of 250,000 ohms for R_1 and R_2 provide a time constant that enables the blocking oscillator to operate at 15,750 cps. The oscillator grid is connected

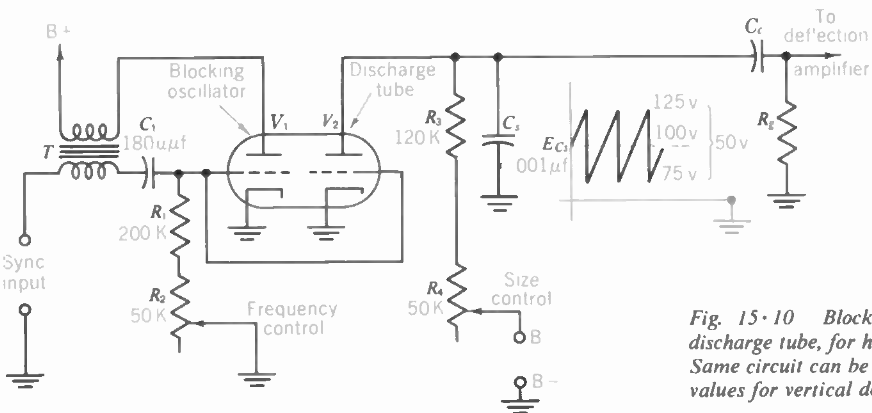


Fig. 15·10 Blocking oscillator and discharge tube, for horizontal scanning. Same circuit can be used with different values for vertical deflection oscillator.

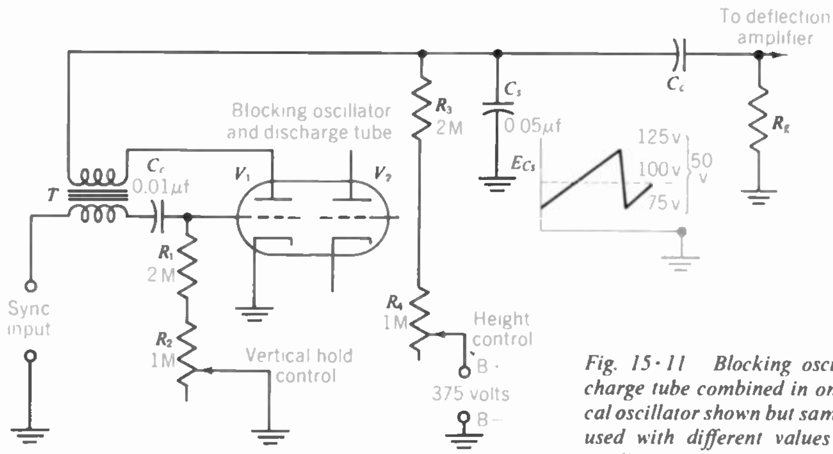


Fig. 15·11 Blocking oscillator and discharge tube combined in one triode. Vertical oscillator shown but same circuit can be used with different values for horizontal oscillator.

directly to the discharge tube grid. Therefore, both tubes have grid voltage consisting of narrow positive pulses followed by relatively long periods of cutoff.

C_s between plate and cathode of the discharge tube is the sawtooth capacitor. While the oscillator grid voltage keeps the discharge tube cut off, C_s charges toward the B+ voltage, through the combined series resistance of R_4 and R_3. This charging of C_s during cutoff time produces the linear rise on the sawtooth voltage output. When the oscillator conducts, the discharge tube also conducts, as both have a sharp positive pulse of grid voltage. Then the sawtooth capacitor discharges fast through the low resistance of the discharge tube. As a result, the voltage across C_s drops rapidly for the retrace part of the sawtooth wave. With continuous output from the blocking oscillator, it drives the discharge tube for sawtooth voltage output at the oscillator frequency.

Single-triode circuit. The grid and cathode voltages of the discharge tube are the same as in the blocking oscillator. Therefore, the blocking oscillator itself can function as the sawtooth generator by connecting the sawtooth capacitor in the plate circuit (see Fig. 15·11). It should be noted that the oscillator is alternately conducting and cut off, as required for the discharge tube.

In Fig. 15·11, the components are shown for a vertical deflection oscillator but the same circuit with different values can be used for horizontal scanning. The single triode is mainly a blocking oscillator. However, the sawtooth capacitor C_s is in the oscillator plate circuit, instead of using a separate discharge tube. While the tube is cut off by its blocking action, C_s charges through the series resistance of R_3 and R_4, toward the B+ voltage. When plate current flows during the oscillator pulse, C_s discharges through the tube. The discharge path for C_s is from cathode to plate and through the plate winding of the transformer. Note that the discharge current of C_s is in the same direction as normal plate current during oscillator conduction. The frequency of the sawtooth voltage output is the frequency of the blocking oscillator.

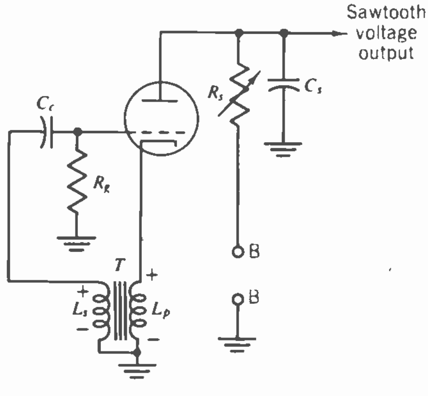
When the blocking oscillator and discharge tube circuit is used for the vertical or horizontal deflection generator, it is generally this single-triode arrangement. The advantage is one stage less without a separate discharge tube. However, changes in average plate voltage vary the oscillator frequency. Higher values of E_b lower the frequency, as more feedback produces more grid-leak bias. This effect occurs when the height control R_4 in the plate circuit is varied to adjust the height of the raster. The resulting change in E_b has the side effect of changing the oscillator frequency, which usually makes the picture roll. However, the frequency can be brought back to normal by readjusting the frequency control R_2.

Cathode feedback. In Fig. 15·12, the blocking oscillator transformer supplies feedback from cathode to grid. The plate current returns through L_p, from the ground side to cathode. Note the voltage polarities shown for increasing cathode current inducing positive grid feedback. The sawtooth capacitor C_s is in the plate circuit without the transformer. While the tube is cut off, C_s charges through R_s. Then C_s discharges through L_p and the tube when it conducts at the blocking oscillator frequency.

Fluctuating d-c waveforms. A fluctuating or pulsating d-c waveform has an average d-c level, serving as an axis for the a-c variations above and below the average value. Although the a-c variations provide the desired waveform in a deflection generator circuit, the average d-c axis should be noted. In the plate circuit, the average plate voltage is the center axis for the a-c sawtooth wave. In the grid circuit, the grid-leak bias is the average d-c level, with negative polarity. This axis is not in the center because of the unsymmetrical waveform of grid voltage. For either case, the average d-c voltage can be measured with a d-c voltmeter. The a-c waveform can be observed with a calibrated oscilloscope to measure peak-to-peak value.

Referring back to Fig. 15·10, as an example, the sawtooth voltage across C_s varies 25 volts above and below the 100-volt axis, which is the average plate voltage of V_2. The peak-to-peak amplitude of the sawtooth component is the difference between the peak at +125 volts and the peak at +75 volts, or 50 volts. This fluctuating d-c voltage is RC-coupled to the deflection amplifier. Although the average d-c level is blocked by C_c, the

Fig. 15·12 Blocking oscillator circuit with cathode feedback. Polarities shown for increasing cathode current.



sawtooth voltage is developed across R_g as an a-c signal to drive the next stage. In its a-c form, the sawtooth wave across R_g has 50 volts peak-to-peak amplitude, with the same variations as in the preceding plate circuit but centered on the zero axis instead of the average plate voltage. These same values also apply to the fluctuating sawtooth voltage output in Fig. 15·11.

As an example of fluctuating d-c grid voltage, note the values indicated for e_c in Fig. 15·9. Here the grid-leak bias of −30 volts is the average d-c axis. The instantaneous values of e_c vary between the peak of +5 volts and the peak at −150 volts. For this example, the peak-to-peak value equal to the difference between the two peaks is 155 volts. Notice that the 5 volts is added to 150 volts because they have opposite polarities. Also, this waveform is unsymmetrical, as the average value axis is not through the center.

15·6 Deflection oscillator controls

The frequency and amplitude of the oscillator output can be adjusted by variable resistances. Referring to Fig. 15·11, R_2 varies the time constant of the grid-leak bias circuit to adjust the oscillator frequency. R_4 in the plate circuit determines the time constant on charge for C_s to adjust the peak-to-peak amplitude of the sawtooth voltage output.

Size control. Decreasing the resistance of R_4 in Fig. 15·11 makes the time constant shorter for charging C_s. Then the voltage across C_s increases at a faster rate. However, the amount of time allowed for charging is equal to the cutoff period, as determined by the oscillator frequency. For any one frequency, therefore, C_s can charge to a higher voltage because of the faster charging rate with a shorter time constant. As a result, decreasing the resistance of the size control increases the amplitude of sawtooth voltage output to increase the size of the scanning raster. Increasing the resistance of the size control decreases the scanning amplitude. Usually, a fixed resistance is in series with the control to limit the range of variations for easier adjustment.

This method of size control is generally used in the vertical deflection generator to adjust the height of the raster. Then it is called the *height control.* The same method could be used for a horizontal generator but varying the grid drive in the horizontal output stage would vary the amount of flyback high voltage. For this reason, a width control is usually in the horizontal output circuit.

Effect of C_s time constant on amplitude and linearity. Figure 15·13 shows the effect on sawtooth voltage amplitude when the height control in Fig. 15·11 is varied. For zero resistance in R_4, the time constant of R_3C_s equals 0.1 sec. This is approximately six times longer than the vertical scanning period of 0.016 sec. Therefore, C_s can take on enough charge to raise its voltage by one-sixth the net charging voltage, since the charging time equals one-sixth of the time constant.

Let us assume 75 volts across C_s at the start of charge, after the first few cycles of charge and discharge. With B+ voltage of 375 volts, the net

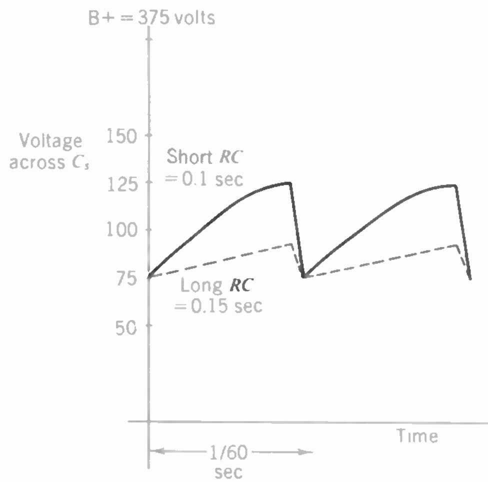


Fig. 15·13 Effect of varying time constant for sawtooth capacitor C_s in Fig. 15·11, with height control at minimum and maximum resistance.

charging voltage equals 375 − 75, or 300 volts. Therefore, C_s charges an additional 50 volts, equal to one-sixth of 300 volts. This voltage equals the peak-to-peak amplitude of sawtooth output, between 75 and 125 volts.

At the opposite extreme, with the height control at maximum resistance, the time constant for charge equals 0.15 sec. This is approximately ten times longer than 0.16 sec. Therefore, C_s takes on additional charge to raise its voltage by one-tenth of 300 volts. Then the peak-to-peak sawtooth voltage equals 30 volts, between 75 and 105 volts. Note the better linearity but smaller amplitude with the longer time constant.

Frequency control. To control the oscillator frequency, the R_gC_c time constant in the oscillator grid circuit is varied. This determines how fast the negative bias voltage can discharge down to cutoff. A shorter time constant with smaller values for R_g and C_c allows a faster discharge for a higher frequency. Making the R_gC_c time constant longer lowers the oscillator frequency.

The oscillator frequency control is adjusted to the point where the sync voltage can lock in the oscillator at the sync frequency to make the picture hold still. For this reason the frequency adjustment is generally called the *hold control*. In the vertical oscillator, the vertical hold control generally is the oscillator frequency adjustment, as in Fig. 15·11. However, since the horizontal oscillator usually has automatic frequency control, the horizontal hold adjustment may be in the AFC circuit.

Effect of R_gC_c time constant on oscillator frequency. We can calculate an example from the vertical deflection generator in Fig. 15·11. When R_2 is at its middle value the total grid resistance R_g equals 2.5 megohms. The time constant with 0.01 μf for C_c equals 0.025 sec.

Let us assume grid-leak bias of −40 volts as the initial capacitor voltage at the start of discharge. Also, consider that the grid-cutoff voltage of the tube is −20 volts. The problem now is to calculate the time for the grid voltage across R_g to decline from the initial negative bias of −40 volts, when the tube is cut off, to the grid-cutoff voltage of −20 volts when plate starts again for the next cycle.

This time can be calculated from the RC discharge formula in Appendix E, as follows:

$$t = 2.3\ RC \log \left(\frac{e_2}{e_1}\right) = 2.3\ RC \log \left(\frac{40}{20}\right)$$
$$= 2.3 \times 0.025 \times \log 2 = 2.3 \times 0.025 \times 0.3$$
$$t = 0.01725 \text{ sec}$$

Therefore, the oscillator is cut off for 0.01725 sec in each cycle, while the grid-leak bias voltage declines to the cutoff grid voltage.

We can add 300 μsec or 0.0003 sec for the short conduction time during vertical retrace. This vertical flyback time corresponds to a little less than five horizontal lines. The total period for one cycle of oscillator grid voltage is 0.01725 sec for trace, plus 0.0003 sec for retrace, which equals 0.01755 sec. Since this period of a complete cycle is slightly more than 1/60 sec or 0.0167 sec, the vertical oscillator frequency in this example is a little lower than 60 cps. Decreasing the resistance of R_2 will raise the frequency; increasing its resistance lowers the frequency. The range of frequency control for the vertical oscillator is usually about 40 to 90 cps.

Effect of oscillator frequency on sawtooth amplitude. It should be noted that the frequency is determined by the oscillator grid circuit, while the sawtooth capacitor in the plate circuit of the discharge tube determines amplitude and linearity of the sawtooth voltage output. However, changing the frequency will affect sawtooth amplitude and linearity. This idea is illustrated in Fig. 15 · 14. With any given time constant for C_s on charge, if the frequency is higher, less time is available for charging. Then C_s is charged to a lower voltage at the time of discharge. For the opposite case, a lower frequency allows C_s to charge to a higher voltage before discharge occurs, producing more sawtooth voltage output.

15 · 7 Synchronizing the blocking oscillator

A circuit like this is called a *soft oscillator* because its frequency is easily changed by variations in the electrode voltages. Its advantage, however, is that the oscillator can easily be synchronized to lock in at the sync

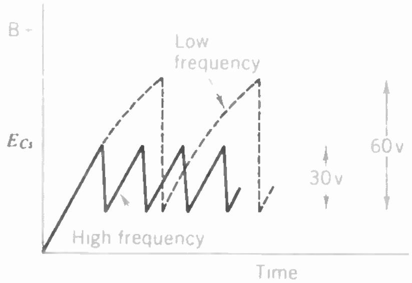


Fig. 15 · 14 Effect of blocking oscillator frequency on amplitude of sawtooth voltage output of C_s.

frequency. The frequency can be synchronized either by sync pulses that trigger the oscillator into conduction at the sync frequency or by d-c voltage to control the grid bias. The vertical deflection oscillator is usually locked in with triggered sync at 60 cps. However, the horizontal oscillator frequency is usually synchronized with d-c control voltage produced by the horizontal AFC circuit.

Triggered sync. The blocking oscillator can be synchronized by small positive pulses injected in the grid circuit to trigger the oscillator into conduction at the frequency of the sync pulses. The sync voltage is applied in series with the grid winding of the transformer. Then the positive sync voltage cancels part of the grid bias voltage produced by the oscillator. For the vertical oscillator, the sync pulse input is about 10 volts peak to peak from the integrator.

In Fig. 15·15, the positive sync pulses arrive at the times marked *S*, when the declining grid voltage is close to cutoff. Then a small sync voltage will be enough to drive the grid voltage momentarily above the cutoff voltage. As soon as plate current starts to flow, the oscillator goes through a complete cycle. With another positive sync pulse applied at a similar point of the following cycle, the oscillator again begins a new cycle at the time of the pulse. As a result, the pulses force the oscillator to operate at the sync frequency. The frequency of the oscillator without sync is called the *free-running frequency;* the synchronized oscillator frequency is the *forced frequency.*

The free-running frequency of the oscillator must be set lower than the sync frequency. Then the sync pulses will drive the grid voltage in the pos-

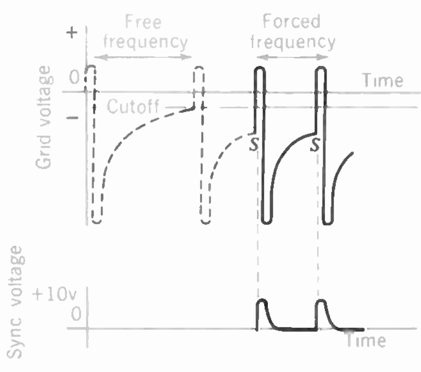


Fig. 15·15 Synchronizing the blocking oscillator with positive trigger pulses injected into grid circuit.

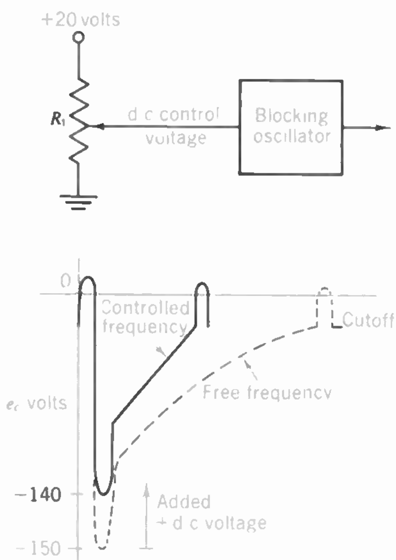


Fig. 15·16 Adding positive d-c control voltage to blocking oscillator grid voltage raises the frequency.

itive direction when the oscillator is ready for triggering. This is the time when the grid bias has declined practically down to cutoff by itself, and needs only a slight additional positive voltage to start the flow of plate current and the beginning of a cycle. A positive sync pulse that occurs in the middle of the oscillator cycle must have a much higher value to drive the grid voltage to cutoff. The peak negative swing of the grid-voltage wave may be more than 200 volts, but toward the end of the cycle a few volts of positive sync voltage can be enough to trigger the oscillator into conduction.

Sync voltage of negative polarity at the grid cannot trigger the blocking oscillator. Also, the oscillator cannot be triggered if the free frequency is slightly higher than the synchronizing frequency. Then the sync pulses will occur after the oscillator has started to conduct by itself and they will have no effect. Operating the oscillator at the same frequency as the synchronizing pulses does not provide good triggering because the oscillator frequency can drift above the sync frequency, resulting in no synchronization. For best synchronization, the free-running oscillator frequency is adjusted slightly lower than the forced frequency, so that the time between sync pulses is shorter than the time between pulses of the free-running oscillator. Then each synchronizing pulse occurs just before an oscillator pulse and forces the tube into conduction, thereby triggering every cycle to hold the oscillator locked in at the sync frequency.

The triggering action can be made less sensitive to noise pulses by returning the grid to a positive voltage, instead of chassis ground. The added positive voltage has two effects on the grid discharge. First, the negative bias declines to cutoff with a sharper slope because a smaller part of the R_gC_c discharge curve is used. With a sharper slope of declining grid voltage, an interfering noise pulse just before the sync pulse will need much more amplitude to trigger the oscillator. The second effect is that the added positive voltage makes the negative bias decline to cutoff in less time, raising the oscillator frequency. However, the frequency control can be adjusted to bring the oscillator to the frequency desired. It should be noted that the oscillator frequency can also be controlled by d-c voltage.

D-c control voltage. In Fig. 15 · 16 positive d-c control voltage is directly coupled to the blocking oscillator grid. R_1 varies the amount of d-c voltage used to control the oscillator frequency. In the waveshape below, the +10 volts added decreases the negative grid voltage. As a result, less time is necessary for the grid leak bias to decline to cutoff to start the next cycle. The blocking oscillator frequency is raised, therefore, by inserting positive d-c control voltage to make the grid voltage less negative.

15 · 8 Multivibrators

The multivibrator is a pulse-generator circuit like the blocking oscillator, but is even more commonly used for the sawtooth generator in television receivers. Two stages are needed for the multivibrator but there is no blocking oscillator transformer. The basic multivibrator circuit consists of two *RC*-coupled stages, with the output of one stage driving the other stage (see Fig. 15 · 17). Since each amplifier inverts its signal, the feedback voltage is in the same polarity as grid input signal and oscillations can take place. The oscillations are in the on-or-off conditions for each stage. When one tube conducts it cuts off the other. As soon as the cutoff tube starts to conduct again it cuts off the opposite tube. The rate at which the tubes are cut off and conducting is the oscillator frequency. One cycle includes cutoff for both tubes.

Multivibrator circuits can be classified according to the method of feedback. In a plate-coupled multivibrator, the plate of each stage drives the grid of the opposite stage. In a cathode-coupled multivibrator the feedback is between the two cathodes. For either case, the multivibrator can be free-running, meaning it operates with or without sync voltage input. If it operates only when sync input voltage drives the cutoff tube into conduction, the circuit is a driven multivibrator, instead of free-running. The many applications of multivibrators include sawtooth generators, square-wave generators, electronic switching, and frequency dividers.

15 · 9 Plate-coupled multivibrator

This basic circuit is illustrated in Fig. 15 · 18. When supply voltages are applied, plate current begins to flow in both tubes and the circuit immediately begins oscillating. The amount of plate-current flow cannot be identical for the two tubes even if both use the same supply voltage and have plate load resistors of the same size. No matter how small this

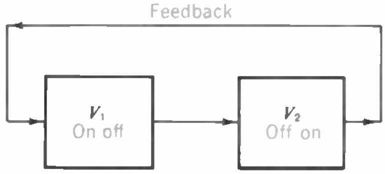
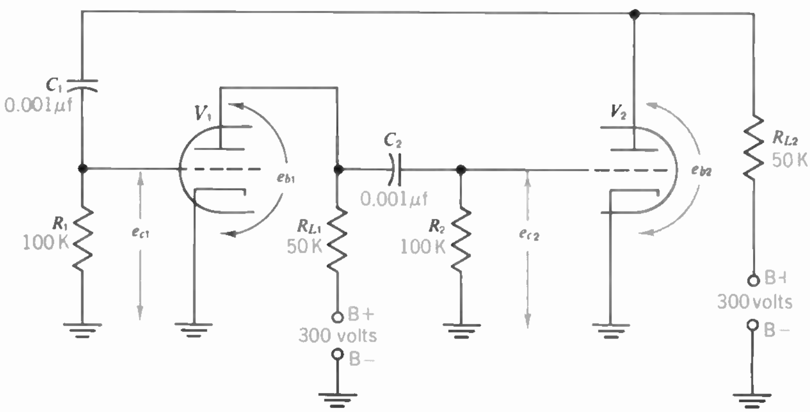


Fig. 15 · 17 Illustrating the switching operation between the two sections of a multivibrator.

Fig. 15 · 18 Basic plate-coupled symmetrical multivibrator circuit. Note that e_{c_1} and e_{c_2} refer to grid-cathode voltages.



difference in plate current may be, it is immediately amplified to produce the result of one tube conducting while the other is cut off. Assume that tube 1 conducts slightly more than tube 2 when plate voltage is applied, driving the grid of tube 2 slightly more negative. This negative signal is amplified and inverted to provide feedback that drives the grid of tube 1 more positive, which in turn allows the first tube to drive the grid of tube 2 still more negative. The amplification of the unbalance in the stage takes place almost instantaneously to drive the grid of tube 2 to cutoff immediately. The tube remains cut off for a period of time that depends upon the R_2C_2 grid time constant.

The grid voltage of tube 2 then declines toward zero. As soon as the grid voltage is reduced to less than cutoff, plate current begins to flow. Then conduction in tube 2 cuts off tube 1. Now the cutoff time depends on the R_1C_1 time constant. As a result, the slight initial unbalance sets up a regenerative switching action with first one tube conducting and then the other. Each tube conducts for a period equal to the time the other is cut off.

It can be helpful in following the multivibrator action to remember the following fundamentals:

1. When the grid is driven in the positive direction, the plate current increases but the plate voltage goes down. The increased voltage across the plate load resistor drops the plate voltage. When the grid is driven more negative plate current decreases but the plate voltage goes up. With the tube cut off, its plate voltage equals B+ as there is no IR drop across the plate load.
2. When the plate voltage goes up, the voltage applied to the RC coupling circuit into the next stage increases. Then the coupling capacitor charges to increase its voltage. The charging current produces a voltage drop across the grid resistor with the grid side positive. With the grid positive, grid current flows. As an example, when e_{b_1} increases in Fig. 15·18, C_2 charges from the B supply, through the grid-cathode circuit of tube 2, repelling electrons through R_{L_1}, back to the B+ terminal.
3. When the plate voltage goes down, the voltage applied to the coupling circuit decreases and the capacitor discharges. In Fig. 15·16, C_2 discharges through R_2 and the internal plate-cathode resistance of tube 1. Remember that tube 1 is conducting now, which is why e_{b_1} decreased to discharge C_2. The discharge current produces voltage across R_2 with the grid side negative.
4. The coupling capacitor cannot charge or discharge instantly but is limited to a rate determined by the RC time constant. When the applied voltage drops or rises, instantaneously the entire voltage change appears across the resistance. Then the maximum voltage across the resistor declines as the capacitor either charges or discharges. Actually, the resistor voltage corresponds to the charge or discharge current, not

the voltage across the capacitor. See Appendix E for more details on RC charge and discharge.

Analysis of multivibrator waveforms. Beginning at time A in Fig. 15·19, tube 2 has just been cut off because of conduction in tube 1. Assume that 4 ma of plate current flows through the 50,000-ohm plate load resistor of the first tube, producing a 200-volt drop in plate voltage. Since the plate-to-cathode voltage applied across the R_2C_2 coupling circuit is reduced abruptly from 300 to 100 volts, C_2 must discharge. Instantaneously, the entire 200-volt drop in applied voltage is developed across R_2, with the grid side negative, cutting off tube 2.

As C_2 discharges, the voltage across R_2 declines toward zero with the typical RC discharge curve shown. When the grid voltage is down to cutoff, tube 2 begins to conduct plate current. The resultant decrease in plate voltage in tube 2 drives the grid of the first tube more negative, reducing its plate current and increasing the plate voltage. This drives the grid of tube 2 more positive, further reducing the plate voltage and allowing tube 2 to drive tube 1 still more negative. This amplification of the unbalance in the stage reverses the action of the two tubes almost instantaneously, with tube 1 now cut off and tube 2 conducting, as shown at time B.

The plate-to-cathode voltage of the first tube, now cut off, rises immediately to B+, driving the grid of tube 2 positive. C_2 charges rapidly through the low resistance of the grid-to-cathode circuit of tube 2, and the grid voltage is reduced to zero very soon as C_2 becomes completely charged. The grid voltage for tube 2 remains at zero and zero-bias plate current flows in tube 2 as long as tube 1 remains cut off.

Meanwhile, the coupling capacitor for the first tube, C_1, is discharging through its grid resistor R_1 and the negative grid voltage of tube 1 declines toward zero. When cutoff voltage is reached at time C in the illustration,

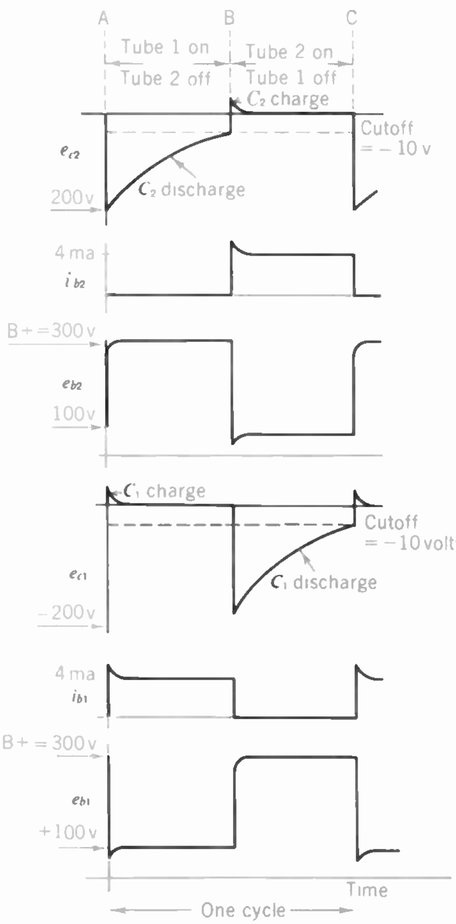


Fig. 15·19 Plate and grid waveforms for plate-coupled multivibrator in Fig. 15·18.

conduction begins again in the first tube, cutting off tube 2 again to repeat the cycle. The waveforms for both tubes are exactly the same but of opposite polarity, since one tube is conducting while the other is cut off. The period of conduction for either tube is equal to the cutoff time of the other tube. It is the change from cutoff to conduction that initiates the switching operation.

The output voltage from either plate is a symmetrical square wave of voltage as the plate voltage rises sharply to B+, remains at that value for a period equal to the cutoff time, and then drops sharply to some low value resulting from plate-current flow. The slight departure from square corners is caused by charging of the coupling capacitors. The output is symmetrical because both tubes are cut off the same amount of time.

Multivibrator frequency. The time from A to C in Fig. 15·19 is one complete cycle, including a complete flip-flop of operating conditions. The frequency may have an approximate range from 1 to 100,000 cps, depending primarily on the RC time constant of the grid coupling circuits. The period of one cycle is exactly equal to the sum of the cutoff periods of both tubes.

The cutoff period for each tube can be calculated as the time it takes the grid voltage to decline from its maximum negative peak to the value required to start plate current to flow. As a specific example, we can calculate the frequency of the symmetrical multivibrator in Fig. 15·18, with the waveshapes in Fig. 15·19. For either tube 1 or 2, the problem is to calculate the time for the grid voltage across R_g to decline from its negative peak of -200 volts to the cutoff value of -10 volts. The R_gC_c time constant for either tube is 0.0001 sec. Using the RC discharge formula:

$$t = 2.3\ RC \log\left(\frac{e_2}{e_1}\right) = 2.3\ RC \log\left(\frac{200}{10}\right)$$

$$= 2.3 \times 0.0001 \times \log 20 = 2.3 \times 0.0001 \times 1.3$$

$$t = 0.00029 = 0.0003 \text{ sec (approx.)}$$

Therefore, the cutoff period of a half cycle is 0.0003 sec. For both tubes, the cutoff time is twice 0.0003, or 0.0006 sec, in this symmetrical multivibrator. The frequency then is 1/0.0006 sec, which equals approximately 1,667 cps.

15·10 Cathode-coupled multivibrator

In Fig. 15·20 the coupling for feedback from tube 2 to tube 1 is produced by the common cathode resistor. Note that plate current for both tubes flows through R_k. However, V_1 also drives V_2 with the usual R_gC_c grid coupling circuit. Therefore, V_2 can be cut off by conduction in V_1 as C_2 discharges through R_2 because of the first tube's drop in plate voltage, just as in the plate-coupled multivibrator. However, V_1 is cut off by the cathode bias voltage produced across R_k when V_2 conducts maximum plate current.

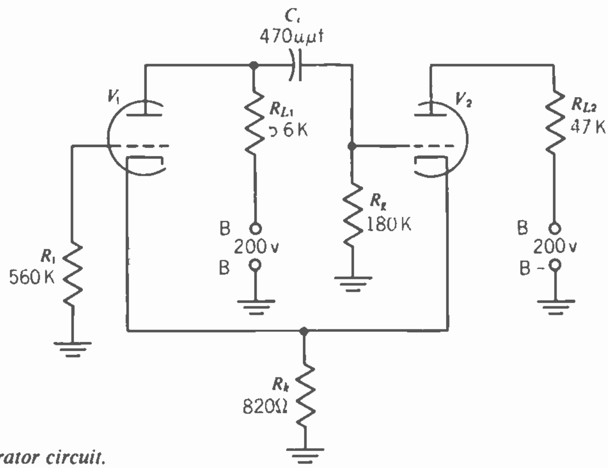


Fig. 15·20 Cathode-coupled multivibrator circuit.

Note that there is no feedback from the V_2 plate to the V_1 grid. Therefore, the grid-cathode feedback voltage for V_1 depends only on the voltage across R_k. The circuit oscillates as a free-running multivibrator, with first one tube cut off and then the other.

How V_2 is cut off. When plate voltage is applied, both tubes start to conduct. The flow of plate current in V_1 reduces its plate voltage, driving the grid of V_2 negative. The plate current in V_2 is reduced because of the negative grid signal, decreasing the voltage across R_k. This allows V_1 to conduct more plate current, driving the grid of V_2 more negative, and the unbalance is amplified to drive V_2 to cutoff almost instantaneously.

V_2 is held cut off during the time C_c discharges through R_g, R_k, and the plate-to-cathode resistance of V_1, which is now conducting. The negative grid voltage across R_2 declines exponentially with the normal capacitor discharge curve until the grid voltage for V_2 has been reduced to the grid cutoff voltage. Then V_2 starts to conduct.

How V_1 is cut off. With the plate current of V_2 now flowing through the common cathode resistor, the cathode bias for V_1 is increased, driving its grid negative. Plate current in V_1 is reduced because of the additional cathode bias, allowing its plate voltage to rise toward the B+ of 200 volts. The increase in plate voltage drives the grid of V_2 more positive as C_c charges from the B supply through the grid-to-cathode circuit of V_2 and R_{L_1}. With V_2 driven more positive, its plate current increases, and more bias is developed across R_k as the cathode voltage follows the applied grid voltage. The action is cumulative and results in V_1 being cut off almost instantaneously by the plate current of V_2. Now V_1 is cut off. This cutoff time depends on how long it takes C_c to charge. Until C_c charges, the grid of V_2 is positive enough to produce the amount of plate current needed to raise the cathode voltage enough to cut off V_1.

It should be noted that the grid of V_1 is at ground potential, since there is no grid-coupling circuit from the other tube. Therefore, V_1 remains cut off as long as the cathode voltage exceeds its cutoff voltage, because this is the only input voltage.

When the cathode voltage drops below the cutoff value, because of decreasing plate current in V_2, then V_1 can start to conduct again. Then conduction in V_1 cuts off V_2 again to repeat the cycle. The circuit operates as a free-running multivibrator, therefore, as each tube alternately conducts to cut off plate current in the other tube. The waveshapes are shown in Fig. 15·21.

Note the following points:

1. The e_{c_1} voltage is zero with respect to ground because there is no input to this grid. Therefore, no waveshape is shown for e_{c_1}.
2. Instead, the e_{gk} voltage is shown for V_1. This is the same as e_k but with inverted polarity.
3. When C_c charges by grid current, in this example e_{c_2} is reduced from $+12$ to $+4$ volts. At $+4$ volts, the plate current of V_2 is reduced enough to drop e_k to 6 volts, which is not enough to keep V_1 cut off. This point is where V_1 conducts to produce negative grid drive that cuts off V_2.

Summary of operation. V_2 is cut off by negative grid voltage when the plate voltage of V_1 drops with conduction. Therefore, the cutoff period for V_2 depends on the grid time constant for C_c to discharge through R_g. V_1 is cut off by the cathode voltage across R_k produced by maximum plate current in V_2. The reason why V_2 is not cut off when the increased voltage across R_k cuts off V_1 is that this is time when the grid of V_2 is driven positive from the plate of V_1. In fact, it is the positive grid drive at V_2 that makes the cathode voltage rise.

The cathode voltage e_k remains high enough to cut off V_1 until C_c is charged by grid current to the point when the positive grid drive on tube 2 drops close to zero. Therefore, the cutoff period for tube 1 depends on the time constant for charge of C_c. This charging path is through the low resistance of approximately 2,000 ohms internal grid-cathode resistance of V_2 when grid current flows, through R_k and R_{L_1} in the V_1 plate cir-

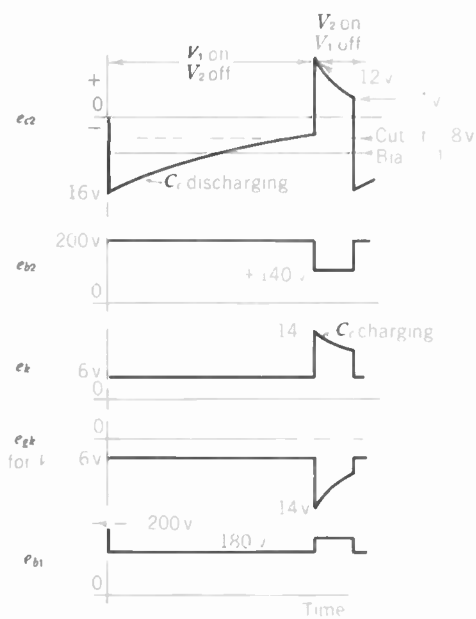


Fig. 15·21 Waveforms for cathode-coupled multivibrator in Fig. 15·20.

cuit while it is cut off. As a result, C_c charges fast for a short period of cutoff in V_1 compared with the slow discharge of C_c through R_2 for a long period of cutoff in V_2. The cathode-coupled multivibrator, therefore, automatically produces unsymmetrical output, as V_2 must be cut off for a much longer time than V_1.

Calculating the frequency. We can determine the frequency of the multivibrator in Fig. 15·20, with the waveshapes in Fig. 15·21, by calculating the cutoff time for both tubes. V_2 is cut off while C_c discharges through R_g from -16 to -8 volts. For discharge, the R_gC_c time constant equals 84.6 μsec. The cutoff time for V_2 equals 59.8 μsec, therefore, calculated from the RC discharge formula, or as 70 per cent of the RC time constant for a 50 per cent change in voltage.

The cutoff period for V_1 is the time it takes C_c to change the voltage across R_g from $+12$ to $+4$ volts. Although C_c is charging now, we are determining the time for a voltage change across R and, therefore, can use the RC discharge formula from Appendix E:

$$t = 2.3\ RC \log\left(\frac{e_2}{e_1}\right)$$

$$= 2.3 \times 8.42 \times 10^3 \times 470 \times 10^{-12} \times \log\left(\frac{12}{4}\right)$$

$$t = 4.35\ \mu\text{sec}$$

The R of 8,420 ohms is taken as 5,600 ohms R_{L_1} and 820 ohms R_k in series with 2,000 ohms internal grid-cathode resistance of V_2 when C_c is charging with grid current. The total cutoff period for both tubes then is 59.8 μsec plus 4.35 μsec, which equal 64.15 μsec. This time is slightly more than the horizontal line period of 63.5 μsec. Therefore, the frequency is a little less than 15,750 cps.

15·11 Multivibrator sawtooth generator

An unsymmetrical multivibrator can be used as a sawtooth generator by connecting a sawtooth capacitor in the plate circuit of the tube that is cut off a long time and conducting a short time. Then this stage functions as a discharge tube. In the cathode-coupled multivibrator this tube is the one with the RC coupling circuit for grid drive from the plate of the opposite tube. Figure 15·22 shows a typical circuit for a horizontal sawtooth oscillator. The same circuit can be used with different values for a vertical deflection oscillator.

In Fig. 15·22, C_2 is the sawtooth capacitor in the plate circuit of V_2, functioning as a discharge tube. While negative drive from V_1 holds V_2 cut off for a relatively long time C_2 charges toward B+ through R_6. The charging action produces a linear rise for the trace part of the sawtooth voltage across C_2. When the grid of V_2 is driven positive, because of cutoff in V_1, then C_2 can discharge fast for the retrace part of the sawtooth wave. The path of discharge current is through R_3 and the low internal

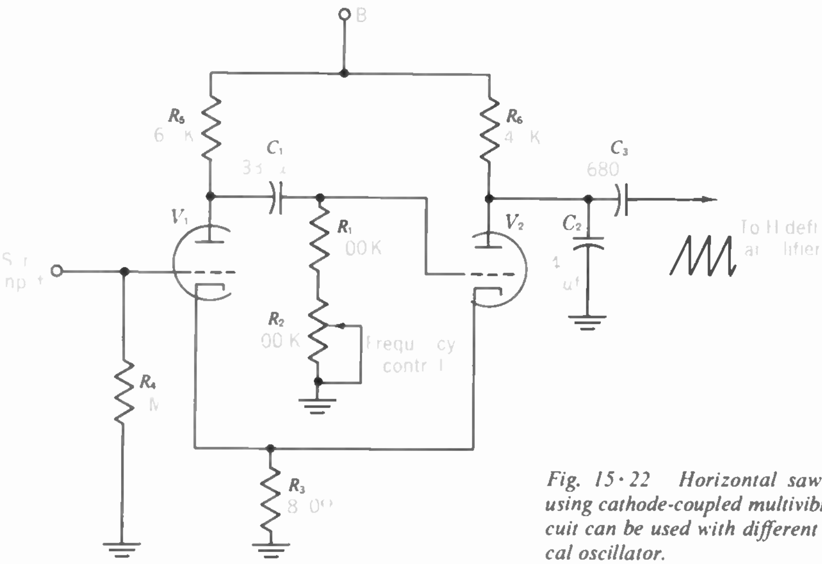


Fig. 15·22 Horizontal sawtooth generator using cathode-coupled multivibrator. Same circuit can be used with different values for vertical oscillator.

plate-cathode resistance in V_2, which is now conducting. The sawtooth voltage output across C_2 is coupled by C_3 to the grid circuit of the horizontal output stage. Figure 15·23 illustrates how the sawtooth voltage output corresponds to cutoff and conduction of V_2.

Note that the variable resistance R_2 in the grid circuit of V_2 serves as the frequency control. Decreasing R_2 reduces the cutoff time for V_2 to raise the free frequency of the oscillator. Increasing R_2 lowers the oscillator frequency. The grid resistance for V_1 does not control the oscillator frequency because its grid voltage is determined by the cathode voltage across R_3. However, this grid is the best connection for injecting sync because of isolation from the oscillator voltages.

15·12 Synchronizing the multivibrator

Either positive or negative sync polarity can be used with multivibrators.

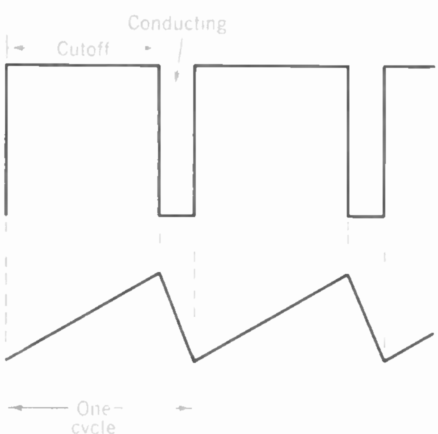


Fig. 15·23 Plate voltage of discharge tube without sawtooth capacitor (above) and corresponding sawtooth voltage with C_2 (below).

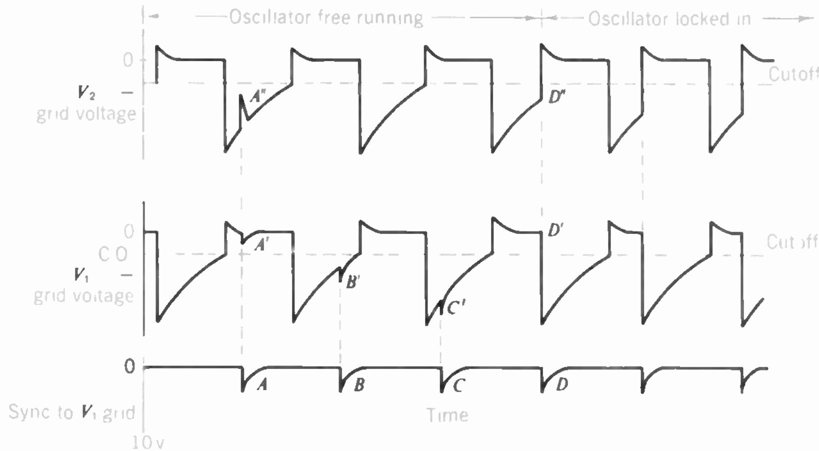


Fig. 15·24 Grid voltage waveforms in multivibrator synchronized by negative trigger pulses. The oscillator is pulled in to the sync frequency at time D.

A positive pulse applied to the grid of a cutoff tube can cause switching action if the pulse is large enough to raise the grid voltage above cutoff. This idea corresponds to triggering the grid of a blocking oscillator with positive pulse. However, negative trigger pulses can be used, particularly in the cathode-coupled multivibrator. The requirement is that the negative pulse applied to the conducting tube be amplified and inverted to produce a positive pulse large enough at the grid of the cutoff tube to make it conduct.

Synchronization with negative trigger pulses is illustrated in Fig. 15·24. At time A, the negative pulse at the grid of V_1 reduces its plate current but the amplified pulse is not positive enough at the grid of V_2 to make it conduct. Therefore, the pulse has no effect on switching. The negative pulses B and C also have no effect because they are applied to V_1 while it is cut off. However, at time D, the negative sync pulse is inverted and amplified enough to drive V_2 from cutoff into conduction. Then operating conditions are switched as V_2 conducts and cuts off V_1. Once V_2 is triggered by the sync voltage, the oscillator operates at the sync frequency as each sync pulse triggers each cycle.

Just as in blocking oscillator synchronization, the free frequency of the multivibrator must be slightly lower than the sync frequency. Then the pulses occur just before the natural switching action would take place by itself, when the oscillator is ready for triggering.

Negative d-c control voltage. The multivibrator frequency can also be varied by controlling its d-c grid voltage. This method applies when the oscillator is controlled by an AFC circuit for horizontal synchronization. In the cathode-coupled multivibrator, negative d-c control voltage at the V_1 grid raises the oscillator frequency. Its added negative grid voltage reduces plate current when V_1 conducts. The result is a smaller drop in plate voltage, and less negative drive at the V_2 grid. Then less time is needed for C_c to discharge down to cutoff for conduction in V_2. With a shorter cutoff time for V_2, the multivibrator frequency is increased.

15·13 Frequency dividers

Either the multivibrator or blocking oscillator can be triggered at a submultiple of the sync frequency to act as a frequency divider. In Fig. 15·25, as an example, the input sync frequency of 600 cps is divided by 3 to provide 200 cps for the frequency of the multivibrator output. The circuit is a divider as the oscillator output frequency is an exact submultiple of the sync frequency input.

The sync voltage is coupled to the multivibrator grid that needs positive pulses for triggering. For minimum sync voltage input, the free frequency of the oscillator is set slightly below the desired submultiple of the sync frequency. Then the oscillator is forced to lock in at the exact submultiple frequency because of the trigger pulses. Note that the third sync pulse labeled *C* in Fig. 15·25 has enough amplitude to trigger the oscillator from cutoff into conduction. Pulses *A* and *B* do not affect the oscillator because they do not have enough amplitude to raise the grid voltage above cutoff. The reason is that the free frequency slightly under 200 cps has been set much lower than the sync frequency of 600 cps. However, the third pulse *C* does trigger the oscillator into conduction. Then every third pulse triggers one oscillator cycle to produce output at one-third the sync frequency.

The voltages in Fig. 15·25 indicate how the frequency-division factor can be changed with more sync input. If the sync is increased to more than 10 volts, pulse *B* will have enough amplitude to trigger the oscillator. Then it divides by 2 with output at 300 cps, locked in by every second sync pulse. Or, sync voltage above 25 volts could lock in the oscillator at the sync input frequency, without any frequency division. This relatively large sync voltage to trigger the oscillator at the sync frequency is necessary because the oscillator free frequency is not close to the sync frequency.

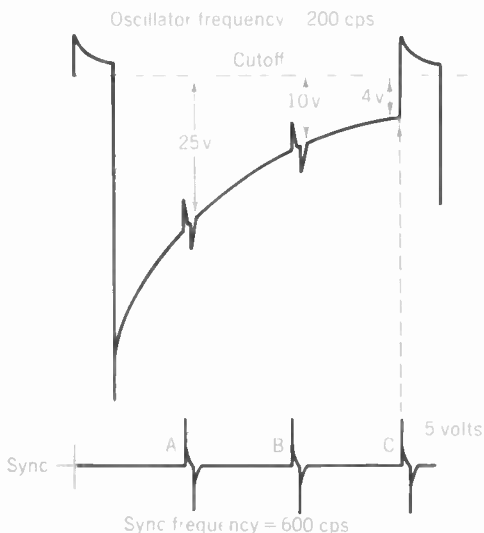


Fig. 15·25 Trigger pulses forcing multivibrator to operate as frequency divider at one-third the sync frequency.

15·14 Trapezoidal voltage waveshape

In magnetic scanning the sawtooth waveform for linear scanning is required for the current in the deflection coils of the yoke, since it is the magnetic field associated with the current that deflects the beam. With sawtooth current through the deflection coils, the voltage waveshape across the coils is not the same as the current waveshape because inductance opposes any change in current.

We can make a comparison with the more familiar example of inductance in sine-wave circuits. The self-induced voltage produced across the inductance by changes in current is 90° out of phase with the sine-wave current through the inductance. This concept of phase angle, however, applies only to sine waves. The corresponding idea in nonsinusoidal circuits is that inductance changes the waveshape of the voltage, compared with the current. More details of waveshaping with *RL* and *RC* circuits are explained in Appendix E but the main facts that apply here are illustrated in Fig. 15·26. With sawtooth current through an *RL* circuit:

1. The self-induced voltage e_L across the inductance has a rectangular waveform. This consists of a constant, relatively low value of induced voltage for the slow rise of sawtooth current with a constant rate of change during trace time, followed by a sharp voltage peak or spike of opposite polarity for the sharp drop in sawtooth current.
2. The *IR* voltage drop e_R has the same sawtooth waveshape as the current because the resistance has no reactance to change the waveform.
3. The combined waveform of voltage e_{RL} across the *RL* circuit is the algebraic sum of the sawtooth and rectangular voltages. This combination is a trapezoidal waveform.

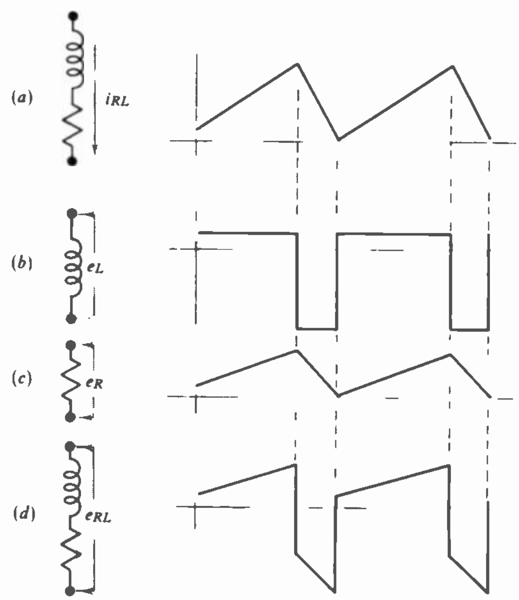


Fig. 15·26 Waveforms of current and voltage for inductance L and resistance R. (a) Sawtooth current in RL circuit. (b) Rectangular self-induced voltage across L. (c) IR voltage across R. (d) Trapezoidal voltage across L and R in series.

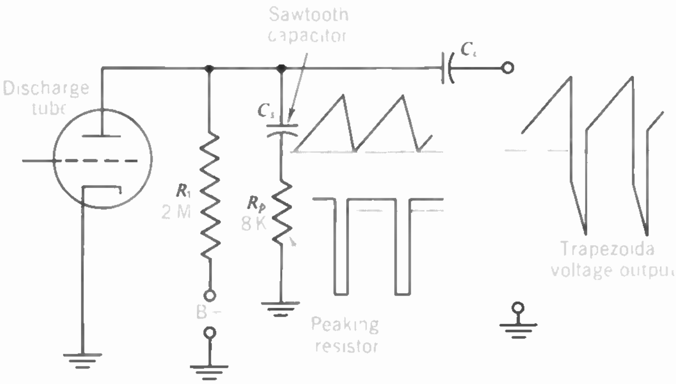


Fig. 15·27 Discharge tube with sawtooth capacitor and peaking resistor to produce trapezoidal voltage waveshape.

Therefore, sawtooth current through an *RL* circuit produces trapezoidal voltage across the combination. Or, trapezoidal voltage must be applied across an *RL* circuit to produce sawtooth current. Remember that for inductance alone, however, the rectangular voltage waveshape corresponds to sawtooth current.

These principles apply to the waveshapes in the horizontal and vertical output circuits for magnetic deflection. The horizontal output circuit is mainly inductive. Because of the large self-induced voltage produced by the fast horizontal flyback, the inductive effect is much greater than the resistance. Therefore, the voltage waveshape is rectangular for sawtooth current through the inductance in the horizontal output circuit. In the vertical output circuit, however, the self-induced voltage across the inductance is smaller at the lower frequency. Both resistance and inductance must be considered, therefore. Since it is an *RL* circuit, the voltage waveshape is trapezoidal for sawtooth current in the vertical output circuit.

As another application of trapezoidal voltage, this waveshape is often required for the output of the deflection oscillator to drive the power-output stage for sawtooth plate current. The negative peak on the trapezoid drives the amplifier grid voltage beyond cutoff to make sure its plate current drops to zero for flyback. At this time, the amplifier plate has high positive voltage induced by the decreasing plate current. Therefore, a high value of negative grid voltage is needed to cut off the tube, particularly with triodes.

The circuit used for generating the trapezoidal voltage waveform is shown in Fig. 15·27. This is the usual circuit for producing sawtooth voltage across C_s, with the addition of a peaking resistor R_p in series with C_s. The peaking resistor produces voltage spikes that are combined with the sawtooth voltage across C_s for trapezoidal voltage output.

The typical value of 8,000 ohms for R_p is small, compared with 2 megohms for R_1. Therefore, the voltage across R_p is small when C_s is charging from the B supply through R_1. On discharge, though, R_p is relatively large in comparison with the low resistance of the conducting discharge tube. When the discharge tube conducts, C_s discharges rapidly, developing a large negative pulse of voltage across R_p. The voltages across R_p and C_s are in series with each other across the plate-to-cathode circuit of the discharge tube. Therefore, the trapezoidal output voltage is the sum of the sawtooth voltage across C_s and the negative voltage spikes across R_p. The spike voltage is negative because it is produced by the discharge current of C_s.

15 · 15 Incorrect oscillator frequency

If the vertical hold control cannot stop the picture from rolling, even for an instant, this indicates the vertical oscillator is operating at the wrong frequency since it cannot be adjusted to 60 cps. With the oscillator frequency lower than 60 cps, the picture rolls upward; above 60 cps the picture rolls downward.

Vertical roll. The reason for the rolling picture is illustrated in Fig. 15 · 28 for the case of vertical oscillator frequency too high. Notice the relative timing of vertical blanking in the composite video signal and vertical retrace in the sawtooth vertical deflection current. When both are at the same frequency, every vertical retrace occurs within vertical blanking time. Then vertical blanking is not visible at the top and bottom edges of the frame. However, with the vertical frequency too high, the sawtooth cycles advance in time with respect to the 60-cycle blanking pulses. Then vertical blanking occurs during trace time, instead of during retrace. Furthermore, each sawtooth advances farther into trace time for succeeding blanking pulses. As a result, the black bar produced across the screen by the vertical blanking pulse appears lower and lower down the screen for successive cycles.

Remember that the information for the top of the picture as it is transmitted always comes immediately after vertical blanking. When the vertical oscillator is locked in sync, each frame is reproduced over the previous frame and then the picture holds still. However, when the picture information and blanking in each frame are reproduced lower on the screen than the previous frame, the picture appears to roll down. The same idea applies to rolling up.

The farther the vertical scanning frequency is from 60 cps, the faster the

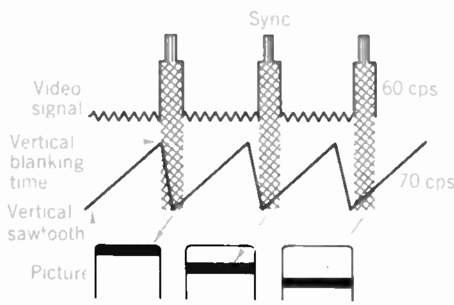


Fig. 15 · 28 Illustrating why the picture with blanking bar rolls up or down when vertical scanning frequency is not at 60 cps. Horizontal blanking omitted for clarity.

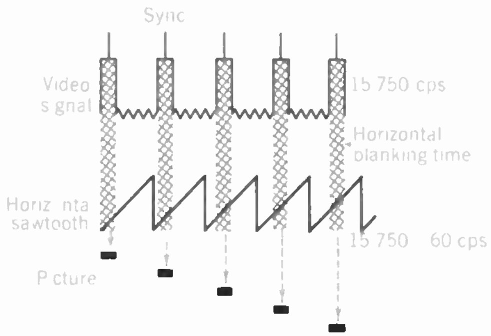


Fig. 15·29 Illustrating why the picture tears diagonally when horizontal scanning frequency is not at 15,750 cps.

picture rolls. The upward rolling is usually slower than the downward rolling because the oscillator frequency changes more gradually at the high-resistance end of the hold control for low frequencies. If the vertical oscillator frequency can be made as low as 30 cps, or 20 cps, which are submultiples of 60 cps, two or three duplicate pictures will be seen one above the other. If the frequency is raised to 120 cps, the bottom of the picture will be superimposed on the top, usually with reduced height in the raster.

The main factor in the frequency of a blocking oscillator or multivibrator is the R_gC_c time constant in the grid circuit. When either R_g or C_c has a value too small, the oscillator frequency is too fast. When the R_gC_c time constant is too long, the oscillator frequency is too slow.

Diagonal bars. When the horizontal oscillator is locked into the sync frequency of 15,750 cps, the line structure holds together to show a complete picture and horizontal blanking is invisible at the left and right edges. If the oscillator is off the correct frequency, the picture will tear into diagonal segments as shown in Fig. 14·2. The diagonal black bars are produced by horizontal blanking pulses. The picture is in segments because the horizontal AFC circuit prevents individual horizontal lines from tearing apart, as the frequency cannot change from line to line. When the number of diagonal bars is continually changing, this shows the AFC circuit is not controlling the oscillator. When the bars are steady, the AFC circuit is holding the oscillator but at the wrong frequency.

In either case, the horizontal blanking pulses produce diagonal black bars when the horizontal oscillator is off the correct frequency. If the frequency differs from 15,750 cps by 60 cps, there will be one diagonal bar. Every 60-cps difference between the oscillator frequency and 15,750 cps results in another diagonal bar. As the bars increase in number, they become thinner with less slope. The bars slope down to the left, when the oscillator frequency is below 15,750 cps, or up to the left above 15,750 cps.

The reason for the diagonal bars is illustrated in Fig. 15·29, for the case of horizontal oscillator frequency too high. Notice how successive sawtooth cycles advance in time with respect to the blanking pulses transmitted at 15,750 cps. The blanking pulse then goes into trace time. Each blanking pulse is 10 μsec wide, reproducing black for about one-sixth of every line. Remember that the left edge of the picture is always immediately after horizontal blanking. Furthermore, the blanking goes farther into trace time for successive sawtooth cycles. For each line, then, the black is farther to the right.

Since vertical scanning is occurring at the same time, the black area moves down as it progresses to the right. Only five scanning lines are illustrated here but if all the lines were shown the result would be one diagonal black bar from the top left corner to the bottom right corner. Then, the same action is repeated over the previous diagonal bar. The same idea applies to the case of oscillator frequency too low but the black would start at the top right corner and progress diagonally down to the left.

Between the black diagonal bars, the picture information is reproduced in the wrong position to such an extent that the picture usually cannot be recognized. Near the bottom of each bar, the picture information is actually reversed in its left-right position. Also, the horizontal flyback during visible time stretches black or white information all the way across the screen.

SUMMARY

1. Sawtooth voltage is produced by charging a capacitor C_s slowly through a high resistance for the slow linear rise and discharging the capacitor fast through a low resistance for the retrace. One complete cycle includes trace and retrace.
2. A peaking resistor in series with C_s on discharge produces a voltage spike that combines with the sawtooth to provide trapezoidal voltage waveshape.
3. In the blocking oscillator circuit in Fig. 15·6, plate current in the transformer induces grid feedback voltage that develops enough grid-leak bias to cut off the tube when feedback ceases. Then the oscillator has cut itself off until C_c discharges through R_g to reduce the bias to cutoff for the next cycle. The blocking rate is the frequency of the oscillator.
4. The blocking oscillator is cut off for a relatively long time and conducting for a short time. Therefore, it can control the charge and discharge of a sawtooth capacitor. In the plate circuit, C_s charges while the oscillator is cut off and discharges through the oscillator while it is conducting. The sawtooth voltage output is at the frequency of the blocking oscillator.
5. The blocking oscillator can be synchronized by positive trigger pulses at the grid. The free-running frequency is adjusted slightly below the sync frequency to allow every pulse to force the tube into conduction.
6. The multivibrator uses two stages that are alternately conducting and cut off. The period of conduction for each depends on the cutoff time of the other. One cycle includes cutoff for both stages.
7. In the plate-coupled multivibrator in Fig. 15·18, the plate circuit of each tube is RC-coupled to drive the grid of the opposite tube. Conduction in one tube drops its plate voltage to drive the next grid negative beyond cutoff.
8. In the cathode-coupled multivibrator in Fig. 15·20, V_1 cuts off V_2 through the R_gC_c coupling circuit but V_2 cuts off V_1 through R_k. The output is unsymmetrical, as V_2 is cut off a relatively long time and conducting a short time. Therefore, a sawtooth capacitor can be connected in the plate circuit of V_2, serving as a discharge tube, to provide sawtooth voltage output at the multivibrator frequency.
9. The cathode-coupled multivibrator can be synchronized by negative pulses at the V_1 grid, which are amplified and inverted to become positive trigger pulses at the V_2 grid.
10. The free frequency of a blocking oscillator or cathode-coupled multivibrator is usually

adjusted by making the grid resistor variable, to set the R_gC_c time constant. Larger values for R_g or C_c lower the oscillator frequency; decreasing R_gC_c raises the frequency.

11. The sawtooth voltage output can be adjusted by varying the resistance in series with C_s on charge. This size control increases the amplitude when the resistance is decreased; more resistance decreases the amplitude. Lower amplitudes allow better linearity, as a smaller part of the *RC* charge curve is used.
12. When the vertical oscillator is off 60 cps, the picture with the black horizontal bar produced by the vertical blanking pulses rolls up or down the screen. When the horizontal oscillator is off 15,750 cps, the picture tears into diagonal segments, with black diagonal bars produced by the horizontal blanking pulses.

SELF-EXAMINATION *(Answers at back of book.)*

Part A

Answer true or false.

1. The sawtooth capacitor charges through the discharge tube while it is cut off.
2. The vertical oscillator cannot operate without sync input.
3. The horizontal oscillator must be operating for the receiver to have flyback high voltage.
4. If the vertical oscillator does not operate, there will be just a bright line across the center of the screen, instead of the raster.
5. The vertical and horizontal oscillators cannot produce deflection unless a station is tuned in.
6. Decreasing plate current induces negative feedback voltage for the blocking oscillator grid.
7. Decreasing the grid resistance raises the frequency in a blocking oscillator.
8. Increasing the capacitance of C_s reduces the sawtooth voltage amplitude, improves linearity, but does not change the frequency.
9. Increasing the charging resistance for C_s reduces the sawtooth voltage amplitude.
10. The free frequency of the oscillator should be set lower than the sync frequency.
11. When plate voltage drops, C_c and R_g couple negative voltage drive to the next grid.
12. The cathode-coupled multivibrator can be used for either a vertical or horizontal sawtooth generator.
13. In the cathode-coupled multivibrator, C_s is in the plate circuit of the tube that has R_gC_c coupling into its grid.
14. In a multivibrator, the sync voltage must drive the cutoff tube into conduction to lock in the oscillator at the sync frequency.
15. In a cathode-coupled multivibrator, the free frequency depends on the R_gC_c time constant.
16. In a cathode-coupled multivibrator, the voltage across R_k cuts off both tubes at the same time.
17. When the picture rolls up the vertical oscillator frequency is above 60 cps.
18. Two diagonal bars sloping down to the left indicate the horizontal oscillator frequency is 120 cps below 15,750 cps.
19. When the rolling can be stopped with the vertical hold control but the picture does not lock in, this indicates no vertical sync.
20. If the oscillator frequency is increased, the sawtooth voltage across C_s will decrease.

Part B

Fill in the required value.

1. A 0.05-μf sawtooth capacitor charging through 3 megohms has a charge time constant of ______ sec.
2. The same capacitor discharging through 8,000 ohms and 1,000 ohms in series has a discharge time constant of ______ sec.
3. A 0.01 coupling capacitor charged to -60 volts will discharge through a 3-megohm grid resistor down to -6 volts in ______ sec.
4. If the vertical blocking oscillator transformer has a resonant frequency of 2,000 cps, the time for one-half cycle is ______ μsec.

5. In the multivibrator of Fig. 15·18, if each tube is cut off for 500 μsec, the oscillator frequency is ______ cps.
6. In Fig. 15·22, if V_1 is cut off 7.5 μsec and V_2 cut off 58 μsec, the sawtooth voltage output frequency is approximately ______ cps.
7. In Fig. 15·22, when R_2 is at zero, the R_1C_1 time constant equals ______ μsec.
8. In Fig. 15·22, when the cathode current through R_3 is 20 ma, the cathode voltage equals ______ volts.
9. In Fig. 15·11, if the Ec_s sawtooth voltage varies between 140 and 110 volts, the peak-to-peak output equals ______ volts.
10. In Fig. 15·25, if the free frequency is set at 295 cps, the oscillator frequency will lock in at ______ cps.

ESSAY QUESTIONS

1. What is the function of the vertical deflection oscillator? The horizontal deflection oscillator?
2. Why is the sawtooth waveshape required for linear deflection? In magnetic deflection, why is the sawtooth waveshape required for current in the coils?
3. Give two factors that affect linearity of the voltage rise across C_s.
4. Give three factors that affect cutoff time of the blocking oscillator. Give one factor affecting conduction time.
5. Referring to the numbered steps in the e_c waveshape in Fig. 15·9, which numbers occur during conduction and which during cutoff?
6. Draw the schematic diagram of a blocking oscillator–discharge tube trapezoidal voltage generator for vertical deflection, using a single triode. Indicate typical values of all components. Label the height and hold controls.
7. Why is the free-running frequency of a blocking oscillator lower than the sync frequency? Why is the polarity positive for sync voltage at the grid? Why does the frequency increase when positive d-c grid voltage is added?
8. Referring to the cathode-coupled multivibrator in Fig. 15·20, what determines cutoff time and conduction time for V_2? Cutoff and conduction time for V_1?
9. Referring to the cathode-coupled multivibrator in Fig. 15·22, draw the voltage waveshapes at each electrode of V_1 and V_2 in a ladder diagram, with one under the other to show the different voltages at similar times.
10. In a cathode-coupled multivibrator, why is the polarity negative for sync voltage at the grid of V_1? Why is the free frequency set lower than the sync frequency? Why does negative d-c voltage added to the V_1 grid increase the frequency?
11. Describe briefly the function of the height control and vertical hold control in a vertical deflection oscillator.
12. Illustrate an example of a blocking oscillator set for a free frequency of 18 cps, with 100-cps sync input, resulting in oscillator output at 20 cps. Why is this circuit called a frequency divider? How much is the division factor here?
13. Why does the multivibrator in Fig. 15·18 have symmetrical output? How could the cutoff time of tube 2 be made shorter for unsymmetrical output?
14. Show the waveform of sawtooth current through a pure inductance without resistance, and the corresponding voltage waveform across the inductance.
15. In a circuit with the sawtooth capacitor in the plate of the blocking oscillator, why does the picture usually roll when the height control is adjusted? Will this also happen with a cathode-coupled multivibrator for the vertical deflection oscillator? Why?
16. Explain briefly why the height of the raster can decrease when the vertical oscillator frequency is much higher than 60 cps.
17. Give the function of the peaking resistor in a trapezoidal voltage generator.
18. If the vertical oscillator locks in at 30 cps, what will be the effect on the picture? What will be the effect at 120 cps?

19. Name five components or voltages that affect the frequency of a multivibrator or blocking oscillator.
20. Referring to Fig. 15 · 11, give the function of all components.
21. Referring to Fig. 15 · 22, give the function of all components.
22. If R_3 in Fig. 15 · 22 opens, what will be seen on the kinescope screen, in a receiver using flyback high voltage?
23. If the primary winding opens in the blocking oscillator transformer in Fig. 15 · 11, what will be seen on the kinescope screen?
24. If the horizontal oscillator is at 15,690 cps, how will the picture look on the kinescope screen?
25. Draw a diagram of a blocking oscillator using an NPN transistor in a common-emitter circuit.

PROBLEMS *(Answers to odd-numbered problems at back of book.)*

1. With 100 volts applied, C initially uncharged, and a time constant of 2 sec, calculate E_c for the following charging periods: (*a*) 0.2 sec; (*b*) 1.0 sec; (*c*) 1.14 sec; (*d*) 2 sec; (*e*) 4 sec; (*f*) 10 sec.
2. For the examples in Prob. 1, draw a graph showing E_c vs. time in seconds.
3. (*a*) A capacitor charged to 10 volts is connected to a 110-volt source. How much is E_c after one time constant? (*b*) Then C is discharged for one time constant. How much is E_c?
4. A capacitor has a time constant of 0.045 sec for charge. A separate discharge path has a time constant of 0.0002 sec. The capacitor is charged from a 100-volt source for 0.009 sec and then discharged for 0.001 sec. Draw the resulting sawtooth voltage across C, to scale, in voltage vs. time. What is the frequency of the sawtooth voltage?
5. A 1-μf capacitor charged to 100 volts is discharged through a 1-megohm resistor. (*a*) How much is the peak discharge current at the start of discharge? (*b*) How much is the resistor voltage then? (*c*) How much is the discharge current after 0.707 sec? (*d*) How much is the resistor voltage now?
6. R_g is 1 megohm with 1-μf C_c charged to an average plate voltage of 150 volts. Then the plate voltage drops to 50 volts and remains at that value for 2 sec. Draw a graph with values of current and voltage across R_g during the discharge.
7. In Fig. 15 · 13, how much will a d-c voltmeter across C_s read for: (*a*) $RC = 0.1$ sec; (*b*) $RC = 0.15$ sec?
8. What is the approximate frequency of the plate-coupled multivibrator in Fig. 15 · 18, with the waveshapes in Fig. 15 · 19, if C_1 and C_2 are doubled in value?
9. Calculate the voltage induced across 0.5 henry inductance when the current: (*a*) increases from 0 to 20 ma in 100 μsec; (*b*) decreases from 20 to 5 ma in 60 μsec; (*c*) decreases from 5 to 0 ma in 50 μsec.
10. A sawtooth waveform of current through an 8-mh inductance increases from 0 to 300 ma during 50 μsec, then drops to zero in 5 μsec. Draw the waveforms of current and corresponding induced voltage, with values.

Chapter 16 Horizontal afc circuits

A deflection oscillator triggered by individual sync pulses for each cycle is capable of exact synchronization, if there is no noise interference. However, interfering noise pulses can be mistaken for synchronizing pulses and trigger the oscillator at the wrong time, causing loss of synchronization. Noise pulses are especially troublesome at the ripe time of the oscillator cycle, just before the desired sync pulse occurs. In order to make the synchronization more immune to noise, an AFC circuit is used for the horizontal deflection oscillator in practically all television receivers. This is generally called *flywheel sync, sync lock, stabilized sync,* or *horizontal AFC.* The AFC circuit is generally the only section used for horizontal sync alone. Therefore, when the picture tears into diagonal bars too easily, the trouble is likely to be in the horizontal AFC. Automatic frequency control is generally not used for the vertical deflection oscillator.

16·1 AFC requirements

The typical arrangement of an AFC circuit for the horizontal deflection oscillator is illustrated in Fig. 16·1. The operation can be considered in the following steps:

1. Horizontal sync voltage and a fraction of the horizontal deflection voltage are coupled into the frequency-comparing circuit. The deflection voltage is needed as a sample of the oscillator frequency. It can be taken either from the oscillator or horizontal output circuit.
2. The frequency-comparing circuit produces d-c output voltage proportional to the difference in frequency between its two input voltages.
3. The d-c control voltage indicates whether the oscillator is on or off the sync frequency. The greater the difference between the oscillator and sync frequencies, the larger is the d-c control voltage.
4. The d-c control voltage is filtered by an *RC* integrator. With a shunt

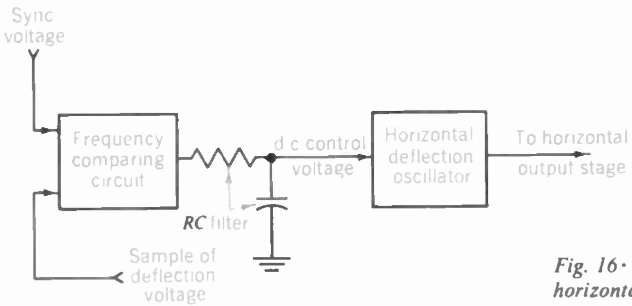


Fig. 16·1 Block diagram of AFC circuit for horizontal oscillator.

capacitor bypassing noise pulses, they cannot affect the control voltage appreciably. Therefore, the low-pass filter prevents fast changes in d-c control voltage.

5. The filtered d-c control voltage changes the oscillator frequency by the amount necessary to make the scanning frequency the same as the sync frequency. With a multivibrator or blocking oscillator, the d-c control voltage is coupled directly to the grid of the horizontal oscillator to correct its frequency.

In a blocking oscillator, making the grid voltage more positive raises the frequency. In a cathode-coupled multivibrator more negative d-c voltage at the sync grid (V_1 in Fig. 15·22) is inverted to more positive voltage at the V_2 grid to raise the frequency. The reverse applies to lowering the oscillator frequency. As a result, the d-c control voltage holds the oscillator at the sync frequency, as any difference between the two is measured by the control circuit to produce the required amount of correction voltage.

Discriminator with multivibrator. The most common type of frequency-comparing circuit is a sync discriminator that compares the frequency of the sync pulses and sawtooth deflection voltage. A duodiode vacuum tube can be used, or the dual crystal diode unit shown in Fig. 16·2. The d-c control voltage from the sync discriminator generally controls the frequency of a cathode-coupled multivibrator circuit for the horizontal oscil-

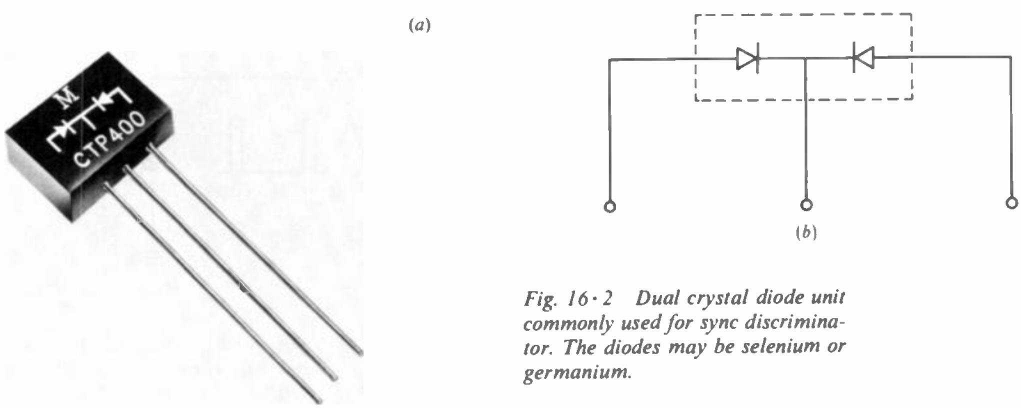


Fig. 16·2 Dual crystal diode unit commonly used for sync discriminator. The diodes may be selenium or germanium.

lator. For typical circuits, negative d-c control voltage raises the multivibrator frequency; positive d-c control voltage lowers the frequency. The sync discriminator is a high-impedance circuit suitable for the multivibrator triode section that does not draw grid current.

D-C control tube with blocking oscillator. With a blocking oscillator for the horizontal oscillator, the AFC circuit generally uses a triode as a d-c control tube. A dual triode is used, with one section for the control tube and the other the blocking oscillator sawtooth generator. Then the output voltage of the control tube can make the oscillator grid voltage more positive to raise the frequency; less positive d-c control voltage lowers the blocking oscillator frequency. This horizontal AFC circuit is often called *Synchro-Guide* or a *pulse-width* control circuit.

16·2 Push-pull sync discriminator

In Fig. 16·3, opposite polarities of sync voltage from a sync inverter stage are coupled to the two diodes. R_2C_2 couples positive sync to the V_2 plate, while R_1C_1 couples negative sync to the V_1 cathode. Remember that negative voltage at the cathode makes diode current flow, just like positive voltage at the plate. As a result, the sync input voltage drives both diodes into conduction.

Sawtooth input for the diodes. Deflection voltage is needed as a sample of the oscillator frequency. Therefore, flyback pulses from a winding on the horizontal output transformer are applied to the R_3C_3 network. The polarity is chosen to make the slope of the flyback on the sawtooth increase in the positive direction. Each positive pulse charges C_3 during the fast flyback. Then C_3 discharges during the relatively long time between pulses. The result is the sawtooth voltage shown across C_3, which is applied to the plate of V_1 and the cathode of V_2.

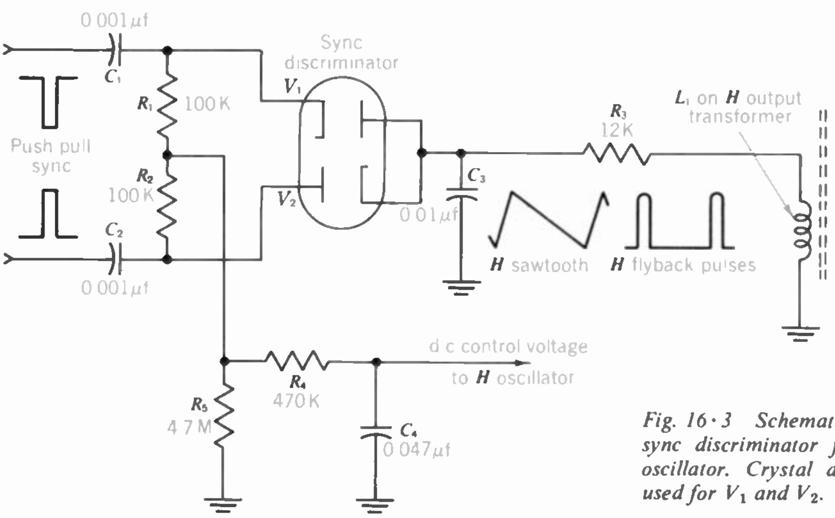


Fig. 16·3 Schematic of push-pull sync discriminator for horizontal oscillator. Crystal diodes can be used for V_1 and V_2.

The sawtooth input voltage is applied in the same polarity to both diodes. Note that when this voltage makes the V_1 plate positive it aids the diode current. However, this positive voltage at the V_2 cathode opposes diode current, like negative plate voltage.

D-C control voltage from the sync discriminator. The a-c input is rectified by the diodes to produce the required d-c control voltage in the output. Each diode is a peak rectifier for the sync pulses. This means the amount of diode conduction depends on the peak value of the a-c input voltage. The amount of conduction depends on how the sync input voltage is phased with respect to flyback on the sawtooth voltage. Note that C_4 in series with R_4 is in a common path for both diodes. R_5 provides a d-c return to chassis ground.

Conduction in V_1 makes the voltage across C_4 more positive. When V_2 conducts the current is in the opposite direction and the voltage across C_4 becomes more negative. When both diodes conduct the same amount of current, the net voltage across C_4 is zero. This condition of zero d-c control voltage occurs when the sync and scanning frequencies are the same. When the oscillator is below or above the sync frequency, C_4 provides negative or positive d-c control voltage to correct the oscillator frequency.

Discriminator waveshapes. Figure 16·4 shows how the sync and sawtooth voltages combine to control conduction in the discriminator diodes. In *a* each sync pulse occurs in the middle of flyback time because the oscillator is on frequency. At this time, the sawtooth input voltage is at the same value of zero for both diodes. The zero level is the average value axis of the a-c sawtooth wave. With equal sync voltages, therefore, the peak input voltage has the same amplitude for either diode. Then they conduct equal currents to provide a net voltage of zero across C_4.

In *b*, the oscillator frequency is too high. As the sawtooth cycle takes

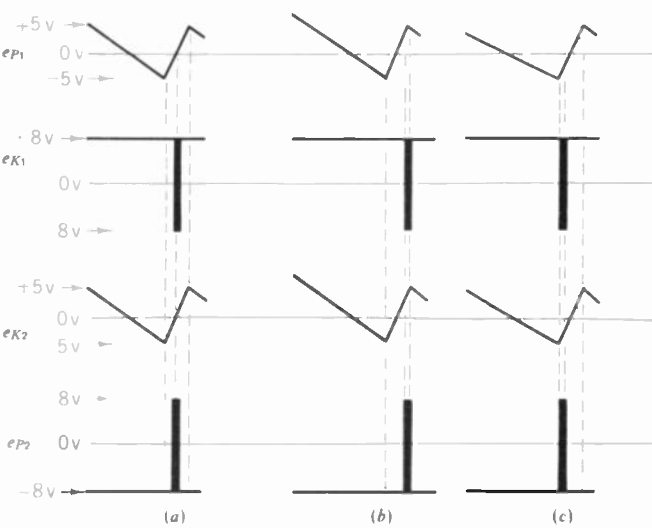


Fig. 16·4 Waveshapes of diode voltages for sync discriminator in Fig. 16·3. (a) Normal control with sync and sawtooth same frequency. Sync pulse in middle of flyback. (b) Oscillator too fast with shorter sawtooth cycle. Sync pulse at end of flyback. (c) Oscillator too slow with longer sawtooth cycle. Sync pulse at start of flyback.

less time, the sync pulse occurs later, at the end of the flyback voltage. Then the sawtooth voltage makes the V_1 plate 5 volts positive. This can be combined with -8 volts for the tip of sync at the V_1 cathode, which is also in the forward direction for diode current. The peak voltage for V_1 then equals $8 + 5$ or 13 volts. For V_2, however, this $+5$ volts of the sawtooth makes the cathode positive. Here the polarity is in the reverse direction, like negative plate voltage. Therefore, the $+5$ volts cancels part of the $+8$ volts for tip of sync at the plate, and the net voltage for V_2 equals $8 - 5$ or 3 volts. V_1 conducts more than V_2 and the result is positive d-c control voltage across C_4.

In *c* the oscillator frequency is too low. As the sawtooth cycle takes more time, the sync pulse now occurs earlier, at the start of flyback voltage. Now the sawtooth voltage makes the V_1 plate 5 volts negative at the time of the sync pulse. Then V_1 has only $8-5$ or 3 volts plate-cathode voltage. However, the -5 volts at the cathode of V_2 aids the $+8$ volts sync at the plate and its plate-cathode voltage is 13 volts. Now V_2 conducts more than V_1 and the result is negative d-c control voltage across C_4.

D-C control voltage. The filtered d-c voltage from the sync discriminator usually controls the frequency of a cathode-coupled multivibrator, as in Fig. 16·5. Then negative d-c control voltage at the synchronizing grid raises the oscillator frequency. The discriminator action in controlling the oscillator can be summarized as follows:

1. When the oscillator frequency is too low, V_2 conducts more than V_1 to produce negative control voltage that raises the multivibrator frequency.
2. When the oscillator frequency is too high, V_1 conducts more than V_2 to produce positive control voltage that lowers the oscillator frequency.

As a result, the sync discriminator continuously measures the frequency difference between the sync and sawtooth waves to produce the d-c correction voltage that locks in the horizontal oscillator at the synchronizing frequency.

16·3 Circuit of multivibrator controlled by sync discriminator

The schematic diagram in Fig. 16·5 illustrates a cathode-coupled multivibrator for the horizontal oscillator controlled by the same sync discriminator circuit as in Fig. 16·3. For the multivibrator, C_{420} is the sawtooth capacitor in the plate circuit of the stage cut off for a longer period of time than it conducts. Its free-running frequency is controlled by the grid time constant of C_{419} with R_{433} and R_{434}. The filtered d-c control voltage from the discriminator is connected to oscillator pin 1, which is the grid for sync input.

Sine-wave ringing coil. Note the coil L_{401} in Fig. 16·5, which produces sine-wave ringing with C_{418}. Although connected to plate pin 2 of V_{405}, the tuned circuit couples sine-wave voltage to grid pin 4. The pulses of plate current shock-excite the tuned circuit to produce sine-wave oscillations. The combined waveform in Fig. 16·6 shows how the grid voltage

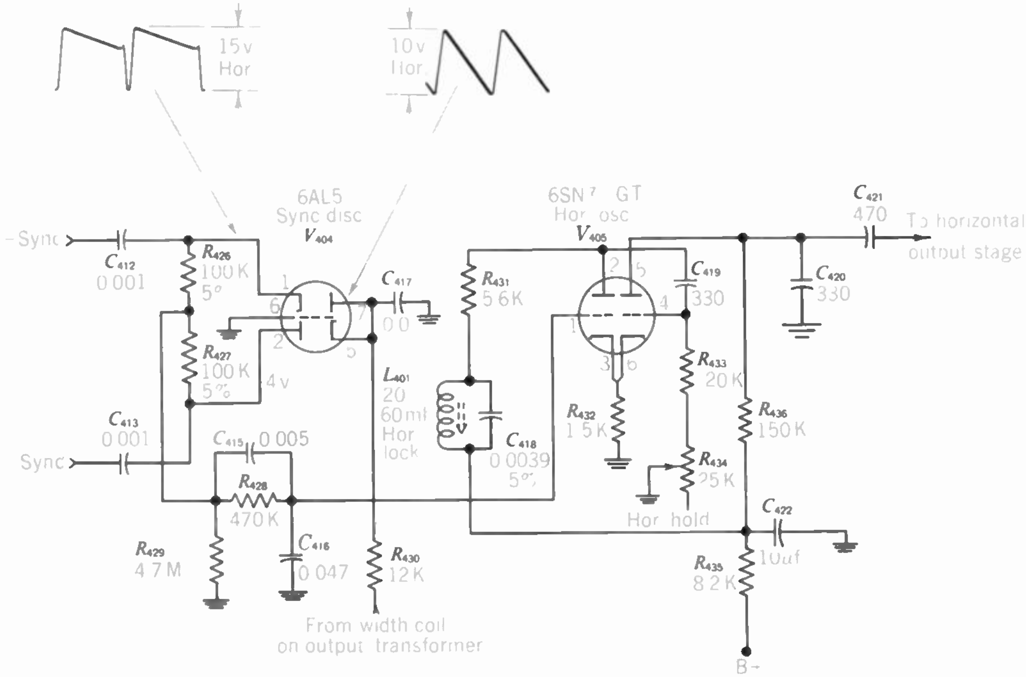
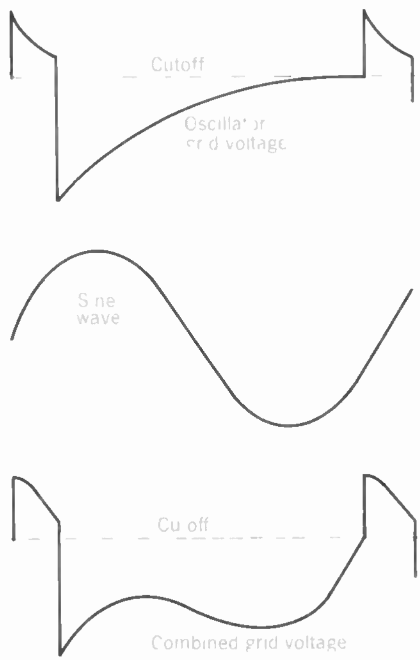


Fig. 16·5 Complete circuit of cathode-coupled multivibrator horizontal oscillator controlled by push-pull sync discriminator. Capacitance values less than 1 in μf and more than 1 in μμf.

approaches cutoff with a much sharper slope because of the sine-wave component. This means that noise pulses must have a much higher amplitude to have any effect on the oscillator. Although the oscillator frequency is controlled by d-c control voltage from the AFC circuit, there may be direct pickup of interfering pulses at the oscillator grid.

The inductance L_{401} is generally called a *stabilizing coil.* It prevents the oscillator from being triggered by noise pulses, allowing only the AFC voltage to control the oscillator frequency. The frequency of the sine-wave stabilizing circuit is set just below 15,750 cps. Its resonant frequency is determined by the inductance of the coil and the capacitance across it. A typical frequency is approximately 15,000 cps.

Fig. 16·6. Effect of sine-wave stabilizing tuned circuit on oscillator grid-voltage waveform.



Oscillator frequency adjustments. The hold control R_{434} in Fig. 16·5 is a variable grid resistor for the oscillator to adjust its free-running frequency independently of the AFC circuit. The stabilizing coil L_{401} must also be adjusted to allow the oscillator to operate at the horizontal scanning frequency of 15,750 cps. An approximate adjustment can be made by adjusting the stabilizing coil to lock in the picture, with the hold control at its middle setting. A more exact procedure is as follows:

1. Short the d-c control voltage temporarily with a jumper wire from sync grid pin 1 to chassis. This is done to operate the oscillator without any d-c control voltage.
2. Connect a resistance of about 1,000 ohms temporarily across the stabilizing coil. This is done to eliminate sine-wave ringing, while maintaining the same d-c electrode voltages.
3. With weak signal input, adjust the hold control until the picture just floats from side to side across the screen. This shows the oscillator is free-running at 15,750, without lock-in and without the sine wave ringing.
4. Remove the resistor across the stabilizing coil. The oscillator may now be off frequency. Adjust the coil until the full picture again floats sideways. This shows the oscillator frequency is at 15,750 cps without lock-in.
5. Remove the jumper wire shorting the d-c control voltage to allow the AFC circuit to hold the oscillator locked in at 15,750 cps. The AFC should hold the picture in sync when changing channels and at least through one-half rotation of the hold control.

16·4 Single-ended sync discriminator

The AFC circuit in Fig. 16·7*a* does not require push-pull sync input. Instead, negative sync voltage is coupled by C_1 to the common cathode of the two diodes D_1 and D_2. Effectively, the diodes are in parallel for the sync input, as shown in *b*. The R_1C_1 coupling circuit provides negative sync voltage across R_1 for D_1. Similarly, R_2C_1 couples sync voltage for D_2. The low side of R_2 is returned to ground through the series combination of C_3 and C_4.

Sawtooth input. Sawtooth voltage is also coupled to the diodes, to compare the oscillator frequency with sync. The slope of flyback voltage on the sawtooth wave should be negative, which is the polarity generated by the sawtooth capacitor in the oscillator circuit. As shown in Fig. 16·7*c* the sawtooth voltage coupled by C_3 is applied across the two diodes in series. Therefore, each diode has one-half the sawtooth input voltage.

Notice that the sawtooth voltage is applied plate-to-cathode for D_2 but cathode-to-plate for D_1. Let us consider the sawtooth as cathode-to-plate voltage for D_2, the same way sync is applied to both diodes. Then the sawtooth voltage for D_2 can be shown inverted, as in Fig. 16·8. In effect, the two diodes have equal and opposite sawtooth input voltages. Notice that this circuit has push-pull sawtooth input, compared with push-pull sync input for the discriminator in Fig. 16 • 3.

D-c control voltage. The negative sync voltage input at the cathode, combined with sawtooth voltage input, makes the diodes conduct. When D_1 conducts, the result is d-c voltage E_{R_1} across R_1 with the polarity shown in *a*. The electron flow illustrated for i_p is opposite from the arrowhead in the rectifier, which shows the movement of positive hole charges. In any case, the rectified d-c voltage output is positive at the cathode of the rectifier. Similarly, conduction in D_2 results in the d-c voltage E_{R_2} shown across R_2.

The d-c control voltage, taken from point C with respect to ground, consists of E_{R_1} and E_{R_2} in series opposition. When the two voltages are equal, the net output at C is zero. E_{R_2} larger than E_{R_1} results in negative output voltage, but when E_{R_1} is larger the control voltage is positive. The d-c control voltage is filtered by R_3C_5 for the grid of the horizontal oscillator, generally a cathode-coupled multivibrator.

Function of each component. Summarizing the functions of the components in Fig. 16 • 7*a*, R_1 and R_2 are the diode load resistors. C_1 is the

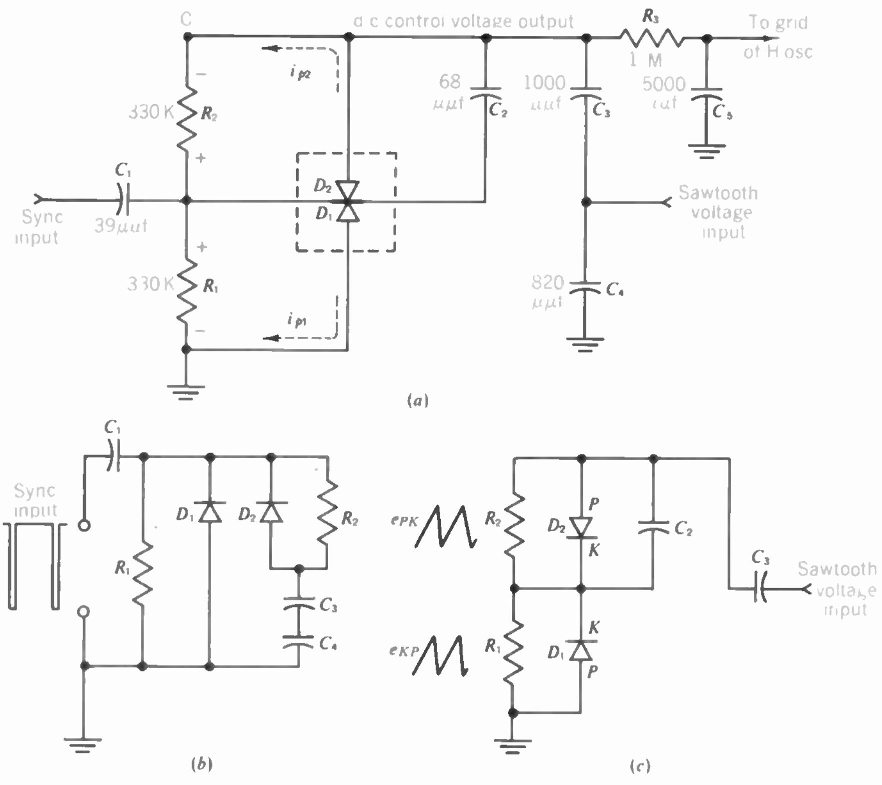


Fig. 16 • 7 Single-ended sync discriminator for horizontal AFC. (a) Circuit with d-c control voltage at point C. (b) Equivalent circuit showing diodes in parallel for sync voltage input. (c) Diodes in series for sawtooth voltage input.

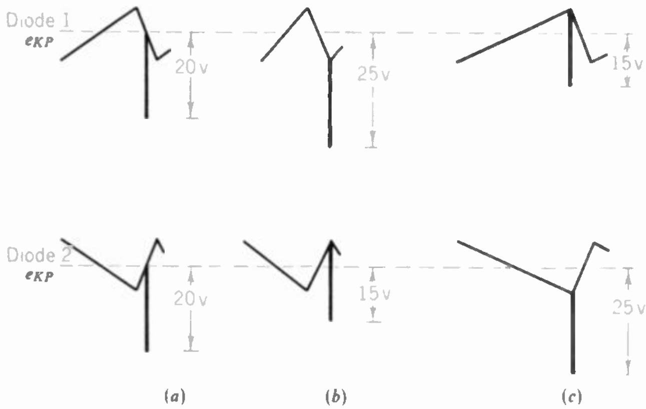


Fig. 16·8 Waveshapes of diode voltages for sync discriminator in Fig. 16·7. (a) Normal control. Oscillator at 15,750 cps. (b) Oscillator too fast. (c) Oscillator too slow.

coupling capacitor for negative sync input to both diodes in parallel. The plate of D_2 is returned to chassis ground through C_3 and C_4 in series. Also, C_4 is needed for a ground return without shorting B+ of the discharge tube supplying sawtooth voltage. C_3 couples sawtooth voltage input to the two diodes in series, without shorting the d-c control voltage. C_2 equalizes the amount of sawtooth voltage across the diodes. In order to help balance the input to the diodes, C_2 is larger than C_1 but smaller than the series combination of C_3 and C_4. Finally, R_1 can be smaller than R_2 but proportioned to compensate for unbalance in the diode circuits.

Discriminator waveshapes. In Fig. 16·8, the sync voltage has negative polarity from cathode to plate for both diodes. The sawtooth voltage is also applied with negative polarity during flyback for the cathode-to-plate circuit of D_1. However, for D_2 its sawtooth voltage has the reverse polarity when considered cathode-to-plate, as shown in Fig. 16·8.

The sawtooth voltage from the horizontal oscillator is at the correct frequency of 15,750 cps in Fig. 16·8*a*. Then the peak amplitude of the negative sync pulse is equal for both diodes. They conduct the same amount, producing equal and opposite d-c output voltages across R_1 and R_2. As a result, the d-c control voltage is zero. In *b*, the oscillator frequency is too high. The sync pulses then occur later in the sawtooth cycle, toward the end of flyback time. Therefore, diode 1 with negative polarity of flyback voltage has more peak negative voltage from cathode to plate to conduct more than diode 2. Since D_1 produces positive d-c output voltage across R_1, the d-c control voltage becomes positive. This polarity of d-c grid voltage reduces the frequency of a cathode-coupled multivibrator. For the opposite case in *c*, the oscillator is too slow. Then the sync pulse occurs toward the start of flyback, allowing more peak voltage input to diode 2. This diode conducts more now to produce negative d-c control voltage to raise the oscillator frequency. As a result, the two diodes continuously measure the difference between the sync and sawtooth frequencies to correct the horizontal oscillator and hold its frequency at 15,750 cps.

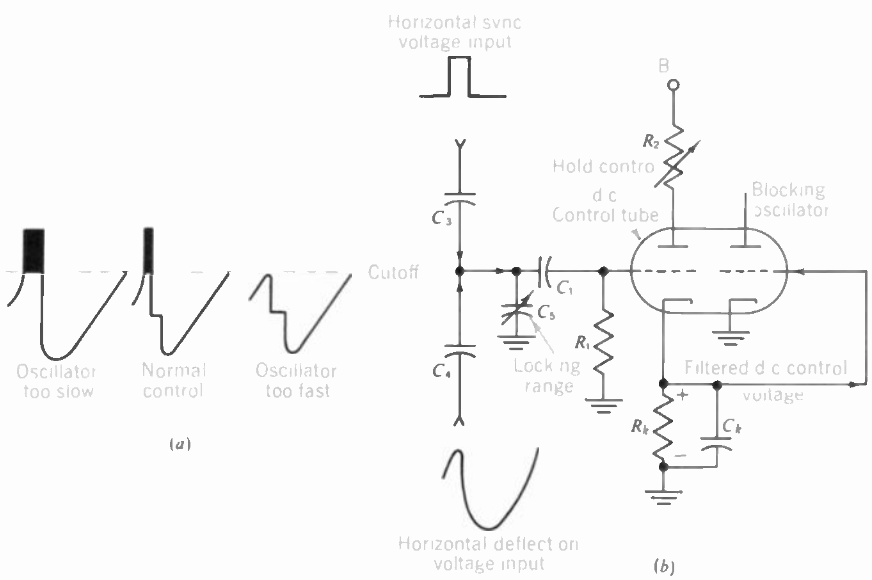
Typical values of d-c control voltage. Because of slight unbalance in the single-ended discriminator circuit in Fig. 16·7, the actual d-c control voltage is approximately −1 volt at the sync grid of the horizontal oscillator when it is on frequency at 15,750 cps. This voltage can vary between +6 volts and −6 volts, approximately, to correct the oscillator frequency. Positive d-c control voltage lowers the multivibrator frequency, while negative voltage increases the frequency. Six volts can correct the multivibrator frequency ±600 cps above or below 15,750 cps.

16·5 D-C control tube (Synchro-Guide)

The AFC circuit illustrated in Fig. 16·9 uses a triode amplifier as a control tube to produce positive d-c control voltage for the grid of a blocking oscillator. Plate current produces the required voltage drop across R_k, filtered by C_k. This cathode voltage is directly connected to the oscillator grid to correct its frequency. More positive d-c control voltage raises the blocking oscillator frequency; less positive control voltage lowers the frequency.

The amount of d-c correction voltage produced by the cathode resistor depends upon how much plate current flows. In order to regulate the amount of plate current through the control tube in accordance with the difference between the sync and scanning frequencies, the sync and scanning voltages are coupled to the control grid. The horizontal sync voltage from the sync circuits is coupled by C_3, while C_4 couples voltage from the horizontal sawtooth generator. The deflection voltage at the grid

Fig. 16·9 D-C control tube for horizontal AFC with blocking oscillator in Synchro-Guide circuit. (a) Waveshapes at grid of control tube. Shaded area produces pulse current and d-c control voltage across R_k. (b) Circuit arrangement.



of the control tube has a parabolic waveshape, like distorted half-sine waves, obtained by integrating the sawtooth voltage. The parabolic waveform is used because it has a sharper slope than the sawtooth wave just before the peak at the start of flyback.

Waveshapes at grid of control tube. The combined grid voltage for the control tube then consists of the horizontal sync pulses atop the peaks on the deflection voltage (see Fig. 16 · 9*a*). As a result, plate current in the control tube depends upon how much of the sync pulse is on the peak of the deflection voltage. When the phase between the two input voltages changes, the grid voltage effective in producing plate current varies. With sync and scanning voltages of the same frequency and phase, approximately one-half the width of the sync pulse is on the deflection voltage peak to produce plate current for the required amount of correction voltage. The remaining one-half of the sync pulse occurs after the peak and produces the step shown at the lower part of the deflection voltage.

When the sync pulse occurs too soon, because the scanning frequency is slightly lower, more of the sync pulse width is effective in producing plate current. The result is a greater positive d-c control voltage from the cathode tube to increase the oscillator frequency. If the scanning frequency is higher than the sync frequency, less control voltage is produced, reducing the oscillator frequency. As a result, the control tube keeps the horizontal deflection oscillator locked in phase with the horizontal synchronizing pulses, as the correction voltage continuously corrects the oscillator frequency. This circuit is often called a *pulse-width control system.*

Synchro-Guide circuit. Figure 16 · 10 shows the complete schematic of a blocking oscillator controlled by a d-c control tube, using a twin triode.

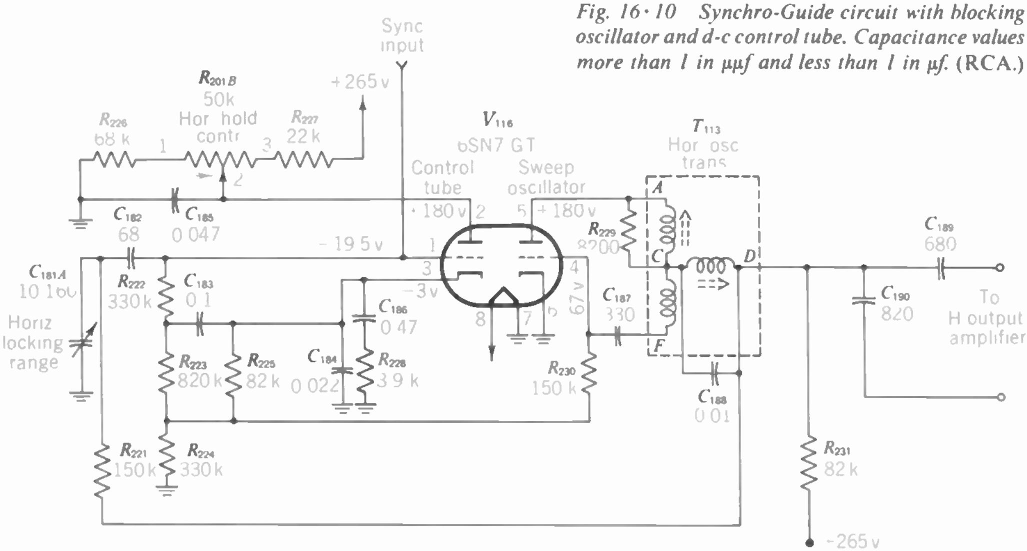


Fig. 16 · 10 Synchro-Guide circuit with blocking oscillator and d-c control tube. Capacitance values more than 1 in μμf and less than 1 in μf. (RCA.)

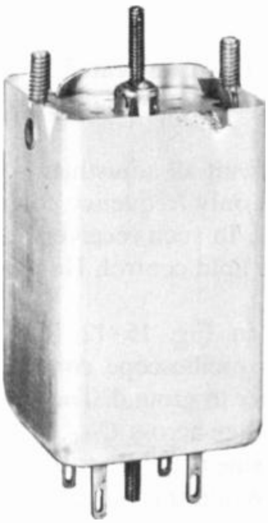


Fig. 16·11 Synchro-Guide transformer in shield can 1½ in. square. Top slug is blocking oscillator frequency adjustment for plate winding AC; bottom is waveform adjustment for stabilizing coil CD. (Merit Coil and Transformer Corp.)

The stage at the right is the blocking oscillator and discharge tube, with C_{190} the sawtooth capacitor. The sawtooth ouput voltage is coupled by C_{189} to the grid-cathode circuit of the horizontal deflection amplifier. Note that the blocking oscillator circuit uses an autotransformer (T_{113}) instead of isolated windings. Current in the plate winding AC induces feedback voltage in the grid winding CF. R_{229} is a damping resistor lowering the transformer Q to prevent double firing of the oscillator. A typical transformer is shown in Fig. 16·11, which includes all the windings in T_{113}.

For the control tube, sawtooth voltage across C_{190} is coupled back from terminal D on the transformer through R_{221} and C_{182}, which blocks B+ voltage from grid pin 1. Horizontal sync voltage is also coupled to the control-tube grid, from the sync separator. The phasing of deflection voltage with respect to sync pulses determines how much plate current flows to develop d-c output voltage across cathode resistors R_{225} and R_{224}. Only the cathode voltage across R_{224} is the d-c control voltage for the oscillator grid to correct its frequency. The voltage across R_{225} is cathode bias, for the desired amount of cathode current without sync input. Negative bias for the grid of the control tube is also obtained from the oscillator grid-leak bias, through R_{230}, R_{223}, and R_{222}. In addition, R_{223} is a decoupling resistor with C_{183} its bypass for oscillator grid signal.

Synchro-Guide adjustments. Notice that the hold control R_{201B} is in the plate circuit of the control tube. Therefore, it can vary the oscillator frequency only when the control tube is operating. The free-running frequency of the oscillator is adjusted by the top slug in the oscillator transformer T_{113}. Varying the inductance of the plate winding AC varies the inductance to change the amount of feedback voltage, which changes the oscillator frequency. The adjustable slug in the coil CD for the sine-wave stabilizing circuit is generally called the *waveform* adjustment.

The *locking-range* control C_{181A} is a variable mica capacitor mounted on the rear apron of the chassis as a screw-driver adjustment. Tightening the mica capacitor increases the capacitance, which reduces the grid-voltage input to the control tube. The amount of sync voltage at the grid determines the range of frequencies through which the control tube can lock in the oscillator to the sync frequency. In general, the locking-range control is adjusted to make the picture pull into sync within the range of frequencies 120 to 180 cps lower than 15,750 cps. This is indicated by two or three diagonal bars sloping down to the left, just before the picture is pulled into sync by varying the hold control.

In simplified versions of the Synchro-Guide circuit all adjustments for the control tube have been eliminated. Then the only frequency control for the oscillator is the slug in plate winding *AC*. In such receivers, this adjustment is at the front panel to be used as the hold control. However, the waveform adjustment is still necessary.

The correct waveform adjustment is shown in Fig. 16·12. This is obtained with the low-capacitance probe of an oscilloscope connected from terminal *C* of the Synchro-Guide transformer to ground. The waveform consists of the sine wave plus sawtooth voltage across C_{190}. Adjust the waveform slug to put the broad peak of the sine wave *S* at the same level as the sharp peak *P* of sawtooth voltage. While this adjustment is being made, the picture must be kept in sync so that the oscillator frequency is at 15,750 cps.

When peak *S* is too low, immunity to noise pulses is reduced. Most important, when *S* is too high the positive peak can trigger the oscillator grid, causing double firing of the oscillator. Then the width and brightness of the raster collapse intermittently because of insufficient horizontal drive for the output stage. Sometimes a receiver has this trouble because of incorrect waveform adjustment in the Synchro-Guide circuit. An approximate adjustment can be made by turning the waveform slug into the sine-wave coil, until the collapsing-width trouble starts; then back off the slug a few turns.

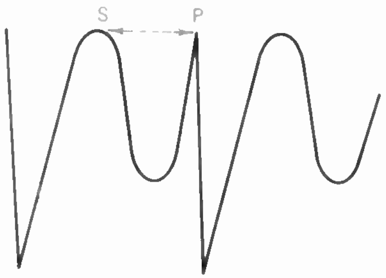
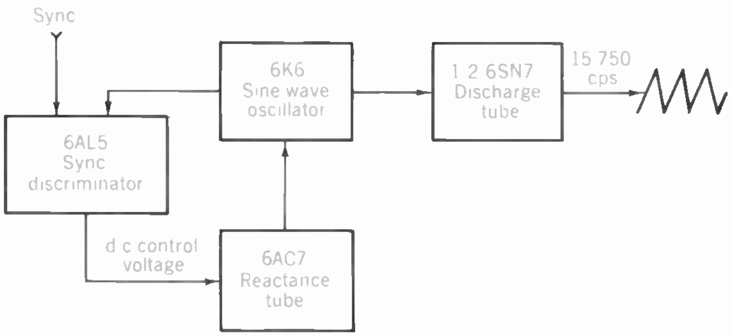


Fig. 16·12 Correct waveform adjustment with oscilloscope at terminal C of Synchro-Guide transformer. Sine-wave peak S at same level as oscillator peak P.

Fig. 16·13 Block diagram of Synchro-Lock AFC circuit for horizontal oscillator.



Synchro-Phase circuit. This is similar to Synchro-Guide but a dual-diode discriminator is used for comparing the sync and sawtooth voltages. Then the d-c output of the discriminator varies the grid voltage of the triode control tube, combined with a triode blocking oscillator, as in the Synchro-Guide circuit. The advantage of the extra discriminator stage is that the control voltage is more independent of noise pulses.

16·6 Sine-wave oscillator with reactance tube (Synchro-Lock)

This circuit was the first developed for horizontal AFC but is seldom used now because of its relatively elaborate arrangement. As shown in Fig. 16·13, the main features are:

1. Instead of a blocking oscillator or multivibrator, the horizontal deflection generator uses a sine-wave Hartley oscillator with a resonant circuit tuned to 15,750 cps. The oscillator output drives a separate discharge tube, which produces sawtooth voltage at the oscillator frequency. This sawtooth voltage is coupled to the grid of the horizontal deflection amplifier.
2. The frequency of the sine-wave oscillator is controlled by a 6AC7 reactance tube[1] in shunt with the oscillator tuned circuit. The reactance tube is necessary because the frequency of a tuned oscillator cannot be controlled effectively by varying its d-c grid voltage. Instead, the d-c control voltage is applied to a reactance tube, which controls the reactance across the tuned circuit of the oscillator.
3. The 6AL5 sync discriminator compares the oscillator frequency with the horizontal sync input. A sync discriminator transformer couples sine-wave voltage from the oscillator to the 6AL5 plates in push-pull. The sync voltage is capacitively coupled to the two plates in parallel. D-c output voltage indicating the difference between the two input frequencies is taken from the cathode load resistors. This d-c control voltage from the discriminator is connected to the grid of the reactance tube, serving as a control tube for the oscillator. As its reactance is varied by the d-c control voltage, the reactance tube corrects the oscillator frequency by changing the reactance across the tuned circuit.

The tube types indicated in Fig. 16·13 are generally used in the Synchro-Lock circuit. The most common trouble is a weak 6AC7, causing weak horizontal hold. Two slugs are provided in the sync discriminator transformer for adjustments. The top slug is the frequency adjustment, which is set to sync the picture. The bottom slug is a phase adjustment that is necessary because of the sine-wave oscillator. In the AFC circuits described previously, the horizontal flyback must occur during the sync and blanking time when the oscillator is locked in. However, the sine-wave oscillator can be locked in by the reactance tube at 15,750 cps but incorrect phase with respect to blanking. Then the horizontal blanking is in the

[1] The reactance-tube circuit is basically the same as those shown for frequency modulation in Sec. 22·4.

picture, as shown in Fig. 16 · 14. To correct this, the phase adjustment is varied to move the blanking bar to the right and out of the picture.

16 · 7 Hold-in range and pull-in range

With AFC for the horizontal oscillator, the picture generally stays in sync through one-half rotation, or more, of the hold control. The d-c correction voltage changes to hold the oscillator at 15,750 cps. When the oscillator frequency is too far off 15,750 cps, though, the picture tears into diagonal segments as shown in Fig. 15 · 2. As an example of the hold-in range, suppose that hold is lost with the oscillator frequency at 16,350 cps at one end of the hold control or 15,150 cps at the other end. Since these frequencies are 600 cps above and below 15,750 cps, the hold-in range equals $\pm$600 cps. When the picture tears apart 600 cps off 15,750 cps, there will be 10 diagonal bars.

As an example of pull-in, assume the picture is out of sync with the oscillator below 15,750 cps. Sync has been lost temporarily when changing channels. To pull the picture into sync, the horizontal hold control is varied to raise the oscillator frequency toward 15,750 cps. At 15,450 cps, which is 300 cps below 15,750 cps, the picture locks in. This pull-in occurs from five diagonal bars. Similarly, at the high-frequency end, the AFC circuit can pull the oscillator into sync from the frequency of 16,050 cps. The pull-in range then is $\pm$300 cps.

The pull-in range is always less than the hold-in range. This means the picture may hold in sync when tuned to a station but tear apart when the channel is changed. Therefore, the horizontal hold control should be set at the middle of its hold-in range, so that the oscillator can be pulled into sync easily when changing channels. For a wide hold-in range and pull-in range, the AFC circuit must have enough sync input to produce the required amount of d-c control voltage.

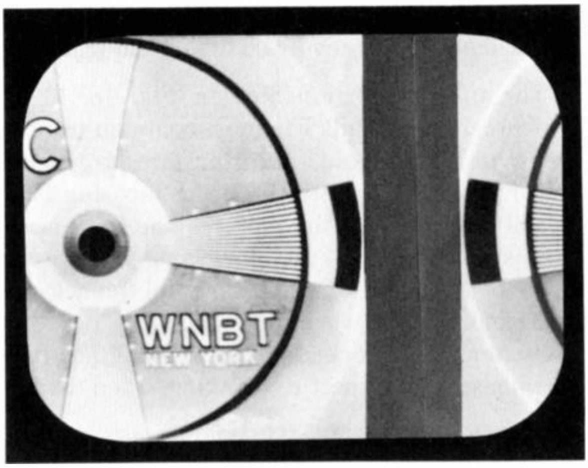


Fig. 16 · 14 Blanking bar in the picture, caused by incorrect phasing of horizontal flyback with respect to blanking. (RCA.)

The cathode-coupled multivibrator requires less d-c control voltage than a blocking oscillator. The required values depend on the amount of oscillator grid-leak bias. For typical circuits, the deflection oscillator can change its frequency by approximately 100 cps per volt for the multivibrator or 50 cps per volt for a blocking oscillator. As an example, -4 volts corrects the multivibrator by 400 cps, increasing the frequency from 15,350 to 15,750 cps. With a blocking oscillator, $+8$ volts would be needed for the same frequency change. Note the opposite polarities required to raise the oscillator frequency.

16 · 8 Filtering the d-c control voltage

The *RC* time constant of the integrating filter determines how fast the d-c control voltage can change its amplitude to correct the oscillator frequency. A time constant much longer than 63.5 μsec is needed for the shunt capacitor to bypass the horizontal sync and sawtooth components in the control circuit, while filtering out noise pulses. However, the oscillator should be able to pull into sync within a fraction of a second when sync is temporarily lost between channels. Also, if the time constant is too long, the d-c control voltage may be affected by vertical sync, causing bend at the top of the picture. A typical value for the AFC filter time constant is approximately 0.005 sec. This time includes about 80 horizontal scanning lines.

Hunting in the AFC circuit. The shunt capacitor in the filter introduces time delay in the effect of d-c control voltage on oscillator frequency. The reason is that the output filter capacitor takes time to charge to the applied d-c control voltage from the AFC circuit, or discharge down to a lower voltage. As a result, the filtered voltage is still correcting the oscillator after it has been pulled in to the correct frequency. Each succeeding step of overcorrection is less until, finally, the oscillator operates at 15,750 cps. This action of the control circuit in varying the oscillator frequency within a smaller and smaller range around the correct frequency is called *hunting.* Excessive hunting in the AFC circuit can cause scalloped edges in the picture, generally called *piecrust* or *geartooth* effect.

In order to prevent hunting, a double-section filter is commonly used for the d-c control voltage, as shown in Fig. 16 · 15. The R_1C_1 time constant of 0.005 sec is long enough to filter out noise and horizontal sync or deflection voltages. The relatively low resistance of R_2 serves as a damping

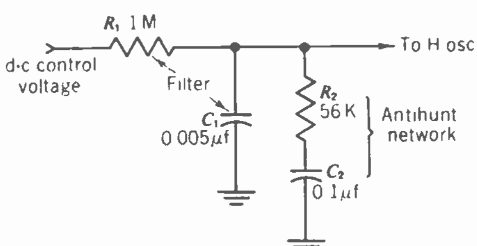


Fig. 16 · 15 AFC filter with antihunt network R_2C_2.

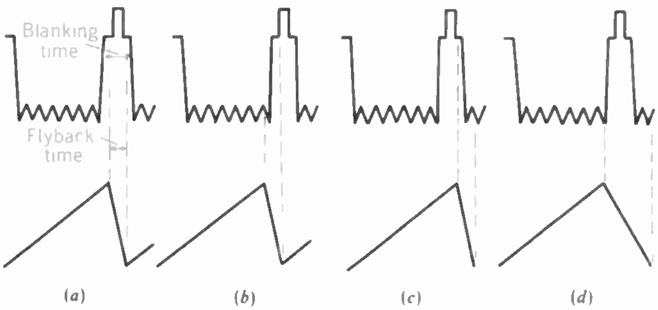


Fig. 16·16 Phasing of horizontal flyback and blanking. (a) Flyback within blanking time. (b) Flyback starts too soon. (c) Flyback starts too late. (d) Flyback time too long.

resistor across C_1, making the output voltage more resistive and less capacitive to reduce time delay in the control voltage. C_2 blocks the d-c control voltage from shorting to chassis ground. The capacitance of C_2 is large to provide reactance low enough to put R_2 in parallel with C_1 when the d-c control voltage is changing. As a result, the R_2C_2 network eliminates hunting in the AFC circuit. In general, the hunting is corrected by inserting either series capacitance or shunt resistance to reduce the time delay associated with lagging d-c control voltage across the shunt filter capacitor.

16·9 Phasing between horizontal blanking and flyback

While the deflection circuits are scanning the raster, the composite video signal is varying the intensity of the electron scanning beam. The blanking on the kinescope screen has the timing of the sync and blanking pulses in the transmitted signal, but the flyback is determined by the deflection circuits in the receiver. In a triggered system, the synchronized flyback starts during blanking time automatically because each sync pulse begins the retrace. With automatic frequency control, however, the horizontal oscillator

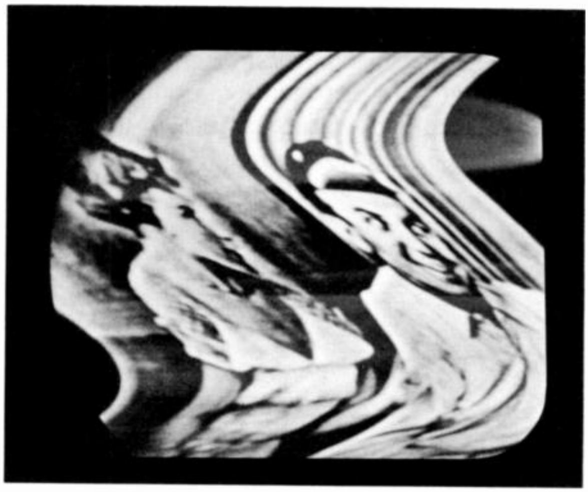


Fig. 16·17 Bend in picture caused by 60-cycle hum in horizontal sync. Note white spear at top right. (RCA.)

produces scanning independently of individual synchronizing pulses. In fact, horizontal retrace can start with the front porch, allowing more time for flyback within blanking.

In Fig. 16 · 16 normal phasing of the flyback within blanking time is shown in *a*. It should be noted that a small change of phasing can put more or less blanking at the left and right edges of the raster. This is why the horizontal centering of the picture usually shifts with respect to the raster when the hold control is varied.

The other examples in Fig. 16 · 16 illustrate incorrect phasing between horizontal flyback and blanking. Since the AFC circuit can lock in the oscillator at 15,750 cps but with retraces not completely within blanking, some picture information may be reproduced during flyback time. In *b*, horizontal flyback starts just before blanking. Then some of the picture information that should be at the extreme right side of the trace in the picture is reproduced during the flyback to the left. With picture information reproduced during both retrace and trace at the right, this side can appear folded over or under itself, usually brighter than normal. When the horizontal flyback starts too late after blanking, as in *c*, the retrace cannot be completed before picture information starts for the trace at the left side of the next line. Then the left side of the picture appears folded. If the horizontal retrace time is too long, as in *d*, the flyback can start before blanking but still continue after blanking. Then both the left and right sides of the picture may be folded.

Whenever the flyback occurs during picture information time there may be a light haze in the background, pointing outward from either side of the picture, like a big spear. This is white picture information spread out by the fast flyback, instead of being reproduced correctly during trace time. The picture usually has this effect just before horizontal hold is lost. As an example, note the white haze in Fig. 16 · 17, extending out from the right side of the picture near the top.

SUMMARY

1. Automatic frequency control is used to lock in the horizontal deflection oscillator at 15,750 cps. The AFC circuit compares the oscillator frequency with the sync frequency to produce d-c output voltage proportional to their difference. After being filtered to eliminate noise, the d-c control voltage corrects the oscillator frequency. The AFC circuit is the horizontal sync section of the receiver.
2. Positive d-c control voltage at the grid of a blocking oscillator raises its frequency. The correction sensitivity is approximately 50 cps change in oscillator frequency for 1 volt change in grid bias.
3. Negative d-c voltage at the synchronizing grid of a cathode-coupled multivibrator raises its frequency. The correction sensitivity is approximately 100 cps change in oscillator frequency for 1 volt change in grid bias.
4. With a dual-diode sync discriminator for the horizontal AFC circuit, each diode has sync input and the sawtooth voltage that indicates oscillator frequency. The sync pulses ride up or down the flyback part of the sawtooth voltage, as the oscillator frequency changes. At 15,750 cps both diodes have equal peak voltage input, producing equal d-c output voltages

that cancel for zero d-c control voltage. Above or below 15,750 cps, one diode produces more output than the other.

5. In the Synchro-Guide AFC circuit, a dual triode is used, with one section a blocking oscillator and the other a d-c control tube. D-c control voltage is taken from the cathode of the control tube to the grid of the blocking oscillator.
6. In the Synchro-Lock circuit, a sine-wave Hartley oscillator is used for the horizontal deflection generator. Its frequency is varied by a reactance tube controlled by the sync discriminator.
7. The time constant for the AFC filter is about 0.005 sec to filter noise but allow fast lock-in when changing channels.
8. Overcorrection of the oscillator frequency can cause hunting in the AFC circuit. In the picture, the result is *piecrust* or *geartooth* effect. A series RC combination is usually connected across the AFC filter capacitor as an antihunt network.
9. With horizontal AFC, some of the picture information may be reproduced during horizontal retrace time. This effect can cause a white haze in the background corresponding to white picture information spread out by the fast flyback.

SELF-EXAMINATION *(Answers at back of book.)*

Choose (*a*), (*b*), (*c*), or (*d*).

1. The advantage of horizontal AFC compared with triggered sync is: (*a*) faster pull-in; (*b*) noise can be filtered from the d-c control voltage; (*c*) shorter blanking time for flyback; (*d*) automatically corrects phasing between flyback and blanking.
2. A blocking oscillator is at 15,700 cps. The required correction voltage to raise the frequency 50 cps is: (*a*) -1 volt; (*b*) $+1$ volt; (*c*) -50 volts; (*d*) $+50$ volts.
3. For the same example as in Question 2, a cathode-coupled multivibrator requires: (*a*) -0.5 volt; (*b*) $+0.5$ volt; (*c*) -50 volts; (*d*) $+50$ volts.
4. A sync discriminator circuit generally uses a: (*a*) twin triode; (*b*) 5U4 full-wave power rectifier; (*c*) single half-wave power rectifier; (*d*) semiconductor dual-diode unit.
5. In the Synchro-Guide AFC circuit: (*a*) The d-c control tube corrects a multivibrator oscillator. (*b*) A sync discriminator controls a blocking oscillator. (*c*) Cathode voltage of control tube controls a blocking oscillator. (*d*) A reactance tube controls a multivibrator oscillator.
6. In the Synchro-Lock AFC circuit a: (*a*) sync discriminator controls a reactance tube; (*b*) reactance tube controls a multivibrator oscillator; (*c*) reactance tube controls a blocking oscillator; (*d*) sine-wave Hartley oscillator is controlled by a multivibrator.
7. In the sync discriminator circuit of Fig. 16·3, sawtooth voltage input to the diodes is developed by: (*a*) C_1; (*b*) C_2; (*c*) C_3; (*d*) R_5.
8. In the single-ended sync discriminator circuit in Fig. 16·7: (*a*) The diodes have push-pull sync input voltage. (*b*) Sawtooth voltage is applied to the diodes in parallel. (*c*) The diodes have opposite polarities of sawtooth voltage cathode-to-plate. (*d*) Sync input is applied to D_1 but not D_2.
9. In Fig. 16·7, sync voltage is coupled to both diodes by: (*a*) C_1; (*b*) C_2; (*c*) C_3; (*d*) C_4.
10. Hunting in the AFC circuit: (*a*) causes a white haze in the picture; (*b*) results from time delay in the AFC filter; (*c*) makes the horizontal blanking time too long; (*d*) makes the horizontal flyback time too short.

ESSAY QUESTIONS

1. Give two advantages and one disadvantage of horizontal AFC.
2. Describe briefly the difference between triggered sync and AFC.
3. List the stages, including horizontal oscillator, for the following AFC circuits, giving the function of each stage: (*a*) Synchro-Lock; (*b*) Synchro-guide; (*c*) push-pull sync discriminator; (*d*) single-ended sync discriminator.

4. Usually, what part of the receiver operates for horizontal sync alone? For vertical sync alone?
5. The picture is in diagonal segments in a receiver having the circuit in Fig. 16·5. Varying the horizontal hold control produces a complete picture but it slips horizontally. Vertical hold is normal. Name three components that can be the cause of the trouble.
6. When horizontal blanking starts in the composite video signal at the kinescope grid, where should the electron scanning beam be on the kinescope screen?
7. (*a*) What is meant by hunting in the AFC circuit? (*b*) What is the effect in the picture? (*c*) What components form the antihunt network in Fig. 16·10?
8. For the push-pull sync discriminator circuit in Fig. 16·3, give the function of each component.
9. For the single-ended sync discriminator circuit in Fig. 16·7*a*, give the function of each component.
10. Redraw the waveshapes in Fig. 16·4, as combined cathode-plate voltages for each diode, similar to Fig. 16·8.
11. Redraw the filter for d-c control voltage in Fig. 16·7*a*, adding an antihunt network.
12. Give the effect on raster and picture for the following two troubles in Fig. 16·5: (*a*) R_{432} open; (*b*) R_{428} open.
13. Give the effect on raster and picture for the following two troubles in Fig. 16·10: (*a*) R_{226} open; (*b*) R_{231} open.

PROBLEMS *(Answers to odd-numbered problems at back of book.)*

1. How much d-c control voltage is necessary to correct a blocking oscillator from 15,675 to 15,750 cps?
2. Referring to Fig. 16·5, make a table listing values of d-c control voltage from 0 to ± 5 volts, in steps of ½ volt, with the corresponding oscillator frequency.
3. (*a*) Calculate the time constant of the $R_{430}C_{417}$ network in Fig. 16·5. (*b*) In how much time will C_{417} charge to one-fifteenth of the applied voltage?
4. Draw to scale the voltage waveshape across C_{417} in Fig. 16·5 for two cycles. Assume the flyback pulse applied has a positive peak amplitude of 150 volts and a pulse width of 8 μsec.
5. A picture is out of horizontal hold, with six diagonal bars. How far off is the oscillator frequency from 15,750 cps?
6. A receiver with single-ended sync discriminator controlling a cathode-coupled multivibrator has a picture in diagonal bars. List the d-c control voltages needed to correct the following: (*a*) five diagonal bars sloping down to the left; (*b*) five diagonal bars sloping up to the left; (*c*) ten diagonal bars sloping down to the left.
7. Referring to Fig. 16·5, calculate the inductance of L_{401} needed for resonance with C_{418} to produce sine waves at the frequency of 15,000 cps.

Chapter 17 Vertical deflection circuits

Both the vertical and horizontal deflection circuits produce the scanning raster on the kinescope screen, as illustrated in Fig. 17 • 1. For horizontal scanning, the horizontal oscillator drives its amplifier to produce the scanning lines at 15,750 cps. At the same time, the vertical oscillator generates 60-cps deflection voltage to drive the vertical deflection amplifier. As a result, the vertical scanning fills the screen from top to bottom with horizontal lines to form the raster.

The deflection amplifier for magnetic scanning is a transformer-coupled power output stage to provide the amount of sawtooth current required in the deflection coils. The width of the scanning raster depends on the peak-to-peak sawtooth current in the horizontal coils, while the height depends on the sawtooth current amplitude in the vertical coils. Although the basic function is similar for the vertical and horizontal deflection amplifiers, the flyback high voltage and damping in the horizontal output circuit are analyzed in the next chapter. Here we can consider the requirements of vertical output circuits to supply 60-cps sawtooth current in the vertical deflection coils.

17 • 1 Triode vertical output stage

For magnetic scanning, the deflection amplifier is generally a single-ended power output stage, rather than the push-pull voltage amplifier commonly used for electrostatic deflection. The required symmetry in magnetic deflection is provided by the scanning coils in the yoke. Practically all television receivers use magnetic scanning because electrostatic scanning would require too much sawtooth deflection voltage for wide-angle large-screen picture tubes with very high anode voltage. In magnetic scanning, the electron beam is deflected by the magnetic field associated with current in the scanning coils. Therefore, it is the current in the yoke that must have the sawtooth waveshape for linear scanning. The function

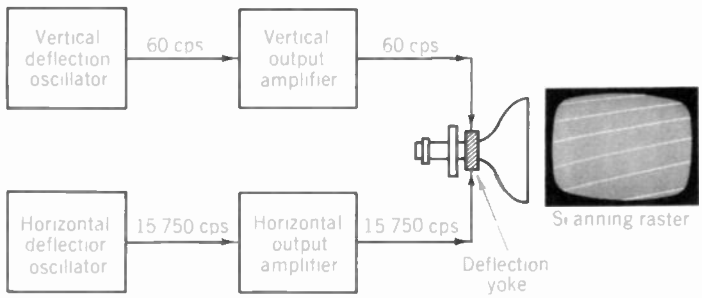


Fig. 17·1 The vertical and horizontal deflection circuits produce the scanning raster on the kinescope screen.

of the deflection amplifier, then, is to supply the required amount of sawtooth output current. The frequency of the amplifier output is the same as its input from the deflection oscillator.

As shown in Fig. 17·2, the main components of a vertical deflection amplifier stage are the power output tube V_2 and the step-down output transformer T_o. Similar to an audio output transformer, T_o matches the relatively high primary impedance with low current in the plate circuit to the low secondary impedance and high current required for the deflection coils. Actually, the vertical output stage is an audio power amplifier for the 60-cps sawtooth deflecting signal. The vertical output tube is usually a triode. Although beam-power pentodes have greater power sensitivity, requiring less grid-voltage drive for equal power output, triodes provide better linearity in the sawtooth output current. The operation of the vertical output stage in providing 60-cps sawtooth current in the vertical deflection coils can be summarized as follows:

1. The deflection voltage from the oscillator provides a linear rise of grid voltage for the amplifier to produce a linear rise of plate current during

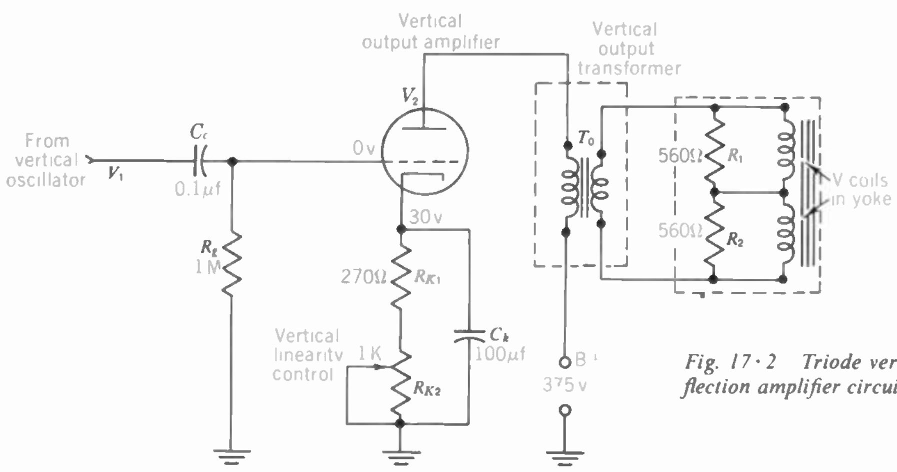


Fig. 17·2 Triode vertical deflection amplifier circuit.

trace time. When the grid voltage drives the amplifier more negative than cutoff for retrace, the plate current drops to zero. The result is sawtooth plate current.

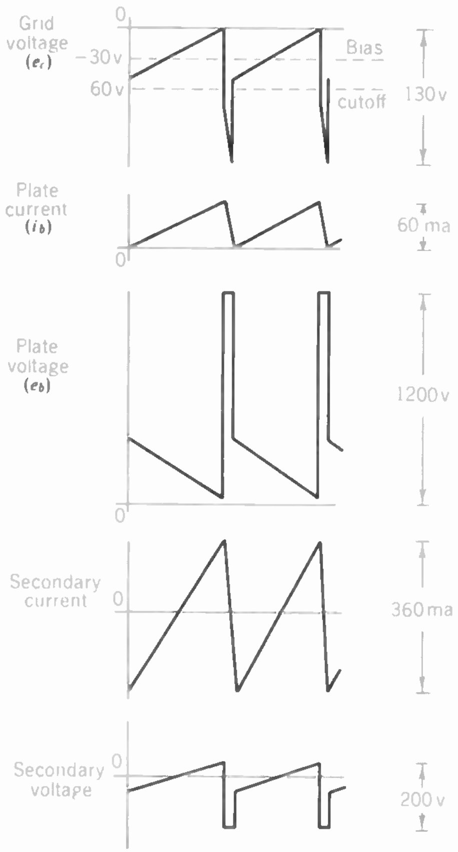
2. The sawtooth plate current of the amplifier flows through the T_o primary winding.
3. The plate current variations in the primary induce voltage in the T_o secondary that produces sawtooth secondary current.
4. Since the T_o secondary winding is across the vertical deflection coils, they also have sawtooth current.

Referring to Fig. 17·2, C_c and R_g couple the 60-cps deflection voltage from the plate of the vertical oscillator to the grid of the amplifier. Cathode bias is provided by R_{k_1} and R_{k_2}, both bypassed by C_k. The variable resistance R_{k_2} adjusts the amplifier bias to control linearity of the sawtooth rise in plate current. Minimum bias is provided by R_{k_1} when the linearity control is at zero resistance. Note that the shunt damping resistors R_1 and R_2 are inside the yoke. Their function is to prevent the oscillations in the horizontal output circuit from inducing appreciable voltage in the vertical deflection coils.

The cathode bias voltage E_k equals $I_k \times R_k$. If we assume R_{k_2} is set for 730 ohms, the total cathode resistance equals 1,000 ohms. With an average plate current of 30 ma, this is also the cathode current for a triode stage without grid current. Therefore, E_k equals $1{,}000 \times 0.03$, or 30 volts.

The cathode bypass capacitor should be large enough to provide an R_kC_k time constant at least five times longer than the period of the a-c sawtooth deflection signal. This idea of a long time constant corresponds to low reactance for sine-wave signal. In both cases, the relatively large capacitance does not develop appreciable a-c voltage across itself. For the case of 60-cycle sawtooth signal, five times the period equals 5/60 sec, or 0.083 sec. The R_kC_k time constant of 0.1 sec, therefore, is long enough for the cathode bypass. Similarly, the R_gC_c time constant of 0.1 sec is long

Fig. 17·3 Waveshapes for vertical output stage in Fig. 17·2. Frequency is 60 cps. Voltage output shown for 6:1 turns ratio in T_1. Polarity of secondary voltage may be as shown or inverted.



enough for the grid coupling circuit. For this case, C_c is in series with R_g. Then C_c should develop little a-c voltage so that most of the signal is across R_g for input to the amplifier. For C_k as a shunt bypass, it should develop little a-c voltage so that the cathode voltage can serve as a steady d-c bias.

Normally there is no grid-leak bias on the vertical output stage. The only bias is the cathode voltage, for minimum amplitude distortion in the sawtooth output current. Grid current would limit the peak plate current, causing poor linearity with a white bar at the bottom of the raster, as shown in Fig. 17 · 14.

Typical voltage and current waveshapes for a triode vertical output stage are shown in Fig. 17 · 3. Note that the negative spike in the trapezoidal grid voltage drives e_c far negative to cut off plate current in the triode, even though its plate voltage rises to a high positive value. This rise in e_b is the self-induced primary voltage in T_o when i_b drops to zero for the retrace. The trapezoidal waveshape of e_b is an amplified and inverted form of e_c. The sawtooth plate current i_b is a fluctuating d-c waveform, but the sawtooth current in the secondary is an a-c waveform, as only the variations of current are coupled by the transformer.

17 · 2 Vertical output transformers

A typical unit is shown in Fig. 17 · 4. The transformer primary provides the impedance needed as the plate load for the vertical output stage. Typical inductance values for L_p are 12 to 40 henrys for the required impedance. As listed in Table 17 · 1, the primary impedance is specified for 60-cps deflection signal superimposed on the rated value of average plate current. The amount of current is important because saturation in the iron core can reduce the effective inductance. Since it is an iron-core transformer with practically unity coupling, the voltage step-down and current step-up are in the same proportion as the turns ratio.

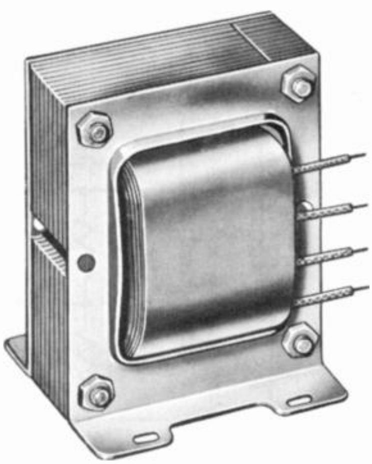


Fig. 17 · 4 Vertical output transformer with isolated secondary. Base 2½ in. square. (Chicago Standard Transformer Corporation.)

Table 17 • 1 Vertical output transformer characteristics

Primary winding			*Secondary resistance, ohms*	*Turns ratio, primary/secondary*	*Remarks*
Impedance, ohms*	*Average plate ma*	*D-c resistance, ohms*			
24,000	15	740	0.8	44:1	Isolated secondary
19,000	13	500	7	10:1	Autotransformer
3,200	40	300	9	6:1	Autotransformer

* For 60-cps signal superimposed on average current listed.

If we assume 12 henrys for L_p and 60 ma peak-to-peak plate current, the values shown in Fig. 17 • 3 can be calculated for the vertical output circuit. The self-induced voltage peak of e_b when i_b drops for retrace is equal to $L\,di/dt$. For a vertical retrace of 600 μsec, a little less than the time of 10 horizontal lines, the vertical flyback voltage is

$$e = L\frac{di}{dt}$$

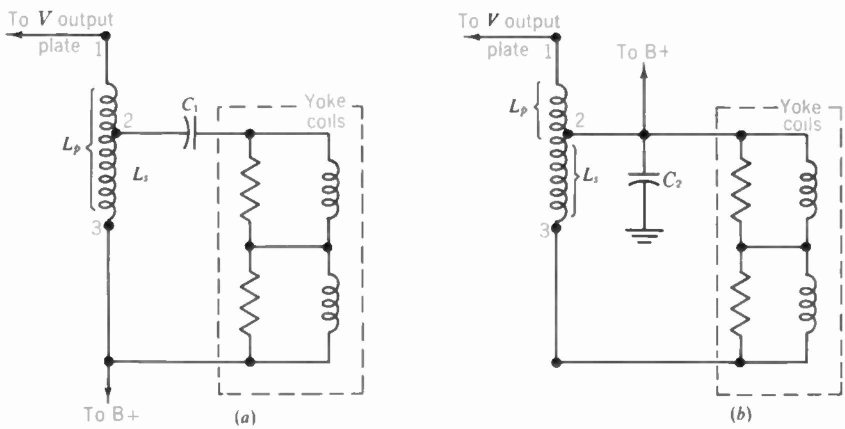
$$e_b = 12 \text{ henrys} \times \frac{0.06 \text{ amp}}{600 \times 10^{-6} \text{ sec}}$$

$$e_b = 1{,}200 \text{ volts}$$

This 1,200 volts in the primary is stepped down to 200 volts in the secondary with a 6:1 turns ratio. Similarly, the primary current of 60 ma corresponds to 360 ma in the secondary. All values are peak-to-peak.

Autotransformer output. Instead of having a separate secondary, an autotransformer uses one tapped winding. Two methods of using an autotransformer in the vertical output circuit are shown in Fig. 17 • 5. Note that

Fig. 17 • 5 Autotransformer connections for vertical output stage. (a) L_p includes all turns with B+ at low end. (b) L_p includes turns between terminals 1 and 2.



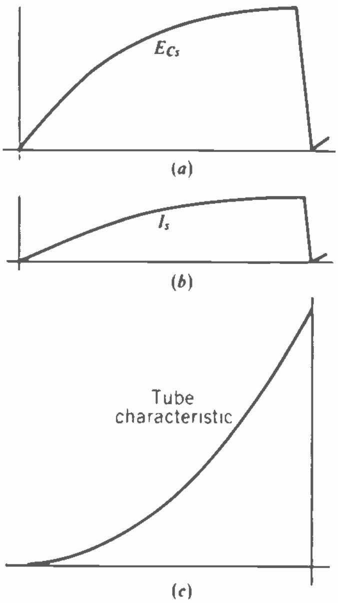


Fig. 17·6 Factors affecting vertical linearity. (a) Voltage across sawtooth capacitor flat because of too short a time constant for charge. (b) Secondary sawtooth current flat because of saturation in output transformer. (c) Opposite curvature of grid-plate operating characteristic of triode output tube.

the primary winding in *a* includes all turns between terminals 1 and 3, while the low-impedance secondary is the tapped winding between 2 and 3. Varying plate current in the primary winding generates self-induced voltage across the secondary winding. The voltage step-down and current step-up are determined by the ratio of primary to secondary turns, the same as with a separate secondary winding. However, the B+ voltage in the primary is not isolated from the secondary. The vertical deflection coils in the yoke then have the B+ voltage to chassis ground. C_1 blocks the average d-c plate current from the coils, though, to prevent a shift in vertical centering.

The autotransformer in *b* has the same windings but L_p includes only the turns between terminals 1 and 2 because B+ voltage is connected to the tap. Then plate current flows through L_p to B+ at terminal 2, without going through L_s. The deflection voltage in the secondary is at terminal 3, with respect to the tap bypassed by C_2 to chassis ground.

17·3 Vertical linearity

The three main factors affecting linearity of the vertical scanning are illustrated in Fig. 17·6. In *a*, the voltage across the sawtooth capacitor E_{c_s} is shown too flat toward the end of the sawtooth wave. This nonlinearity results from using too much of the exponential *RC* charge curve, because of too small a time constant. In *b* the current in the secondary of the output transformer I_s is also too flat toward the end of the sawtooth wave. This distortion can occur because of saturation in the iron core of the transformer with high values of plate current. Then the secondary current does not increase in proportion to the increase in primary current.

Both these nonlinear distortions are opposite from the curvature of the grid-plate operating characteristic for the output tube, as illustrated in *c*. The amplifier expands the amplitude values near zero grid voltage, while compressing amplitudes toward the grid-cutoff voltage. Therefore, the operating characteristic of the vertical output stage compensates for nonlinearity in the sawtooth wave to allow linear scanning. By shifting the bias on the amplifier, we can adjust its operating characteristic for best

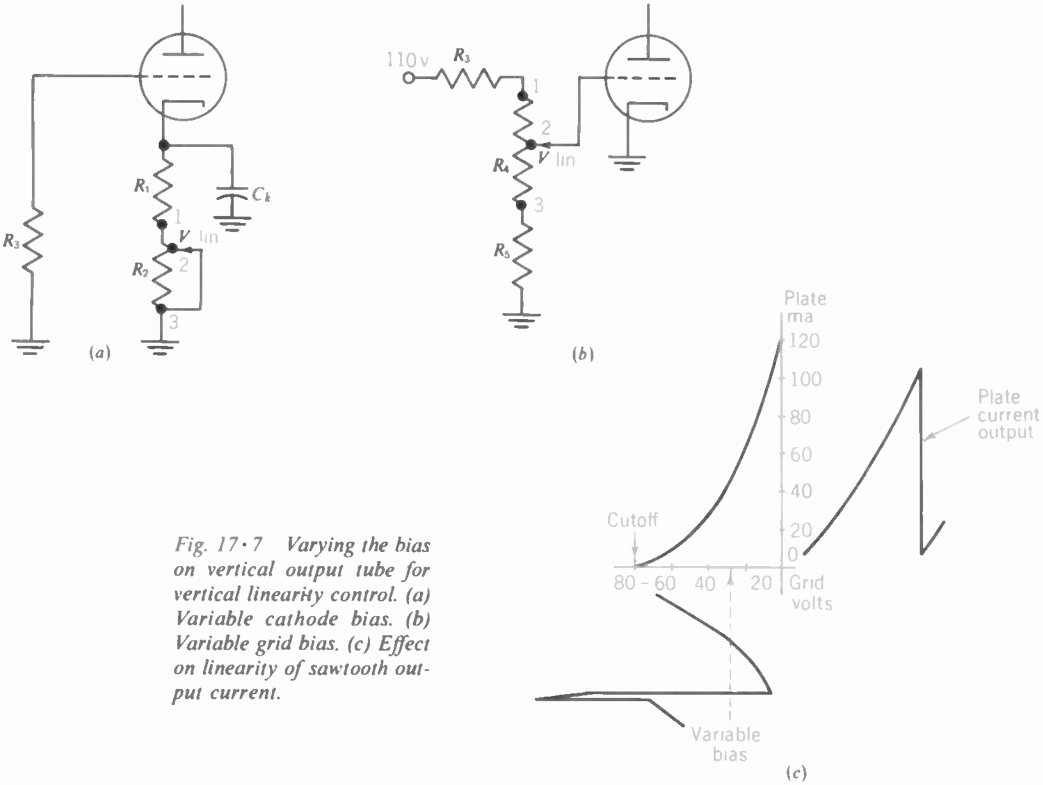


Fig. 17·7 Varying the bias on vertical output tube for vertical linearity control. (a) Variable cathode bias. (b) Variable grid bias. (c) Effect on linearity of sawtooth output current.

linearity. This variable bias control in the vertical deflection amplifier is called *vertical linearity*. The end result required is linear spacing of the scanning lines from top to bottom in the raster.

Two common methods of varying the bias for vertical linearity control are shown in Fig. 17·7*a* and *b*, while *c* illustrates how the variable bias shifts the operating point for the deflection amplifier. In *a*, the cathode bias is varied by R_2. Moving the variable arm closer to terminal 1 reduces the bias, as more of the resistance in R_2 is shorted to chassis ground. Minimum cathode bias is provided by R_1. In *b*, the cathode is grounded but variable grid bias is provided by R_4 from a negative d-c voltage source, which is often the grid-leak bias voltage on the vertical oscillator. Moving R_4 closer to terminal 1 increases the bias. An advantage of grounding the cathode is that the plate-cathode voltage is not reduced by the cathode bias voltage.

The vertical linearity control is usually mounted on the rear of the chassis as a servicing or installation adjustment. Linearity of the vertical scanning can be checked by observing crowding or spreading from top to bottom in the raster and picture. For instance, the vertical scanning current may rise to three-fourths its peak amplitude during the time one-half the field is scanned. Then picture information that should fill the top half of the screen is stretched to three-quarters of the raster. The remaining half of the picture is crowded into a quarter of the raster at the bottom.

This distortion is shown in Fig. 17·8*a*. Note that the finish at the top of the sawtooth wave corresponds to the bottom of the raster, just before vertical flyback. Circles in the picture are egg-shaped, instead of round, being flattened toward the bottom. In a test pattern, the bottom wedge is shorter than the top wedge. With people in the picture, they appear with their heads too long and legs too short. These distortions result because the scanning lines are crowded at the bottom of the raster and spread out at the top. When the white scanning lines are crowded enough to be superimposed one on the other, a white bar can appear across the bottom of the raster. The opposite type of vertical nonlinearity is shown in *b*, with crowding at the top and spreading at the bottom. However, crowding at the bottom is a more common trouble, resulting from insufficient peak plate current in the vertical output stage.

Correct adjustment of vertical linearity is indicated by equal top and bottom wedges in the test pattern. Or, if an audio signal generator is available to produce black and white bars across the kinescope screen, the linearity can be judged by equal vertical spacing between the bars. Most conveniently, the vertical linearity can be set by adjusting for uniform spacing of the scanning lines in the raster. Another convenient way of testing vertical linearity is to check the vertical blanking bar as you watch it drift on the screen. Turn the vertical hold control slightly off frequency to make the picture roll slowly. The black blanking bar across the screen should have the same height at any position on the raster.

Since the gain of the vertical output stage varies when the bias is changed, adjusting the vertical linearity control also changes the height of the raster. Similarly, adjusting the height control changes the linearity. Both controls should be adjusted together. However, it can be helpful to note that usually the vertical linearity control is more effective in adjusting the top of the raster. Adjusting the linearity control for less bias increases height and stretches the top. Increasing height with the size control

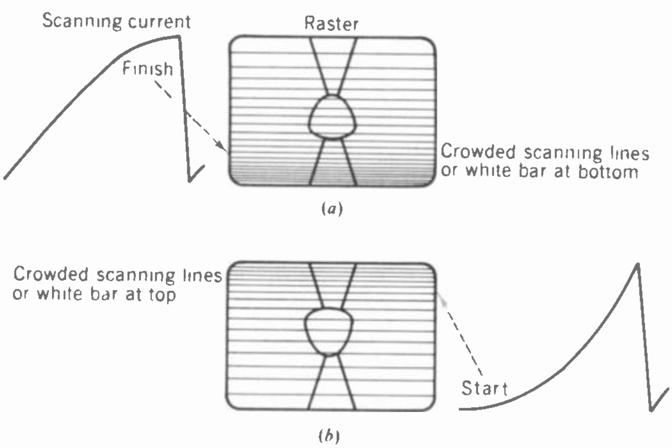


Fig. 17·8 Typical examples of vertical nonlinearity. (a) Flat finish on sawtooth causes crowding at bottom of raster. (b) Flat start on sawtooth causes crowding at top of raster.

stretches the bottom slightly. A common problem is that the height control cannot pull the bottom of the raster down enough. Then good linearity can be obtained only with insufficient height, but the correct height can be produced only with stretching at the top. This trouble is actually insufficient height, rather than nonlinearity. Try a new vertical output tube. Another point to keep in mind is that it may be necessary to readjust the vertical centering to fit the top and bottom halves of the scanning raster equally on the screen.

17 · 4 Internal vertical blanking

In most receivers, the voltage pulse produced during flyback in the vertical output transformer is coupled to the kinescope grid-cathode circuit for the purpose of providing additional blanking during vertical retrace time. The circuit is also called a *retrace suppressor.* This internal vertical blanking is in addition to the blanking voltage present at the kinescope grid as part of the composite video signal. The advantage of using the additional vertical blanking voltage is that the retrace lines produced during vertical flyback do not appear on the kinescope screen for any setting of the brightness control. Actually, the internal vertical blanking is necessary in receivers that do not have the d-c component of video signal at the kinescope. Otherwise, the vertical retrace lines would show when the brightness changed for a darker scene.

Either positive or negative pulses can be used for the internal vertical blanking. A negative pulse coupled to the kinescope grid drives it more negative than cutoff to reduce beam current to zero during the vertical retrace time. Similarly, a positive vertical flyback pulse coupled to the kinescope cathode can cut off beam current. About 50 to 100 volts at the kinescope is required for the internal blanking. After vertical flyback, the grid has its normal video signal to reproduce the picture information.

The two methods of internal vertical blanking with opposite polarities for the flyback pulses are shown in Fig. 17 · 9. In *a*, positive flyback pulses at the plate of the vertical output tube are coupled by R_1C_1 to the kinescope cathode circuit. The flyback pulse has positive polarity because it is the self-induced voltage across the primary produced by a drop in plate current. C_1 is a d-c blocking capacitor, to keep the B+ voltage off the kinescope cathode. R_1 is a decoupling resistor to isolate the video signal circuits from the vertical output stage. In *b*, the vertical flyback pulse is taken from the secondary of the output transformer. The polarity can be either positive or negative, depending on which end is grounded, but the circuit is shown with negative flyback pulses for the kinescope grid. This blanking method is most common as it fits the usual case of cathode drive for video signal at the kinescope. Then the kinescope grid is grounded through R_2, which has the negative flyback voltage for blanking. Note that the secondary returns to chassis ground for 60-cycle signal through the B supply, instead of a direct ground, in order to minimize arcing to the chassis.

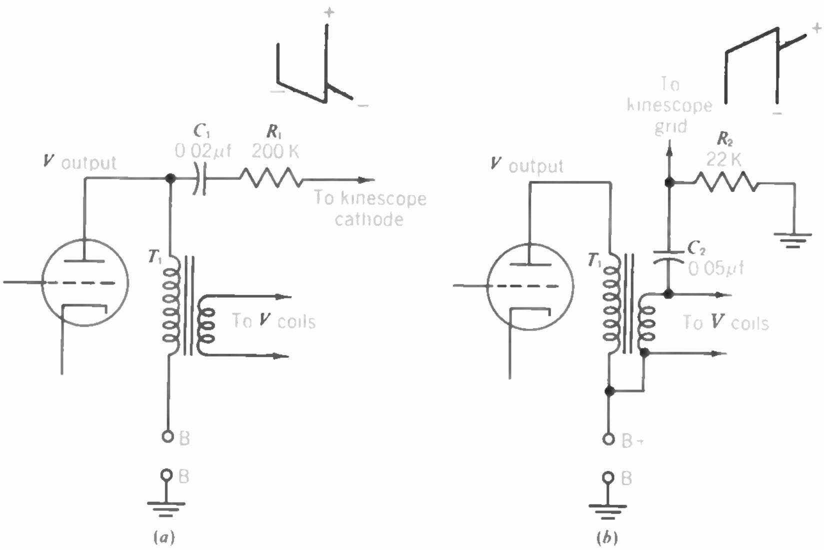


Fig. 17·9 Circuits for internal vertical blanking. (a) Positive flyback pulses from plate of output tube coupled to kinescope cathode. (b) Negative flyback pulses from transformer secondary to kinescope grid.

Another circuit for internal vertical blanking uses the voltage spikes across the peaking resistor in the vertical oscillator. These negative flyback pulses can be coupled to the kinescope grid, but their amplitude is relatively low. Finally, some receivers have an extra winding on the vertical output transformer to provide pulses for internal vertical blanking.

17·5 Vertical deflection circuit with blocking oscillator

Figure 17·10 illustrates how the oscillator and output stages fit together to provide a complete deflection circuit for vertical scanning. Triode V_1 functions as a blocking oscillator and discharge tube. C_2 is the sawtooth capacitor in series with the peaking resistor R_5 to provide trapezoidal grid voltage for the triode output tube V_2. The sawtooth plate current of V_2 is transformer-coupled to the vertical scanning coils in the deflection yoke. For internal vertical blanking C_5 couples negative flyback pulses to the kinescope grid.

Note the thermistor in the yoke. Its resistance decreases with temperature to compensate for increased coil resistance as the yoke heats up. The result is constant raster height.

There are three controls to adjust vertical scanning. The hold control R_2 varies the oscillator grid time constant to set the free frequency slightly below 60 cps so that the sync voltage can lock in the oscillator. Positive sync voltage is applied to the oscillator grid to trigger it into conduction for the flyback. The height control R_4 varies the time constant on charge for C_2, to adjust the sawtooth amplitude. The linearity control R_8 varies the cathode bias on the output tube. Both the height and linearity controls

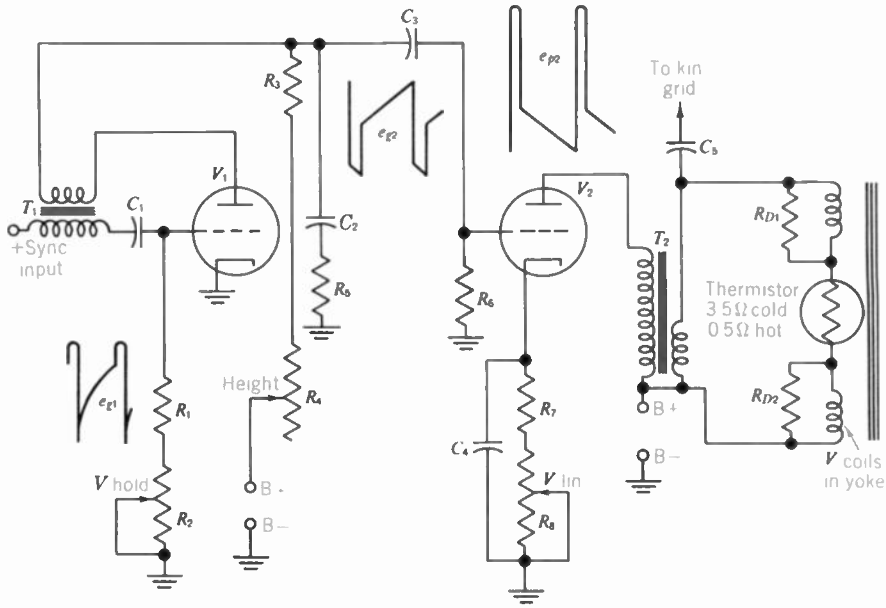


Fig. 17·10 Complete vertical deflection circuit. V_1 is blocking oscillator and discharge tube while V_2 is power output stage.

are adjusted together, with the height control used mainly for the bottom of the raster and the linearity control for the top.

The voltage waveshapes in Fig. 17·10 can be observed with an oscilloscope. When its internal sweep frequency is at 30 cps, there should be two cycles of 60-cps trapezoidal voltage with negative spikes at the plate of V_1 or at the grid of V_2. This amplitude is about 100 volts peak to peak. At the V_2 plate, the trapezoidal voltage across the T_2 primary is inverted with positive spikes of about 1,000 volts. In the secondary of T_2, the trapezoidal voltage has negative spikes with about 150 volts amplitude.

17·6 Combined vertical oscillator and output circuit

In Fig. 17·11, V_1 is the trapezoidal-voltage generator, as before, but it needs feedback from the output stage to function as the vertical oscillator. Note the line feeding plate voltage of V_2 back to the grid of V_1. This circuit is used in practically all television receivers because it is a compact arrangement that eliminates the need for a separate blocking oscillator transformer. Also, no separate oscillator stage is used. The two tubes together can be considered a plate-coupled multivibrator. V_2 conducts during trace time, while V_1 is cut off. Then V_1 conducts and V_2 is cut off during flyback time. Since the sawtooth capacitor with its peaking resistor are in the plate circuit of V_1, this stage functions as the discharge tube. The power output stage is V_2, transformer-coupled to the vertical deflection coils.

Cutoff and conduction in the two tubes. We can start when the discharge tube V_1 is cut off. At this time the sawtooth capacitor C_2 is charging. The

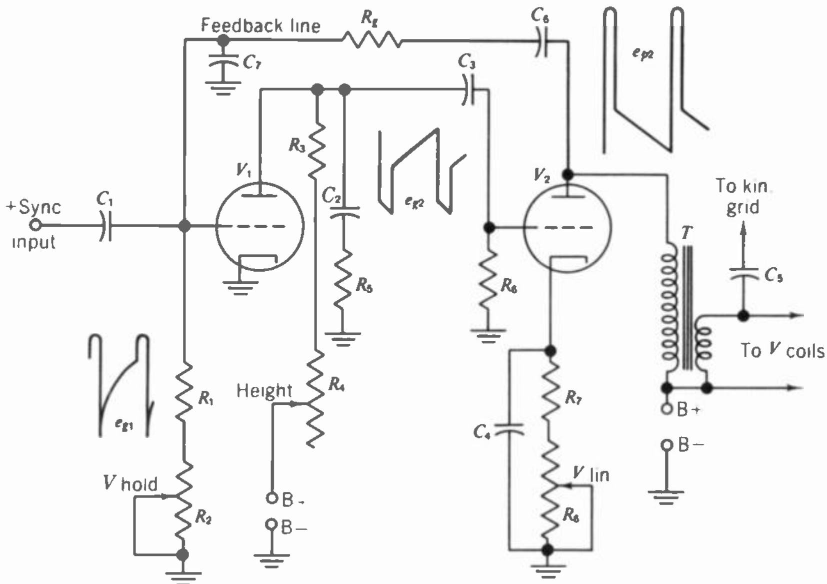


Fig. 17·11 Combined vertical oscillator and output circuit, with feedback from plate of output tube V_2 to grid of discharge tube V_1.

resulting sawtooth voltage at the grid of the output tube V_2 produces sawtooth plate current for the output circuit. V_2 is conducting for the trace because V_1 is cut off, allowing the sawtooth capacitor to charge. This period of time, with the output tube conducting, produces the trace.

How long V_1 is cut off depends on its grid RC time constant. When the grid bias declines to cutoff, V_1 starts conducting. Then the plate voltage of V_1 drops, driving the grid of V_2 negative. This negative grid drive is amplified by the two tubes in the feedback loop to cut off V_2. The period of cutoff for V_2 and conduction in V_1 corresponds to flyback time.

Cutoff in V_2 results in a large positive flyback pulse at its plate that is fed back to the grid of V_1. There are two effects here. The positive grid feedback voltage makes V_1 conduct saturation plate current for discharge of the sawtooth capacitor C_2. At the same time, grid-leak bias is produced for V_1, proportional to the peak positive pulse amplitude. This grid-leak bias is what keeps V_1 cut off after the pulse has gone.

The cutoff time for V_1 depends on how long it takes the grid-leak bias to decline to cutoff, as C_6 discharges through R_g, R_1 and R_2. While C_1 is discharging the bias in the grid circuit of V_1 is being discharged, C_2 is charging in the plate circuit to produce the sawtooth voltage drive for the grid of V_2. During this time the output tube is conducting a linear rise of plate current for the trace.

Conduction in V_1 is what cuts off V_2. With V_1 conducting, C_2 discharges. This is the time when the peaking resistor R_5 produces its negative voltage spike. The spike produced by C_2 discharge combines with the sawtooth rise produced during C_2 charge. The result is trapezoidal voltage output from the V_1 plate as it oscillates between cutoff and conduction. This grid drive for V_2 makes it produce sawtooth plate current, with a sawtooth rise while it conducts and a fast flyback when it is cut off. While conducting, V_2 supplies the sawtooth current amplitude needed for the vertical deflection coils to fill the height of the raster. The on-off conditions for both tubes are summarized in the following table.

V_1 *off = trace*	V_1 *on = flyback*	V_2 *on = trace*	V_2 *off = flyback*
Sawtooth C charges Grid C discharges	Sawtooth C discharges Grid C charges	Sawtooth rise of plate current	Plate current drops to zero

Feedback. C_6 couples the feedback and blocks B+ at the plate of V_2 from the grid of V_1. Bypass capacitor C_7 filters out horizontal flyback pulses from the feedback line. Horizontal pulses can be present at the plate of the vertical output tube because of coupling to the horizontal deflection coils in the yoke. Excessive horizontal pulse amplitude in the vertical deflection circuits can cause poor interlace. R_g provides the desired amount of feedback voltage. Also, it isolates the plate circuit of V_2 from the low resistance in the grid circuit of V_1 when it conducts during retrace.

The RC time constant in the feedback line is critical. This affects the cutoff period of V_2 and the damping of its output circuit during flyback. Both factors must be matched for the desired retrace with good linearity at the start of the sawtooth when the output tube starts to conduct for scanning the top of the raster. Vertical retrace that is slow because of excessive damping can cause crowding. Too fast a retrace because of insufficient damping can cause stretching with white bars across the top of the raster. Combinations of shunt and series RC filter may be used in the feedback line to provide the desired linearity. In some circuits, the linearity control is a variable resistor in the feedback line. The component values in the feedback line also affect the frequency of the oscillator by varying the amount of feedback voltage. In some cases, the vertical hold control is in the cathode circuit of the discharge tube, leaving the grid free for sync input.

Sync input. Either positive sync voltage can be applied to the grid of the discharge tube, or negative sync at the plate. The negative sync at the plate of V_1 is coupled to the V_2 grid, amplified by the conducting output tube, and fed back as positive sync to the V_1 grid.

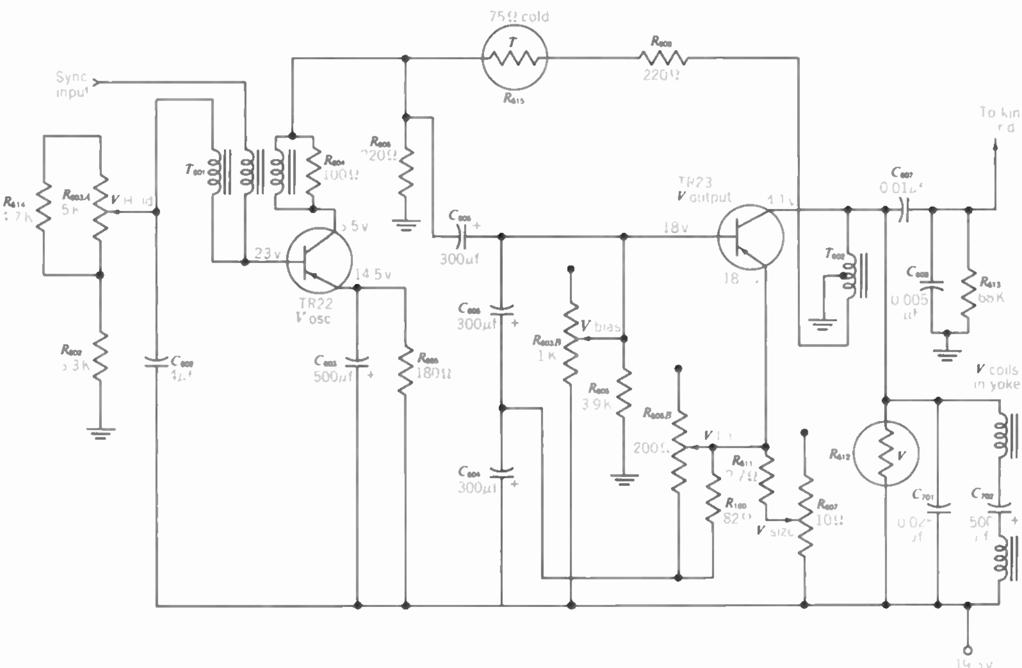
Typical tubes. Usually, a duotriode is used for the vertical deflection circuit, with dissimilar units. The 6EH7, 6EM7, 6DN7, and 6EW7, as examples, feature a medium-mu triode for the discharge tube and a low-mu triode for the output stage with 175 ma rating for maximum peak cathode current. In some circuits a beam-power pentode may be used in the vertical output stage for greater power sensitivity. Then sawtooth grid drive can be used. Also, negative feedback in the output stage is used to improve linearity. The positive flyback pulses at the plate of a pentode vertical output tube are about 40 per cent higher than for a triode.

17·7 Transistorized vertical deflection circuit

Because of their high-current capabilities, transistors are especially useful in deflection circuits for magnetic scanning. However, precautions may be necessary against high-voltage pulses, which can cause breakdown in the transistor. Also nonlinear amplification may require the use of negative feedback.

In Fig. 17·12 two PNP transistors are used in the common-emitter arrangement for the vertical oscillator and output stages. TR22 is in a blocking oscillator circuit, with T_{601} the oscillator transformer providing feedback from collector to base. A third winding couples integrated vertical sync into the oscillator base circuit. The sawtooth voltage output across C_{605} drives the base-emitter circuit of TR23, which supplies the required output current in the vertical deflection coils. Note the negative feedback from one end of T_{602} to the base of TR23, through C_{606}, for the purpose of improving linearity. The thermistor R_{615} reduces its resistance when hot to allow more inverse feedback voltage, stabilizing the transistor against excessive collector current. In the output circuit, the varistor R_{612} has a minimum resistance of 2,500 ohms at 20 d-c volts. At higher voltages, the varistor decreases its resistance to protect the transistor against excessive collector voltage.

Fig. 17·12 Transistorized vertical deflection circuit, with blocking oscillator and choke-coupled output stage. Typical d-c voltages listed. Polarity is indicated for electrolytic capacitors. (Motorola chassis TS-432C.)



Note the d-c voltages for the PNP transistors. The +19.5-volt supply is a common-emitter return for both stages. Then the collector is less positive than the emitter in each stage, providing the negative collector-emitter voltage needed for reverse bias. For TR23, its base at 18 volts has 0.2 volt negative forward bias, with respect to 18.2 volts at the emitter. In the oscillator stage, positive feedback produces base current that charges C_{602} for signal bias. This equals 23 volts minus 19.5 volts, or 3.5 volts positive bias between base and emitter, to keep the oscillator stage cut off between flyback pulses.

Transistor bias polarities. The oscillator is cut off with positive voltage at the base, with respect to emitter, because this polarity is reverse bias for a PNP transistor. To help remember the polarity requirements for transistor bias, the following rule can be helpful: *Voltage opposite from the polarity of the electrode provides reverse bias.* Although reverse bias in the input cuts off the transistor forward current, keep in mind the fact that reverse collector bias is needed for output. As examples, with a PNP transistor in the common-emitter circuit, positive voltage at the N base is reverse bias; also negative voltage at the P collector is reverse bias. For forward bias at the N base, negative voltage is needed. Remember that a PNP schematic shows the emitter arrow for positive hole charges attracted into the N base. With an NPN transistor, all the polarities are opposite. However, the rule still applies that voltage of the same polarity as the electrode is forward bias and opposite voltage is reverse bias.

17·8 Vertical deflection troubles

Any defects in the height of the raster are caused by the vertical deflection circuits. In addition, the internal vertical blanking circuit can cause troubles in kinescope brightness. In this case, the brightness effect changes as the vertical deflection controls are varied.

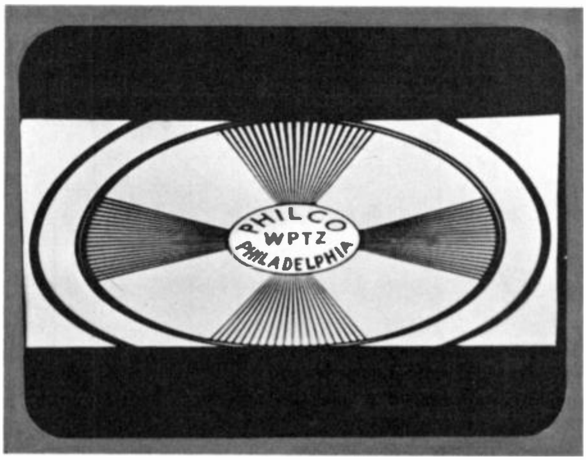


Fig. 17·13 Insufficient height in raster caused by weak vertical output.

Horizontal line only. Just a bright line across the middle of the screen, as in Fig. 6 · 4*b*, shows there is no output from the vertical deflection circuits. The reason for no height can be either no output from the vertical oscillator or trouble in the vertical amplifier and its output circuit. In the plate-coupled multivibrator arrangement, remember that a trouble in the vertical output stage can cause no oscillator output because the feedback loop includes both stages.

Insufficient height. When the raster does not have enough height, the picture appears with black bars at the top and bottom edges, as in Fig. 17 · 13, corresponding to the unused screen area not scanned and therefore not illuminated. This trouble means the current in the vertical deflection coils does not have enough peak-to-peak amplitude to scan the full height of the raster. In many cases, more height can be obtained only with poor linearity and a white bar across the bottom. These effects often indicate a weak vertical output tube.

White bar across bottom of raster. This effect in Fig. 17 · 14 results from insufficient peak plate current in the vertical output tube. Then the end of the sawtooth trace is too flat, compressing lines at the bottom to form the white bar. The cause is generally a weak vertical output tube or insufficient bias. When the grid-voltage peak drives the grid positive to draw grid current, the peak plate current is limited to compress the bottom of the raster. A typical trouble is a leaky coupling capacitor, which adds positive d-c grid voltage that cancels part of the cathode bias. The resulting amplitude distortion of plate current with low bias causes the white bar. In addition, a white bar across the top of the raster indicates amplitude distortion caused by cutoff at the start of the sawtooth rise.

Open sawtooth capacitor. If either the capacitor or its peaking resistor opens, the stray capacitance serves as the vertical sawtooth capacitor. Instead of having no vertical deflection, the oscillator operates with extremely nonlinear output caused by a very small sawtooth capacitor. The

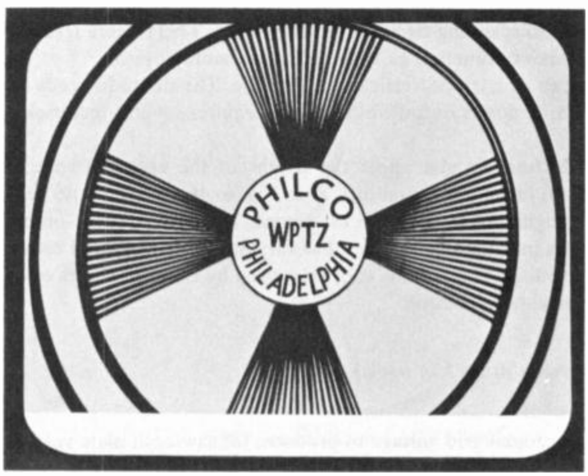


Fig. 17 · 14 White bar across bottom of raster, caused by compression at end of sawtooth wave.

result then is widely spaced scanning lines, crossed by retrace lines, in an effect often called *basketweave.*

Incorrect vertical oscillator frequency. If varying the vertical hold control cannot stop the picture from rolling up or down, even for an instant, this indicates the free frequency cannot be set to 60 cps. Since there is no AFC for the vertical oscillator, frequency troubles are generally caused by an incorrect *RC* time constant for the vertical hold control. Too large a value for R_g or C_c makes the oscillator frequency too low. Then the picture rolls upward, with possibly two or three frames top to bottom separated by black blanking bars. Too small a value for R_g or C_c makes the oscillator frequency too high. Then the picture rolls downward. Also, the bottom of the picture may be superimposed on the top. Furthermore, the scanning lines are spread apart much more than normal, giving the impression of poor linearity.

SUMMARY

1. The vertical deflection circuit consists of an oscillator stage driving the power output stage transformer-coupled to the vertical deflection coils in the yoke.
2. For a triode output stage, its grid drive from the vertical oscillator is trapezoidal voltage consisting of a positive-going sawtooth wave for trace plus a sharp negative spike for flyback. The spike is needed to cut off the triode. The plate voltage waveshape is an inverted trapezoid about 1,000 volts peak to peak.
3. The output transformer couples the sawtooth plate current in the primary to the vertical deflection coils connected across the secondary. The iron-core transformer has a step-down turns ratio to supply the required secondary current of about 0.5 amp peak to peak.
4. The vertical linearity control varies the bias on the output stage to compensate for nonlinear grid drive and saturation in the output transformer. Both the height control in the oscillator stage and the linearity control in the amplifier are adjusted together to fill the raster with uniformly spaced scanning lines.
5. For internal vertical blanking, the vertical flyback pulses are coupled to the kinescope. Positive pulses applied to the cathode or negative pulses at the grid cut off beam current during vertical retrace so that the vertical retrace lines are not visible.
6. A basic circuit for vertical scanning consists of a blocking oscillator driving a triode output stage, as in Fig. 17 · 10. The same arrangement using transistors is shown in Fig. 17 · 12.
7. The most commonly used vertical scanning circuit is the one in Fig. 17 · 11. Here feedback from V_2 to V_1 enables both tubes to function as a plate-coupled multivibrator.
8. A pentode may be used instead of a triode vertical output tube. The pentode needs less grid-driving voltage for the same power output, but usually requires inverse feedback to improve linearity.
9. Troubles in the vertical deflection circuits affect the height of the raster. Two common amplitude troubles are: (*a*) just a horizontal line, as in Fig. 6 · 4*b*, caused by no vertical output; (*b*) insufficient height, as in Fig. 17 · 13, caused by weak vertical output. Incorrect oscillator frequency is indicated by the fact that varying the vertical hold control cannot stop the picture from rolling. This trouble is often caused by changed values of the *RC* time constant for the vertical hold control.

SELF-EXAMINATION *(Answers at back of book.)*

Choose (*a*), (*b*), (*c*), or (*d*).

1. A triode output tube has trapezoidal grid voltage to produce: (*a*) sawtooth plate voltage; (*b*) trapezoidal plate current; (*c*) sawtooth grid current; (*d*) sawtooth plate current.

2. While the raster is being scanned vertically to produce the required height: (*a*) the output tube is on and discharge tube off; (*b*) the output tube is off and discharge tube on; (*c*) both tubes are on; (*d*) both tubes are off.
3. With 40 ma peak to peak I_p and a turns ratio of 1:24, I_s equals: (*a*) 24 ma; (*b*) 40 ma; (*c*) 960 ma; (*d*) 24,000 ma.
4. Current dropping from 20 ma to 0 in 400 μsec through a 25-henry inductance produces self-induced voltage equal to: (*a*) 400; (*b*) 1,250; (*c*) 15,000; (*d*) 20,000.
5. The top of the picture is stretched with too much raster height. To correct this: (*a*) vary vertical hold control; (*b*) reduce height with vertical linearity control; (*c*) increase height with size control; (*d*) replace vertical oscillator tube.
6. Flyback pulses at plate of vertical output tube are for internal blanking at the kinescope: (*a*) cathode; (*b*) control grid; (*c*) screen grid; (*d*) anode.
7. The voltage waveshape across R_5 in Fig. 17·11 is: (*a*) sawtooth; (*b*) trapezoid; (*c*) negative spikes; (*d*) integrated vertical sync.
8. Increasing the resistance of R_2 in Fig. 17·11: (*a*) increases oscillator frequency; (*b*) decreases oscillator frequency; (*c*) reduces height of raster; (*d*) decreases feedback voltage.
9. The forward bias on TR23 in Fig. 17·12 equals: (*a*) 0.2 volt; (*b*) 18 volts; (*c*) 18.2 volts; (*d*) 19.5 volts.
10. The output tube plate voltage is maximum positive: (*a*) at the end of trace; (*b*) during retrace; (*c*) at the start of trace; (*d*) when plate current is maximum.

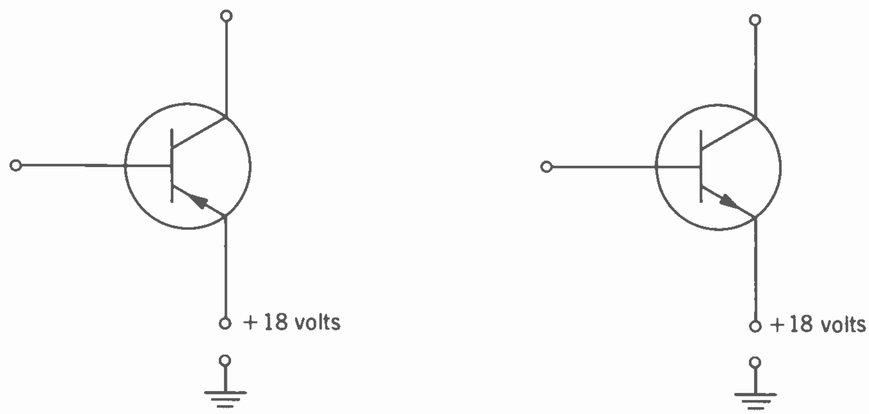
ESSAY QUESTIONS

1. Describe briefly how the scanning raster is produced. Which stages determine height? Which stages determine width?
2. Why can the deflection circuits produce the raster with or without sync?
3. How is the vertical oscillator synchronized at 60 cps with: (*a*) positive sync; (*b*) negative sync?
4. Why does the plate current of the vertical output tube have the sawtooth waveshape? What makes the plate current drop to zero for retrace?
5. Give two requirements of the vertical output transformer.
6. List the three controls for vertical deflection, giving the function of each, and describe briefly how to adjust them.
7. Draw the schematic diagram of a triode output stage, with autotransformer coupling to the vertical deflection coils. Include internal vertical blanking for kinescope. Give typical values of all components. Show voltage waveshapes with typical amplitudes at grid, plate, and across yoke coils.
8. Give two component troubles that can cause a white bar at the bottom of the raster, with insufficient height.
9. Give two component troubles that can make the oscillator frequency too high.
10. Give two component troubles that can cause no vertical scanning, with just a white bar across center of raster.
11. Referring to Fig. 17·11, list function and typical value for each of the components in the complete circuit.
12. What is the function of a thermistor? A varistor?
13. Referring to Fig. 17·12, list the voltages at base and collector, giving polarity, with respect to emitter, for TR22 and TR23.
14. Make a table summarizing the voltage polarities required for forward bias or for reverse bias in the input and reverse bias in the output, for a PNP transistor in a common-emitter circuit. Do this also for an NPN transistor in a common-base circuit.

PROBLEMS *(Answers to odd-numbered problems at back of book.)*

1. How much cathode bias voltage is produced by a cathode resistance of 4,000 ohms with 8 ma average current?

2. How much cathode resistance is needed for cathode bias of: (*a*) 30 volts with 12 ma cathode current; (*b*) 20 volts with 80 ma cathode current; (*c*) 45 volts with 50 ma cathode current; (*d*) 18 volts with 60 ma cathode current?
3. A coil has an impedance of 5,000 ohms for 60-cps sine-wave current, with resistance of 1,000 ohms. Calculate: (*a*) the inductive reactance in ohms; (*b*) the inductance in henrys.
4. Referring to Table 17 · 1, for the three vertical output transformers listed: (*a*) Calculate each primary inductance, assuming the impedances listed are for 60-cps sine waves. (*b*) Calculate the self-induced voltage across each of these inductances when the primary current varies from zero to twice the average plate current listed. Assume the current drops in 600 μsec vertical flyback time.
5. Two resistors with one twice the resistance of the other are connected in series across a 100-volt source. How much is the voltage across the smaller resistance?
6. For the three vertical output transformers listed in Table 17 · 1, calculate the d-c voltage drop across each primary winding.
7. For each of the two transistors below, in a common-emitter circuit, indicate the d-c electrode voltages for 0.3 volt forward bias in the input and 18 volts reverse bias in the output.



(*a*) (*b*)

Chapter 18 Horizontal deflection circuits

The horizontal output tube is a beam-power amplifier capable of high current output. With 15,750-cps grid-driving voltage from the horizontal oscillator, the plate current of the output tube is transformer-coupled to the horizontal deflection coils for scanning the horizontal lines. Approximately 1 amp peak-to-peak sawtooth current in the scanning coils is required to fill the width of the raster. The horizontal output circuit is primarily inductive because of the large self-induced voltage produced with 15,750-cps sawtooth current. For this reason, the inductive output circuit can generate flyback pulses of 15 kv or more for the high-voltage rectifier that produces the kinescope anode voltage. Furthermore, high values of self-induced voltage can make the output circuit oscillate. To control this sine-wave ringing, a diode damper is used in the horizontal output circuit. Operation of the horizontal deflection amplifier must be considered with the diode damper, as both stages are needed to produce horizontal scanning. Also, it is important to remember that without horizontal scanning there is no high voltage, and no brightness on the kinescope screen.

18 · 1 Functions of the horizontal output circuit

As illustrated in Fig. 18 · 1, this circuit includes three stages, the amplifier, damper, and high-voltage rectifier. Note the tube types commonly used. Operating in a high-efficiency circuit for maximum deflection with minimum d-c power input, the stages have the following functions.

1. The amplifier V_1 is the output stage transformer-coupled to the horizontal deflection coils in the yoke to provide sawtooth current for scanning. The plate current of the output tube flows in the primary of the output transformer. Its peak value is about 400 ma. This primary current is stepped up by the transformer for the amount of secondary current required by the scanning coils.

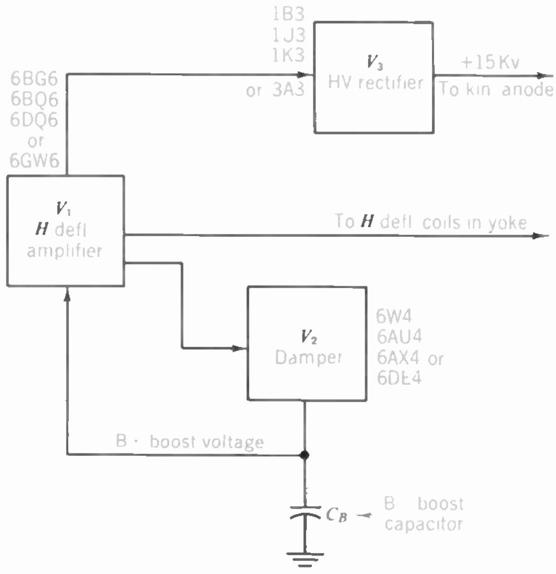


Fig. 18·1 Block diagram of horizontal output circuit.

2. The damper diode V_2 has the primary purpose of damping oscillations. Immediately after the fast flyback, the diode conducts to serve as a low shunt resistance that stops the ringing of the inductance in the output circuit. Damping is needed because the oscillations produce vertical white bars at the left side of the raster. The damper does not conduct during flyback, however, because its plate is negative then. At this time, damping is not desired in order to allow a fast flyback and maximum amount of high voltage.
3. For greater efficiency in horizontal scanning, the damped current is used to produce about one-third the trace at the left side of the raster. During this trace time immediately after flyback, the diode damper is conducting but the deflection amplifier is cut off. The average plate current of the output tube is reduced appreciably, therefore, improving the efficiency. Since the damped current produces part of each horizontal trace, both the damper and amplifier stages produce horizontal deflection. This system is often called *reaction scanning.*
4. When it conducts, the damper allows current to charge a capacitor in series with the B+ supply and the diode. The voltage across C_B in Fig. 18·1 then becomes higher than B+ by the amount of rectified deflection voltage. As an example, 300 volts B+ can be boosted to 500 volts across C_B. This boosted B+ is the plate-supply voltage for the amplifier, which is the reason why the damper must be operating for horizontal output.
5. The high-voltage pulse produced across the primary of the horizontal output transformer during flyback is stepped up, rectified by V_3, and filtered to provide anode voltage of about 15 kv for the kinescope. The high-voltage rectifier is the only tube conducting in the horizontal output circuit during retrace time.

These five functions illustrate how the output tube, damper, and high-voltage rectifier work together to provide high-efficiency horizontal scanning for full width in the raster and flyback high voltage for the kinescope anode.

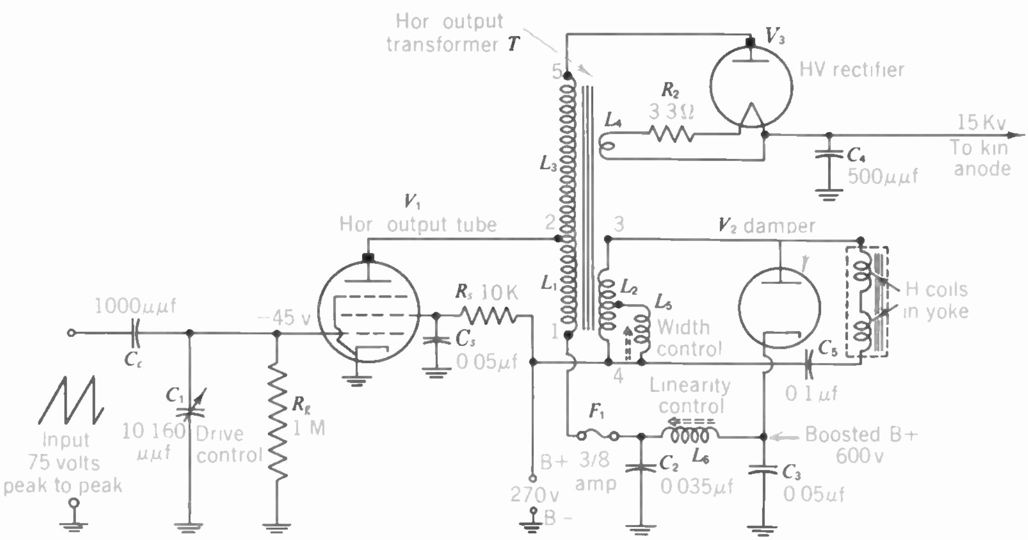
18 · 2 Horizontal amplifier circuit

The block diagram in Fig. 18 · 1 corresponds to the schematic diagram in Fig. 18 · 2. V_1 is the amplifier, transformer-coupled by the horizontal output transformer to the deflection coils in the yoke. V_2 is the damper to prevent excessive ringing, produce boosted B+ voltage across C_3, and provide reaction scanning. V_3 is the high-voltage rectifier for kinescope anode voltage. Figure 18 · 3 shows the high-voltage cage on the receiver chassis.

In Fig. 18 · 2, C_c and R_g couple sawtooth voltage from the horizontal oscillator to the output tube. The grid-driving voltage is about 75 volts peak to peak. Its amplitude is varied by adjusting C_1, which is the horizontal drive control. Grid-leak bias is developed because grid current flows when the positive peak of input voltage drives the grid positive. In fact, measuring the grid-leak bias with a d-c voltmeter is a good check to see if the output tube has grid drive from the oscillator. In Fig. 18 · 2, for instance, a d-c voltmeter at the control grid will read -45 volts to ground, indicating normal drive from the oscillator. With less grid drive, the grid-leak bias is less. If the bias is zero, this means there is no output from the oscillator.

The plate current of V_1 can flow through the output transformer primary L_1, the fuse F_1, through L_6 to C_3, which provides boosted B+ as the

Fig. 18 · 2 Schematic diagram of horizontal output circuit.



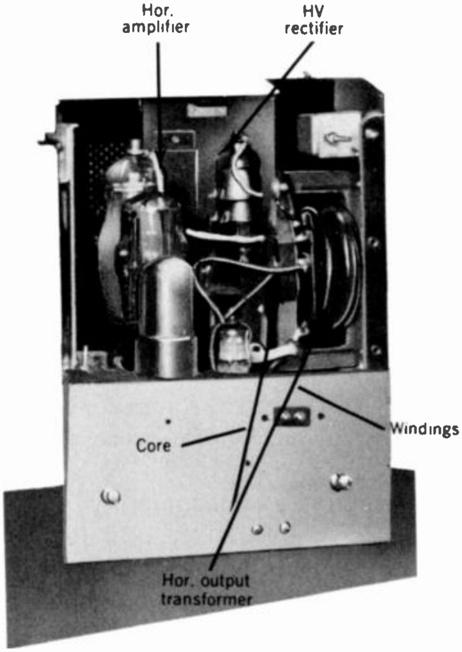


Fig. 18·3 Components of horizontal output circuit in high-voltage cage. (RCA Institutes, Inc.)

source of plate-supply voltage. Current in the output tube is a load on C_3 making it discharge. However, C_3 is recharged when the damper conducts at the start of trace. Note that C_5 blocks direct current from the deflection coils, which would shift the horizontal centering. Since C_5 has little reactance at 15,750 cps, the scanning coils are effectively connected across the secondary winding L_2 for the alternating horizontal scanning current.

As the linear rise of sawtooth input voltage drives the grid in the positive direction the amount of plate current increases, providing a sawtooth rise of current in L_1. This primary current induces voltage in L_2 to produce a sawtooth rise of current through the scanning coils in the secondary circuit. At the peak of the sawtooth input, the grid voltage drops sharply to a value more negative than cutoff. This negative grid drive cuts off the amplifier. As a result, V_1 stops supplying current to the output circuit. The output tube continues cut off while the grid-driving voltage completes its swing in the negative direction for the flyback. Also, the tube remains cut off until the linear rise in the positive direction makes the instantaneous grid voltage less negative than the cutoff voltage. Then the sawtooth rise in grid voltage produces a sawtooth rise of current again in the output circuit to produce the next cycle of operation.

Figure 18·4 shows the grid-plate transfer characteristic curve of the horizontal output tube. Note the sawtooth grid-driving voltage but plate-current pulses corresponding to part of the sawtooth voltage input. While the grid voltage due to the bias and driving voltage is more negative than cutoff, no plate current flows. For approximately two-thirds the cycle, however, the sawtooth input voltage drives the grid voltage less negative than cutoff to produce a linear rise of plate current. This time is indicated by the shaded area of grid voltage in Fig. 18·4. As an example, when the a-c input voltage drives the grid voltage 30 volts more positive than the bias of -40 volts, the instantaneous grid voltage is -10 volts, allowing plate current to flow.

The horizontal output tube is a beam-power amplifier to supply the required amount of scanning current. Typical maximum ratings are: 17.5 watts plate dissipation, 175 ma average cathode current, 550 ma peak

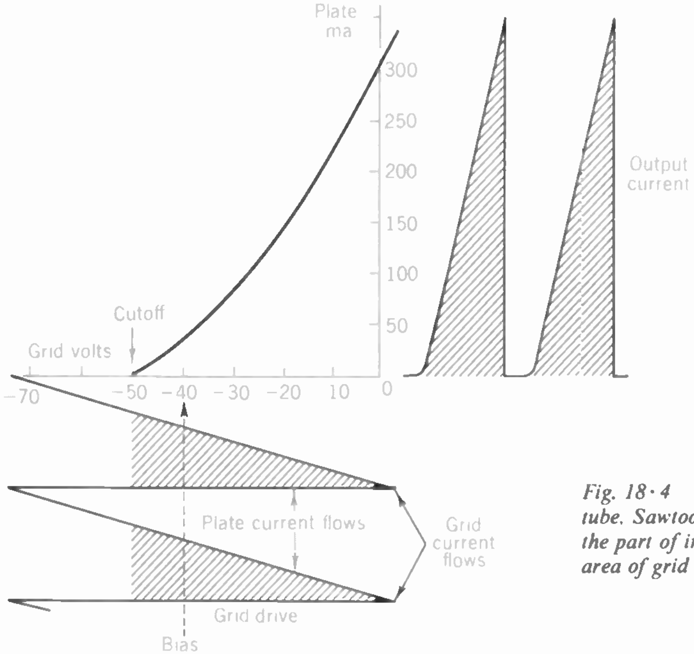


Fig. 18·4 Operation of horizontal output tube. Sawtooth plate current flows for only the part of input cycle indicated by shaded area of grid voltage.

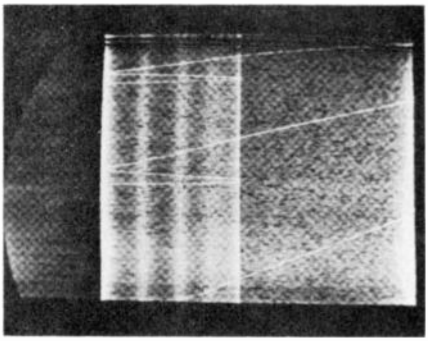
cathode current at end of trace, and 6,500 volts peak positive pulse at the plate during flyback. Maximum d-c plate supply is 770 volts, including boosted B+.

18·3 Damping in the horizontal output circuit

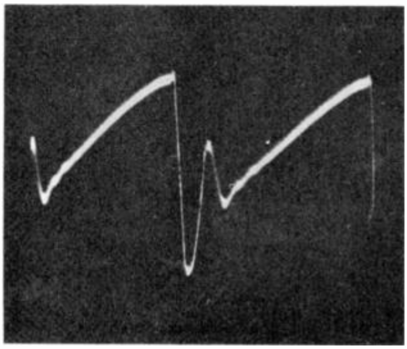
The inductance of the output transformer and scanning coils, with their distributed and stray capacitances, provide a tuned output circuit that can oscillate at its resonant frequency. Oscillations occur because the flyback on each sawtooth wave of current produces a rapid change in current that generates a high value of induced voltage across the inductance. The output circuit then oscillates at its natural resonant frequency by shock excitation. This effect is called *ringing.*

The oscillations, at about 70 kc in the horizontal output circuit, would continue past flyback time and produce ripples on the scanning-current waveform, as shown in Fig. 18·5*a*. Since the scanning current produces deflection, the electron beam also oscillates back and forth in accordance with the oscillatory ripples. As a result, the oscillations can produce one or more white bars at the left side of the raster, as shown in *b*. The bars are at the left because the oscillations occur immediately after flyback time. The bars are white because the electron beam scans these areas several times during every horizontal line traced, as the oscillations make the horizontal scanning current repeat equal amplitudes at different times. Ringing occurs in the horizontal output circuit, primarily, rather than in vertical scanning, because the fast horizontal flyback produces high values of induced voltages.

Shock-excited oscillations. The action of shock-exciting a tuned circuit into oscillations is illustrated in Fig. 18·6. In *a* the switch *S* has been



(b)



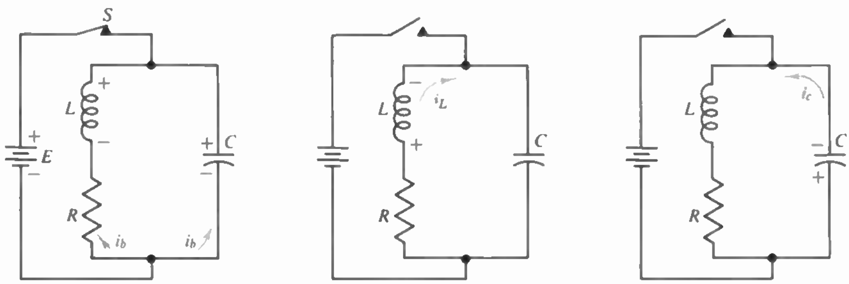
(a)

Fig. 18·5 Effect of insufficient damping in horizontal output circuit. (a) Oscillogram of horizontal scanning current with oscillations. (b) White bars at left side of raster caused by oscillations. (RCA Institutes, Inc.)

closed and current flows through L and R, which represent an inductive-resistive output circuit for magnetic scanning, while the distributed capacitance C charges to the applied voltage. When the switch is opened in b, the voltage E is disconnected from the tuned circuit, removing the source of applied voltage and its internal resistance. Then the change in applied voltage can make the tuned circuit start oscillating.

The tuned circuit oscillates without the battery because the energy stored in the electromagnetic field of the inductance and the electrostatic field of the capacitance, while the battery was connected, provides the required current and voltage. The current in the tuned circuit and the voltage across it oscillate with decreasing amplitudes until the stored energy is dissipated in the resistance. The switch S corresponds to the action of the

Fig. 18·6 Shock-exciting a tuned circuit into oscillation. (a) Switch is closed for battery to supply current and voltage for L and C. (b) Switch is opened and self-induced voltage across L acts as the generator. (c) Reversed polarities one-half cycle later with voltage across C as the generator.



(a) (b) (c)

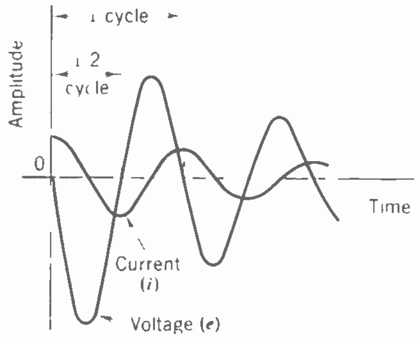


Fig. 18·7 Current and voltage waveforms in oscillating LC circuit. Note i and e are 90° out of phase with each other.

deflection amplifier, as it conducts during trace time to supply current to the output circuit but then is cut off at flyback time by the grid-driving voltage.

The current in the coil decays toward zero when the switch is opened. However, the sharp change as the current starts to decline produces a large self-induced voltage across L. Now the coil is a generator producing voltage that keeps the current flowing in the same direction through the inductance as when the switch was closed. This is illustrated in *b* of Fig. 18·6. At the same time the capacitor discharges through L and R, reducing the voltage across C. Its discharge current is in the same direction as the current in the coil.

A little later the capacitor is completely discharged and the voltage across C is zero. The current i_L produced by the coil still is in the same direction, however, charging C to produce a voltage of opposite polarity from the original battery voltage. As the current in the coil decays to zero, the voltage across C charges to its maximum value.

Then the capacitor voltage becomes the generator as it discharges to produce the current i_c through L and R, as shown in *c*. This current through the coil is in the opposite direction from the original current from the battery. When the capacitor voltage is down to zero, the coil again acts as the generator to maintain the current. C then charges with its original polarity. As a result, the inductance and capacitance interchange energy, making the tuned circuit oscillate.

The waveforms of the oscillations are shown in Fig. 18·7. Notice that the current and voltage are 90° out of phase with each other, as the current is maximum when the voltage is zero. During the first half cycle of oscillations, the current reverses from maximum in one direction to maximum in the opposite direction, while the voltage reaches its maximum negative value and returns to zero.

The LC circuit oscillates at its resonant frequency, with reduced amplitude for successive cycles until the stored energy is dissipated in the resistance. The frequency is approximately $1/(2\pi\sqrt{LC})$, when R is small. How long the oscillations continue depends upon the Q of the circuit. The higher the series resistance R, the lower the Q and the sooner the oscilla-

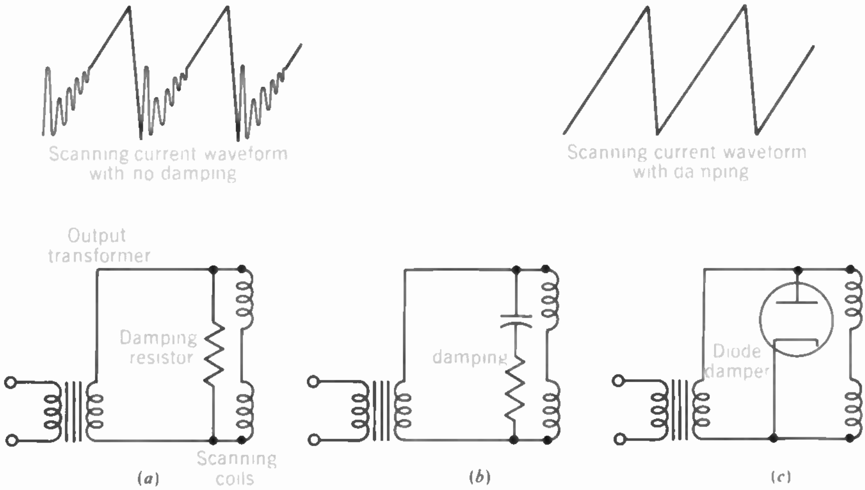


Fig. 18·8 Methods of damping. (a) Shunt resistor. (b) RC circuit. (c) Diode damper.

tions decay to zero. A low value of resistance in parallel with the circuit has the same effect as a high series resistance in reducing the Q and damping the oscillations.

Damping methods. The undesired oscillations can be eliminated from the linear rise on the sawtooth wave by connecting a damping resistance in parallel with the scanning coils as shown in Fig. 18·8. Because of its low value, which is about 1,000 ohms, the parallel damping resistor lowers the Q of the output circuit, reducing the amount of self-induced voltage. With just a simple resistance for damping, however, some of the scanning current is shunted through the parallel resistor. By insertion of a capacitor in series with the damping resistor, as in *b*, the scanning current can be prevented from flowing through the damping resistor, while the high-frequency oscillations are damped out.

Either of these two methods is usually employed for damping in the vertical output circuit. They are not suitable for the horizontal output circuit, however, because the parallel damping resistor would increase the current drain on the horizontal output tube, and make the flyback time too long.

For horizontal damping, a diode damper tube is generally used, as illustrated in Fig. 18·8*c*. The diode can be an open circuit when it is not conducting, to allow a fast horizontal flyback, and then conduct to act as a low damping resistance during trace time. The first half cycle of oscillations makes the diode plate voltage negative, preventing conduction in the damper. Meanwhile, the current in the horizontal deflection coils drops rapidly to produce the fast horizontal flyback. After the first half cycle of oscillations, the damper plate voltage is positive, causing conduction in the diode. Then the diode damper is a relatively low resistance across the horizontal deflection coils that can damp out the oscillations.

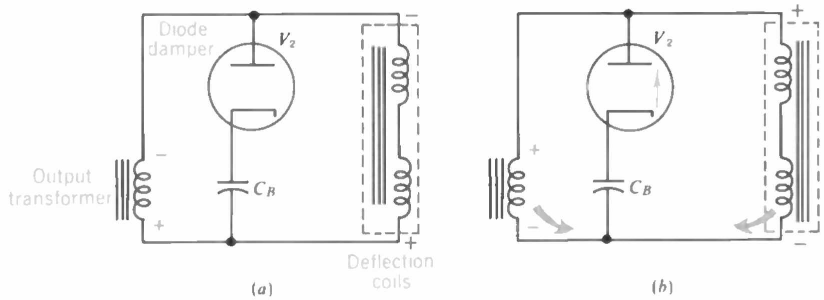


Fig. 18·9 (a) Damper plate negative to prevent conduction during retrace. (b) Damper plate positive and conducting at start of trace.

18·4 Horizontal scanning and damping

When the horizontal output tube conducts to supply current for the horizontal scanning coils, the linear rise in current deflects the electron beam to the right side of the raster. During this time, energy is stored in the inductance and capacitance of the output circuit. Then, when the deflection amplifier tube is cut off by its grid-driving voltage, the output circuit starts oscillating.

The first half cycle of oscillations is allowed to continue undamped for a fast flyback from the right side of the raster to the left side. After the retrace, though, the damper conducts to make the oscillations decay to zero. The damped current in the horizontal scanning coils deflects the electron beam from the left side of the raster toward the center. Just before the damped current decays to zero, the deflection amplifier starts conducting again to finish the trace to the right edge of the raster.

The operation of the diode damper in the horizontal output circuit is illustrated in Fig. 18·9. The first half cycle of oscillations makes the damper plate negative, as shown in *a*. Then the diode cannot conduct. As a result, the undamped current in the deflection coils drops sharply from its maximum value to zero and reverses to its maximum in the opposite direction very quickly for a fast flyback. With a frequency of 70 kc for the oscillations in the horizontal output circuit, the period of a half cycle is 1/0.14 μsec, resulting in a flyback time of approximately 7 μsec. This half cycle is from maximum positive *i* to maximum negative *i* in Fig. 18·10.

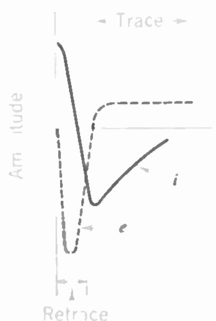


Fig. 18·10 Waveforms of voltage and current for deflection coils while output tube is cut off.

The voltage across the oscillating circuit is 90° out of phase with the current, as shown in Fig. 18·10. Therefore, the polarity of damper plate voltage reverses to become positive just as the current starts to decrease from its maximum negative value. This is the time for the start of the trace. Now the damper tube conducts because of positive plate voltage.

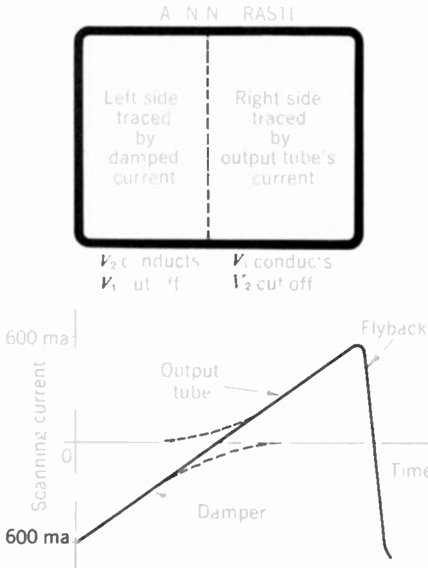
Conduction in the diode is illustrated in Fig. 18·9*b*. Note that C_B is charged by the damper current at this time. As the damped current declines to zero, the electron beam is deflected from the left side of the raster toward the center. Finally, the voltage across C_B biases the diode damper out of conduction. With zero current in the scanning coils, the electron beam would be undeflected at the center. Before the damped current declines to zero, however, the output tube starts conducting. Its linear rise of current then completes the trace to the right side of the raster.

Figure 18·11 illustrates how the current supplied by the deflection amplifier and the damped current in the output circuit combine. The result is sawtooth current in the deflection coils for the full horizontal trace from left to right across the scanning raster. Note that both components of the scanning current for the trace deflect the electron beam toward the right. The decrease in damped current of negative polarity varies in the same direction as the positive rise of current produced by the output tube.

In summary, the horizontal scanning occurs in three steps:

1. The output tube supplies current for the horizontal scanning coils to deflect the electron beam to the right edge of the raster.
2. The undamped oscillation in the horizontal output circuit, while the output tube and damper are not conducting, produces the rapid reversal of scanning current required for the fast flyback from right to left.
3. Immediately after the flyback, the damper conducts while the output tube is cut off, to provide the reaction scanning current for the trace at the left side.

Fig. 18·11 Illustrating how the damped current and output current combine to produce horizontal scanning in the raster.



The damper tube must withstand the high negative voltage pulse at the plate, have good cathode-to-heater insulation because of the B+ boost voltage at the cathode, and be able to conduct the peak values of scanning current. Typical maximum ratings for the 6AU4 half-wave rectifier as a diode damper are: 4,500 volts peak inverse plate voltage, 1,050 ma peak

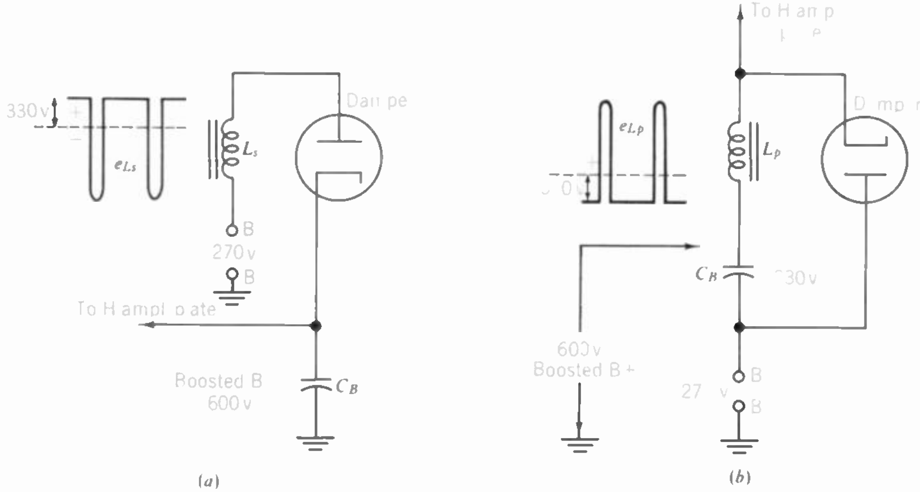


Fig. 18 · 12 Producing boosted B+ voltage. (a) Diode damper in secondary rectifying positive polarity of a-c deflection voltage. (b) Inverted diode damper in primary rectifying negative polarity of a-c deflection voltage.

plate current, and 900 volts average positive cathode voltage.

18 · 5 Boosted B+ voltage

The damper tube is a half-wave rectifier for the a-c deflection voltage produced during trace time in the horizontal output circuit. Its d-c output voltage then is added, in effect, to the B supply voltage. In Fig. 18 · 12*a*, the B+ voltage is applied to one side of the secondary winding L_s, while the other side is connected to the damper plate. Therefore, when the damper conducts, it charges C_B in the cathode circuit to the B+ voltage. The path of charging current is through L_s and from cathode to plate in the damper, charging C_B to 270 volts, with the cathode side positive. In addition, the a-c deflection voltage e_{L_s}, which produces sawtooth current in the scanning coils, makes the diode plate 330 volts more positive during horizontal trace time. As a result, C_B charges to a voltage 330 volts higher than B+, or to 600 volts. The path of charging current is the same as when the B supply charges C_B. In effect, the a-c deflection voltage is in series with the B supply voltage and the damper tube, to produce the boosted B+ voltage across C_B.

The damper circuit in Fig. 18 · 12*a* is for an output transformer with an isolated secondary, which can have the polarity of deflection shown with negative flyback pulses. In an autotransformer, however, any part of the winding must have deflection voltage with positive flyback pulses, the same polarity as the primary winding. Then the damper circuit shown in *b* with an inverted diode is used.

In Fig. 18 · 12*b*, the diode is inverted, for damping in the primary of the output transformer. Then the a-c input voltage is applied to the cathode. This connection is necessary because the deflection voltage is negative in the primary during trace time. The negative polarity of a-c deflection voltage at the cathode produces plate current in the diode to charge C_B. Note that the rectifier plate current still charges C_B with its cathode side positive. The d-c voltage across C_B is approximately equal to the peak negative input to the cathode, which equals 330 volts here. However, C_B is actually in series with the B supply voltage. The potential difference between the positive side of C_B and chassis ground, therefore, is 600 volts. This value of boosted B+ voltage is the plate supply for the horizontal amplifier. Note that the voltage rating of C_B need only be 330 volts in this circuit, which is the potential difference across the two ends of the capacitor, compared with 600 volts across C_B in *a*.

The boosted B+ voltage is commonly used for the horizontal output tube, horizontal oscillator, vertical oscillator, and vertical amplifier, because the higher plate-supply voltage allows more power output with improved linearity. In addition, the screen-grid supply of the kinescope is generally boosted B+ voltage. It is important to note that the B+ boost circuit must be operating in order to produce normal plate voltage for all these stages.

A question often asked is "how can the horizontal amplifier operate when the receiver is first turned on, before boosted B+ voltage is available across C_B?" In effect, this is a *bootstrap circuit* where the output stage increases its own plate supply voltage. Notice in Fig. 18·12 that for the circuits in both *a* and *b* the amplifier plate circuit can return to B+ through the cathode-to-plate path in the conducting damper tube. Therefore, the horizontal output tube starts conducting with just B+ voltage for the plate supply. After a few cycles of operation, boosted B+ voltage is available across C_B.

18·6 Flyback high voltage

The voltage pulse produced in the horizontal output circuit during the first half cycle of oscillations for flyback provides the d-c high voltage required for the kinescope anode. As illustrated in Fig. 18·13, operation of the flyback high-voltage supply can be summarized as follows:

1. The flyback pulse has positive polarity at the plate of the horizontal amplifier in the primary of the output transformer, as shown by waveshape 1. Since both the damper and amplifier tubes are cut off for the retrace, the first half cycle of undamped oscillations produces the high-voltage flyback pulse. A typical value is 5,000 volts at the V_1 plate.
2. The high-voltage primary winding L_3 steps up the flyback pulse for the plate of the high-voltage rectifier V_3, as shown by waveshape 2. This amplitude is 15 kv because of the 3:1 voltage step-up.
3. With the stepped-up a-c deflection voltage at the plate of V_3, it conducts to produce positive d-c output voltage at the cathode. This voltage is filtered to provide the steady d-c output shown by waveshape 3. The d-c output voltage is approximately equal to the peak positive a-c input of 15 kv.

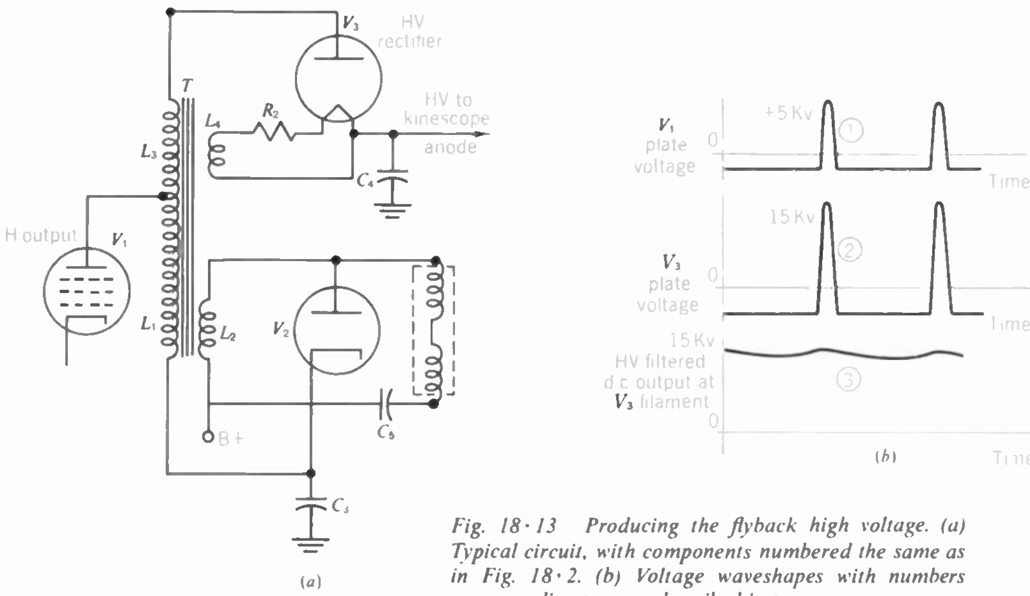


Fig. 18·13 Producing the flyback high voltage. (a) Typical circuit, with components numbered the same as in Fig. 18·2. (b) Voltage waveshapes with numbers corresponding to steps described in text.

When the high-voltage rectifier conducts, current can flow from cathode to plate in V_3, through L_3 and L_1 in the primary of the output transformer, discharging slightly the B+ boost capacitor C_3 and charging the filter C_4. Since the ripple frequency is 15,750 cps for the half-wave rectifier, 500 $\mu\mu$f is enough capacitance for C_4 to provide the required filtering. R_2 is a voltage-dropping resistor for the heater of V_3. Its heater power is supplied by L_4, which is a single-turn loop on the transformer core.

It is important to note that, with a flyback high-voltage supply, the d-c output for kinescope anode voltage cannot be produced without the a-c deflection voltage input to the plate of the high-voltage rectifier. Therefore, the horizontal deflection circuits must be operating to have brightness on the kinescope screen. The amount of high voltage is maximum when the brightness control is at minimum, because of less load current on the high-voltage rectifier.

18·7 Horizontal deflection controls

The drive, linearity, and width controls for the horizontal output circuit are setup adjustments on the rear apron of the chassis. In many receivers, however, some or all of these adjustments for horizontal scanning are omitted. The reason is that incorrect adjustments can cause abnormally high values of average d-c plate current in the horizontal output tube.

Horizontal drive control. Referring back to Fig. 18·2, C_1 is a variable mica capacitor to adjust the amount of grid-driving voltage for the output tube. Since the grid voltage controls plate current, adjusting C_1 varies the output of the deflection amplifier. In addition, more drive increases the grid-leak bias. Note that C_1 forms a capacitive voltage divider with the coupling capacitor C_c for a-c grid voltage. The smaller capacitance develops more voltage. Therefore, C_1 is adjusted for less capacitance to increase the grid drive. In some circuits, C_c is made variable to vary the grid voltage.

More horizontal drive increases the width of the raster. The brightness also increases because of more high voltage. However, excessive drive usually causes a white vertical line near the center, as in Fig. 18·14. The drive can generally be set by adjusting for more width and brightness until the white line appears and then backing off slightly. This is the point of best efficiency, as maximum grid-leak bias allows minimum average plate current.

Width control. The width coil L_5 in Fig. 18·2 is in parallel with part of the secondary winding that provides deflection current for the horizontal deflection coils. Reducing the inductance of L_5 results in more shunt current through the width coil, reducing the horizontal deflection current in the yoke. However, the total stored energy at the end of the scan remains approximately the same to allow the same amount of flyback high voltage. Therefore, L_5 can vary width in the raster without changing its brightness. A typical width coil is shown in Fig. 18·15.

Additional methods of width control include varying either the screen-grid voltage or cathode bias for the output tube. These adjustments affect the high voltage. In some receivers a metal sleeve inside the yoke housing is moved in or out to vary the width. Acting as a shorted-turn secondary, its effect is mainly on width because of the higher frequency of horizontal scanning compared with vertical scanning.

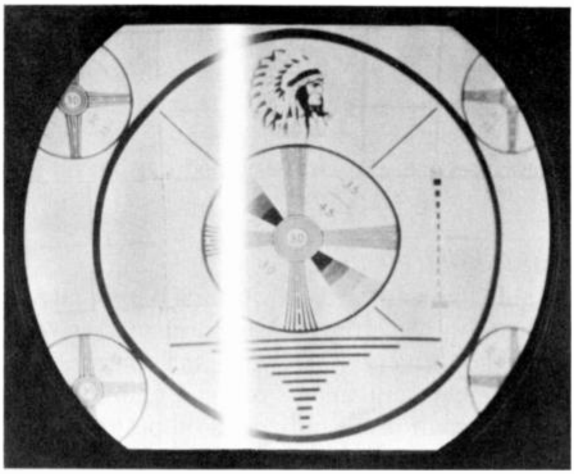


Fig. 18·14 White vertical bar near center of raster, caused by excessive horizontal drive. (Admiral Corporation.)

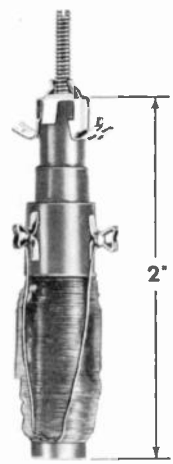


Fig. 18·15 Width coil. Inductance is 3 to 16 mh.

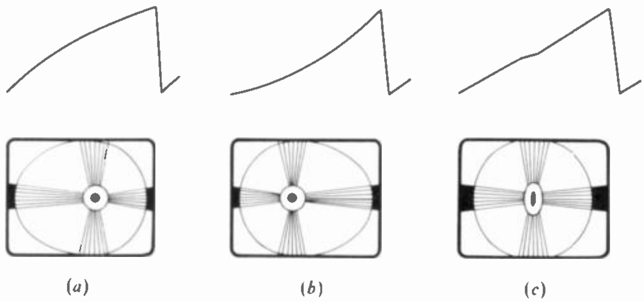


Fig. 18·16 Nonlinear horizontal scanning. (a) Stretching at left and crowding at right. (b) Reverse nonlinear distortion. (c) Crowding at center.

Horizontal linearity. Three examples of nonlinear horizontal scanning are illustrated in Fig. 18·16. To relate the current waveshapes to the scanning, remember that:

1. Damped current produces the left side of the trace.
2. The beam is at the center with zero current, when the damped current approaches zero and the output tube starts to conduct.
3. Plate current of the output tube completes the trace at the right side.

How fast the damped current decays to zero depends on the damper tube and B+ boost capacitor. The time when the output tube starts to conduct depends on its grid-leak bias and plate voltage. Conduction is delayed until the driving voltage is positive enough to make the instantaneous grid voltage less than cutoff. How far the beam is deflected to the right edge depends on the peak plate current in the output tube.

In Fig. 18·16*a*, the picture information is stretched at the left and crowded at the right. With people in the picture, they look too broad at the left or too thin at the right. This type of nonlinearity can be caused by too much horizontal drive and too little inductance in the width coil. A weak horizontal output tube produces the same nonlinearity, but with reduced width. The reverse happens for the nonlinear scanning in *b*, where the left side is crowded and the right side stretched. This effect can be caused by a weak damper tube or too much load current for the B+ boost capacitor. With either type of nonlinear horizontal scanning, a person at the center of the screen appears to have one shoulder broader than the other.

The linearity coil L_6 in Fig. 18·2 forms a parallel resonant circuit with C_3 and C_2, tuned to 15,750 cps, approximately. The inductance of L_6 can be varied from 1 to 8 mh. Adjusting L_6 shifts the phase of 15,750-cps a-c ripple in the B+ boost voltage. There are several effects. First, parallel resonance allows minimum average current in the output tube plate circuit, for maximum efficiency. In addition, adjusting L_6 affects the output tube plate voltage and damper cathode voltage at the transition period when V_2 finishes conduction and V_1 starts. Therefore, L_6 adjusts linearity at the

Table 18 · 1 Deflection yokes

Horizontal coils		*Vertical coils*		*Deflection*
Inductance, mh*	*R, ohms*	*Inductance, mh*	*R, ohms*	*angle, deg*
8.3	13.5	50	66	53°
13.5	17.5	50	53.5	70°
20	28	45	45	90°
20	38	40	39	110°

* The listed values are common but there are many different types.

center of the raster. Figure 18 · 16*c* illustrates crowding at the center. This can be caused by damper current decaying to zero before the output tube starts to conduct.

18 · 8 Deflection yokes

Since the current in the deflection coils deflects the electron beam in the kinescope, the yoke is rated in terms of deflection angle. The yoke rating should correspond to the kinescope. Typical angles are 53°, 70°, 90°, and 110°, as listed in Table 18 · 1. A 90° yoke is shown in Fig. 18 · 17. Sweep current requirements for this yoke are approximately 1,200 ma for the 20-mh horizontal coils and 600 ma for the 50-mh vertical coils, both in peak-to-peak values, for 16 kv anode voltage.

Kinescopes with the same angle and high voltage require equal deflection for a full-sized raster, regardless of screen size. As an example, a 90° yoke can fill a 19-in. screen or 23-in. screen when both kinescopes have the same deflection angle of 90°. It should be noted, though, that yokes

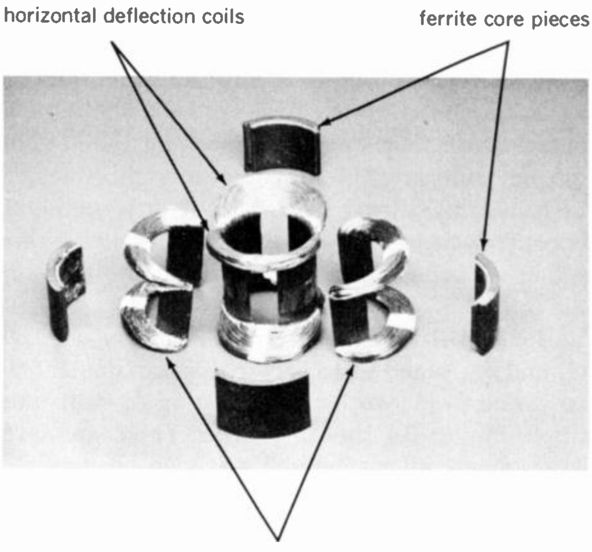


Fig. 18 · 17 Cutaway view showing deflection coils in yoke. (RCA Institutes Home Study School.)

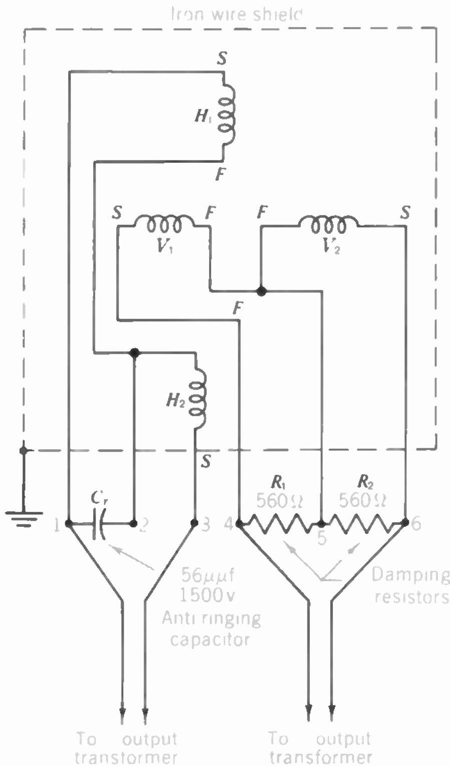


Fig. 18·18 Wiring connections in deflection yoke. S is start and F finish of winding.

with the same deflection angle can have different electrical characteristics. Some units have an internal thermistor connected between the vertical coils. The 110° yokes have a smaller center hole for narrow-neck kinescopes. This feature allows more deflection for the same sweep current.

The inductance of the horizontal coils is usually low to provide a high resonant frequency for a fast flyback in the horizontal output circuit. Although their inductance is lower than the vertical coils, the horizontal coils are an inductive load because of the large self-induced voltage at the 15,750-cps scanning frequency. With sawtooth current in the horizontal coils, the voltage across the inductive circuit has the rectangular waveshape, with sharp pulses for flyback. In fact, this rectangular voltage with flyback pulses is the voltage waveshape across any part of the inductive horizontal output circuit.

Wiring connections for the terminals at the back of the yoke are illustrated in Fig. 18·18. The vertical coils are in series with each other, as are the horizontal coils. Note the damping components mounted in the yoke housing. The small capacitor C_r across only one horizontal coil balances the capacitances to ground. Without this capacitor, or with the wrong value, excessive ringing in the horizontal coils causes white vertical bars at the left in the raster. R_1 and R_2 across the vertical coils prevent the horizontal oscillations from affecting vertical scanning. Without these damping resistors, the top and bottom of the raster have ripples at the left side. Note that these components are located in the yoke.

18·9 Horizontal output transformers

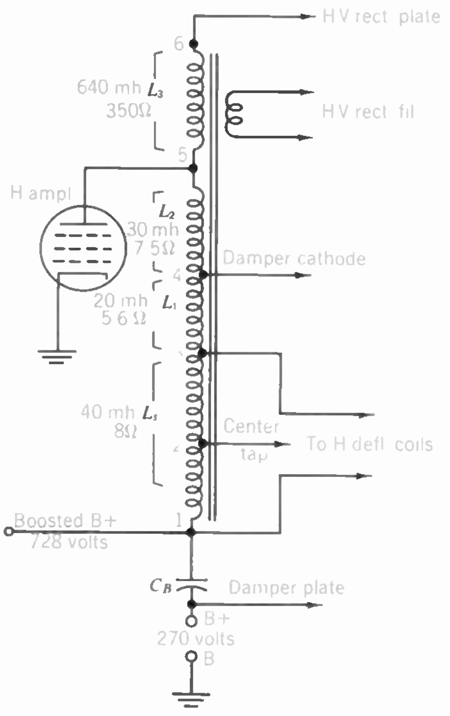
A typical unit is shown in Fig. 18·19. There are three main types, classified according to the way the horizontal deflection coils are connected into the output circuit:

1. Isolated secondary. This method is illustrated schematically in Fig. 18·2. The secondary has negative flyback pulses, requiring the damper connections shown.

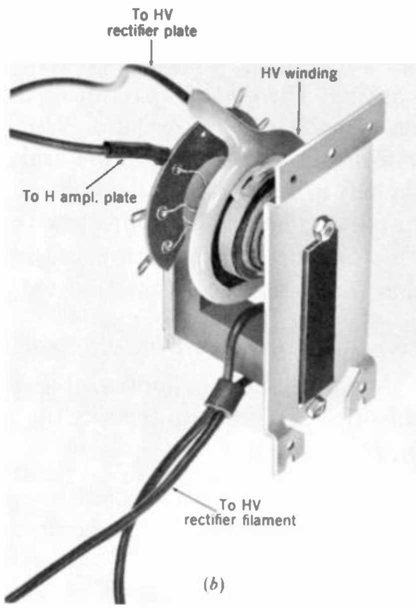
2. Direct drive. In this method, the deflection coils are directly in series in the plate circuit of the output tube. Direct drive is seldom used, however, because the required high-impedance deflection coils lower the resonant frequency of the output circuit.

3. Autotransformer. This method, illustrated in Fig. 18 · 19, is the most common type of horizontal output circuit. The total winding between terminals 5 and 1 is the primary for plate current of the amplifier. The tapped winding between terminals 3 and 1 is the secondary to step down the voltage for the deflection coils. Note that the turns ratio is between all the turns in the primary and the number of turns tapped for the secondary. As an example, a tap at the center of an autotransformer provides a 2:1 turns ratio.

Since there is no isolated secondary winding for polarity inversion, all the taps on an autotransformer have voltage of the same polarity with respect to chassis ground. In Fig. 18 · 19*a*, for instance, the voltage at terminal 3, 4, or 5 has the rectangular waveshape with positive flyback pulses. The damper diode must be inverted, therefore, for flyback pulses to drive its cathode positive, keeping the damper out of conduction during retrace time. In addition, the damper cathode can be tapped up on the autotransformer for more negative deflection voltage during trace time, to be rectified for higher boosted B+ voltage.



(*a*)



(*b*)

Fig. 18 · 19 Autotransformer type of horizontal output transformer. (a) Schematic with typical values. (b) Photograph. (Triad Transformer Corporation.)

The functions of the horizontal output transformer can be summarized in terms of the autotransformer in Fig. 18 · 19*a*.

1. *Horizontal scanning.* The output transformer matches the horizontal amplifier plate circuit to the low-impedance deflection coils, in order to produce the required sweep width. Plate current flows through L_2, L_1, and L_s. Their combined value is the primary inductance. The voltage across the secondary L_s produces the required scanning current in the yoke.

2. *Flyback high voltage.* The winding L_3 at the top between terminals 6 and 5 steps up the primary voltage to supply high voltage a-c input to the high-voltage rectifier. Its filament power is also taken from the output transformer. Positive flyback pulses of about 5 kv amplitude at the amplifier plate can be stepped up for 15 kv at the high-voltage rectifier plate. The amount of step-up is limited by the fact that increased inductance and stray capacitance reduce the resonant frequency of the output circuit. L_3 is a separate winding with many turns of fine wire, which is the reason for its relatively high resistance.

3. *Boosted B+ voltage.* The a-c secondary voltage between terminals 4 and 1 is rectified by the damper to provide boost voltage across C_B. This voltage is series-aiding with the B+ voltage. Therefore, terminal 1 is the source of boosted voltage higher than B+ by the amount of rectified deflection voltage.

Although not shown in Fig. 18 · 19, the horizontal output transformer usually supplies positive flyback pulses of about 500 volts amplitude for a keyed AGC stage. In addition, extra windings can be used for flyback pulses of either positive or negative polarity. These pulses may be necessary in some receivers for the horizontal AFC circuit or for internal horizontal blanking in the kinescope.

18 · 10 Analysis of horizontal output circuit

The inductance values in Fig. 18 · 19*a* can be used for some approximate calculations to illustrate more details of how the horizontal output circuit functions. Unity coupling between windings is assumed because of the powdered-iron or ferrite core. Also, the coil resistances can be neglected, as the induced voltages are much greater than the *IR* drops.

In analyzing the inductance values, we can start with the secondary winding L_s for the deflection coils, as they require a specified amount of current for full deflection. The secondary inductance here is made 40 mh to be twice the total inductance of 20-mh deflection coils connected across L_s. Higher values of L_s allow more inductance in the primary for a given turns ratio. However, too much inductance can make the flyback time too long. The value of 20 mh, then, is suitable for the secondary inductance L_s. In the primary, its inductance L_p for the autotransformer includes all the turns of L_2, L_1, and L_s. The total primary inductance L_p then is 90 mh, equal to 30 mh for L_2 plus 20 mh for L_1 plus 40 mh for L_s.

The turns ratio for voltage step-up to L_p from L_s is equal to the square root of their inductance ratio, or $\sqrt{90/40}$. The square root is based on the fact that inductance of a coil increases as the square of its turns. This ratio of $\sqrt{90/40}$ equals $\sqrt{2.25}$, which is 1.5 for the turns ratio of L_p to L_s. With one and one-half the number of turns for L_p, its voltage is 50 per cent more than the voltage across L_s.

In order to calculate the induced voltages we can assume a flyback of 8 μsec. This time corresponds to one-half cycle of undamped oscillations at 62.5 kc. Trace time equals 8 μsec subtracted from the line period of 63.5 μsec. Then the remainder of time for trace is 55.5 μsec.

Let the peak-to-peak yoke current be 1,200 ma in the 20-mh horizontal deflection coils for a deflection angle of 90°. This value is derived from 400 ma peak plate current in the output tube stepped up to 600 ma in the yoke by the turns ratio. The opposite peak of 600 ma is a result of oscillation in the output circuit after the amplifier cuts off. During trace, the rate of change of current in the yoke is 1,200 ma during 55.5 sec. However, for flyback the rate of change is 1,200 ma in 8 μsec. These values are necessary to calculate the induced voltages as equal to $L\,di/dt$. The voltages induced by the rise of current during trace are negative with respect to chassis ground, at all terminals on the autotransformer. During flyback, the fast drop of current induces positive voltages of much higher amplitude. These voltages are shown in Fig. 18·20.

Induced voltages during retrace. The flyback voltage induced across the yoke coils by a 1,200-ma change in 8 μsec is

$$+e_y = L\,di/dt$$

$$= 20 \times 10^{-3} \times \frac{1{,}200 \times 10^{-3}}{8 \times 10^{-6}}$$

$$+e_y = 3{,}000 \text{ volts}$$

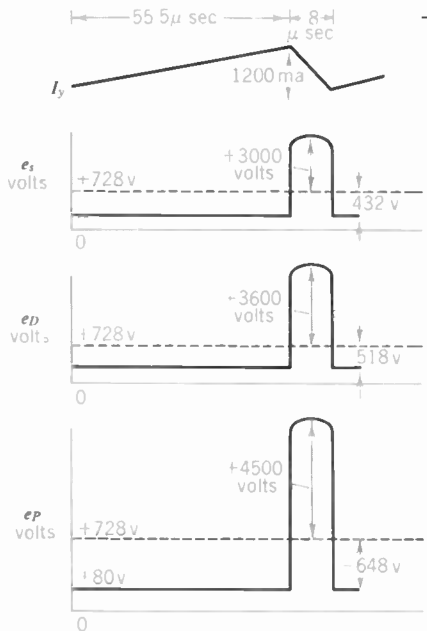


Fig. 18·20 *Induced voltages produced by yoke current, calculated for autotransformer in Fig. 18·19a.*

This voltage is also across L_s. Since the turns ratio of L_p to L_s is 1.5:1, the voltage across L_p is stepped up by 1.5 to be 4,500 volts at the plate of the output tube.

The high-voltage winding L_3 develops 4 × 3,000, or 12,000 volts during flyback. This voltage is four times the secondary voltage because L_3 is sixteen times greater than L_s. The 12,000 volts across L_3 is in series with the 4,500-volt flyback pulse across L_p to provide 16,500 volts peak positive a-c input to the plate of the high-voltage rectifier.

Induced voltages during trace. These are much smaller because of the slower rate of change of current. Also, the increasing current induces negative voltage. The amount of voltage induced across the yoke coils by a 1,200-ma change in 55.5 μsec is

$$-e_y = L\,di/dt$$

$$= 20 \times 10^{-3} \times \frac{1{,}200 \times 10^{-3}}{55.5 \times 10^{-6}}$$

$$-e_y = 432 \text{ volts approx}$$

This voltage is also across L_s during trace time. Across L_p the voltage is stepped up by 1.5, for -648 volts. The net plate voltage is still positive, however, because of the boosted B+ supply of 728 volts at terminal 1.

Notice that the damper cathode at terminal 4 is tapped higher than L_s. This connection results in more boosted B+ voltage, as more deflection voltage is applied to the damper for rectification. The combined turns of L_1 and L_s provide a voltage step-up of approximately 1.2 with respect to L_s. Therefore, the damper voltage e_D is 20 per cent more than e_y, or e_D equals 518 volts. This much voltage is available to drive the damper cathode negative for diode current to charge C_B. Note that this voltage is for the boost only. In addition, C_B is in series with the B+ of 270 volts. Therefore, the boosted voltage at terminal 1 with respect to chassis ground could be as high as 270 + 518, or 788 volts. With a typical load of 10 ma average current on the boost supply, however, its d-c voltage is usually a little less than the peak a-c input. In this example, the boost voltage alone is 458 volts. This value added to the B+ of 270 volts provides the 728-volt average axis shown in Fig. 18·20. Then 728 volts is the amount of boosted B+ voltage available from terminal 1 at the low end of the autotransformer.

Output tube plate dissipation. This rating is important for power tubes. The d-c power dissipated at the plate, equal to the product of average plate current and voltage, should not exceed the maximum specified for the tube. A typical rating is 15 watts for the 6DQ6 horizontal output tube.

The average current of the horizontal output tube can be taken as approximately one-fourth the peak plate current. This factor results from considering the average axis of the sawtooth waveform at one-half the peak value and taking one-half again because the output tube is cut off for almost 50 per cent of the cycle, including flyback time. For the example of 400 ma peak plate current, therefore, the average is 100 ma. A d-c milliammeter in series in the cathode-to-ground circuit would read this value of 100 ma average plate current, plus approximately 25 ma grid current, for a total of 125 ma average cathode current.

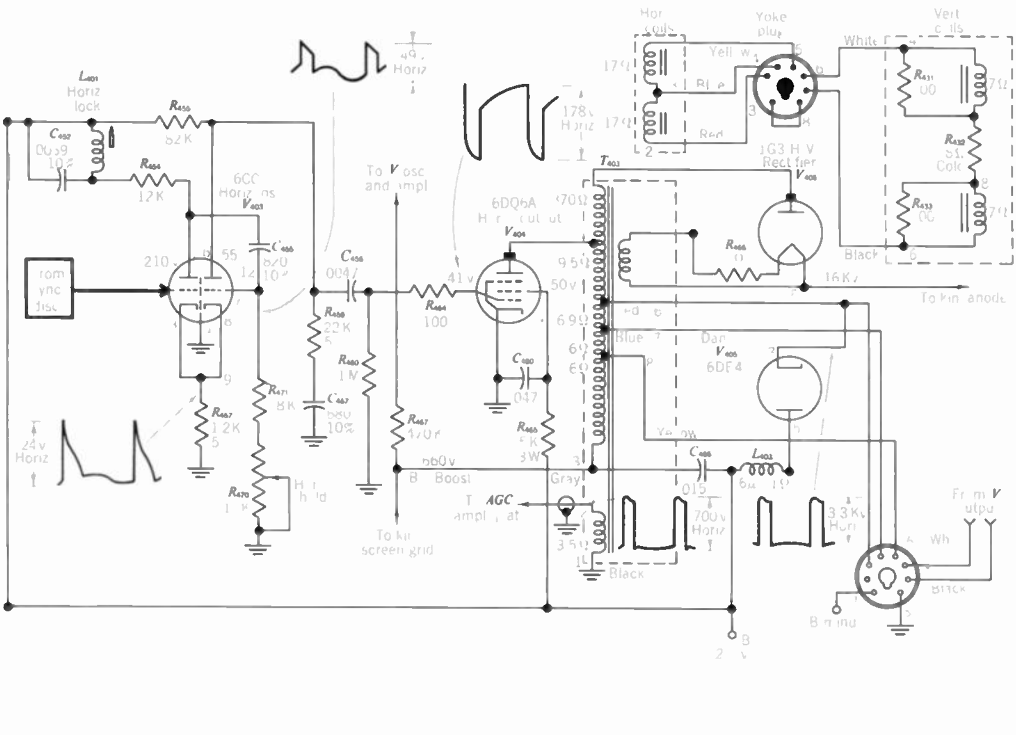
The average plate voltage of the horizontal output tube equals the positive boost supply voltage minus the negative induced voltage across the primary during trace time. For the example given, let the boosted supply voltage be 728 volts positive, with 648 volts negative induced voltage across the primary. The net plate voltage then is 728 − 648, which equals 80 volts. This is the average plate voltage while the output tube is conducting. See the waveform for e_p at the bottom in Fig. 18·20. Therefore, the plate dissipation is 80 volts × 0.1 amp, or 8 watts.

18·11 Typical horizontal deflection circuit

In Fig. 18·21, the horizontal oscillator is a cathode-coupled multivibrator driving the deflection amplifier with autotransformer coupling to the horizontal deflection coils in the yoke. The damper is an inverted diode rectifier producing boosted B+ voltage of 660 volts. The high-voltage rectifier supplies 16 kv anode voltage for the kinescope, with its anode capacitance as the high-voltage filter. Note that the octal yoke plug has a jumper between pins 1 and 8, which closes the B− return to chassis ground for operation of the power supply. Removing the yoke plug makes the entire receiver dead, therefore, because the B-supply circuit is open.

The oscillator is locked in sync by d-c control voltage from a sync discriminator, connected to grid pin 2. In the plate circuit, R_{454} is the plate load resistor in series with the sine-wave stabilizing circuit consisting of L_{401} and C_{452}. The voltage peak across the common cathode resistor R_{457} cuts off this triode section. Plate voltage output from pin 1 is coupled by C_{455} to the opposite triode section, serving as a discharge tube. Its grid

Fig. 18·21 Typical horizontal deflection circuit. Capacitance values more than 1 in μμf. Note yoke plug at top right and its socket connections at bottom.



resistance is varied by the hold control to adjust the free frequency of the oscillator. Plate pin 6 connects to B+ through R_{455}. The sawtooth capacitor is C_{457}. Note the peaking resistor R_{459} to provide trapezoidal grid voltage for driving the output tube. The peaking resistor can be on either side of the sawtooth capacitor, as they are in series. C_{456} is the coupling capacitor, with grid resistor R_{460}. The series grid resistor R_{464} limits the peak grid current charging the coupling capacitor for grid-leak bias, in order to reduce radiation of horizontal pulses.

The grid-leak bias of -41 volts on the output tube is produced by drive from the horizontal oscillator. There is no cathode bias. Screen-grid voltage is obtained from the 270-volt B+ line, with R_{465} the screen-dropping resistor and C_{460} its bypass. The boosted B+ voltage of 660 volts is used for the plate of the output tube, the vertical deflection circuit, through the isolating resistor R_{467} and the kinescope screen grid.

In the plate circuit of the output tube, the top winding of the autotransformer T_{403} is the tertiary supplying positive flyback pulses for the plate of the high-voltage rectifier. The primary winding for the amplifier plate consists of all turns between terminals 5 and 3. Note that terminal 3 at the low end is the source of boosted B+ voltage of 660 volts. C_{466} is the boost capacitor. It is in series with terminal 3 and B+ connected to the damper plate. The damper cathode is tapped at terminal 6. This negative voltage during trace makes the damper diode conduct to charge C_{466}. The 6-μh choke L_{403} prevents radiation of horizontal pulses by the peak damper current.

Terminals 6, 7, and 8 on the autotransformer connect to the horizontal coils in the yoke, through pins 3, 4, and 5 on the yoke plug. Pin 4 is a center-tap connection. In this circuit, the a-c deflection voltage for boost is higher than the yoke voltage by the amount of voltage across terminals 8 and 3 on the transformer. Finally, the extra winding at the bottom of T_{403} supplies positive flyback pulses for the plate of the keyed AGC stage. A separate winding is necessary so that one end can be grounded.

All waveshapes in the horizontal circuits are obtained with the oscilloscope internal sweep at 15,750/2 cps. Although not shown here, the positive flyback pulses at the plate of the output tube can be observed by clipping the oscilloscope input cable to the insulation of the wire to the plate cap.

18·12 Transistorized horizontal deflection

The block diagram in Fig. 18·22 illustrates some circuit requirements for the low-voltage high-current characteristics of transistors. First, the horizontal deflection coils have very low impedance, in order to prevent induced voltages that are too high for typical transistor ratings. Note the scanning coils are in parallel, with a combined inductance of approximately 80 μh. The required scanning current equals 15 amp peak to peak, for a 90° kinescope and 15 kv anode voltage.

The yoke coils are impedance-coupled to the power output stage TR3, without any voltage step-up. However, the high-voltage winding L_2 has a

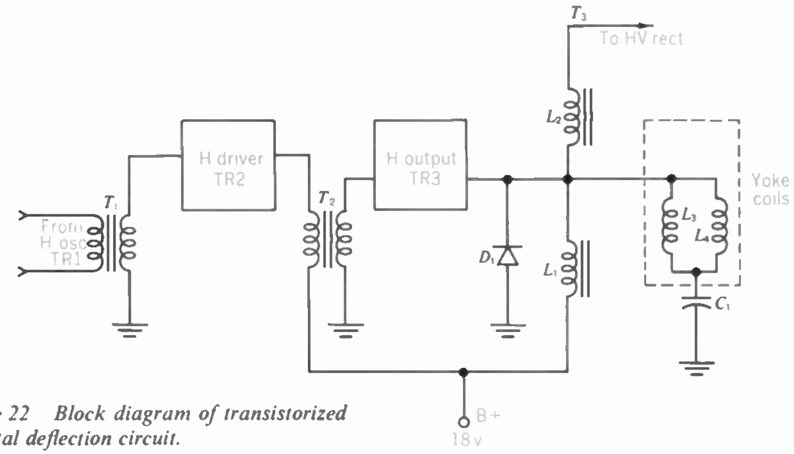


Fig. 18·22 Block diagram of transistorized horizontal deflection circuit.

step-up ratio of 100:1. For 15 kv across L_2, the voltage across L_1 equals 150 volts. Two transistors in series may be necessary then for the power output stage. Each divides the primary voltage to reduce the inverse voltage at the collector. The driver TR2 is a buffer stage for the oscillator, while supplying the relatively large amount of power needed to drive the output stage. Transformer coupling is used between the transistors to match the collector output impedance to the low impedance of the base-emitter circuit.

Damping is less of a problem because the transistor reverse resistance is much lower than for vacuum tubes. However, the silicon diode D_1 is used as a damper to improve linearity. Boosted B+ voltage is not necessary in the transistorized circuit. C_1 is a blocking capacitor, which also improves linearity of the scanning current. The high-voltage rectifier is usually a vacuum tube, but silicon diode units are available with the required peak inverse voltage rating.

18·13 Troubles in the horizontal deflection circuits

This section of the receiver causes troubles in the width of the raster. In addition, no brightness on the kinescope screen results from no horizontal scanning because of the flyback high-voltage supply.

It should be noted that excessive average d-c plate current can cause frequent failure of the horizontal output tube. The average cathode current can be checked with a d-c milliammeter for a reading within the recommended ratings. Also, do not remove the oscillator when you want to disable the horizontal output temporarily. Without the grid-leak bias produced by oscillator drive, the output tube can have excessively high plate current, especially if there is no safety bias in the cathode circuit. The best way is to remove the damper or the plate cap of the amplifier.

Insufficient width. The common case of too much black space at the left and right sides of the screen, without any white fold-over bars, indicates insufficient horizontal output (see Fig. 18·23). This trouble is often caused by a weak horizontal output tube. Also, the screen-grid voltage may be too low, or insufficient bypass capacitance allows degeneration to reduce

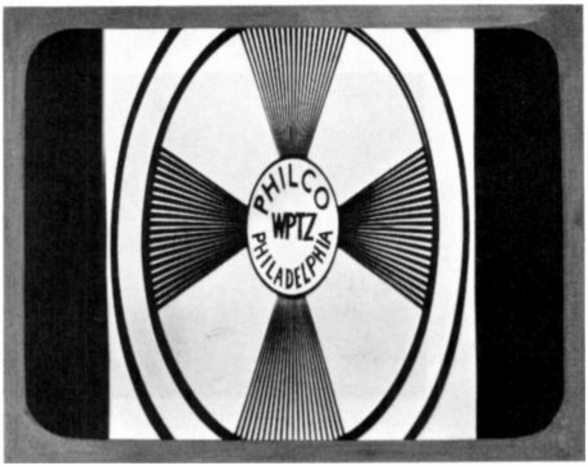


Fig. 18·23 Insufficient width. (Philco Corporation.)

the gain. A leaky bypass capacitor in the screen-grid circuit will have 50 volts or more of sawtooth ripple, instead of 10 volts or less of parabola waveshape.

Usually, insufficient B+ from the low-voltage supply reduces both width and height, resulting in a small raster. In some receivers, however, just the width may be affected because the horizontal deflection circuits require more power.

When the width of the raster is reduced but with distortion at the left, this effect indicates trouble in the damper circuit (see Fig. 18·24). This trouble is caused by an open B+ boost capacitor. With an open damper tube, however, the trouble is no brightness because the output tube plate circuit must return to B+ through the damper.

White vertical bars. Bright bars at the left side are caused by oscillations in the horizontal scanning current at the start of trace, immediately after flyback. The trouble can be localized by noting whether the bars are wide or thin. In Fig. 18·25, the thin bars result from excessive ringing that cannot be completely damped, although the damper circuit is normal. The

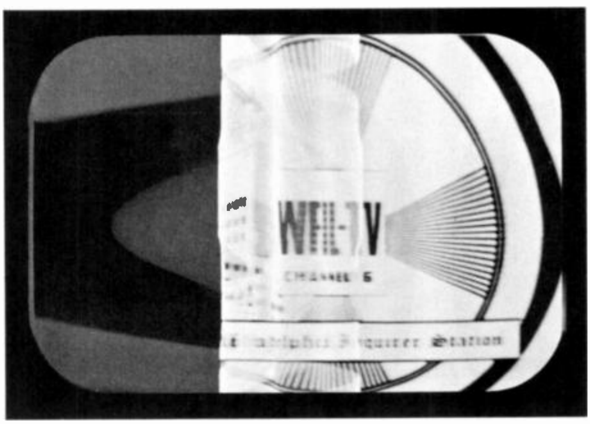
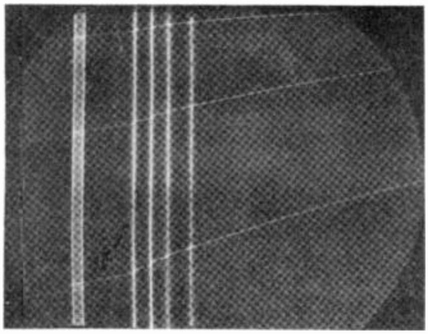
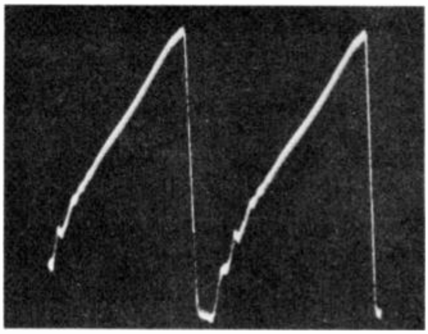


Fig. 18·24 Open B+ boost capacitor. (RCA.)





(a) (b)

Fig. 18·25 (a) Excessive horizontal drive. (b) Ripples in scanning current. (RCA Institutes, Inc.)

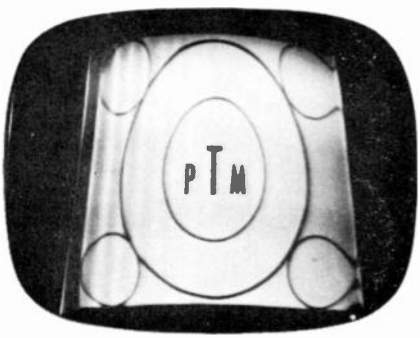
width of the raster is not reduced in this case. Notice that the bars are bright where the oscillatory ripples in the scanning current make the electron beam scan the same area more than once. The excessive ringing that produces thin bright bars can be caused by too much grid drive for the output stage, or troubles in the damping network for the yoke. Figure 18·26*a* shows excessive yoke ringing. Although some ringing at the left side is normal for most receivers, the bright bars should not be obvious when the picture is on the raster.

A wide bright bar at the left, as in Fig. 18·24, indicates trouble in the damper circuit. In this case, the trouble is an open boost capacitor. Such a bar corresponds to a high-amplitude oscillation that should be reduced by the damper. Note the reduced width because the damper is not supplying its normal current.

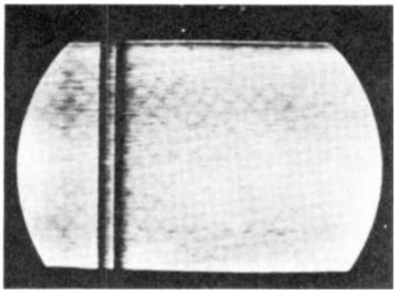
A bright bar at the right edge of the raster can result from compression of sawtooth deflection current at the finish of trace. This time is when the output tube conducts peak plate current.

Fold-over. This effect results when the same area on the kinescope screen is scanned more than once during a scanning cycle. With picture informa-

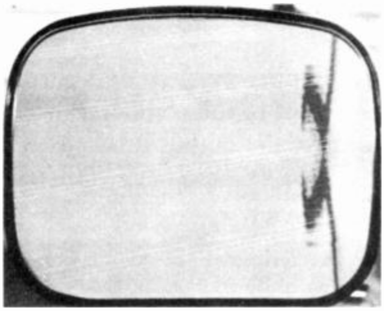
Fig. 18·26 (a) Excessive yoke ringing. (b) Horizontal keystoning caused by defective yoke. (Triad Transformer Corp.)



(a) (b)



(a)



(b)

Fig. 18 · 27 (a) Barkhausen-oscillation black lines at left. (b) Black "snivets" at right. (RCA.)

tion on the raster, part of the image appears folded over or under itself, as though the folded part were wrapped around a cylinder (see Fig. 18 · 24). This case illustrates fold-over in the raster, which is also evident in the picture. Fold-over at the left indicates trouble in the damper circuit. Fold-over at the right indicates trouble in the output tube circuit.

When the fold-over can be seen in the picture but is not in the raster this indicates either incorrect phasing between flyback and blanking or a retrace time longer than blanking time. Some receivers use internal horizontal blanking to eliminate fold-over at the left and right edges when flyback is more than 10 μsec.

Black lines. See Fig. 18 · 27*a*. This effect is caused by spurious oscillations produced in the horizontal output tube. The lines are at the left because the oscillations occur immediately after flyback when the output tube plate voltage is low. Then the screen grid serves as the anode for VHF *Barkhausen* oscillations. The bars are black because the oscillations are radiated to the signal circuits in the receiver with enough amplitude to drive the kinescope grid voltage to cutoff. In some receivers a small magnet is mounted on the output tube to prevent the internal electron oscillations. Or, the suppressor grid may have about +45 volts applied.

The black "snivets" at the right in Fig. 18 · 27*b* are caused by radiation of harmonics of the horizontal scanning current. R-f chokes of 1 to 10 μh may be used in the output tube circuit to prevent snivets at the right. A choke is generally used in the damper tube circuit to prevent snivets at the left. In general, the snivets result from sharp changes in scanning current.

Keystoned raster. An open or shorted deflection coil in the yoke produces keystoning or trapezoidal raster. This effect is shown in Fig. 18 · 26*b* for horizontal keystoning. It is important to note that any geometrical distortion of the raster can be caused only by the yoke or its associated damping network, since the yoke provides the symmetry in deflection. These effects include vertical or horizontal keystoning and pincushion or barrel distortion.

No high voltage. This is the most common cause of no brightness, while the sound is normal. Either the high-voltage rectifier is defective, or there is no high voltage from the horizontal output transformer. Remember that the damper must be operating for the output tube to supply high-voltage a-c input to the rectifier. Also, the horizontal oscillator must be operating to drive the output tube.

SUMMARY

1. The horizontal deflection circuits include the oscillator driving a beam-power amplifier with its diode damper. In addition, the flyback high-voltage supply is in the horizontal output circuit.
2. The oscillator output voltage at 15,750 cps provides the amplifier grid voltage needed for sawtooth plate current in the output.
3. Autotransformer coupling is generally used from the plate of the output tube to the horizontal deflection coils. The amplifier conducts for two-thirds the cycle to scan the right side of the raster.
4. At the end of trace, the output tube is cut off. Then the output circuit oscillates. The damper is cut off by the first half cycle of oscillations, allowing a fast flyback.
5. The high-voltage rectifier conducts during flyback to produce anode voltage for the kinescope.
6. Immediately after flyback, the damper conducts. As the damped current decays to zero, the left side of the raster is scanned while the amplifier remains cut off. Then the output tube starts conducting again to finish the trace.
7. While the damper is conducting, its current charges the B+ boost capacitor. The boosted B+ voltage is the plate supply for the horizontal amplifier.
8. The voltage waveshape at all points in the horizontal output circuit is rectangular with sharp flyback pulses. For the usual case of an autotransformer, all taps have positive polarity of flyback pulses. Therefore, the diode damper is inverted, with positive pulses at the cathode for cutoff during flyback but negative deflection voltage for conduction during trace.
9. The frequency of the horizontal scanning current is set by the oscillator and its AFC adjustments. The amplitude adjustments to fill the width of the raster are in the horizontal output circuit. These may include drive, width, and linearity controls.
10. The horizontal amplifier operates with grid-leak bias, approximately equal to the peak positive drive from the oscillator.
11. A typical value of sawtooth current is 1,200 ma peak to peak in 20-mh horizontal yoke coils. The induced voltages can be calculated as equal to $L\,di/dt$.
12. In transistorized horizontal deflection circuits, more scanning current is needed at lower voltage for the required power output.
13. Insufficient width without distortion indicates weak horizontal output. Reduced width with fold-over at the left is caused by trouble in the damper circuit.
14. Thin white bars at the left indicate excessive drive. Damping trouble causes a wide white bar at the left.
15. A defect in the yoke causes unsymmetrical deflection, producing a keystoned raster.
16. The most common cause of no brightness on the kinescope screen is no high voltage. Either the rectifier circuit is defective or there is no horizontal output. Check the horizontal oscillator, amplifier, and damper.

SELF-EXAMINATION *(Answers at back of book.)*

Choose (*a*), (*b*), (*c*), or (*d*).

1. Which stage is not necessary for producing horizontal output? (*a*) Horizontal oscillator; (*b*) damper; (*c*) horizontal amplifier; (*d*) horizontal AFC.
2. The frequency of sawtooth plate current in the horizontal amplifier is: (*a*) 60 cps; (*b*) 10,500 cps; (*c*) 15,750 cps; (*d*) 70 kc.
3. When the horizontal amplifier is conducting peak plate current, the electron scanning

beam is at the: (*a*) left edge of raster; (*b*) right edge of raster; (*c*) center of trace; (*d*) center of flyback.

4. Which of the following is conducting during flyback? (*a*) High-voltage rectifier; (*b*) output stage; (*c*) inverted diode damper; (*d*) conventional diode damper.
5. With a resonant frequency of 50 kc for the horizontal output circuit, the flyback time equals: (*a*) 5 μsec; (*b*) 10 μsec; (*c*) 20 μsec; (*d*) 50 μsec.
6. With autotransformer coupling, the voltage waveshape at the cathode of the inverted diode damper is: (*a*) sawtooth; (*b*) trapezoidal; (*c*) rectangular with positive flyback pulses; (*d*) rectangular with negative flyback pulses.
7. With 60 volts peak-to-peak sawtooth drive at the grid of the horizontal amplifier, its grid-leak bias equals: (*a*) zero; (*b*) 10 volts; (*c*) 27 volts; (*d*) 80 volts.
8. The voltage induced across an 80-μh inductance by 15-amp current dropping to zero in 8 μsec equals: (*a*) 80 volts; (*b*) 150 volts; (*c*) 2,000 volts; (*d*) 15,000 volts.
9. The B+ boost capacitor charges when the: (*a*) high-voltage rectifier conducts; (*b*) beam retraces; (*c*) damper is cut off; (*d*) damper is conducting.
10. The damper tube has an open heater. The result is: (*a*) wide white bar at left; (*b*) thin white bars at left; (*c*) fold-over at right; (*d*) no brightness.

ESSAY QUESTIONS

1. What three stages in the receiver determine the width of the raster? What stages determine the height?
2. How is the horizontal oscillator synchronized at 15,750 cps?
3. Will there be any high voltage if the horizontal oscillator operates at 15,810 cps? Explain.
4. What is meant by ringing? What two factors determine the ringing frequency?
5. Give six functions of the horizontal output circuit.
6. In horizontal deflection, describe briefly how the scanning current is produced for the right side of trace, the retrace, and the left side of trace.
7. Why is the damper cut off for the flyback? How is an inverted diode cut off?
8. What two voltages add to produce the boosted B+ voltage?
9. What three stages must be operating to produce a-c input voltage for the high-voltage rectifier?
10. Why is a smaller sawtooth capacitor used for the horizontal oscillator, compared with the vertical oscillator?
11. What is the polarity of flyback pulses for the plate of a keyed AGC stage?
12. Draw the schematic diagram of a horizontal amplifier circuit with autotransformer output. Include the damper and high-voltage rectifier circuits and show boosted B+.
13. Name three circuits that may use boosted B+ as a voltage source.
14. Describe briefly one method of removing the picture to see the raster alone.
15. Give two methods of checking with a d-c voltmeter to see if the horizontal oscillator is operating.
16. Draw a sine wave of current and of voltage on one graph, showing the two waveforms 90° out of phase. State the main characteristic of the 90° phase difference.
17. What is the function of a horizontal drive control? Explain briefly how to adjust the drive.
18. Referring to Fig. 18·4, would you consider operation for this amplifier class A, B, AB, or C? Explain. Why does grid current flow?
19. Draw one cycle of sawtooth current in the horizontal deflection coils. Mark the trace and retrace periods in microseconds. Indicate which tubes in the output circuit are conducting for three parts of the cycle.
20. Draw a circuit showing boosted B+ of 550 volts, using an inverted diode damper with its plate returned to +250 volts.
21. Describe briefly three methods of width control.
22. Referring to Fig. 18·18: (*a*) What is the function of C_r? (*b*) Describe the effect in the raster when C_r is open. (*c*) When C_r is shorted. (*d*) What is the function of R_1 and R_2?
23. Give two causes of insufficient width.

24. Give two causes of thin white bars at the left.
25. Give one trouble causing a wide white bar at the left and reduced width.
26. Give one cause of a thin dark bar at the left.
27. How can you distinguish between fold-over in the raster and fold-over only in the picture?
28. Referring to Fig. 18 · 21, give the function of: C_{457}; R_{459}; C_{455}; R_{471}; C_{456}; C_{460}; and R_{466}.
29. Referring to Fig. 18 · 21, give the effect on the kinescope screen for the following troubles: (*a*) R_{457} open; (*b*) R_{455} open; (*c*) L_{401} open; (*d*) C_{466} open; (*e*) C_{460} shorted.
30. The fuse F_1 in Fig. 18 · 2 is open. Give the effect on raster, picture, and sound.

PROBLEMS *(Answers to odd-numbered problems at back of book.)*

1. (*a*) Calculate the time constant of a 0.005-μf coupling capacitor with a 1-megohm grid resistor. (*b*) How many times longer than the horizontal scanning period is this time constant? (*c*) Can grid-leak bias be developed when grid current flows?
2. How much grid-leak bias is produced for the following peak-to-peak values of grid drive: 30 volts, 40 volts, 80 volts, and 160 volts?
3. Referring to Fig. 18 · 4: (*a*) How much is the average plate current? (*b*) How much is the plate dissipation with 100 volts average plate voltage?
4. Redraw Fig. 18 · 4 with sawtooth grid voltage reduced to 40 volts peak to peak. Cutoff remains the same. (*a*) How much is the grid-leak bias? (*b*) What is the class of operation? (*c*) How much is the peak plate current? (*d*) For how much of the cycle does plate current flow? (*e*) How much is the average plate current? (*f*) How much is the plate dissipation with 100 volts average plate voltage?
5. Referring to Fig. 18 · 21, calculate: (*a*) power dissipated in R_{465} with 8 ma screen-grid current, (*b*) voltage across R_{457} with 20 ma cathode current.
6. Refer to tube manual for average plate characteristics of 6DQ6 or similar horizontal output tube. (*a*) List values of plate current for each 5 volts of negative grid voltage from zero to cutoff. Assume a constant plate voltage of 300 volts. (*b*) Draw a graph plotting these values of plate current on the vertical axis against grid voltage on the horizontal axis.
7. How much capacitance resonates with 40 mh inductance at 80 kc?
8. Referring to Table 18 · 1, assume 1 amp peak-to-peak sawtooth current for each of the four examples of horizontal coils shown. Calculate the flyback voltage across each, for a retrace time of 8 μsec. Do the same for the vertical coils listed, assuming 0.5 amp peak-to-peak sawtooth current and a vertical retrace time of 300 μsec.
9. For the table below, indicate which stages are on or off during retrace and any part of trace.

	Trace	*Retrace*
Horizontal amplifier		
High-voltage rectifier		
Damper		

Chapter 19 *Picture i-f amplifiers*

As shown in Fig. 19 · 1, the superheterodyne television receiver generally uses three i-f stages to amplify the signal output of the mixer stage in the r-f tuner. Although the sound i-f signal is amplified along with the picture i-f signal in the common i-f section for intercarrier-sound receivers, this does not change the requirements of picture i-f amplification. With 0.5 mv picture i-f signal from the mixer, the i-f amplifier can provide an overall voltage gain of 10,000 for 5 volts peak amplitude into the video detector. Practically all the receiver gain and selectivity is obtained in the i-f section. Also, the high-frequency emphasis caused by vestigial-side-band transmission is corrected in the receiver by the frequency response of the i-f amplifier.

19 · 1 Picture i-f response

The graph in Fig. 19 · 2 compares how much overall voltage amplification the i-f section has for different frequencies in the pass band. Note the following features of this ideal i-f response curve.

1. The picture i-f carrier frequency is 45.75 Mc, corresponding to the r-f carrier frequency for all stations tuned in by the front end. The sound i-f carrier frequency at 41.25 Mc is separated by 4.5 Mc from the picture i-f carrier. In older receivers, 25.75 and 21.25 Mc are the i-f picture and sound carrier frequencies.
2. The picture i-f carrier frequency at the side of the response curve has 50 per cent of maximum i-f gain. This reduced i-f response for the picture carrier, and side frequencies close to it, is opposite to the effect of vestigial-side-band transmission.
3. The sound i-f signal at 41.25 Mc is amplified with the picture signal for intercarrier sound. However, the relative gain is only 5 to 10 per cent for the sound signal in the picture i-f amplifier section.

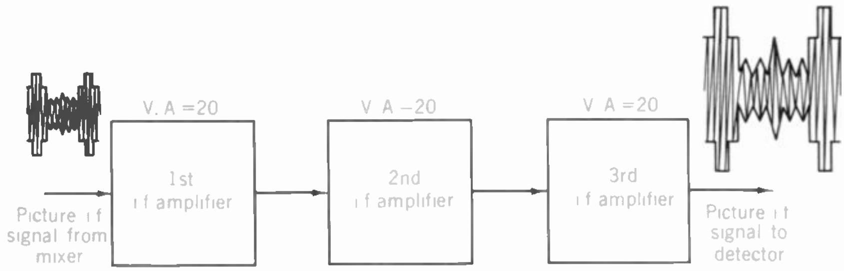


Fig. 19·1 The picture i-f stages amplify the signal from the mixer to supply enough signal for the video detector. Overall voltage amplification here equals 8,000. Typical miniature glass pentodes used are 6AG5, 6AU6, 6CB6, 6CF6, 6BZ6, and 6GM6.

4. The ideal bandwidth of 4 Mc corresponds to 4 Mc response for the video modulating frequencies, allowing maximum detail in the reproduced picture. In typical receivers the i-f bandwidth may be as little as 2.5 Mc or as much as 4 Mc. Less bandwidth allows more gain per stage. Usually, the i-f bandwidth is limited to 3.2 Mc in monochrome receivers, to prevent interference from the 3.58-Mc color subcarrier signal transmitted for programs broadcast in color.

Inversion of i-f side frequencies. As an example, the r-f sound carrier frequency is transmitted 4.5 Mc higher than the r-f picture carrier frequency but in the receiver i-f section the i-f sound carrier at 41.25 Mc is 4.5 Mc lower than the picture carrier at 45.75 Mc. This frequency inversion results only because the local oscillator in the tuner beats above the r-f signal frequencies, which is the usual case. Then higher r-f frequencies are closer to the oscillator frequency, resulting in lower values for the i-f difference frequencies. Additional examples are listed in Table 19·1.

The same frequency separation is maintained between the picture carrier and its side frequencies in the r-f and i-f signals. However, those frequencies above the picture carrier in the r-f signal are below in the i-f signal. Also, the lower r-f side band becomes an upper i-f side band.

How the i-f side frequencies correspond to video frequencies. The highest video modulating frequencies produce side frequencies farthest from the picture carrier in the transmitted signal. Although the i-f side bands are inverted, the same frequency difference compared with the carrier is maintained. Therefore, the i-f side frequencies farthest from the picture carrier correspond to the highest video frequencies in the output of the video detector. In the i-f response curve, these frequencies have lower

Table 19·1 Inversion of i-f signal frequencies

Channel 4, 66–72 Mc	*Transmitted r-f signal frequency, Mc*	*Local oscillator frequency, Mc*	*Intermediate frequency, Mc*
Upper edge of channel	72	113	41
Sound carrier	71.25	113	41.25
Side frequency for 2 Mc video modulation	69.25	113	43.75
Picture carrier	67.25	113	45.75
Lower edge of channel	66	113	47

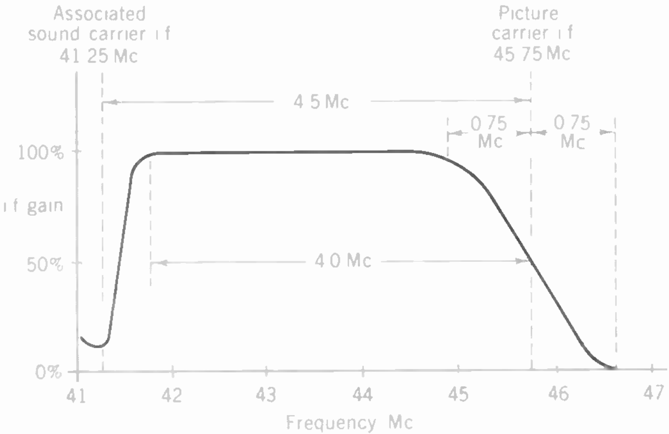


Fig. 19·2 Desired i-f response curve, for intercarrier-sound receiver with picture i-f carrier of 45.75 Mc.

numerical values, near the sound i-f carrier. However, the i-f gain for these frequencies determines the response for the high video frequencies that reproduce fine detail in the picture. At the opposite part of the curve, side frequencies close to the picture carrier represent the low video frequencies. The continuity between r-f, i-f, and video signal frequencies is illustrated by the resolution chart in Fig. 19·3.

In terms of frequency components in the detected signal, we can consider that each i-f side frequency beats with the picture carrier in the video detector to produce the difference frequency. As an example, 42.75 Mc beats with 45.75 Mc in the i-f signal to produce the difference frequency of 3 Mc in the video signal out of the detector. Therefore, 4 Mc bandwidth from 41.75 to 45.75 Mc in the i-f curve corresponds to 4 Mc response for the detected video signal, from 0 cps or direct current up to 4 Mc.

Compensation for vestigial-side-band transmission. The lower video modulating frequencies up to 0.75 Mc are transmitted as double-side-band signals, while the higher video frequencies are transmitted as single-side-

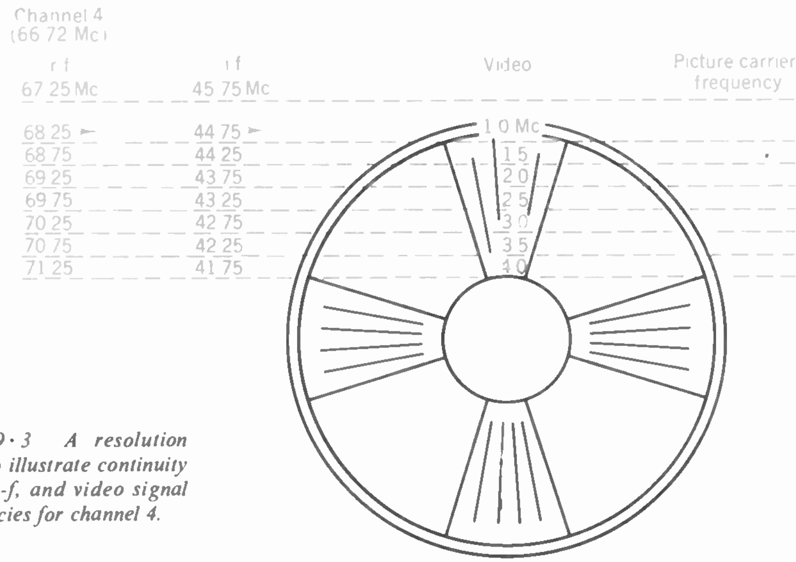


Fig. 19·3 A resolution chart to illustrate continuity of r-f, i-f, and video signal frequencies for channel 4.

band signals, as explained in Sec. 5 · 2. Therefore, video frequencies above 0.75 Mc have only one-half the effective modulation of the lower frequency double-side-band signals. If the receiver i-f response were the same for all signal frequencies, the demodulated output from the video detector would be twice as great for video signals below 0.75 Mc as for the higher video-frequency signals.

In order to equalize the effect of vestigial-side-band transmission, the overall i-f response is aligned to give the picture carrier approximately 50 to 60 per cent of maximum response, as shown in Fig. 19 · 2. As a result, the two side bands of a double-side-band signal are given an average response of 50 per cent, compared with 100 per cent response for the single-side-band frequencies. Therefore, the output from the video detector will be the same for all video modulating signals having the same amplitude, whether they are transmitted with single or double side bands.

When the picture carrier is too high on the i-f response curve, with more than 50 to 60 per cent relative gain, the low video frequencies up to 0.75 Mc are emphasized, while higher video frequencies are attenuated. The extra gain for low video frequencies increases the contrast in the reproduced picture but usually with excessive smear. Attenuation of the high video frequencies reduces the high-frequency detail that makes the picture sharp and clear. When the i-f gain is too low for the picture carrier frequency, the reduced response attenuates low video frequencies and the picture carrier itself. The result is insufficient video signal and weak contrast in the picture. With zero response for the i-f picture carrier there would be no picture at all.

Associated sound signal. This is the sound associated with the picture being received. The two carrier frequencies are always separated by 4.5 Mc. In an i-f amplifier with 4 Mc bandwidth for the picture signal, therefore, sharp cutoff is needed at the sound carrier side of the picture i-f response curve. The required attenuation of the sound signal in the picture i-f amplifier is obtained by wave traps tuned to the sound i-f carrier frequency. In a split-sound receiver, the sound traps reject the sound i-f signal to produce practically zero response. With intercarrier-sound receivers, the response of the sound i-f signal in the common i-f amplifier is approximately 5 per cent. This is the relative strength required to obtain the 4.5-Mc sound signal as the beat frequency between the picture and sound carriers in the second detector. See Fig. 19 · 4.

19 · 2 The intermediate frequency

The picture i-f carrier frequency has progressed from 12.75 Mc in the first commercial television receivers, up to 25.75 and 45.75 Mc, which is now the standard value specified by EIA. The sound i-f carrier frequency is automatically 4.5 below. In general, high values for the intermediate frequencies are desired, for the following reasons:

1. Better image rejection. The image frequency equals twice the i-f plus

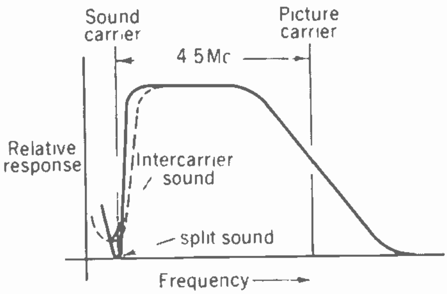


Fig. 19·4 Sound i-f response for split-sound and intercarrier-sound receivers.

the desired r-f signal frequency. With a higher i-f, the image is separated from the r-f signal by a greater amount.

2. Less signal is coupled from the local oscillator, through the r-f amplifier, to the antenna circuit, reducing interference caused by oscillator radiation. With a higher i-f, there is a greater separation between the resonant frequencies for the oscillator and r-f signal circuits.
3. The bandwidth required for the picture i-f stages can be obtained more easily. At higher frequencies, the bandwidth is a smaller percentage of the resonant frequency.
4. Filtering is easier in the video detector output, to remove the i-f variations from the desired video signal. With a higher i-f, there is a greater separation between the video and intermediate frequencies.

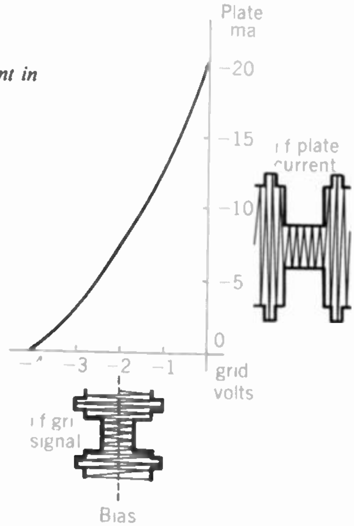
In any case, the intermediate frequency should not be close to a band of frequencies assigned to other services. They can produce excessive interference if these signal frequencies or their harmonics are in the i-f pass band of the television receiver. Also important is the position of the image frequency in the frequency spectrum when the receiver is tuned to different channels. If the image is in another television channel, or in the FM broadcast band of 88 to 100 Mc, these signal frequencies can produce r-f interference in the reproduced picture.

Another factor is the local oscillator frequency for each channel. If the oscillator is in the frequency range of a television channel, a receiver operating on one channel can produce r-f interference in a nearby receiver tuned to a higher channel. Consideration of these factors has led to the present standard values of 45.75 and 41.25 Mc for the picture and sound i-f carrier frequencies in television receivers. Important advantages of these i-f values are that, for any VHF channel from 2 to 13: (1) one VHF channel cannot be an image of another VHF channel; (2) any frequency in the FM broadcast band of 88 to 108 Mc cannot be an image of a television channel; (3) the local oscillator frequency cannot be in any VHF channel; (4) less interference is received from the amateur band of 21 to 24.5 Mc.

19·3 I-F amplification

Each i-f stage is a tuned amplifier, generally using a miniature glass pentode. The amplifier operates class A for minimum distortion of the modulation envelope. As illustrated in Fig. 19·5, the modulated picture i-f

Fig. 19·5 I-f grid signal varying plate current in i-f amplifier operating class A.

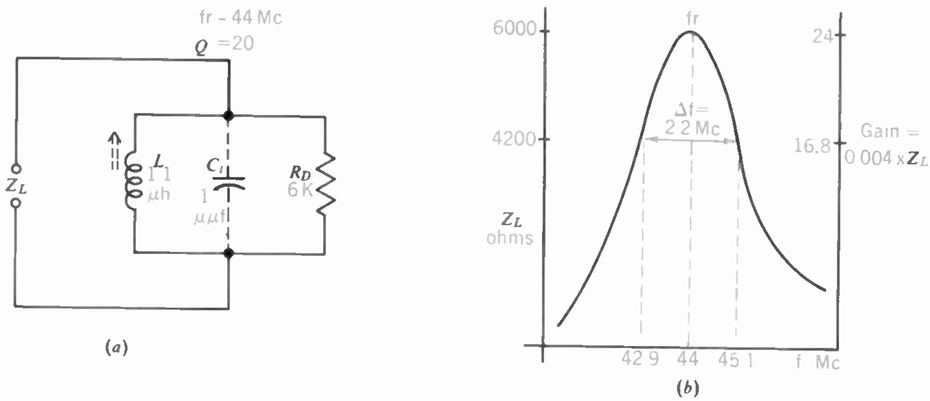


carrier signal varies the amplifier grid voltage to control the plate current variations. The peak i-f signal voltage at the grid of the last i-f stage is about 1 volt or less.

Either a double-tuned transformer may be used for coupling the i-f signal between stages, as in Fig. 19·8, or single-tuned amplifiers as in Fig. 19·10. In either case, each i-f stage has a tuned circuit as the plate load impedance for a-c signal. Therefore, the amplifier has gain only for the i-f signal frequencies, because the tuned circuit has its highest impedance at parallel resonance. An external shunt damping resistor is usually connected across the tuned circuit to obtain the required bandwidth. This is the function of R_D in Fig. 19·6*a*. To provide a high L/C ratio in the tuned circuit, the inductance resonates with the stray shunt capacitances. For alignment, the coil has an adjustable slug to tune each circuit to the required resonant frequency.

Tuned plate load impedance. Some important characteristics of tuned amplifiers can be illustrated by the resonant circuit in Fig. 19·6. For a single-tuned stage, this circuit is the load impedance for alternating signal current in the amplifier plate circuit. It is an example of parallel resonance because the tube is the signal source outside the tuned circuit. The resonant frequency f_r is $1/(2\pi\sqrt{LC_t})$. Note that C_t is not a lumped capacitance but combines the output capacitance of the plate circuit and the

Fig. 19·6 Parallel resonant circuit as tuned plate load impedance. (a) Single-tuned circuit with shunt damping resistance R_D. (b) Resonance curve with maximum impedance at resonant frequency f_r.



input capacitance of the grid circuit in the next stage. Typical values are 4 $\mu\mu f$ plus 8 $\mu\mu f$ for a total of 12 $\mu\mu f$. The required value of L is 1.1 μh for the f_r of 44 Mc. The reactance of either L or C_t at 44 Mc is approximately 300 ohms. This is $2\pi fL$, with 44 Mc for f and 1.1 μh for L.

The tuned circuit provides a resonant rise in impedance, as shown by the response curve in *b*. Specifically, the maximum impedance at parallel resonance equals the Q of the tuned circuit times either reactance. With a Q of 20 for the damped circuit, Z_L is 20 × 300, which equals 6,000 ohms.

Shunt damping resistance. The i-f coil generally has a Q of 80 to 100 by itself. However, the Q of the tuned circuit must include the effect of the shunt damping resistance. R_D lowers the Q, as its resistance allows a resistive branch current that cannot be canceled by the opposite reactances at resonance. For the damped tuned circuit,

$$Q_D = \frac{R_D}{X_L} \tag{19·1}$$

This formula with parallel resistance is the reciprocal of the Q formula for the coil itself with series resistance. For the example here,

$$Q_D = \frac{R_D}{X_L} = \frac{6{,}000 \text{ ohms}}{300 \text{ ohms}}$$

or

$$Q_D = 20$$

Lower values of R_D lower the Q of the tuned circuit, which increases its bandwith, but decreases Z_L (see Fig. 19·7). The curve for infinite R_D means no damping resistance is used. Then the Q of the tuned circuit is the same as the Q of the coil, plotted here for a Q of 80. At the opposite extreme, zero resistance for R_D would be a short circuit, which would eliminate the resonance effect completely.

The value of 6,000 ohms for R_D in Fig. 19·6 is chosen for a Q of 20 to provide the desired bandwidth. R_D can be calculated by transposing Eq. (19·1) to $R_D = Q \times X_L$, where X_L is the reactance at f_r. In this example R_D is 20 × 300 ohms, or 6,000 ohms. Note that the value of R_D is the same as Z_L at resonance, which also equals $Q \times X_L$. This identity means that

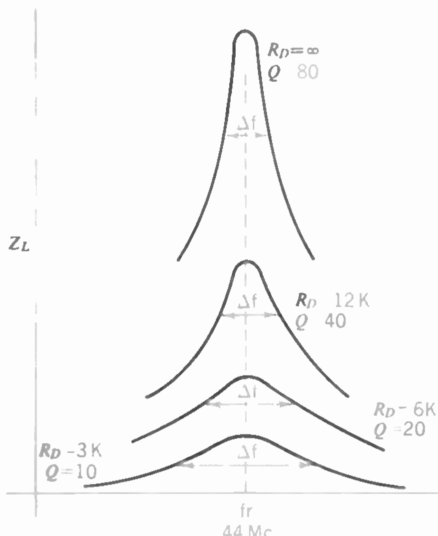


Fig. 19·7 Effect of changing shunt damping resistance R_D in Fig. 19·6.

the impedance of the shunt-damped resonant circuit at f_r is equal to the resistance of R_D, as the equal reactive branch currents cancel. However, below or above the resonant frequency, the reactive branch currents are not equal. Then the combined reactance reduces the parallel impedance to a value less than R_D.

Bandwidth. The total bandwidth of the resonance curve between 70.7 per cent response points, or the half-power frequencies, is

$$\Delta f = \frac{fr}{Q} \tag{19·2}$$

For the example here,

$$\Delta f = \frac{44 \text{ Mc}}{20}$$

or

$$\Delta f = 2.2 \text{ Mc}$$

In Fig. 19·6, Z_L is 70.7 per cent of maximum, or more, between the frequencies of 42.9 and 45.1 Mc for a total bandwidth of 2.2 Mc. These values are included in the pass band of the tuned amplifier. Frequencies outside this band have little gain because the *LC* circuit provides low impedance off resonance. Note that more bandwidth can be obtained with higher values of f_r, for the same Q. Also, lower values of Q increase the bandwidth. However, lower Q reduces the resonant impedance, which reduces the gain.

Voltage amplification. Since a pentode with high r_p is used with a relatively low Z_L for the i-f amplifier, its voltage amplification or gain equals g_mZ_L, as for a video amplifier. The plate load impedance Z_L is provided by parallel resonance of the *LC* circuit. The g_m value depends on the tube and its bias. High values of g_m are important for high gain with the relatively small Z_L needed for bandwidth.

A frequency response curve showing voltage amplification or gain of the tuned amplifier corresponds to the resonance curve of impedance Z_L. The reason is that the gain depends on Z_L of the parallel resonant circuit, which is multiplied by the g_m of the tube. In Fig. 19·6*b*, note the gain values at the right, equal to g_mZ_L, calculated for a g_m of 4,000 μmhos or 0.004 mho. For the 44 Mc f_r at the center of curve, Z_L equals 6,000 ohms and the gain is 0.004 mho $\times$ 6,000 ohms, which is 24. At the edges of the pass band, for 42.9 Mc and 45.1 Mc, the gain is 16.8, equal to 0.0004 $\times$ 4,200. The gain of 16.8 here is down 3 db from 24 because the Z_L of 4,200 ohms is 70.7 per cent of 6,000 ohms. The gain is still less for frequencies farther from f_r, as Z_L drops. At frequencies completely removed from resonance, there is practically no gain because Z_L is close to zero.

The amount of voltage gain depends on the value of g_m, which is varied by AGC bias. As an example, we can assume the gain of 24 for Z_L of 6,000 ohms and g_m of 4,000 μmhos is obtained with -4 volts AGC bias. These values are for f_r at 44 Mc in Fig. 19·6*b*. If the AGC bias changes to -5 volts and reduces g_m to 3,000 μmhos, the gain will drop to 18, for the same Z_L of 6,000 ohms. Similarly, the gain for other frequencies in the response curve will drop proportional to the reduced g_m value. Then the entire response curve has lower values of absolute gain, although the shape of the curve remains essentially the same.

The overall gain of cascaded stages equals the product of individual gain values. The value of 24 for the gain at 44 Mc in Fig. 19·6*b* is for one stage. If a second stage has a gain of 10 and a third stage a gain of 20 at this frequency, the overall i-f gain at 44 Mc equals $24 \times 10 \times 20$, or 4,800. Similarly, the overall i-f response curve is the product of the individual response curve for each of the cascaded stages.

19·4 Double-tuned i-f amplifiers

A typical circuit is shown in Fig. 19·8. The primary and secondary inductances L_p and L_s in the i-f transformer T_i tune the shunt capacitances C_{out} and C_{in} to the desired frequency in the i-f range. L_p and L_s each have a slug for aligning the primary and secondary tuned circuits. Both primary and secondary are resonant at the same frequency. The required bandwidth is provided by the shunt damping resistors R_4 and R_5, in addition to the close coupling used between L_p and L_s. The circuit components in Fig. 19·8 have the following functions. The decoupling resistor R_1 isolates the tuned circuit from the common B+ line, while supplying +120 volts for plate and screen. Note that C_1 has the dual functions of screen-grid bypass and ground return to cathode for the primary tuned circuit. R_2C_2 is a decoupling filter to isolate the grid circuit from other stages connected to the AGC line for bias. Similar to C_1, the bypass C_2 is the ground return for the secondary tuned circuit. The cathode resistor R_k is unbypassed in order to provide degeneration. This effect stabilizes the input capacitance

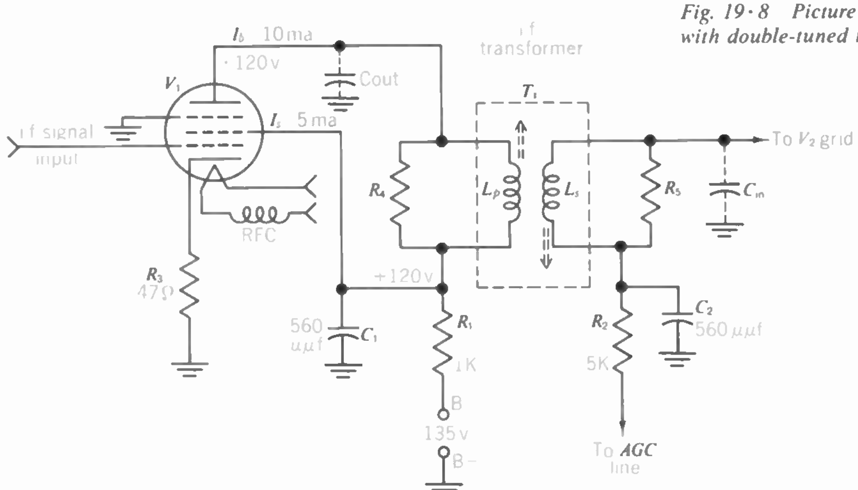


Fig. 19·8 Picture i-f amplifier with double-tuned transformer.

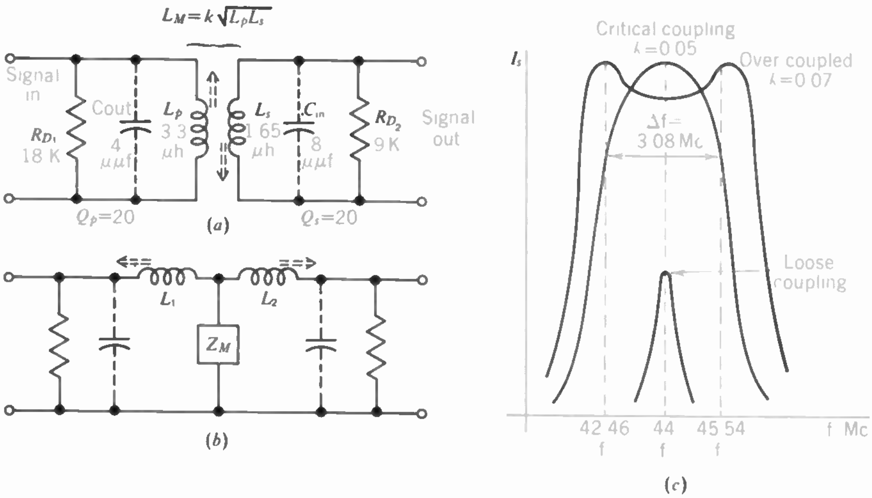
and resistance against changes of bias to minimize detuning when the AGC voltage varies. The heater choke prevents coupling of cathode signal between stages through the common heater line. The suppressor grid is grounded externally at the socket to reduce regeneration.

Note the d-c voltages in Fig. 19·8. With 10 ma average plate current and 5 ma screen-grid current the total through R_1 is 15 ma. Therefore, the voltage drop across R_1 equals $0.015 \times 1{,}000$, or 15 volts. This value subtracted from 135 volts equals 120 volts for the screen-grid and plate-supply voltage. The voltage at the plate is also 120 volts because the i-f coil has practically zero d-c resistance. It is the resonance in the i-f transformer that provides the plate load impedance for a-c signal.

Double-tuned coupling. More details of resonance in the double-tuned coupling arrangement are shown in Fig. 19·9. Both primary and secondary are tuned to the same frequency, which is 44 Mc here for comparison with the single-tuned circuit of Fig. 19·6. Coupling in the double-tuned transformer in *a* results from the mutual inductance L_M that links L_p and L_s, providing a mutual coupling impedance common to both circuits. Similar results can be obtained with other types of mutual coupling impedance, indicated by Z_M in *b*. Physically, Z_M can be a coupling coil or capacitor, or combine both to function as a wave trap in addition to the mutual coupling impedance between primary and secondary circuits.

To tune primary and secondary to 44 Mc in Fig. 19·9*a*, L_p is 3.3 μh and L_s 1.65 μh. The reactance of L_p at 44 Mc equals 900 ohms. Therefore, the primary circuit Q_p is 20, equal to its R_D/X_{L_p}. The secondary L_s has a reactance of 450 ohms at 44 Mc. Similarly, its value of R_D/X_{L_s} equals 20 for Q_s. Note that splitting the total shunt capacitance into two smaller

Fig. 19·9 Double-tuned coupling. (a) Transformer with L_p and L_s tuned to f_r. (b) The general form of double-tuned circuit with mutual coupling impedance Z_M. (c) Effect of coupling on response of double-tuned transformer.



parts allows more inductance with a higher L/C ratio for the same resonant frequency. Then higher values of R_D can be used for the same Q. Equal values of Q are used in this example, but the primary and secondary need not have the same Q.

Coefficient of coupling. The effect of coupling in a double-tuned transformer is illustrated in Fig. 19·9*c*. The values of coupling coefficient k indicate the fraction of total flux from one coil linking the other coil. For loose coupling, the primary and secondary are far apart with little flux linkage between L_p and L_s. The secondary response then has a single peak at f_r like a single-tuned circuit, with low amplitude. As the coupling is increased by having L_p and L_s closer together, the primary produces more secondary output. Also, the secondary current has a greater effect on the primary. With close coupling, two peaks result in the secondary output to broaden the frequency response. Although L_p and L_s are both tuned to f_r, the coupled reactance between them produces separate peaks at two different frequencies f_1 and f_2, below and above the resonant frequency of each circuit by itself. However, the response dips for f_r at the center. If the coupling is increased still further the two peaks will spread farther apart with a more pronounced valley for f_r. Note that, for the critical value of coupling k_c, the secondary has its maximum amplitude and greatest bandwidth without double peaks.

Critical coupling. The coefficient of coupling required for maximum gain at f_r without double peaks is

$$k_c = \frac{1}{\sqrt{Q_pQ_s}} \tag{19·3}$$

In this example, $$k_c = \frac{1}{\sqrt{20 \times 20}} = \frac{1}{20}$$

or $$k_c = 0.05$$

This value of coupling means 5 per cent of the flux in one coil links the other coil. Then the mutual inductance

$$L_m = k\sqrt{L_pL_s} = 0.05\sqrt{3.30} \times 1.65 = 0.12\mu\text{h}$$

Loose coupling corresponds to less L_M; tight coupling produces more L_M. For the case of k equal to 0.07, L_M is then 0.17 μh.

At critical coupling, the response has a single peak at f_r with a bandwidth

$$\begin{aligned} \Delta f &= 1.4\, k_c \times f_r \qquad (19·4) \\ &= 1.4 \times 0.05 \times 44 \text{ Mc} \\ \Delta f &= 3.08 \text{ Mc} \end{aligned}$$

The bandwidth for this example is shown by the curve for critical coupling in Fig. 19·9*c*. Note that this bandwidth for critical coupling in the double-tuned transformer is 1.4 times greater than the bandwidth of the corresponding single-tuned circuit in Fig. 19·6.

The absolute gain at f_r with critical coupling is given by the formula

$$A_v = \frac{g_m \times Q \times X_L}{2} \qquad (19\cdot5)$$

where Q is $\sqrt{Q_pQ_s}$ and L is $\sqrt{L_pL_s}$. If L were the same as in the single-tuned circuit, the gain would be one-half for the double-tuned case. However, higher values of L can be used. In this example, L is approximately double the inductance of the single-tuned circuit. As a result, the gain is about the same for the two examples, but the critically coupled double-tuned transformer has 1.4 times more bandwidth.

Close coupling. When the coupling is more than k_c, the bandwidth increases but the gain is less for f_r. The result is a double-peak response with a dip in the middle. Note the curve for $k = 0.07$, equal to 1.4×0.05. For this case, when k is $1.4 \times k_c$, the two peaks on the overcoupled curve are at the same frequencies having 70 per cent response with critical coupling. In other words, the increased coupling moved the gain for f_1 and f_2 up to peak response, instead of being 3 db down. However, the dip of f_r is down about 10 per cent.

For the general case of close coupling but k still much less than one, the peaks are symmetrical and not far from f_r. Specifically, the frequency separation between peaks equals $f_r \times k$. For k of 0.07, the frequency separation is 44×0.07, which equals 3.08 Mc, as in Fig. 19·9*c*. If the coupling is increased, the peaks will move farther apart. Also, the dip at f_r becomes more severe. However, the dip at the center of an overcoupled response curve can be corrected by adding a single-tuned amplifier with its peak gain at f_r.

A special case of damping. The effect of Q on the double-tuned response can be illustrated by an example that has useful applications. Suppose that either the primary or secondary is damped with a parallel resistor of a few hundred ohms for a Q of 1. With Q values of 20 and 1, the value of k required for critical coupling is $1/\sqrt{20}$, or 0.22. However, the coupling

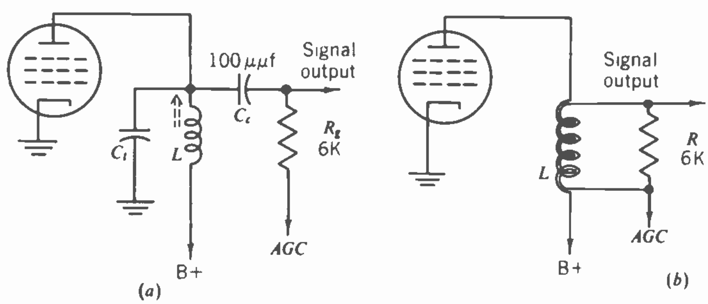


Fig. 19·10 Single-tuned picture i-f amplifiers. (a) With R_gC_c coupling. (b) With bifilar coil.

between L_p and L_s in the transformer is fixed at a value we can take as 0.07. This coupling normally produces double peaks in the response because k is more than k_c, without the extra damping. Now, though, the value of 0.07 provides loose coupling compared with a k of 0.22 needed for critical coupling. As a result, the extra damping gives the double-tuned transformer a single-peaked response at f_r. For the purpose of alignment, therefore, it is possible to peak each side of the transformer at f_r when a small damping resistor of about 330 ohms is temporarily shunted across the opposite side.

19·5 Single-tuned i-f amplifiers

In Fig. 19·10*a* the R_gC_c network couples i-f signal from the plate of the single-tuned amplifier to the input of the next stage. R_g also serves as a damping resistance effectively in parallel with the plate-tuned circuit because C_c has low reactance for the i-f signal frequencies. The parallel resonant circuit formed by L with the total shunt capacitance C_t is the plate load impedance for the amplifier to develop the output signal voltage. Its relative gain is the same as the single-tuned response curve in Fig. 19·6*b*. The numerical gain equals g_mZ_L.

The single-tuned amplifier in Fig. 19·10*b* does not require any coupling capacitor because of the bifilar winding of the i-f coil. As shown in Fig. 19·11, the bifilar coil is wound with twin conductor wires, each insulated from the other. One conductor is the plate winding connected to B+, while the other conductor is the grid winding. Since the i-f plate signal voltage is inductively coupled into the grid winding for the next stage, no

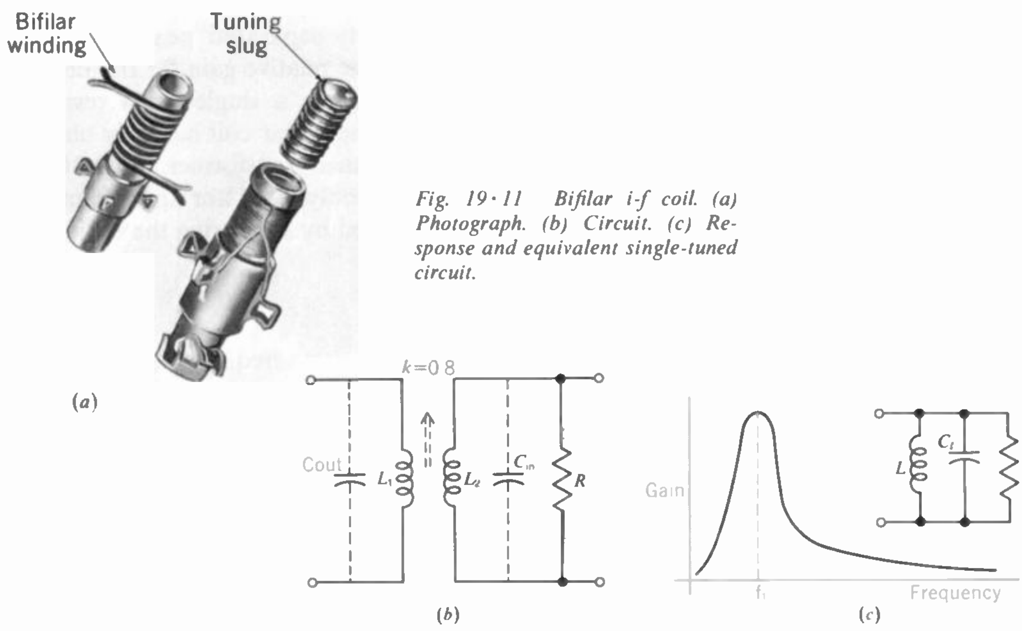


Fig. 19·11 Bifilar i-f coil. (a) Photograph. (b) Circuit. (c) Response and equivalent single-tuned circuit.

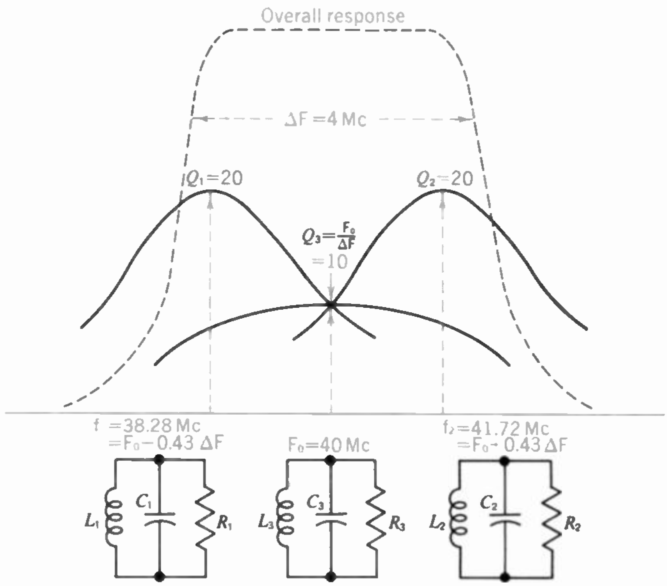


Fig. 19·12 Stagger tuning of three single-tuned circuits for overall bandwidth of 4 Mc, centered at 40 Mc.

coupling capacitor is needed. A grid load resistor is not required either, as the grid winding provides a d-c return path to cathode. However, R is used as a damping resistor for the tuned circuit.

The advantage of the bifilar coil is that it provides the response of a single-tuned amplifier, without the need for a coupling capacitor. Elimination of C_c is helpful because it can be charged by noise pulses, causing too much negative grid bias trailing after noise peaks in the signal. This effect can cause white tails at the right of horizontal black streaks produced in the picture by noise pulses.

The bifilar coil is actually an overcoupled transformer with k close to one. Then its response includes two widely separated peaks, with the lower frequency peak in the i-f pass band. The relative gain for frequencies around the peak is approximately the same as a single-tuned response curve, as shown in Fig. 19·11*c*. Note that the bifilar coil has only one adjustment, compared with two for a double-tuned transformer. Single-tuned stages, each with a bifilar coil, are commonly used for the picture i-f amplifier. The required bandwidth is obtained by staggering the individual resonant frequencies.

19·6 Stagger-tuned stages

When cascaded amplifiers are tuned to the same frequency (synchronous tuning) the overall bandwidth shrinks drastically. This effect occurs because the overall gain equals the product of individual gain values in each stage. Then the peak values become more peaked, while the low-gain values become more attenuated. The sharp peak with narrow bandwidth is undesirable for the wide pass band needed in the picture i-f amplifier. However, the required overall response can be obtained by staggering the resonant frequency of individual single-tuned stages. This procedure is illustrated in Fig. 19·12 for the example of three single-tuned stages,

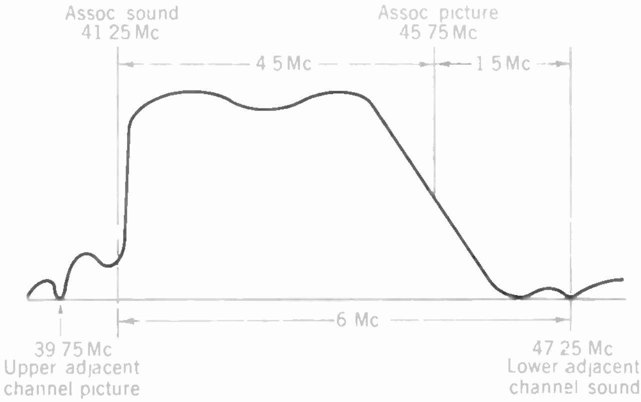


Fig. 19·13 Location of carrier frequencies for associated sound and picture, lower adjacent-channel sound, and upper adjacent-channel picture, on i-f response curve.

staggered around 40 Mc, with an overall bandwidth of 4 Mc.

The values in Fig. 19·12 are calculated as follows: One stage is tuned to the center frequency F_o of the desired overall response curve and has the same bandwidth (ΔF). Its required Q is $F_o/\Delta F$, which equals 40/4, or 10, in this example. Also, L_3 has the value required to resonate with C_3 at 40 Mc. The damping resistor R_1 provides the required Q of 10. The other two circuits have double the Q and one-half the bandwidth. Each is staggered at the resonant frequency of $F_o \pm 0.43\ \Delta F$. This means f_1 and f_2 are above and below F_o by 43 per cent of the bandwidth. The order in which the stages are staggered does not affect the overall response.

Staggered tuning can be used for pairs of stages, triplets, quadruplets, or quintuplets.[1] Also, either single-tuned stages or double-tuned stages can be staggered. For picture i-f amplifiers, a common arrangement is one double-tuned transformer coupling i-f signal from the mixer plate to the i-f amplifier, with three single-tuned stages in a staggered triplet.

There are several advantages in the staggered tuning. Since successive stages are resonant at different frequencies, the possibility of regeneration is reduced. Furthermore, alignment of staggered single-tuned stages is much easier, compared with double-tuned circuits. Each single-tuned circuit is simply peaked at its resonant frequency. Referring to Fig. 19·12, notice that, although each stage peaks at only one frequency, other signal frequencies in the i-f pass band are also applied. As a result, the signal generator used for alignment can be in the mixer grid circuit, set for the individual frequencies required to peak all the single-tuned stages. Since the individual responses are multiplied in the cascaded amplifier, its overall response has the required relative gain for all frequencies in the i-f pass band. The sharp attenuation needed at the edges of the response to reject undesired frequencies is provided by wave traps.

[1] For more details of stagger-tuning see R. B. Dome, Television Principles, McGraw-Hill Book Company, Inc., New York, 1951.

19·7 Wave traps

These are resonant circuits tuned to reject an undesired frequency. Because of the wide pass band in the picture i-f amplifier, wave traps are generally used to reduce adjacent channel interference and to provide the i-f response required for intercarrier sound. Figure 19·13 shows the rejection frequencies to which the wave traps are tuned. Note that the wave traps determine the slope of the skirts at the edges of the overall i-f response curve.

The associated sound is the signal associated with the picture in the channel to which the receiver is tuned. This frequency at 41.25 Mc is 4.5 Mc below the i-f picture carrier at 45.75 Mc.

The lower adjacent sound is the signal for the channel below the one tuned in. For instance, when the receiver is tuned to channel 4, the sound signal of channel 3 can interfere with the picture. The lower adjacent channel sound intermediate frequency at 47.25 Mc is 6 Mc above the associated sound intermediate frequency at 41.25 Mc, or 1.5 Mc above the picture i-f carrier at 45.75 Mc. Since 47.25 Mc is relatively close to 45.75 Mc, aligning the adjacent sound trap can also change the picture i-f response.

For the upper adjacent channel, its picture i-f carrier frequency at 39.75 Mc is close to the i-f pass band. As an example, when the receiver is tuned to channel 3, the picture carrier of channel 4 can cause interference. Although adjacent channels are not assigned in the same city, in some locations there will be reception of signals from transmitters in different areas and on adjacent channels. The effects of interference between channels are described in Sec. 23·3.

Sound bars in the picture. This interference of the associated sound in the picture is shown in Fig. 19·14. The bar pattern is a result of audio signal from the sound in the video amplifier circuits for the kinescope. When the sound i-f response is too high on the picture i-f response curve,

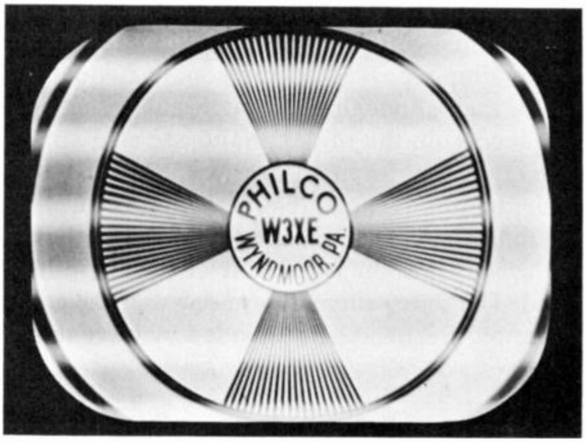


Fig. 19·14 Sound bars in the picture. The frequency of this audio signal interference at the kinescope grid is 360 cps, producing six pairs of horizontal dark and light bars. (Philco Corporation.)

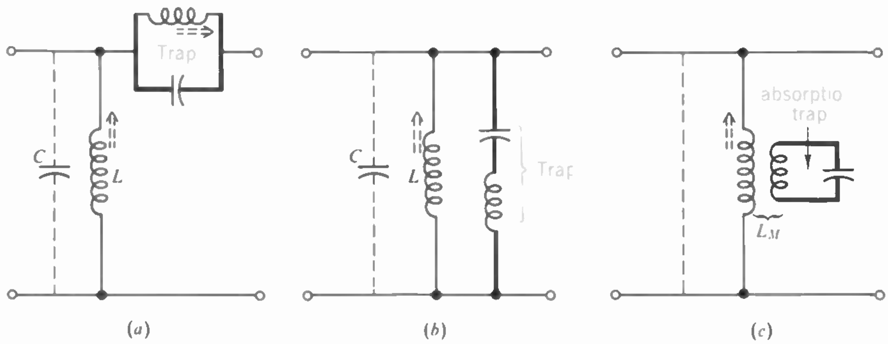


Fig. 19·15 Three types of wave-trap circuit for rejecting an undesired frequency. (a) Parallel resonant trap circuit in series with output voltage. (b) Series resonant trap circuit in shunt with output voltage. (c) Absorption trap inductively coupled to plate load impedance.

the sloping response converts frequency variations in the FM signal to a corresponding AM signal. The video detector rectifies this AM signal to produce an interfering audio signal in the video amplifier. The audio signal in the kinescope grid-cathode circuit produces horizontal pairs of dark and light bars. This effect can be recognized by the fact that the bars vary in width and number in accordance with the audio modulation as people in the picture speak, and the bars disappear when there is no voice. It should be noted that the FM signal still has its frequency variations for the sound circuits and the video signal provides the picture information needed for the kinescope. The sound bars are just superimposed on the picture, while the sound is louder than normal, usually with 60-cycle sync buzz.

Wave-trap circuits. Three main types are illustrated in Fig. 19·15. In *a*, the parallel resonant trap is connected in series with the input to the next stage. Parallel resonance provides a very high resistance at the rejection frequency. The series connection for the trap makes it a voltage divider with the following input circuit. Most of the undesired signal voltage is developed across the high resistance of the resonant trap circuit, therefore, with little voltage coupled to the next grid circuit at the rejection frequency. The trap in *b* is a series resonant circuit connected in shunt with the next grid circuit. Series resonance provides a very low resistance at the rejection frequency. At this frequency, the grid input circuit is effectively shorted by the series resonant trap. The result is very little voltage in the grid circuit at the trap frequency. The absorption trap in *c* is inductively coupled to the plate load inductance. At resonance, maximum current flows within the trap circuit. As a result, maximum current is coupled from the primary. This absorption of power from the primary causes a sharp decrease in the Q of the tuned plate load impedance, only for the trap frequency. The result is reduced gain in the i-f amplifier stage for the undesired signal.

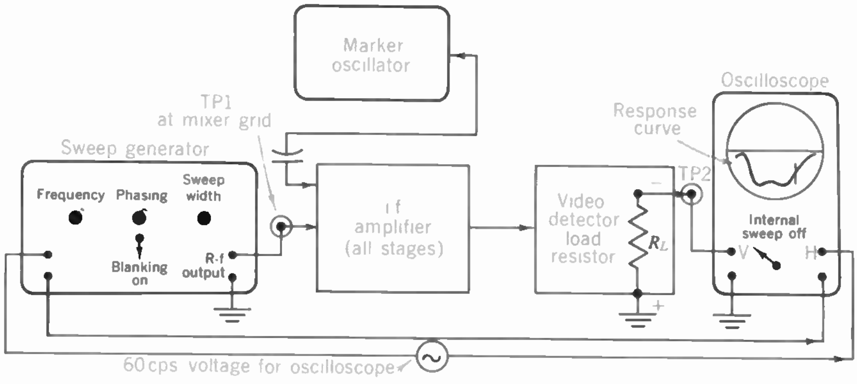
In all cases, the trap is tuned to the frequency to be rejected. As examples, the associated sound traps are aligned at 41.25 Mc to attenuate this frequency in the picture i-f response, while the lower adjacent channel sound traps are tuned to reject 47.25 Mc. The coil in the wave-trap circuit has a variable slug to tune the trap for minimum output. Each of the traps can be aligned by coupling in i-f signal at the rejection frequency and tuning for minimum d-c voltage output across the video detector load resistor. For this part of the i-f alignment, the desired result is a dip in the output because the traps have the function of reducing the gain for the rejection frequencies.

19·8 Picture i-f alignment

Correct alignment of the picture i-f stages is very important in obtaining a good picture, since the composite video signal for the kinescope is the envelope of the picture carrier signal amplified in the i-f stages. In the modulated i-f signal, the side-band frequencies farthest from the picture carrier frequency correspond to the highest video frequencies. If these are missing or attenuated, the picture will lack horizontal detail. The frequencies close to the i-f carrier correspond to the low video frequencies that reproduce the large areas of picture information. If the i-f response for the picture carrier frequency is incorrect, this can cause weak contrast or smear in the picture. Sound bars in the picture may result from improper adjustment of the sound traps.

Visual response curve. The visual alignment method will be used generally to align double-tuned picture i-f stages and to check the over-all response of staggered single-tuned stages. Figure 19·16 shows the test equipment and connections used for obtaining a visual response curve. A sweep generator is necessary to supply i-f signal input at all frequencies in the pass band of the amplifier. The oscilloscope indicates the amount of video detector output voltage at the different i-f signal frequencies. Preferably, the local oscillator in the r-f tuner should be disabled for i-f alignment to prevent spurious responses. Fixed bias of about -3 volts should be used in place of the AGC bias.

Fig. 19·16 Test equipment and connections for visual i-f response curve. TP1 is test point at mixer grid in r-f tuner; TP2 is detector output.



The average oscilloscope has high-frequency response that is good enough for alignment. With the vertical input of the oscilloscope connected across the d-c load resistance of the detector, the input to the oscilloscope vertical amplifier is not the i-f signal but only a d-c voltage fluctuating at the 60-cycle modulation rate of the sweep generator. The amount of d-c voltage depends on the amplitude of the signal input to the detector, which is proportional to i-f gain. Therefore, detector output indicates the i-f amplifier response. Actually, the low-frequency response of the oscilloscope is important as the visual response curve shows a graph of detector d-c output voltages.

The procedure for i-f alignment can be as follows:

1. Connect the oscilloscope vertical input to the video detector output circuit. A test point for this connection is usually at the top of the chassis.
2. Connect the sweep generator to the mixer grid. A test point for this connection is usually at the top of the r-f tuner.
3. The oscilloscope internal sweep is not used for horizontal deflection. If there is a phasing control on the oscilloscope, switch the horizontal amplifier selector switch to 60-cps line. If not, switch to the external position and connect the horizontal input to the oscilloscope terminals on the sweep generator. Then the phasing control on the generator is used.
4. Set the frequency of the sweep generator to the i-f range to obtain an i-f response curve on the oscilloscope screen. Use about 10 Mc sweep width to see the entire i-f response. The exact frequency of the sweep generator does not matter, as the frequencies must be indicated with an accurate marker. The i-f response curve is normally downward on the oscilloscope screen for negative d-c voltage output from the video detector. Use little generator output to avoid overload distortion. Turn the oscilloscope gain up to make the curve fill the screen.
5. Two response curves will be produced by the sweep generator, without blanking. First, make the curves overlap by adjusting the phasing control. Then turn on the blanking to provide one curve with a zero base line.

The wave traps are usually adjusted first, for minimum response at the rejection frequency, to provide the required attenuation at the edges of the i-f pass band. The i-f tuned circuits then are adjusted for maximum output, indicated by increased height in the curve. Adjusting the i-f tuned circuits and the traps changes the shape of the response curve to fit the markers correctly. Several examples are shown in Fig. 19·20.

Marking the frequencies on the final response curve should show (1) picture i-f carrier frequency 50 to 60 per cent up on the sloping side, with about 1.5 Mc between the bottom and top of the slope; (2) sufficient bandwidth across the flat top, without any dip more than about 10 to 20 per cent down; and (3) the required response of 5 to 10 per cent for the associated sound i-f carrier frequency, in intercarrier-sound receivers.

Peaking staggered single-tuned stages. This alignment can be accomplished with just a d-c voltmeter and an r-f signal generator, such as the marker oscillator. A sweep generator is not needed for test signal at one frequency. The oscilloscope is not necessary because the d-c voltmeter serves as the output indicator.

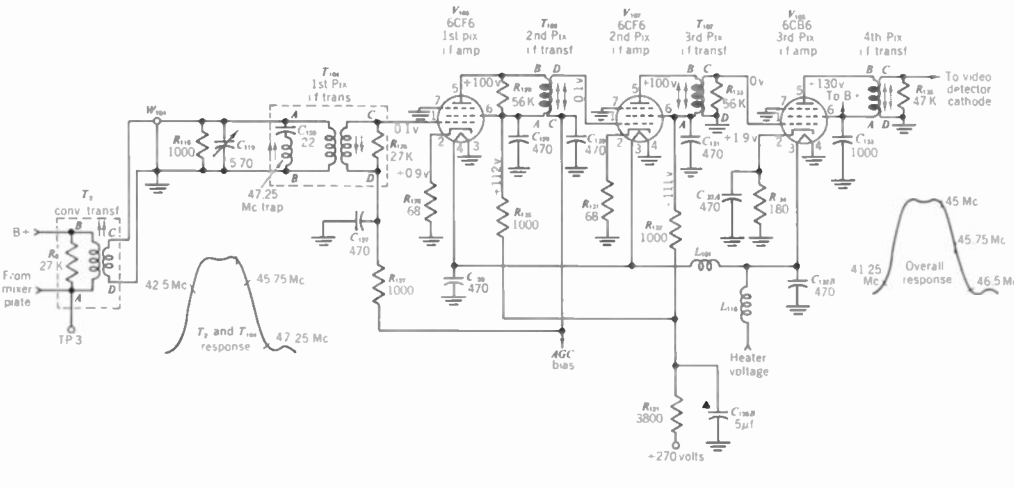
The voltmeter, set to its lowest d-c voltage scale, is connected across the video detector load resistor for all adjustments. The signal generator remains connected to the mixer grid to supply test signal at the individual frequencies for each i-f stage. With the signal generator frequency set to each of the rejection frequencies, the wave traps are adjusted first for minimum output on the d-c voltmeter. Then the signal generator is set for the individual peaking frequencies, as each single-tuned i-f circuit is adjusted for maximum output at its resonant frequency, indicated by maximum on the d-c voltmeter.

19·9 Picture i-f amplifier circuits

Figure 19·17 shows the schematic diagram of a typical three-stage i-f amplifier, with a double-tuned coupling circuit from the mixer stage and three staggered single-tuned stages using bifilar coils. The printed wiring commonly used for i-f circuits is shown in Fig. 19·18, with the i-f alignment adjustments.

Starting at the left in the diagram, the converter transformer T_2 in the plate circuit of the mixer stage is mounted on the r-f tuner subchassis. The i-f output of the mixer is link-coupled by means of a short length of 75-ohm coaxial line to T_{104} in the grid circuit of the first i-f stage V_{106}. The link coupling has the advantages of minimizing radiation of the local oscil-

Fig. 19·17 Typical picture i-f amplifier circuit, with one double-tuned coupling and three staggered single-tuned stages using bifilar coils.



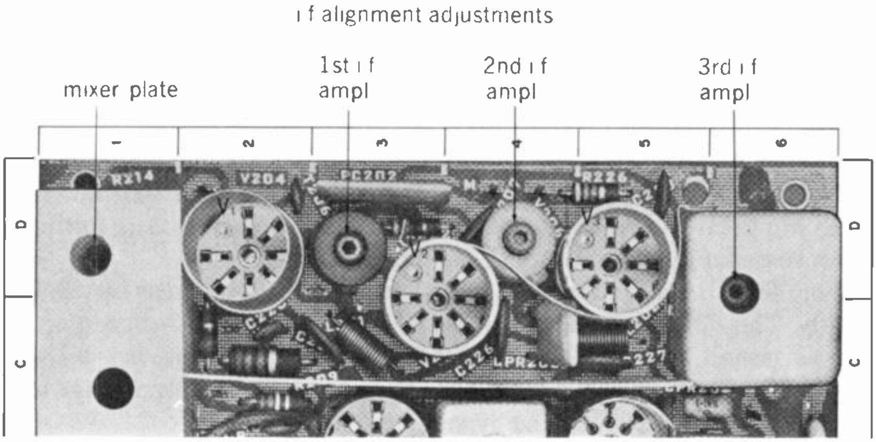


Fig. 19·18 Picture i-f amplifier strip with printed wiring. (RCA.)

lator signal from the main chassis, while allowing the r-f tuner to be located farther from the i-f amplifier. Transformer coupling in T_{104} provides the i-f signal voltage across the secondary winding *C-D* in the grid circuit of the first i-f stage. R_8 in the plate circuit of the mixer stage and R_{126} in the grid circuit of V_{106} are shunt damping resistors.

The response of T_1 and T_{104} with the link coupling is the same as though their primary and secondary windings were transformer-coupled. Because of the sharp skirt selectivity with the double-tuned coupling circuit, only one wave trap is used in the i-f amplifier. This series resonant circuit across the link winding *A-B* in T_{104} attenuates the lower adjacent channel sound i-f carrier frequency of 47.25 Mc. The variable trimmer C_{119} adjusts the slope of the i-f response curve at the low-frequency side to provide the relative gain required for the associated sound. The overall i-f response is the combined response of the double-tuned coupling circuit provided by T_1 and T_{104} with the response of the three single-tuned i-f stages using the bifilar coupling coils.

In the plate circuit of the first i-f stage, T_{106} tunes with the shunt capacitances to form a parallel resonant circuit, which is the plate load impedance. R_{129} is a shunt damping resistor. The required screen-grid voltage is provided by R_{130}, with the bypass C_{129}. No coupling capacitor is needed with the bifilar coil. Similarly, V_{107} and V_{108} are single-tuned stages with peak response at their individual resonant frequencies. For V_{107}, the resistor R_{133} provides shunt damping across T_{107}. The screen dropping resistor is R_{132}, with the bypass C_{131}. In these two stages, the cathode resistors are unbypassed to allow some degeneration, which stabilizes the input capacitance with changes in AGC bias. In the third i-f stage, the shunt damping resistor across T_{108} is R_{136}. As the i-f signal output is coupled to the cathode of the second detector, its relatively low resistance provides additional damping across the last i-f tuned circuit. Cathode bias is provided by R_{134} with C_{132A} for the last i-f stage.

19·10 Transistorized i-f stage

In Fig. 19·19, the PNP transistor TR6 is in a double-tuned i-f amplifier circuit, which is the last of three i-f stages. No AGC bias is used in this stage but the previous two transistors are controlled by forward AGC. The i-f input signal is coupled by C_{116} and R_{123} to the base of TR6, operating in a common-emitter circuit. This base input circuit has a trap tuned to 41.25 Mc to attenuate the associated sound signal. C_{118} with L_{108} form the series resonant trap in shunt with the input circuit.

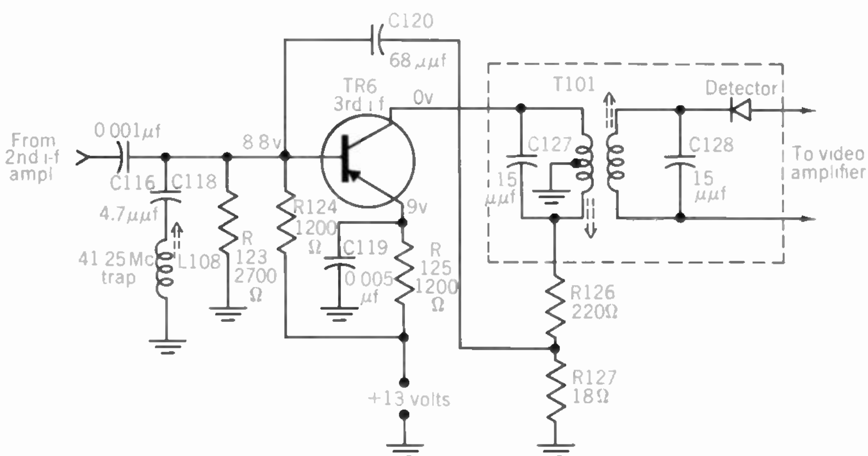
Note that the base resistor R_{123} is in series with R_{124} across the 13-volt supply. This voltage divider provides 8.8 volts at the base with respect to chassis ground. However, the forward-bias voltage at the base is -0.2 volt with respect to the emitter at 9 volts. R_{125} drops the supply voltage to 9 volts at the emitter. C_{119} is the bypass for R_{125}.

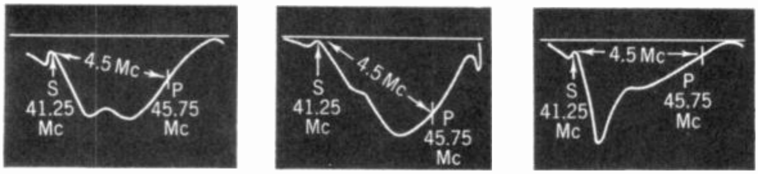
In the collector output circuit, T_{101} is a double-tuned transformer supplying i-f signal to the video detector. The primary of T_{101} is tapped to provide negative feedback to the base for minimizing regeneration. The feedback voltage is divided down by R_{126} and R_{127}, while C_{120} couples the i-f feedback to the base input circuit. Although the d-c voltage at the collector is zero, its reverse bias is -9 volts with respect to the emitter.

19·11 Troubles in the picture i-f amplifier

Since these stages amplify the modulated carrier that is detected to provide video signal for the kinescope, troubles in the picture i-f section can cause no picture, weak picture, or poor picture quality, while the raster is normal. Also, there will be no sound without any output from the common i-f amplifier section in intercarrier-sound receivers. The amount of i-f signal can be checked by measuring the rectified d-c output voltage across the video detector load resistance.

Fig. 19·19 Transistorized i-f stage. (Motorola chassis TS 432C.)





(a) (b) (c)

Fig. 19·20 Picture i-f alignment curves. Note marker at picture carrier frequency. (a) Normal response. (b) Excessive response for picture carrier with narrow bandwidth. (c) Insufficient gain at 45.75 Mc with excessive peak in response near 41.25 Mc.

Tunable smear. This is smear in large areas of the picture that moves when the fine tuning control is varied. As the control varies the local oscillator frequency, the i-f picture carrier frequency changes, resulting in a different response for the i-f picture carrier signal. Tunable smear in the picture generally indicates a distorted response curve for the i-f amplifier, with incorrect relative gain and excessive phase distortion. Examples are shown in Fig. 19·20.

Picture carrier too high on i-f response curve. See Fig. 19·20*b*. Excessive gain for the i-f picture carrier and side-band frequencies close to it corresponds to excessive low-frequency response in the video amplifier. The result is strong contrast but phase distortion causes smear in large areas of the picture. This effect is the same as excessive gain for low video frequencies as illustrated in Fig. 9·16.

Narrow bandwidth. In Fig. 19·20*b*, the i-f response curve does not have enough bandwidth to provide normal gain for the side-band frequencies farthest from the picture carrier. Although their numerical values are lower than the picture carrier in the i-f response, these side frequencies correspond to the high video frequencies. The result is poor horizontal resolution in the picture, with insufficient detail. This effect is the same as insufficient gain for high video frequencies.

Picture carrier too low on i-f response curve. See Fig. 19·20*c*. Low gain for the i-f picture carrier and side-band frequencies close to it corresponds to insufficient low-frequency response in the video amplifier. The result is weak contrast, with highlights in the picture being reproduced as gray instead of white. These effects are illustrated in Fig. 9·13, which shows insufficient gain for the low video frequencies. There may also be sound bars in the picture, if the i-f response is too high for the associated sound signal. Poor synchronization can result because of weak video signal.

Excessive peak in i-f response. In Fig. 19·20*c*, the i-f response has too much relative gain for frequencies close to 41.25 Mc. The excessive gain can cause multiple outlines in the picture, as shown in Fig. 19·21, corresponding to ringing for the high video frequencies. Although the multiple outlines appear similar to ghosts in the picture caused by reflections in the antenna signal, the ringing effect is the same on all stations and the mul-

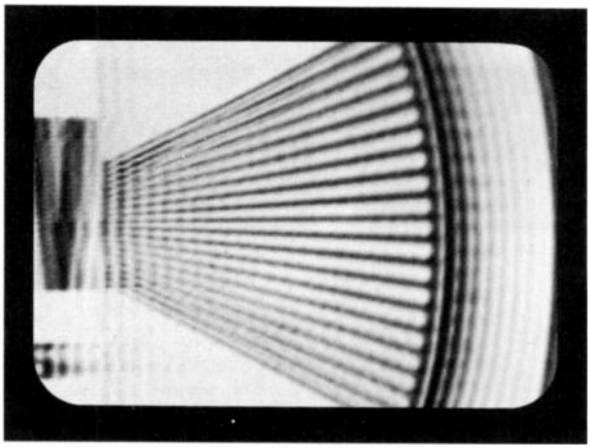


Fig. 19·21 Multiple outlines caused by ringing. (RCA.)

tiple outlines are regularly spaced. In addition, multiple outlines caused by the i-f amplifier move in the picture when the fine tuning control is varied.

Oscillations in the i-f amplifier. Regeneration can make the i-f amplifier oscillate, especially when there is an excessive peak in the i-f response at one frequency. The effect in the picture is usually multiple outlines, with black streaks caused by radiation of i-f signal back to the r-f tuner. To check whether the i-f amplifier is oscillating, measure the rectified i-f output across the video detector load resistor. Normally, this d-c output is about 3 volts with a station tuned in. When the fine tuning control is varied, the d-c voltage varies as the picture carrier moves up or down the side of the i-f response curve. Also, the detector output drops practically to zero when the receiver is switched to an unused channel. However, if the detector output voltage is very high and stays about the same regardless of the r-f tuning, these effects indicate the i-f amplifier is oscillating. The trouble can be caused by misalignment or a defective component that alters the alignment. It should be noted that an open in one of the i-f bypass capacitors in the decoupling filters for the tuned circuits usually affects the resonant response.

SUMMARY

1. The picture i-f section usually includes three stages to amplify the i-f signal output from the mixer to supply enough i-f signal into the video detector. Overall i-f gain is about 10,000. The i-f picture carrier has 50 per cent response to compensate for vestigial-side-band transmission. The i-f sound carrier signal is also amplified but with 5 to 10 per cent response, to produce the 4.5-Mc intercarrier-sound signal in the video detector output. The standard frequencies are 45.75 Mc for i-f picture carrier and 41.25 Mc for i-f sound carrier, or 25.75 and 21.25 Mc, respectively, in older receivers.
2. The i-f bandwidth may be as little as 2.5 Mc or as high as 4 Mc, to accommodate the side bands of the modulated picture carrier signal. Side frequencies close to the carrier correspond to the low video frequencies that determine contrast in the reproduced picture. Side frequencies farther from the picture carrier correspond to the high video frequencies up to 4 Mc that determine horizontal detail.

3. Wave traps are generally used in the picture i-f amplifier to attenuate the associated sound carrier at 41.25 Mc, and the lower adjacent channel sound at 47.25 Mc. Too much gain for the associated sound carrier causes audio sound bars in the picture and buzz in the sound.
4. Each i-f stage is a tuned amplifier using a miniature glass pentode tube with either cathode bias or AGC bias for class A operation. The gain of the stage equals $g_m Z_L$, where Z_L is the a-c plate load impedance. The tuned amplifier has gain only for frequencies that provide a resonant rise in impedance. The d-c voltage drop across the tuned circuit is practically zero.
5. With a single-tuned circuit, its resonant frequency f_r equals $1/(2\pi\sqrt{LC})$ where C is the total shunt capacitance across the i-f coil L. The relative gain of the single-tuned amplifier is the same as the impedance-frequency curve of the parallel resonant circuit. The maximum impedance at f_r equals $Q \times X_L$. The total bandwidth at 70 per cent response is f_r/Q.
6. A shunt damping resistor R_D is generally used for the required bandwidth. Lower values of R_D provide more bandwidth because of lower Q. The tuned circuit with shunt damping has Q_D equal to R_D/X_L.
7. In a double-tuned i-f transformer, both L_p and L_s are tuned to the same frequency. Compared with a single-tuned circuit, the double-tuned amplifier can provide more bandwidth for the same gain, with sharper skirt selectivity.
8. In staggered single-tuned stages, each is resonant at a different frequency to broaden the overall response.
9. The overall picture i-f response curve can be observed with an oscilloscope connected to the video detector load resistance, while a sweep generator at the mixer grid provides the required frequencies in the i-f pass band. When the i-f picture carrier is too high on the curve, the result is more contrast in the reproduced picture but less detail and possibly smear. Too low a response means little contrast. The required bandwidth provides full detail in the picture. A typical i-f bandwidth is 3.2 Mc.
10. No i-f output means no picture and no sound, while the raster is normal. You can check for i-f output by measuring the d-c voltage across the video detector. The normal amount of rectified i-f signal is about 3 to 5 volts.

SELF-EXAMINATION *(Answers at back of book.)*

Part A

Fill in the missing answer.

1. The i-f output of the mixer includes picture carrier signal at 45.75 Mc and sound carrier signal at ______ Mc.
2. The output of the video detector includes video signal and intercarrier FM sound signal at ______ Mc.
3. If the i-f gain at 43 Mc is 8,000, the gain for 45.75 Mc should be ______?
4. With 1 mv input and a gain of 5,000 the amplified voltage output equals ______ volts.
5. If one i-f stage has a gain of 30 and two others a gain of 15 each, all at the same frequency, the overall voltage amplification equals ______.
6. With picture i-f carrier at 25.75 Mc, the lower adjacent channel sound carrier is at ______ Mc.
7. With picture i-f carrier at 45.75 Mc, the i-f side frequency for 3-Mc video is ______ Mc.
8. With picture i-f carrier at 45.75 Mc, the upper adjacent channel picture carrier is at ______ Mc.
9. A single-tuned circuit resonant at 40 Mc with a Q of 20 has a bandwidth of ______ Mc.
10. If X_L in Question 9 is 300 ohms, Z_L at 40 Mc is ______ ohms.
11. If this circuit is used with a tube having g_m of 4,000 μmhos the gain at 40 Mc equals ______.
12. The shunt damping resistor R_D for this circuit is ______ ohms.
13. If R_D is reduced by one-half, the lower Q will be ______.

14. Then the increased bandwidth will be ______ Mc.
15. In a double-tuned i-f transformer, when Q_p is 40 and Q_s is 10, critical coupling results at k_c equal to ______.

Part B

Answer true or false.

1. Any type of wave-trap circuit is always tuned for minimum output at the trap frequency.
2. The first two picture i-f stages often have AGC bias.
3. Each i-f stage is a class A amplifier.
4. The d-c voltage drop across the primary of an i-f transformer is practically zero.
5. The plate load impedance for an i-f stage is a result of resonance in a tuned circuit.
6. Insufficient bandwidth in the picture i-f response results in poor horizontal detail.
7. Insufficient gain for the i-f picture carrier causes weak contrast.
8. When a common i-f amplifier is cut off by excessive AGC bias there is no picture and no sound.
9. A single-tuned stage is aligned by peaking for maximum output at its resonant frequency.
10. Although the associated sound is an FM signal, it can produce audio interference in the video signal.
11. If the capacitance is reduced one-half the required inductance for the same resonant frequency will be double.
12. Lower Q in a tuned circuit allows more bandwidth.
13. Less bandwidth in a tuned amplifier allows more gain.
14. A smaller shunt damping resistor across a tuned circuit increases its bandwidth.
15. In a double-tuned i-f transformer, primary and secondary are resonant at the same frequency.

ESSAY QUESTIONS

1. Draw the desired i-f response curve for a bandwidth of 3.2 Mc, for intercarrier sound, with picture i-f carrier at 45.75 Mc. Mark the following carrier frequencies: associated sound and picture, lower adjacent channel sound, and upper adjacent channel picture.
2. Repeat Table 19·1 but with frequencies listed for channel 2, 54 to 60 Mc.
3. Draw the schematic diagram of a double-tuned i-f stage with a wave trap.
4. Draw the schematic diagram of a single-tuned i-f stage with a wave trap. Use a bifilar i-f coil.
5. With i-f picture carrier at 25.75 Mc, list the i-f side frequencies for video signal frequencies of 1 Mc, 2 Mc, and 3 Mc.
6. What is the image frequency of channel 2 picture carrier (54 to 60 Mc) in a receiver with picture i-f carrier at 25.75 Mc?
7. For picture i-f carrier at 45.75 Mc, use numerical values to prove: (*a*) no image frequency is in the commercial FM broadcast band; (*b*) local oscillator frequency is not in any VHF television channel.
8. What is a wave trap? Give three types. Why is a wave trap adjusted for minimum i-f output at the trap frequency?
9. What causes audio sound bars in the picture? Give two differences between sound bars and hum bars.
10. What two signals are necessary for the video detector to produce a 4.5-Mc beat in the output signal?
11. Give two functions of a 4.5-Mc wave trap in the video amplifier.
12. What is meant by a stagger-tuned amplifier? What is the purpose of the stagger tuning?
13. For more bandwidth, should the resistance of a shunt damping resistor be increased or decreased? Why?
14. What is the advantage of a single-tuned stage using a bifilar coil, instead of a single i-f coil?

15. Define loose coupling, critical coupling, and overcoupling.
16. Explain briefly why a double-tuned transformer can have a single-peak response when either side is temporarily damped with a 330-ohm resistor.
17. The gain of a tuned i-f stage is $g_m Z_L$. Give two factors affecting g_m. Give two factors affecting Z_L.
18. Describe five possible effects in the picture resulting from an incorrect i-f response curve.
19. What is meant by an overloaded picture? How can loss of AGC bias cause this effect?
20. Explain briefly how excessive heater-cathode leakage in an i-f stage can cause horizontal pulling in the picture, with one pair of hum bars.
21. Describe briefly how an i-f amplifier with staggered single-tuned stages is aligned, using an r-f signal generator and a d-c voltmeter.
22. Describe briefly how to obtain the visual i-f response curve showing overall i-f gain, with provision for marking frequencies on the curve.
23. How does the marker oscillator produce a marker pip on the curve?
24. Why is the signal generator at the mixer grid for i-f alignment, instead of being connected into the i-f circuits?
25. How would you provide fixed bias of -3 volts instead of AGC bias for i-f alignment?
26. List five controls or output terminals on a sweep generator and give their functions.
27. In Fig. 19·17, give the function of T_{106}, R_{127}, R_{134}, C_{131}, and R_{132}.
28. In Fig. 19·17, what will be the effect of the following troubles: (*a*) in the screen-grid circuit of V_{107}, R_{132} opens; (*b*) in the screen-grid circuit of V_{106}, C_{129} shorts; (*c*) in the cathode circuit of V_{106}, R_{128} opens; (*d*) in the plate circuit of V_{107}, coil *A-B* in T_{107} opens; (*e*) in the grid circuit of V_{106}, C_{127} shorts.
29. In Fig. 19·19 give the functions of C_{116}, C_{120}, R_{123}, and L_{108}.
30. In Fig. 19·19, give the value and polarity of base-emitter forward bias and collector-emitter reverse bias on TR6.

PROBLEMS *(Answers to odd-numbered problems at back of book.)*

1. A single-tuned circuit has X_L of 300 ohms at f_r, and a shunt damping resistor of 4,500 ohms. Calculate the Q, bandwidth Δf, and Z_L at the resonant frequency f_r of 30 Mc.
2. For the circuit in Prob. 1 draw four separate graphs showing: (*a*) values of Q for 10 values of R_D from 1,000 to 10,000 ohms; (*b*) values of Z_L for the 10 values of Q; (*c*) values of bandwidth Δf for the 10 values of Q; (*d*) absolute gain values at f_r for the 10 values of Z_L, with g_m equal to 3,000 μmhos.
3. Calculate the image frequency of channel 2 picture carrier for a receiver with: (*a*) picture intermediate frequency at 45.75 Mc; (*b*) picture intermediate frequency at 25.75 Mc.
4. (*a*) List the local oscillator frequency with r-f and i-f carrier frequencies for picture and sound of channel 3 and channel 4 when a receiver with picture intermediate frequency of 45.75 Mc is tuned to channel 4. (*b*) Do the same for the case of the receiver tuned to channel 3.
5. For the i-f response curve in Fig. 19·2, assume the overall gain at 43 Mc is equal to 8,000. How much is the gain at 41.25 Mc, 42 Mc, and 45.75 Mc?
6. (*a*) Draw the response curve in Fig. 19·2 showing the gain values from Prob. 5. (*b*) Show the curve that would be obtained with gain reduced one-half by AGC bias.
7. A tuned circuit has X_L of 628 ohms at f_r of 25 Mc. (*a*) Calculate L. (*b*) Calculate C. (*c*) If C is reduced one-half, how much is L for the same f_r?
8. The i-f transformer T_i in Fig. 19·8 is resonant at 43 Mc, with C_{out} 5 $\mu\mu$f and C_{in} 10 $\mu\mu$f. Calculate: (*a*) L_p and L_s; (*b*) R_4 and R_5 for Q of 18; (*c*) critical coupling factor k_c; (*d*) value of L_M at critical coupling.
9. A stagger-tuned triple is centered at 30 Mc with overall bandwidth of 3 Mc. What are the resonant frequencies and values of Q for the three single-tuned circuits?
10. Draw the schematic diagram for the triple in Prob. 9. Label all values of L, with 10 $\mu\mu$f shunt capacitance in each stage, and values of R_D for the required Q. Show the response curve for each stage and the overall response.

Chapter 20 The R-F tuner

The r-f amplifier, local oscillator, and mixer stages in the superheterodyne television receiver form the r-f tuning section, often called the *front end* or tuner. This section selects the picture and sound carrier signals of the desired channel by converting the r-f signal frequencies of the selected station to the intermediate frequencies of the receiver, as illustrated in Fig. 20 · 1. It is the local oscillator that tunes in the station, since the oscillator frequency determines which r-f signal frequencies are converted to frequencies in the pass band of the i-f section. The fine tuning control varies the oscillator frequency slightly for exact tuning. This control is set for the best picture in intercarrier-sound receivers or best sound in split-sound receivers. The main function of the r-f amplifier is to provide a good signal-to-noise ratio for the amplified r-f signal coupled to the mixer, for minimum snow in the picture.

20 · 1 Operation of the r-f tuner

In Fig. 20 · 1, the r-f stage is tuned to the frequencies of the selected channel to amplify its r-f picture and sound carrier signals coupled from the antenna input circuit. The amplified r-f output, still at the channel frequencies, is coupled to the grid circuit of the mixer. In the mixer stage, the r-f signal frequencies beat with the local oscillator output for frequency conversion to the intermediate frequencies. The one oscillator frequency beats with the two r-f carrier signals of the selected channel to produce the desired i-f picture and sound signals. When the station-selector switch is varied, the resonant circuits for the r-f amplifier and mixer grid are tuned to the channel frequencies for each selected station. At the same time, the resonant circuit of the local oscillator is set to the frequency required for converting the selected channel's r-f signal frequencies to the corresponding i-f signal frequencies. The mixer and local oscillator stages, which function as the frequency converter, are generally combined in one twin-

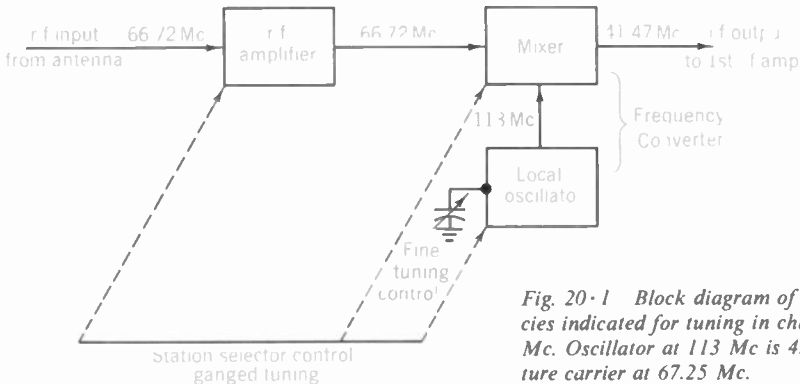
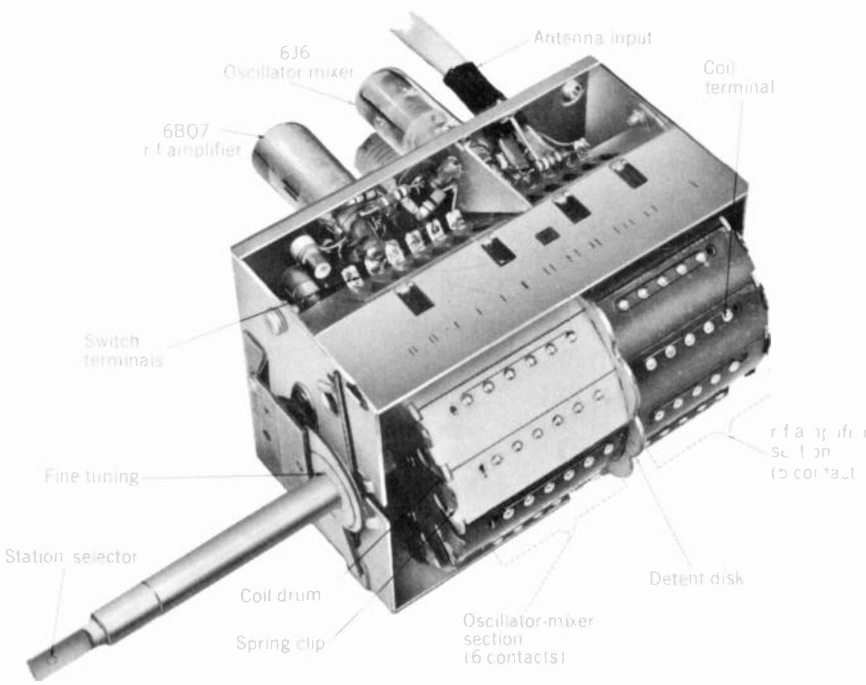


Fig. 20·1 Block diagram of r-f tuner. Frequencies indicated for tuning in channel 4 at 66 to 72 Mc. Oscillator at 113 Mc is 45.75 Mc above picture carrier at 67.25 Mc.

triode tube or a triode-pentode. Therefore, the front end usually has only two tubes, the r-f amplifier and the oscillator-mixer.

The r-f tuner is a separate subchassis, with connections to the main chassis for B+ voltage, heater power, AGC bias for the r-f stage, and i-f signal output. Antenna signal input is coupled to the r-f amplifier. In construction, there are three main types of front end. These include the turret tuner in Fig. 20·2, the rotary-switch tuner in Fig. 20·3, and continuous tuners with variable inductors or capacitors on a common shaft. The turret

Fig. 20·2 Turret tuner, with individual coil strips for each channel. Bottom shield cover and tube shields removed to show components. (Standard Kollsman Industries.)

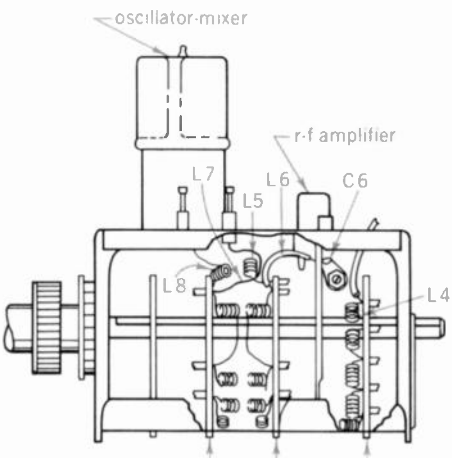


and switch types are most common. In Fig. 20·2, note the individual coil strips held by spring clips on the drum, which is the turret to be rotated. Each coil is on the underside of the plastic strip, with silver-plated terminals on the top for the connections. As the drum is turned by the station-selector control, the coil terminals contact the stationary switch connections, to tune in the desired channel. The turret has 12 positions for 12 channels. At each position, individual coils are connected into the circuits of the r-f amplifier, mixer grid, and local oscillator for ganged tuning.

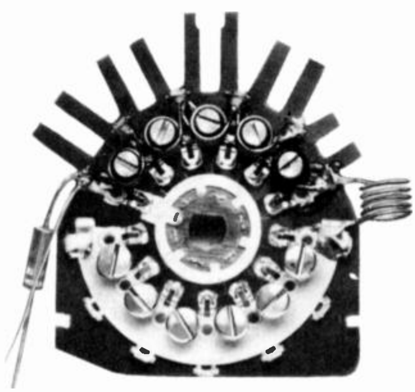
In Fig. 20·3, the ganged tuning is provided by three switch wafers on a common shaft. Individual coils for each of the 12 channels are mounted between terminals on the three wafers. The local oscillator section is to the front, for convenience in aligning the oscillator frequency on each channel.

Figure 20·4 shows a combination VHF-UHF tuner. The shaft for fine tuning on VHF channels provides continuous tuning through all the UHF channels, when the station selector switch is on the UHF position. More details of UHF tuning are described in Secs. 20·7 to 20·9.

Also necessary is a mechanical *detent,* which is a spring arrangement to hold the station selector firmly as it clicks into position on each channel, for tight contacts at the switch connections. Two common mechanical troubles are a loose detent and dirty switch contacts. Indications of poor contact are: (1) moving the station selector switch slightly improves the picture; (2) rotating the switch one way tunes in the station better than the opposite rotation. Since the contacts are silver-plated they tarnish and must be cleaned, preferably with a commercial contact cleaner.



(a)



(b)

Fig. 20·3 Rotary-switch tuner. (a) Ganged wafer switches. (b) Oscillator section; screws are frequency adjustments for each channel. (RCA.)

Fig. 20·4 Combination VHF-UHF tuner.
(Sarkes Tarzian, Inc.)

20·2 The r-f amplifier stage

The tuned r-f amplifier, or *preselector,* is optional in a superheterodyne circuit but practically all television receivers have one t-r-f stage. Its main purpose is to improve the signal-to-noise ratio for the picture and sound signals into the mixer. There are two important considerations. First, the antenna signal is relatively small, compared with radio broadcasting at frequencies lower than the VHF range. Second, the noise is relatively high, compared with receivers having less bandwidth. The result of a poor signal-to-noise ratio is snow in the reproduced picture, as shown in Fig. 20·5. Since the r-f amplifier is the first stage in the receiver, where the signal level is lowest, the r-f gain is the most important factor in determining the signal-to-noise ratio of the receiver. There must be enough r-f signal to override the high level of tube noise generated in the mixer stage.

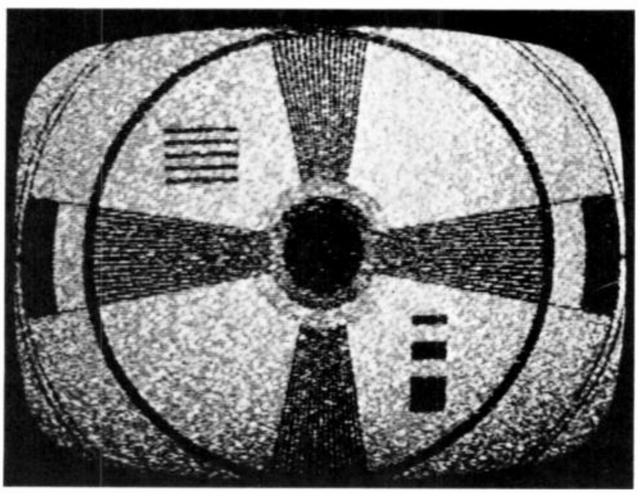


Fig. 20·5 White speckled background of snow, produced by noise voltages generated in the receiver. (RCA Institutes Home Study School.)

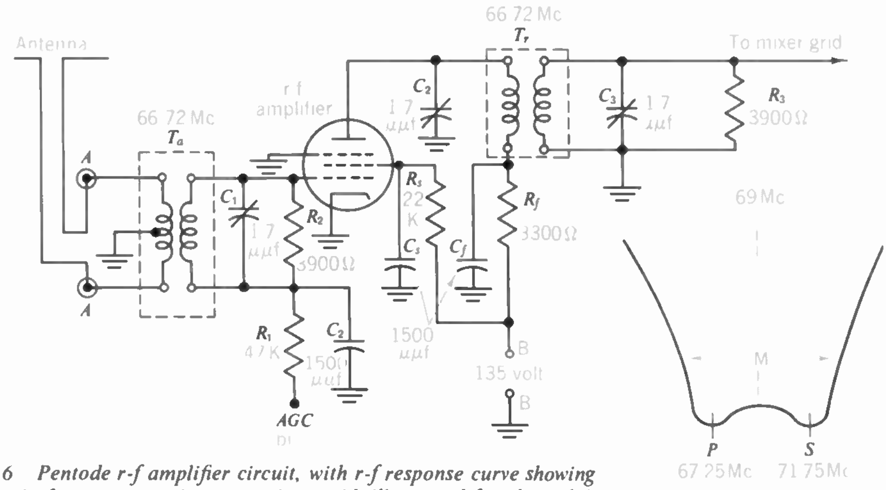


Fig. 20·6 Pentode r-f amplifier circuit, with r-f response curve showing relative gain from antenna input to mixer grid, illustrated for channel 4. The r-f transformers T_a and T_r are changed for each channel.

As illustrated in Fig. 20·6, double-tuned circuits are generally used for the r-f amplifier. The inductances resonate with the shunt capacitances at the desired channel frequencies. The trimmer capacitances C_1, C_2, and C_3 are provided to align the r-f tuned circuits, usually at channel 10. The amount of coupling and the shunt damping resistance provide the required 6-Mc bandwidth. However, the low internal resistance of the grid-cathode circuit can provide the damping. A pentode is illustrated here but triodes or tetrodes are commonly used because they produce less tube noise. The r-f stage is a class A amplifier, usually with AGC bias. However, the r-f bias is less than the i-f bias to increase r-f gain. The voltage amplification of a typical r-f stage is about 10.

The typical r-f response curve in Fig. 20·6 shows relative voltage amplification of the tuned r-f amplifier, from the antenna input to the output at the mixer grid. Frequencies are marked for channel 4, but the r-f amplifier must provide a similar response curve for each selected channel. Note that the curve has 6 Mc bandwidth, symmetrical about the center frequency of the channel, in order to accept both the picture and sound r-f carrier signals with approximately equal response.

Oscillator radiation. The r-f amplifier provides the only isolation between the local oscillator and the antenna, which can radiate the oscillator signal to produce r-f interference in nearby receivers. Therefore, the use of an r-f amplifier is advantageous in providing a buffer stage between the local oscillator and the antenna. Pentode r-f stages provide more isolation than triodes because of the smaller plate-to-grid capacitance. It should be noted, though, that the local oscillator output can also be radiated by the receiver chassis. This is why a separate chassis ground return is generally used for the tuner circuits. When the frequency of oscillator radiation is in the channel to which a nearby receiver is tuned, it can cause diagonal lines in the picture of the other receiver, like the r-f interference shown in

Fig. 23·3*a*. To minimize this problem, the EIA rating for television receivers specifies that the field strength of a radiated oscillator signal 1,000 ft away shall not exceed 25 μv per meter with an antenna height of 30 ft. In any case, receivers with 45.75 Mc intermediate frequencies do not have the oscillator frequency in any VHF channel.

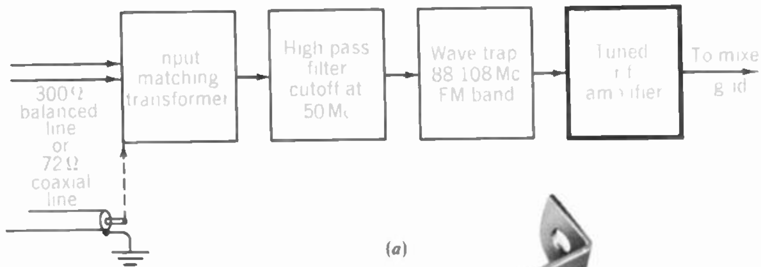
R-F selectivity. Most of the receiver selectivity to reject adjacent channel frequencies is in the i-f section, but the use of r-f amplification helps in rejecting interfering r-f signals that can produce beat frequencies within the i-f pass band of the receiver. This is especially important in rejecting image frequencies. The first-order image frequency is equal to twice the intermediate frequency plus the desired frequency, when the local oscillator operates above the r-f signal frequencies. Another function of the r-f amplifier is to prevent reception of spurious signals not related to the desired signal frequency. The r-f preselection is especially important for rejecting frequencies in the i-f pass band.

Input impedance. A definite value of impedance must be provided at the antenna terminals on the receiver, which is the input circuit for the r-f amplifier, in order to match the characteristic impedance of the transmission line from the antenna. Without this impedance match, the length of the transmission line would be critical. The standard input impedance for television and FM receivers is 300 ohms. This value matches 300-ohm twin-lead transmission line. Some receivers also have 72-ohm input connections for coaxial line. The input impedance is resistive because of resonance but usually it cannot be measured with an ohmmeter. Its value is relatively constant for all channels. With the receiver input impedance matched to the transmission line, it can be cut at any convenient length without affecting the picture.

R-F input circuit. Figure 20·7 illustrates the impedance matching and filter circuit requirements of the antenna input circuit to the r-f tuner. The input transformer couples the antenna signal to the grid circuit of the r-f amplifier, while matching the impedance of the grid circuit to the impedance of the transmission line from the antenna. In order to improve the r-f selectivity and reject interfering r-f signals, the r-f input circuit usually includes a high-pass filter and a wave trap. The wave trap is a resonant circuit that can be tuned to reject one frequency within the FM broadcast band of 88 to 108 Mc. The high-pass filter is an *LC* band-pass network that attenuates a wide range of frequencies lower than 50 Mc, approximately, cutting off just below channel 2. Interfering frequencies in this range include diathermy equipment and some amateur transmitter bands. The desired signal frequencies for channel 2 or higher are passed by the filter to provide the picture and sound signals for the grid input circuit of the r-f amplifier.

20·3 R-F amplifier circuits

Two basic types of amplifiers are shown in Fig. 20·8. The conventional pentode circuit in *a* where the input signal is applied to the control grid,



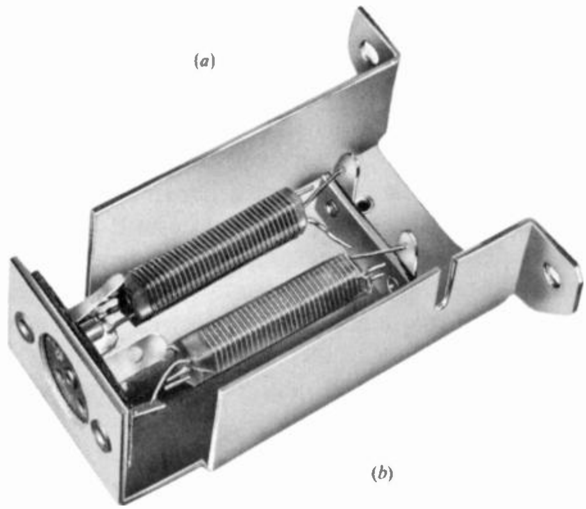
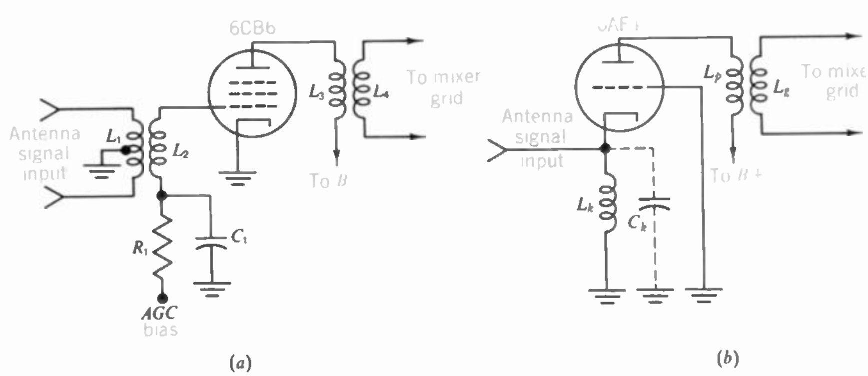


Fig. 20·7 Illustrating impedance matching and filter requirements for antenna input circuit of r-f tuner. (a) Block diagram of circuits. (b) Input matching transformer. (RCA.)

with the cathode grounded, is called a *grounded cathode amplifier.* In *b*, the input signal is applied to the cathode, while the grid of the triode is grounded. This circuit, called a *ground grid amplifier,* is useful for a triode r-f stage to reduce feedback through the grid-plate capacitance C_{gp}. For both amplifiers, the output signal is taken from the plate circuit. Either type can be used for an r-f amplifier stage. The pentode grounded cathode stage has little C_{gp} but tube noise is much higher than in a triode. This circuit is also used with low-noise tetrodes, which have minimum screen-grid current. The small C_{gp} for a screen-grid tube not only prevents feedback that can make the r-f amplifier oscillate but also provides isolation between

Fig. 20·8 Signal input and output for two basic types of r-f amplifiers. (a) Conventional pentode grounded cathode amplifier. (b) Triode grounded grid amplifier.



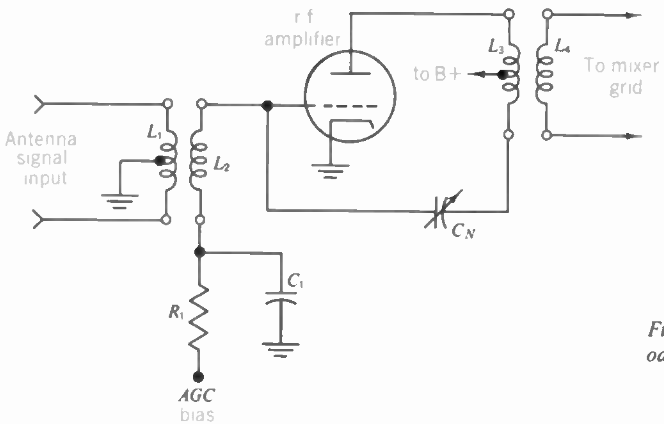


Fig. 20·9 Neutralized triode r-f amplifier.

the local oscillator and antenna to reduce oscillator radiation. When a triode is used for minimum tube noise, either the grounded grid circuit in Fig. 20·8*b* is necessary to shield the input circuit from the output signal, or the triode amplifier must be neutralized, as in Fig. 20·9. The neutralizing circuit provides feedback of opposite phase from the signal coupled through C_{gp}, to prevent oscillations in the r-f amplifier.

Grounded cathode amplifier. In Fig. 20·7*a*, the r-f input signal is coupled to the control grid by the antenna transformer L_1L_2. Since the cathode is grounded, the signal input voltage is applied between control grid and cathode. The tuned grid circuit returns to the AGC line for bias, through the decoupling filter R_1C_1. The bypass C_1 provides a signal return to the grounded cathode. It should be noted that, even with a cathode bias resistor, the cathode is essentially at ground potential for the r-f signal. The amplified output signal in the plate circuit is coupled by the r-f transformer to the mixer grid.

Grounded grid amplifier. In Fig. 20·8*b*, the r-f input signal is applied to the cathode while the grid is grounded. If necessary to maintain a d-c potential at the grid, it can be grounded only for r-f signal by means of a bypass capacitor. Note that the grounded grid is actually connected to the low side of the cathode load inductance L_k. Therefore, the input signal voltage across R_k is connected between cathode and control grid. The plate current is varied by the input signal voltage to produce amplified output in the plate circuit, which is coupled by L_pL_g to the mixer grid. The advantage of this arrangement is that the grounded grid shields the plate signal output from the cathode signal input, preventing feedback through C_{gp}. A disadvantage, however, is the difficulty in controlling the amplifier gain by AGC bias.

Neutralized triode amplifier. The r-f amplifier in Fig. 20·9 uses the grounded cathode circuit but the triode is neutralized to prevent oscillations. This method is plate neutralization. Some signal from the plate circuit is fed back to the grid through the adjustable neutralizing capacitor C_N. The neutralizing signal has opposite polarity from the plate signal, because of the tap on L_3. Therefore, the signal coupled by C_N opposes the feedback through C_{gp} from the plate. When both signals have the same amplitude, the amplifier is neutralized. If the tap on L_3 is at the center, the required value of C_N equals C_{gp} to neutralize the amplifier.

Cascode amplifier. This circuit illustrated in Fig. 20·10 combines a grounded grid stage V_2 driven by a grounded cathode stage V_1, in a twin-triode tube. Although triodes are used for low noise, the V_1 stage is usually not neutralized because its gain is low with a low value of plate load impedance. Most of the gain is in the grounded grid stage, which needs no neutralization. The combined gain of the cascode amplifier is 1.5×8, approximately, or 12. As a result, the cascode r-f amplifier has the low noise figure of a triode but with the gain and selectivity of a pentode. AGC bias is applied to the grounded cathode section.

In Fig. 20·10, note the d-c coupling from the plate of V_1, through L_k to the cathode of V_2. As a result, the internal plate-cathode resistances r_p of the two tubes form a d-c voltage divider across the B supply. The plate circuit of V_2 connects to +250 volts, but the plate of V_1 obtains its plate voltage, approximately equal to one-half B+, by the voltage division across the r_p of the two tubes. Note that the direct plate current of V_1 flows through V_2 to return to the B supply. Since the plate of V_1 connects to the cathode of V_2 directly through L_k, this cathode has a d-c potential of +125 volts with respect to chassis ground. Therefore, the R_2R_3 voltage divider is used to supply +123 volts to the control grid of V_2, so that its grid-to-cathode d-c voltage is −2 volts. Otherwise, V_2 would be cut off by an effective cathode bias of 125 volts, negative at the control grid,

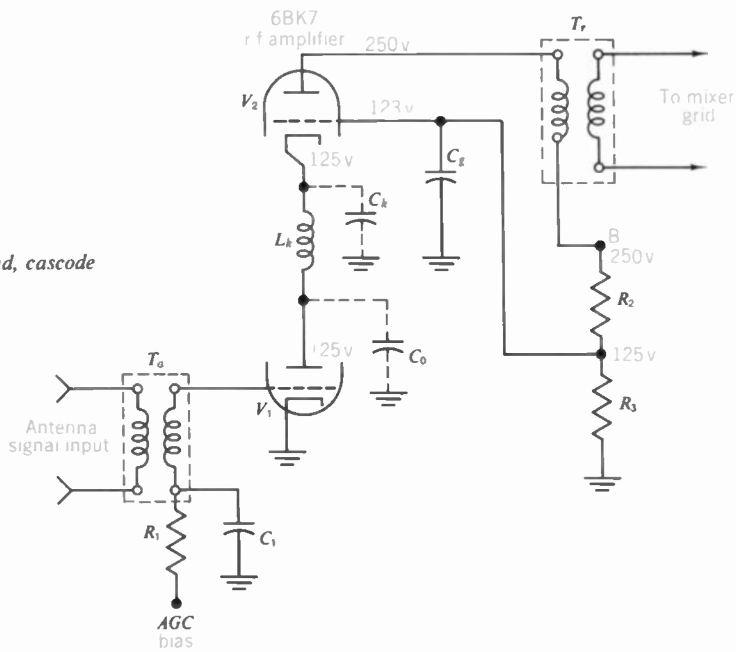


Fig. 20·10 D-c coupled, cascode r-f amplifier circuit.

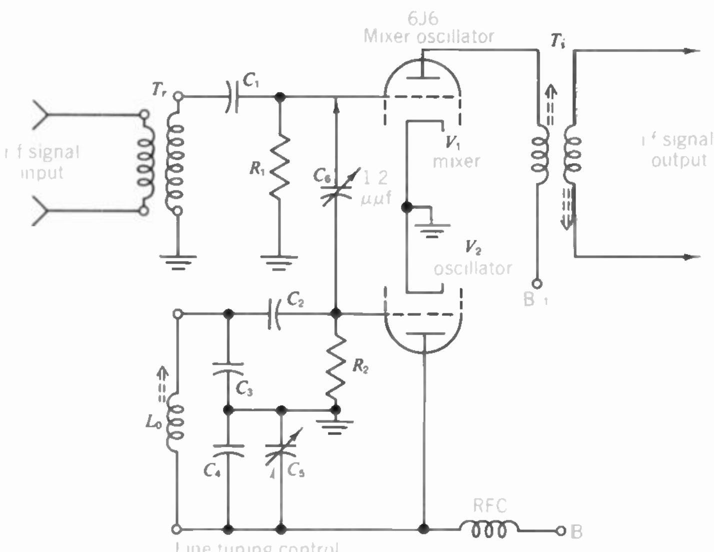
if it were connected directly to ground. C_g is an r-f bypass to return the grid to ground for r-f signal while maintaining the required d-c grid voltage. It should be noted that the cascode amplifier needs relatively high B+ because of the voltage division between the two sections.

The r-f signal input is applied by T_a to the grid of V_1. Its plate load impedance is the resonant circuit formed by L_k, C_k, and C_o, which is the output capacitance from plate to cathode of V_1. They form a π-type filter coupling circuit. The resonant rise in voltage across C_k provides input signal for the grounded grid stage V_2. In the output circuit, the amplified r-f signal in the plate circuit of V_2 is coupled by T_r to the mixer grid. The r-f tuning for each station is necessary only for the transformers T_a and T_r, as the interstage coupling network between V_1 and V_2 provides a resonant response broad enough to include all 12 VHF channels.

20·4 The mixer stage

Figure 20·11 illustrates a typical mixer-oscillator circuit that converts the r-f signal frequencies of the selected channel to the intermediate frequencies of the receiver. Using the 6J6 twin triode, the V_2 section operates as the local oscillator while V_1 is the mixer stage. The input to the mixer includes the desired r-f signal from the r-f amplifier coupled by T_r, and output from the local oscillator coupled by C_6 to the mixer grid. The difference frequency between the r-f picture carrier and the oscillator frequency is the picture carrier intermediate frequency for the picture i-f amplifier, while the difference between the r-f sound carrier and the

Fig. 20·11 Mixer-oscillator circuit for converting r-f signal to i-f signal. The r-f transformer T_r and oscillator coil L_o are changed for each channel.



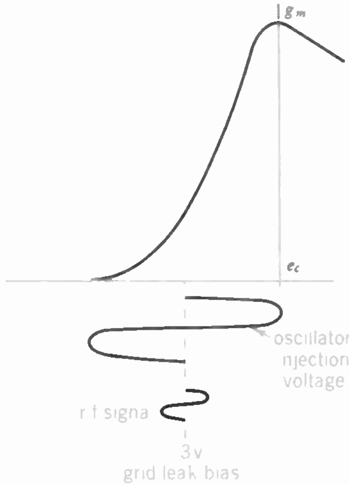


Fig. 20·12 Operation of the mixer stage as a detector for combined waveform of oscillator and r-f signals.

oscillator is the sound i-f carrier frequency. Note that the same oscillator signal beats with both the picture and sound carriers. Therefore, the separation between picture and sound carrier frequencies in the i-f output of the mixer is the same 4.5 Mc as between the two r-f carriers. T_i is the mixer plate i-f transformer, coupling the picture and sound i-f carrier signals, with their side-band frequencies, to the grid of the first i-f stage.

It is preferable to use separate oscillator and mixer stages in the television receiver because of the high signal frequencies and wide range of the television band. In UHF applications, the mixer is often a crystal diode because converter noise is reduced and less oscillator injection voltage is required.

Interaction between the r-f signal circuits and the oscillator is minimized by using loose coupling from the oscillator to the mixer grid, as illustrated by the 1- to 2-$\mu\mu$f capacitance for C_6 in Fig. 20·11. This adjustment varies the oscillator injection voltage at the mixer grid. The required amplitude is usually about 3 volts peak value. This produces grid-leak bias of approximately 3 volts, which is a d-c voltage, negative at the mixer grid. Note that the oscillator injection of 3 volts is very large compared with r-f signal of about 1 mv at the mixer grid.

Rectification in the converter circuit. The mixer is the first detector in the superheterodyne receiver, rectifying the combination of r-f signal and oscillator voltage at the grid to produce i-f signal output in the plate circuit. The nonlinear operation required for detection is illustrated in Fig. 20·12. Because of the large oscillator signal, it determines the amount of grid-leak bias. For the example here, the bias reduces the transconductance to approximately one-third the maximum g_m. As the large oscillator voltage swings the grid voltage, the g_m of the tube varies for the small r-f signal. The result is rectification of the combined waveform of the two grid signals. The combined waveform has an envelope corresponding to the difference frequencies desired as the i-f signal. The rectification allows the envelope signal to produce maximum output across the tuned i-f load impedance in the plate circuit of the mixer.

Conversion transconductance. This is defined as

$$g_c = \frac{\Delta i_p \text{ (for i-f signal)}}{\Delta e_g \text{ (for r-f signal)}} \qquad (20\cdot 1)$$

As an example, if the i-f signal current in the plate current is varied 3 μa by a change of 2 mv r-f grid signal, g_c equals 1,500 μmhos.

$$g_c = \frac{3 \times 10^{-6}}{2 \times 10^{-3}} = 1.5 \times 10^{-3}$$

or

$$g_c = 1{,}500 \times 10^{-6} \text{ or } 1{,}500\ \mu\text{mhos}$$

In general, g_c is about one-third the maximum g_m of the tube, the exact value depending on the amount of oscillator injection voltage.

Conversion gain. This is the ratio of i-f signal voltage output to r-f signal voltage input. This gain depends on g_c and Z_L of the mixer plate circuit tuned to the i-f signal. Specifically,

$$\text{Conversion gain} = g_c \times Z_L \qquad (20 \cdot 2)$$

With g_c of 1,500 μmhos and an assumed Z_L of 4,000 ohms,

$$\text{Conversion gain} = 1{,}500 \times 10^{-6} \times 4 \times 10^3$$

or

$$\text{Conversion gain} = 6$$

With an r-f signal input of 2 mv at the mixer grid, therefore, the i-f signal voltage is 12 mv at the mixer plate, with a conversion gain of 6.

20·5 The Local Oscillator

Its function is to generate unmodulated r-f sine-wave voltage at the frequency required for beating with the r-f signal to produce the difference frequencies equal to the receiver's intermediate frequencies. For any one station, the oscillator operates at only one frequency. The oscillator may beat either above or below the r-f signal frequencies but usually is above.

Table 20·1 lists the oscillator frequency required for each of the 12

*Table 20·1 Local oscillator frequencies**

Channel number	*Frequency band, Mc*	*Picture carrier, Mc*	*Sound carrier, Mc*	*Oscillator, Mc*
2	54–60	55.25	59.75	101
3	60–66	61.25	65.75	107
4	66–72	67.25	71.75	113
5	76–82	77.25	81.75	123
6	82–88	83.25	87.75	129
7	174–180	175.25	179.75	221
8	180–186	181.25	185.75	227
9	186–192	187.25	191.75	233
10	192–198	193.25	197.75	239
11	198–204	199.25	203.75	245
12	204–210	205.25	209.75	251
13	210–216	211.25	215.75	257

* For VHF channels, with picture i-f carrier of 45.75 Mc in receiver.

VHF channels. These values are figured as 45.75 Mc above the r-f picture carrier for each channel, or 41.25 Mc above the r-f sound carrier. The same method can be used to calculate the oscillator frequency for each of the UHF channels listed in Appendix A. If the receiver has the picture i-f carrier at 25.75 Mc, the oscillator frequency is 25.75 Mc above the r-f picture carrier frequency.

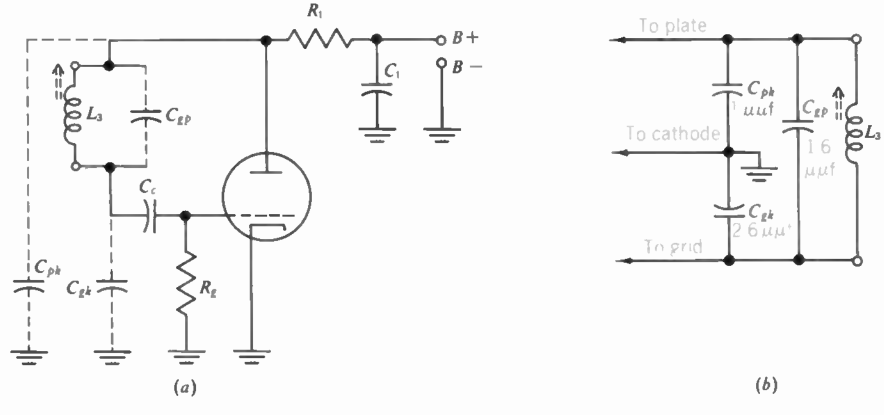
Assuming separate oscillator coils for each channel, there is usually a slug to align each coil for the correct oscillator frequency as the station selector is set for different stations. In addition, the oscillator generally has a fine tuning control for exact adjustment of the oscillator frequency. The fine tuning control can also be arranged to vary the oscillator slug, especially in tuners for remote control. Then with exact tuning set manually for each channel, retuning is not necessary when channels are changed by remote operation. With this pre-set fine tuning, the control is pushed in or out temporarily to make the adjustment.

The oscillator output voltage for the mixer is about 2 to 5 volts peak value. Since the oscillator operates with grid-leak bias produced by the feedback voltage, its operation can be checked by measuring the d-c bias. The frequency of the oscillator output is the resonant frequency of its tuned circuit.

Oscillator circuit. Most common is the circuit in Fig. 20 · 13, called an *ultraudion* oscillator. This is a modified Colpitts circuit, using the tube capacitances as the voltage divider for feedback. Note that the oscillator coil L_3 from plate to grid is across C_{gp}. Furthermore, C_{pk} and C_{gk} form a capacitive voltage divider across L_3. The oscillator voltage across C_{gk} is the feedback voltage for the grid.

Using the resonant frequency formula $f_r = 1/(2\pi\sqrt{LC})$, we can calculate f_r for a given value of L_3 in Fig. 20 · 13. Assume L_3 is 1 μh. The total C

Fig. 20 · 13 (a) Ultraudion oscillator, a modified Colpitts circuit. (b) How tube capacitances form capacitive voltage divider for feedback. Values given for 6J6 triode section with shield cover.



equals C_{gp} in parallel with the series combination of C_{gk} and C_{pk}. For 2.6 $\mu\mu$f in series with 1 $\mu\mu$f, the combination has a capacitance of 0.7 $\mu\mu$f, approximately. This value added to the parallel C_{gp} of 1.6 $\mu\mu$f makes a total capacitance of 2.3 $\mu\mu$f across L_3. The resonant frequency of this oscillator-tuned circuit with 1 μh inductance and 2.3 $\mu\mu$f capacitance, therefore, is 105 Mc.

Similarly, the inductance of L_3 can be calculated for a specific resonant frequency, using the formula $L = 1/(4\pi^2 f_r^2 C)$. As an example, for the oscillator frequency of 113 Mc to tune in channel 4, the required value of L_3 is

$$L = \frac{1}{4\pi^2 f_r^2 C} = \frac{0.025}{(113)^2 \times 10^{12} \times 2.3 \times 10^{-12}}$$
$$= 0.85 \times 10^{-6} \text{ or } 0.85\ \mu\text{h}$$

Fine tuning control. A typical arrangement is illustrated by the fine tuning control C_5 in Fig. 20·11. This capacitance is across only part of the tuned circuit, to reduce its effect on oscillator frequency. Varying the control changes the frequency within the range of about $\pm$1.5 Mc.

The fine tuning control is adjusted for the best sound in split-sound receivers. The oscillator is then at the frequency that produces the correct sound i-f carrier frequency for the sound i-f stages. This automatically results in the best picture also, when the picture and sound i-f amplifiers are correctly aligned with the required 4.5-Mc separation and the receiver has the normal amount of antenna signal. With intercarrier-sound receivers, however, the fine tuning control is adjusted for the best picture. This is determined by adjusting the oscillator frequency to the point of sound bars in the picture and then backing off a little for more contrast and a clear picture with maximum detail. These effects on the picture result because changes in oscillator frequency move the i-f picture carrier up or down the slope of the i-f response curve. Since the correct 4.5-Mc sound signal is produced as the difference-frequency beat between the picture and sound carriers, the sound is relatively independent of the exact local oscillator frequency for intercarrier-sound receivers.

20·6 R-F alignment

This consists of two separate jobs:

1. The local oscillator must be set to the correct frequency for tuning in each channel. Usually, there is an alignment adjustment for every channel to obtain the exact local oscillator frequency.
2. The r-f signal circuits, from antenna input to mixer grid, are aligned for maximum r-f gain with the required bandwidth. See the desired r-f response curve in Fig. 20·6.

Realignment of the r-f signal circuits is seldom necessary. However, the local oscillator frequency may need readjustment for each channel when

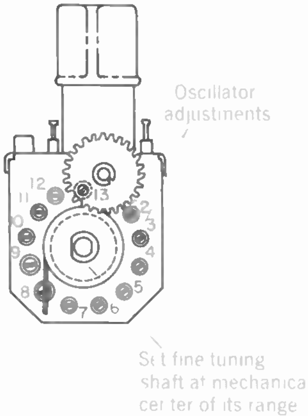


Fig. 20·14 Front view of tuner, showing local oscillator frequency adjustments. (RCA.)

the oscillator tube is replaced. Also, excessive fine tuning can be eliminated by more exact oscillator alignment.

Oscillator alignment. R-f tuners that switch inductances for each selected channel have individual oscillator coils with a slug that can be adjusted to set the oscillator frequency for each station. These are provided in addition to the fine tuning control, as installation or servicing adjustments. The oscillator coil slugs can usually be adjusted through the front panel of the receiver, without removing the chassis (see Fig. 20·14). The oscillator coils are adjusted for the correct oscillator frequency on each selected channel, within the middle range of the fine tuning control.

When the desired stations are broadcasting, the oscillator frequency can be set by adjusting for the best sound and picture. This method is the easiest and best because the broadcast stations supply accurate carrier frequencies. In split-sound receivers, adjust the oscillator for the best sound, with maximum volume and minimum background noise, which indicates balance in the FM sound detector. Since the bandwidth of the sound i-f stage is very narrow compared with the oscillator frequency, an accurate adjustment can be made. With the sound and picture i-f sections correctly aligned, this also results in the proper response for the i-f picture signal.

In intercarrier-sound receivers, however, the best picture is used to indicate the correct frequency of the oscillator, since the sound output is independent of the exact oscillator frequency. The best picture is indicated by good contrast with maximum detail. This is slightly off the point of sound in the picture. Refer back to Fig. 19·14 to see how the audio sound bars look. Check both the sound and picture for each channel while adjusting each oscillator coil.

If no broadcast stations are available to supply r-f input to beat with the oscillator, or if the tuner is not connected to the receiver to check i-f output, the oscillator frequency can be measured directly. One method is to heterodyne the oscillator output against a crystal-calibrated marker generator for zero beat. The unit in Fig. 20·15 has its own audio section

Fig. 20·15 Crystal-calibrated marker, oscillator. (RCA.)

so you can hear the beat and adjust for zero output or a very low-frequency beat.

In most switch-type tuners the oscillator frequency is decreased for lower channels by adding series inductances. Therefore, the higher channels must be adjusted first. If the oscillator adjustment is changed for one channel, all the lower channels are affected. In some tuners with parallel inductances the lower channels are adjusted first. These tuners may have one oscillator adjustment for two or three consecutive channels. With turret tuners, the oscillator adjustments can be made in any order because each is independent of the others.

R-F response curve. The r-f circuits tuned to the signal frequencies of each selected channel, from antenna input to mixer grid, are aligned by the visual response curve method. Make sure the generator and receiver are set to the same channel for r-f alignment.

The procedure is similar to the method used for obtaining the i-f response curve. However, the sweep generator is connected to the antenna input terminals of the receiver and set to sweep the band of r-f signal frequencies in the channel to which the station selector is set. Also, the vertical input terminals of the oscilloscope are connected to the mixer grid circuit to indicate the relative amplitude of the r-f signal frequencies. A test point is usually provided on the tuner for this oscilloscope connection, through a decoupling resistor of about 50,000 ohms to the mixer grid. The grid-leak detector action here results in rectified r-f signal for the oscilloscope. A typical r-f response curve is shown in Fig. 20·6.

There are generally two or three trimmer capacitors to align the tuned circuits for grid and plate of the r-f stage and the mixer grid. They are adjusted for maximum height on the response curve, with the required bandwidth. Generally the r-f trimmers are adjusted on channel 10. With

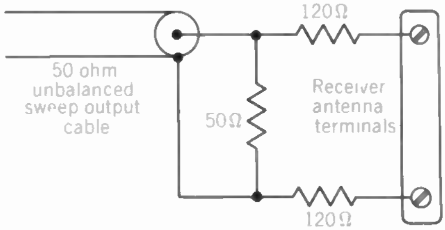


Fig. 20·16 Resistance pad for matching between 50 ohms unbalanced and 300 ohms balanced impedances.

correct r-f alignment, the required r-f response is normally obtained for all channels, although every channel does not have exactly the same curve. If necessary, the coils for individual channels can be squeezed or spread for best alignment. When neutralizing adjustments are needed, adjust for minimum height on the response curve, which indicates maximum neutralization. Generally, a triode r-f amplifier has a neutralizing capacitor adjustment.

In order to avoid distortion of the r-f response curve by the sweep generator connection, its output impedance must be matched to the receiver input. This is necessary because the response of the antenna input circuit is included in the r-f response curve. Figure 20·16 shows a typical resistance pad. In addition to matching between 50- and 300-ohm impedances, the series resistances isolate the shunt capacitance of the generator cable, which can detune the r-f input circuit. The matching pad is needed here because the antenna circuit is the only part of the receiver where the signal generator cannot be connected ahead of the tuned circuits to be aligned. In general, do not align a tuned circuit connected to the signal generator unless a matching pad is used. Otherwise the circuit will be aligned incorrectly.

20·7 Conversion methods for UHF channels

A practical question for the r-f tuner is how to provide for UHF channels[1] 14 to 83, in addition to VHF channels 2 to 13. A typical method is illustrated by the combination VHF-UHF tuner in Fig. 20·17. Note that there are two separate tuners, but both are used for UHF operation. For the VHF channels, the VHF tuner operates alone. For UHF operation, the VHF tuner is an i-f amplifier for the output from the UHF tuner. The fine tuning control on the VHF tuner can operate the UHF tuner for continuous tuning through the entire UHF band from 470 to 890 Mc. Or, the UHF tuning can be a separate control. In either case, continuous tuning for UHF channels operates like the tuning in a radio, as you tune through the band for the selected channel. Then the UHF channel numbers are on the fine tuning control.

The UHF setting on the station selector is an extra switch position between channels 2 and 13. On this position, B+ voltage is supplied to the 6AF4

[1] As of 1963, all television receivers shipped in interstate commerce must be able to tune in VHF and UHF channels.

UHF oscillator, while B+ is removed to disable the VHF oscillator. Also, the r-f circuits of the VHF tuner are resonated for the i-f range of 41 to 47 Mc by adding the required inductances. In this way, the r-f amplifier and mixer stages of the VHF tuner are used for additional i-f amplification when a UHF channel is tuned in. The single frequency conversion in the crystal diode mixer enables the UHF oscillator to heterodyne the UHF signal frequencies down to the i-f range of 41 to 47 Mc for the selected channel. This i-f output from the UHF mixer is connected to the VHF tuner by a short length of coaxial line.

As an example, suppose that UHF channel 14 (470 to 476 Mc) is to be tuned in. The station selector is set to the UHF position, to apply B+ to the UHF oscillator, disable the VHF oscillator, and tune the VHF r-f circuits to 41 to 47 Mc. The fine tuning control is varied to drive the UHF tuning assembly by means of a belt or gear drive. For this example, then, the UHF r-f circuits are tuned to 470 to 476 Mc for the channel 14 frequencies. Most important, though, the UHF oscillator is tuned to 517 Mc, which is 45.75 Mc above the channel 14 picture carrier at 471.25 Mc. The r-f signal from the UHF antenna input terminals and the UHF oscillator output are both coupled into the diode mixer. Its output, then, is i-f signal at 41 to 47 Mc, corresponding to the difference frequencies between the oscillator at 517 Mc and the channel 14 frequencies at 470 to 476 Mc.

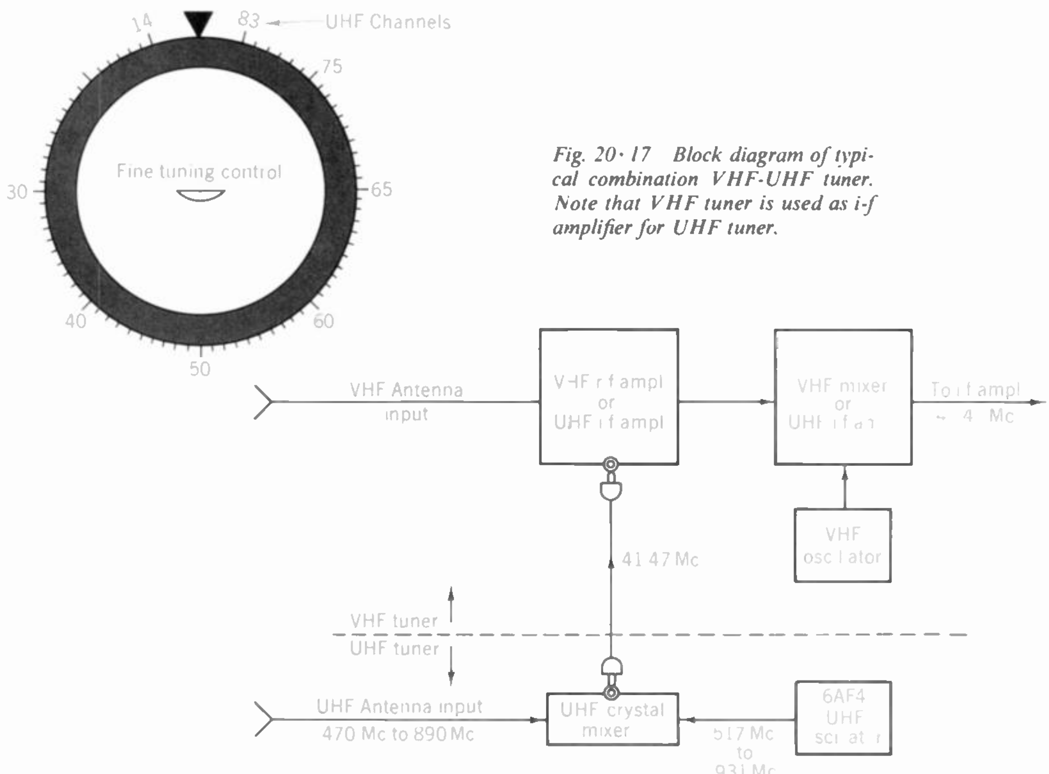


Fig. 20·17 Block diagram of typical combination VHF-UHF tuner. Note that VHF tuner is used as i-f amplifier for UHF tuner.

The i-f signal from the UHF mixer then goes to the VHF tuner, where the r-f amplifier and mixer input circuits are tuned to 41 to 47 Mc. As a result, the selected UHF channel is amplified as an i-f signal by these two stages, in addition to the normal gain in the i-f amplifier section of the receiver.

There is no r-f amplifier for the UHF tuner because the signal-to-noise ratio is better with just the tuned preselector circuits, compared with a conventional UHF amplifier. Similarly, a diode is used for the UHF mixer for minimum converter noise. The crystal diode also has much less capacitance than a vacuum tube. Note that the additional i-f amplification at 41 to 47 Mc compensates for less r-f amplification on UHF operation.

Continuous tuning through all the UHF channels is no problem because of the relatively small frequency ratio. For the oscillator, the highest frequency at 931 Mc is only 1.8 times greater than the lowest frequency at 517 Mc. The UHF oscillator beats above the channel frequencies, assuming that the VHF oscillator does also, in order to have the same inversion of i-f side frequencies for either VHF or UHF channels.

A receiver with a combination tuner has two sets of input terminals, one for a VHF antenna and the other for a UHF antenna. When a combined VHF-UHF antenna is used with one transmission line, it can be connected to the receiver by the method shown in Fig. 20 · 18. The 3¼-in. quarter-wave stubs are shorted at the end to isolate the VHF terminals from the UHF terminals for the UHF channel frequencies. Twin-lead transmission line is generally used for minimum losses.

The VHF antenna can also be used for the UHF band, when there is enough signal strength and no reflections. At harmonic frequencies, the directional pattern splits into many lobes around the antenna, with the main lobes directed toward the ends. In fact, the VHF antenna may have appreciable gain for the UHF band, compared with a UHF dipole. However, there may be a problem with ghosts because of the omnidirectional response. Also, it should be noted that, although the UHF antenna usually has more antenna gain compared with VHF reception, there is less antenna signal voltage.

UHF conversion strips. Turret tuners can be adapted for UHF reception by inserting UHF conversion strips in place of the VHF strips for unused channels. Up to four UHF strips may be installed, but there should be a

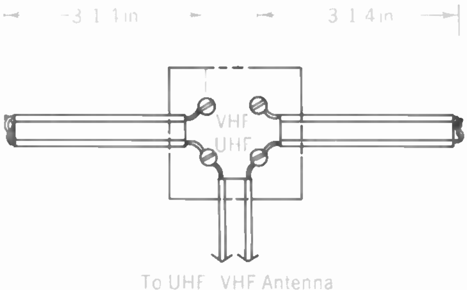


Fig. 20 · 18 Receiver input connections for combination VHF-UHF antenna.

spacing of at least two channel positions between UHF strips. These strips are for 12-step tuners that do not have the extra UHF position.

In a turret tuner having 13 positions, the extra position is for UHF operation, with the addition of a UHF tuner operated by the fine tuning control. On the UHF position, the strips tune the VHF r-f amplifier and mixer stages to the 41- to 47-Mc signal output from the UHF tuner.

For VHF switch-type tuners that have a UHF position but no UHF tuner installed, a kit is generally available to add UHF tuning. In addition to the UHF tuner, the adapter kit includes an extra switch wafer for the wiring needed on the UHF position.

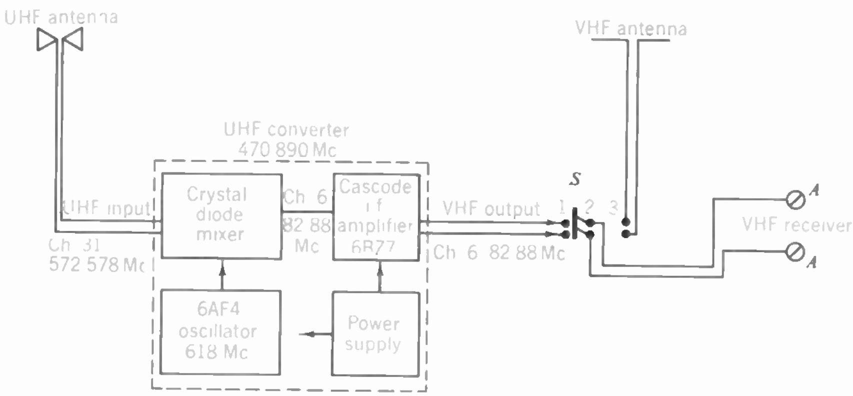
UHF converters. For a 12-position switch tuner without the extra UHF position, the receiver can be adapted to receive UHF channels by adding a frequency converter. This is an external UHF tuning unit connected between the antenna and the input to the VHF receiver (see Fig. 20·19). In this case, the converter must heterodyne the UHF channel down to the frequencies of a VHF channel for input to the VHF tuner. Generally, the converter output is on either channel 5 (76 to 82 Mc) or channel 6 (82–88 Mc), depending on which is not used for a station in the area. The VHF tuner is then set to channel 5 or 6 for the converter output.

The stages in the converter include its own power supply, r-f preselector circuits for the UHF channels, UHF oscillator, crystal diode mixer, and a fixed-tuned stage for the mixer output on either channel 5 or 6 frequencies. This stage is considered an i-f amplifier in the converter, as the same VHF channel frequencies are amplified for any converted UHF channel. A cascode amplifier is often used here for low noise. The oscillator in the converter operates below the UHF signal frequencies so that there will be no inversion of the side frequencies. Then the converted signal has the same r-f side bands as a standard VHF channel.

20·8 *Types of R-F tuner circuits*

Figure 20·20 illustrates the four main types of VHF tuners on the basis

Fig. 20·19 Operation of frequency converter, illustrated for UHF channel 31 converted to VHF channel 6.



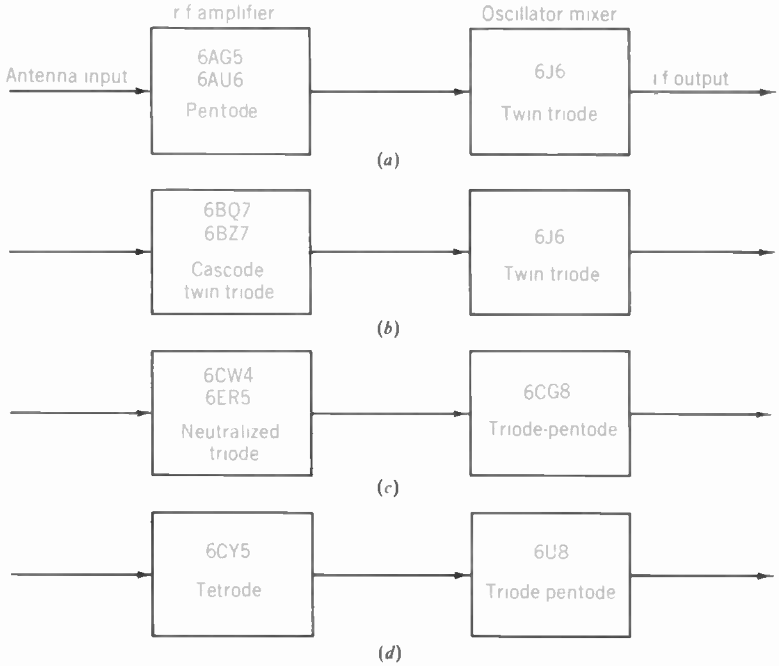


Fig. 20·20 Types of VHF tuner circuits with typical tubes.

of tubes used and the different r-f amplifier circuits. The conventional pentode stage in *a* was used in older tuners. To improve the signal-to-noise ratio, later tuners used the cascode amplifier for the r-f stage, as in *b*. However, the cascode circuit is not so well adapted to the additional function of i-f amplifier in combination VHF-UHF tuners, as the switching is simpler for a triode or tetrode stage. Special low-noise r-f amplifier tubes maintain a good signal-to-noise ratio. In addition, a pentode mixer, as in *c* and *d*, provides more gain than a triode when the mixer functions as an i-f amplifier for UHF operation.

A triode for the r-f amplifier, as in *c*, features low tube noise. However, neutralization is necessary to compensate for feedback through the internal grid-plate capacitance. The 6CW4 is a miniature nuvistor type, though, with very small interelectrode capacitances (see Fig. 6·10). The 6ER5, 6FH5, and 6ES5 feature a guided grid construction, with internal beam-forming plates at cathode potential to reduce C_{gp}. Finally, the 6CY5 in *d* is a low-noise tetrode with only 0.03 $\mu\mu$f for C_{gp}, eliminating the need for any neutralizing adjustment.

The schematic diagram of a typical VHF tuner is shown in Fig. 20·21. This circuit uses a nuvistor for the r-f amplifier and a triode-pentode for the oscillator-mixer. For each channel strip, L_8 is the oscillator coil; L_6 with L_7 form the r-f transformer, coupling signal from the r-f amplifier plate to the mixer grid; and L_5 is the antenna input coil. The trimmer capacitors for r-f alignment are C_8 and C_{12}. The neutralizing capacitor is C_{21}. Because of lead inductance, some r-f signal voltage is available at the low end of the plate coil L_6 to couple back to the grid for neutralization.

Test point 1 is for the oscilloscope to observe the r-f response curve at the mixer grid in r-f alignment. Test point 2 is for injecting i-f signal in i-f alignment. Separate connections are necessary because test point 1 is bypassed for r-f signal by C_{18}. In the mixer plate circuit, L_{11} tunes with C_{28} and C_{16} to provide the converted i-f signal at the output jack J_1.

The oscillator is an ultraudion circuit with L_8 between plate and grid. Grid-leak bias is provided by C_{15} with R_{10}.

The symbol for C_{18} in Fig. 20·21 indicates a *feed-through capacitor.* A typical unit is shown in Fig. 20·22. This type of capacitor is generally used for bypassing in the r-f tuner. These are silvered ceramic disk capacitors. The two end terminals are connected internally to one capacitor plate, through a small hole in the center of the disk. The other capacitor plate is connected to the metal base, which is mounted on the chassis. This type of construction features low inductance and extremely high resonant frequency. In addition, the feed-through capacitor prevents radiation of signal from the leads that must go through the shield cover.

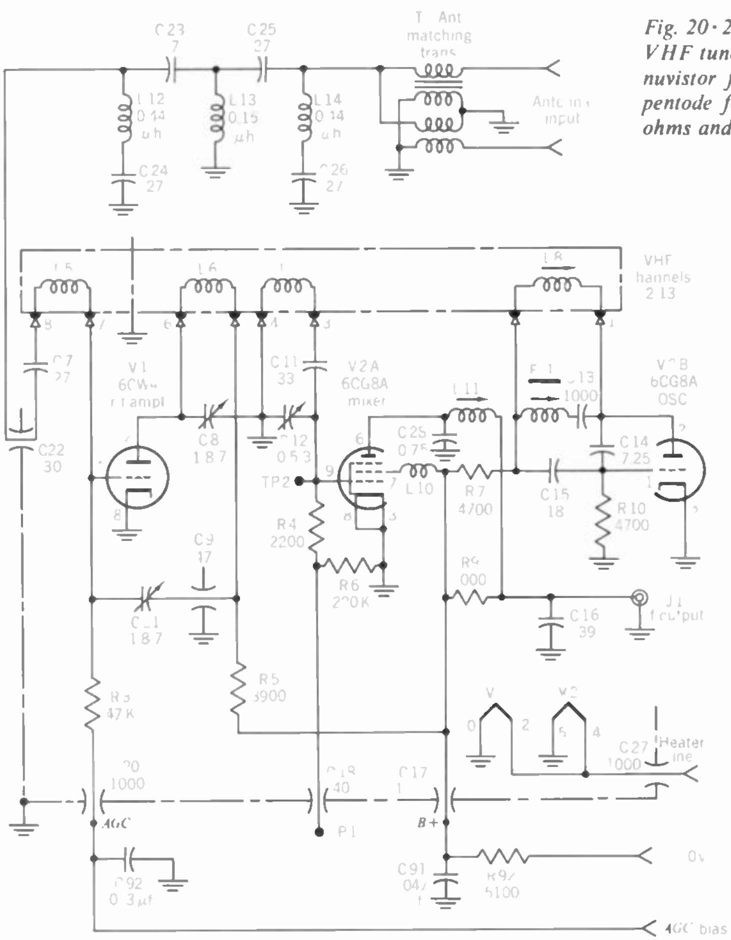


Fig. 20·21 Schematic diagram of VHF tuner with neutralized triode nuvistor for r-f stage and triode-pentode for oscillator-mixer. R in ohms and C in μμf. (RCA.)

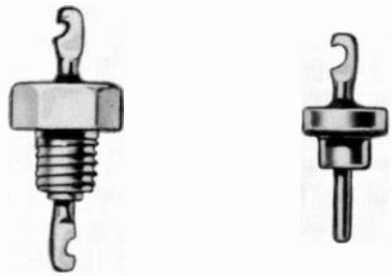


Fig. 20·22 Feed-through type of capacitors. Height ¾ in. (Allen-Bradley Co.)

20·9 *UHF tuner circuit*

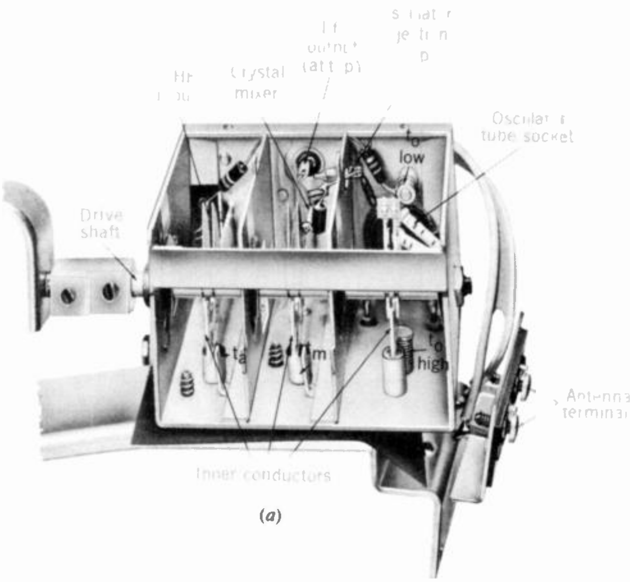
In Fig. 20·23, the metal cover is removed for an inside view. Each of the three sections is a resonant cavity for the r-f preselector, diode mixer, and 6AF4 oscillator circuits. The sides and partitions form a wall which is the outside conductor of the resonant cavities. The inner conductor for each cavity is the metal bar between the two plates of the variable capacitor. For the UHF band the required length is about 2½ in. Each capacitor varies the resonant frequency of the cavity by varying the electrical length of the inner conductor. In the schematic diagram, the three capacitors are C_{2A}, C_{2B}, and C_{2C}, ganged for continuous tuning through the UHF band. The three inner conductors are L_2, L_3, and L_6.

In the preselector and mixer cavities, the inner conductor is attached at one end to the metal wall as chassis ground. The opposite end has series capacitive coupling to chassis ground, through the tuning capacitors. In the oscillator section, neither end of the inner conductor is grounded directly, so that one side can be connected to the 6AF4 control grid for signal voltage. The oscillator is an ultraudion circuit with the inner conductor between control grid and the plate grounded for signal voltage by C_9. In the cathode circuit, L_7 is resonated below the oscillator frequency to provide capacitive reactance that cancels the effect of the internal cathode lead inductance.

The UHF input signal from the antenna terminals is coupled by a small loop L_1 to the preselector cavity for r-f tuning with L_2 and C_{2A}. Additional r-f selectivity is provided by the mixer tuning with L_3 and C_{2B}. The r-f signal is coupled into the mixer circuit by the open slot between the two cavities. The oscillator output is injected into the mixer cavity by the pickup loop L_5. With rectification of the combined r-f and oscillator input to the diode mixer, i-f output is produced at the difference frequencies of 41 to 47 Mc. In the mixer output circuit, the i-f coil L_4 forms a π-type low-pass filter, with the shunt capacitance of the feed-through capacitor C_3 and the capacitance of the i-f output cable at test point X. This i-f signal for the desired UHF channel then goes to the VHF tuner for more amplification at 41 to 47 Mc on UHF operation.

20·10 *Wireless remote control*

Many receivers have provision for remote operation in changing channels, varying the volume, and turning the set on or off. The ultrasonic system



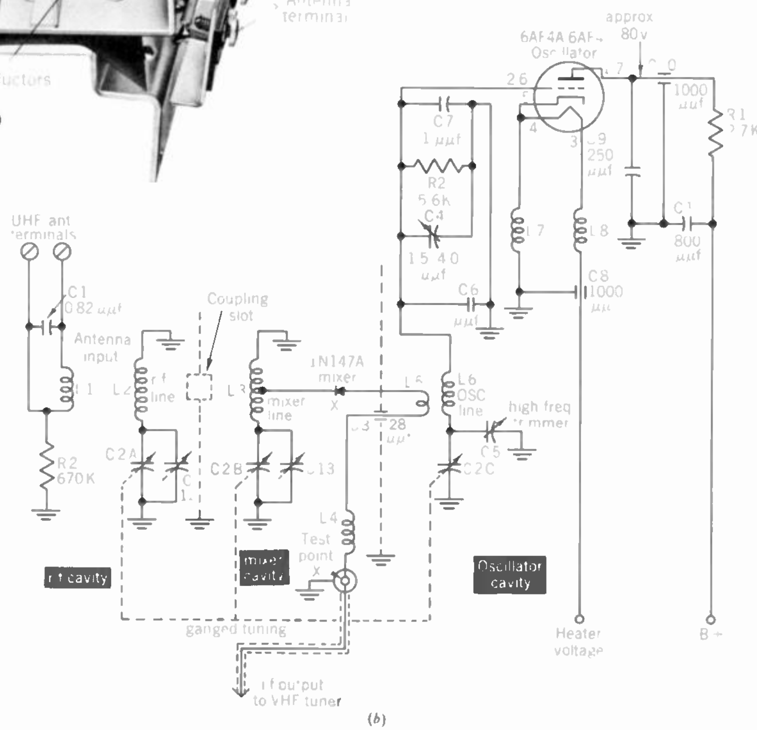


Fig. 20·23 UHF continuous tuner. (a) Photograph. (b) Schematic diagram. (Zenith Radio Corp.)

illustrated in Fig. 20·24 is generally used. The small hand-held remote transmitter radiates ultrasonic waves at frequencies around 40 kc. These vibrations are sound waves of compression and rarefaction traveling through the air but their frequency is too high for detection by the human ear. The velocity of propagation for sound waves is approximately 1,130 ft per sec.

A microphone at the receiver intercepts the ultrasonic waves, converting them to electric signal for the remote-control chassis. This chassis is in the cabinet to operate the receiver from the remote transmitter. Amplification and detection of the remote signals are necessary for enough d-c control

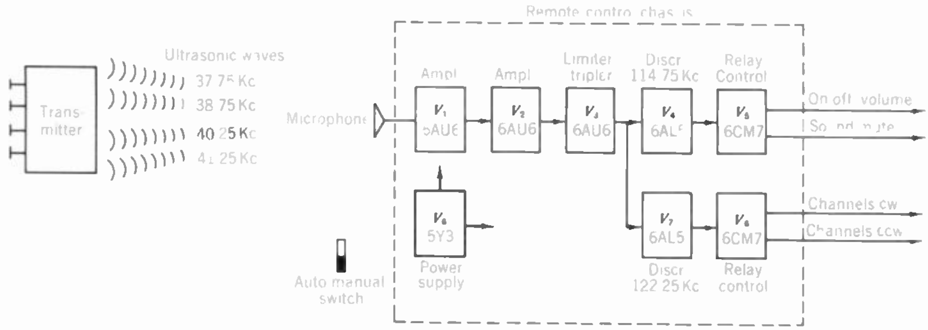


Fig. 20·24 Wireless remote-control system for four receiver functions.

voltage to operate the relays. Separate relays provide the required switching functions and operate the drive motor for the channel selector.

Ultrasonic waves are generally used as the radiated signal for remote control, in preference to radio transmission, to minimize interference problems. It should be noted that sound waves do not pass easily through the walls in a room. The remote transmitter can usually operate the control unit from a distance of at least 20 ft.

Control frequencies. Different control functions are possible by using specific frequencies for each. In Fig. 20·24, the transmitter produces four frequencies for the following four functions:

37.75 kc......Turn set on or off and vary volume in steps
38.75 kc......Sound on or off (muting)
40.25 kc......Turn channel selector clockwise (cw)
41.25 kc......Turn channel selector counterclockwise (ccw)

If the muting function is not used and the channels are changed in only one direction, then only two control frequencies need be used.

Transmitter unit. The remote transmitter unit may be either a mechanical vibrating source of ultrasonic waves or the electronic oscillator type. In the mechanical type, there is a separate metal tuning rod, similar to a tuning fork, for each frequency (see Fig. 20·25). When the control button is pushed, a spring-loaded hammer strikes the rod and it rings at its resonant ultrasonic frequency. The ultrasonic waves are radiated directly through an open grille at the top of the unit. No power supply is needed for this transmitter, as its operation is entirely mechanical.

The electronic transmitter unit contains transistorized oscillator circuits. The operating frequencies are also around 40 kc. Only two oscillators need be used, as four different beat frequencies can be obtained for the four separate functions. The oscillator output at ultrasonic frequencies drives a ceramic transducer to produce inaudible sound waves transmitted to the receiver. A small battery is the power supply for the transistors.

Control chassis. At the receiver, the ultrasonic waves are converted to

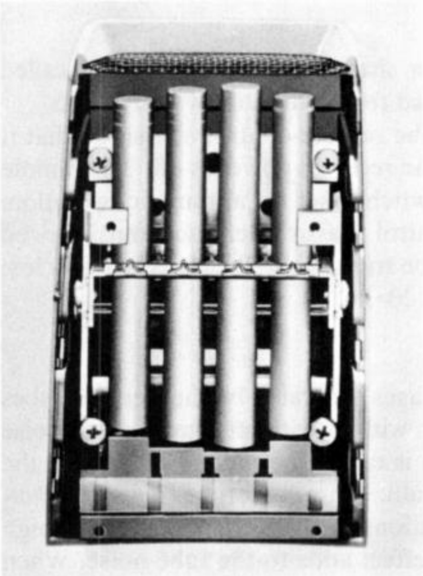


Fig. 20·25 Internal view of supersonic transmitter unit for remote control. (Zenith Radio Corp.)

electric signals by a ceramic microphone, with peak response at about 40 kc. Then the control signals can be amplified and detected in the remote-control chassis, as illustrated in Fig. 20·24. V_1 and V_2 are broad-band amplifiers, centered at 40 kc with enough bandwidth for any of the control frequencies. The tripler stage V_3 multiplies the input frequency for greater separation between the control frequencies for each function. This stage is also an amplitude limiter for interfering noise pulses.

Although there is no frequency modulation, the discriminator is used to detect the control signal according to whether its frequency is below or above the discriminator center frequency. As examples, the third harmonic of 41.25 kc at 123.75 kc is 1.5 kc above the center frequency of 122.25 kc for V_7; however, the third harmonic of 40.25 kc at 125.75 kc is 1.5 kc below center frequency. If one signal produces positive d-c output voltage from the discriminator, the other will produce negative d-c output. The same idea applies to the two control frequencies for V_4. There are then four different d-c output voltages.

The d-c control voltage is then integrated and coupled to a relay control tube. Twin triodes provide four stages for the four functions.

Each control tube has fixed bias more negative than cutoff. Conduction occurs when the discriminator voltage is enough to overcome the cutoff bias. As typical values, the bias on each control tube is about -20 volts. Then 30 volts added by the discriminator can allow plate current of about 10 ma. Each control tube has a relay in series with the plate circuit to B+. When plate current flows, the relay is energized to close or open its switch for the desired function.

The drive assembly for the channel selector includes an a-c motor operating from the 60-cps power line. When the appropriate control relay is closed, input power is applied to the motor, to rotate the channel selector. If desired, unused channels can be skipped by setting tabs on an index wheel that turns with the tuner shaft. This arrangement is called *programming*. A reversible motor is used for rotation in two directions.

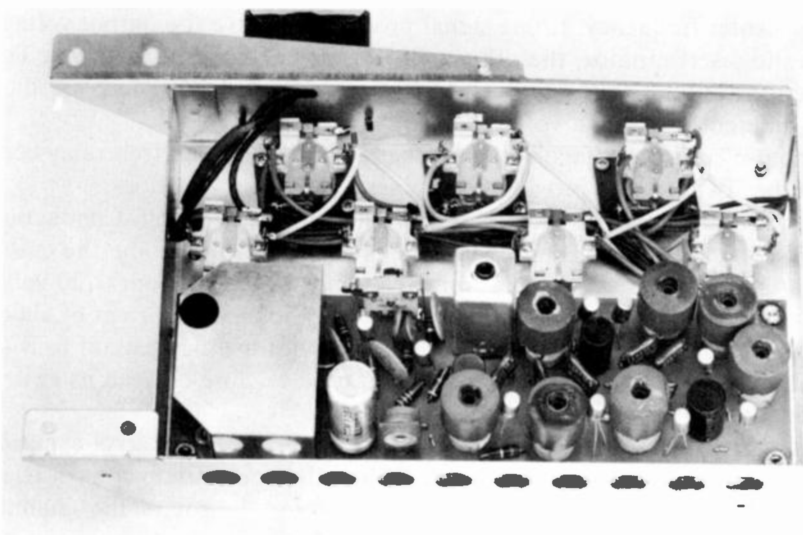
The auto-manual switch turns on the remote-control chassis so that it can operate the receiver even when the receiver power is off. The remote chassis has power on as long as the switch is set for automatic operation. It should be noted that the remote-control chassis often uses transistorized circuits. In addition to compactness, the transistorized chassis requires less power. A typical unit is shown in Fig. 20·26.

20·11 Receiver noise

This noise results from random voltages generated by the electron tubes and circuit components in the receiver, with or without signal input. Noise voltage produced by the components is called *thermal noise* because the source is thermal agitation. The main source of tube noise is *shot-effect noise* caused by random fluctuations of plate current. At very high frequencies, though, the transit-time effect adds to the tube noise. When the receiver has high gain, the amplified random noise voltages at the kinescope grid produce the white speckles called *snow*, as shown in Fig. 20·5. The snow is a visual reproduction of the continuous hissing or frying sound of receiver noise in an audio system.

Receiver noise is always present but may not be noticeable with strong r-f signal input. The noise has a fixed level, depending on the receiver circuits. The signal may be weak or strong. Furthermore, AGC action reduces the receiver gain for strong signals. Therefore, a strong signal makes the receiver noise level insignificant, as relatively little receiver gain

Fig. 20·26 Transistorized remote-control chassis mounted in receiver cabinet. Note relays at top, with transistors and 40-kc transformers on chassis. (RCA.)



is necessary to reproduce the desired signal. With a weak signal, however, the receiver operates at maximum gain. Then the noise generated in the first r-f stage is amplified enough to make it just as evident as the small r-f signal. As a result, receiver noise is primarily a problem in low-level stages, to obtain a good signal-to-noise ratio with weak signal input.

Thermal noise. The heat in a conducting material, even at room temperature, agitates the molecules and atoms in the conductor. The result is a random electron motion equivalent to a small noise current. This current through the resistance produces random noise voltages. The higher the temperature, resistance, and bandwidth, the greater is the generated noise voltage. Since there is no specific frequency for the random noise voltages, more noise is passed by a circuit with more bandwidth. The amount of rms noise voltage generated by any resistive impedance at room temperature can be calculated from the formula.

$$E_R = 1.28 \times 10^{-10} \sqrt{R \times \Delta F} \qquad (20 \cdot 3)$$

where ΔF is in cps and R in ohms. The constant factor 1.28×10^{-10} converts E_R to the volt unit, at room temperature equal to 20°C. As an example, the receiver input resistance of 300 ohms, with a bandwidth of 6 Mc, generates a thermal noise voltage:

$$E_R = 1.28 \times 10^{-10} \sqrt{300 \times 6 \times 10^6} = 1.28 \times 10^{-10} \times \sqrt{18 \times 10^8}$$
$$= 1.28 \times \sqrt{18} \times 10^{-10} \times 10^4 = 1.28 \times 4.24 \times 10^{-6}$$
$$E_R = 5.4 \times 10^{-6} \text{ or } 5.4\ \mu\text{v}$$

Shot-effect noise. This is caused by random fluctuations in the electron flow from cathode to plate, producing random noise voltages across the plate load impedance. The shot-effect noise voltage in the plate circuit increases with higher values of plate load impedance and bandwidth. Also, the more electrodes that can collect electrons, the greater is the shot effect because the increased partition of the emission current results in a more random effect in plate current. Triodes are less noisy than pentodes, therefore. The higher the transconductance (g_m) the lower is the shot-effect noise. Also, the noise is less with a low ratio of screen-grid current to plate current.

For a pentode r-f amplifier with bandwidth of 6 Mc, the shot-effect noise generated in the plate circuit is about 100 μv. Assuming a voltage gain of 10, this 100-μv noise voltage in the plate circuit is equivalent to 10 μv noise input to the grid circuit. For the twin-triode cascode r-f amplifier, the equivalent shot-effect noise in the grid circuit is about 7 μv. The shot-effect noise in the plate circuit is generally considered as equivalent noise voltage in the grid circuit, where it can be combined with the noise of the input circuit.

A mixer stage generates more shot-effect noise than the same tube used as an r-f amplifier because of low conversion transconductance. Pentagrid converter tubes produce the most noise because of the multigrid structure and low transconductance. Crystal diode mixers produce less noise than tubes but have a conversion loss. However, when the signal frequencies are so high that no conversion gain can be obtained with a tube, the crystal mixer can provide a better signal-to-noise ratio. If the antenna signal is coupled directly to a crystal diode mixer without an r-f stage, as is done in UHF circuits, the first i-f stage must be a low-noise amplifier.

Minimum r-f signal level. Receiver noise is the limitation on how much sensitivity is useful. High gain without a corresponding reduction in receiver noise can only produce a picture masked by snow. This is why more antenna signal or less receiver noise must be obtained to reduce snow in the picture.

An average value of receiver sensitivity is 32 μv r-f signal input to the r-f amplifier for 1 volt at the video detector. A typical value for total receiver noise at the input is 16 μv. The signal-to-noise ratio then is 2:1 with 32 μv signal. However, for a relatively noise-free picture, the desired signal-to-noise ratio should be about 30:1. For this case, the required r-f signal input is 480 μv. It should be noted, however, that receiver noise does not include external interference such as ignition noise. With typical external noise in city areas, an r-f input signal of several millivolts may be necessary for a good picture.

Excessive snow in the picture. Snow that is visible on the kinescope screen is caused primarily by shot-effect noise generated in the mixer tube. Therefore, the snow can be used as an indicator to localize a trouble between the r-f circuits preceding the mixer tube and the i-f section.

Suppose that the trouble is no picture and no sound, or weak picture and sound. If the trouble occurs on some channels only, this indicates a defect in the front end, but if the picture and sound are affected on all channels the trouble can be in either the r-f section or the common i-f amplifier section. To localize between the r-f and i-f sections, turn up the contrast and volume controls to maximum and note whether the mixer-tube noise can be seen as snow on the kinescope screen and heard as a hissing sound. When an increase in receiver noise is evident, this indicates that the mixer stage is operating and all the succeeding stages are amplifying the mixer-tube noise. Therefore, if the desired signal is not going through the receiver but the mixer noise is being amplified, the trouble probably is in the circuits before the mixer tube. Either the signal from the antenna is too weak or the r-f amplifier does not have enough gain. If there is little or no increase in receiver noise evident, this indicates the trouble is in the mixer tube or succeeding stages.

SUMMARY

1. VHF tuners are generally the turret or switch type for channels 2 to 13. The two tubes on this separate subchassis are the r-f amplifier and mixer-oscillator stages. Ganged tuning is provided by the station selector. The fine tuning control varies the local oscillator frequency.

2. UHF tuners are continuous tuners for channels 14 to 83. The two stages generally used are local oscillator and crystal diode mixer. The i-f output of the UHF mixer is amplified by the VHF tuner on UHF operation.
3. The r-f amplifier must have low tube noise for a high signal-to-noise ratio to minimize snow in the picture with weak antenna signal. Low-noise circuits for this stage are the cascode amplifier (Fig. 20·10) and the neutralized triode (Fig. 20·9).
4. Triode r-f amplifiers must be neutralized to prevent oscillations caused by feedback through the relatively large C_{gp}.
5. The input impedance of the tuner is generally 300 ohms This means 300-ohm twin-lead transmission line from the antenna should be used for an impedance match at the receiver on all channels.
6. Conversion transconductance g_c is the ratio of i-f signal plate current in the mixer to r-f signal voltage at the grid. Conversion gain is the ratio of i-f signal voltage output to r-f signal voltage input.
7. The local oscillator circuit most often used is a modified Colpitts type called the *ultraudion.* This oscillator circuit features one tuned circuit between plate and control grid, with interelectrode capacitances serving as the capacitive voltage divider for feedback.
8. The r-f amplifier can be aligned by the visual response curve method, for a symmetrical response 6 Mc wide.
9. The local oscillator generally has individual adjustments for each channel (see Fig. 20·14). Adjust the oscillator frequency for the best picture, in intercarrier-sound receivers.
10. Wireless remote control for changing channels is generally accomplished by the ultrasonic system illustrated in Fig. 20·24. The operating frequencies are around 40 kc.
11. Receiver noise produces snow in the picture with weak signal input. The two main sources are thermal noise in all components and shot-effect noise caused by random fluctuations of plate current in tubes. Thermal noise increases with more bandwidth. Shot-effect noise is greater for pentodes and tetrodes compared with triodes.
12. Two common mechanical troubles in tuners are a broken detent or dirty switch contacts. In both cases, the station-selector setting is too critical when tuning in a channel.

SELF-EXAMINATION *(Answers at back of book.)*

Choose (*a*), (*b*), (*c*), or (*d*).

1. The fine tuning control varies the frequency of the: (*a*) r-f amplifier; (*b*) mixer grid circuit; (*c*) antenna input circuit; (*d*) local oscillator.
2. With an i-f picture carrier frequency of 45.75 Mc, to tune in channel 13 the local oscillator frequency is: (*a*) 211.25 Mc; (*b*) 215.75 Mc; (*c*) 257 Mc; (*d*) 302.75 Mc.
3. A pentode mixer has a conversion gain of 10. With 5 mv r-f input, the i-f signal output equals: (*a*) 5 mv; (*b*) 10 mv; (*c*) 50 mv; (*d*) 2 to 3 volts.
4. An ultraudion local oscillator circuit tuned to 160 Mc has an inductance of 1 μh. If this is increased to 4 μh the new resonant frequency will be: (*a*) 80 Mc; (*b*) 160 Mc; (*c*) 320 Mc; (*d*) 400 Mc.
5. The oscillator frequency in intercarrier-sound receivers is adjusted for: (*a*) maximum sound; (*b*) minimum sound; (*c*) sound in the picture; (*d*) best picture.
6. Which of the following will provide the best tuned circuit for the UHF band? (*a*) Series resonance with low Q; (*b*) parallel resonance with low L/C ratio; (*c*) resonant cavity shorted at one end; (*d*) quarter-wave line open at both ends.
7. A matching pad for signal generator input is necessary when aligning the r-f amplifier because this stage: (*a*) has weak signal; (*b*) is the first amplifier; (*c*) may need neutralization; (*d*) drives the mixer stage.
8. In Fig. 20·21 the oscilloscope connection for the r-f response curve is at: (*a*) J1; (*b*) TP1; (*c*) TP2; (*d*) pin 2 of V_1.
9. Which of the following replacements may require oscillator realignment? (*a*) R-f amplifier; (*b*) oscillator; (*c*) first i-f stage; (*d*) antenna and transmission line.

10. Which of the following stages needs neutralization? (*a*) Triode oscillator; (*b*) triode r-f amplifier; (*c*) cascode r-f amplifier; (*d*) pentode r-f amplifier.

ESSAY QUESTIONS

1. Give the function of each of the three stages in a VHF tuner. What is the function of the station selector? The fine tuning control?
2. Give the function of the four stages used for UHF operation in a combination VHF-UHF tuner. How are the station selector and fine tuning controls used on UHF operation?
3. Draw the desired r-f response curve for channel 10, marking the picture and sound carrier frequencies.
4. State three requirements of the r-f amplifier stage.
5. What is the advantage of a cascode r-f amplifier, compared with a pentode stage? Compare a triode stage with a pentode and with a cascode circuit.
6. List the operating stages and their resonant frequencies in a combination VHF-UHF tuner for channel 13 and for channel 14.
7. What is one possible effect of local oscillator radiation interfering in a nearby receiver?
8. Draw a block diagram showing equipment and connections for obtaining a visual response curve of the r-f section for channel 10. Give two precautions important for r-f alignment.
9. Explain how the local oscillator in a turret tuner can be aligned for channel 5 by: (*a*) using the transmitted signal from a broadcast station; (*b*) using a signal generator. To what frequency will the generator be set?
10. How can the local oscillator be checked to see if it is operating? What is the effect on picture, sound, and raster if the local oscillator does not operate?
11. Why are resonant cavities used as tuned circuits at ultra-high frequencies?
12. Compare the operation of quarter-wave and half-wave resonant line sections, both shorted at the end. (Hint: See Sec. 21 · 13).
13. Describe the effect on picture and sound when the oscillator frequency is varied to cause: (*a*) sound carrier signal too high on the i-f response curve, with picture carrier too low on opposite side of curve; (*b*) picture carrier signal at 100 per cent response on the i-f response curve with practically zero response for the sound i-f carrier signal.
14. Referring to the combination VHF-UHF tuner in Fig. 20 · 17, give the resonant frequencies for all operating stages to tune in channel 4. Do the same for channel 31.
15. Referring to the UHF converter in Fig. 20 · 19, give the resonant frequencies for all stages to tune in channel 83.
16. Describe briefly three methods of providing for reception of the UHF channels.
17. Give two differences between supersonic waves at 40 kc and r-f signal at the same frequency.
18. Give the function of the following in a wireless remote-control system: transmitter unit, microphone, 40-kc amplifier, channel-control relay.
19. List three types of r-f amplifier circuits for the r-f tuner, with one feature of each.
20. Why is minimum tube noise more important in the r-f amplifier than in the last i-f stage? Why is overload distortion a bigger problem in the last i-f amplifier than in the r-f amplifier?
21. What is the purpose of neutralization in the r-f amplifier?
22. For each of the following, state the probable trouble and explain why: (*a*) No picture and no sound on channel 11 only. Other channels normal. (*b*) No picture and no sound on all channels. No snow in picture. (*c*) No picture and no sound on all channels. Picture is very snowy when contrast control is turned up. (*d*) Picture weak on some channels but becomes perfect when station-selector switch is moved slightly.
23. Referring to the schematic diagram of the r-f tuner in Fig. 20 · 21, give the function of T_1, L_5, L_6, L_7, L_8, L_{11}, C_8, C_{12}, C_{21}, C_{20}, R_{92}, C_{91}, C_{15}, R_{10}, L_{14}, and C_{26}.
24. Give the effect on picture and sound for the following component troubles in Fig. 20 · 21: (*a*) Open heater in 6CW4 r-f amplifier. (*b*) C_{91} shorted. What effect will this have on R_{92}?

(*c*) R_7 open in plate circuit of oscillator. (*d*) R_8 open in plate circuit of mixer. (*e*) L_5 coil for channel 10 is open. (*f*) L_8 coil for channel 10 is open.

25. In Fig. 20·23*b*, which three components correspond to the center conductors of the three resonant cavities?
26. Give one advantage and one disadvantage of a crystal diode mixer.
27. Give three factors that determine the amount of snow in the picture.
28. Which stage in a superheterodyne receiver produces the largest amount of tube noise? Which stage has the smallest amount of signal?
29. What is meant by programming a tuner, in remote operation?
30. Could remote operation be used to receive one UHF channel in addition to the VHF channels? Explain.
31. Describe briefly two mechanical troubles in tuners.

PROBLEMS *(Answers to odd-numbered problems at back of book.)*

1. Calculate the lowest frequency and highest frequency for a VHF oscillator beating 45.75 Mc above the r-f picture carrier for channels 2 to 13.
2. Calculate the ratio of highest to lowest frequency for a UHF oscillator beating 45.75 Mc above the r-f picture carrier for channels 14 to 83. Compare this ratio with a local oscillator for the AM radio band of 535 to 1,605 kc, beating 455 kc above the r-f signal frequencies. Which band is easier to cover by continuous tuning?
3. Calculate the length of a quarter-wave stub at 100 Mc, using twin lead with a velocity factor of 0.85. (Hint: See Sec. 21·13.)
4. What is the resonant frequency of each stub across the antenna input terminals in Fig. 20·18?
5. Calculate the resonant frequency of the wave trap with L_{14} and C_{26} in Fig. 20·21.
6. What oscillator frequency is needed to tune in channel 3? Calculate the oscillator inductance L required for this frequency with a capacitance of 5 $\mu\mu$f.
7. In a receiver with 25.75 Mc i-f picture carrier and oscillator beating above the r-f signal, what frequency is the image of the picture carrier for channel 2?
8. With an i-f picture carrier of 45.75 Mc, do the same as in Prob. 7 for channels 2, 6, 7, 13, and 14. Which i-f value is better for minimum interference from image frequencies?
9. Calculate the wavelength of a supersonic signal at 40 kc.
10. What is the frequency of a radio signal with the same wavelength as the supersonic signal at 40 kc in Prob. 9?
11. Calculate the thermal noise voltage generated by a resistive input impedance of 3,000 ohms, with a bandwidth of 60 kc.
12. Make a table listing thermal noise voltages generated by 300 ohms resistive impedance for bandwidth values of 6 cps, 6 kc, 6 Mc, 12 Mc, and 60 Mc. Also, plot a graph showing how thermal noise voltage increases with bandwidth.

Chapter 21 Antennas and transmission lines

A receiving antenna or aerial can be simply a length of wire or any conductor which has current induced in it by the transmitted electromagnetic wave. To have sufficient signal, however, either the antenna must be very long or it should have a resonant length that magnifies the effect of particular frequencies. For the VHF and UHF bands, where one-half wavelength is a practical size, the most common type of antenna is the half-wave resonant dipole illustrated in Fig. 21 · 1. In this antenna system for a television receiver, the dipole intercepts the radiated electromagnetic wave to provide induced signal current in the antenna conductors. The transmission line conducts the signal current to the antenna input terminals of the receiver, with minimum losses and no pickup by the line itself. The antenna signal includes both the picture and sound carrier signals, which are received by the same antenna.

The receiver needs enough antenna signal to reproduce a good picture. It is the picture carrier signal from the antenna that is amplified and detected to recover the composite video signal for the kinescope and for the sync circuits. A strong antenna signal enables the picture to have good contrast without snow, and no ghosts from multiple signals. Also the picture holds steady with good sync. The required amount of antenna signal is about 200 to 2,000 μv.

21 · 1 Resonant length of an antenna

The frequency, wavelength, and velocity of propagation of the radiated electromagnetic wave are related to each other by the equation

$$\lambda = \frac{\text{velocity}}{\text{frequency}} = \frac{3 \times 10^{10} \text{ cm per sec}}{f} = \frac{186{,}000 \text{ miles per sec}}{f} \qquad (21 \cdot 1)$$

where lambda (λ) is wavelength, f is frequency, and the velocity in

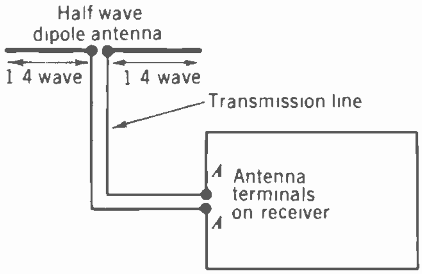


Fig. 21 · 1 Antenna and transmission line connected to receiver.

free space is equal to the speed of light. When the physical length of the antenna is cut to equal either one-half or one-quarter the wavelength corresponding to the signal frequency, the antenna is resonant at that frequency to provide a resonant rise in antenna current. Multiples of these lengths are considered harmonic resonances.

The two basic types of resonant antennas are the grounded quarter-wave Marconi antenna used at lower frequencies and the half-wave Hertz antenna in Fig. 21 · 1. The half-wave antenna usually consists of two quarter-wave elements insulated from each other, which add to provide a half wavelength. This is called a *dipole*. The dipole antenna operates independently of ground and therefore may be installed far above the earth or other absorbing bodies. Because of this, and because the physical length of a half wave is a practical size at high frequencies, the dipole is the basic antenna in television. More elaborate antenna arrays are usually just combinations of dipoles.

On the basis of Eq. (21 · 1), the formula for the length of a half wave in feet is derived as $L = 492/f$, where f is in megacycles and L gives the half wavelength in feet, in terms of the electromagnetic field traveling with the speed of light. However, the resonant length of a half-wave conductor is slightly less than a half wave in free space because the antenna has capacitance that alters the current distribution at the ends of the antenna. This end effect requires foreshortening of the conductor length, to provide the resonant current distribution that corresponds to the length of a half wave in free space. Wider antenna conductors and higher frequencies require more foreshortening, but it can be taken as approximately 6 per cent for the antennas used in television. Therefore, the length of the half-wave dipole is computed from the formula

$$L\,(\text{ft}) = \frac{462}{f\,(\text{Mc})} \qquad (21 \cdot 2)$$

L gives the length of the half-wave dipole directly in feet. This is the actual physical length of the half-wave antenna corresponding to an electrical half wave. As shown in Fig. 21 · 1, one-half this value is used for the length of each of the two quarter-wave poles. The small insulation distance between the two poles can be considered negligible. For a dipole tuned to 60 Mc, as an example, the length of the half wave is 7.7 ft, and each section is made 3.85 ft long.

21 · 2 Definition of antenna terms

Wave polarization. The moving electromagnetic field, which is the radio signal, consists of two components: a magnetic field associated with the current in the transmitting antenna and the electric field associated with its potential. The two fields are perpendicular to each other in space, and both are perpendicular to the direction of travel of the wave. When the electromagnetic wave passes the receiving antenna, it induces antenna current with the same variations as the transmitted radio signal.

The polarization is arbitrarily defined as the direction of the electric field. This is determined by the physical position of the antenna in space. A horizontal dipole is horizontally polarized. Then the magnetic lines of force are in the vertical plane around the conductors and the electric field is horizontal between the conductors. An antenna vertical with respect to earth is vertically polarized. Horizontal polarization is specified by the FCC for transmission in the television and FM broadcast bands. Therefore, the receiving antenna is mounted horizontally for maximum pickup of a horizontally polarized wave. Horizontal polarization is chosen because experimental results show more signal strength and less reflection for frequencies in the VHF and UHF spectrum. Also, the horizontal directivity of the receiver dipole helps reduce ghosts.

Microvolts per meter. This unit is a measure of field strength of the electromagnetic wave. A meter is slightly less than 40 in. As an example, when a resonant half-wave dipole 40 in. long provides 300 μv signal to the transmission line, the field strength is 300 μv per meter. The antenna height is standardized, often at 30 ft, and the antenna polarization is the same as the wave polarization. The values of field strength required for VHF service are 500 μv per meter for rural or residential areas and 5,000 μv per meter for city areas.

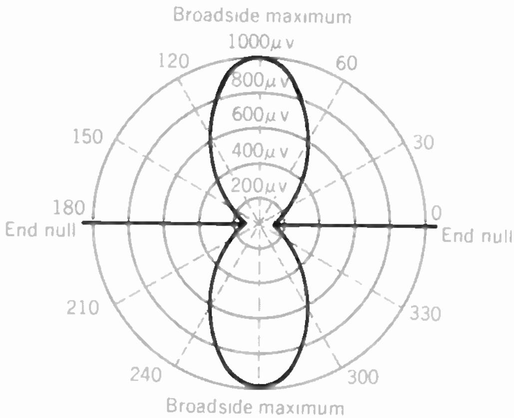


Fig. 21 · 2 Polar directivity pattern in horizontal plane for half-wave dipole antenna.

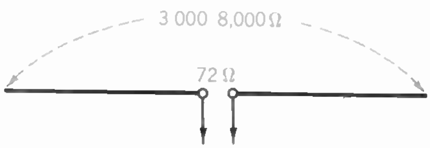


Fig. 21 · 3 Impedance of half-wave dipole antenna.

Note that the same field strength produces more antenna signal in a longer antenna, which is resonant for lower frequencies. As an example, assume a field strength of 800 μv per meter with a half-wave dipole 4.62 ft long for resonance at 100 Mc. The antenna signal then is 1,120 μv, because the length is 1.4 times more than a meter. A half-wave dipole 2.31 ft long for resonance at 200 Mc in the same field provides one-half the antenna signal, equal to 560 μv, since this antenna has one-half the length.

The field strength of the r-f carrier signal at the receiver antenna depends on radiated power and propagation characteristics for the carrier frequency. Even though several stations may transmit from one location, the field strength at the receiver is generally not the same for different channels. In addition, the characteristics of the receiver antenna vary for different frequencies.

Polar directivity patterns. As shown in Fig. 21 · 2, signal strength is plotted in polar coordinates to show magnitude and direction. The angle gives the direction for which the signal strength is plotted while the length of the radial arm defines the magnitude. The polar diagram shows the horizontal directivity pattern of the antenna, as determined by the current distribution of the antenna conductor. For a transmitting antenna, the pattern shows in which direction the antenna radiates the most signal; for a receiving antenna the pattern shows the direction from which most signal is intercepted by the antenna conductor. Rotating the antenna for the direction of best signal pickup is called *orientation.*

A half-wave dipole at its fundamental resonant frequency has the *figure-eight* directivity pattern in Fig. 21 · 2. The antenna receives best from the front and back, broadside to the antenna conductor, with little signal received from directions off the ends. As examples, in Fig. 21 · 2 the antenna provides 1,000 μv for signal in the broadside direction, about 500 μv signal at 60°, and a null of practically zero signal at 0° and 180°. This pattern applies only for half-wave resonance. For frequencies off resonance, the directional pattern changes because of a different current distribution on the antenna.

Antenna impedance. This impedance Z_a varies with the values of current at different points along the antenna. For a resonant half-wave dipole, Z_a is approximately 72 ohms at the center, as shown in Fig. 21 · 3. At the ends, Z_a is several thousand ohms. Intermediate points have intermediate values. Furthermore, the value of Z_a at the center is higher than 72 ohms for frequencies off resonance. These values are a-c impedances corresponding to an E/I ratio, which cannot be measured with an ohmmeter.

Antenna bandwidth. The half-wave antenna is equivalent to a resonant circuit with resistance and reactance. Therefore, the antenna can be considered as having a value of Q, which determines its bandwidth. Larger diameter for the antenna conductors decreases the reactance, allowing lower Q and wider frequency response. For this reason, metal tubing of ¼ to ½ in. diameter is generally used for the receiving antenna.

Antenna gain. This is a term used to express the increase in signal for one antenna over a standard antenna, usually a half-wave dipole having the same polarization. Antenna gain is generally used in connection with directional antenna systems with multiple elements and is measured in the optimum direction. The gain is usually stated in decibels. An antenna with a gain of 3 db, as an example, has a power gain of 2 or voltage gain of 1.4. Typical curves of antenna gain are shown in Fig. 21 · 17*b*.

Front-to-back ratio. This indicates the amount of signal the antenna receives from the front compared with signal received from the back. As an example, if the antenna intercepts 1,000 μv of signal from a transmitter in front, but only 500 μv for signal of the same frequency arriving from the back, the front-to-back ratio is 2 times in voltage, or 6 db.

Summary of desirable antenna characteristics. Multiple dipole elements can provide many types of combinations. In general, more dipoles intercept more signal, but the antenna characteristics determine how much useful signal is delivered by the transmission line to the receiver. The antenna should have high gain to provide enough signal for a good signal-to-noise ratio, especially in fringe areas. However, high gain is generally associated with narrow bandwidth, which may limit reception to one channel. An antenna impedance of 72 to 300 ohms is desirable for matching to the transmission line. Very important is a good front-to-back ratio to prevent ghosts in the picture caused by receiving the same signal from two different directions. In addition, a narrow forward lobe for the antenna directivity pattern helps reduce ghosts. Such an antenna with a sharp directional pattern generally has high gain and narrow bandwidth.

21 · 3 Ghosts

A duplicate image of the reproduced picture, a little to the side of the original as shown in Fig. 21 · 4, is called a *ghost*. Such ghosts are commonly

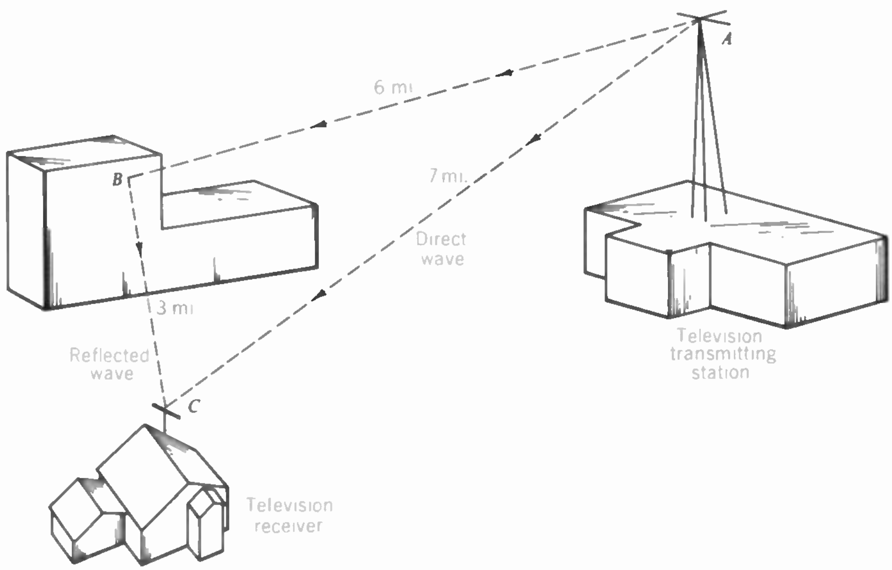
Fig. 21 · 4 Ghost caused by reflected signal.

caused by multipath reception of reflected signals. Referring to Fig. 21·5, the antenna at point *C* can receive picture signal by two separate paths from the transmitter at point *A*. The path *ABC* for the signal reflected from the building at point *B* is longer than the direct path *AC* by 2 miles, in this case. Since the velocity of the radiated signal is 186,000 miles per sec and the reflected wave path is 2 miles longer than the direct path, the reflected wave is delayed by 2/186,000 sec in reaching the antenna. This is approximately equal to 11 μsec.

The electron beam scanning the screen of the picture tube requires about 55 μsec to scan across one horizontal line. On a picture having a width of 20 in., then, it requires only 5.5 μsec to scan 2 in. The reflected signal, delayed in reception by 11 μsec, will produce a second image displaced from the original by 4 in. The ghost is displaced to the right, in the direction of scanning, because the reflected signal arrives at the receiving antenna later in time than the direct signal. Reflection can be produced by any conducting material, as induced current in the conductor reradiates the signal.

With multiple reflections there may be multiple ghosts. The intensity of the ghost may be nearly as strong as the original image or just noticeable. Any difference in relative intensity is the result of more attenuation of the reflected wave in its longer travel. The ghost may be a positive or negative image, depending on the relative phase between the multipath r-f signals. Reflection by a conductor generally shifts the phase and polarization of the reradiated signal.

Fig. 21·5 Reception of multipath signals. The reflected-wave distance ABC is 2 miles longer than the direct path AC.



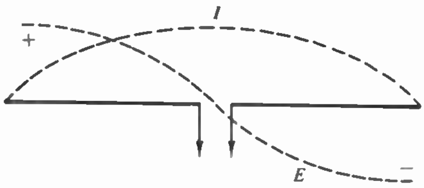


Fig. 21·6 Current and voltage distribution on half-wave antenna.

The interference in the picture may vary from definite ghost images to reflections that are not noticeable as duplicate images because of insufficient time delay but which cause the picture to appear fuzzy. A delay distance of about 50 ft or less can be considered negligible, since the resultant horizontal displacement on the screen of the picture tube is then considerably less than the width of a picture element.

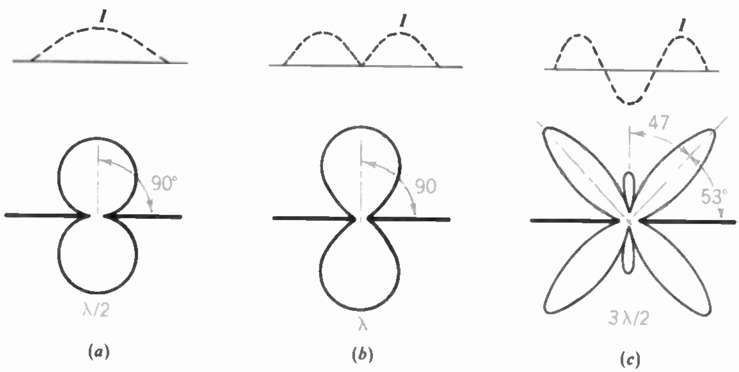
For the problem of multipath reception, an antenna that has a good front-to-back ratio and a narrow forward lobe with minimum side responses can be rotated horizontally to minimize ghosts. Sometimes changing the antenna location reduces the intensity of the ghost. Once the multipath signals are coupled from the antenna to the receiver, it is impossible to eliminate the ghost. The only remedy is in the antenna system.

21·4 Straight dipole

The antenna illustrated in Figs. 21·1 to 21·3 is a Hertz, doublet, half-wave, or simply straight dipole. Its distribution of current and voltage for half-wave resonance is illustrated in Fig. 21·6. With r-f excitation by the radiated electromagnetic field, current is induced in the antenna with the same variations as the applied voltage. The dotted lines for *E* and *I* indicate an envelope of peak values for the varying a-c signal at different points along the antenna.

The electron flow in the antenna conductor is not instantaneous but

Fig. 21·7 Directional characteristics of center-fed dipole antennas at fundamental and harmonic frequencies. (a) Figure-eight pattern at half-wave resonance. (b) Broadened response for full-wave operation. (c) Lobes split because of harmonic operation at three times the resonant frequency.



travels along the wire in free space with approximately the speed of light. When the electrons reach the end of the wire, the resulting accumulation of charge at the end provides a potential for moving the charge in the opposite direction, reversing the direction of current flow. The resultant current is zero at the ends, with two currents of equal amplitude flowing in opposite directions. Farther back on the wire, the outgoing and returning currents are not the same, because the charges causing the currents have been supplied to the antenna at different parts of the r-f cycle. Maximum current is at the center, where the reflected current from the ends adds to the original current. The ends of the antenna in free space are points of maximum voltage and zero current. Because of capacitance of the ends, however, the current is normally not zero at the ends but has a definite value. Therefore, the antenna must be foreshortened to give the same current distribution that would be obtained for a half wave in free space.

The voltage and current distribution illustrate why the physical length of antenna makes it resonant for a particular frequency. When the electron charges in the antenna conductor can travel from the center out to the ends and back to the center in the time of one-half cycle, the current and voltage values are maximum. At other frequencies, partial cancellation of the incident and returning electrons reduces the amount of antenna signal.

Response off resonance. When the television receiving antenna is used to receive several channels, the directional characteristics and impedance of the dipole at frequencies off half-wave resonance vary as the antenna current and voltage distribution change. Referring to Fig. 21 · 7*a*, the dipole has the figure-eight directivity pattern at its fundamental resonant frequency, with a resistive impedance of 72 ohms at the center, when the antenna length is one-half wave. Maintaining the same physical length, at double this frequency the dipole is a full-wave antenna. It still has the figure-eight directivity pattern, but the impedance at the center, where the transmission line connects, is now a maximum equal to several thousand ohms. At three times the fundamental resonant frequency, the same physical antenna length is three half waves. The center is again a point of minimum antenna resistance, equal to about 100 ohms. However, notice that the directional pattern splits into four major lobes. There is a gain of 1 db in the direction of maximum reception but this is 53° off the ends, with little pickup from the broadside direction. At 4λ the response of the dipole from the broadside direction is practically zero. In order to utilize the broadside response, a center-fed straight dipole is limited to operation over a frequency range of about 2 to 1 or less.

V dipoles. In order to maintain broadside response over a wider frequency range, the dipole can be angled in the form of a V. For instance, assume the two poles in Fig. 21 · 7*c* are moved in 53° to line up with the two forward lobes shown. Then both top lobes would overlap in the forward direction. The result, therefore, is to maintain the broadside response when the antenna operates at harmonics of its fundamental half-wave resonant frequency. The smaller the angle of the V in the dipole, the higher is the frequency at which the dipole maintains its forward response. Note that the V dipole also has a better front-to-back ratio, as the back lobes spread farther apart. However, "V-ing" the dipole for better response at high frequencies lowers its gain for low frequencies because then the antenna extends less than a half wavelength in the space across the ends of the poles.

21 · 5 Folded dipole

As shown in Fig. 21 · 8, the folded dipole is constructed of two half-wave rods joined at the ends, with one rod open at the center where it connects to the transmission line. The spacing between the rods is small compared with a half wavelength. This half-wave folded dipole has the same directional characteristics as the half-wave straight dipole, with the same amount of signal pickup. However, the antenna resistance of the folded dipole is approximately 300 ohms, which is a convenient value for matching to 300-ohm transmission line. The center of the closed section of the half-wave folded dipole is a point of minimum voltage, allowing direct mounting at this point to a grounded metal mast without shorting the signal voltage. Another difference from the straight dipole is that, with operation at twice the fundamental half-wave resonant frequency, the full-wave folded dipole does not have the figure-eight polar pattern, receiving little signal from the broadside direction.

The resistance of the half-wave folded dipole at the center where it connects to the transmission line is higher than the value for a straight dipole because only part of the total antenna current flows in the open section. As a result, the folded dipole antenna resistance equals 72 ohms multiplied by the square of the ratio of the total diameter of all conductor sections to the diameter of the open section. As an example, in Fig. 21 · 8 with the same diameter for the two conductor sections, the total diameter is twice the diameter of the open section. Therefore, the antenna impedance is 4×72, or 288, ohms, which is generally considered as approximately equal to 300 ohms. In applications where a higher resistance is desired for the folded dipole, the diameter of the closed conductor section is

Fig. 21 · 8 Folded half-wave dipole.

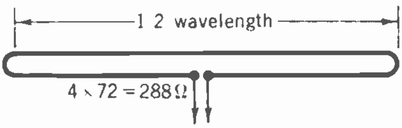
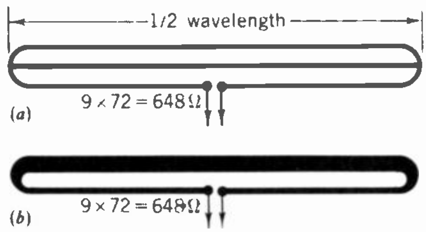


Fig. 21 · 9 High-impedance folded dipole antennas. (a) With additional closed conductor section. (b) With wider diameter closed conductor section.



increased. In Fig. 21 · 9*a*, an additional closed conductor of the same diameter is added, making the total conductor diameter triple the diameter of the open section, to provide an antenna resistance of 9 × 72, or 648, ohms. The same impedance transformation can be obtained by increasing the diameter of the closed conductor, as in *b*, instead of adding sections.

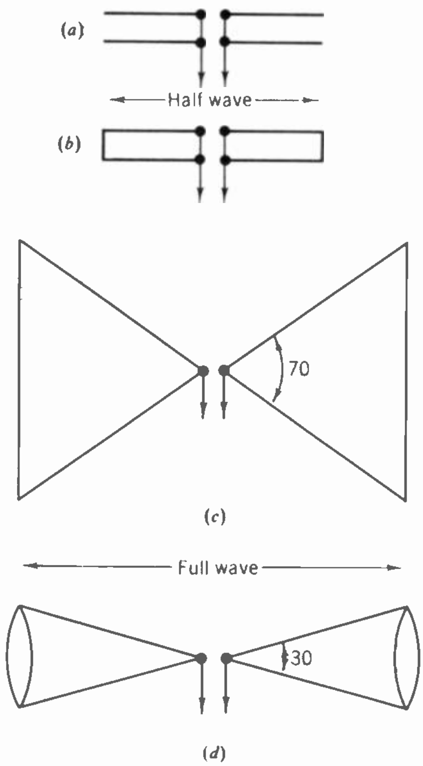


Fig. 21 · 10 Broad-band dipole antennas. (a) and (b) Half-wave dipoles. (c) Full-wave triangular dipole, or bowtie antenna. (d) Full-wave dicone antenna.

21 · 6 Broad-band dipoles

A *thick dipole* antenna, which has a cross-sectional dimension approximately 0.1λ or greater, can provide more uniform response over a wider band of frequencies, compared with a *thin dipole* conductor having negligible diameter. Figure 21 · 10 shows several types of thick broad-band dipoles. These can be constructed of wire conductors, metal sheets, wire screening, or metallic foil. The increased thickness has the following three effects on the dipole characteristics:

1. The antenna resistance decreases and the reactance is lowered even more, resulting in an antenna with lower Q that has more uniform impedance values over a wide band of frequencies.
2. The broadside response of the directional pattern is maintained over a wider frequency range, before splitting into multiple lobes.
3. More foreshortening is needed with increasing thickness to provide physical lengths equivalent to the electrical wavelengths.

Referring to Fig. 21 · 10, the antennas in *a* and *b* are operated as half-wave dipoles. When the conductors are separated by 0.1 or less they can be considered as one uniform antenna of wider cross section. The centers are joined in *a* and the ends can also be joined, as in *b*, since the end points of the two antennas are at the same potential. The triangular dipole in *c* and conical dipole in *d*, however, are operated as full-wave antennas because they have too low a resistance at the center at half-wave resonance.

As a full-wave dipole, the antenna has a gain of approximately 3 db. The included angles shown provide an antenna resistance of approximately 300 ohms at the center, for full-wave resonance. Smaller angles result in higher impedance. The overall physical length can be foreshortened about 10 per cent for the triangular dipole and 25 per cent for the dicone antenna. The triangular dipole, often called a *bowtie antenna,* is commonly used as a broad-band receiving antenna to cover all the UHF channels. The non-uniform cross section, with wider diameter at the ends, improves the broad-band characteristics.

21 · 7 Long-wire antennas

Compared with a half-wave dipole, a long-wire antenna, which is several wavelengths, has the advantages of increased signal pickup and sharper directivity. As the antenna wire is made longer in terms of the number of half waves, the directional pattern changes because of the current distribution, increasing the directivity along the line of the antenna wire itself.

V antenna. When two long wires are combined in the form of the horizontal V antenna shown in Fig. 21 · 11, the major lobes of the directional pattern for each wire reinforce along the line bisecting the angle between the two wires. Therefore, the V antenna receives best along the line of the bisector. The greater the electrical length of the conductor legs, the smaller the included angle of the V should be for maximum antenna gain.

Rhombic antenna. A more efficient arrangement is the rhombic antenna shown in Fig. 21 · 12 which consists of two horizontal V sections. To make it unidirectional, the rhombic antenna can be terminated with a resistor of 470 ohms, for an approximate match to 300-ohm transmission line. Both the V and rhombic antennas are mounted horizontally for horizontal polarization as television antennas. Each leg should be at least two wavelengths at the lowest operating frequency, the gain and directivity of the antenna increasing with the length. The angle of 50° is a compromise value suitable for leg lengths of 2 to 6λ. Longer legs should have a smaller angle. With each leg four wavelengths, the V antenna has a gain of 7 db, while the rhombic antenna gain is 10 db, approximately. At ultra-high frequencies these lengths are practicable, to provide a high-gain antenna for the UHF television band. The rhombic can provide a uniform value

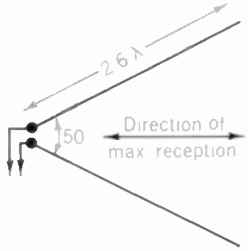


Fig. 21 · 11 V antenna.

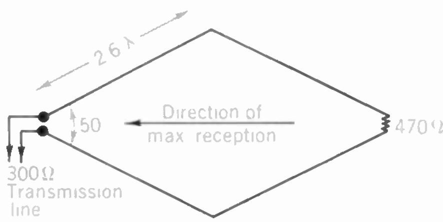


Fig. 21 · 12 Rhombic antenna.

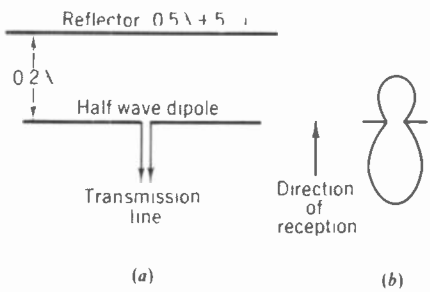


Fig. 21 · 13 Dipole with reflector in back.

of antenna impedance over a total frequency range of 3 to 1, with high gain and sharp directivity.

21 · 8 Parasitic arrays

When current flows in the receiving antenna it radiates part of the intercepted signal, as in a transmitting antenna. If a conductor approximately one-half wave long is placed parallel to the half-wave dipole antenna but not connected to it, as illustrated in Fig. 21 · 13, the free wire will intercept some of the signal radiated by the antenna. This signal is reradiated by the free wire to combine with the original antenna current. As a result, part of the intercepted signal lost by radiation in the receiving antenna is recovered by using the free wire, providing increased gain and directivity. The free wire is called a *parasitic* element because it is not connected to the dipole antenna. A parasitic element placed behind the antenna is a *reflector;* a parasitic element in front of the antenna is a *director.* The dipole antenna itself, to which the transmission line is connected, is the *driven element.* This can be either a straight dipole or a folded dipole.

A dipole antenna with one or more parasitic elements is a *parasitic array.* This is the most common type of television receiving antenna because it is simple to construct, can be oriented easily, provides enough gain for average signal strengths, and increases the directivity compared with a dipole alone. The main directional effect of the parasitic element is to reduce the strength of signals received from the rear of the antenna, making its response unidirectional. Therefore, an antenna with a parasitic element is useful for reducing the strength of multipath reflected signals arriving from directions behind the antenna, to eliminate ghosts in the picture.

Dipole and reflector. Referring to Fig. 21 · 13, the reflector is usually placed approximately 0.2λ, or a little less than a quarter wave, behind the dipole, to reinforce signals arriving from the front. Also, the reflector is about 5 per cent longer than the dipole, or a little more than a half wave. The antenna response depends on the spacing between the elements, and the tuning of the parasitic, which can be adjusted by changing its length. Closer spacing lowers the antenna impedance and narrows the frequency response. For the typical arrangement shown in Fig. 21 · 13 the dipole and reflector have a voltage gain of approximately 5 db. The antenna impedance is about one-half the impedance of the antenna itself, resulting in

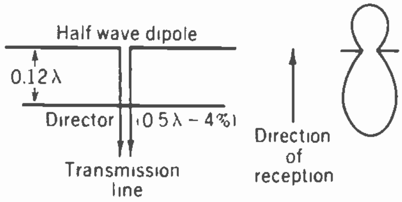


Fig. 21 · 14 Dipole with director in front.

150 ohms for the folded dipole and reflector and 36 ohms for the straight dipole and reflector. The front-to-back ratio is approximately 3 db, resulting in about 1.4 times more signal voltage from the front than from the back.

It is important to note that the parasitic element is effective only within the frequency range for which it is approximately tuned to half-wave resonance. Furthermore, the increase in gain and directivity with a reflector cuts off sharply at frequencies lower than the resonant frequency. For this reason, a dipole with reflector is usually cut for the lowest frequency in the range to be covered. The reflector can then be effective up to a frequency about 30 per cent higher than the resonant frequency.

The response of the dipole with reflector can be explained as follows. Signal from the front produces current in the reflector 90° later than in the dipole. The 90° lag results because of the reflector spacing and its extra length. Current in the reflector reradiates signal to the dipole, which arrives an additional 90° later. Meanwhile, the signal at the dipole has varied through 180° of its cycle. The reradiated signal from the reflector, which has been delayed 180° also, then combines with the antenna current in the dipole to provide more signal for the transmission line. This increase in signal provides a forward gain up to 6 db, or double the signal voltage of a dipole alone. For signal arriving from the back, however, the reradiated signal from the reflector is just 90° out of phase with antenna current on the dipole. The combined signal from the back, therefore, is less than the combined signal from the front.

Dipole with director. Referring to the parasitic director element in Fig. 21 · 14, it is placed 0.12λ in front of the dipole and is about 4 per cent shorter than the driven element. The gain and directivity of the dipole with director drop off sharply at frequencies higher than the resonant frequency, which is opposite to the operation of the dipole and reflector off resonance. Practically all television antennas with one parasitic element use a reflector, instead of a director, because the director needs closer spacing for the same gain and front-to-back ratio, which reduces the antenna resistance and narrows the frequency response. For an antenna operating over a relatively narrow frequency, however, directors are combined with a dipole and reflector. A dipole with one reflector and one director, having the same spacings as for a single parasitic element, pro-

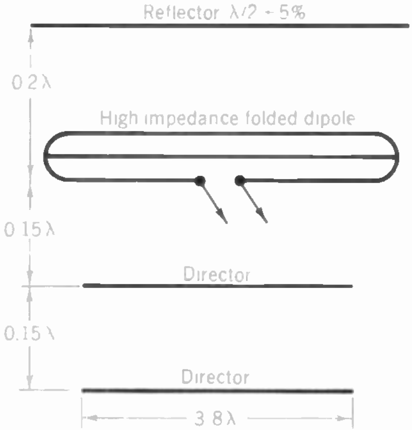


Fig. 21 · 15 Yagi antenna, with dipole, reflector, and two directors.

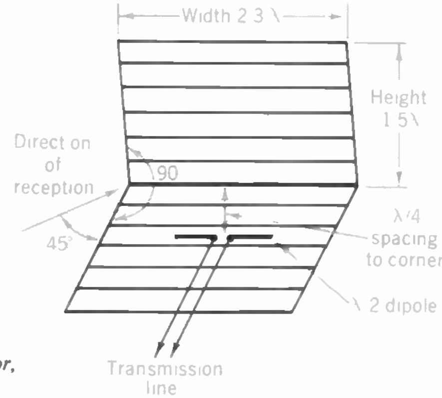


Fig. 21 · 16 Dipole with corner reflector, for UHF band.

vides about 7 db gain, while the antenna resistance is approximately one-eighth the value of the driven element by itself.

Yagi antenna. A dipole with one reflector and two or more directors, as illustrated in Fig. 21 · 15, is called a *Yagi antenna.* This is a compact high-gain array, with a sharp forward broadside lobe and narrow bandwidth, often used in low-signal areas to cover one television channel or several adjacent channels. The gain of the Yagi antenna with three parasitic elements is about 10 db, with a front-to-back ratio of approximately 15 db. A high-impedance folded dipole is generally used for the driven element so that the reduced value of antenna resistance with the parasitic elements can be about 150 to 300 ohms. More directors can be added but generally only one reflector is used because additional reflectors do not improve the response.

Dipole with Corner Reflector. The antenna in Fig. 21 · 16 has a reflector constructed as a corner conducting sheet behind the half-wave dipole. The corner reflector can be either a solid metal sheet or a grid consisting of wires or wire screening, provided that the spacing between grid wires is 0.1λ or less. The dipole antenna, insulated from the parasitic reflector, is mounted along the line bisecting the 90° corner. Maximum signal is received from the front along this line. The construction of this antenna is practical for the UHF band because of the small size at these frequencies.

21 · 9 *Multiband antennas*

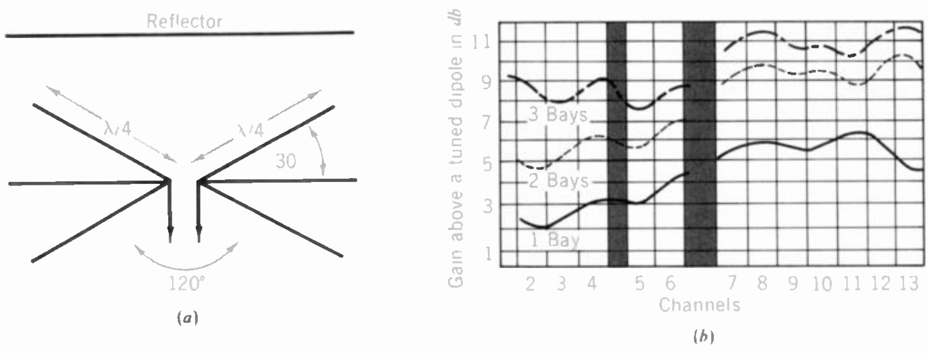
All the television broadcast channels can be considered in three bands: 54 to 88 Mc for the low-band VHF channels, 174 to 216 Mc for the high-band VHF channels, and 470 to 890 Mc for the UHF channels. The main problem in using one dipole for both VHF bands is maintaining the broadside response, as the directional pattern of a low-band dipole splits into side lobes at the third and fourth harmonic frequencies in the 174- to 216-Mc band. A high-band dipole cut for a half wavelength in the 174- to 216-Mc band is not suitable for the 54- to 88-Mc band because of insufficient signal pickup at the lower frequencies. As a result, antennas for both VHF bands generally use either separate dipoles for each band or a dipole for the 54- to 88-Mc band modified to provide broadside unidirectional response in the 174- to 216-Mc band also. For the UHF band, a VHF antenna can operate as a long-wire antenna or be used as a reflector sheet behind a broad-band dipole added in front. It should be noted that higher gain is necessary for a UHF antenna to provide the same signal strength as on the VHF channels because the reference dipole is shorter at ultra-high frequencies.

Conical dipole. The VHF dual-band antenna illustrated in Fig. 21 · 17 is generally called a *conical* or *fan* dipole. This antenna consists of two half-wave dipoles inclined about 30° from the horizontal plane, similar to a section of a cone, and usually a horizontal dipole in the middle. All the dipoles are tilted inward toward the wavefront of the arriving signal at an angle approximately 30° from the broadside direction, resulting in a total included angle of 120° between the two conical sections. Smaller values of included angle reduce the amount of signal intercepted at low frequencies, as the distance across the front is decreased. Either a straight reflector or conical reflector can be used behind the conical dipole with approximately the same results.

Cut for a half wavelength at channel 2, the conical dipole with reflector is commonly used as a receiving antenna to cover both VHF bands. The antenna resistance is about 150 ohms. For the 54- to 88-Mc band, the antenna is a conical-type half-wave dipole with a parasitic reflector, providing the desired unidirectional broadside response. For the 174- to 216-Mc band, the tilting of the dipole rods shifts the direction of the split lobes produced at the third and fourth harmonic frequencies, so that they combine to produce a main forward lobe in the broadside direction.

In-line antenna. As shown in Fig. 21 · 18, this antenna consists of a half-

Fig. 21 · 17 Conical dipole with reflector. (a) Spacing of elements. (b) Response curves.



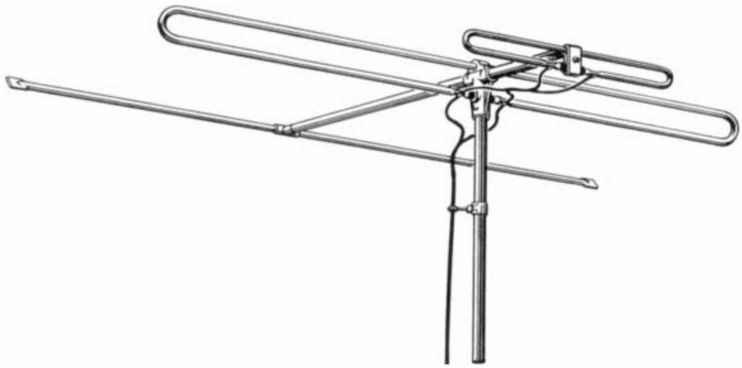


Fig. 21 · 18 In-line dual-band dipole antenna.

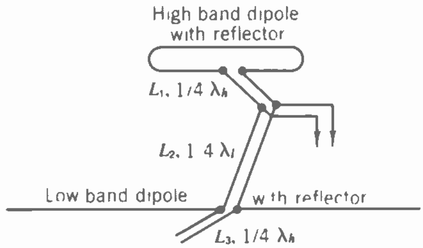


Fig. 21 · 19 High-low-band dipoles connected to common transmission line. Reflectors omitted for clarity. λ_h is high-band wavelength; λ_L is low-band wavelength.

wave folded dipole with reflector for the 54- to 88-Mc band, in line behind the shorter half-wave folded dipole for the 174- to 216-Mc band. The distance between the two folded dipoles is approximately one-quarter wavelength at the high-band dipole frequency. This is the length of line connecting the short dipole to the long dipole, where the transmission line to the receiver is connected. For the low-band channels, the long folded dipole with reflector supplies antenna signal to the transmission line, as the short dipole has little pickup at these low frequencies. For the high-band channels, the short folded dipole supplies signal to the transmission line, with the long folded dipole operating as a reflector. The directivity pattern of the in-line antenna is relatively uniform on all VHF channels, with a unidirectional broadside response. The average antenna gain on the low-band channels is approximately 2 db, and 5 db on the high-band channels. The antenna resistance is about 150 ohms.

High-low antenna. The VHF dual-band antenna in Fig. 21 · 19 uses a separate dipole and reflector for the 54- to 88-Mc band, mounted on the same mast as the dipole and reflector for the 174- to 216-Mc band. This allows separate orientation of each dipole, which can be an advantage when stations in the two bands are in different directions. A folded dipole or straight dipole can be used for either antenna or both antennas. The short dipole is usually mounted at the top of the mast for mechanical considerations, although it can be in a separate location if this provides more signal. The distance above the long dipole is not critical, but approximately the same half wavelength as the short dipole is a suitable spacing. Best results are obtained when separate transmission lines are connected to each antenna, with a double-pole double-throw switch at the receiver to select either one. A knife switch is preferable for minimum capacitance. How-

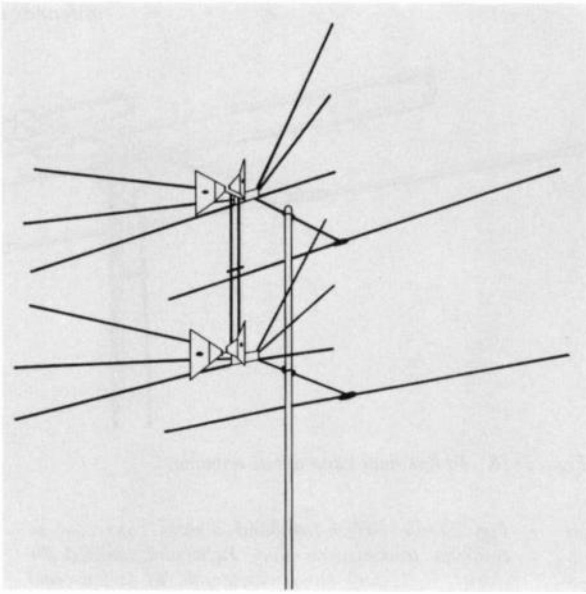


Fig. 21 · 20 VHF conical or fan antenna with UHF triangular dipole or bowtie antenna. (Channel Master Corporation.)

ever, the two dipoles can be connected to a common transmission line, as shown in Fig. 21 · 19.

VHF and UHF antenna. The antenna in Fig. 21 · 20 combines a VHF conical dipole with reflector and UHF triangular dipole, or bowtie antenna, to cover all the television channels. For the low-band channels 2 to 5, the antenna is a conical dipole with parasitic reflector, while for the high-band channels 7 to 13 it operates as a V-type antenna. On the 470- to 890-Mc band, the conical elements form a reflector for the bowtie antenna, covering the UHF channels 14 to 83 with an average antenna gain of 5 to 6 db.

21 · 10 Stacked arrays

In order to increase the antenna gain and directivity, two or more antennas of the same type can be mounted close to each other and connected to a common transmission line. The individual antennas are called *bays* and the combined unit is a *stacked array.* Mounting the antennas one above the other, as illustrated in Fig. 21 · 21, is vertical stacking. This can be considered a *broadside array,* as the antenna receives or transmits best in the direction broadside to the stacking.

In general, stacking can provide an additional gain up to 3 db for each antenna bay added, since the larger the area of the array the greater is the amount of signal that can be intercepted. To utilize the signal, however, the individual bays must be phased correctly with respect to the common junction for the transmission line, so that the antenna signals can add to provide the required antenna gain, with the desired directional response. Rods used for interconnecting the bays in the stacked array are called *phasing rods.* In phasing separate antennas to a common transmission line, the following points should be noted:

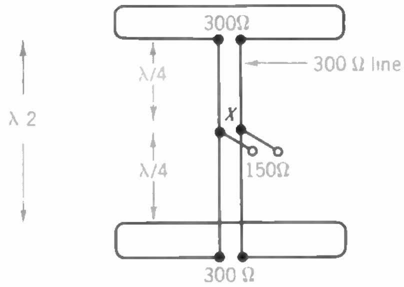


Fig. 21·21 Two vertically stacked antenna bays in broadside array. Reflectors omitted for clarity.

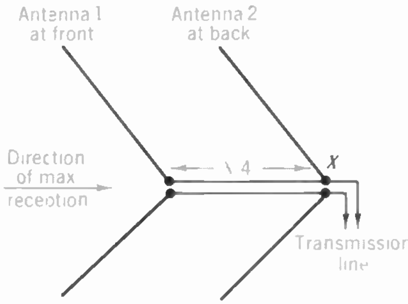


Fig. 21·22 Double-V antennas mounted one behind the other in horizontal end-fire array. No reflectors needed for unidirectional response.

1. A difference of one-quarter wavelength in the distance the antenna signal must travel is equivalent to a phase-angle difference of 90°.
2. A difference of one-half wavelength is equivalent to a phase-angle difference of 180°.
3. Reversing the connections is equivalent to a phase-angle difference of 180°. With twin lead, a half twist of the line reverses the connections.

Vertical stacking. This is often done in weak signal areas to increase gain and sharpen directivity in the vertical plane, which reduces pickup of external noise from sources usually below the antenna. Figure 21·21 illustrates vertical stacking and how the antenna bays are usually connected to the common transmission line. Folded dipoles are shown, without reflectors for simplicity, but the same principles apply for any other antenna. The half-wave spacing generally used between stacked antenna bays is convenient for interconnecting them with two quarter-wave sections. Because of the symmetry, the antenna signals from both bays travel the same distance to point X and are in phase at the common junction for the transmission line to the receiver. The impedance here is one-half the impedance of either section, as the two lines are in parallel. With more than two antennas in the array, they can be grouped in adjacent pairs, and the pairs are then interconnected to the common transmission line. The horizontal directivity pattern of the vertically stacked array has the same broadside response as the individual antennas, but with more gain.

Undirectional end-fire array. In Fig. 21·22, two antennas are stacked one behind the other a quarter wavelength apart. They are connected to the common transmission line at point X to form an array that has the unidirectional response shown, without any parasitic reflector or director. The arrangement is an end-fire array because it has maximum response through the line of the array. This double V is often used as a VHF antenna.

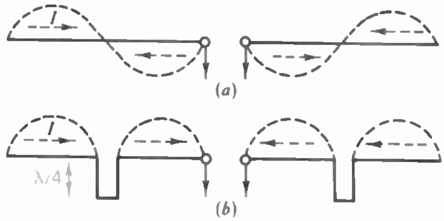


Fig. 21·23 Collinear arrangement. (a) Full-wave dipole with current distribution I. (b) λ/4 stub shifts I by 180°.

Each V is a dipole, cut for the low-band channels but angled in for broadside response on the high-band channels.

The end-fire directivity results from making the currents in the individual antennas out of phase with each other. With 90° phasing, the response is unidirectional off the end farthest from the transmission line. The antenna to which the transmission line connects is then back of the array (see Fig. 21·22). When receiving from the front, antenna 1 intercepts the signal a quarter cycle sooner than antenna 2. However, the quarter-wave line connecting the two antennas delivers this signal at point X in the same phase as the signal intercepted by antenna 2. The array provides a gain of 3 db, therefore, for signal arriving from the front.

Signal from the back is intercepted by antenna 1 a quarter cycle later than antenna 2. With the additional quarter wavelength of the connecting line, the signal delivered by antenna 1 arrives at point X 180° out of phase with the signal from antenna 2. The two signals cancel, therefore, resulting in minimum reception from the back.

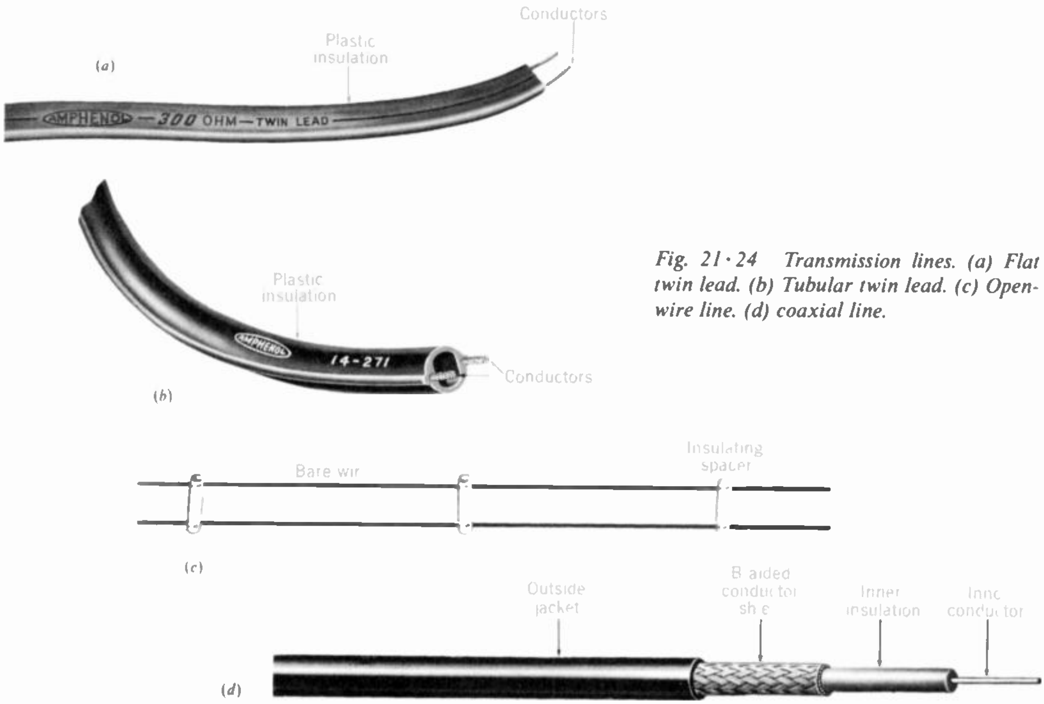


Fig. 21·24 Transmission lines. (a) Flat twin lead. (b) Tubular twin lead. (c) Open-wire line. (d) coaxial line.

Collinear array. Some antennas use the collinear principle illustrated in Fig. 21·23 to maintain broadside response at harmonic frequencies. Assume a dipole is operating at its fourth harmonic, which makes each pole one full wavelength. The current distribution in *a* then cancels for broadside response as each pole has half waves in opposite directions. This effect shifts the directional pattern toward the ends. However, the quarter-wave stub in *b* alters the current distribution. Then the current is in the same direction for both half waves on each pole. This distribution allows maximum antenna response in the broadside direction.

Note that the shorted $\lambda/4$ stub uses a half wave of conductor for a complete path. This accounts for the missing half wave that would be inverted on each pole. The stub does not contribute to the directivity pattern, since its conductors are close together, with opposite directions of the same current. The collinear arrangement, therefore, allows more than a half wave of conductor to be used to increase the antenna gain while maintaining the broadside response. The gain is a little less than 2 db per half wave in the array.

21·11 Transmission lines

The transmission line has the function of delivering the antenna signal to the receiver with minimum loss. The line itself should not pick up any stray signal. For this reason the line should be either balanced or shielded or both. A line is balanced when each of the two conductors has the same capacitance to ground. Usually, a balanced line is connected to opposite ends of a center-tapped antenna input transformer. Then the in-phase signal currents in the two conductors, due to stray pickup by the line, are canceled with a balanced receiver input circuit. A shielded line is completely enclosed with a metallic braid that is grounded to serve as a shield for the inner conductor.

The main types of transmission line are the concentric or *coaxial* line and the two-wire parallel conductor, as illustrated in Fig. 21·24. Parallel-wire line constructed in the form of a plastic ribbon, as shown in *a* and *b*, is generally called *twin lead*, either flat or tubular. The type in *c* is *open-wire* line. These are balanced lines but not shielded. The coaxial line in *d* is shielded but unbalanced. Shielded lines generally have more capacitance

Table 21·1 Transmission lines

Type	*Characteristic impedance Z_0, ohms*	*Attenuation, db per 100 ft, at 100 Mc*	*Capacitance per foot, uuf*	*Velocity factor V†*
Flat twin lead	300*	Dry, 1.2; wet, 7.3	6	0.8
Tubular twin lead	300	Dry, 1.1; wet, 2.5	5	0.8
Open-wire line	300–600	0.2	1	0.98
Coaxial line, RG-59U	72	3.7	21	0.65

* Flat twin lead is also available in 75- and 150-ohm characteristic impedance.

† V is ratio of velocity in dielectric to velocity in free space.

and higher losses, as shown in Table 21 · 1, which lists the characteristic impedance and attenuation for the common types of transmission lines. The attenuation is caused by I^2R losses in the a-c resistance of the line, reducing the amplitude of antenna signal delivered by the line to the receiver. The longer the line and the higher the frequency, the greater the attenuation. The characteristic impedance of the line, which results from the uniform spacing between the two conductors, is the same regardless of the length of line.

Flat twin lead. This is the most popular type of line because it has low losses, is available in 300 ohms characteristic impedance, costs less, and is flexible for ease of handling. The plastic ribbon is a low-loss dielectric, such as polyethylene, with the parallel conductors embedded in the plastic. The 300-ohm twin-lead spacing is approximately ⅜ in.; the 75- and 150-ohm twin lead have closer spacing. Since most television receivers have a balanced input impedance of 300 ohms, the 300-ohm twin lead is convenient for matching the transmission line to the receiver. For the same 300-ohm Z_0, there are many different qualities of twin lead. Stronger line is preferable for outside use as it lasts longer.

Because twin lead is unshielded it should not be run close to power lines, in order to avoid pickup of 60-cps hum voltage, and should be kept away from large metal structures, which can alter the balance of the line. Because of unbalance, there may be stray pickup of interfering noise voltages or signals from a nearby unshielded transmission line. The losses of flat twin lead are much greater when the line is wet, increasing from 1.2 to 7.3 db per 100 ft at 100 Mc. Taking a 6-db loss as an example, which is a voltage ratio of 1:2, this means that the signal-voltage output at the end of 100 ft of line is one-half the signal-voltage input; for 200 ft the signal-voltage output is one-fourth the input.

Tubular twin lead. In this type of line the two parallel conductors are embedded in the polyethylene plastic formed as hollow tubing, with air the dielectric for most of the inside area. The inside of the tubing is protected from rain, snow, salt, or dirt; as a result, deposits on the outside of the line do not lower the leakage resistance so much as in flat twin lead. Therefore, the tubular twin lead has much less attenuation when wet, as indicated in Table 21 · 1. The top end of the tubular line should be sealed.

Open-wire line. As shown in Fig. 21 · 24*c*, this line is constructed with low-loss insulating spacers between the bare-wire parallel conductors. The open-wire line has the least attenuation because the dielectric is air. However, its characteristic impedance is relatively high.

Coaxial line. As illustrated in Fig. 21 · 24*d*, this line consists of a center conductor in a solid dielectric completely enclosed by a metallic covering, which is often a flexible copper braid. An insulating jacket, usually of vinylite plastic, is molded over the entire line as a protective outside coating. This line is relatively immune to stray pickup because the outer conductor acts as a grounded shield. Therefore, coaxial line is used in noisy locations, or where stray pickup of interfering signals by the line is a problem, when unshielded lines would not be suitable. The coaxial line is unbalanced, with the inner conductor supplying antenna signal and the outer conductor connected to the receiver chassis. Compared with twin lead, coaxial line is stronger and unaffected by moisture but costs more and has higher losses. Coaxial line generally has a characteristic impedance of 50 to 75 ohms.

21 · 12 Characteristic impedance

When the transmission line has a length comparable with a wavelength of the signal frequency carried by the line, the line has important properties other than its resistance. The small amount of inductance of the conductor and the small capacitance between conductors are distributed over the entire length of the line. The result is a distributed inductance and capacitance which can make the line a reactive load, equivalent to lumped reactance in an ordinary circuit. Furthermore, the uniform spacing between conductors provides constant inductance and capacitance per unit length. Therefore, the characteristic impedance can be defined in electrical terms as

$$Z_0 = \sqrt{\frac{L}{C}} \qquad (21 \cdot 3)$$

Z_0 is in ohms, L the inductance per unit length in henrys, and C the shunt capacitance per unit length in farads. As an example, assume a line with 0.54 μh inductance for both conductors per foot and 6 $\mu\mu$f capacitance. Then

$$Z_0 = \sqrt{\frac{0.54 \times 10^{-6}}{6 \times 10^{-12}}} = \sqrt{0.09 \times 10^6} = 0.3 \times 10^3$$

$$Z_0 = 300 \text{ ohms}$$

The closer the conductor spacing, the greater the capacitance and the smaller the Z_0 of the line.

Physical characteristics. In terms of physical construction, Z_0 depends on the conductors and their spacing. For parallel-conductor line with air dielectric,

$$Z_0 = 276 \log \frac{s}{r} \quad \text{ohms} \qquad (21 \cdot 4)$$

where s is the distance between centers and r is the radius of each conductor, with s and r both in the same units. As an example, No. 12 gage wire has a radius of 0.04 in. With 1-in. spacing, the ratio of s/r equals 1/0.04 or 25. The log of 25 is 1.398. Therefore, Z_0 is 276 $\times$ 1.398, which equals 386 ohms.

For air-insulated coaxial line,

$$Z_0 = 138 \log \frac{d_o}{d_i} \quad \text{ohms} \qquad (21 \cdot 5)$$

where d_o is the diameter of the outside conductor and d_i is for the inside conductor. As an example, No. 18 wire for the inside conductor has a diameter of 0.08 in. For ¼-in. diameter of the outside conductor, the ratio of d_o/d_i then is 0.250/0.08 or 3.125. The log of 3.125 is 0.4949. Therefore, Z_0 equals 138 $\times$ 0.4949, or 68 ohms.

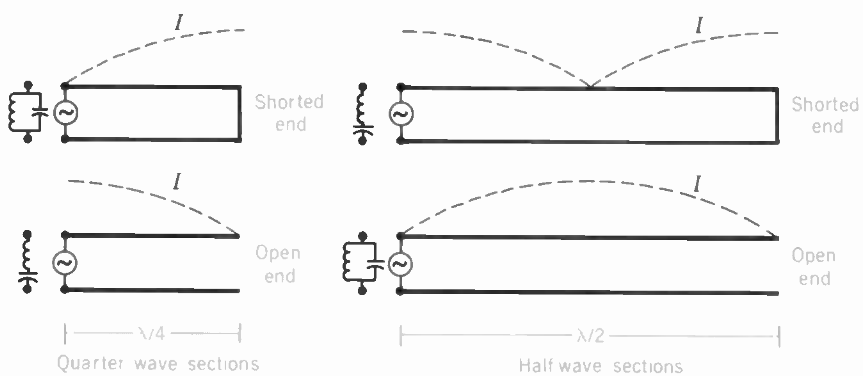
Resonant and nonresonant lines. If the line is terminated with a resistive load equal to Z_0 all the energy traveling down the line will be dissipated in the load. Maximum power transfer is accomplished and no energy is reflected back into the line. Termination of the line in its characteristic impedance makes it in effect an infinitely long line because of the continuity of the line and load impedance. Therefore, a line terminated in its Z_0 is a nonresonant or flat line, because there are no reflections. Then the length of the line is not critical.

If the end of the transmission line is open or short or has any other termination not equal to the characteristic impedance, the current and voltage waves on the line will be reflected from the end of the line, and standing waves are set up on the line just as on an antenna. The ratio of either current or voltage at a maximum point to the value at a minimum point is defined as the *voltage standing-wave ratio,* abbreviated *VSWR*. For a nonresonant line, its VSWR is 1; a line not terminated in Z_0 has a VSWR greater than one. A resonant antenna has a high VSWR.

21 · 13 Transmission-line sections as resonant circuits

When the transmission line is not terminated in its characteristic impedance, the values of current and voltage change along the line, the magnitudes varying with a wave motion that is the same as for an antenna. Therefore, the impedance for different points on the line varies from maximum at the point of highest voltage on the line to minimum at the point of highest current, as the impedance at any point equals the ratio of voltage to current. This is illustrated in Fig. 21·25 showing transmission lines being used as resonant circuits. Since the action of the line in such

Fig. 21 · 25 Transmission-line sections as resonant circuits.



an application depends on the existence of reflections, the lines are not terminated in their characteristic impedance but are either shorted or open at the end in order to produce the maximum standing-wave ratio and the highest Q for the equivalent resonant circuit.

In analyzing the action of the transmission-line sections it should be noted that an open end must be a point of maximum voltage, minimum current, and maximum impedance. Conversely, a shorted end must be a point of maximum current, minimum voltage, and minimum impedance. For each length equal to a quarter wave back from the end of the line the voltage and current values are reversed from maximum to minimum or vice versa. Intermediate values of impedance are obtained for points along the line between the maximum and minimum points.

In Fig. 21·25 it is shown that a quarter-wave section shorted at the end is equivalent to a parallel-tuned circuit at the generator side because there is a very high impedance across these terminals at the resonant frequency; for a length shorter than a quarter wave the line is equivalent to an inductance. Conversely, the open quarter-wave section provides a very low impedance at the generator side of the line; a length less than a quarter wave makes the line appear as a capacitance. The half-wave sections, however, repeat the impedance at the end of the line to furnish the same impedance at the generator side, with a phase reversal of the voltage and current.

Transmission-line sections are often called *stubs.* These can be used for impedance matching, as an equivalent series resonant circuit for shorting an interfering r-f signal, or to phase antenna signals correctly in multielement antennas. For phasing sections, a quarter wave produces a 90° change in phase angle between the signal at the input and output ends, while a half-wave section shifts the phase by 180°. To reduce interference, an open stub one-quarter wavelength at the interfering signal frequency can be used. One side is connected across the antenna input terminals on the receiver, while the end of the quarter-wave stub is left open to produce a short at the receiver input one-quarter wave back from the open end. The same results can be obtained with a half-wave stub shorted at the end.

21·14 Impedance matching

In order to obtain maximum efficiency and eliminate reflections in the antenna system that may produce ghosts or reduce the detail in the picture, the impedance of the antenna, transmission line, and receiver input circuit should be matched. Matching impedances means making the impedance of the load circuit equal to the impedance of the generator producing signal for the load. This is the condition for maximum transfer of power from the generator to its load because no energy is reflected from the load back to the generator. In the receiver antenna system, the transmission line is the load for the antenna and the receiver input circuit is the load for the transmission line. The effect of an impedance mismatch at either end of the transmission line, therefore, is a loss in power transfer and signal level. However, it should be noted that only one end of the transmission line need be connected to its characteristic impedance to eliminate traveling reflections in the line, since there cannot be any reflections from the matched end.

Usual practice is to terminate the transmission line in its characteristic impedance at the receiver end, because the receiver input impedance is designed to be approximately constant for all channels. Therefore, the transmission line used should have a characteristic impedance equal to the receiver input impedance. An impedance match at the receiver end of the line is maintained throughout the television band, as a result, to provide maximum transfer of signal from the transmission line to the receiver and eliminate reflections in the line. Matching the value of the antenna impedance at resonance to the characteristic impedance of the transmission line usually is not critical, since the antenna impedance may vary over a wide range for different television channels as the electrical length of the antenna changes with the operating frequency. An impedance mismatch of 2.5 to 1 results in a 1-db loss of signal. When it is necessary to match impedances, matching sections of transmission line can be employed, or resistance networks are suitable when there is enough signal.

Quarter-wave matching section. When a quarter-wave section of transmission line with a characteristic impedance Z_0 is neither shorted nor open at the end but has an impedance Z_1 connected across one end, the impedance at the other end Z_2 is

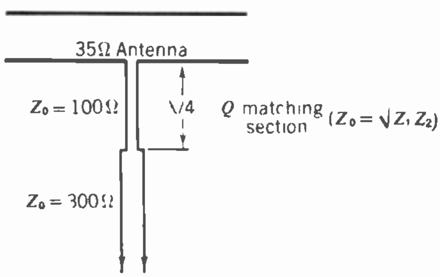
$$Z_2 = \frac{Z_0^2}{Z_1}$$

or

$$Z_0 = \sqrt{Z_1 Z_2} \qquad (21 \cdot 6)$$

Z_0 is the geometric mean of Z_1 and Z_2. Therefore, if a quarter-wave section of line having an impedance equal to $\sqrt{Z_1 Z_2}$ is used to couple two unequal impedances Z_1 and Z_2, the section will provide an impedance match at both ends. Referring to Fig. 21 · 26, a 35-ohm antenna is used with 300-ohm line, and the matching section has an impedance equal to

Fig. 21 · 26 Quarter-wave section matching antenna to transmission line.



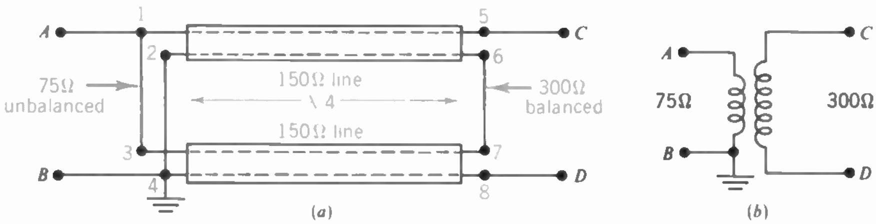


Fig. 21·27 Balancing unit to match between 75-ohm unbalanced impedance and 300-ohm balanced impedance. (a) With two 150-ohm quarter-wave sections. (b) Equivalent circuit.

$\sqrt{35 \times 300}$, or approximately 100 ohms. This equivalent quarter-wave matching transformer is often called a *Q section.*

The length of the quarter-wave section is calculated for the desired frequency from the formula

$$\frac{\lambda}{4}\ (\text{ft}) = V \times \frac{246}{f(\text{Mc})} \qquad (21 \cdot 7)$$

V is the *velocity factor,* which depends on the velocity of propagation along the line. This is less than the speed of light because of the reduced velocity in solid dielectric materials. Values of the velocity factor V are listed in Table 21·1 for different types of line. As an example, a quarter-wave matching stub at 67 Mc using twin lead would have a length of 2.9 ft.

The quarter-wave matching section has the advantage of producing an impedance match with very little attenuation of the signal, but the section can provide a match only for frequencies at which it is approximately resonant. However, the stub functions in the same way when its length is ¾λ or any odd multiple of a quarter wave. Therefore, it can be cut to serve for both the low- and high-frequency VHF television bands.

Balancing unit. Two Q sections can be combined to make the balancing and impedance-transforming unit illustrated in Fig. 21·27. This is called a *balun,* for matching between balanced and unbalanced impedances. At one end, the two 150-ohm quarter-wave transmission-line sections are connected in parallel, resulting in 75-ohm impedance across points A and B. Either A or B can be grounded to provide an unbalanced impedance at the ungrounded point with respect to ground. At the other end, the two 150-ohm quarter-wave sections are connected in series to provide 300 ohms impedance between points C and D. The quarter wavelength of line isolates the ground point from C or D, allowing a balanced impedance with respect to ground.

Either side of the balun can be used for input, with the opposite side for

Attenuation	R_1	R_2
6 *db*	47 Ω	390 Ω
10 *db*	82 Ω	220 Ω
20 *db*	120 Ω	68 Ω

Attenuation	R_3	R_4
6 *db*	22 Ω	100 Ω
10 *db*	33 Ω	51 Ω
20 *db*	56 Ω	15 Ω

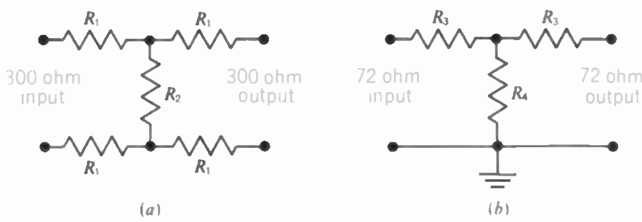


Fig. 21·28 Resistance pads for attenuating antenna signal. (a) Balanced H pad for 300-ohm twin lead. (b) Unbalanced T pad for 72-ohm coaxial line.

output. Useful applications are matching 72-ohm coaxial line to 300-ohm receiver input, or 300-ohm twin-lead to 75-ohm unbalanced input.

Resistance attenuator pads. In cases where excessive antenna signal causes overloading, the signal can be attenuated without introducing any impedance mismatch by using the resistance networks, called *pads,* illustrated in Fig. 21 · 28. The H pad in *a* is for 300-ohm balanced input and output impedances; the T pad in *b* for a 72-ohm unbalanced arrangement. The input terminals of the pad are connected to the transmission line, while the output terminals connect to the antenna terminals. The entire pad is mounted at the back of the receiver. Carbon resistors of the smallest wattage are used; wire-wound resistors are not suitable because of

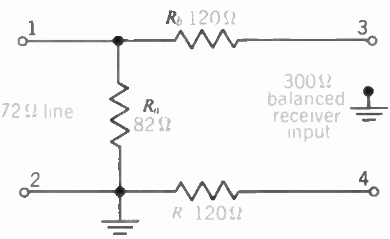


Fig. 21·29 Balanced L pad for matching between 72-ohm unbalanced impedance and 300-ohm balanced impedance.

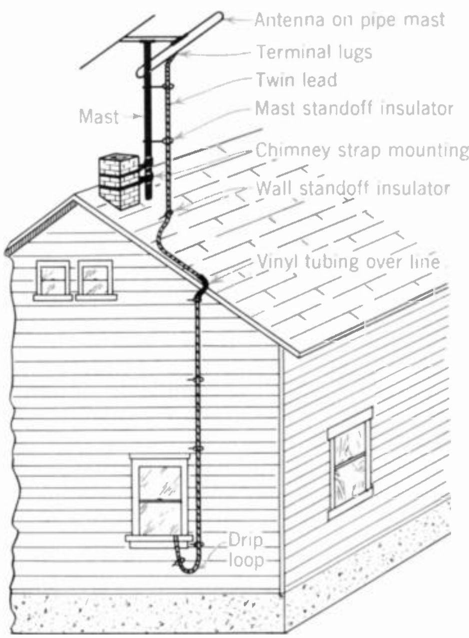


Fig. 21·30 An antenna installation.

their inductance. The amount of attenuation required is usually about 6, 10, or 20 db, which correspond to voltage-loss ratios of ½, ⅓, and 1/10, respectively.

Resistance matching pad. When it is desired to attenuate the signal and match a 72-ohm coaxial line to 300-ohm receiver input, the resistance pad shown in Fig. 21 · 29 can be used. The balanced L pad provides an impedance match in both directions, from the 72-ohm transmission line into the pad and from the 300-ohm receiver input circuit into the pad. Regarding the transmission line as a generator supplying signal, it is terminated at points 1 and 2 in a resistance of 82 ohms shunted by 540 ohms, equal to approximately 72 ohms. Looking from the receiver into the pad, the impedance across the receiver input terminals at points 3 and 4 is 240 ohms in series with the parallel combination of 82 ohms and 72 ohms, equal to approximately 300 ohms. Matching from the receiver side into the pad avoids detuning the r-f input circuit, which can change the r-f gain and bandwidth.

The use of a resistance matching pad has the advantage of allowing an impedance match that does not vary with frequency. However, the attenuation inserted by the pad in Fig. 21 · 29 is approximately 11 db. Without the pad, the attenuation caused by the 4:1 mismatch is only 2 db, but the reflections on a long transmission line may result in ghosts in the reproduced picture. To match from a 72-ohm unbalanced impedance to a 300-ohm balanced impedance without attenuation of the signal, the balun or an equivalent transformer is generally used.

21 · 15 Antenna installation

In locations of average signal strength within about 25 miles of the transmitter, usually typical of suburban areas, the dipole with reflector has adequate gain and directivity. A typical installation is illustrated in Fig. 21 · 30. When channels in different bands must be received, a multiband antenna is necessary. To obtain more signal, an array of two antenna bays stacked vertically is commonly used. In fringe areas far from the transmitter, where the field strength is very low, arrays of three or four bays may be necessary. It should be noted that locations in crowded city areas close to the transmitter but surrounded by tall buildings can have very weak signal with severe reflections.

Antenna mounting. The main requirements for a typical outdoor installation are to mount the antenna near a line of sight broadside to the transmitter, if possible, and as high as possible. Increased height often results in more antenna signal and may also reduce pickup of external noise and interference. The antenna should be at least 6 ft away from other antennas and large metal objects. It is important to note that changing the antenna placement only a few feet either horizontally or vertically sometimes can make a big difference in the amount of antenna signal because of standing waves of the radio signal in areas where there are large conductors nearby, such as a location inside a steel building or between buildings.

Antenna orientation. Before being clamped tight in its mounting the antenna is rotated to the position that results in the best picture and sound. For suburban locations and fringe areas, this is usually broadside to the transmitter location. All the antennas in one neighborhood generally face the same way. In city areas, however, the antenna may receive more signal over a reflected path from a high building nearby. When several channels must be received in different directions, it is helpful to use an *antenna rotator*, which is a motor-driven arrangement to turn the antenna mast from a remote-control switch at the receiver location.

Selecting the transmission line. In most cases, 300-ohm twin-lead ribbon line is used for the run from the antenna to the receiver, which generally has 300-ohm input impedance. For long runs, heavy-duty ribbon line should be used for greater strength. Where the picture quality is affected seriously by wet weather, the tubular twin lead is preferable. When there is a problem of pickup of noise and interference by the line, shielded cable should be used. In locations where the transmission line must be run down a corner of an apartment house parallel to many other lines, for instance, shielded line is preferable in order to minimize pickup of interference from the other lines. For long line runs that must be strong and have minimum attenuation, the open-wire type of line can be used.

The transmission-line run. The run should be as short as possible. Excess line is cut off in order to minimize attenuation of the signal. Do not coil extra line because the inductance acts as an r-f choke to reduce the signal. Twin lead is generally twisted about one turn per foot to improve the electrical balance and strengthen the line mechanically so that it does not flap too much in the wind.

Grounding. To prevent accumulation of static charge on the mast, it can be grounded by connecting a heavy ground wire to a cold-water pipe or to a metal stake driven into the earth. In some areas, local regulations require grounding the mast.

Lightning arrestor. When the antenna is a high point in the area, a lightning arrestor should be installed on the transmission line. The lightning arrestor provides a high-resistance discharge path that prevents static charge from accumulating on the antenna; and, if lightning should strike the antenna, the arrestor is a short circuit to ground that protects the receiver. Indoors, the lightning arrestor is usually mounted by a grounding strap on a cold-water pipe. For use outdoors, a weatherproof type of arrestor can be installed on a grounded mast.

Impedance-matching applications. It is best to use transmission line having a characteristic impedance the same as the input impedance of the receiver, since then a match at one end of the line is approximately maintained for all channels. This generally means using 300-ohm line. The impedance match can be checked by noting the effect of added capacitance when you touch the line. When your hand capacitance at any position on the line decreases the picture strength, this indicates the impedance match is correct.

At the antenna, it is preferable not to match to a higher value of line impedance when the antenna is used for several channels and is cut for the lowest frequency. The antenna impedance then increases with frequency and the mismatch allows a broader frequency response in terms of antenna signal delivered at the receiver. If the antenna is used for only one channel, however, more signal can be obtained by matching the antenna to the transmission line, using a quarter-wave matching section. When 72-ohm coaxial cable is used in order to have a shielded transmission line, it can be matched to 300-ohm receiver input by means of a balun.

Indoor antennas. These can be used in strong-signal locations without ghosts. Most common is a simple dipole assembly with telescoping rods of adjustable length, generally placed on the receiver cabinet. The rods are extended to full length for the low-band VHF channels. Shortening the rods provides half-wave resonance for the high-band channels. By tilting the antenna on its side, the extended rods can serve as a long-wire V antenna for the UHF channels. As another antenna type, a wide-band triangular dipole can be made of metal foil or screening. A convenient indoor antenna for just two or three channels is a folded dipole formed from 300-ohm twin lead, as shown in Fig. 21·31.

The location of an indoor antenna is usually determined by trial and error to find the spot that provides the most signal. The best position is not always horizontal because the antenna pickup is generally reflected signal from nearby conductors, which is usually not horizontally polarized.

Another possibility is a power-line antenna, where one antenna input terminal is connected to the high side of the a-c power line through an r-f coupling capacitor of about 100 $\mu\mu$f. The signal return path is to earth ground, through stray capacitance. However, the antenna signal is usually weak, depending on peaks and nodes for standing waves of r-f signal on the power line.

Built-in antennas. Most receivers have an antenna mounted on the cabinet. In wood cabinets, a folded dipole made of twin lead is stapled around the inside frame or a short triangular dipole made of metal foil is tacked to the underside of the cabinet top. With portable receivers or any metal cabinet, the antenna cannot be inside. Then a telescoping dipole assembly is often mounted to the back. When a single pole is used for a *monopole antenna*, it operates as a grounded quarter-wave type, connected to the receiver input with a balun. Some sets have a single-turn loop antenna for UHF reception. The diameter is about 10 in., making the loop conductor several wavelengths for the UHF channel frequencies.

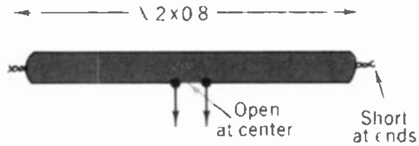


Fig. 21·31 Folded dipole antenna made with 300-ohm twin lead.

If the built-in antenna is used, the location and orientation of the receiver cabinet can make a big difference in the amount of antenna signal. When it is not being used, the built-in antenna should be disconnected from the receiver input terminals, using only the external antenna.

Attic antennas. In a private house, an outdoor type of antenna can usually be mounted in the attic with good results. Since wood and brick are insulators, they have very little effect on the antenna signal.

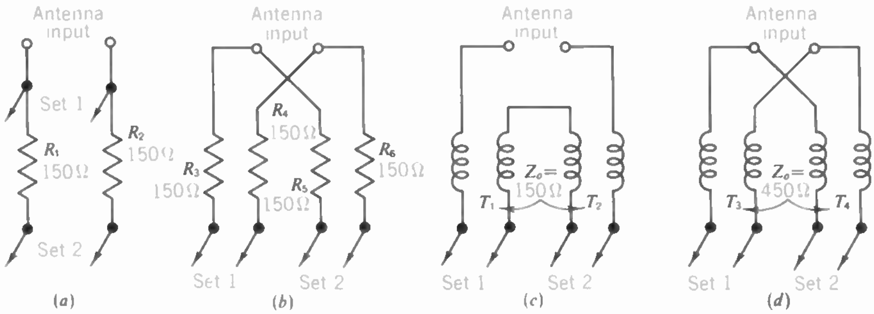
Antenna booster. This is a separate unit containing an r-f preamplifier of one or two stages. The antenna is connected to the input of the booster, while its output goes to the receiver input terminals. The booster must provide a good signal-to-noise ratio to reduce snow in the picture. It may be necessary to install the booster directly on the mast, to amplify the antenna signal before attenuation by line losses. Transistorized boosters are available, with the advantages of compactness and low power requirements. The extra r-f amplification of antenna signal may be needed in fringe areas because of weak signal or the booster can be used to provide enough antenna signal for multiple receivers.

21 · 16 Multiple installations

When multiple receiver outlets from a common antenna system are needed, there are three main requirements:

1. The amount of antenna signal available for each receiver must be at least several hundred microvolts for a good picture without excessive snow. In addition, the signal from the antenna distribution system must be much greater than the amount of signal picked up directly by the transmission line or receiver chassis, without the antenna. Otherwise, there can be ghosts in the picture caused by duplicate signals.
2. Isolation should be provided between receivers to attenuate the local oscillator signal, which can produce beat-frequency interference patterns

Fig. 21 · 32 Two-set antenna couplers. (a) Series resistors for isolation. (b) Parallel branches with resistors for isolation and matching. (c) Elevator transformers T_1 and T_2 in series. (d) Elevator transformers T_3 and T_4 in parallel.



consisting of diagonal bars in other receivers on the common distribution line.

3. The impedance of the transmission line should be matched at the receiver end. This is important with a long line, to prevent ghosts caused by reflections on the line, or when the impedance match is necessary for maximum signal to provide a satisfactory picture.

Considering these requirements, it may be useful to note that several receivers can simply be connected in parallel to one transmission line, with adequate results if there is enough antenna signal, local oscillator interference is no problem, and the transmission line is less than about 50 ft long.

Multiset couplers. These are small, inexpensive units designed to provide multiple 300-ohm output terminals to two to four receivers, with 300-ohm input for the line from the antenna. Since no amplifier is included, a strong antenna signal is necessary because the coupler attenuates the distributed signals for each set. Types of circuits for a two-set coupler are illustrated in Fig. 21 · 32. In *a*, R_1 and R_2 isolate set 2 from set 1. With 150 ohms for R_1 and R_2, set 2 has one-half the antenna input signal. Set 1 has all the input signal. However, the input impedance is less than 300 ohms because of the added line for set 2, connected across the terminals of set 1. Higher values for R_1 and R_2 reduce the mismatch but with more attenuation of the signal for set 2.

A parallel distribution system is shown in *b*. Note that the isolation resistors provide a combined impedance of 300 ohms for the antenna input signal. Here the two parallel paths are 600 ohms each, including each receiver input impedance of 300 ohms. The two impedances of 600 ohms in parallel provide an equivalent 300-ohm impedance for the antenna input connection. Each set then has one-half the antenna input signal. This parallel arrangement can be used for three, four, or five receivers with values of 300, 450, or 600 ohms, respectively, for each of the resistors. Each set would then have one-third, one-fourth, or one-fifth the antenna input signal. Therefore, the antenna signal must be strong enough to allow for the attenuation produced by distribution to the multiple sets. More than five or six receivers should not be connected to one antenna, even with a very strong antenna signal, because the attenuated signal distributed to each set can then be less than the signal picked up directly by the distribution line.

Bifilar quarter-wave elevator transformers are used in *c* and *d*, instead of resistors, for the required impedance matching and isolation. In *c*, the 150 ohms Z_0 of T_1 and T_2 are effectively in series across the 300-ohm terminals. In *d*, the 300-ohm input of each receiver is stepped up to 600 ohms at the antenna input by the 450-ohm Z_0 of T_3 and T_4. The two parallel impedances of 600 ohms provide 300-ohm input for the antenna signal.

Amplified distribution systems. When there is not enough antenna signal to operate several receivers, or when many receivers must be fed from a single antenna system, as in apartment houses, hotels, and motels, a

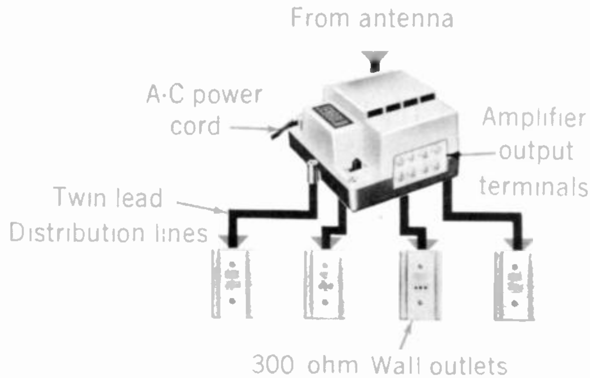


Fig. 21 · 33 Amplifier antenna distribution system for four outlets. Size of amplifier is 5½ by 4½ by 3 in. (Jerrold Electronics Corp.)

master antenna distribution system using r-f amplifiers is necessary. Furthermore, if individual antennas must be used for different channels, the amplifiers provide a means of mixing the antenna signals for common distribution. Signal for the FM broadcast band of 88 to 108 Mc is usually provided also to feed FM receivers. Coaxial cable is generally used for long distribution lines, as it is shielded against noise and direct pickup of signal.

The requirements of an amplified antenna distribution system are illustrated by the arrangement in Fig. 21·33 for the home. Because of the amplifier, enough antenna signal may be provided by an antenna in the attic. Similar to a booster, the amplifier uses a double triode to provide enough signal for distribution to four sets of 300-ohm impedance. Twin lead can be used for the relatively short distribution lines to different rooms in the home. The broad-band amplifier covers 54 to 108 Mc and 174 to 216 Mc to include all VHF television channels and the FM band. Required input signal from the antenna is 100 to 1,000 μv. Minimum gain at each distribution outlet is 0.5 db, with 13 db isolation between outlets. This means that each of the outlets has almost the same amount of signal as one receiver would have connected directly to the antenna. Without the amplified distribution system, all the receivers would have one-quarter of the original signal, or less.

Community antenna systems. In fringe areas where reception of the nearest television broadcast station requires elaborate and costly antenna structures, a master antenna and distribution system is often used to supply signal by coaxial cable to subscribers who pay for this service. Arrays of high-gain receiving antennas are mounted on tall towers at a high point in the terrain, to pick up antenna signal from distant stations. The signals are amplified and combined in a distribution amplifier at the antenna site. Usually, high-band channels are converted to low-band frequencies for less attenuation in the distribution line. A coaxial-cable line feeds the receivers in the community. The community television systems are not licensed by the FCC, as each is considered a private distribution of antenna signal.

21 • 17 Troubles in the antenna system

The trouble symptom of weak picture with excessive snow, on either some channels or all channels, is often caused by insufficient antenna signal. Also, ghosts in the picture can be eliminated only by providing antenna signal without any reflections.

When the transmission line is open, the line serves as an antenna. The signal may actually be stronger than normal on some channels but very weak on most channels. If the signal is better with only one side of the line connected to the receiver input, this effect shows the line is open.

The transmission line can be checked for continuity by an ohmmeter. Remember that the ohmmeter cannot indicate characteristic impedance but only d-c resistance. With a folded dipole antenna connected to the line, it should have practically zero resistance at the receiver end to indicate continuity. With other antenna types, the line can be temporarily shorted at one end to check continuity at the opposite end.

If the transmission line is intermittently open, it will produce flashing in the picture, especially in windy weather. Flashing can also be caused by static discharge of the antenna when there is no lightning arrestor. If the signal is weak only in rainy weather, this usually means worn insulation on the line, which absorbs too much water, causing excessive losses for the antenna signal input to the receiver.

For the UHF channels, vibration of the antenna may produce fading of the picture, because of the very short wavelengths. Clean, tight contacts at the antenna are especially important for ultra-high frequencies to avoid intermittent connections and loss of signal caused by leakage across the terminals.

SUMMARY

1. The length of a half-wave dipole equals $462/f$, where f is in Mc to calculate the half wavelength in feet. The antenna is horizontal for signals with horizontal polarization. The impedance at the center of a straight dipole is 72 ohms. The characteristics of a half-wave folded dipole are similar to the straight dipole but the impedance at the center is 300 ohms.
2. The polar directivity pattern of a half-wave dipole is a figure eight, showing maximum reception broadside and minimum off the end of the antenna. At harmonic frequencies, however, the broadside response decreases and reception increases off the ends. For this reason, dipoles are often angled in a V to improve the broadside response for higher frequencies.
3. Front-to-back ratio compares how much more signal the antenna receives from the front than the back. A ratio of 2:1 equals 6 db.
4. Antenna gain indicates how much more signal an antenna array can receive, compared with a half-wave resonant dipole. A voltage gain of 2 equals 6 db.
5. Ghosts in the picture are generally caused by reflected signal received by the antenna, in addition to direct signal from the transmitter. A directional antenna, with a narrow forward lobe, can eliminate ghosts.
6. A parasitic reflector mounted behind the dipole increases the antenna gain and front-to-

back ratio, to minimize ghost problems. See Fig. 21 · 13 for length and spacing. However, a parasitic director is mounted in front of the dipole. See Fig. 21 · 14.

7. A Yagi antenna is a parasitic array with one reflector and two or more directors. Its features are high gain and good front-to-back ratio with a narrow forward lobe, but relatively narrow bandwidth.
8. Antennas can be stacked vertically one above the other to increase the broadside response of the array. See the broadside array in Fig. 21 · 21.
9. Antennas can be stacked horizontally one behind the other to increase the response through the line of the array. See the double-V end-fire array in Fig. 21 · 22.
10. The main types of transmission line are flat twin lead and coaxial cable. The coaxial line is shielded but unbalanced. Flat twin lead is most common because it is economical, has low losses, and its 300-ohm impedance matches the receiver input impedance.
11. The characteristic impedance $Z_0 = \sqrt{L/C}$ where L is the inductance and C the capacitance per unit length of line. Z_0 is independent of length.
12. Insufficient antenna signal results in a weak, snowy picture. Then interference in the picture is more noticeable and synchronization is poor because of the weak signal.

SELF-EXAMINATION *(Answers at back of book.)*

Choose (*a*), (*b*), (*c*), or (*d*).

1. The length of a half-wave dipole at 100 Mc is: (*a*) 4 meters; (*b*) 4.62 ft; (*c*) 100 ft; (*d*) 100 meters.
2. To eliminate ghosts in the picture: (*a*) Use a longer transmission line. (*b*) Connect a booster. (*c*) Change the antenna orientation or location. (*d*) Twist the transmission line.
3. In Fig. 21 · 2, the antenna signal for a transmitter 60° off the ends equals: (*a*) 100 μv; (*b*) 300 μv; (*c*) 600 μv; (*d*) 800 volts.
4. A Yagi antenna features: (*a*) stacked V dipoles; (*b*) two or more driven elements; (*c*) high forward gain; (*d*) very large bandwidth.
5. Which of the following is false for a dipole with reflector? (*a*) It should be cut for the lowest channel. (*b*) Gain is increased in the forward direction. (*c*) Front-to-back ratio is increased. (*d*) The line should be connected to the reflector.
6. Which of the following is true? (*a*) Angling a dipole in a V improves the response for low frequencies. (*b*) A folded dipole cannot be angled in a V. (*c*) The impedance of a folded dipole is 300 ohms. (*d*) The impedance of a straight dipole is 300 ohms.
7. Which of the following applies to 300-ohm twin lead? (*a*) Low losses, shielded and unbalanced; (*b*) high losses, unshielded and unbalanced; (*c*) low losses, shielded and balanced; (*d*) low losses, unshielded and balanced.
8. A nonresonant or flat line has no reflections regardless of length because it: (*a*) is usually very short; (*b*) is usually very long; (*c*) has low value of Z_0; (*d*) is terminated in Z_0.
9. Which of the following is true? (*a*) A reflector is in back of the antenna. (*b*) A director is in back of the antenna. (*c*) A reflector is connected to the line. (*d*) A director is connected to the line.
10. Which of the following is true? (*a*) Characteristic impedance can be measured with an ohmmeter. (*b*) Twin lead has lower line losses when wet. (*c*) An open twin-lead line can serve as an antenna. (*d*) Shielded line picks up more signal than unshielded line.

ESSAY QUESTIONS

1. Define the following terms: antenna gain, front-to-back ratio, parasitic element, harmonic antenna, broadside array, end-fire array.
2. Show a straight dipole antenna with length cut for 54 Mc. Also for 174 Mc. Do the same for folded dipole antennas.
3. What is the cause of a ghost in the picture? Why is the ghost usually to the right of the main image?

4. Show the polar directivity pattern of a half-wave dipole alone, and with a reflector. List two differences between the patterns.
5. A dipole is cut for half-wave resonance at 54 Mc. Show its polar directivity pattern at 54 Mc and at its third harmonic frequency of 162 Mc.
6. Which is more desirable to minimize ghosts: narrow bandwidth or a narrow forward lobe for the antenna response? Explain why.
7. List the velocity factors for twin lead, coaxial cable, and open-wire line. Why is the velocity factor for twin lead less than for open-wire line?
8. A ghost caused by multipath reception is 2 in. to the right of the main image, on a raster 20 in. wide. What is the difference in length between the direct and reflected signal paths?
9. Explain briefly how multiple ghost images can be caused by a long transmission line. What is a minimum line length that will produce a noticeable ghost?
10. Explain briefly how a ghost could be produced to the left of the main image. Assume the ghost image varies in strength when a person walks near the receiver.
11. Give three methods of reducing ghosts caused by multipath antenna signals.
12. Give two advantages of dipole with reflector, compared with a dipole alone.
13. Give two differences between a parasitic reflector and parasitic director.
14. Give two advantages and one disadvantage of a Yagi antenna.
15. Draw a folded dipole with an impedance of approximately 1,200 ohms at the center.
16. Describe briefly five types of television receiving antennas, giving one important feature of each.
17. Describe briefly how the directional pattern of a dipole cut for 54 Mc changes from channel 2 to channel 6 to channel 13.
18. Explain briefly why a long-wire V antenna receives best along its center line.
19. Explain briefly why an end-fire double-V array does not need a reflector.
20. Give one example where coaxial line would be preferable to twin lead for a line length of 100 ft to the external antenna.
21. Why is the transmission line terminated in its characteristic impedance?
22. Compare the following lines, giving Z_0 and one additional feature of each: flat twin lead, tubular twin lead, coaxial line.
23. Distinguish between a resonant line and a flat (nonresonant) line. Why is a resonant line critical in length, while a flat line is not? How do they compare in VSWR?
24. Describe briefly five main parts of an antenna installation job.
25. What are two problems to consider when connecting multiple receivers to a common antenna?
26. Why does a weak picture with snow indicate the trouble may be insufficient antenna signal? Give one other cause of this trouble symptom.
27. Give two indications of an open transmission line.

PROBLEMS *(Answers to odd-numbered problems at back of book.)*

1. Referring to Fig. 21·2, tabulate the values of antenna signal at 90°, 120°, 150°, and 180°.
2. Redraw Fig. 21·2 for an antenna with 6 db gain, compared with the antenna shown, and front-to-back ratio of 6 db.
3. Calculate the length of a half-wave dipole with reflector and give spacing, in feet, for 54 Mc.
4. Draw a high-low antenna, with lengths and spacings, using a straight dipole with reflector cut for 54 Mc and folded dipole with reflector cut for 174 Mc. Show the two antennas with separate lines to a double-pole double-throw knife switch.
5. Calculate the length in feet of a half-wave folded dipole made of twin lead for 98 Mc.
6. Show a $\lambda/4$ Q section made of twin lead, matching 300-ohm line to a 72-ohm antenna. Calculate length and Z_0 for the Q section.
7. Calculate the length in feet of a folded dipole for the center frequency of channel 11.
8. (*a*) Draw a Yagi antenna for channel 11, showing lengths and spacings for the folded dipole,

reflector, and two directors. (*b*) Show two of these Yagi antennas stacked vertically, with two alternate methods of phasing the antennas to a common transmission line.

9. What is the length in feet for each leg of a rhombic antenna 2λ long at 470 Mc?
10. (*a*) Draw a double-V end-fire array, showing length and spacing for 69 Mc. (*b*) Show two of these double-V antennas stacked vertically, with two alternate methods of phasing the antennas to a common transmission line.
11. If the signal into the line is 1,000 μv and the signal out is 100 μv how much is the db loss?
12. Show an H pad between 300-ohm input and output giving resistance values to attenuate the signal by one-half.
13. A voltage ratio of 1:4 is a loss of how many db?
14. Draw the diagram of a parallel distribution system with resistance-matching pads for 300-ohm line from the antenna feeding four 300-ohm receivers. What fraction of the input signal is available at each receiver?
15. Using Eq. (21.5) for air-insulated coaxial line, what is the Z_0 for $d_o = 0.64$ in. and $d_i = 0.2$ in.?
16. Calculate Z_0 for the following examples of parallel-conductor line with air insulation: (*a*) $L = 0.225$ μh per ft, $C = 1$ $\mu\mu$f per ft. (*b*) Conductors of 0.04 in. radius with 0.6-in. spacing. (*c*) No. 18 gage wire conductors spaced 0.5 in. apart.

Chapter 22 The FM sound signal

The sound associated with the picture is transmitted on a separate carrier as a frequency-modulated signal. In frequency modulation (FM) the instantaneous frequency of the modulated carrier is made to vary with the amplitude of the audio modulating voltage, instead of varying the carrier amplitude. FM has many advantages over the conventional amplitude-modulation (AM) system, and there is easily enough space in the 6-Mc television channel for the extra bandwidth needed for FM sound signal. The biggest benefit is freedom from noise and interference in FM reception. For broadcasting the picture signal in the standard television channel, however, AM is used in preference to FM. The main reason is that multipath reception of FM picture signals produces severe distortion of the picture, instead of ghosts.

22 · 1 Frequency changes in an FM signal

The idea of frequency modulation can be illustrated by the *wobbulator* circuit in Fig. 22 · 1. The inductance L and capacitance C form the tuned circuit for the oscillator. In parallel with the tuned circuit is the variable air capacitor C_T. Its shaft is driven by a motor to rotate the plates in and out of mesh to vary the oscillator frequency. The result will be frequency modulation of the r-f output from the oscillator.

Assuming that the oscillator is at 100 kc with C_T halfway in mesh, the frequency varies above and below this center value of 100 kc as the capacitor is driven in and out of mesh. With C_T completely in mesh, the added capacitance in the tuned circuit is maximum and the output from the oscillator is at its lowest frequency. When C_T is all the way out of mesh, the oscillator is at its highest frequency. For values of capacitance between the two extremes, the oscillator frequency varies continuously between its highest and lowest values around the center frequency of 100 kc. If the time for one complete revolution of the trimmer capacitor C_T is taken as

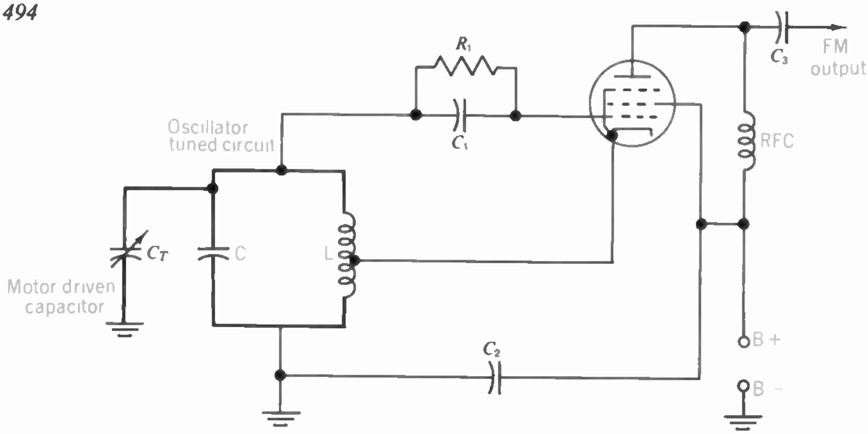


Fig. 22·1 The wobbulated oscillator. The shaft of variable air capacitor C_T rotates in and out of mesh to change the oscillator frequency.

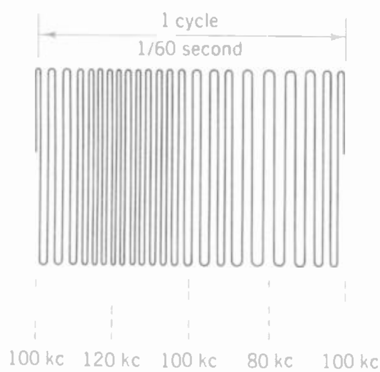
1/60 sec, the r-f output of the oscillator will appear as in Fig. 22·2. The amplitude remains the same at all times but the frequency is changing continuously.

Note the difference between the repetition rate of the frequency swings, which is 60 cps here, and the amount of frequency change, which is ± 20 kc. In this example, the 60-cycle repetition rate is the frequency of shaft rotation to produce complete cycles of capacitance variation in C_T. The 20-kc frequency variation from center frequency depends on how much the capacitance of C_T changes. It is important to note that the frequency swing can be made almost any amount and has no relation to the repetition rate. For the same 60-cps repetition rate, the carrier frequency could be varied by ± 30 kc or ± 50 kc, depending on the amount of capacitance variation. Or, the same frequency swing of ± 20 kc could be obtained at a slower or faster rate than 60 cps, by changing the speed of the shaft rotation.

22·2 *Audio modulation in an FM signal*

Figure 22·3 illustrates four examples of audio voltage producing fre-

Fig. 22·2 The FM signal. Its amplitude is constant but the instantaneous frequency is continuously changing.



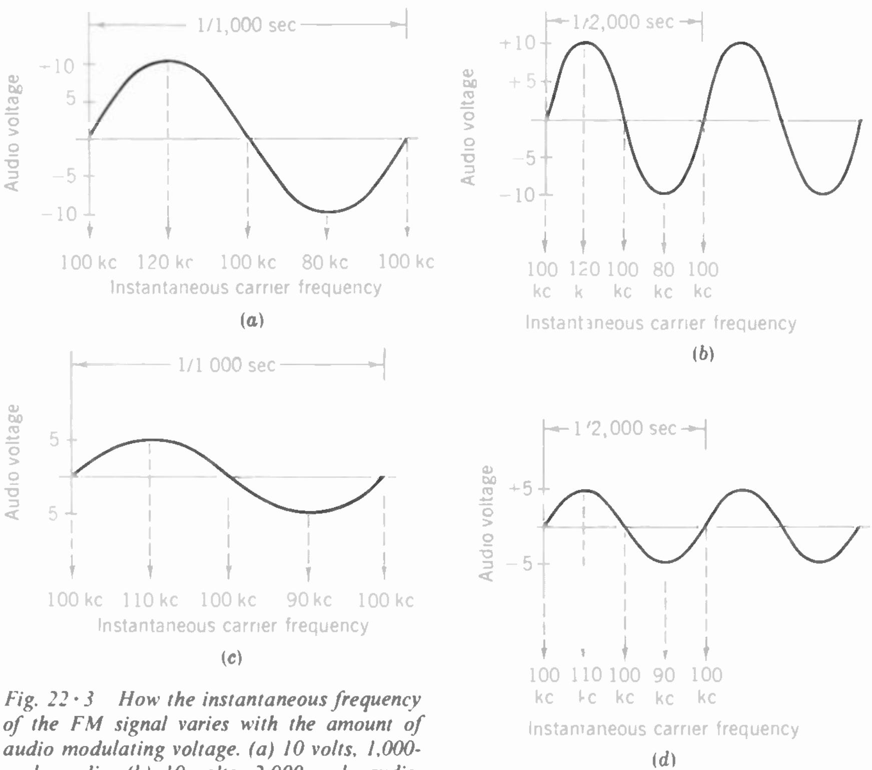


Fig. 22·3 How the instantaneous frequency of the FM signal varies with the amount of audio modulating voltage. (a) 10 volts, 1,000-cycle audio. (b) 10 volts, 2,000-cycle audio. (c) 5 volts, 1,000-cycle audio. (d) 5 volts, 2,000-cycle audio.

quency modulation of an r-f carrier. An audio modulation amplitude of 10 volts is assumed to produce a frequency change of 20 kc with a 100-kc carrier. Also, we assume linear modulation so that one-half the audio voltage produces one-half this frequency swing, or twice the audio voltage would double the frequency swing. Notice that in all cases the amount of audio voltage decides the amount of frequency swing while the repetition rate of the frequency swing is the audio frequency.

For the 1,000-cycle audio voltage in Fig. 22·3*a* the output frequency of the transmitted signal is 100 kc when the audio modulating voltage is at its zero value because this is the carrier frequency with no modulation. With modulation, the transmitted frequency continuously varies in value between the values of 100 ± 20 kc. If the frequency increases for positive values of modulating voltage, it will decrease for negative values. Thus, for the positive half cycles of audio the instantaneous frequency increases from 100 kc to the maximum value of 120 kc, with frequencies between 100 and 120 kc for values of audio voltage between 0 and $+10$ volts. During the negative half cycles the output frequency varies between 100 and 80 kc as the audio voltage varies between 0 and -10 volts.

For the audio voltage in *b* the amount of frequency change is the same as in *a* because the audio amplitude is the same 10 volts. However, the rate at which the transmitted signal goes through its complete frequency swing is now 2,000 cps because of the 2,000-cps audio modulating frequency. Note that the carrier modulated by 2,000-cps audio voltage goes through two complete cycles of frequency swing while the carrier modulated by 1,000 cycles goes through one complete cycle. A complete cycle of frequency swing is from the center frequency up to maximum, down to the middle frequency, decreasing to the lowest frequency value, and returning to center frequency.

For the audio modulating voltage shown in *c* the maximum frequency change is now only 10 kc instead of 20 kc, because the peak value of the audio is 5 volts instead of 10 volts. The rate at which the output signal swings about 100 kc is 1,000 cps. For the modulating voltage of *d* the maximum frequency change is still 10 kc for 5 volts audio. However, the repetition rate is 2,000 complete swings per second, which is the audio frequency.

The amount of frequency change in the transmitted carrier varies with the amplitude of the audio and should not be confused with the audio frequency. The frequency of the audio modulating voltage is the rate at which the carrier goes through its frequency swings. This determines the pitch of the sound as it is reproduced at the receiver. The amount of audio voltage determines the amount of frequency swing, and this determines the intensity or loudness of the sound reproduced at the receiver. These characteristics of FM are summarized in Table 22 · 1.

22 · 3 Definition of FM terms

Center frequency. This is the frequency of the transmitted r-f carrier without modulation, or the output frequency at the time when the modulating signal voltage is at its zero value. The center frequency is also called *rest* frequency.

Frequency departure. This states the instantaneous change of the transmitted signal frequency from the center frequency. For instance, if a transmitted carrier wave having a center frequency of 100 kc is changed to 108 kc by the modulating voltage, the frequency departure is 8 kc. The frequency departure results from modulation, and the amount of frequency change varies with the amplitude of the modulating voltage.

Frequency deviation. The maximum frequency departure from center frequency, at the peak value of the modulating voltage, is the frequency deviation. As an example, an audio modulating voltage having a peak

Table 22 · 1 Comparison of FM and AM signals

FM	AM
Carrier amplitude constant	Carrier amplitude varies with modulation
Carrier frequency varies with modulation	Carrier frequency constant
Modulating-voltage *amplitude* determines r-f carrier *frequency*	Modulating-voltage amplitude determines r-f carrier amplitude
Modulating frequency is rate of frequency change of carrier	Modulating frequency is rate of amplitude change of carrier

value of 5 volts might produce the frequency departure of 10 kc when the modulating voltage amplitude is 1 volt, 20 kc at 2 volts, and 50 kc for its peak value of 5 volts. The frequency deviation in this case is 50 kc. The amount of frequency deviation depends on the peak amplitude of the audio modulating voltage.

Frequency swing. With equal amounts of frequency change above and below center, the frequency swing is twice the deviation. As an example, when the audio modulating voltage has a peak amplitude on either its positive or negative half cycle of 5 volts to produce a frequency deviation of 50 kc, the frequency swing is ±50 kc, or 100 kc. The amount of frequency swing depends on the amplitude of the audio modulating voltage, just like frequency departure and deviation.

Per cent modulation. This is the ratio of actual frequency swing to the amount defined as 100 per cent modulation, expressed in per cent. For commercial FM broadcast stations, ±75 kc is defined by the FCC as 100 per cent modulation. For the aural or sound transmitter of commercial television broadcast stations, 100 per cent modulation is defined as ±25 kc. If, for example, the audio modulating voltage for the associated sound signal in television produces a frequency swing of ±15 kc, the per cent modulation is 15/25, or 60 per cent. Less swing is used for the FM sound in television, compared with FM broadcasting, so that the response of the sound i-f circuits in the receiver can be made broader than the bandwidth of the signal, in order to minimize the problem of tuning in the sound with the picture.

The percentage of modulation varies with the intensity of the audio voltage. For weak audio signals the audio voltage is small and there is little frequency swing with a small per cent modulation. The audio voltage is greater for the louder signals and there is more frequency swing, producing a higher percentage of modulation. The frequency swing produced for the loudest audio signal should be that amount defined as 100 per cent modulation.

Phase modulation (PM). In a phase modulator, the phase angle of the r-f carrier is shifted in proportion to the amplitude of the audio modulating voltage. The varying phase causes changes in carrier frequency. Therefore, phase modulation results in an equivalent FM signal or *indirect FM*. An audio correction filter is used to provide the same frequency swings as in direct FM.

22 · 4 Reactance-tube modulator

A common method of producing FM directly is the system of varying the frequency of an oscillator by means of a reactance tube, as illustrated in Fig. 22 · 4. The oscillator can be of any type except crystal-controlled. Similar to the variable capacitor in the mechanical wobbulator arrangement of Fig. 22 · 1, the plate-to-cathode circuit of the reactance tube is in parallel with the oscillator-tuned circuit. The basic idea of the reactance-tube modulator is the same as the motor-driven wobbulator, with the

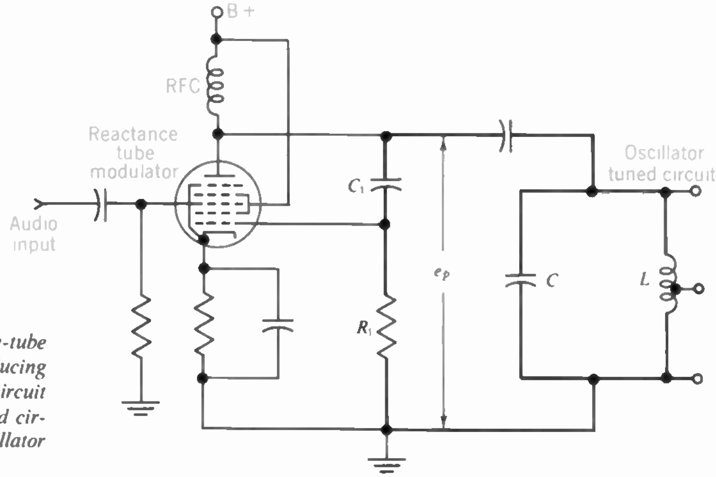


Fig. 22·4 Reactance-tube modulator for producing FM. Its plate-cathode circuit is across the LC tuned circuit to vary the oscillator frequency.

reactance tube substituted for the reactance of the variable trimmer capacitor to vary the oscillator frequency.

The fundamental characteristic of any reactance is that the current flowing through such a circuit element be 90° out of phase with the voltage across it. This effect can be produced by the reactance tube. Whatever reactance the plate-to-cathode circuit of the reactance tube may have is in parallel with the oscillator-tuned circuit and affects the frequency of the oscillator.

The reactance tube can be made to appear as either a capacitance or an inductance. When the plate-to-cathode voltage lags the plate current, the tube appears as a capacitance in parallel with the tuned circuit. When the plate-to-cathode voltage leads the plate current, the tube is in effect an inductance. Furthermore, changing the amount of inductance or capacitance changes the frequency of the oscillator. Since the amount of reactance added by the reactance tube depends upon its transconductance, the injected reactance can be made to vary at the audio rate by applying the audio modulating voltage to the control grid of the reactance tube. In this way, the r-f output of the oscillator is made to vary above and below rest frequency by an amount proportional to the amplitude of the modulating voltage. The repetition rate of the frequency swings is the audio modulating frequency.

Circuit arrangement. Figure 22·4 illustrates the circuit for a reactance-tube modulator. The tube shown is convenient because it has two control grids. One grid can be used for the audio modulating voltage and the other for feedback voltage from the plate. The R_1C_1 branch is connected across the plate-to-cathode circuit of the reactance tube to provide the feedback voltage from plate to grid that makes the tube appear as a reactance. The voltage across R_1 is the feedback voltage. It must be in quadrature (90° out of phase) with the plate-to-cathode voltage. Note that the quadrature network R_1C_1, the tuned circuit LC, and plate-cathode circuit of the reactance tube are effectively in parallel, with the voltage e_p across all three branches. At the operating frequency of the r-f oscillator the coupling capacitor C_c and cathode bypass have no appreciable reactance.

Functions. A reactance tube can be used two ways. When audio signal voltage modulates the bias, the tube serves as an FM modulator, varying the r-f oscillator frequency in step with the modulating voltage amplitude. However, if a d-c control voltage is applied to the grid to shift its bias, then the reactance tube can control the frequency of the oscillator. In this way, the reactance tube serves as a control tube in an AFC circuit to hold the oscillator at its correct frequency.

Quadrature feedback. The capacitance of C_1 is small enough to make its reactance at center frequency at least ten times the resistance of R_1. Then R_1 and C_1 form a series capacitive circuit. Since the R_1C_1 branch is capacitive, its current i_f leads the applied voltage e_p by 90°. The feedback voltage e_{R_1} across R_1 has the same phase as i_f because current and voltage are in phase with each other in a resistance. Therefore, the feedback voltage e_{R_1} leads the plate-cathode voltage e_p by 90°.

This voltage across R_1 is the feedback voltage coupled to the control grid. In a tube, its plate current varies in step with the control-grid voltage. Then the plate current also leads e_p by 90°. Finally, we can say the tube is operating as a capacitive reactance because the current through the tube leads by 90° the voltage e_p across the tube. The plate-to-cathode circuit therefore appears as a capacitive reactance across the oscillator-tuned circuit.

To summarize the phase relations:

1. The current in R_1 and its voltage e_{R_1} lead e_p by 90° because R_1C_1 is a capacitive branch circuit.
2. Plate current through the reactance tube also leads e_p by 90°, since plate current varies in phase with control-grid voltage.
3. Therefore, plate current leads the plate-cathode voltage by 90° and the reactance tube is capacitive.

Types of quadrature networks. Four variations are shown in Fig. 22·5. In all cases the impedance of the grid-to-plate element must be at least ten times the impedance of the grid-to-cathode element. Then the grid feedback voltage is approximately 90° out of phase with the plate-to-

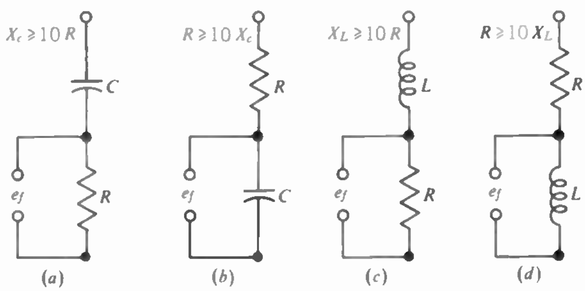


Fig. 22·5 Four forms of quadrature feedback networks. (a) e_f leads 90°. (b) e_f lags 90°. (c) e_f lags 90°. (d) e_f leads 90°.

cathode voltage to obtain a quadrature component of plate current. Because of its quadrature network, the reactance tube is also called a *quadrature tube.*

The quadrature networks in Fig. 22·5 can make the reactance tube capacitive or inductive. When e_f is fed back to the control grid, the reactance tube is capacitive in *a*, inductive in *b*, inductive in *c*, or capacitive in *d*.

22·5 Advantages and disadvantages of FM

The greatest advantage of FM is its ability to eliminate the effects of interference from the desired signal. The interference can be a modulated carrier from another FM or AM station, atmospheric or man-made static, or receiver noise. In any case, the effects of the interference on the desired signal can be made negligible in an FM system. This important advantage is an inherent part of the FM system because the instantaneous frequency variations of the modulated carrier correspond to the desired signal, while the dominant effect of an interfering signal is to change the amplitude of the carrier.

Transmitter efficiency. Because the carrier amplitude is constant in FM transmission, low-level modulation can be used, with class C operation permissible for maximum efficiency in amplifying the FM signal. These factors allow a smaller, more economical, and more efficient transmitter for FM. Furthermore, additions can be made to an FM transmitter just by adding r-f power stages to increase the output.

Broadcast channels. Any FM service with wide frequency swings must be in the VHF or UHF band to allow for the greater bandwidth of the modulated signal. For the commercial FM broadcast band of 88 to 108 Mc, the assigned channel is 200 kc wide for each station. For the FM sound in television broadcasting, 50 kc of the 6-Mc channel is used. Use of these higher broadcast frequencies has the disadvantage of reduced transmission distance, compared with the standard AM radio broadcast band of 535 to 1,605 kc.

Audio-frequency range. In general, the use of the VHF band allows a wider audio modulating frequency range, because at the higher frequencies a wider channel is feasible for accommodating the resultant side-band frequencies. This is true of either FM or AM. An AM radio station is limited to approximately 5 to 10 kc as the highest audio modulating frequency because of the close spacing of assigned carrier frequencies in the standard broadcast band. Given a wide enough channel, though, an AM system can use as wide a modulating frequency range as FM. The picture carrier in television, for example, is amplitude-modulated with video frequencies as high as 4 Mc.

The audio modulating frequency range is 50 to 15,000 cps for commercial FM broadcast stations and the FM sound in television. However, the extent to which this greater audio-frequency range is made useful depends on the quality of the audio system in the receiver.

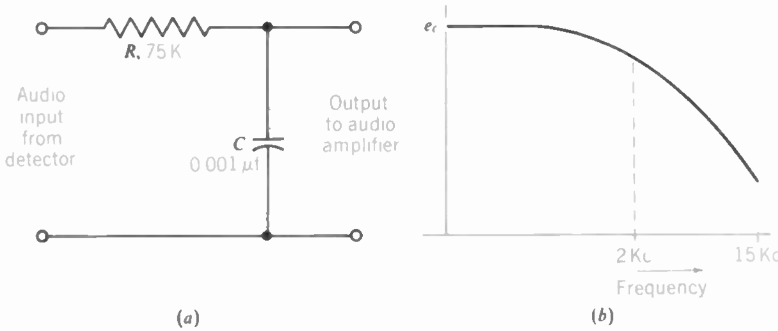


Fig. 22·6 Deemphasis circuit in FM receiver. (a) Low-pass RC filter with 75 μsec time constant. (b) Frequency response.

Multipath reception. Since the instantaneous frequency of an FM signal varies with time, the multipath signals at the receiver generally will have different frequencies at any instant. As a result, heterodyning action between the FM multipath signals at the receiver produces interfering beats that continuously change in frequency. The changing beat frequency can produce garbled sound, similar to the effect produced by nonlinear amplitude distortion in an audio amplifier. In picture reproduction, the interfering FM beat would produce a bar interference pattern in the image with a shimmering effect, as the bars continuously change with the beat frequency. This is why AM is preferable to FM for broadcasting the picture signal, as multipath AM signals simply produce multiple ghost images.

FM is generally used for transmitting the picture and sound signals between microwave radio-relay stations. In this service, however, multipath reception is not a problem because directive antennas beam the signal directly from transmitter to receiver.

22·6 Preemphasis and deemphasis

Any PM interference produces more equivalent FM for higher audio frequencies. Also, the desired signals in the higher audio-frequency range usually have relatively low amplitude because they are harmonics of the fundamental tones and produce little frequency swing. Therefore, it is desirable in an FM system to preemphasize the amplitude of higher audio modulation frequencies at the transmitter. This is done by increasing their relative values before modulation. The result is more frequency swing in the transmitted signal and a higher signal-to-noise ratio. In order to restore the original relative amplitudes, however, the audio signal is deemphasized at the receiver.

For the transmitter, FCC standards specify that preemphasis shall be employed in accordance with the impedance frequency characteristic of an *LR* network having a time constant of 75 μsec. This characteristic increases the audio modulating voltage for higher frequencies. At the receiver the deemphasis should have the same time constant of 75 μsec, but with an opposite characteristic that reduces the amplitude of higher audio frequencies. A low-pass *RC* filter, as in Fig. 22·6, is often used in

FM receivers, generally in the detector output circuit. The deemphasis network attenuates high audio frequencies fed to the audio amplifier. Since the shunt capacitance *C* has less reactance as the frequency increases, the deemphasized audio voltage has less amplitude for the higher audio frequencies. With a 75 μsec time constant, the filter attenuates frequencies at about 1,000 cps and above.

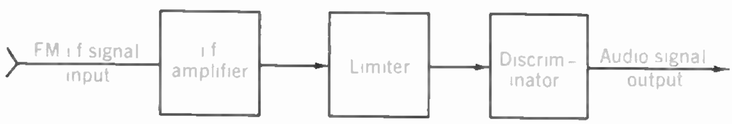
While it would seem that no progress is made if the audio voltage is deemphasized to the same extent that it is preemphasized, a great improvement in signal-to-noise ratio actually is accomplished. The reason is that the preemphasis precedes the effect of the interference. When the signal and noise are both reduced by deemphasis, the signal returns to normal while the noise is reduced below normal. This is more effective in FM than in AM because the noise level in FM increases for higher audio frequencies and the deemphasis attenuates the highest audio frequencies the most.

22 · 7 *Receiver requirements for an FM signal*

An FM receiver is a superheterodyne like a typical AM receiver. The fact that the signal is FM does not alter the heterodyning process, as the original frequency swing is maintained around a lower center frequency corresponding to the i-f sound carrier. Figure 22 · 7 illustrates the requirements of the i-f section. In FM receivers for the FM broadcast band (88 to 108 Mc), the FM i-f signal at 10.7 Mc is obtained from the converter of the r-f tuner. In television receivers, the FM i-f signal is the associated sound for the selected channel. This FM signal for the sound i-f section is obtained from the sound take-off circuit. For intercarrier-sound receivers, the 4.5-Mc sound take-off circuit in the video detector or video amplifier section couples the 4.5-Mc frequency-modulated sound signal to the 4.5-Mc sound i-f section.

I-F bandwidth. In receivers for the FM broadcast band, the maximum frequency swing is ± 75 kc, requiring an i-f bandwidth of at least 150 kc. The maximum frequency swing for the FM sound signal in television is ± 25 kc, which requires an i-f bandwidth of 50 kc or more. While these values may seem high, the required bandwidth is easily attained because it is a relatively small percentage of the intermediate frequency. In receivers for the FM broadcast band, the intermediate frequency generally used is 10.7 Mc. Then a bandwidth of 150 kc is only 1.4 per cent of center frequency. For the 4.5-Mc sound i-f carrier in intercarrier-sound receivers, a bandwidth of 50 kc is 0.9 per cent of center frequency. In AM radio re-

Fig. 22 · 7 I-f section of an FM receiver.



ceivers a bandwidth of 5 kc is 1.1 per cent of the 455-kc intermediate frequency. In all three cases, therefore, the bandwidth is only a small percentage of the center resonant frequency.

An important difference between FM and AM circuits, though, is the effect of narrow bandwidth on distortion of the signal. In an AM receiver, insufficient i-f bandwidth results in loss of high-frequency response in the detected audio signal. This effect of frequency distortion is not too obvious in the detected signal. In an FM receiver, however, narrow i-f bandwidth causes amplitude distortion on loud signals because of insufficient response for maximum frequency deviation. This effect corresponds to clipping and limiting in an audio amplifier, which can actually make the audio signal unintelligible. Therefore, sufficient bandwidth is an absolute necessity for an FM signal.

AM rejection. Eliminating amplitude modulation of the FM signal is possible in the FM receiver because the desired signal is a variation in frequency. Therefore, limiting the amplitude does not distort the FM signal variations. Referring to Fig. 22·7, the last i-f amplifier is operated as a limiter stage to reject amplitude modulation in the FM signal. The limiter is similar to the preceding i-f stages but the d-c operating potentials in the tube enable the stage to function as a saturated amplifier. Constant amplitude of the i-f output signal is maintained, as a result, over a wide range of variations in the amplitude of the input signal.

FM detection. Following the limiter in Fig. 22·7 is the discriminator stage, which is a balanced double-diode detector. Either crystal diodes or vacuum tubes can be used. The discriminator is the second detector to convert frequency changes of the FM signal into corresponding variations in audio signal voltage.

Instead of the limiter-discriminator combination, however, many FM receivers use an FM detector that does not produce audio output voltage for undesired amplitude variations in the FM signal. As an example, the ratio detector circuit does not need a limiter stage.

22·8 Slope detection of an FM signal

Remember that the amount of frequency deviation from the carrier center frequency varies with the amplitude of the audio modulating voltage during the audio cycle. When the carrier frequency of the FM signal falls on a sloping side of the i-f response curve, therefore, the frequency variations of the carrier signal are converted to equivalent amplitude variations (see Fig. 22·8). These AM variations result from the unequal amounts of i-f gain above and below the carrier frequency. Then the i-f output varies in amplitude at the audio rate, in addition to its continuously changing frequency.

The resultant amplitude variations can be coupled to an AM detector to recover the audio voltage. With i-f filtering of the rectifier output, the voltage across the diode load resistor is the desired audio voltage because it varies in amplitude with the amount of input voltage, which in turn

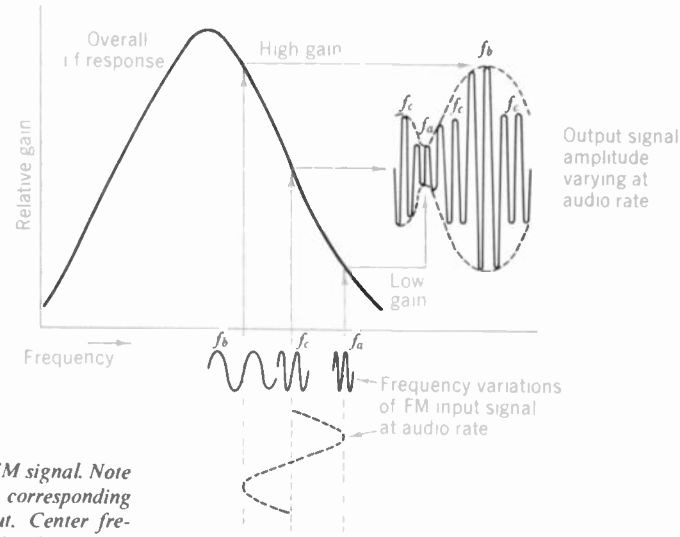


Fig. 22·8 Slope detection of FM signal. Note amplitude variations of output corresponding to frequency variations of input. Center frequency is f_c; f_a is above center; f_b below center.

varies with the amount of frequency deviation. The frequency of the audio output voltage is the same as the original modulating audio signal, since both are equal to the rate at which the FM signal goes through its frequency swing. Note that the frequency variations of the signal cannot disappear. What happens is that the output signal also has amplitude variations corresponding to its frequency variations.

The principle of slope detection is important because it illustrates the two basic requirements of FM detection: (1) converting the FM signal into equivalent amplitude variations for i-f signal input to a diode rectifier and (2) rectifying the audio variations in i-f signal amplitude.

In addition, slope detection is the reason why an FM signal can produce audio output in an AM receiver. As an example, the FM sound in the AM picture channel of the television receiver can be detected in this way, producing an interfering audio signal in the video amplifier and the resultant interference pattern of sound bars in the picture. In general, slight mistuning in an AM receiver can allow slope detection of an FM signal.

22·9 *Triple-tuned discriminator*

A discriminator uses two diode detectors in a balanced circuit as shown in Fig. 22·9. Each diode is a rectifier with its own load resistor. When positive signal voltage is applied between plate and cathode of D_1, plate current flows from cathode to plate, through the transformer secondary, and back to the cathode through the diode load resistor R_1. Its bypass C_1 acts as an i-f filter, just as in the ordinary AM detector. With positive signal voltage applied to D_1, therefore, the rectified current produces an IR drop across R_1. The cathode side is positive as indicated in the diagram. In the same way, when signal voltage is applied to D_2, plate current flows from cathode to plate, through the secondary L_2, and back to cathode through R_2. This produces an IR drop across R_2 with the cathode side positive, resulting in a voltage across R_2 of opposite polarity from the rectified voltage across R_1. Therefore, we can consider the two diodes of a discriminator as being connected in series opposition for d-c output voltage.

Assume that the i-f signal voltages applied to the two diodes are equal, at 5 volts, for example. Then the voltages across R_1 and R_2 will be equal, with a value close to 5 volts for each. However, the output voltage is taken from the cathode load resistors between chassis ground and the point marked *A* for audio output. The output here includes the voltages across both R_1 and R_2. These are in series opposition. Therefore, the net output voltage available at *A* with respect to chassis ground is equal to the difference between the two rectified voltages. If the voltages across the diode load resistors are equal, the net output voltage will be zero. When equal i-f signal voltages are applied to the two diode rectifiers, therefore, the rectified output of each is the same and the net output voltage is zero.

Audio output voltage. Assume now the signal voltages applied to the

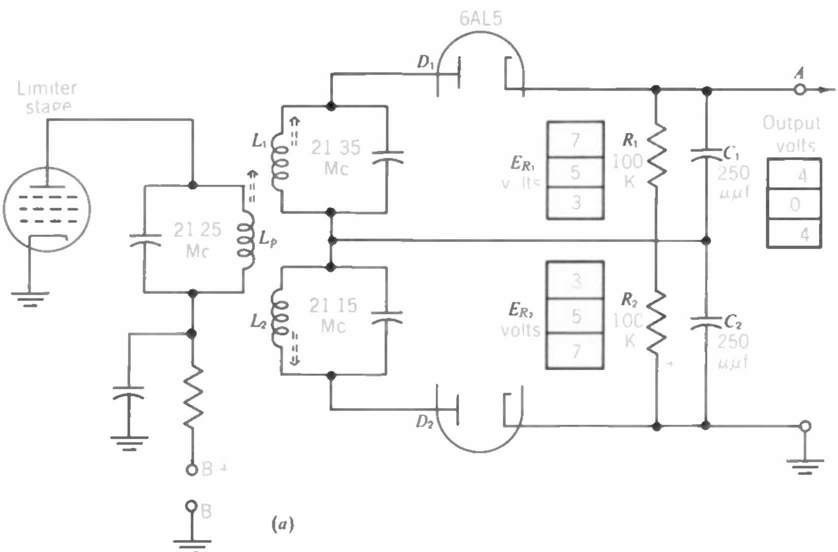
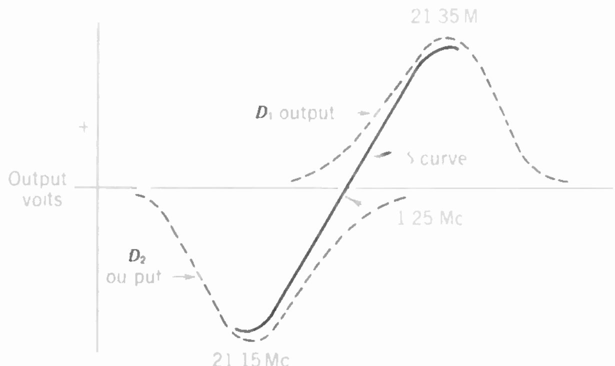


Fig. 22·9 Triple-tuned discriminator for an i-f center frequency of 21.25 Mc. (a) Circuit. (b) S-response curve.



two diodes vary. For example, the plate voltage for D_1 increases from 5 to 7 volts while the plate signal for D_2 decreases correspondingly to 3 volts. Then V_1 will conduct more plate current, producing a larger IR drop across R_1. However, D_2 conducts less current to produce a smaller voltage drop across R_2 than was obtained with 5 volts applied.

Now the voltages are unbalanced. The voltage across R_1 rises to 7 volts while the voltage across R_2 decreases to 3 volts. The net output at point A is the difference between 7 and 3 volts, or 4 volts, positive with respect to ground because the positive voltage is greater.

When the applied signal voltages are reversed, 7 volts is applied to V_2 and 3 volts to V_1. Then the voltage drop across R_2 is 7 volts with only 3 volts across R_1. The circuit is again unbalanced with an output voltage of 4 volts, but this is now negative.

I-F input voltage. The frequency swings are converted to corresponding amplitude variations in the signal applied to the two diodes by means of the triple-tuned coupling circuit for the i-f signal. Assume first that the instantaneous frequency of the FM signal deviates above center frequency. Then the secondary tuned circuit L_1C_1 develops a greater signal voltage because the signal frequency is closer to the resonant frequency of this tuned circuit. The secondary voltage developed across L_2C_2, however, is now less than at center frequency because the signal frequency is farther removed from its resonant frequency. With more signal voltage applied to D_1 and less to D_2 for a frequency deviation above rest frequency, diode 1 provides a greater rectified output voltage than diode 2. A net output voltage of positive polarity results. When the instantaneous frequency of the FM signal deviates below center frequency, more signal voltage is developed across the tuned circuit L_2C_2 and a negative output voltage is obtained. Thus, with the primary tuned to center frequency and having a frequency response broad enough to provide uniform response for the total frequency swing of the FM signal, the two secondary circuits alternately provide more or less i-f signal voltage for the two diode rectifiers. These amplitude variations correspond to the frequency variations in the FM signal.

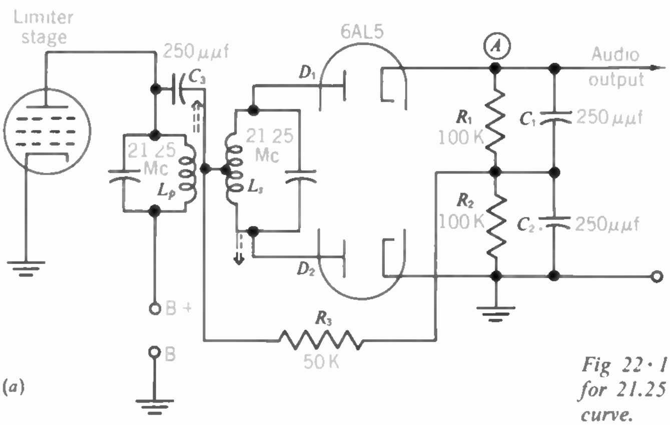
Discriminator response curve. The typical S-shaped discriminator response curve is shown in Fig. 22·9*b* for this balanced detector. The output at center frequency is zero. For frequencies above center, the output voltage is positive and increases progressively for an increasing swing away from center frequency. Similarly, the output voltage is negative when the input signal is below center frequency. It should be noted that the same S-shaped curve can be obtained with opposite polarity by reversing connections to the diodes.

The composite response curve is the net resultant of the sloping responses of the two individual secondary tuned circuits, with each providing output of opposite polarity because of the balanced arrangement of the diodes. The positive and negative peaks on the discriminator response curve occur at the resonant frequency for each of the two secondary tuned circuits. The separation between peaks should be great enough to accommodate the total frequency swing over a linear portion of the discriminator response curve. The linear portion includes about one-half the distance between peaks. With FM signal input, we can consider the instantaneous frequencies to be swinging up and down the linear part of the S curve to provide amplitude variations corresponding to the audio modulation. This S-shaped response is characteristic of any FM detector circuit.

This triple-tuned discriminator is also called a *stagger-tuned circuit* or a *Travis discriminator.* It features good linearity of the S curve to provide linear detection for faithful recovery of the audio modulation. However, the triple-tuned arrangement is seldom used because it is difficult to align with mutual coupling between three tuned circuits.

22·10 Center-tuned discriminator

Figure 22·10 shows the circuit of a balanced discriminator using only two tuned circuits. Both are tuned to center frequency. This is the most frequently used discriminator circuit, generally called the *phase-shift,* or *Foster-Seeley,* discriminator. L_p and L_s are the primary and secondary of the i-f transformer used to couple the signal inductively from the limiter stage. In addition, signal voltage from the primary is coupled to the center tap of the secondary by the coupling capacitor C_3, which has negligible reactance at the signal frequency. The load resistor R_3 provides a d-c re-



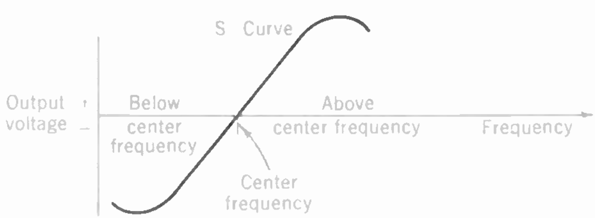


Fig 22·10 Center-tuned discriminator for 21.25 Mc. (a) Circuit. (b) S-response curve.

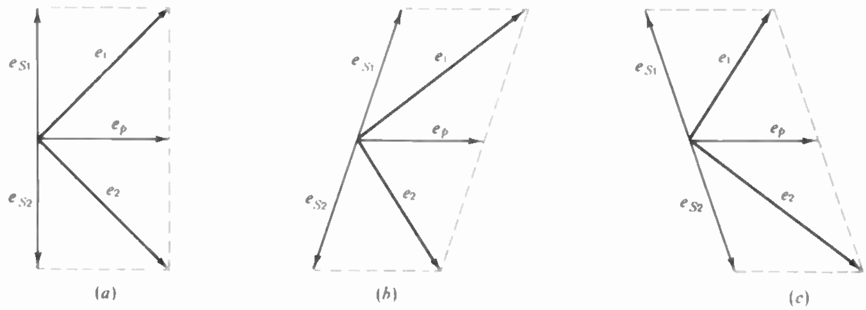


Fig. 22·11 Vector diagrams for center-tuned discriminator. (a) At resonance; 90° phase between e_p and e_{s_1}; $e_1 = e_2$. (b) Above resonance, 70° phase between e_p and e_{s_1}; e_1 larger than e_2. (c) Below resonance, 110° angle between e_p and e_{s_1}; e_1 smaller than e_2.

turn for diode plate current while maintaining a load impedance across the primary. An r-f choke can be used instead of the load resistor R_3.

Balanced detection. The two diode rectifiers are balanced in exactly the same way as in the triple-tuned discriminator. When equal signal voltage is applied to the two diodes, the rectified output is the same for each cathode load resistor and the net output voltage across the two equal series opposing voltages is zero. As the applied voltage for each of the diodes alternately increases, while decreasing for the other, an output voltage is obtained of either positive or negative polarity.

Coupling circuit. As the FM signal varies in frequency, the i-f signal voltage applied to the diodes is made to vary in amplitude by means of the coupling arrangement between the limiter stage and discriminator. Note that signal from the limiter stage is simultaneously coupled to the discriminator in two ways: one by induction across the i-f transformer, and the other by C_3, which is independent of the mutual induction between L_p and L_s. Furthermore, the induced voltage across the secondary is applied to the diode plates in push-pull by means of the center-tap return to cathode. The directly coupled voltage is the primary voltage itself and is applied in parallel to the two diodes, both of which are connected to the primary at the same point. Therefore, the i-f signal voltage for the two diodes is the resultant sum of an induced secondary voltage that is applied in push-pull and the primary voltage applied in parallel to the two diodes.

The secondary voltage e_s is 90° out of phase with the primary voltage e_p at center frequency. This phase relation results because the secondary is tuned to resonance. In general, for a transformer where the secondary is tuned, the voltage across the secondary tuned circuit at resonance is 90° out of phase with the voltage across the primary. Also, the phase angle varies above and below 90° as the applied signal varies above and below the resonant frequency. Since the primary and secondary of the discriminator transformer are tuned to the i-f carrier frequency, which is center frequency for the FM signal, the secondary and primary voltages are 90°

out of phase with each other at center frequency. Both these voltages are simultaneously applied to the diode rectifiers, with the phase relations illustrated by the vector diagrams in Fig. 22 · 11.

In *a* is shown the case for resonance, with the FM signal at center frequency. When the secondary tuned circuit is resonant, e_p is 90° out of phase with e_s. Note that e_s is divided into two equal voltages of opposite polarity, e_{s_1} and e_{s_2}, by the center-tap connection. Therefore, the 90° phase between primary and secondary voltage results in a 90° lagging angle for e_p to e_{s_1} but a 90° leading angle for e_p to e_{s_2}.

The voltage e_1, which is the resultant of its two components e_{s_1} and E_p, is the combined signal voltage applied to diode 1 for rectification. Similarly, the voltage e_2 is the resultant of e_{s_2} and e_p, which is the voltage applied to diode 2. The two applied voltages e_1 and e_2 are equal in this case because of the 90° angle between their components. The 90° phase of these components is illustrated by the sine waves in Fig. 22 · 12.

With equal amplitudes of signal voltage applied to the two diodes, the rectified output of each is the same. Then the net output voltage from the balanced cathode circuit is zero. In the vector diagram, when the i-f signal is above center frequency, however, e_s is less than 90° out of phase with e_p. Then the secondary is tuned below the signal frequency. As a result, e_1 increases as the component voltages of e_1 come closer to being in phase, while e_2 decreases. See Fig. 22 · 11*b*. More signal is applied to D_1 than to D_2 and a positive output voltage is obtained. For the opposite case of a signal frequency below center frequency, e_s is more than 90° out of phase with e_p. The signal voltage e_2 is then greater than e_1 and the rectified output voltage is negative.

Response. The phase-shift discriminator response curve shown in Fig. 22 · 10*b* is the same as for the triple-tuned discriminator. How this response recovers the desired audio voltage from the FM signal can be seen by reviewing the action of the phase-shift discriminator as the transmitted signal is modulated.

When the audio modulating voltage is at its zero value the transmitter output is the r-f carrier at center frequency. Tuned in at the receiver, the

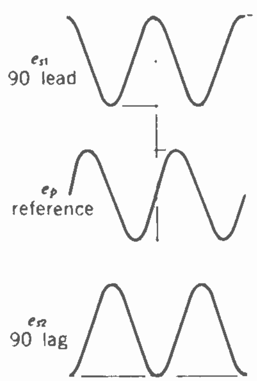


Fig. 22 · 12 Sine waves corresponding to vector phase angles in Fig. 22 · 11a.

signal is converted to the intermediate frequency, amplified in the i-f section, and coupled to the discriminator for rectification. With the signal at center frequency, the i-f transformer coupling the signal to the discriminator is resonant and the secondary voltage is 90° out of phase with the primary voltage. The signal voltages applied to the discriminator diodes then are equal and the output is zero.

As the transmitter is modulated, the audio modulating voltage produces frequency deviations above and below center frequency. Following the modulation through a positive half cycle, the amount of frequency deviation is small when the audio voltage is a little greater than zero. Then there is little discriminator output voltage. While the audio modulating voltage increases to its peak value, the amount of frequency deviation also increases to its maximum. As the circuit becomes more unbalanced the discriminator output voltage increases to its peak value. While the audio modulating voltage declines from its peak value toward zero, the frequency deviation decreases and the discriminator output also declines to zero.

Similarly, on the negative half cycle of audio modulating voltage the FM signal varies below center frequency. Now the signal frequency is below resonance, however, and a negative output voltage is obtained. This output also varies in amplitude with the amount of frequency deviation. As a result, the audio voltage modulating the r-f carrier is reproduced at the output of the discriminator as a changing d-c voltage that varies in amplitude with the frequency deviations in the FM signal.

Effect of interfering amplitude modulation. Since the diodes are balanced at center frequency, the net output voltage is zero regardless of the amplitude of the FM signal. For signal frequencies other than center frequency, however, any variation in amplitude of the i-f signal is reproduced in the output of the discriminator. Therefore, a discriminator is preceded by a limiter stage, which has the function of eliminating AM interference in the FM signal coupled to the discriminator.

22 · 11 The limiter

This circuit is essentially a class C amplifier in which the plate and screen voltages are low, allowing saturation with small values of signal voltage applied to the control grid (see Fig. 22 · 13*a*). A sharp-cutoff pentode tube is used. Also, the plate and screen voltages are lowered in order to reduce the amount of negative grid voltage required for plate-current cutoff. Grid-leak bias is necessary. With varying signal amplitudes, the bias can automatically adjust itself to a value that allows just the positive tip of the signal swing to drive the grid positive and cause grid current to flow. Note the R_2R_3 voltage divider to stabilize the screen voltage with changes in grid bias.

How the limiter functions is illustrated in *b* of Fig. 22 · 13. Note that the signal has a peak amplitude greater than cutoff. The grid-cutoff value depends on the tube and its d-c voltages. The amount of grid-leak bias, though, depends on the amount of signal. The grid-leak bias voltage de-

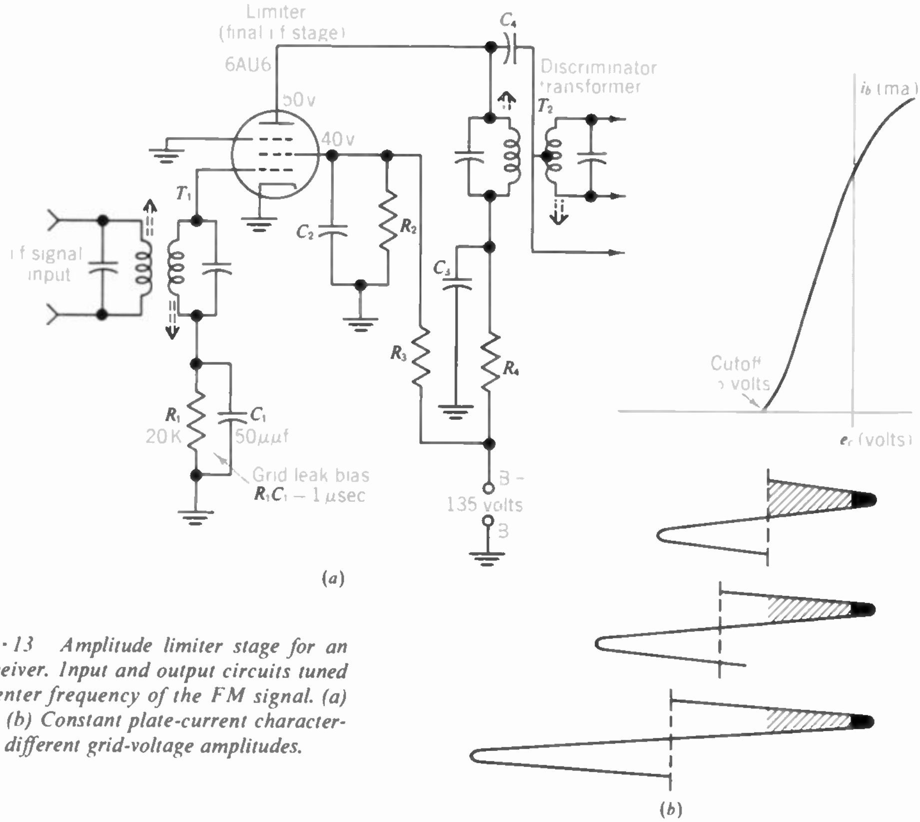


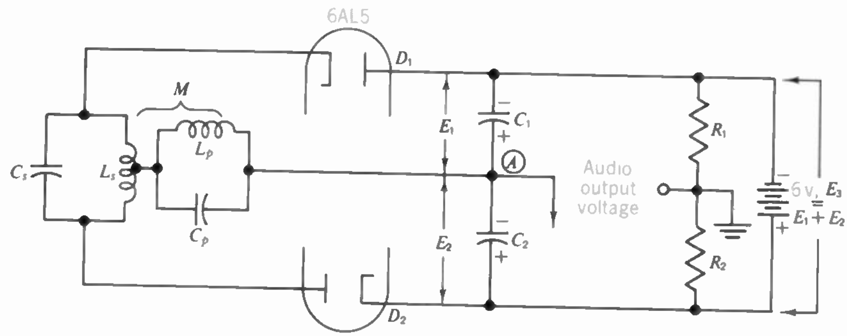
Fig. 22·13 Amplitude limiter stage for an FM receiver. Input and output circuits tuned to i-f center frequency of the FM signal. (a) Circuit. (b) Constant plate-current characteristic for different grid-voltage amplitudes.

veloped will be approximately equal to the peak value of the signal swing. The bias results from grid current, which flows for a very small part of the positive half cycle at the tip of the positive signal swing. Plate current, however, flows for almost the entire positive half cycle, as indicated by the shaded area in the illustration.

When the amplitude of the input grid signal increases, a greater negative bias voltage is developed. The grid-cutoff voltage remains the same, however, and the average plate current changes very little. Therefore, the amount of plate-current flow in the limiter stage is approximately constant for all signals having an amplitude great enough to develop grid-leak bias greater than the cutoff voltage.

With a relatively uniform value of average plate current, then, the output voltage across the tuned circuit is constant. Note that the frequency variations in the FM signal are maintained in the output, since the plate-current pulses are produced at the grid signal frequency.

For satisfactory limiting action, the time constant of the grid-leak resistor and capacitor should be much larger than the period of one cycle of the i-f input signal. This is necessary to maintain the grid bias constant as long as the signal amplitude is constant. However, the time constant must be short enough to allow the bias to change with amplitude variations caused by noise and interference. The bias should follow any amplitude



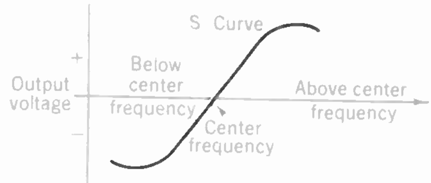


Fig. 22 · 14 Basic ratio detector circuit and its S-response curve. Primary and secondary of the coupling transformer are tuned to i-f center frequency.

variations of the signal in order to keep the peak positive signal swing clamped at zero grid voltage if the plate current is to be constant. Time-constant values of 1 to 4 μsec are commonly used.

22 · 12 *Ratio detector*

This is an FM detector circuit insensitive to amplitude variations in the FM signal. Because no limiter stage is necessary, allowing fewer i-f amplifiers, the ratio detector circuit is often used in FM receivers. As shown in Fig. 22 · 14, the circuit is almost like a discriminator, but the two diodes are connected series-aiding for a ratio detector.

Circuit arrangement. In Fig. 22 · 14, the input coupling transformer has the same function as in the phase-shift discriminator. Both the primary and secondary tuned circuits of the input coupling transformer are resonant at the i-f center frequency. The secondary is center-tapped to produce equal voltages of opposite polarity for the diode rectifiers, while the primary voltage is applied in parallel to both diodes. In the ratio detector circuit, however, one diode is reversed so that the two half-wave rectifiers are in series with the secondary voltage in the input circuit for charging the stabilizing voltage source E_3 in the output side of the circuit. Another difference is that audio output is taken from point A at the junction of the two diode capacitors C_1 and C_2, with respect to the center tap on the stabilizing voltage.

Stabilizing voltage. In order to make the ratio detector insensitive to AM interference effects in the audio output, the total voltage E_3 equal to the diode output voltages $E_1 + E_2$ must be stabilized so that it cannot vary at the audio-frequency rate. Then audio output is obtained at point A only when the ratio between E_1 and E_2 changes, as their sum voltage E_3 remains fixed by the stabilizing voltage source. Since the ratio of the two diode output voltages depends only on the amount of frequency change in the i-f input, audio output is obtained only for the FM signal variations. The 6-volt battery illustrates the stabilizing voltage source connected across both diodes. However, a large capacitor actually is used so that the amount of stabilizing voltage will be determined by the i-f signal level.

The polarity of the stabilizing voltage makes one diode plate negative and the other diode cathode positive. Furthermore, the amount of bias on each diode equals one-half the stabilizing voltage. The i-f signal applied to each diode must have enough amplitude to overcome the bias, therefore, to produce diode plate current.

It is important to note that the stabilizing voltage prevents variations caused by amplitude changes in the total secondary signal voltage across both diodes. This action rejects undesired amplitude modulation in the i-f signal. However, it is the proportion of combined primary and secondary signal voltage applied to each diode that produces the audio output.

Audio output signal. As each diode conducts it produces the rectified output voltage E_1 or E_2 across C_1 or C_2, approximately equal to the peak value of the i-f signal applied to each rectifier. At center frequency, the input transformer proportions the i-f signal voltage equally for the two diodes, resulting in equal voltages across C_1 and C_2. The output voltage at the audio take-off point A is zero, therefore, since E_1 and E_2 are equal. Note that these two voltages have opposite polarity at point A with respect to chassis ground. When the FM signal input is above center frequency, however, D_1 has more i-f signal input than D_2 and the rectified diode voltage E_1 is greater than E_2. This makes point A more positive, producing audio output voltage of positive polarity. Below center frequency, D_2 has more i-f signal input and E_2 is greater than E_1, resulting in audio output voltage of negative polarity at point A. This response is illustrated by the ratio detector S-curve in Fig. 22 · 14, which is essentially the same as the discriminator response curve.

An important characteristic of the ratio detector output circuit can be illustrated by numerical examples. Assume that the frequency deviation above center frequency increases E_1 by 1 volt. This makes point A more positive by 1 volt. At the same time, E_2 decreases by 1 volt. This makes point A less negative by 1 volt, which is the same change as 1 volt more positive. The two rectified diode voltages E_1 and E_2 then produce the identical voltage change of 1 volt in the positive direction at point A. Since the audio signal is taken from point A, the amount of output is the same as though only one diode were supplying audio voltage corresponding to the frequency variations in the FM signal. As a result, the audio voltage ouput of the ratio detector is one-half the output of a discriminator, where the audio signal voltages from the two diodes are combined in series with each other at the audio take-off point. The output in the ratio detector must be taken from the junction of the two diode loads because there is no audio signal voltage across the stabilizing voltage source.

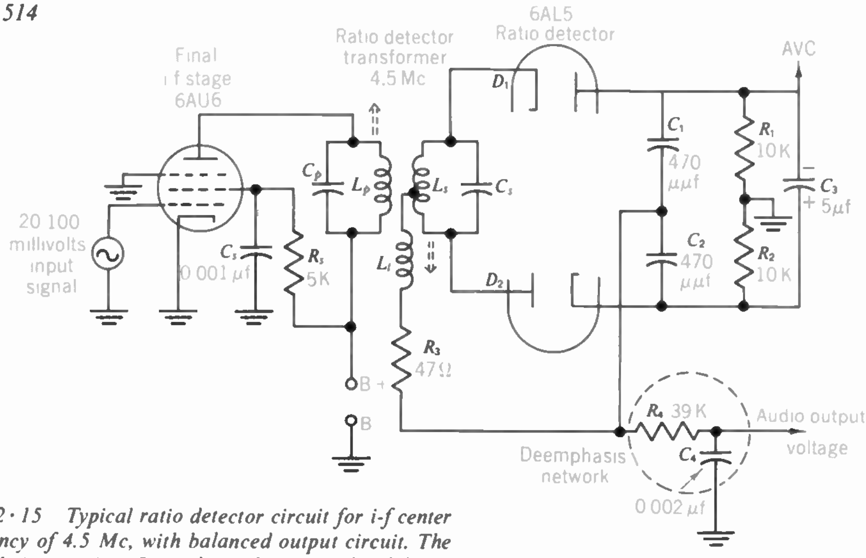


Fig. 22·15 Typical ratio detector circuit for i-f center frequency of 4.5 Mc, with balanced output circuit. The electrolytic capacitor C_3 produces the required stabilizing voltage.

Typical circuit. If a battery were used for the stabilizing voltage, the diodes would operate only with a signal at least great enough to overcome the battery bias on each diode. A large capacitor is used instead, as illustrated by C_3 in the schematic diagram of a typical ratio detector circuit in Fig. 22·15. C_3 charges through the two diodes in series, automatically providing the desired amount of stabilizing voltage for the i-f signal level. The capacitance of C_3, generally called the *stabilizing capacitor,* is large enough to prevent the stabilizing voltage from varying at the audio-frequency rate. A discharge time constant of about 0.1 sec is required, as provided by R_1 in series with R_2 across C_3.

Since the voltage across the stabilizing capacitor is proportional to the amount of i-f signal input, the negative side of C_3 is a convenient source of AVC voltage, to control the gain of preceding i-f and r-f stages, if this is desired. For this connection, the AVC bias is one-half the stabilizing voltage. Whether AVC is used or not, though, the d-c voltage at this point indicates the i-f signal amplitude into the ratio detector.

The tertiary winding L_t of the ratio detector transformer is used to couple the primary signal voltage in parallel to the two diodes. L_t is wound directly over the primary winding for very close coupling, so that the phase of the primary signal voltage across L_p and L_t is practically the same. The construction of a ratio detector transformer with the tertiary winding is illustrated in Fig. 22·16. This arrangement with the tertiary winding, instead of direct coupling to the secondary center tap, is commonly used with the ratio detector circuit in order to match the high-impedance primary and the relatively low impedance secondary, which is loaded by the conduction in the diodes. The resistor R_3 in series with L_t limits the peak diode current, to improve the balance of the dynamic input capacitance of the diodes at high signal levels.

Because of the 90° phase relation between the voltages across the secondary and across L_t at resonance, the circuit provides detection of the FM signal in the same manner as the phase-shift discriminator. Both the primary and secondary of the ratio detector transformer are tuned to the i-f center frequency, which is generally 10.7 Mc in FM broadcast receivers or 4.5 Mc for intercarrier-sound television receivers. The audio output voltage of the ratio detector is taken from the junction of the diode load capacitors C_1 and C_2, in series with the audio deemphasis network, to supply the desired audio signal for the first audio amplifier.

Note the deemphasis circuit for the detector output, consisting of R_4 and C_4 with a time constant of 78 μsec. Since the high audio frequencies are preemphasized for a better signal-to-noise ratio in transmission, the deemphasis circuit attenuates the high frequencies to provide the original audio modulating signal. C_4 serves as a bypass for high audio frequencies, as illustrated by the response curve in Fig. 22·6*b*.

Single-ended ratio detector. Figure 22·17 illustrates a ratio detector circuit that has an unbalanced output circuit. The input circuit, which is shown in its equivalent form, is the same as in the balanced circuit. However, only the one diode load capacitor C_2 is used in the audio output circuit and the stabilizing voltage across R_1 and C_1 is not center-tapped. It is not necessary to ground the center point of the stabilizing voltage source, since the only effect of the different ground connection is to change the d-c level of the audio output signal with respect to chassis ground. The same audio signal variations are obtained from the audio take-off point, about a d-c voltage axis equal to one-half the stabilizing voltage, instead of varying around the zero axis of chassis ground. It should be noted, though, that the audio output always has a zero d-c level with respect to

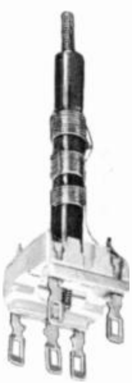
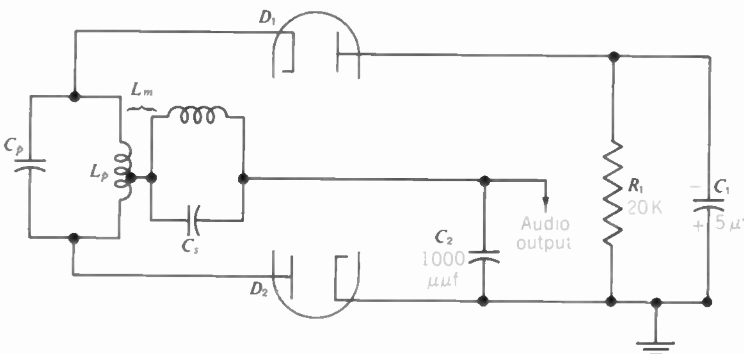


Fig. 22·16 Construction of ratio detector transformer. (Meissner Mfg. Division.)

Fig. 22·17 Ratio detector with single-ended, or unbalanced, output circuit.



the mid-point of the stabilizing voltage source. With the unbalanced arrangement, the entire stabilizing voltage is available for AVC.

Only the one diode load capacitor C_2 is necessary in the output circuit to obtain the audio signal because it serves as the load for both diodes. Each diode charges C_2 in proportion to the i-f signal input for each rectifier, but in opposite polarity. Therefore, the voltage across C_2 is the same audio output signal as in the balanced ratio detector. The capacitance of C_2 is doubled in the single-ended circuit because it replaces two capacitors that are effectively in parallel for signal voltage in the balanced circuit.

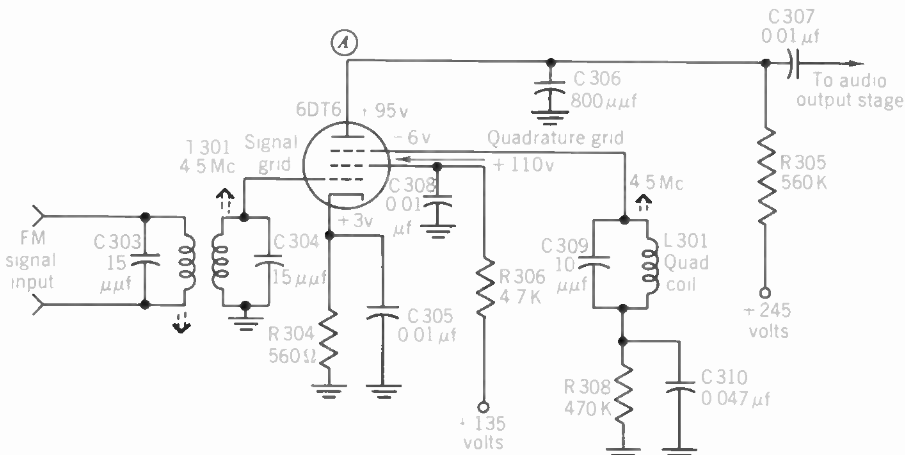
Ratio detector receiver characteristics. Referring to Fig. 22·15, note that only 20- to 100-mv signal is applied to the grid of the i-f stage driving the ratio detector. With a gain of about 100, the i-f output of 2 to 10 volts will be enough to drive the diodes. Since no limiter is required, less i-f gain is needed as there is no fixed threshold voltage that must be exceeded, the ratio detector automatically adjusting itself to the i-f signal level. Fewer i-f stages are necessary, therefore, making the ratio detector receiver more economical, compared with the limiter-discriminator arrangement.

22·13 Quadrature-grid FM detector

This circuit uses a pentode combining the functions of FM detection, AM limiting, and first audio amplifier. As shown in Fig. 22·18, the audio signal output voltage in the plate circuit has enough amplitude to drive the audio power output stage.

Either of two types of tubes can be used. One type is a gated-beam tube such as the 6BN6. The other is a dual-control pentode such as the 6DT6. In either case, the tube is equivalent to a pentode with two sharp-cutoff control grids to control plate current. For the 6DT6 in Fig. 22·18, grid 1 is the usual control grid, with a typical cutoff of −4 volts. The FM signal

Fig. 22·18 Quadrature-grid FM detector circuit. D-c voltages shown at tube pins. (From GE chassis LW.)



input is applied to this grid. In addition, grid 3 is made to serve as a control grid, also with a typical cutoff of -4 volts. This grid has signal by space-charge coupling to grid 1 within the tube.

Quadrature grid. The characteristics of grid 3 are important in the operation of this FM detector circuit. In tubes with multiple grids, when one grid has positive voltage with respect to cathode and the next grid is negative, this grid develops its own space charge corresponding to a virtual cathode. In the 6DT6 here, grid 3 is at -6 volts while grid 2 is the screen grid at $+110$ volts. Therefore, electrons in the space current to the plate are slowed down enough near grid 3 to form an outer space charge, in addition to the normal space charge near the cathode. The outer space charge serves as a source of electrons for the outer electrodes, including grid 3 and the plate.

With signal input voltage to grid 1, the space current is varied. Then the amount of space charge at the virtual cathode near grid 3 also varies. As a result, electrostatic induction by the varying space charge causes signal current to flow in the grid 3 circuit. The resonant circuit here provides signal voltage corresponding to the induced current.

The induced current in the grid 3 circuit lags by $-90°$ the signal voltage at grid 1. At resonance, the voltage across the tuned circuit for grid 3 also lags the grid 1 voltage by $-90°$. This is why grid 3 is called the *quadrature grid.* The $-90°$ phase applies at center frequency of the FM input. When the signal deviates below and above center frequency, the phase of the quadrature-grid signal voltage varies below and above $-90°$.

Gating action. In a tube with two control grids, each controls the plate current. Also, either one can cut off the output. Therefore, the grids serve as voltage-controlled gates to turn the output on or off. When the grid voltage allows plate current, the gate is open; when the grid voltage is more neg-

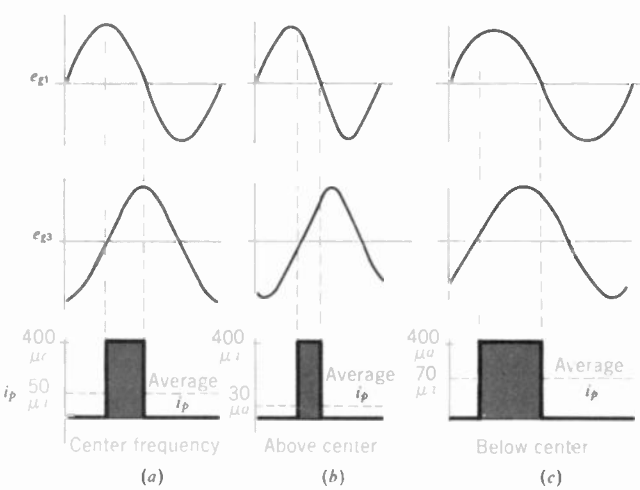


Fig. 22·19 Gating action of grids 1 and 3 on plate current output of quadrature-grid detector. Average i_p varies with FM deviation. (a) At center frequency, e_{g_3} lags e_{g_1} by $-90°$. (b) Above center frequency e_{g_3} lags e_{g_1} by more than $-90°$. (c) Below center frequency e_{g_3} lags e_{g_1} by less than $-90°$.

ative than cutoff the gate is closed. Either grid can gate the plate current on or off. When one gate is closed there is no plate current, even if the other gate is open. Therefore, both gates must be on at the same time to have output plate current. This feature is the basis of detecting the FM signal at grid 1 which is gated by the quadrature signal at grid 3 (see Fig. 22 · 19). Such operation is often called a *gated-beam detector.*

Circuit operation. In Fig. 22 · 18, the 4.5-Mc sound signal is coupled by T_{301} to the signal grid of the 6DT6 FM detector. This signal is also in the quadrature-grid circuit, because of space-charge coupling, but 90° out of phase with the input signal. The quadrature coil L_{301} is adjusted to resonate with C_{309} at the i-f center frequency of 4.5 Mc. Grid-leak bias is provided by R_{308} with C_{310}.

The phasing between the signals at grid 1 and grid 3 is shown in Fig. 22 · 19. At center frequency, in *a*, the two signal voltages are 90° out of phase. Plate current flows only for part of the cycle, when both grids gate i_p on. This time is indicated by the shaded area of the i_p waveform. Above center frequency, in *b*, e_{g_3} lags e_{g_1} by more than $-90°$. Then the two signals are closer to being of opposite phase. Therefore, the on-time for i_p is shorter, resulting in narrower pulses of plate current. Although the amplitude is the same, equal to saturation i_p, the narrower pulses mean a smaller value of average plate current. Below center frequency in *c*, e_{g_3} lags e_g by less than $-90°$. Then the two signals are closer to being in the same phase. Therefore, the on-time for i_p is longer, resulting in wider pulses, with a higher average value.

Remember that these phase changes for e_{g_3} result from frequency deviations in the FM signal, which follow the audio modulating voltage. Similarly, the variations in average i_p follow the modulation in the FM signal. Although the plate-current pulses are at the i-f signal frequency, the average plate current varies with the audio signal. In the plate circuit, the integrating capacitor C_{306} bypasses the i-f variations, allowing the audio output voltage to be developed across the plate load resistor R_{305}. The audio signal amplitude at the audio take-off point *A* is about 14 volts peak value, for 25 kc deviation from the center frequency of 4.5 Mc. This voltage is enough to drive the audio power output stage, without the need for a first audio voltage amplifier.

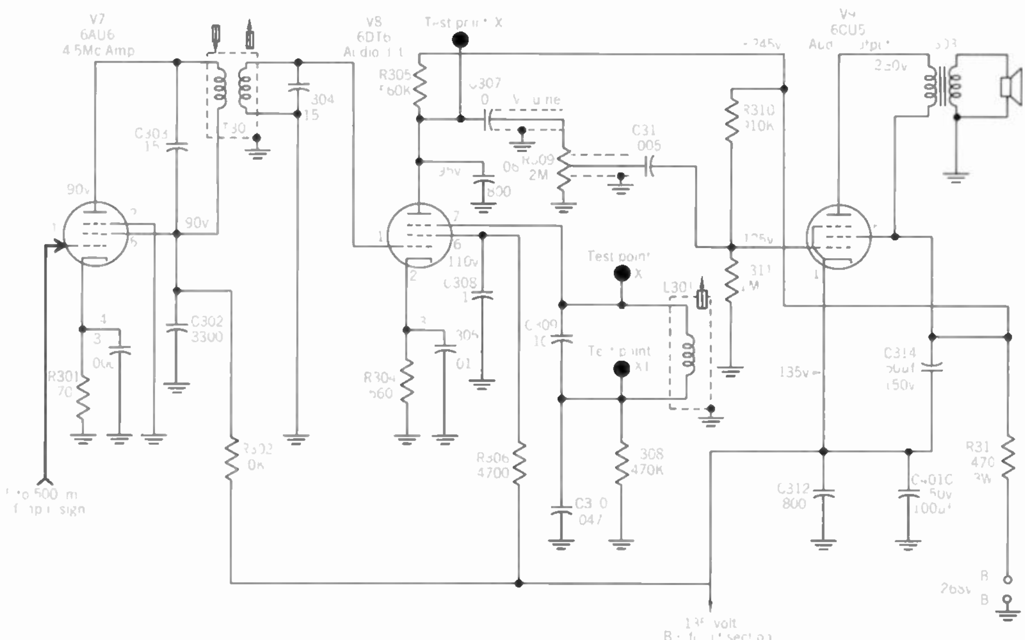
AM rejection. Undesired amplitude modulation is attenuated approximately -30 db, because of the following factors. The cathode bias resistor R_{304} is bypassed for i-f but not audio frequencies. The resultant degeneration of the audio signal reduces the effect of any high-amplitude interference. Also, the i-f transformer T_{301} is damped by grid current when signal amplitude at grid 1 exceeds the cathode bias of 3 volts. The shunt damping lowers the Q of the tuned transformer, reducing the amount of signal voltage. Finally, grid 3 produces saturation plate current when it is driven positive. This grid has grid-leak bias, which increases with higher signal amplitudes to clamp the positive peak at zero voltage. The result is saturation plate current at a peak value of approximately 0.4 ma, for typical signal input, with 3 volts bias on grid 1. In some circuits a variable cathode bias resistor is used as a *buzz control,* which is adjusted for best limiting.

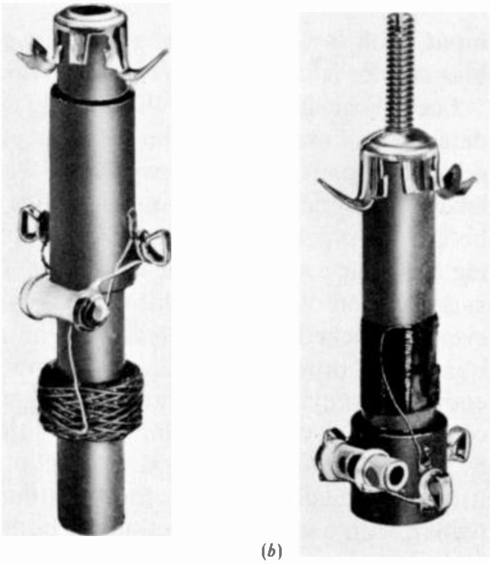
Locked-oscillator mode. Without any input, or for weak signal, the 6DT6 detector will oscillate at relatively low amplitude. The circuit is a tuned-grid tuned-plate oscillator, with grid 3 as the oscillator plate supplying feedback to grid 1 through interelectrode capacitance. Then the FM input becomes a synchronizing signal for the oscillations to lock in the oscillating frequency at the signal frequency. Detection occurs because of the gating action of the two control grids in quadrature, as described. However, the locked-oscillator feature is an advantage in rejecting AM with weak signal input. Then, the signal amplitude in the detector is always high enough to provide the required limiting action. For weak signal input, the oscillations provide the required amplitude of grid voltage, synchronized by the FM input signal. With normal or strong signal, the input voltage itself has enough amplitude for AM limiting. The result is quadrature detection with good AM rejection for both weak and strong signals, in this locked-oscillator quadrature-grid detector circuit.

22·14 Complete sound i-f circuit.

Figure 22·20 shows the complete sound section for an intercarrier-sound receiver using the 6DT6 detector circuit. There are only three

Fig. 22·20 Complete 4.5-Mc sound i-f section with quadrature-grid FM detector. D-c voltages shown at tube pins. Capacitance values less than 1 in μμf (From GE chassis LW.)





(a) (b)

Fig. 22·21 Coils for 4.5-Mc sound i-f signal. Height 2 in. (a) Quadrature coil; L = 125 μh, C = 10 μμf. (b) Take-off trap; L = 27 μh, C = 47 μμf. (Merit Coil and Transformer Corp.)

stages, including one 4.5-Mc sound i-f amplifier as the driver stage for the detector, which is coupled to the audio output stage driving the loudspeaker. Typical coils for the 4.5-Mc sound i-f signal are shown in Fig. 22·21.

Input signal for the 6AU6 driver is 4.5-Mc sound i-f signal from the 4.5-Mc sound take-off trip in the video amplifier (not shown). T_{301} in the 6AU6 plate circuit is a double-tuned transformer resonant at 4.5 Mc to couple the sound i-f signal to grid 1 of the detector. The 6DT6 circuit is the same locked-oscillator quadrature-grid detector shown in Fig. 22·18. Test point X is to check signal voltage at the quadrature grid. Its grid-leak bias can be measured at test point XI, with a d-c voltmeter. Test point XII at the plate is the audio take-off. These test points are available at the top of the chassis for convenience in alignment. C_{307} couples the detected audio output signal to the volume control R_{309}.

Note the d-c voltages for the audio output stage V_9, which is in a stacked B+ arrangement. The V_9 cathode at 135 volts is the d-c supply for plate and screen of the sound i-f stage and two of the picture i-f amplifiers (not shown).

22·15 Sound i-f alignment

This consists of two parts: tuning the i-f amplifiers for maximum output and balancing the detector at center frequency. The final adjustments, however, should be made on the overall i-f response from mixer grid to detector output. Then you are sure the amplifier and detector stages are aligned at exactly the same frequency.

Discriminator-limiter combination. The i-f amplifiers are aligned for maximum output at the limiter grid. A visual i-f response curve here looks like Fig. 22·22*a*. However, the i-f circuits can also be aligned for maximum grid-leak bias, measured by a d-c voltmeter at the limiter grid.

For the discriminator transformer, its S curve can be obtained with an oscilloscope at the audio take-off point. However, the discriminator can also be balanced with a d-c voltmeter, as explained later.

Ratio detector receiver. The i-f amplifiers are aligned for maximum d-c output voltage across the stabilizing capacitor. This i-f alignment includes the primary of the ratio detector transformer but not the secondary. A visual i-f response curve like Fig. 22 · 22*a* can be obtained. However, the stabilizing capacitor must be disconnected temporarily because it bypasses the 60-cps oscilloscope input signal produced by the sweep generator.

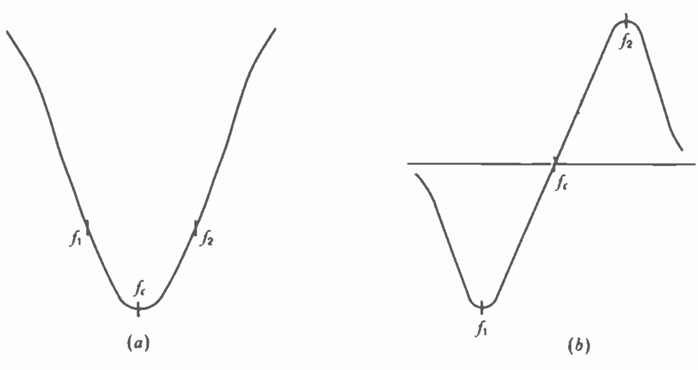
The secondary of the ratio detector transformer is aligned by checking the audio output voltage, not the stabilizing voltage. Either an S curve can be obtained at the audio take-off point, as in Fig. 22 · 22*b*, or the ratio detector can be balanced with a d-c voltmeter, similar to the discriminator alignment.

Quadrature-grid detector. This circuit is usually aligned with FM input signal from a broadcast station. In intercarrier receivers, the sound i-f signal is always 4.5 Mc, which is exactly the center frequency needed for alignment. The procedure is to adjust the quadrature coil, 4.5-Mc i-f transformer, and 4.5-Mc sound take-off trap for best sound, with increasing volume and minimum distortion. In addition, a d-c voltmeter to check grid-leak bias on the quadrature grid will read maximum. This voltage can be measured across R_{308} at test point XI in Fig. 22 · 20.

The quadrature coil is aligned with normal signal level but the other circuits should be tuned with very weak signal input to prevent limiting action during alignment. When the signal is below the limiting level you can hear the hissing sound of receiver noise, with the volume control at maximum. To reduce the signal level, couple the antenna input loosely to the receiver, instead of making a direct connection.

Balancing the discriminator. Adjustment of the discriminator is essentially a problem of tuning the input transformer to the i-f center frequency.

Fig. 22 · 22 Visual response curves for FM sound i-f section. Marker at f_c is center frequency; f_1 and f_2 mark bandwidth. (a) Response of i-f amplifiers. (b) S curve of FM detector.



This can be done without a visual response curve, using just a d-c voltmeter and a conventional r-f signal generator to supply unmodulated output voltage at a signal frequency. A d-c voltmeter is used because the output is a steady voltage when the input signal is not frequency-modulated, since there are no frequency variations to produce variations in the output. The meter should have an impedance of 20,000 ohms per volt or higher to avoid detuning the discriminator.

The phase-shift discriminator is aligned by tuning the primary and secondary circuits of the discriminator transfer to the i-f center frequency. With the generator supplying signal at the i-f center frequency, and the d-c voltmeter connected across the audio output terminals:

1. Tune the primary of the discriminator transformer for maximum output.
2. Tune the secondary of the discriminator transformer for a sharp drop to zero.

It should be possible to produce either a positive or negative output voltage when adjusting the secondary. Therefore, it should be tuned for zero indication at the balance point where the output voltage starts to swing from one polarity to the other. A gradual decrease to zero is the wrong indication, as this only means the circuit is being detuned farther away from center frequency.

When the signal generator frequency is varied manually above and below center frequency, the d-c output voltage should vary from zero at center frequency to a maximum value at both sides of center frequency, with opposite polarities below and above center. The response should be symmetrical about center frequency with equal output voltages produced for the same amount of frequency change below or above center frequency. The two points of maximum output voltage, corresponding to the two peaks on the discriminator response curve, should have the required frequency separation. The actual polarity of the output voltage for a frequency departure above or below center does not matter. It is only required that the output voltages be of opposite polarity for frequency departures above and below center frequency.

Balancing the ratio detector. With the generator supplying signal at the i-f center frequency:

1. Tune the primary of the ratio detector transformer for maximum output on the d-c voltmeter across the stabilizing capacitor.
2. Move the d-c voltmeter to the audio take-off point. Then tune the secondary for zero balance.

When the ratio detector has a single-ended output circuit, however, the audio output has a d-c level equal to one-half the stabilizing voltage. Therefore, usual practice is to insert two balancing resistors temporarily, as shown in Fig. 22 · 23. This converts the single-ended arrangement to a balanced output circuit, so that the ratio detector secondary can be aligned for balance at zero with the d-c voltmeter.

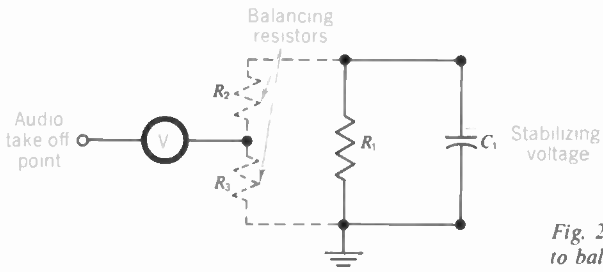


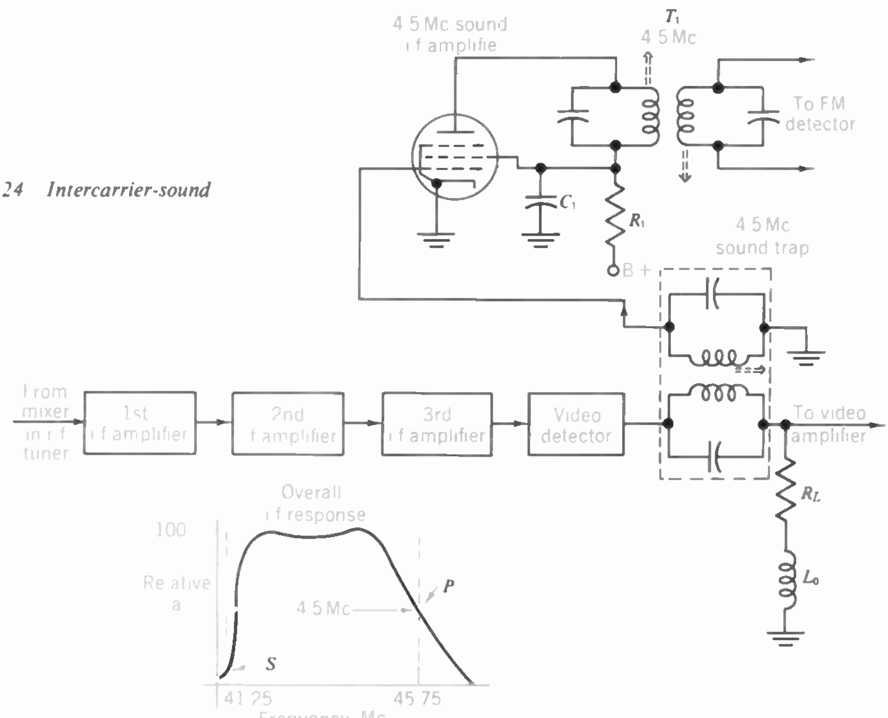
Fig. 22·23 Temporary resistor connections to balance audio take-off point to ground for aligning secondary of single-ended ratio detector with a d-c voltmeter.

22·16 Intercarrier sound

Practically all television receivers use the intercarrier system of converting the associated sound into a frequency-modulated sound i-f signal with the center frequency of 4.5 Mc, equal to the difference between the transmitted sound and picture carrier frequencies (see Fig. 22·24).

The frequency difference between the sound and picture carriers is standardized at 4.5 Mc for any channel and is held accurately within close tolerances at the transmitter. Therefore, intercarrier-sound receivers do not depend upon any precise local oscillator frequency for reception of the associated sound signal. For instance, suppose the local oscillator frequency drifts by 100 kc to produce 41.35 Mc for the sound i-f carrier

Fig. 22·24 Intercarrier-sound system.



output from the mixer. Then the picture i-f carrier will also change by 100 kc to 41.85 Mc. However, the difference between the two carriers is still 4.5 Mc. This is an important advantage over the split-sound system with a separate sound i-f amplifier section, where it is difficult to maintain the local oscillator frequency within the narrow limits required to heterodyne the r-f sound signal down exactly to the intermediate frequencies within the relatively narrow sound i-f pass band. The higher the oscillator frequency, the more difficult is the problem of drift. Therefore, the intercarrier-sound system is practically a necessity for receiving the sound signal for UHF television stations.

The 4.5-Mc output from the video detector can be considered the second sound i-f carrier frequency, to distinguish this from the i-f output of the mixer or first detector stage. With respect to the sound signal, the intercarrier receiver is a double superheterodyne, beating the transmitted r-f sound carrier signal down to the 4.5-Mc i-f sound signal in two steps. The first frequency conversion occurs in the mixer, which produces an i-f sound signal equal to the difference between the local oscillator and r-f sound carrier frequencies. This is then converted to the lower second sound i-f carrier frequency at 4.5 Mc in the video detector, with the picture carrier serving as a local oscillator signal that can beat with the sound signal.

In order to produce the 4.5-Mc sound signal, the first sound i-f carrier must have much less amplitude than the picture carrier in the input to the video detector. This is necessary because, when two signal voltages of slightly different frequencies are mixed, the resultant wave varies in amplitude at the difference frequency and has the original modulation of the weaker of the two input signals. The idea is the same as in any frequency converter where local oscillator output of several volts beats with much weaker r-f signal of a few millivolts or less. To preserve the modulation of the sound signal, therefore, the first sound i-f carrier is made much weaker than the i-f picture carrier amplitude. The resultant beat is obtained because of rectification of the combined waveform.

The 4.5-Mc sound signal is still an r-f FM signal with variations about the i-f center frequency of 4.5 Mc that correspond to the desired audio signal. The maximum frequency swing is ± 25 kc, as in the r-f and first i-f sound signals. When the 4.5-Mc sound signal is coupled into an FM detector, the audio voltage recovered in the detector output can then be coupled to an audio amplifier for driving the loudspeaker. The FM detector circuit must be aligned exactly at 4.5 Mc, as the 4.5-Mc sound intermediate frequency cannot be changed by the oscillator fine tuning control. Rejection of amplitude modulation in the 4.5-Mc signal is important to minimize AM distortion in the audio output.

22 · 17 Intercarrier buzz

Since the 4.5-Mc beat depends upon the picture carrier, excessive amplitude variations of the AM picture carrier signal can be transferred to the 4.5-Mc sound i-f signal by cross modulation in the video detector. This is especially evident with a predominantly white picture. Then the picture carrier amplitude varies between the extremes of the white level close to zero amplitude and the 75 per cent pedestal level for horizontal and vertical blanking pulses. As a result, the 4.5-Mc sound i-f signal has severe amplitude variations corresponding to the blanking and sync pulses. Without enough AM rejection, these extreme amplitude variations can produce a 15,750-cycle hiss in the audio output, corresponding to the horizontal blanking rate, and 60-cycle buzz at the vertical blanking rate. Since the 60-cycle buzz produced by voltage variations of the vertical blanking and sync pulses in the 4.5-Mc sound i-f signal is usually more evident, it is generally called *intercarrier buzz*.

Intercarrier buzz is minimized by (1) keeping the relative amplitude of the first sound i-f carrier very low to reduce amplitude modulation of the 4.5-Mc sound i-f signal by the picture carrier signal and (2) providing sufficient AM rejection in the 4.5-Mc sound i-f circuits, with the FM detector exactly aligned for balance at the 4.5-Mc center frequency.

Typical troubles that cause intercarrier buzz are incorrect balance in alignment of the FM detector, defective stabilizing capacitor in a ratio detector, and overload distortion in any of the i-f stages. The overload can be a result of weak amplifier tubes or insufficient bias.

22 · 18 Multiplexed stereo sound

Since stereophonic broadcasting with separate left (L) and right (R) audio signals has been approved for the commercial FM broadcast band, similar methods are being considered to multiplex two audio signals in the FM sound signal for television. The problem is a little different, however, because of the smaller deviation of 25 kc, compared with 75 kc. The general idea of multiplexing two signals on one carrier is indicated by the technique for color television, where the luminance signal and the chrominance modulation on a 3.58-Mc subcarrier are combined before modulating the picture carrier, as described in Chap. 24. For one suggested method of stereo sound in television,[1] the subcarrier frequency is 23.625 kc. The L + R signal, corresponding to monaural audio, has the frequency range of 50 to 15,000 cps. The L − R signal for the subcarrier includes the frequency range of 50 to 7,875 cps, which is enough to give the extra dimension of stereo sound.

SUMMARY

1. In FM, the *frequency* of the transmitted carrier varies with the *amplitude* of the audio modulating voltage. Center or rest frequency is the carrier frequency with zero modulation voltage. Deviation is the frequency change from center frequency, at the peak of the audio modulating voltage. Swing is the total deviation above and below center frequency, for positive and negative half cycles of audio modulating voltage.

[1] R. B. Dome. Television Stereophonic System, *IRE Trans. Broadcast Television Receivers*, July, 1962.

2. The FCC defines 100 per cent modulation as ± 25 kc swing for the FM sound in television broadcasting and ± 75 kc for commercial FM broadcast stations in the 88 to 108-Mc band.
3. A reactance-tube circuit provides either X_L or X_c that can be varied by its control-grid voltage. The reactance tube can then vary the frequency of a tuned oscillator.
4. Preemphasis in FM transmission increases the audio amplitude and the resulting carrier deviation to improve the signal-to-noise ratio for higher audio frequencies. At the receiver, deemphasis after the FM signal is detected reduces the amplitude of higher audio frequencies to provide the original audio-frequency response. The specified time constant is 75 μsec.
5. For slope detection of an FM signal, the i-f center frequency is at the side of the i-f response curve, to produce amplitude variations corresponding to the frequency variations. A diode detector can then recover these variations corresponding to the audio modulation signal. Slope detection is the reason why an AM receiver can receive FM signals.
6. A discriminator is a balanced FM detector using two diodes. The tuned input transformer changes the frequency variations of the i-f signal to amplitude variations that are detected to produce audio output signal. The response curve of a discriminator is an S curve, consisting of two i-f response curves of opposite polarity (see Fig. 22·9).
7. In the center-tuned or phase-shift discriminator both primary and secondary are tuned to center frequency (see Fig. 22·10).
8. A limiter stage is a saturated class C amplifier to reduce amplitude variations in the FM i-f signal.
9. The ratio detector uses a center-tuned input transformer with two diodes in series to charge the stabilizing capacitor (see Fig. 22·15). No limiter stage is necessary, as the ratio detector rejects AM interference.
10. In the quadrature-grid detector, the input i-f transformer and quadrature coil are tuned to center frequency (see Fig. 22·18). No limiter stage is necessary, as the detector rejects AM interference. The amplified audio output in the plate circuit is enough to drive the audio output stage.
11. In intercarrier-sound receivers, the difference frequency of 4.5 Mc is produced as a beat in the video detector output, with the modulation of the FM sound signal. The sound i-f amplifier and FM detector then are always aligned at the center frequency of 4.5 Mc.
12. Intercarrier buzz is produced by 60-cycle vertical blanking pulses from the picture signal in the 4.5-Mc intercarrier-sound signal. The buzz can be reduced by AM rejection circuits and exact alignment of the FM detector. In addition, the sound i-f response must be low in the common i-f amplifier.

SELF-EXAMINATION *(Answers at back of book.)*

Part A

Fill in the missing answer.

1. The assigned carrier frequency of an FM broadcast station is 96.3 Mc. Its FM signal has a rest frequency or center frequency of ______ Mc.
2. If 4 volts peak audio amplitude changes the r-f signal frequency from 200 to 210 kc the frequency deviation is ______ kc.
3. The total swing then for 8 volts peak-to-peak audio is ______ kc.
4. For 100 per cent modulation of the associated sound signal, the total frequency swing is ______ kc.
5. If 2 kc deviation is produced by 2 volts audio, then modulating voltage of 4 volts will produce a deviation of ______ kc.
6. If the audio modulation in Question 5 has the same amplitude but double the frequency, the deviation will be ______ kc.
7. The term quadrature indicates a phase angle of ______.
8. The time constant for preemphasis is ______ μsec.

9. The d-c output of a balanced discriminator for center frequency is ______ volts.
10. The frequency of intercarrier buzz is ______ cps.

Part B

Answer true or false.

1. In FM, the loudest sounds produce maximum frequency deviation.
2. In FM, the audio modulating frequency determines the rate of frequency swing.
3. A 2-Mc carrier swings ± 100 kc. Therefore, the instantaneous carrier frequency varies between 1.9 and 2.1 Mc.
4. A frequency swing of ± 5 kc corresponds to 5 per cent modulation.
5. Deemphasis of the detected signal provides the most attenuation for the highest audio frequencies.
6. A reactance-tube circuit can provide either X_L or X_c.
7. In a reactance-tube circuit, the grid feedback voltage is 90° out of phase with the plate voltage.
8. AM rejection circuits in the FM receiver reduce noise and interference.
9. Grid-leak bias clamps the positive peak of signal input at approximately zero grid voltage.
10. In a limiter stage, a typical time constant for grid-leak bias is 0.2 sec.
11. Slope detection enables an AM receiver to detect an FM signal.
12. In an intercarrier-sound receiver with a ratio detector, the primary and secondary are aligned at 4.5 Mc.
13. In the center-tuned discriminator, the primary and secondary voltages are 90° out of phase at center frequency.
14. In the quadrature-grid detector, the signal grid and quadrature grid are 90° out of phase at center frequency.
15. The audio output voltage of a quadrature-grid detector is usually 2 to 5 μv.
16. In a quadrature-grid detector, the audio output voltage is taken from the plate circuit.
17. The i-f stages in a ratio detector receiver can be aligned for maximum stabilizing voltage.
18. If the sound i-f response is too high in the common i-f amplifier, the result can be sound in the picture and buzz in the sound.
19. The sound take-off trap is always tuned to 4.5 Mc in intercarrier-sound receivers.
20. The S-response curve of a discriminator shows two opposite polarities of detected output voltage.

ESSAY QUESTIONS

1. Compare how the carrier wave varies with audio modulation for AM and FM.
2. Define center frequency, deviation, swing, and per cent modulation.
3. What characteristic of the FM signal determines loudness of the reproduced audio signal? What determines the pitch or tone of the reproduced sound?
4. Define phase modulation. Why does PM produce equivalent FM?
5. Give two reasons for the improved noise reduction in an FM system.
6. Give two advantages of FM over AM, and one disadvantage of FM, with both in the VHF band.
7. Why is deemphasis necessary in the FM receiver?
8. Draw the schematic diagram of a reactance-tube circuit using an R – L quadrature network that makes the reactance tube inductive. (*a*) Outline briefly how the circuit provides an electronic reactance. (*b*) Why is the reactance in this case inductive rather than capacitive? (*c*) What is the effect on the frequency of a tuned *LC* circuit shunted by the tube if its reactance is increased by less negative control-grid voltage?
9. When an FM signal has a modulation index 0.4 or less, why is its required bandwidth the same as an AM signal with the same audio modulating frequency, regardless of the amount of frequency swing? (See Bibliography at end of book for references in FM theory).

10. What are two important differences between an FM receiver and an AM receiver?
11. Give numerical values for the beat frequencies in the output of the mixer showing that the amount of frequency swing of a 41.25-Mc i-f signal in the receiver is the same as the FM sound signal transmitted in channel 2 (54 to 60 Mc).
12. Give three requirements for detecting an FM signal.
13. What is meant by slope detection?
14. Why is AM rejection an essential requirement of an FM receiver?
15. Draw the schematic diagram of a phase-shift discriminator circuit for an i-f center frequency of 41.25 Mc and briefly describe how the circuit operates. Label the audio take-off point.
16. Draw the schematic diagram of a balanced ratio detector circuit for an i-f center frequency of 4.5 Mc. Indicate the audio take-off point and the stabilizing voltage.
17. Draw the schematic diagram of a grid-leak bias limiter stage for an i-f carrier frequency of 41.25 Mc and explain briefly how limiting is accomplished.
18. In the quadrature-grid detector circuit, where is the i-f signal applied and where is the audio take-off point?
19. Give two advantages of the quadrature-grid detector, compared with the discriminator.
20. Describe how to align the quadrature-grid detector at 4.5 Mc.
21. How is the second sound i-f signal of 4.5 Mc produced in intercarrier-sound receivers?
22. Give two reasons why most television receivers use intercarrier sound. Give one disadvantage of intercarrier sound.
23. When aligning by the voltmeter method, what is the required indication when tuning the secondary in a phase-shift discriminator and in a ratio detector? Where is the meter connected in both cases?
24. Show the connections of the d-c voltmeter and balancing resistors for aligning the secondary in a single-ended ratio detector.
25. In a sound i-f circuit consisting of two i-f stages and a ratio detector, describe how to align the entire section by: (*a*) d-c voltmeter method with a signal generator. (*b*) visual response curve method. Show the i-f and detector curves, with marker frequencies for an i-f center frequency of 4.5 Mc.
26. In an i-f circuit consisting of two i-f stages, limiter, and center-tuned discriminator, describe how to align the entire section by: (*a*) d-c voltmeter method with a signal generator; (*b*) visual response curve method. Show curves and marker frequencies for an i-f center frequency of 10.7 Mc.
27. Explain how the FM sound signal can produce horizontal sound bars in the picture.
28. Referring to the schematic diagram in Fig. 22·20, give the function of each of the following components: R_{301}, C_{301}, R_{302}, C_{302}, T_{301}, R_{304}, R_{306}, C_{308}, R_{305}, C_{306}, R_{209}, L_{301}, R_{310}, with R_{311}, C_{401c}, and T_{303}.
29. In Fig. 22·20 give the effect for each of the following troubles: (*a*) R_{302} open; (*b*) R_{505} open; (*c*) C_{308} shorted; (*d*) T_{303} primary open.

PROBLEMS *(Answers to odd-numbered problems at back of book.)*

1. A 3-volt 1,000-cycle audio modulating voltage produces a frequency deviation of 15 kc. Assuming linear modulation, calculate the deviation for the following examples of audio modulation: (*a*) 1 volt, 1,000 cps; (*b*) 3 volts, 1,000 cps; (*c*) 6 volts, 500 cps; (*d*) 9 volts, 100 cps; (*e*) 0.3 volt, 10 kc.
2. Do the same as Prob. 1 for a modulation system where 3 volts audio produces 10 kc deviation.
3. For the *RC* deemphasis circuit in Fig. 22·6 calculate the exact frequency at which X_c equals R. How much is the attenuation in db at this frequency, compared with the response at 100 cps?
4. A deemphasis circuit has R of 75,000 ohms and C of 0.001 μf. Consider this a series a-c circuit with 10 volts rms audio signal applied. (*a*) Make a table listing the values of X_c, I, and E_c at 100 cps, 500 cps, 1 kc, 5 kc, 10 kc, and 15 kc. (*b*) Draw a graph showing the E_c values plotted against frequency.

5. If 4-Mc modulating voltage produces 100-kc frequency deviation, what bandwidth is required to include the upper and lower side frequencies?
6. In Fig. 22 · 18, calculate the inductance required for L_{301} to resonate with C_{309} plus 6 $\mu\mu$f capacitance of grid 3 to cathode. Also, calculate primary and secondary inductances of T_{301}, assuming 15 $\mu\mu$f plus 5 $\mu\mu$f capacitance on each side of the transformer.
7. Referring to the quadrature networks in Fig. 22 · 5, calculate the value required for L or C in *a*, *b*, *c*, and *d*, when the reactance tube is across an oscillator-tuned circuit operating at 1 Mc. Use 1,000 ohms for R in all cases and a factor of 10 or 0.1 for the ratio between reactance and resistance.
8. In Fig. 22 · 20 calculate the following average d-c values: (*a*) current through R_{302}; (*b*) cathode current of V_7, (*c*) cathode current of V_8, (*d*) plate current of V_8, (*e*) voltage across R_{310}. How much is the grid-cathode bias on V_7?
9. Referring to Fig. 22 · 20: (*a*) If Z_L for V_7 is 9,000 ohms at 4.5 Mc, and g_m is 4,000 μmhos, how much is the voltage amplification? (*b*) With 300 mv signal input to V_7, how much is the output in volts?

Chapter 23 Receiver servicing

The required work may include antenna installation, adjustment of setup controls, and troubleshooting defective circuits. Alignment is seldom necessary, compared with localizing troubles in tubes and components. When a defective component is replaced, the following points may be helpful. Keep the same positions for ground returns and connecting leads. This *lead dress* is often critical with high frequencies to prevent feedback, also with high voltages to prevent corona and arcing. When replacing a resistor, a higher wattage rating than the original can be used. With capacitors, a higher d-c voltage rating is permissible. However, do not replace a mica or ceramic capacitor with tubular capacitors as their inductance will alter the coupling or bypassing for high frequencies. Also, do not use too high a voltage rating with electrolytic capacitors; the normal forming voltage is needed for the rated capacitance. On printed-wiring boards, individual capacitors and resistors usually can be removed easily by clipping the leads close to the component. Then the new component is just soldered to the old leads, without disturbing the eyelet connections on the board.

23 · 1 Receiver adjustments

Figure 23 · 1 shows a typical chassis. Notice that the power input receptacle is open, as the safety interlock disconnects the power cord when the back panel is removed. Therefore a substitute line with a female plug, called a *cheater cord,* is needed to operate the receiver when making adjustments with the back panel off. Also, a mirror may be necessary to view the kinescope screen from the back of the set. If the kinescope uses an ion-trap magnet, adjust it for maximum brightness. However, most kinescopes have an aluminized screen and a straight gun, which does not require the ion-trap magnet.

Tilted raster. See Fig. 23 · 2*a*. This is corrected by rotating the yoke slightly. Usually, there is a wing nut to hold the yoke tight in its housing.

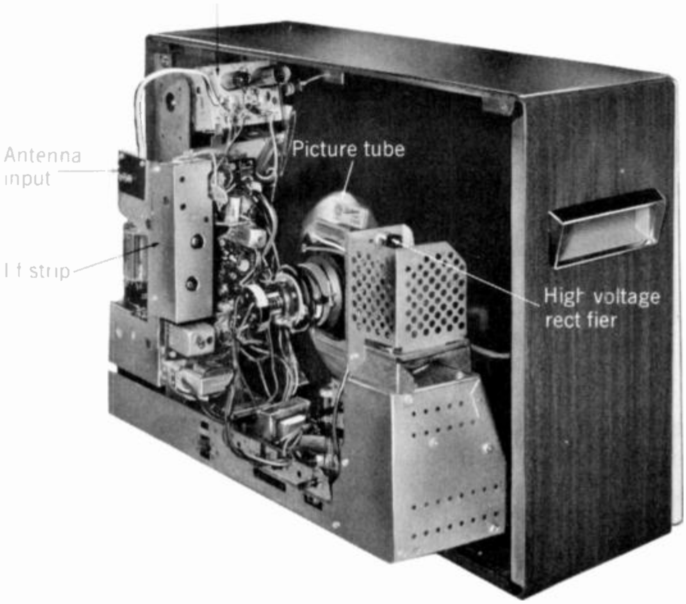


Fig. 23·1 Rear view of typical receiver chassis. Back cover off to show location of tubes and adjustments. Schematic diagram of this receiver is in Fig. 23·16. (GE chassis LW.)

Loosen the nut, turn the yoke for a straight raster, and tighten. The yoke should be as far forward as possible against the wide bell of the kinescope to scan the corners of the screen. Make sure the metal grounding spring attached to the yoke housing makes good contact to the Aquadag coating on the kinescope bell. The yoke housing in Fig. 23·1 also has pincushion magnets, which may need adjusting for straight edges at the top and bottom of the raster.

Centering. How the picture looks with the raster off center is shown in Fig. 23·2*b* and *c*. To center the raster on the screen vertically and horizontally, a kinescope with electrostatic focus generally uses two centering magnet rings (see Fig. 7·3). If a focusing magnet is used, it usually has a tab to adjust a magnetic centering ring (see Fig. 7·2). Also, the focus magnet can be tilted slightly to adjust centering. In some receivers, electrical centering controls are on the back of the chassis to vary the amount of direct current through the deflection coils.

Focusing. With many electrostatic-focus kinescopes there are no adjustments for focusing. Where the focus-grid voltage can be adjusted, this may be provided by (1) moving a wire jumper between pins on the kinescope base or (2) moving a plug or a potentiometer to change the focusing voltage. These adjustments are usually not critical. Where magnetic focusing is used, the position of the focus magnet is varied and the field strength is adjusted by a ring on the magnet (Fig. 7·2). With an electromagnetic focus coil, the amount of direct current can be varied by a control on the

Fig. 23·2 (a) Tilted raster. (b) Raster off center horizontally and (c) vertically. (RCA.)

(*a*)

back of the chassis. Adjust for sharp scanning lines and then check for sharp focus of the fine details in a picture.

Raster height. The vertical linearity and height controls are on the chassis. Adjust both to fill the screen top to bottom with uniformly spaced scanning lines. Usually, the linearity control affects the top more and the height control stretches the bottom. If the controls cannot be adjusted for full height with good linearity, this indicates the trouble of weak vertical output.

Raster width. There may not be any controls to adjust width, or just one of the following. A horizontal drive control is set for maximum width, without any white drive bar near the center. A width control is adjusted for the desired width. The width adjustment may be a movable sleeve in the yoke on the kinescope neck. A horizontal linearity control should be adjusted for minimum average direct current in the cathode circuit of the horizontal output tube, with good linearity in the center of the raster. Insufficient width or crowding at either side indicates a weak horizontal output or low B+ voltage. Too much width may cause excessive power dissipation in the horizontal output tube.

Horizontal lock. Usually, there is just a slug to adjust the stabilizing coil. With the horizontal hold control set to the center of its rotation, adjust the slug to the position where the picture locks in and stays in sync when changing channels, even for weak signals. If another type of AFC is used, see Sec. 16·6 for the Synchro-Lock circuit or Sec. 16·5 for the Synchro-Guide circuit.

AGC control. Set for maximum contrast, without overload, on the strongest station.

23·2 Types of ghosts

Reception of multipath signals by the antenna is the most common cause of ghosts. However, there are other possibilities.

Built-in ghosts. Multiple images produced by the receiver or by the transmission line are generally called *built-in ghosts.* These can be recognized by the fact that usually there are three, four, or more images uniformly spaced to the right. Excessive response for the high video frequencies

(b)

(c)

in the i-f amplifier or video amplifier can cause built-in ghosts on all channels. If the excessive response is in the picture i-f amplifier, varying the fine tuning will affect the ghosts. Ghosts caused by reflections of the antenna signal in a long transmission line change when hand capacitance is added by holding the line, and are not the same on all channels.

Leading and trailing ghosts. A ghost due to multipath reception by the antenna is usually to the right of the main image, producing a *trailing ghost.* However, a *leading ghost* can be produced to the left of the main image. One cause of leading ghosts is direct pickup of signal, especially in strong signal areas. In this case, the direct pickup can provide a picture before the antenna signal delivered by a long transmission line. With direct pickup, the ghost will vary when people walk near the receiver.

Minimizing ghosts. For the problem of multipath reception, an antenna that has a good front-to-back ratio and a narrow forward lobe, with minimum side responses, can be oriented carefully to minimize the ghost. Sometimes, changing the antenna location reduces the intensity of the ghost. Or trying vertical polarization, especially with indoor antennas, may help. Separate antennas cut for the individual channels may be necessary.

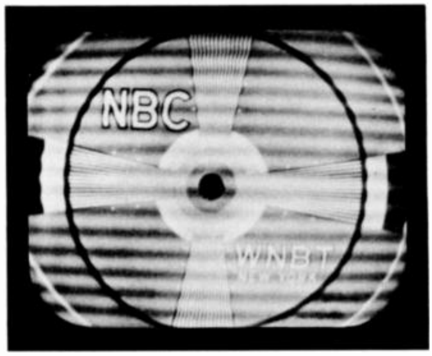
A leading ghost, to the left of the image, is a problem of reducing stray pickup. Try shielding the transmission line and r-f tuner, if necessary, and supplying more signal from the antenna. The antenna system and the r-f amplifier can be checked for the possibility of a trouble that causes weak antenna signal. With an antenna distribution system, it must supply much more antenna signal than the direct pickup, at each receiver outlet.

23 · 3 R-F interference

Interfering r-f signals in the television receiver are heterodyned to produce video frequencies in the output of the second detector that cause interference patterns on the kinescope screen. Typical examples are shown in Fig. 23 · 3.

Diagonal lines. The uniform diagonal bars in *a* are caused by an unmodulated carrier wave (c-w) signal. The bars are of uniform thickness because there is no modulation of the c-w interference. Usually, the bars shift slowly from one diagonal position, through the vertical, and then to

(a)



(b)

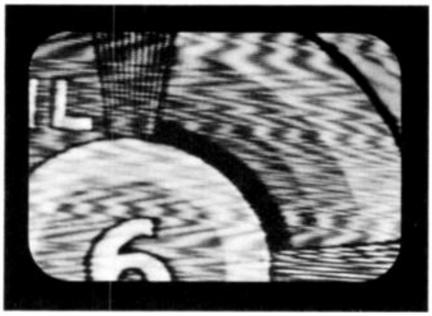
Fig. 23·3 Interference effects in the picture. (a) Uniform diagonal bars caused by beat of unmodulated r-f wave. (b) Horizontal bars caused by audio amplitude modulation. (c) Herringbone weave caused by frequency modulation. (d) Diathermy interference.

the opposite diagonal, as the interfering carrier wave drifts in frequency. This type of interference can be caused by local oscillator radiation from a nearby receiver. The number of bars and their thickness depend on the beat frequency produced by the c-w interference. If the beat frequency resulting from the r-f interference is less than 15,750 cps, it will produce uniform horizontal bars, in a venetian-blind effect.

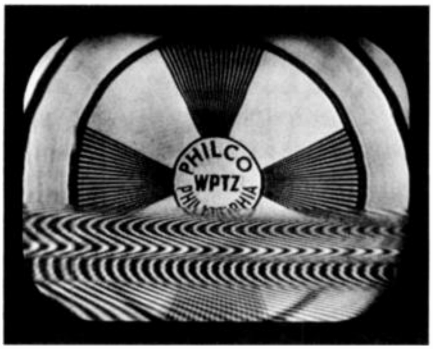
Horizontal bars. An interfering r-f carrier wave that varies in amplitude with audio modulation produces in the second detector output audio interference in the video signal. This results in a horizontal-bar pattern, as illustrated in Fig. 23·3*b*. The number, width, and intensity of the bars will vary with the audio modulation.

Herringbone weave. An interfering FM carrier wave produces in the second detector output interfering beats that vary in frequency. With a center frequency high enough to produce a fine-line interference pattern, the frequency modulation adds a herringbone weave to the vertical or diagonal bars. An example of this effect is the 4.5-Mc FM sound signal in the picture shown in Fig. 23·5. When the beat frequencies are too low for a diagonal-line pattern, instead of horizontal bars the FM interference produces a watery effect through the entire picture, which then looks as if it were covered with a shimmering silk gauze (see Fig. 23·3*c*).

Diathermy interference. Diathermy machines and other medical or industrial equipment usually produce the r-f interference pattern shown in Fig. 23·3*d*. There may be two dark bands across the screen instead of the one shown, and the bars will be darker if the interference is stronger. This r-f interference pattern is produced because diathermy equipment is effectively a transmitter. The diagonal lines at the bottom indicate c-w interference, and the waving effect is caused by frequency variations. The single band shows a strong 60-cps component in the interference; with 120-cps modulation, there would be two bands.



(*c*)



(*d*)

The frequency ranges of r-f equipment for industrial, scientific, and medical use are 13.553 to 13.566 Mc, 26.957 to 27.282 Mc, and 40.659 to 40.700 Mc. Since these frequencies are below channel 2, diathermy interference that enters the receiver through the antenna can be eliminated by inserting a high-pass filter in series with the transmission line at the receiver input. The front end in most receivers has a high-pass filter cutting off below 50 Mc but an external filter can be used if necessary.

Strong r-f interference. In addition to bar patterns in the picture, the light values are altered by the r-f interference when it has enough amplitude to raise the white level of the picture carrier signal closer to the black level. A very strong interfering signal, therefore, can produce a negative picture or black out the picture completely.

How interference enters the receiver. An interfering r-f signal can enter the receiver through the antenna and transmission line, direct pickup by the chassis, or from the power line. R-f interference at the antenna input of the receiver must go through the r-f tuner and, therefore, usually appears on specific channels. Direct pickup by the chassis of i-f interference results in the same interference on all channels. If the r-f interference is from the power line, reversing the plug and adding hand capacitance by holding the line should affect the interference. This can be reduced by an r-f filter in the power line.

FM broadcast interference. An FM interference pattern, as illustrated in Fig. 23 · 3*c*, only on channel 2 is usually caused by an FM broadcast station in the 88- to 108-Mc band. You may also hear the audio modulation. An FM trap in the antenna input circuit can be tuned to reject the interfering frequency and eliminate the interference.

Sound i-f harmonics. Harmonics of the associated sound i-f signal in the receiver can cause an FM interference pattern. Sound i-f harmonic interference in the picture can be identified by noting that it varies with the audio modulation of the associated sound signal and will disappear if a sound i-f tube is removed.

Picture i-f harmonics. Harmonics of the picture i-f carrier signal, usually generated in the video detector, also can be coupled back to the front end and cause interference in the picture. This may appear as diagonal lines, depending on the beat frequency, with a grainy effect due to amplitude modulation by the high-frequency components of the video signal. The frequency of the beat interference will vary widely with a slight change in the oscillator fine tuning control.

Co-channel interference. Stations broadcasting on the same channel are separated by 150 miles or more, but in fringe-area locations between cities co-channel stations can interfere with each other when the ratio of the desired signal to the interference is less than 45 db, approximately. If the interfering signal is strong enough, its picture will be superimposed on the desired picture. In addition, there is usually a bar pattern resulting from the beat between the two picture carrier frequencies. The beat is an audio frequency that produces a horizontal bar pattern, generally called *venetian-blind* effect, which is similar to sound bars in the picture. The remedy for co-channel interference is a more directional antenna, especially with respect to the front-to-back ratio, as interfering co-channel stations are often in opposite directions.

Adjacent channel interference. When the picture signal of the upper adjacent channel is strong enough, the side bands corresponding to low video frequencies can beat with the desired picture carrier, producing picture information of the interfering station superimposed on the desired picture. Most noticeable is the vertical bar produced by horizontal blanking, as it usually drifts from side to side because of the slight difference in horizontal phasing between the two signals. This is generally called *windshield-wiper* effect. The remedy is a more directional antenna, but in addition the i-f selectivity can be improved by wave traps tuned to reject the adjacent channel intermediate frequencies because they are outside the required i-f pass band.

Television receiver interference in radio receivers. Harmonics of the 15,750-cps horizontal deflection current in the television receiver can cause beat-frequency whistles in nearby radio receivers. The whistles appear every 15 kc with maximum points every 70 to 100 kc, approximately, on the dial of the radio receiver. This interference can be reduced by inserting an r-f filter in series with the power cord of the television receiver.

23 · 4 External noise interference in the picture

Noise pulse voltages from automobile ignition systems and motors produce short horizontal black streaks in the picture, as illustrated in Fig. 23 · 4. The picture may also skip frames vertically and tear apart horizontally if the noise is strong enough to interfere with synchronization. Similar effects can be caused by elevator motors, neon signs, cash registers,

Fig. 23 · 4 Short horizontal black streaks caused by auto-ignition interference.

or any device that produces sparking, which generates r-f energy in the VHF range modulated at the sparking rate. However, a device operated from the 60-cps power line produces noise streaks in a cluster that stays still like a horizontal hum bar, while ignition noise streaks occur at random throughout the picture.

External noise interference can enter the receiver as pickup by the antenna or by an unshielded transmission line, direct pickup by the chassis, or through the a-c power line. Minimizing the interference effect in the picture is a problem of supplying more antenna signal or reducing the noise pickup, or doing both, to provide a suitable signal-to-noise ratio. This may be accomplished by using a high-gain directive antenna, with vertical stacking. Moving the antenna out of the noise field, by either increasing the antenna height or just finding an antenna placement farther from the noise source, is often helpful. It may be necessary to use shielded transmission line to prevent noise pickup by the line. To reduce pickup from the power line, a low-pass filter can be used. This unit consists of 1-mh r-f chokes in series in each side of the line cord and 0.01-μf bypass capacitors across the line. Connect the filter at a point closest to the interference source in order to minimize radiation from the power line.

23 · 5 Sound in the picture

The associated sound signal can produce horizontal sound bars corresponding to the audio modulation in the picture, as shown in Fig. 23 · 3*b*, or the fine beat pattern in Fig. 23 · 5 caused by 4.5-Mc intercarrier beat.

Sound bars. Slope detection converts the modulation of the FM sound signal to audio voltage in the video detector output, which is coupled to the kinescope grid-cathode circuit to produce the horizontal sound bars. The sound bars can be recognized as they vary with the audio modulation and disappear when there is no voice. The cause of sound bars is incorrect response for the i-f sound signal in the picture i-f amplifier. This can often be corrected by tuning the associated sound traps in the i-f amplifier to

Fig. 23 · 5 Fine herringbone-weave or wormy effect of 4.5-Mc beat in picture. (RCA.)

eliminate the sound bars. When the sound bars are present only at high volume levels, this indicates microphonics caused by the vibrating loudspeaker. The microphonics may be in the local oscillator, r-f, i-f, or video amplifier. Sound bars only at high volume levels can also be the result of insufficient filtering of the plate-supply voltage in the audio output stage.

4.5-Mc beat. The 4.5-Mc output of the video detector has the FM sound signal variations. Coupled to the kinescope grid-cathode circuit, the 4.5-Mc signal produces a fine beat pattern in the picture consisting of about 225 pairs of thin black-and-white lines with small "wiggles" like a fine herringbone weave. This effect in Fig. 23 · 5 is also called a "wormy" picture. The fine lines are caused by the 4.5-Mc carrier, while the herringbone weave is the result of the frequency variations in the FM sound signal. The 4.5-Mc beat produced by the associated sound signal can be recognized by observing the pattern closely to see that the herringbone effect disappears when there is no voice, leaving just the straight-line pattern corresponding to the c-w interference of the 4.5-Mc carrier without modulation.

Excessive 4.5-Mc sound signal at the kinescope is the result of insufficient rejection in the video amplifier circuits. The 4.5-Mc traps in the video amplifier can be tuned for minimum 4.5-Mc signal at the kinescope grid-cathode circuit and minimum beat interference in the picture. If necessary, a 4.5-Mc trap can be added to the video amplifier, as illustrated in Fig. 23 · 6. The parallel resonant trap circuit is inserted in series with C_c and L_c, on either side. Adjust the trap for minimum 4.5-Mc signal at the kinescope grid, and maximum sound signal if it is also the sound take-off circuit.

23 · 6 Localizing hum troubles

Table 23 · 1 indicates how different symptoms of hum on the kinescope screen can be distinguished to localize the trouble. Hum in the picture means the symptom is present only when a picture is on the screen but disappears with just a raster. The picture can be removed by switching to

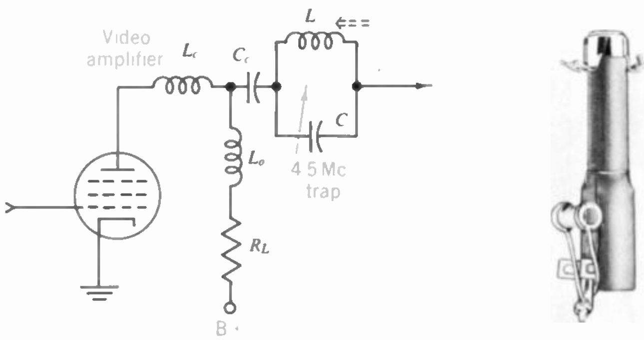


Fig. 23 · 6 (a) Parallel-resonant 4.5-Mc trap in plate circuit of video amplifier. (b) Photograph of typical trap. (RCA.)

an unused channel or shorting the antenna terminals to see the raster alone. In order to see whether bend is in the picture or raster, shift the raster off center horizontally and increase the brightness to see the side of the raster and edge of the picture where horizontal blanking begins. Effects that are only in the picture result from modulation hum, which requires signal input for amplification of the hum voltage. Additive hum is in the raster and picture. A special case is hum in the vertical sync or oscillator circuit, which makes the picture lock in the wrong position (see Fig. 14 · 1*a*).

It should be noted that 60-cycle hum produces one pair of hum bars or one sine-wave bend from top to bottom on the screen, while 120-cycle hum results in two pairs of bars or two cycles of bend. Heater-to-cathode leakage in a tube introduces 60-cycle hum but 120-cycle hum is caused by excessive ripple in the B supply voltage from a full-wave power supply.

Figure 23 · 7 shows 120-cycle hum bars and bend in the picture. There are two pairs of dark and light bars across the screen, with two cycles of bend from top to bottom, which indicate hum at 120 cps. The bars show hum in the signal. If there is more hum voltage, the bars will be darker. The bend shows hum in the horizontal sync, resulting from hum injected into the signal. If the bars are present without a picture, this must be additive hum injected in the video section, which is evident with or without signal. If the bars are present only with a picture, this shows the hum is injected in the r-f or i-f stages, which require signal input for hum modulation.

Table 23 · 1 Localizing hum troubles

Symptom	*Cause*	*Source*
Hum bend in picture but not in raster	Hum in H sync	AFC circuit, if no hum bars; r-f, i-f, and video circuits, if hum bars also
Hum bend in both raster and picture	Hum in H deflection	H oscillator, amplifier, or damper
Hum bars in picture but not in raster	Hum in picture signal	R-f and i-f circuits; can have bend also
Hum bars in both raster and picture	Hum in video signal	Video amplifier; can have bend also

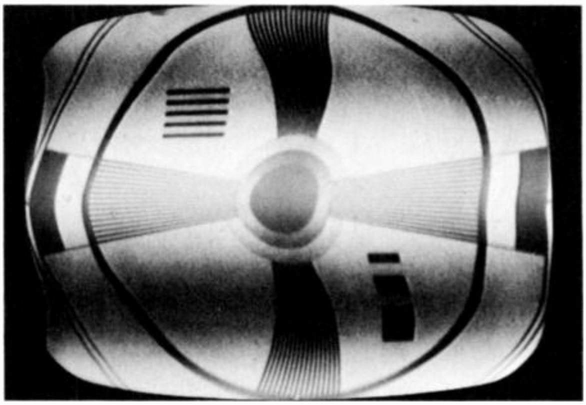


Fig. 23·7 120-cycle hum bars and bend in picture. (RCA Institutes Home Study School.)

23·7 Testing scanning linearity with bar patterns

Linearity of the vertical and horizontal scanning can be checked by producing bar patterns on the kinescope screen, with a variable-frequency audio signal generator. This method does not require any transmitted picture signal. Instead, the output from the signal generator is coupled into the video amplifier to supply signal voltage for the kinescope grid-cathode circuit. As the generator signal varies the kinescope grid voltage, while the deflection circuits are producing scanning, pairs of dark and light bars are formed on the raster (see Fig. 23·8).

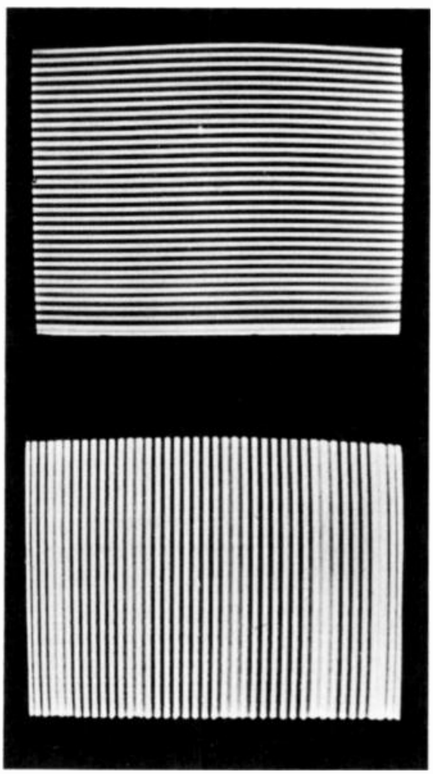
Horizontal bars are produced when the frequency of the signal at the kinescope grid is less than 15,750 cps; above 15,750 cps the bars are vertical or diagonal. Since the synchronizing voltage for the deflection oscillators is usually taken from the video amplifier in the receiver, the generator signal also provides synchronization. The synchronization can hold when the synchronizing frequency is an exact multiple of the scanning frequency. Just vary the signal generator frequency to obtain the desired number of bars and adjust the receiver hold control to make the bars stay still. A signal generator made specifically for this job is called a *linearity checker;* a *crosshatch generator* produces both vertical and horizontal bars at the same time.

Suppose that a 60-cps sine-wave signal is varying the kinescope control-grid voltage in synchronism with the vertical scanning motion at the field frequency of 60 cps. During the positive half cycle the signal makes the grid more positive, increasing the beam current and screen illumination; the negative half cycle reduces the beam current and screen illumination. Since it takes $\frac{1}{120}$ sec for a half cycle of the 60-cps signal, the scanning beam moves approximately halfway down the screen during this time. Therefore, if the positive half cycle of the sine-wave signal coincides with the first half of the vertical scan, the top half of the picture will be brighter than the bottom half. The pattern on the screen then is a pair of horizontal bars, one bright and the other dark.

When the signal generator output frequency is increased to multiples of 60 cps, additional pairs of narrower horizontal bars will be formed on the screen, as shown in Fig. 23·8*a*. The number of pairs of bars is equal to the signal generator frequency divided by the vertical scanning frequency, minus any bars that may be produced during vertical retrace time if the signal frequency is high enough to produce more than about 20 pairs of bars. As an example, a frequency of 240 cps results in four pairs of horizontal bars when the vertical scanning frequency is 60 cps. With an audio-frequency output of 400 cps from the signal generator, six pairs of bars can be obtained by adjusting the vertical hold control to the vertical scanning frequency of $^{400}/_{6}$ cps.

Vertical scanning linearity is indicated by the spacing between the parallel horizontal bars. If the vertical scanning motion is linear, the bars will be equally spaced. Otherwise, the bars will be spread out or crowded together. Adjustments can then be made with the vertical linearity and height controls to obtain the most uniform distribution of the bars.

When the frequency of the modulating signal becomes equal to the horizontal line-scanning frequency, vertical bars are formed instead of the horizontal bars. Consider the case of a 15,750-cps sine-wave signal varying the kinescope grid voltage in phase with the horizontal scanning. During one horizontal line the screen is made brighter for approximately one-half the picture width, as the positive half cycle of the grid voltage increases the beam current; the negative half cycle makes the screen darker. The same effect occurs for succeeding horizontal lines and the result is a pair of vertical bars on the screen, one bar white and the other dark. If the frequency of the signal generator is increased to multiples of 15,750 cps, additional pairs of narrower vertical bars will be produced on the screen, as shown in Fig. 23·8*b*. Their spacing indicates linearity of the horizontal scanning motion.



(*a*)

(*b*)

Fig. 23·8 Bar patterns on kinescope screen produced by signal generator. (a) Horizontal bars to check vertical scanning linearity. (b) Vertical bars to check horizontal scanning linearity.

The number of pairs of vertical bars is equal to the frequency of the applied grid signal divided by the horizontal scanning frequency. As an example, a frequency of 157.5 kc results in 10 pairs of vertical bars when the horizontal scanning frequency is 15,750 cps. However, all the bars may not be visible. Some are formed during the horizontal retrace time, when the signal frequency is high enough to produce more than about 10 pairs of bars. These bars formed during the flyback are wider because of the fast retrace and appear as variations of shading in the background, as can be seen in Fig. 23·8*b*. It is possible to determine the retrace time by counting the bars visible during the trace time and comparing this with the total that should be produced.

Diagonal bars are produced when the frequency of the grid modulating voltage is higher than the horizontal line-scanning frequency but is not an exact multiple. In this case the light and dark parts of each line are regularly displaced in successive order at different positions with respect to the start of the trace, instead of lining up one under the other. The diagonal bars usually do not stay still because the signal frequency is not an exact multiple of the horizontal scanning frequency and does not synchronize the deflection oscillator.

23·8 Signal injection

In this method, your own test signal is injected to see the effect on the kinescope screen. Figure 23·9 illustrates how a conventional signal generator can be used. Each amplifier stage can be checked, working back from the kinescope toward the antenna input, to localize a defective stage that cannot pass the signal. The 400-cps audio output of the signal generator is coupled into the video amplifier circuit to see if the signal is amplified to produce horizontal bars on the kinescope screen. In the i-f amplifier the modulated r-f test signal is injected at the i-f picture carrier frequency. If the i-f and video sections are operating, the amplitude-modulated test signal will be detected to provide 400-cps video signal that produces bars. In the r-f section, the modulated r-f test signal at the r-f picture carrier frequency of the selected channel can be injected into the mixer grid circuit and antenna input. If the local oscillator stage, i-f sec-

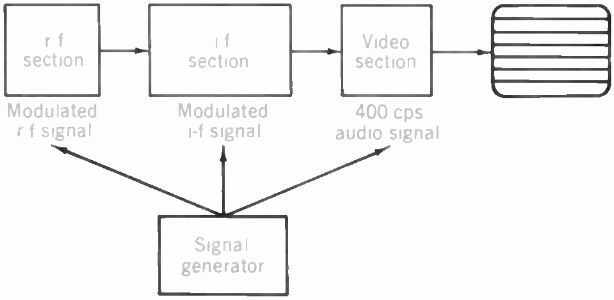


Fig. 23·9 Using generator to inject test signal in picture circuits.

tion, and video amplifier are operating, the 400-cps modulation will produce horizontal bars. The test signal usually cannot be substituted for the local oscillator because of insufficient r-f output from the signal generator.

In the vertical deflection circuits, 60-cps filament voltage can be injected into the grid circuit of the vertical amplifier to check operation of the output stage when there is no vertical scanning. Vertical deflection with the injected voltage indicates the output stage is operating and, therefore, the trouble is no deflection voltage from the vertical oscillator.

To test the horizontal deflection circuits, the horizontal oscillator output from an operating receiver can be injected at the grid of the horizontal output tube in a receiver with trouble. Also, the d-c high voltage for the kinescope anode can be "borrowed" from an operating receiver. Some types of servicing test equipment, called *analyzers,* provide video and deflection voltages for signal injection in all the receiver circuits.

23·9 Localizing receiver troubles

The key to quick efficient trouble shooting in the television receiver is the ability to use the raster, picture, and sound as indicators to localize the trouble to a particular section. This procedure is illustrated by the trouble-shooting charts in Tables 23·2 to 23·5, which summarize the most common causes of typical troubles. The analysis is based on a typical receiver circuit using intercarrier sound with the 4.5-Mc sound take-off circuit in the video detector output.

Parallel heaters are assumed. With series heaters, an open in one makes the entire string open. The tubes for parallel heaters generally have 6- or 12-volt heater ratings, as in the 6AQ5 or 12AU7 with two parallel heater sections. Tubes for a series string have heater ratings of 3, 5, 19 volts, etc.

Table 23·2 illustrates troubles that affect more than just one indicator. As examples, the low-voltage power supply can produce troubles in the

Table 23·2 Receiver troubles

TROUBLE INDICATION	ANALYSIS
Set completely dead. No raster, picture, or sound	Check power line into receiver, safety interlock switch, and power-line fuse if any. An open heater in series string. If heaters light, trouble is no B+
No sound and no picture but raster is normal	Check AGC, r-f tuner, common i-f amplifier, video detector, and audio output tube in stacked B+ circuit
Trouble on one channel, but operation normal on other channels	Check front end. Incorrect local oscillator frequency. Dirty contacts on station-selector switch. Weak antenna signal on one channel

raster and signal; the circuits common to the picture and sound signals affect both the picture and sound.

Table 23 · 3 lists common raster troubles. These are also evident in the picture, but the fact that the trouble is in the raster circuits can be determined by removing the picture to see the trouble symptom in the raster alone. This table also includes troubles evident in the picture but caused by a defect in the raster circuits. For instance, crowding and spreading in the picture are the result of nonlinear scanning.

Table 23 · 3 Raster troubles

TROUBLE INDICATION	ANALYSIS
No brightness. Sound is normal	No high voltage; check high-voltage circuits and fuse. Check kinescope voltages; ion-trap magnet
Insufficient brightness	Insufficient high voltage. Weak kinescope
Raster too small	Low B+
Raster too narrow. Insufficient width	Weak horizontal output; check oscillator, amplifier, and damper stages. Low B+
Raster too small vertically. Insufficient height	Weak vertical output; check oscillator and output stages
Only a bright horizontal line on screen	No vertical deflection; check vertical oscillator and output stages
Raster tilted	Rotate deflection yoke
Trapezoidal raster	Defective deflection yoke
Pincushion or barrel distortion of raster	Defective deflection yoke or misadjustment of correction magnets
Shadowed corner in raster	Check focusing magnet and ion-trap magnet
Blooming with defocusing at high brightness levels	Poor high-voltage regulation; weak high-voltage rectifier
Wide bright bar at left side of raster, with fold-over and reduced width	Insufficient damping in horizontal output circuit; check damper tube and circuit
White vertical bar at left or right side of raster	Crowded nonlinear horizontal scanning; check horizontal amplifier and damper circuits
White vertical bar near middle of raster	Excessive horizontal drive; readjust drive control; check input voltage and bias on horizontal amplifier
White horizontal bar at top or bottom of raster	Crowded nonlinear vertical scanning; check height and linearity controls; trouble in vertical oscillator or amplifier circuit

TROUBLE INDICATION	ANALYSIS
Crowding and spreading of picture at left and right sides of raster	Nonlinear horizontal scanning; check width, drive, and linearity controls. Weak horizontal oscillator, amplifier, or damper
Crowding and spreading of picture at top and bottom of raster	Nonlinear vertical scanning; check height and linearity controls; weak vertical oscillator or amplifier
Reversed direction of scanning, right to left or bottom to top	Reverse connections to horizontal or vertical deflection coils

Table 23·4 indicates common types of trouble that are in the picture but not in the raster or in the sound. Therefore, the trouble is in a circuit for the picture alone.

Sync troubles are considered as troubles in the picture, rather than the raster, when the picture can be made to stay still temporarily by varying

Table 23·4 Picture troubles (raster normal)

TROUBLE INDICATION	ANALYSIS
No picture. Normal raster and normal sound	Trouble in circuit only for picture; video amplifier in most receivers
Weak picture with low contrast; no snow	Insufficient video signal; low gain in i-f and video amplifiers; weak kinescope
Snowy picture	Insufficient antenna signal; weak r-f amplifier
Very dark or reversed picture, out of sync	Overloaded picture; excessive antenna signal, insufficient grid bias, or weak tubes; check AGC bias circuit
Picture intensity fades in and out	Check for loose transmission line that flaps in the wind
Streaks in picture with antenna disconnected	Arcing in high-voltage supply. Intermittent coupling capacitor in signal circuit
Picture rolls vertically and slips horizontally	Weak sync; check common sync circuits; check video signal for normal sync
Picture rolls vertically. Can be stopped but does not hold	Weak vertical sync; check vertical sync circuits; check video signal for normal sync
Picture tears horizontally. Can be stopped but does not hold	Trouble in horizontal AFC circuit
Bend in picture but not in raster	Hum in horizontal sync, or weak horizontal sync
Picture locks in vertically at wrong position	Hum in vertical sync or weak vertical sync
One or two pairs of wide black-and-white bars across the picture	Hum in the video signal

the deflection oscillator frequency. This indicates the oscillator operates normally but there is trouble in the synchronizing pulses obtained from the composite video signal. If the picture does not hold at any setting of the frequency control, the trouble is incorrect frequency of the deflection oscillator, which is a defect in the raster circuits.

Troubles in the sound alone must be in the 4.5-Mc second sound i-f circuits or the audio section, as indicated in Table 23 · 5, when the trouble is the same on every channel.

Tubes are the most common source of trouble and should be checked first, preferably by substitution of a tube known to be good or substituting the suspected tube in a chassis that is operating normally. When the same tube type is used in different sections of the receiver, switching the tubes will often indicate whether the suspected tube is bad. In the local oscillator stage, it may be necessary to try several tubes because one tube may not work well where another of the same type will. This may also happen in the horizontal oscillator control stage. Microphonic or noisy tubes can be checked by tapping them lightly and noting any indication in the picture and sound. The power-supply rectifiers and the output amplifier stages are common sources of trouble, since the components subject to high voltages or high currents are more likely to fail.

Fuses. Most receivers have one or more fuses on the top or back of the chassis, in the high-voltage cage, or possibly on the underside of the chassis. An open fuse in the horizontal output circuit results in no brightness, with normal sound. An open fuse or circuit breaker in the B+ line causes no sound, picture, or raster but the heaters light. With series heaters, there is usually a wire fuse just for the heater line. An open fuse in the input power line means no input power; there is no B+ and the heaters do not light.

Multiple troubles. The low-voltage power supply affects the raster, picture, and sound. With series heaters, particularly, one open heater or an open fuse in the heater line means none of the tubes can light. Even with the tubes on, if there is no B+ from the power supply the result is no raster, picture, or sound.

The AGC circuit is a common cause of no picture and no sound, while

Table 23 · 5 Sound troubles (picture normal)

TROUBLE INDICATION	ANALYSIS
No sound, or weak sound, on all channels	Check 4.5-Mc sound i-f section and audio amplifier
Distorted sound	Check FM detector alignment and audio amplifier bias
Buzz in sound	Check AM rejection circuit and FM detector alignment for balance

the raster is normal. Also, a defective audio output stage can cause no picture and no sound, when a stacked B+ circuit is used. Finally, the i-f stages affect both picture and sound in intercarrier receivers.

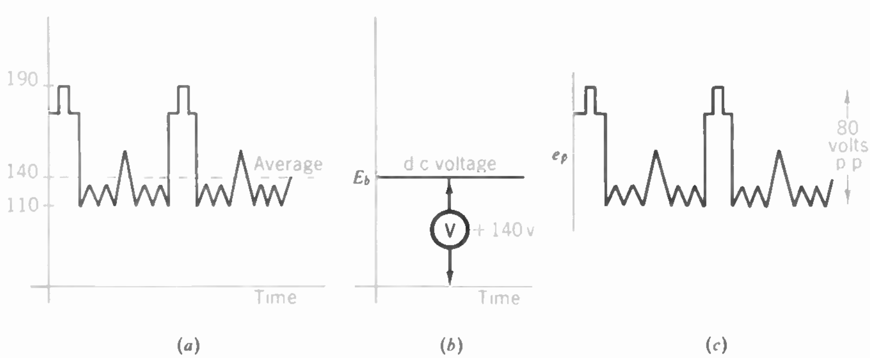
Checking high voltage. The usual cause of no brightness with normal sound is no high voltage for the kinescope anode. To check, you can arc the high-voltage lead to the anode connector. This lead from the rectifier filament is the d-c high-voltage output. Then check for an arc at the plate of the high-voltage rectifier for high-voltage a-c input. If there is a-c input but no d-c output, the trouble is in the high-voltage rectifier tube or circuit. If there is no a-c input, the trouble is the horizontal deflection circuits. Check the output stage, damper, and oscillator.

23 · 10 D-C voltage measurements

The manufacturer's schematic diagram usually gives normal d-c voltages measured with voltmeter of 20,000 ohms per volt sensitivity or a VTVM. In Fig. 23 · 16, as an example, the video output stage has a normal d-c plate voltage of 120 volts, with 130 volts at the screen grid and 2.6 volts cathode bias. These are d-c voltages measured to chassis ground or B−. They are not a-c signal voltages. In fact, the specified d-c voltages are usually measured with no signal input. About 10 to 20 per cent tolerance in the measured d-c voltages is usually permissible.

As described in the next section, the a-c signal voltages are usually given as oscilloscope waveforms, in peak-to-peak values. The oscilloscope waveforms include video signal, sync, and deflection voltages. The actual signal voltage in an amplifier is a pulsating d-c voltage, with a-c variations from an average d-c axis. The d-c voltmeter measures the average level. The oscilloscope shows the signal variations. This combination of d-c and a-c voltages is illustrated in Fig. 23 · 10 for video signal. If the d-c voltages are far from normal, the a-c signal voltages usually will not be correct. This

Fig. 23 · 10 (a) Pulsating d-c voltage e_b at plate of video amplifier. (b) Average d-c plate voltage E_b of +140 volts measured by d-c voltmeter. (c) A-c signal voltage e_p of 80 volts peak to peak as observed on oscilloscope screen.



is why d-c voltage measurements can be useful for localizing troubles in the a-c signal.

Plate voltage. The two extreme cases are full B+ or zero voltage at the plate. Either trouble can prevent signal output but for opposite reasons. Zero plate voltage means the d-c path for plate current to the B+ supply is open. Look for an open plate load or open decoupling resistor. However, full B+ at the plate means the plate circuit is not open. It should be noted that in tuned amplifiers, such as an i-f stage, there may not be any resistance to drop the B+ voltage for the plate. Then full B+ voltage at the plate is normal.

With about 1,000 ohms or more in the plate circuit, though, full B+ at the plate means no plate current. The plate voltage is connected but without plate current there is no *IR* voltage drop. Stages with a resistance plate load include the video amplifier, sync separator or amplifier, first audio amplifier, and multivibrator deflection oscillators.

Common causes of zero plate current with full B+ at the plate are (1) defective tube, (2) open cathode circuit, (3) zero screen-grid voltage, and (4) control-grid bias more negative than cutoff.

Zero screen-grid voltage. This often results from a shorted screen bypass capacitor, causing excessive current through the screen-grid resistor, which then burns open. In most tubes designed for positive screen-grid voltage, without this accelerating voltage there is no plate current and no output.

Cathode bias. An open cathode circuit opens the return for the plate circuit, resulting in zero plate current and full B+ at the plate. The indication of an open cathode resistor is a voltage reading much higher than normal at the cathode. When the voltmeter is connected, its high resistance completes the cathode circuit, giving a reading approximately equal to the cutoff voltage of the tube.

D-C coupled stages. In this case, high positive d-c voltages at the control grid and cathode are normal because there is no coupling capacitor to block the B+ voltage. When using a VTVM, measure from each electrode to chassis ground. Then subtract readings to obtain the potential difference between electrodes. If any one electrode has the wrong voltage, it will change all the other d-c voltages in the amplifier because of the direct coupling. Circuits where d-c coupling may be used include cascode r-f amplifier, video amplifier with keyed AGC stage, and audio output tube in a stacked B+ arrangement with several i-f stages.

Grid-leak bias. This negative d-c voltage indicates the amount of a-c signal input. In an oscillator, the grid-leak bias indicates feedback voltage is being generated, which means the oscillator is operating; no bias indicates no feedback and the oscillator is not operating. Oscillator stages generally use grid-leak bias. In an amplifier that uses grid-leak bias, this negative d-c voltage indicates the amount of a-c signal input from the previous stage. The amplifiers that usually have grid-leak bias are the horizontal output, mixer, and sync separator stages. Stages generally using grid-leak bias are listed in Table 23·6, with typical bias values. These voltages are preferably measured with a VTVM for high impedance.

Table 23·6 Typical grid-leak bias voltages

Stage	Bias, d-c volts	Notes
Horizontal amplifier	−35	Shows drive from horizontal oscillator
Sync separator	−30	Receiver must have signal input
Horizontal multivibrator	−20	At V_2 grid. Blocking oscillator bias is higher
Vertical multivibrator	−75	At V_1 grid.
Mixer	−2	Produced by oscillator injection voltage
Local oscillator	−3	May be less on high channels

Detected signal voltage. The d-c voltmeter can also be used to measure rectified signal output of the video detector and ratio detector or discriminator for the FM sound. Typical values of rectified signal voltage are −3 to −5 volts. No d-c voltage output indicates no a-c signal input. Similarly, when a limiter stage is used for the FM sound signal, grid-leak bias of −5 to −15 volts indicates signal input. In a quadrature-grid detector, the quadrature grid normally has about −5 volts grid-leak bias with input to the signal grid.

Meter probes. These are used as follows:

1. D-C probe for a VTVM. This generally has a 1-megohm resistor inside. The high input resistance of the VTVM allows use of a series resistance to isolate the capacitance of the meter. Since the d-c voltage scales are calibrated for this series resistance, the probe is used for all d-c voltage measurements.
2. Direct probe. This is just an extension of the shielded meter lead. The purpose is simply to omit the isolating resistor used for d-c voltage measurements. Although called a direct probe, it is used for a-c voltage measurements. The frequency range of the meter is 30 cps to 3 Mc, as a typical example, for a-c voltages measured with the direct probe. For higher frequencies, use the r-f probe.
3. R-f probe. This includes a crystal diode detector circuit for a-c voltages from about 1 to 250 Mc. Although the probe input is a-c voltage, its output for the meter is d-c voltage, usually of negative polarity. Therefore, the meter is set to negative d-c volts when measuring a-c voltage with the r-f probe. Its applications include measuring video and i-f signal voltages. However, the minimum amount of input signal required for linear detection is about ½ volt. Also, the maximum input voltage is limited by the breakdown rating of the diode in the probe. Finally, the probe capacitance may detune the circuit being checked.
4. High-voltage probe. This includes an external multiplier resistor for measuring very high d-c voltages. The main application is measuring the kinescope anode voltage of 15 to 20 kv. The series multiplier resistance is generally 99 times the input resistance of the d-c voltmeter to extend the range by the factor of 100:1. As an example, with the meter range switch on 500 volts and the pointer on 180 volts, the voltage

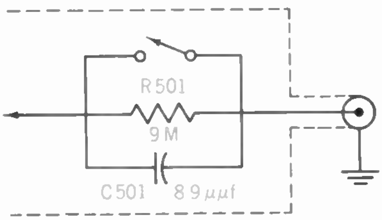
measured with the high-voltage probe is 180 × 100, or 18,000 volts. Note that this probe is for d-c voltages only.

23 · 11 Oscilloscope measurements

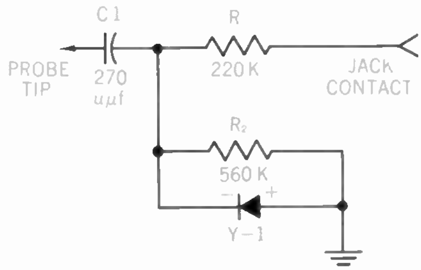
The oscilloscope is generally used for nonsinusoidal voltages because you can observe the waveshape, measure peak-to-peak amplitude, and check the frequency. Waveshapes are often checked for the video signal, from detector to kinescope, for the sync circuits and in the vertical and horizontal deflection circuits. The pattern on the oscilloscope screen shows peak-to-peak amplitude of the a-c signal. Referring to the example of video signal in Fig. 23 · 10, a d-c voltmeter can measure the average plate voltage of 140 volts, while the oscilloscope shows the a-c video signal with 80 volts peak-to-peak amplitude.

Usually, the oscilloscope is set for two cycles of the input signal as in Fig. 23 · 10*c*, in order to see the complete beginning and end of a typical cycle. For two cycles on the screen, you must set the internal sweep frequency to one-half the frequency of the signal input to the vertical amplifiers of the oscilloscope. This means set the oscilloscope internal sweep to 60/2, or 30, cps to see two cycles of the 60-cps vertical sync pulses or vertical deflection voltage. Similarly the internal sweep is at 15,750/2, or 7,875, cps for two cycles of the 15,750-cps horizontal sync pulses or horizontal deflection voltage.

Fig. 23 · 11 Typical 5-in. oscilloscope. (RCA.)



(a)



(b)

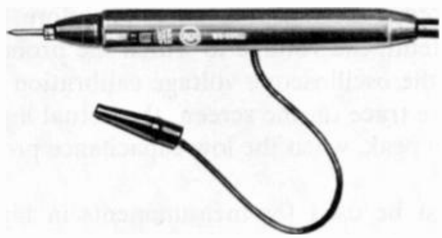


Fig. 23 · 12 Probes for the oscilloscope in Fig. 23 · 11. (a) Direct and low capacitance. (b) Demodulator. (RCA.)

When observing composite video signal, it can appear with vertical sync pulses if the oscilloscope sweep is at 30 cps, or with horizontal sync pulses if the oscilloscope sweep is at 7,875 cps. The technique is simply to set the sweep frequency to the desired range and then vary the fine frequency until you see two cycles on the screen. Many oscilloscopes have a switch that sets the internal sweep frequency to either 30 or 7,875 cps.

A typical 5-in. oscilloscope is shown in Fig. 23 · 11. This unit has V and H positions on the frequency-range switch to set the internal sweep at 30 or 7,875 cps. Trimmer adjustments at the side calibrate the sweep frequency for these two positions. The frequency response for signal input to the vertical amplifiers in the oscilloscope extends up to 4 Mc in the wide-band position. This allows observing waveforms at 3.58 Mc, which is the color subcarrier frequency. Also, the oscilloscope has built-in voltage calibration to allow peak-to-peak voltage readings for the signal input. The following three probes are used with the lead for signal input to the vertical amplifiers: direct probe, low-capacitance probe, and demodulator or detector probe (see Fig. 23 · 12).

Direct probe. This is just an extension of the shielded lead, without any isolating resistance. Shielding is needed to prevent pickup of interfering signals. One particular problem is hash in the oscilloscope trace resulting from pickup of horizontal deflection signals when observing waveforms at the vertical deflection frequency. If necessary, the horizontal deflection can be disabled temporarily to eliminate the hash.

It should be noted that a shielded lead has relatively high capacitance. A typical value is 60 $\mu\mu$f. Also, the oscilloscope has an input capacitance of about 10 $\mu\mu$f. The total input capacitance then is 70 $\mu\mu$f. Therefore, the direct probe can be used only where this capacitance has little effect on the circuit to which the lead is connected. This usually means audio-frequency circuits with relatively low resistance of several thousand ohms or less. In visual alignment, the direct probe can be connected to the video detector output or mixer grid to see the response curve. The oscilloscope input in this case is detected sweep signal, producing d-c voltage varying at the 60-cps sweep rate.

Low-capacitance probe. This probe has an internal series resistance to isolate the capacitance of the cable and oscilloscope. Usually the isolating resistance is nine times the oscilloscope internal resistance. Therefore, the oscilloscope input voltage is one-tenth the voltage to which the probe is connected. As an example, when the oscilloscope voltage calibration indicates 8 volts peak to peak for the trace on the screen, the actual input signal amplitude is 80 volts peak to peak, when the low-capacitance probe is used.

The low-capacitance probe must be used for measurements in high-frequency or high-resistance circuits. Furthermore, this probe is needed to observe the correct waveform of nonsinusoidal voltages. This means use the low-capacitance probe to check video signal, sync signal, horizontal deflection, and vertical deflection waveforms, especially since there is usually enough signal to allow for the 1:10 voltage division. Keep in mind, though, that the higher the input impedance is, the more it picks up interfering signals from stray fields.

Demodulator probe. This has an internal crystal diode rectifier. Therefore, the probe can be used to observe the envelope of amplitude-modulated i-f and r-f signals. This probe must be used for the visual response curve of the video amplifier, to demodulate the video sweep signal at the kinescope grid-cathode circuit. The polarity of the d-c output voltage of the probe is usually negative.

The oscilloscope demodulator probe and VTVM r-f probe both have diode detectors but they are not interchangeable. The r-f probe for a meter has a long-time-constant filter for the d-c output voltage that would eliminate the variations needed to observe any envelope signal.

Voltage measurements. The oscilloscope is very useful as an a-c voltmeter for nonsinusoidal waveforms. First, the oscilloscope must be calibrated in terms of how much height is produced in the trace on the screen by the vertical amplifier. In general, the procedure is to compare the amplitude of a nonsinusoidal waveform with a sine wave that can be measured accurately with an a-c voltmeter.

To calibrate the oscilloscope for peak-to-peak voltage measurements of all waveforms, or to check the built-in calibration:

1. Connect, to the vertical input of the oscilloscope, a sine-wave voltage from either the a-c power line or an audio oscillator.
2. It is preferable to shut off the internal sweep so there will be just

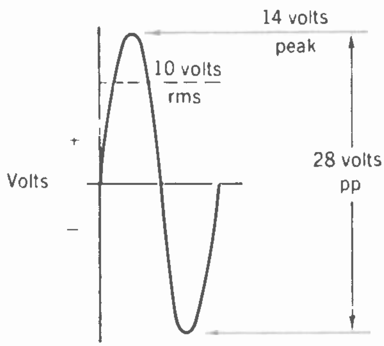


Fig. 23·13 Peak-to-peak sine-wave voltage is 2.8 times the rms value.

an easily measured vertical line on the screen, without horizontal deflection.

3. Vary the amount of a-c input voltage for a trace 1 in. high. Do not change the setting of the oscilloscope vertical gain controls.
4. Measure the amount of sine-wave input with an a-c voltmeter. Read the peak-to-peak value or multiply the rms value by 2.8. As an example, input voltage of 10 volts rms is equal to 2.8 × 10 or 28 volts peak to peak (see Fig. 23·13).
5. The a-c voltmeter reading is the oscilloscope sensitivity in peak-to-peak volts per 1-in. deflection. In this case the sensitivity is 28 volts per inch.

Now the peak-to-peak amplitude of any trace on the screen can be measured. Suppose that sawtooth voltage with a trace 2 in. high is the oscilloscope input voltage. Its amplitude is 2 × 28, or 56, volts peak to peak. Similarly, a 3-in. trace measures 84 volts peak to peak. This calibration is good for just the one setting of the oscilloscope vertical gain. When the unknown voltage to be measured produces a trace too large or too small, reset the oscilloscope gain for a trace of convenient size and calibrate again with the same procedure. The voltage calibration applies for all frequencies having flat response in the oscilloscope vertical amplifier.

Current measurements. Although the oscilloscope is a voltage-operated device, current waveforms can be observed and peak-to-peak measurements made the same way. This can be accomplished by inserting a known resistance in series with the circuit in such a manner that the current to be measured flows through the resistor. The voltage across this resistor is then coupled to the vertical binding posts of the oscilloscope so that its waveshape can be observed and peak-to-peak amplitude measured as usual. With the voltage measured, the current I can be calculated as E/R. The current amplitudes are then in peak-to-peak values, like the voltage amplitudes. As an example, suppose sawtooth deflection voltage across a 10-ohm R measures 6 volts peak to peak. The current I then is 6/10, equal to 0.6 amp or 600 ma peak to peak. The current waveshape is identical with the observed voltage waveshape, since current and voltage are in phase with each other in the resistive element and there is no change in waveshape.

The resistance to be inserted temporarily for the current measurement should be as small as possible so that it will not change the current in the circuit appreciably. In high-current circuits such as the deflection coils a resistance of 3 to 10 ohms is enough.

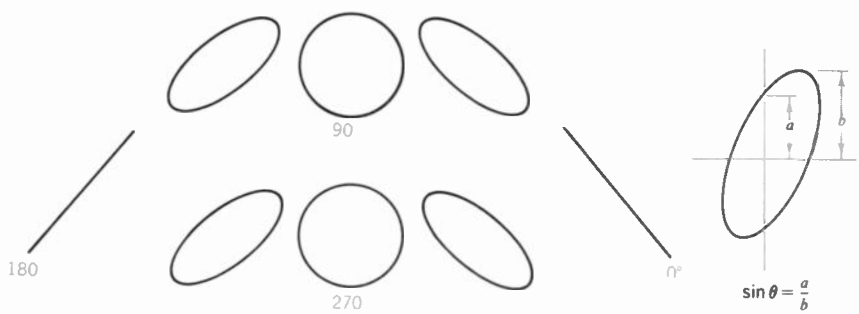
Frequency comparisons. In order to check frequencies with the oscilloscope, the unknown signal voltage is compared with a voltage source of known frequency. Since we are usually interested in the scanning frequencies of 60 and 15,750 cps, which are audio frequencies, an audio signal generator with an accurate frequency calibration is used. The procedure can be as follows:

1. Connect the voltage of unknown frequency to the vertical input of the oscilloscope. Adjust internal sweep frequency for a trace of two cycles on the screen. It is preferable not to use any internal sync by switching the oscilloscope switch to external sync. Adjust for the least amount of motion, if the trace drifts across the screen.
2. Disconnect the voltage to be measured and connect in its place the output of the audio oscillator.
3. Adjust the audio oscillator frequency for two cycles on the screen. Do not change the oscilloscope frequency controls.
4. The frequency of the audio oscillator is now equal to the frequency of the unknown voltage previously connected, since they both produce a trace of two cycles with the same sweep frequency.

The setting of the gain controls does not matter in the frequency comparison and they can be adjusted for a pattern of convenient size. The results can be very accurate when no sync is used for the oscilloscope sweep and the signal generator frequency is correct.

This same procedure can be used to check the accuracy of the signal generator at 60 and 15,750 cps. For 60 cps, compare the generator frequency with the 60-cycle power line. For 15,750 cps, compare the gen-

Fig. 23·14 Lissajous patterns on oscilloscope screen comparing phase of two sine waves of the same frequency. The phase angle is θ.



erator frequency with the output of the horizontal deflection oscillator in a receiver where the picture is in horizontal sync.

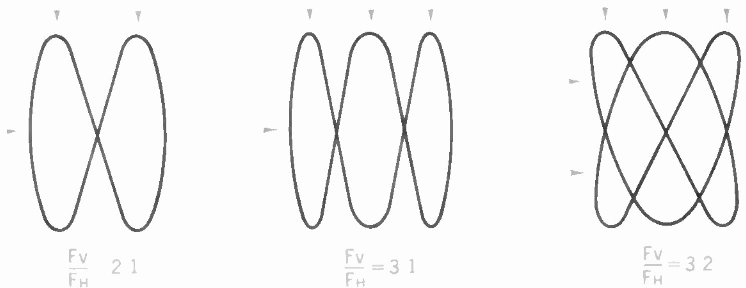
Phase-angle measurements. The oscilloscope can also be used to compare the phase of two sine-wave voltages of the same frequency. One sine-wave voltage is applied to the vertical input of the oscilloscope. At the same time, the other sine-wave voltage is applied to the horizontal input. The oscilloscope internal sweep is not used. Both the vertical and horizontal voltages should produce the same amplitude in the oscilloscope trace. The combined trace of two sine-wave voltages will look like one of the patterns in Fig. 23 · 14. The phase angle θ can then be calculated by the ratio of the lengths a and b, as shown, which determine the sine of the angle. As an example, for a circle a/b equals 1 for the sine and the angle is 90°. When a is one-half b, the sine equals 0.5 and θ is 30°. Remember, though, that phase-angle measurements apply only for sine waves.

Lissajous patterns. The phase-angle patterns actually show a 1:1 frequency ratio. For higher ratios, the patterns in Fig. 23 · 15 are produced. Count the loops across either the top or bottom of the trace for the value of F_v, and count the loops at either side for F_H. The frequency ratio is then equal to F_v/F_H. Be sure to count only closed loops; open loops do not count at all.

As an example, assume the horizontal input is 60-cps power-line voltage as a reference frequency. The vertical input is signal from an audio oscillator to check the calibration of its frequency dial. For a circle or line pattern, where F_v/F_H is 1, the audio signal generator frequency is 60 cps. With a ratio of 2 for F_v/F_H, the vertical input frequency is 120 cps. Furthermore, frequencies that are not exact multiples can be compared with this method. For the case of F_v/F_H equal to 3/2, in this example the vertical input is 90 cps.

How to reproduce the picture on oscilloscope screen. The cathode-ray tube in the oscilloscope can be used as a kinescope to see the picture on a green screen, if this is desired to check a receiver. Note the following requirements:

Fig. 23 · 15 Lissajous patterns on oscilloscope screen comparing frequencies of two sine waves. F_v is frequency of signal applied to oscilloscope vertical input; F_H is frequency of horizontal input. Arrows indicate closed loops that are counted for frequency ratio.



1. The video signal output from the receiver is connected to the Z axis or intensity-modulation terminal of the oscilloscope. This terminal couples the signal to the control grid of the cathode-ray tube.
2. Sawtooth deflection voltage at 60 cps from the receiver is connected to the vertical amplifier of the oscilloscope.
3. Sawtooth deflection voltage at 15,750 cps is coupled to the horizontal amplifier of the oscilloscope. Either the oscilloscope internal sweep can be used or external horizontal deflection voltage from the receiver. With oscilloscope internal sweep, synchronization of the picture is difficult. With deflection voltage from the receiver, use a decoupling resistor to prevent detuning of the horizontal oscillator.

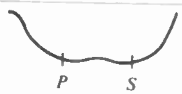
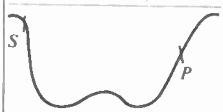
The oscilloscope vertical and horizontal amplifiers then provide a scanning raster on the screen. Use the oscilloscope gain controls to adjust for convenient height and width. The video signal may produce a positive or negative picture, depending on polarity. A minimum of about 25 volts peak-to-peak signal is necessary.

23 · 12 Alignment precautions

The main points to be noted when using the oscilloscope and sweep generator to obtain a visual response curve are summarized in Table 23 · 7. The curves are usually down from the base line, as shown here, with negative polarity of rectified signal input to the oscilloscope. The wrong polarity for the response curve can result from oscillations or overload distortion. Which side of the curve corresponds to sound or picture signal is checked with the marker oscillator.

For correct results the alignment must be done without changing the

Table 23 · 7 Visual response curves

Test equipment	i f response 	r f response 	r f and i f response 
Sweep generator connected to	Mixer grid	Antenna input	Antenna input
Marker oscillator connected to	I-f circuit	R-f circuit	I-f circuit
Oscilloscope connected to	Video detector load resistance	Mixer grid circuit	Video detector load resistance
Sweep generator frequency at	I-f	R-f, same as selected channel	R-f, same as selected channel
Marker frequency at	I-f	R-f, same as selected channel	I-f

operating conditions of the amplifier being tested. Use typical bias and signal voltage. Most important, do not connect the test equipment directly to any circuit being aligned. The oscilloscope is connected to the video detector load resistor, to measure rectified i-f signal without detuning the i-f amplifier. For i-f alignment, the signal generator is connected to the mixer grid to avoid detuning the i-f amplifier when feeding in i-f signal. When the oscilloscope is connected to the mixer grid to check r-f response, a decoupling resistance is necessary to prevent detuning the converter grid circuit. For this reason rectified r-f output is sometimes checked at the screen grid of the mixer. When r-f signal is fed to the antenna input, the signal generator cable is matched to the receiver impedance to avoid detuning the r-f input circuit. If the r-f input signal from the generator is radiated to the receiver antenna, no impedance match is necessary.

The following checks can be made to be sure the curve you see is the actual receiver response:

1. Reduce the signal input to see that the curve just becomes smaller, without changing its shape. A curve that is perfectly flat probably results from overload distortion, caused by too much signal, either in the i-f amplifier or in the oscilloscope. Also, the curve should disappear when you turn off the signal generator. If the curve remains, the i-f amplifier is probably oscillating.
2. Reduce the marker signal to zero to see that the curve remains the same with or without the marker. Incidentally, if you see two markers moving in opposite directions when the marker frequency is varied, this probably means the curves were not phased correctly before the blanking was turned on. Any spurious markers that do not move with the marker frequency can usually be eliminated by disabling the local oscillator. When the marker is too broad, it can be made narrower by connecting a 0.01-μf bypass capacitor across the vertical input of the oscilloscope.
3. If the curve changes its shape when you touch the chassis, check the ground connections.
4. With weak r-f signal input, the oscilloscope trace has "grass," corresponding to snow in the picture produced by receiver noise.

In summary, the response curve should change its shape only when you change the alignment adjustments.

23 · 13 Typical receiver circuit

Figure 23 · 16 illustrates how the circuits for individual sections fit together in a television receiver. Typical waveshapes are shown in red for video, sync, and deflection voltages, with normal peak-to-peak values for these a-c voltages measured with an oscilloscope. The d-c voltages in black are with no signal input. Note the test points, which are available at the top of the chassis to check all sections of the receiver. These test points are listed in Table 23 · 8. The i-f picture carrier frequency is 45.75 Mc and associated sound 41.25 Mc, as in practically all receivers, with the intercarrier-sound signal at 4.5 Mc.

The r-f circuits are not included because either of two tuners is used with the same chassis. One is a rotary-switch tuner for VHF channels only, while the other is a turret tuner for VHF and UHF. In either case the tuner is a separate unit requiring the following input connections from the receiver chassis: heater voltage, B+ voltage, and AGC bias. The i-f output from the plate of the converter stage in the tuner is coupled to the grid of the first i-f amplifier. Note that test point I is at the grid of the mixer or converter, either for injecting i-f signal for i-f alignment or as the looker point for the oscilloscope in r-f alignment.

There are three i-f stages with bifilar coils to amplify the picture and sound signals for the crystal diode video detector. V_3 is the only i-f stage with AGC bias because its plate-cathode circuit is in series with V_4 in a stacked B+ circuit. The d-c cathode voltage of V_4 is the plate supply for V_3.

The AGC bias is derived from the negative grid-leak bias voltage of the sync clipper V_{6B}. Test point II is used to check the AGC bias, which is -3.2 volts with normal signal input at the control grid of the first i-f amplifier V_3. The r-f bias is less because of the positive clamp voltage

Table 23·8 Test points in Fig. 23·16

	TEST POINT	USE
I	Mixer grid	Inject i-f signal for i-f alignment; connect oscilloscope for r-f response curve
II	AGC bias	Measure AGC bias voltage; should vary from -1 to -15 volts with signal
III	Video detector output	Measure detected i-f output; -2 to -5 d-c volts or 3 to 10 volts peak-to-peak a-c signal. Connect oscilloscope for i-f response. Inject signal here for video amplifier input
IV	Kinescope input	Check video signal drive for kinescope; typical value 70 to 100 volts peak to peak
VI	Horizontal AFC	Measure d-c control voltage for horizontal oscillator. Short to chassis to adjust oscillator frequency without AFC
VIII	Stabilizing coil	Check sine-wave stabilizing voltage
IX	B+ voltage	Check B+ for oscillator. Short to VIII to adjust hold control without stabilizing coil
X	Quadrature-grid signal	Check quadrature signal of FM detector
XI	Quadrature-grid bias	Check grid-leak bias produced by quadrature-grid signal
XII	Audio take-off	Check audio signal output of detector. Measure d-c plate voltage

4

Fig. .

applied through R_{160}. For weak signals, the grid-cathode circuit of the r-f amplifier serves as a clamp diode to keep the r-f bias close to zero.

Test point III is for video detector output voltage. Here you can measure about -2 volts d-c output or 3 volts peak-to-peak video signal, with normal i-f signal input. Also, the oscilloscope is connected to this point for a visual response curve of either the i-f amplifier or the combined r-f and i-f response. The load resistor for the crystal diode Y_{151} is R_{168}, with L_{159} and L_{160} the peaking coils. The small r-f chokes L_{157} and L_{158} reduce interference from harmonics of the detected video signal.

The video signal output of the detector is coupled by C_{168} to the single video amplifier stage V_{6A}. Note the inversion of sync polarity by the amplifier. Its composite video signal with positive sync polarity for the kinescope cathode is 85 volts peak to peak, measured at test point IV, with normal contrast. The plate load resistance R_{172} is in parallel with the contrast control R_{173}.

The picture tube uses low-voltage electrostatic focus. Pin 4 is the focus grid, with a potential of zero, 268, or 580 volts, as determined by the focus jumper. The spark gap at pin 3 eliminates internal arcing of the screen-grid voltage, which is 580 volts from the boosted B+ line. Both vertical and horizontal retrace suppression are used by coupling flyback pulses to the kinescope control-grid pin 6. Horizontal flyback pulses come from the winding between terminals 7 and 8 on the output transformer T_{251}. In addition, the lower winding of the vertical output transformer connects to terminal 8 on T_{251}, to couple vertical flyback pulses to the kinescope grid. The purpose of using internal horizontal blanking is to eliminate faint vertical lines at the left that may be produced by color sync voltage when a color program is transmitted.

Referring back to the plate circuit of the video amplifier, the primary of T_{154} is a 4.5-Mc trap to keep sound out of the picture, while the secondary couples the 4.5-Mc sound i-f signal to V_7. The sound section is at the bottom of the diagram. This circuit with a gated-beam FM detector is the same as in Fig. 22·20. Note the stacked B+ arrangement with the audio output tube V_9 to supply 135 volts. All points in the diagram marked +135 volts return to the V_9 cathode for this supply voltage.

The video amplifier plate circuit also provides signal for the sync clipper V_{6B}, taken from the top of L_{161}. With 70 volts peak-to-peak composite video signal of positive sync polarity at control-grid pin 2, the separated sync output is 60 volts peak to peak, as the positive sync pulses are clipped and amplified. Note the grid-leak bias of -40 volts produced by the input signal. This negative d-c voltage is used for AGC bias, reduced by the voltage division between R_{162} and R_{161} in the r-f bias line. The sync clipper also has video signal input of negative sync polarity from the detector, through R_{162}, C_{164}, and R_{165}. This line provides sync voltage of opposite polarity to cancel the effect of high-amplitude noise pulses in the signal.

The separated sync voltage is coupled to both the vertical oscillator and the horizontal AFC circuit. Vertical sync of negative polarity is integrated by the printed unit *RC* 201 for coupling to the plate of V_{10A} and the grid of V_{10B}. The integrated sync voltage is amplified and inverted in V_{10B} to provide positive sync through the feedback line to the grid of V_{10A}. In addition, the negative sync output of the clipper is coupled through the 39-$\mu\mu$f C_{251} to the diodes for horizontal AFC. This circuit is a single-ended discriminator controlling the frequency of the cathode-coupled horizontal multivibrator V_{12}.

The vertical deflection circuits include two triode sections combining the oscillator and output stages, with feedback from the plate of V_{10B} to the grid of V_{10A}. Both stages use the boosted B+ of 580 volts for plate voltage. C_{204} is the coupling capacitor between stages. The sawtooth capacitor is C_{206} in series with the peaking resistor R_{212}. Note the typical value of 140 volts peak-to-peak trapezoidal voltage at the grid of the output tube. Bias for this amplifier is taken from the grid-leak bias of the oscillator section. R_{204} taps off the desired bias, filtered by R_{206} and C_{205}. The plate voltage for the oscillator is varied by R_{208}. Both the height and linearity controls must be adjusted for the desired raster height with good linearity.

The horizontal deflection circuits include the multivibrator V_{12}, power output stage V_{13}, damper V_{15}, and the flyback high-voltage supply. In the oscillator output, C_{263} is the sawtooth capacitor with the peaking resistor R_{260} to produce horizontal drive of 130 volts peak to peak at the grid of the output tube. Note the value of -35 volts grid-leak bias, which indicates normal drive. The autotransformer T_{251} provides output for the horizontal deflection coils between terminal 6 and the width control. The horizontal coils in the yoke are in parallel, but the vertical coils are in series. The damper cathode is connected to terminal 4 on T_{251}, to provide boosted B+ of 580 volts at terminal 1 with respect to ground. C_{265} is the boost capacitor connected between terminal 1 and B+.

SUMMARY

1. The receiver setup adjustments include height, width, and linearity for the raster; AGC to prevent overload; and horizontal AFC to sync the picture.
2. Multipath signals received by the antenna are the usual cause of ghosts. However, built-in ghosts can be produced by misalignment of the receiver, a long transmission line incorrectly terminated, or direct pickup by the receiver chassis. With direct pickup, the signal strength changes when people walk near the receiver.
3. Types of interference patterns are shown in Figs. 23·3 to 23·5.
4. Hum in the video signal produces horizontal bars; hum in the horizontal sync makes the picture bend; hum in the vertical sync can lock in the picture at the wrong position.
5. Typical troubles are grouped in Tables 23·2 to 23·5.
6. The d-c voltmeter is used to measure electrode voltages to localize a defective component.
7. The oscilloscope is used for a-c voltage measurements in the video, sync, and deflection circuits. Typical peak-to-peak values are shown by the waveshapes in red in Fig. 23·16.
8. The three oscilloscope probes are used as follows: (*a*) Low-capacitance probe whenever there is enough signal to allow for the 1:10 voltage division. This probe must be used to see nonsinusoidal signals with correct waveshapes. (*b*) Direct probe where its high capaci-

tance is not critical and the signal level is low. This probe can generally be used for i-f and r-f visual response curves. (*c*) Detector probe to see the envelope of an AM signal, or to see a visual response curve for circuits that do not end with a detector, as in checking video amplifier response up to the kinescope.

9. The oscilloscope is used to check frequency or phase. See the Lissajous patterns in Figs. 23·14 and 23·15.
10. The main precautions in alignment are: (*a*) do not allow the test equipment to detune the circuits being aligned, (*b*) do not use too much signal to avoid overload, and (*c*) use an accurate frequency marker.

SELF-EXAMINATION *(Answers at back of book.)*

Match the items in the right column with those in the left column.

1. Overload distortion	*a.* Mixer grid test point
2. Venetian-blind effect	*b.* Horizontal output tube
3. Poor vertical linearity	*c.* Kinescope cathode
4. Bowed edges on raster	*d.* Co-channel interference
5. Wormy picture	*e.* Small raster
6. Bars and bend in picture	*f.* Weak r-f signal
7. +135 volts at audio output tube cathode	*g.* Horizontal AFC
8. No high voltage	*h.* Video detector test point
9. Low B+	*i.* 580 volts
10. Snowy picture	*j.* AGC adjustment
11. Picture tears into diagonal bars	*k.* Weak vertical output tube
12. I-f alignment	*l.* No brightness
13. 4 volts peak-to-peak video signal	*m.* Pincushion magnets
14. −35 volts grid-leak bias	*n.* Hum in video signal
15. Boosted B+ voltage	*o.* 10:1 voltage divider
16. 100 volts peak-to-peak video signal	*p.* 90° phase angle
17. Low-capacitance probe	*q.* 28 volts peak to peak
18. 40 volts per inch	*r.* Oscilloscope sensitivity
19. Circle pattern on oscilloscope screen	*s.* Stacked B+
20. 10 volts rms	*t.* 4.5-Mc beat

ESSAY QUESTIONS

1. Describe briefly how to adjust: (*a*) ion-trap magnet; (*b*) pincushion magnet; (*c*) deflection yoke; (*d*) focus magnet.
2. Describe briefly how to adjust the height of the raster.
3. Why is the picture size slightly smaller than the raster?
4. Give two possible causes of ghosts and a remedy for each.
5. Describe briefly how to adjust the AGC level control.
6. Describe briefly how to obtain 10 vertical bars and 10 horizontal bars on the kinescope screen to check linearity.
7. What is the distinguishing feature of FM interference in the picture?
8. How does a television receiver cause whistles in nearby radio receivers?
9. How can horizontal streaks in the picture be distinguished between ignition interference and motor noise?
10. How can you recognize sound bars in the picture?
11. How can you recognize 4.5-Mc intercarrier beat in the picture?
12. Explain how an interfering carrier at 67.251 Mc can interfere with a carrier at 67.250 Mc to produce venetian-blind effect in the picture.
13. Describe briefly the interference pattern resulting from oscillator radiation.
14. What is the required internal sweep frequency of the oscilloscope to see two cycles of the following waveshapes: (*a*) vertical oscillator output; (*b*) video signal with vertical sync;

(*c*) horizontal oscillator output; (*d*) video signal with horizontal sync?

15. Describe the equipment setup for the following alignment curves for the receiver in Fig. 23 · 16: (*a*) i-f response; (*b*) r-f response; (*c*) overall r-f and i-f response.
16. Which meter probe would you use for the following: (*a*) heater voltage; (*b*) grid-leak bias; (*c*) screen-grid voltage; (*d*) cathode bias on the video amplifier; (*e*) 15 kv anode high voltage?
17. Which oscilloscope probe would you use for the following: (*a*) 1,000-cps sine-wave audio; (*b*) grid drive for horizontal output stage; (*c*) grid drive for vertical output stage; (*d*) r-f response curve?
18. Draw a block diagram of the receiver circuit in Fig. 23 · 16, with a cascode r-f tuner.
19. For the receiver circuit in Fig. 23 · 16: (*a*) list all fuses with function of each; (*b*) list all setup adjustments.
20. In Fig. 23 · 16, why is the control grid of V_2 at +130 volts with respect to chassis ground?
21. For the receiver circuit in Fig. 23 · 16, localize each of the following troubles: (*a*) Set completely dead. No raster picture or sound. (*b*) No picture and no sound with normal raster. (*c*) No raster and no picture with normal sound. (*d*) No sound, with normal raster and picture. (*e*) Insufficient width in raster. (*f*) Only a thin horizontal line across center of screen. (*g*) 4.5-Mc beat pattern in picture. (*h*) Picture does not hold horizontally. Vertical hold is normal.
22. Referring to the receiver circuit in Fig. 23 · 16, indicate where a d-c voltmeter would be connected to measure: (*a*) detected i-f signal; (*b*) grid-leak bias on horizontal oscillator; (*c*) grid-leak bias on horizontal output tube; (*d*) AGC bias.

PROBLEMS *(Answers to odd-numbered problems at back of book.)*

1. Referring to the video amplifier V_6 in Fig. 23 · 16, calculate the cathode current through R_{171}.
2. For the i-f amplifier V_5 in Fig. 23 · 16, calculate cathode current, screen-grid current, and plate current.
3. For the horizontal multivibrator V_{12} in Fig. 23 · 16, calculate the peak cathode current through R_{259}.
4. Draw an i_b-e_c curve at plate voltage of 580 volts, for a typical horizontal output tube such as type 6DQ6 or 6FW5. Show the d-c bias and a-c drive using the values shown in Fig. 23 · 16 for V_{13}.
5. How much is the voltage across R_{266} in the screen-grid circuit of V_{13} in 23 · 16?
6. Calculate the total heater current through fuse F_{402} in Fig. 23 · 16.
7. Referring to the oscilloscope pattern in Fig. 23 · 14, the ellipse has a height a of 1 in. and 2 in. for b. Calculate the phase angle.
8. For the ellipse in Fig. 23 · 14 list the ratios of a/b for the phase angles of 15°, 30°, 45°, 60°, and 90°.
9. For the Lissajous patterns in Fig. 23 · 15, the horizontal input frequency is 1,000 cps. What are the vertical input frequencies for the three patterns shown?
10. Draw the Lissajous pattern for 400 cps vertical input and 1,200 cps horizontal input, both sine waves.

Chapter 24 The color television system

A reproduction in natural colors is much more pleasing than a black-and-white (monochrome) picture, which is a variation in shades of white only. For a color picture, the three primary colors in television are red, green, and blue, used either alone or in combinations. Practically all natural colors as well as white, gray, and black can be obtained by mixtures of these three primary colors. In fact, the range of colors possible in television is greater than the color range now used in printing. A typical color television picture is shown in Plate I.

24·1 Color signals

When the image is scanned at the broadcast station, video signals corresponding to the desired picture information are obtained for the red, green, and blue information in the scene by means of optical color filters (see Fig. 24·1). Picture information for the red content of the image is in the red video signal, while the green picture information is in the green video signal and the blue parts of the picture produce the blue video signal. At the color television receiver, the red, green, and blue video signals are used for reproduction of the picture in its natural colors as mixtures of red, green, and blue by means of a tricolor kinescope arrangement. Although Fig. 24·1 shows three separate picture tubes for a projection system, the color receiver generally uses one tricolor kinescope. This tube has three electron guns but one screen with red, green, and blue phosphors. Details of color kinescopes are described in Secs. 25·8 to 25·10.

In color television broadcasting, however, the red, green, and blue video signals are not transmitted. Instead, they are combined to produce a composite color signal and a monochrome signal. These two signals are transmitted to the receiver. The monochrome signal is produced by adding the color video signals in the proportions necessary to indicate only the bright-

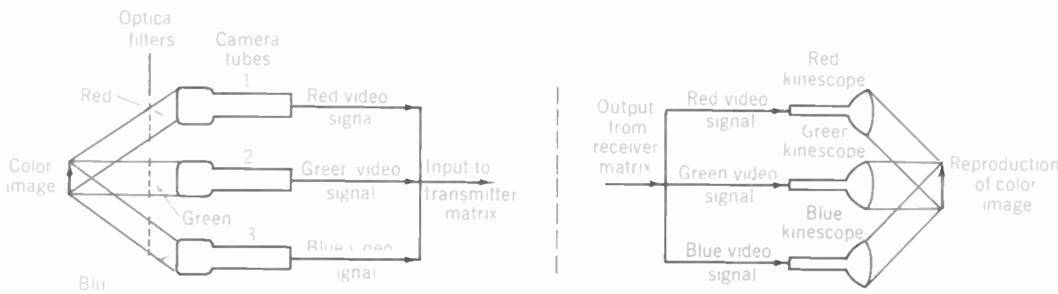


Fig. 24·1 Televising a scene in color.

ness variations in the picture information. For this reason, it is the *luminance signal.* Conversion to the luminance signal is necessary for compatibility with monochrome television broadcasting. The compatibility feature means monochrome receivers can reproduce in black and white pictures that are televised in color. The color signal with the red, green, and blue information is the *chrominance signal.* This signal for color is multiplexed with the luminance signal so that both can modulate the picture carrier. Color television receivers utilize both the luminance and chrominance signals.

In summary, then, color television consists of transmitting a luminance signal essentially the same as the monochrome signal in black-and-white television broadcasting, with the addition of the chrominance signal for color. In this way, any station can use its assigned 6-Mc channel for either monochrome or color television broadcasting.

24·2 Color addition

In color television, reproduction of the many different colors in the scene is based upon the addition of primary colors. The process is *additive* because individual color images are produced by the kinescope and combined in an arrangement that allows the eye to integrate the individual colors.

The additive effect can be obtained by superimposing the color images, as illustrated in Fig. 24·1. In this arrangement the screen of each color kinescope has a phosphor that produces a red, green, or blue image. By means of an optical projection system, the three primary color images are projected onto one common viewing screen. Since the phosphor of each kinescope is a light source producing a color image, and the color images are viewed together, the observer sees the picture on the viewing screen in all its natural colors as additive mixtures of the three primary colors. The same results are obtained with a single kinescope having a tricolor phosphor screen.

Additive color mixtures. This is shown in color in Plate VII. There are three individual circles in red, green, and blue here, which partially overlap. Where the circles are superimposed, the color shown is the mixture obtained by adding the individual colors. At the center, all three color circles overlap, resulting in white. Therefore, the white area at the center is a mixture of red, green, and blue in the proper proportions.

Notice that, where only green and blue add, the resultant color is a greenish-blue mixture generally called *cyan*. The red-purple color shown by the addition of red and blue is called *magenta*; more blue with less red produces purple. Yellow is an additive color mixture of approximately the same amount of red and green; more red with less green produces orange. Similarly, practically all natural colors can be produced as additive mixtures of the three primary colors red, green, and blue.

Primary colors and complementary colors. Colors that can be combined to form different color mixtures are primary colors, the only requirement being that none can be matched by any mixture of the other primaries. Red, green, and blue are the primary colors used in television because they produce a very wide range of color mixtures when added. Therefore, red, green, and blue are *additive primaries*.

The color that produces white light when added to a primary is called its *complementary color*. For instance, cyan added to red produces white light. Therefore, cyan is the complement of the red primary. The fact that cyan plus red equals white follows from the fact that cyan is a mixture of blue and green, so that the combination of cyan and red actually includes all three additive primaries. Similarly, magenta is the complement of green, while yellow is the complement of blue. Sometimes the complementary colors, cyan, magenta, and yellow, are indicated as minus-red, minus-green, and minus-blue, respectively, as each is equal to white light minus the corresponding primary.

The complementary colors are also known as the *subtractive primaries*. In a reproduction process such as color photography, where color mixtures are obtained by subtracting individual colors from white light by the use of color filters, cyan, magenta, and yellow are the subtractive primary colors used to filter out red, green, and blue.

24 · 3 Definition of color television terms

The most important terms used in color television are defined as follows:

White. For practical purposes, white light can be considered a mixture of the red, green, and blue primary colors in the proper proportions. The reference white for color television is a mixture containing 30 per cent red, 59 per cent green, and 11 per cent blue, which combine to produce a bluish white like daylight.

Hue. The color itself is its hue. Green leaves have a green hue; a red apple has a red hue; the color of any object is distinguished primarily by its hue. Different hues result from different wavelengths of the light producing the visual sensation in the eye.

Saturation. This indicates how little the color is diluted by white light, distinguishing between vivid and weak shades of the same hue. Pastel blue, for instance, has little saturation, while vivid blue is highly saturated. The more a color differs from white the greater is its saturation. Saturation is also indicated by the terms *purity* and *chroma*. High purity and chroma correspond to high saturation and vivid color.

Chrominance. This term is used to indicate both hue and saturation of a color. *Chromaticity* is also used for chrominance. As will be shown, the transmitted chrominance signal indicates hue and saturation of the red, green, and blue color information.

Luminance. This indicates the amount of light intensity, which is perceived by the eye as brightness. In addition to the more familiar luminance variations in black-and-white monochrome, different colors are perceived with different brightness values by the eye. As illustrated in the relative luminosity curve in Plate VIII, the green hues between cyan and orange have maximum brightness.

Registration. This refers to the positioning of individual color images to make the resultant picture of color mixtures have the correct color. With superimposed images, as an example, if each does not exactly overlay the others incorrect color mixtures will be produced because the primary colors will be in the wrong position with respect to the original colors in the picture.

Compatibility. Color television is compatible with black-and-white television because essentially the same scanning standards are used and the luminance signal enables a monochrome receiver to reproduce in black and white a picture televised in color. In addition, color television receivers can use a monochrome signal to reproduce the picture in black and white. Color television broadcasting uses the same 6-Mc broadcast channels as monochrome transmission. Also, the same picture carrier frequency is used.

The three distinguishing features of any color are its hue, saturation, and brightness. Most obvious is the hue corresponding to the wavelength of the light. Remember that light energy can be considered a form of electromagnetic radiation but with much higher frequencies than radio waves. The visible light frequencies are above 430×10^{12} cps. Because of these high values, light waves are usually considered in terms of wavelength, using the millimicron unit equal to 1×10^{-9} meter. As shown in Plate VIII, blue has the shortest wavelength of 400 millimicrons, red has the longest wavelength of 700 millimicrons, with green hues around 550 millimicrons. Each color has a *dominant wavelength,* which determines its hue.

Vivid red when mixed with white light will appear as a pink color. The hue of the two colors is the same because the dominant wavelength has not changed. However, the pink is less saturated. The vivid red, as an example of a color with 100 per cent saturation, has no white light. As white light is added, the per cent saturation decreases and the color is weakened or desaturated.

The last identifying characteristic of any color is its brightness. This really indicates how the color will look in a black-and-white reproduction. Consider a scene being either photographed on black-and-white film or televised in monochrome. The picture includes a colorful costume with a dark red skirt, yellow blouse, and light blue hat. For the same illumination, these different hues will have different brightness values and will therefore be reproduced in different shades of monochrome. As shown by the graph in Plate VIII for relative brightness values of different hues, dark red has low brightness, yellow has high brightness, and blue is medium in relative brightness. Therefore, the monochrome picture reproduction will show a white blouse (for yellow) with black skirt (for red) and a gray hat (for blue). Actually, it is the relative brightness variations for different hues that make it possible to reproduce scenes that are naturally in color as similar pictures in black and white.

24 · 4 Color television broadcasting

The camera receives red, green, and blue light corresponding to the color information in the scene, to produce the primary red, green, and blue color video signals indicated as *R*, *G*, and *B* in Fig. 24 · 2. This illustrates the color video signal voltages that would be obtained in scanning one horizontal line across the color bars indicated. The 100 per cent amplitude represents maximum color video voltage, which is produced by a saturated color.

Notice the values shown for yellow, as an example of a complementary color consisting of two primary colors. Since yellow includes red and green, video voltage is produced for both these primary colors, as the red and green camera tubes have light input through their color filters. Furthermore, the red and green video voltages are at the 100 per cent level because the yellow is 100 per cent saturated. However, there is no blue in yellow. This is why the blue video voltage is at zero for the yellow bar.

The last bar at the right is white, as an example that includes all three primaries. Then all three camera tubes have light input and there are color video signals for red, green, and blue. Also, each color video signal is at the 100 per cent level because the white is made up of saturated primary colors. The red, green, and blue video voltages are then combined in order to encode the primary color voltage signals as brightness and chrominance signals for transmission to the receiver. This is illustrated in Fig. 24 · 3.

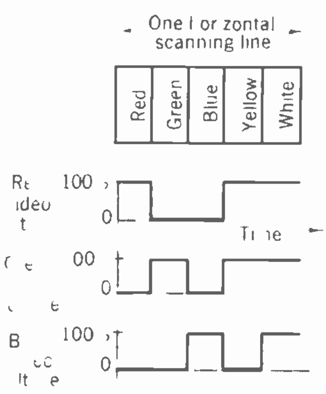


Fig. 24 · 2 Red, green, and blue color video signal voltages corresponding to color-bar pattern. All colors 100 per cent saturated.

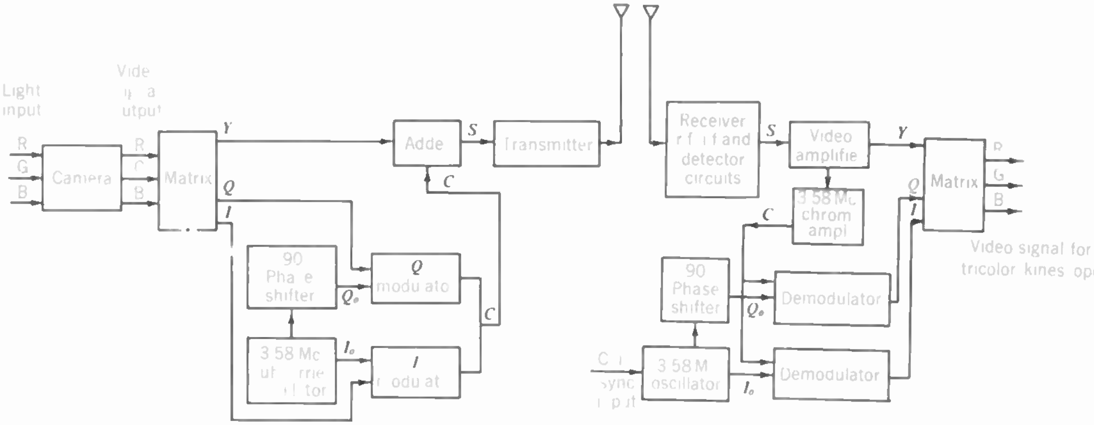


Fig. 24·3 Functions of the transmitter and receiver in the color television system.

Matrix section. The matrix is essentially a resistive voltage-divider circuit that proportions the primary color signals as required to produce the brightness and chrominance signals. With the red, green, and blue color video voltages as input to the matrix, the three video signal output combinations formed are the following:

1. Luminance signal, designated the *Y signal,* which contains the brightness variations of the picture information.
2. A color video signal, designated the *Q signal,* which corresponds to either green or purple picture information.
3. A color video signal, designated the *I signal,* which corresponds to either orange or cyan picture information.

The *I* and *Q* signals are combined to form the color information for the chrominance signal. It should be noted that the *Y* signal is not for yellow, as it is actually an equivalent monochrome signal.

Color subcarrier. The *I* and *Q* signals are transmitted to the receiver as the modulation side bands of a 3.58-Mc subcarrier. A subcarrier is a relatively low frequency carrier wave, within the range of modulation frequencies, which, in turn, modulates the main carrier wave. As an example, the picture carrier wave at 67.25 Mc for channel 4 can be modulated by the 3.58-Mc video-frequency color subcarrier. Note that the color subcarrier has the same video frequency of 3.58 Mc for all stations, although the assigned picture carrier frequency is different for each channel.

The method of using a subcarrier to transmit two modulation signals on one main carrier is called *multiplexing.* Here the 3.58-Mc chrominance signal *C* is multiplexed with the *Y* signal so that both chrominance and luminance can be transmitted on the picture carrier.

Chrominance modulation. Referring to Fig. 24·3, the output from the 3.58-Mc color subcarrier oscillator is coupled to the *I* and *Q* modulators. They also have *I* and *Q* video signal input from the matrix. Each circuit produces amplitude modulation of the 3.58-Mc subcarrier by I or Q video signal. Notice the separate inputs for the I and Q modulators but the common output combines the I and Q modulation. This combined output is the chrominance signal C.

The 3.58-Mc oscillator input to the Q modulator, labeled Q_o, is shifted by 90°, with respect to I_o. This is the reason for the designation Q signal, since the Q modulation product is *in quadrature*, or 90° out of phase, with the I signal. Modulating the subcarrier in these two different phases makes it possible to broadcast the color information of the I and Q signals simultaneously, without loss of identity. Each can be recovered as a separate signal, with the same phasing between the demodulators for the I and Q signals at the receiver.

The I and Q signals modulate the 3.58-Mc subcarrier in a *balanced modulator*, which is a circuit that cancels out the modulated carrier, leaving only the side bands. This is done to reduce interference. With the color subcarrier suppressed, therefore, it must be reinserted at the receiver to detect the modulated chrominance signal. This is why the color receiver has a 3.58-Mc oscillator.

Total video signal. Both the chrominance signal C containing the color information and the Y signal with the luminance information are coupled to the adder section, or colorplexer. This stage combines the Y signal and the 3.58-Mc chrominance signal to form the total colorplexed video signal transmitted to the receiver, here indicated as S. The oscilloscope waveshape for colorplexed composite video signal can be seen in Fig. 24 · 12. This signal is transmitted by amplitude modulation of the broadcast station's assigned picture carrier. Vestigial-side-band transmission, negative polarity of transmission, and 4.5-Mc separation from the sound carrier frequency are used for the picture carrier signal, in accordance with the standard characteristics of the 6-Mc television broadcast channel. Also, the S modulation is a composite video signal including deflection sync and blanking pulses, in addition to color synchronizing signals to time the color information correctly.

Color sync. Because the chrominance signal includes the I and Q color information as two-phase modulation, without the subcarrier itself, a sample of the 3.58-Mc subcarrier must be transmitted to indicate the correct frequency and phase for the 3.58-Mc signal reinserted in the receiver for the demodulators. This color synchronization is accomplished by transmitting 8 to 10 cycles of the 3.58-Mc r-f subcarrier on the back porch of each horizontal blanking pulse.

Decoding the primary color signals. To recover the red, green, and blue video signals for the color kinescope in the receiver, essentially the same functions must be performed as at the transmitter, but the decoding at the receiver is in reverse order. Starting with the receiving antenna, the modulated picture carrier signal of the selected channel is amplified in the r-f and i-f stages and rectified in the video detector to provide the total video signal S. Following the detector, the video circuits divide into two separate paths, one for monochrome and the other for chrominance. In

Table 24 · 1 Sequence of color signals

Transmitter	Receiver
1. Red, green, and blue video from camera	1. Antenna signal is r-f picture carrier modulated by colorplexed video
2. Y, I, and Q video	2. Colorplexed video with Y and 3.58-Mc chrominance, from video detector
3. 3.58-Mc chrominance with I and Q modulation	3. Demodulated chrominance may be I, Q, $R - Y$, $B - Y$, $G - Y$, or other color video
4. Colorplexed video with Y and 3.58-Mc chrominance	4. Color video and Y signals for kinescope input
5. Antenna signal is r-f picture carrier modulated by colorplexed video	5. Red, green, and blue on kinescope screen

Fig. 24 · 3, the Y signal in the output of the video amplifier includes only the luminance component of the S signal input. The original color video modulation is in the 3.58-Mc chrominance signal, but it must be demodulated.

To recover the I and Q signals, the total video signal S is also coupled to the chrominance band-pass amplifier for the color demodulators. This stage amplifies the chrominance signal because it is tuned to 3.58 Mc. The chrominance signal output is then coupled to the I and Q demodulators. Note that the 3.58-Mc local oscillator reinserts the subcarrier frequency needed in the demodulators to detect the I and Q signals. The color signals from the demodulators, with the Y signal from the video amplifier, are then combined in the matrix unit. The receiver matrix forms the original red, green, and blue primary color video signals for the kinescope reproduction of the picture in color. As a summary of the different signals, Table 24 · 1 lists their sequence from camera tube to antenna at the transmitter and from antenna to kinescope at the receiver.

The color picture is generally reproduced by a tricolor kinescope, which has a screen that produces red, green, and blue light. In a kinescope having a tricolor screen and three electron guns to produce an electron beam for each of the color phosphors, each color video signal voltage is applied to one of the three guns.

24 · 5 Y signal

Now we can consider more details of the luminance signal, which contains the brightness variations of the picture information. The Y signal is formed in the matrix at the transmitter by adding red, green, and blue primary color video signals in the proportions

$$Y = 0.30R + 0.59G + 0.11B \qquad (24 \cdot 1)$$

These proportions correspond to the relative brightness of these three primary colors. Therefore, a scene reproduced in black and white by the Y signal looks the same as when it is televised in monochrome.

Figure 24·4 illustrates how the Y signal voltage in d is formed in the matrix from the specified proportions of the primary color voltages shown in a, b, and c, corresponding to a standard color-bar pattern. Notice that the bars include the primary colors, their complementary mixtures of two primaries, and white for all three primaries.

Fig. 24·4 Waveforms, corresponding to color-bar pattern. All colors 100 per cent saturated.

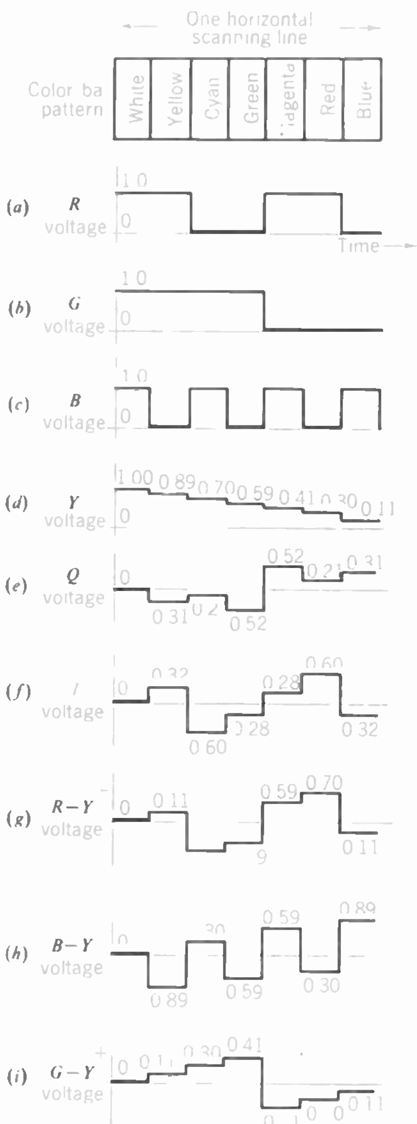
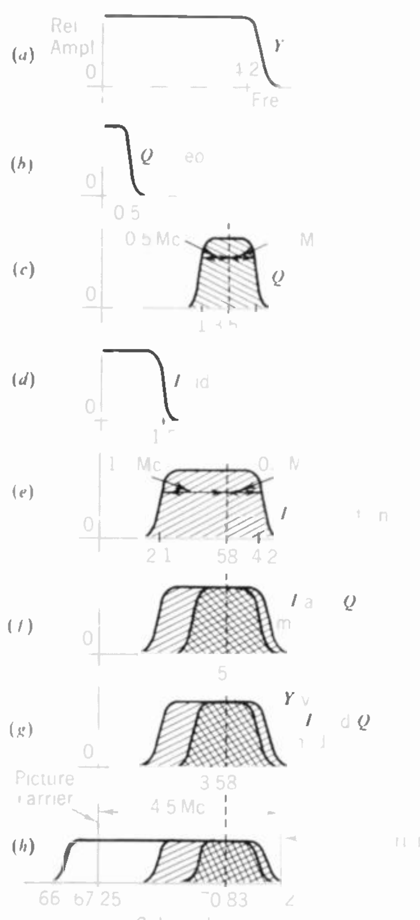


Fig. 24·5 Bandwidth requirements of luminance and chrominance signal frequencies. Graph at bottom is for colorplexed composite video signal modulating channel 4 picture carrier.



The Y signal for white has the maximum relative amplitude of unity, equal to 30 per cent red, 59 per cent green, and 11 per cent blue. The yellow color bar, as another example, produces red and green but no blue video voltage. Adding 30 per cent of the red voltage and 59 per cent of the green voltage results in luminance video signal voltage having the relative amplitude of 0.89 for yellow. If the *Y* signal alone were used to reproduce this pattern, it would appear as monochrome bars shading off from white at the left to gray in the center and black at the right, corresponding to the relative brightness of these color bars.

The *Y* signal has the full video-frequency bandwidth of 4.2 Mc. This is illustrated in *a* of Fig. 24·5, which shows the bandwidth requirements for the signal voltages in Fig. 24·4.

24·6 Types of color video signals

The main color signals are red, green, and blue video because the system starts and finishes with these primary hues. At the camera, the color information is disassembled into red, green, and blue video signals and at the receiver these three video signals are used by the kinescope to reproduce the picture in its natural colors. For a closed-circuit color television system, just the red, green, and blue video signals could be used. For broadcasting, however, these three primary video signals are not transmitted because too much bandwidth would be needed and the system would not be compatible with monochrome television. Therefore, the primary color signals are converted to *Y* signal for luminance and chrominance signal for color. The chrominance signal contains all the color information in terms of hue and saturation.

The transmitted chrominance signal consists of *I* and *Q* color video signals. The *I* signal has been chosen because it represents either orange or cyan, which are the hues best for color reproduction of small details. The *Q* signal is automatically 90° out of phase with the *I* signal, for quadrature modulation of the 3.58-Mc color subcarrier. The hue of *Q* video signal is either purple or green.

At the receiver, however, any two hues in the chrominance signal can be detected by the demodulators. The hue of the output depends on the phase of the reinserted 3.85-Mc color subcarrier signal coupled into the demodulator. Remember, though, that the tricolor kinescope needs red, green, and blue video signals for a phosphor screen in these three primary colors.

In a receiver designed for maximum color detail, the *I* and *Q* video signals are detected and then converted by a matrix circuit to red, green, and blue video for the kinescope. The bandwidth of color video with *I* signal is approximately 1.5 Mc, compared with 0.5 Mc for the other color video signals.

However, the extra color detail of the *I* signal is seldom necessary. Therefore, most color receivers simplify the circuits in the chrominance section to a great extent by using red, green, or blue color-difference video signals instead of *I* and *Q* signals. In this case, the two demodulators can detect the hues corresponding to $R - Y$ and $B - Y$. Color-difference signals are used because two of them have color information for all three primaries. Remember that the Y signal includes red, green, and blue video.

Voltage waveforms. Each of these types of color video signal is described in more detail in the following sections. Referring to the voltage waveforms in Fig. 24·4, note the following general facts about the relative amplitudes:

1. The Y luminance signal has its maximum value for white.
2. All the color video signals for white are at zero. The proportions of red, green, and blue in the color signals are chosen to make the primary components cancel for white.
3. In each color video signal, a primary color and its complement have equal values but opposite polarity. For instance, in the Q video signal green has the relative amplitude of -0.52 while its complement magenta is $+0.52$.

Bandwidth and color resolution. The frequencies needed for the voltage waveforms in Fig. 24·4 are illustrated by the bandwidth response curves in Fig. 24·5. Note that the Y signal has full bandwidth of 4.2 Mc but this is not necessary for the color video signals. The reason is that for very small details the eye can perceive only brightness, rather than hue and saturation. Therefore, the color information can be transmitted with restricted bandwidth much less than 4 Mc. This feature allows the narrow-band chrominance signal to be multiplexed with the wide-band luminance signal in the standard 6-Mc broadcast channel, without too much interference between the two signals. Utilizing the fact that the eye does not require color for very fine details of picture information, the color resolution can be considered in the following three parts:

1. Full three-color reproduction as additive mixtures of the red, green, and blue primaries for large areas of picture information. The full color applies for resolution up to 0.5 Mc, corresponding to 1/25 of the width of the picture, approximately. This color information is available with any two of the color-difference video signals.
2. Two-color reproduction as cyan or orange for smaller areas. This color information is in the I signal, with its 1.5-Mc bandwidth. The video-frequency range from 0.5 to 1.5 Mc corresponds to details 1/25 to 1/150 of the picture width.
3. Monochrome reproduction in black and white for the smallest details of picture information, up to the video-frequency limit of 4.2 Mc, which cannot be perceived in color by the eye. This picture information is in the Y signal.

The color characteristics of the color television picture are illustrated by the NTSC[1] flag shown in Plate IX. It should be noted, however, that

[1]The National Television Systems Committee (NTSC) of the Electronics Industries Association (EIA) prepared the standard specifications approved by the FCC, December, 1953, for commercial color television broadcasting.

with a televised image the entire picture appears in color because full three-color reproduction is provided for the largest areas in the scene. In a televised picture of an automobile, for instance, the entire body of the car would be in full color; narrow vertical strips on the frame might be reproduced less exactly in orange or cyan, while the detail of the outline of the car where it joins the background would be in black-and-white monochrome. The background would be in full natural colors.

24·7 Q signal

This color video signal is produced in the transmitter matrix as the following combination of red, green, and blue:

$$Q = 0.21R - 0.52G + 0.31B \qquad (24\cdot2)$$

It should be noted that the minus sign merely indicates the addition of video signal voltage of negative polarity. For instance, $-0.52G$ means 52 per cent of the total green video signal but inverted from the polarity out of the green camera tube.

Opposite polarities of the Q video signal represent the complementary colors purple and green. We can see this by combining the primary color components. Positive Q signal voltage contains minus-green, or magenta, with red and blue, which combine to produce magenta or purple hues. For the complementary color, negative Q signal voltage is primarily green.

Comparative values of the Q video signal for the standard color-bar pattern are shown in Fig. 24·4*e*. Relative voltage values for each bar are indicated according to the red, blue, and green components. For magenta. as an example, the Q video voltage consists of 21 per cent red plus 31 per cent blue. This equals 0.52 for the magenta component in the Q signal. Note that the Q signal has its peak positive and negative amplitudes for the magenta and green bars, indicating that the positive values of Q voltage represent mainly purple hues and the negative ones mainly green.

The range of video frequencies from 0 to 0.5 Mc is used for the Q signal. This is illustrated in Fig. 24·5*b*, showing the Q video signal frequencies. In *c*, the Q signal frequencies are shown as side bands of the 3.58-Mc suppressed subcarrier, which is the way the Q signal is transmitted in the broadcast channel. Both the upper and lower side bands of the Q signal, extending approximately 0.5 Mc above and below the 3.58-Mc subcarrier frequency, are included.

24·8 I signal

This color video signal is formed from the primary color signals in the matrix at the transmitter, as follows:

$$I = 0.60R - 0.28G - 0.32B \tag{24·3}$$

Opposite polarities of the I video signal represent the complementary colors orange and cyan. Combining the primary color components, positive I signal contains primarily red combined with minus-blue, or yellow, which add to produce orange. Negative I signal combines green plus blue to form cyan, with minus-red, which is cyan.

The bar pattern is shown in Fig. 24·4*f* with relative I voltage values for each bar indicated according to the red, blue, and green components. Consider the cyan bar as an example. In terms of its component blue and green primary color voltages, it produces 28 per cent negative green and 32 per cent negative blue voltage, which total -0.60 for cyan. Note that the I signal has its peak positive and negative amplitudes for the red and cyan bars. These values show that positive I voltage corresponds mainly to red-orange hues and negative I voltage is mainly cyan.

The I signal is transmitted with the range of video frequencies from 0 to 1.5 Mc. Compared to the Q signal, more bandwidth is used for the I signal because orange and cyan color information can be resolved by the eye for small color details.

Figure 24·5*d* shows the video frequencies for the I signal, while in *e* the side-band frequencies of the I signal are shown as the modulation product of the 3.58-Mc suppressed subcarrier. Note that vestigial-side-band transmission with respect to the 3.58-Mc subcarrier is used for the I signal, only part of the upper-side-band frequencies being transmitted. The lower side band of the I signal includes the full 1.5-Mc bandwidth, while the upper side band extends approximately 0.5-Mc above the subcarrier frequency. Upper-side-band frequencies more than 0.5 Mc above the 3.58 Mc subcarrier cannot be transmitted because they would interfere with the sound carrier signal. Therefore, the 1.5-Mc bandwidth for the I signal can be considered in two parts: (1) modulating frequencies from 0 to 0.5 Mc, approximately, are transmitted with double side bands, as for the Q signal; (2) modulating frequencies of 0.5 to 1.5 Mc in the I signal are transmitted with the lower side band only.

Referring to *f* in Fig. 24·5, this shows the video frequencies transmitted for both the I and Q signals, as side bands of the 3.58-Mc subcarrier. The crosshatched area extending 0.5 Mc above and below 3.58 Mc indicates where the frequencies utilized for the I and Q signals overlap. Therefore, both I and Q color information are provided for the video frequency range from 0 to 0.5 Mc. The I signal alone can be utilized for color information corresponding to video frequencies in the range of 0.5 to 1.5 Mc.

The video frequencies for the colorplexed total video signal including Y, I, and Q are shown in Fig. 24·5*g*. The corresponding r-f signal frequencies transmitted to the receiver are illustrated in *h*, using channel 4, 66 to 72 Mc, as an example.

24·9 R − Y signal

The −Y signal has opposite polarity from Eq. (24·1). Adding 100 per cent R to $-Y$, the result is

$$R - Y = 1.00R - 0.30R - 0.59G - 0.11B$$

therefore,
$$R - Y = 0.70R - 0.59G - 0.11B \qquad (24\cdot4)$$

The hue of $R - Y$ signal is mainly red and magenta, for minus-green, which are equivalent to purplish red. The voltage waveform for $R - Y$ signal is shown in Fig. 24·4*g* for the standard color-bar pattern. The advantage of $R - Y$ signal in the receiver is the simplicity of converting to red video voltage for the kinescope. When $+Y$ signal is added to $R - Y$ signal, the result is red video alone for the kinescope reproduction.

24·10 B − Y signal

Combining $-Y$ signal and 100 per cent B, the result is

$$B - Y = -0.30R - 0.59G + 0.89B \qquad (24\cdot5)$$

The hue of $B - Y$ signal is mainly blue. The magenta for $-G$ and cyan for $-R$ make the hue a purplish blue. Voltage waveforms for $B - Y$ signal are shown in Fig. 24·4*h*. When $+Y$ signal is added to $B - Y$ signal the result is blue video alone for the kinescope reproduction.

24·11 G − Y signal

Combining $-Y$ signal and 100 per cent G, the result is

$$G - Y = -0.30R + 0.41G - 0.11B \qquad (24\cdot6)$$

The hue of $G - Y$ signal is mainly green and cyan, forming a bluish green. The voltage waveforms for $G - Y$ signal are shown in Fig. 24·4*i*.

In the receiver, $G - Y$ video can be obtained as the combination of $R - Y$ and $B - Y$ in the following proportions:

$$G - Y = -0.51(R - Y) - 0.19(B - Y) \qquad (24\cdot7)$$

Then, with $+Y$ signal added to the $G - Y$ signal, the result is green video alone for the kinescope reproduction.

24·12 Desaturated colors

The relative voltage values shown in Fig. 24·4 are for vivid colors that are 100 per cent saturated. In this case, there is no primary color video for hues not included in the color. As examples, saturated R has zero B and G video voltage; saturated yellow (red-green) has zero B video voltage. This follows from the fact that with zero light input to a given color camera there is no signal output.

In natural scenes, however, most colors are not 100 per cent saturated. Then any color diluted by white light has all three primaries. The following example illustrates how to take into account the amount of desaturation for weaker colors. Assume 80 per cent saturation for yellow. Now this color has two components: 80 per cent saturated yellow and 20 per cent white. First consider the primary color video signals produced by each camera tube:

80 per cent yellow (red-green) produces	$0.80R$	$0.80G$	$0.00B$
20 per cent white (red-green-blue) produces	$0.20R$	$0.20G$	$0.20B$
Total camera output is	$1.00R$	$1.00G$	$0.20B$

These percentages of primary color video voltages can then be used for calculating relative amplitudes of the Y signal and color video signals for 80 per cent saturated yellow. As examples, this desaturated color has the Y value of 0.912, Q value of -0.248, and I value of 0.256. Compare these with the values of 0.89, -0.31, and 0.32 shown in Fig. 24·4 for 100 per cent saturated yellow. Note that the addition of white to desaturate a color increases the luminance value and decreases the chrominance value, compared with 100 per cent saturation.

24·13 The transmitted chrominance signal

Since the phase of the 3.58-Mc subcarrier is shifted 90° for the Q modulator, the Q component is 90° out of phase with the I component in the chrominance signal. Therefore, the chrominance signal consists of the I and Q signals in quadrature.

When two signal voltages 90° out of phase with each other are combined, the resultant is the vector sum of the quadrature components, just as in combining reactive and resistive a-c voltages. The amplitudes of the two quadrature components determine the magnitude of the resultant voltage. The phase angle of the resultant varies with the relative value of one quadrature voltage compared with the other.

In terms of the I and Q quadrature voltages, the instantaneous value of the resultant chrominance signal varies in amplitude and phase with varying I and Q signals. The amplitude of the resultant chrominance signal indicates the saturation of the color information in the picture, while the phase of the resultant chrominance signal corresponds to the hue. This is illustrated in Fig. 24·6. The top row shows fully saturated colors, while below are the same hues with 50 per cent saturation.

Color values of the chrominance signal. Taking the chrominance signal for yellow in Fig. 24·6*a* as an example, the relative voltages of 0.32 for I and -0.31 for Q, compared with 1 for white, are obtained with equal amounts of saturated red and green but no blue. Combining I and Q vectorially results in the chrominance vector C shown for yellow, with a phase angle between I and $-Q$. With less I and $-Q$ but in the same proportions, the phase angle is the same for the resultant yellow chrominance

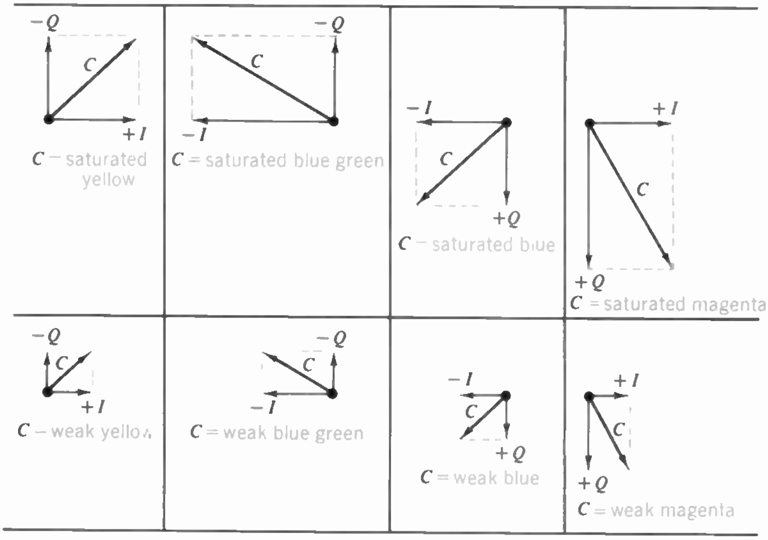


Fig. 24·6 Vector addition of I and Q signal voltage for four different hues. The phase angle of the resultant signal C indicates the hue while its length corresponds to saturation. From left to right the hues are yellow, blue-green, blue, and red-purple. Bottom row has same hues but only 50 per cent saturation.

signal. However, there is less amplitude for the resultant, corresponding to less saturation. The directions of the *I* and *Q* vectors correspond to the color wheel in Fig. 24·7. This shows the different hues for phase angles of the *C* signal.

The different hues and saturation values that can be obtained with the four main combinations of positive or negative *I* and *Q* voltages are illustrated in Fig. 24·6*a*, *b*, *c*, and *d*. Notice that opposite polarities of chrominance signal voltage correspond to complementary colors. For instance, the phase angles for yellow and blue differ by 180°.

In terms of the reproduction of color information by the kinescope, positive signal voltage can be considered as adding the color, while negative signal voltage is equivalent to subtracting from the white reproduced on the kinescope screen as the sum of the primary colors. White results from the red, green, and blue phosphors excited by all three electron guns. For yellow, only the red and green guns are on. Or, minus-blue voltage can cut off beam current of the blue gun to allow production of red-green for yellow. Similarly, the chrominance signal phase angle can be considered a specific hue in terms of the polarity and amplitude of the component primary signal voltages for the kinescope.

In summary, then, the hue and saturation of the picture information are in the chrominance signal, while the luminance is in the *Y* signal. Therefore, the transmitted signal includes three specifications of the picture in-

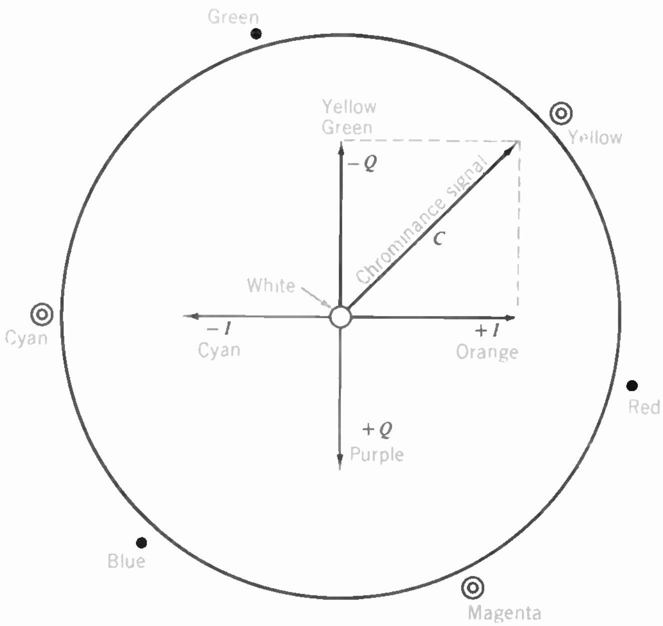


Fig. 24·7 Color wheel showing approximate hues for different combinations of I and Q signal voltage. The −Q voltage is shown upward to indicate green at the top.

formation, converted from the original red, green, and blue signals, as follows:

1. Hue is transmitted as the instantaneous phase angle of the modulated 3.58-Mc color subcarrier.
2. Saturation is transmitted as the instantaneous amplitude of the modulated 3.58-Mc color subcarrier.
3. Luminance is transmitted as the instantaneous amplitude of the Y signal.

Since these characteristics of the transmitted signal include all the required picture information, the receiver can convert back to red, green, and blue video signals for the color kinescope. The conversion of red, green, and blue information to hue, saturation, and luminance, and then back again, is necessary for compatibility with monochrome broadcasting.

24·14 Matrix circuits

The matrix has the function of adding several input voltages in the desired proportions to form new combinations of output voltage. At the transmitter, the matrix has three input voltages corresponding to the red, green, and blue primary colors in the televised scene and forms the Y, Q, and I output signals for transmission. In the receiver, the matrix is used to

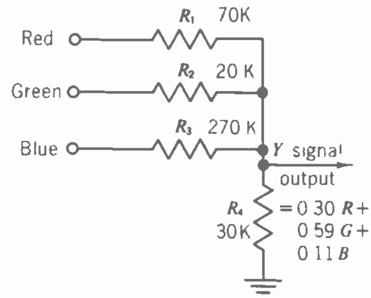


Fig. 24·8 Resistive matrix circuit for forming Y signal at transmitter. Resistance values approximate.

form the original R, G, and B primary color video signals for the tricolor kinescope. Usually the kinescope itself serves the matrix by combining input signal voltages coupled to the three grid-cathode circuits. Then the matrix operation is in terms of the effect of the three signal voltages on the amount of beam current.

The required proportions of the component voltages in a separate matrix circuit can be obtained by means of resistive voltage dividers. Adding the proportioned voltages is accomplished by combining them across a common load resistance. Subtraction of a signal voltage is done by adding it in negative polarity to a signal voltage of positive polarity.

An example is illustrated in Fig. 24·8. The resistive voltage divider proportions the R, G, and B primary color video signals, in accordance with the resistances of R_1, R_2, and R_3 compared with the total resistance. The result is Y signal output across the common load resistor R_L. Similarly, different resistance values for separate voltage dividers can be used for proportioning R, G, and B in the amount necessary to produce the Q and I signals at the transmitter.

24·15 Color subcarrier frequency

This must be a high video frequency, in the range between 2 and 4 Mc, approximately. If the color subcarrier frequency is too low, it can produce excessive interference with the luminance signal. At the opposite extreme, the chrominance signal can interfere with the sound signal. The choice of approximately 3.58 Mc for the color subcarrier is a compromise that allows 0.5-Mc side bands for chrominance information below and above the subcarrier frequency. Also, there is room for the extra 1 Mc of lower side frequencies of the I signal. Most important for compatibility, 3.58 Mc is a video frequency high enough to have little response in monochrome receivers. These sets use the luminance signal alone, with practically no effect from the 3.58-Mc chrominance signal.

The exact frequency of the color subcarrier is based on the following additional factors:

1. The transmitted picture carrier and sound carrier frequencies cannot be changed, in order to preserve the 4.5-Mc beat for intercarrier-sound receivers.

2. There will be an interfering beat frequency of approximately 0.92 Mc or 920 kc between the color subcarrier frequencies near 3.58 Mc and the intercarrier sound at 4.5 Mc.
3. There will be interfering beat frequencies between the chrominance signal and higher video frequencies of the luminance signal.

In order to minimize these interference effects, the color subcarrier frequency is made exactly 3.579545 Mc. This frequency is determined by harmonic relations for the color subcarrier, the horizontal line-scanning frequency, and the 4.5-Mc intercarrier beat. Specifically, 4.5 Mc is made to be the 286th harmonic of the line frequency. Therefore,

$$f_H = \frac{4.5 \text{ Mc}}{286} = 15{,}734.26 \text{ cps}$$

where f_H is the horizontal line-scanning frequency for color television broadcasting. Notice that 286 is the even number that will make f_H closest to the value of 15,750 cps used for horizontal scanning in monochrome television. The slight difference has practically no effect on horizontal scanning and sync in the receiver because of the horizontal AFC circuit.

The vertical scanning frequency is also changed slightly, since there must be 262.5 lines per field. Then the vertical field-scanning frequency is

$$f_V = \frac{15{,}734.26}{262.5} = 59.94 \text{ cps}$$

The slight difference of 0.06 cps below 60 cps has practically no effect on vertical scanning and sync in the receiver, because an oscillator that can be triggered by 60-cycle pulses can also be synchronized by 59.94-cps pulses. It should be noted that, when the scanning frequencies are shifted slightly for color television, the transmitted sync is also changed to the new frequencies for f_V and f_H.

With the horizontal line-scanning frequency chosen, now the color subcarrier frequency can be determined. This value is made to be the 455th harmonic of $f_H/2$:

$$C = 455 \times \frac{15{,}734.26 \text{ cps}}{2} = 3.579545 \text{ Mc}$$

Because of the odd-line interlaced scanning pattern, picture information for video frequencies that are odd multiples of $f_H/2$ tends to cancel in its effect on the kinescope screen. The cancellation results because these frequencies have opposite voltage polarities for the picture information on even and odd scanning lines. You can prove this canceling effect by coupling from a signal generator output voltage at 2 to 4 Mc into the video amplifier to produce a diagonal-bar interference pattern on the kinescope screen. By adjusting the generator frequency carefully, and watching the screen pattern closely, at certain frequencies the interference pattern will disappear. These frequencies are odd multiples of one-half the horizontal line-scanning frequency.

The technique of interlacing odd and even harmonic components of two different signals in order to minimize interference between them is called *frequency interlace.* As a result of the frequency interlace, the chrominance signal can be transmitted in the same 6-Mc channel as the luminance and sound signals with practically no interference.

The 3.579545-Mc frequency is made an exact multiple of $f_H/2$ by using the color subcarrier signal to lock in the sync generator at the transmitter. The slight change in scanning frequencies is accomplished by the sync generator so that no changes are required at the receiver for programs broadcast in either monochrome or color. However, the fine tuning control must be set exactly to prevent excessive 920-kc beat between the sound signal and chrominance signal.

24·16 Color synchronization

Figure 24·9 shows details of the 3.58-Mc color sync burst transmitted as part of the total composite video signal for color synchronization in color receivers. Specifically, the color burst synchronizes the phase of the 3.58-Mc local oscillator. This stage reinserts the subcarrier needed for the demodulators in the receiver to detect the color-difference signals. The color synchronization is necessary to establish the correct hues corresponding to the phase of the chrominance signal. Then the color AFC circuit in the receiver can hold the hue values steady.

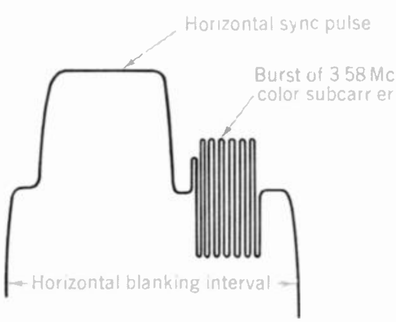


Fig. 24·9 The color sync burst.

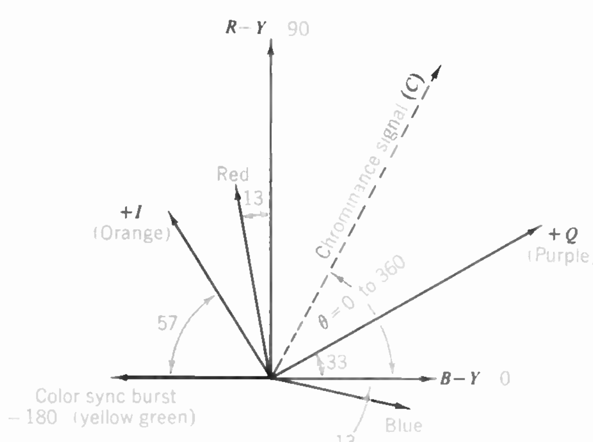


Fig. 24·10 Phases of different hues with respect to color sync burst phase. Relative amplitudes not to scale.

The burst is 8 to 11 cycles of the 3.58-Mc subcarrier transmitted on the back porch of each horizontal sync pulse. There is no color sync during vertical blanking time, in order to minimize the effect of the color burst in the sync circuits of monochrome receivers. Notice that the average value of the color burst coincides with the pedestal level of the composite video signal voltage. This prevents the horizontal sync circuits from interpreting the color burst as an increase in signal voltage. Then the color burst will not be mistaken for a horizontal sync pulse.

Reference hue phase. Figure 24 · 10 illustrates how the hues of the modulated chrominance subcarrier signal are determined by its varying phase angle with respect to the constant phase angle of the color sync burst. Note that the hue of color sync burst corresponds to yellow-green. As an example of how the hue of the chrominance signal is determined by its phase angle, when greenish-yellow picture information is scanned at the transmitter, the phase angle of the chrominance signal is made the same as the phase of the color sync burst. For other hues, the chrominance signal has different phase angles. How much the phase angle differs from sync burst phase determines how the hue differs from yellow-green.

24 · 17 Colorplexed composite video signal waveforms

Formation of the total video signal, combining the luminance signal Y and chrominance signal C, is illustrated in Fig. 24 · 11 in successive steps. Starting with the primary color video signals in *a*, *b*, and *c*, the red, green, and blue video voltages are shown for the time of scanning one horizontal line containing the color bars indicated. The relative voltage

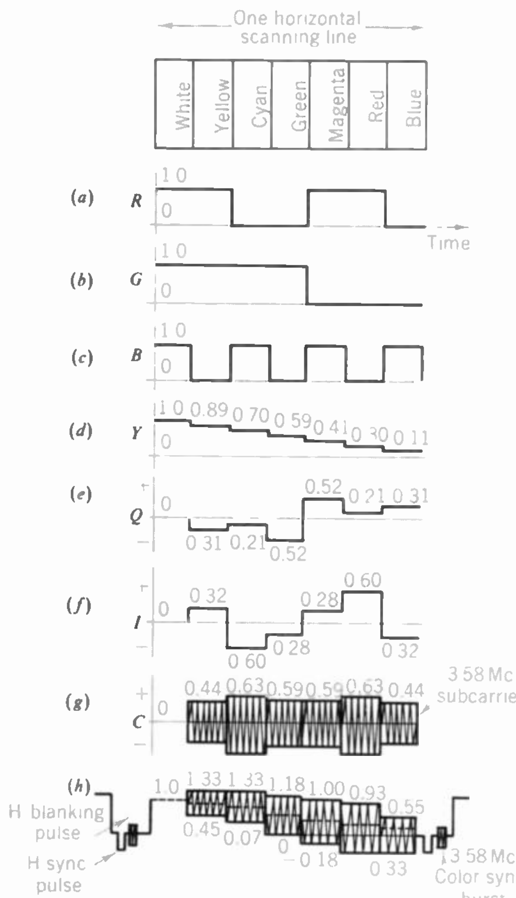


Fig. 24 · 11 Construction of colorplexed composite video signal from Y, I, and Q voltages. (a) Red video. (b) Green video. (c) Blue video. (d) Luminance video. (e) Q video. (f) I video. (g) 3.58-Mc subcarrier chrominance signal C modulated by I and Q in quadrature. (h) Colorplexed total video signal S including Y, I, and Q with standard blanking, deflection sync, and color sync burst.

amplitudes are indicated in terms of 100 per cent color video voltage for a fully saturated color.

The luminance, or Y, signal in d shows the brightness component for each color bar. The I and Q signals in e and f have the relative voltage values indicated according to their proportions of primary colors. Note that I and Q voltages can have either positive or negative values, because the components include both positive and negative primary colors.

Figure 24 · 11g illustrates the 3.58-Mc subcarrier amplitude-modulated by the I and Q signals. The amplitude of this 3.58-Mc modulated C signal is obtained by vector addition of the quadrature I and Q relative voltage amplitudes. For blue as an example, vector addition of -0.32 for I and 0.31 for Q gives the resultant amplitude of 0.44 in the C signal. There is no polarity for the C signal because it is an a-c carrier wave with both positive and negative half cycles.

Note that the peak amplitude of 0.44 for blue C signal means that it varies 0.44 units above and below the zero axis of this modulated a-c waveform. For S signal in h, however, the Y amplitude for luminance is included. The result is to shift the C signal variations to the axis of the Y signal amplitude, instead of the zero axis. Then blue in the S signal is shifted to the 0.11 axis, which is the Y amplitude for blue. In the colorplexed S signal for blue, therefore, the positive peak goes up to $0.44 + 0.11$, or 0.55, and the negative peak goes down to $0.44 - 0.11$, or 0.33. The same idea applies to all the color bars shown. The peak amplitude of the C signal indicates saturation while its average level indicates Y amplitude for luminance.

In the adder section at the transmitter the modulated 3.58-Mc chrominance signal can then be combined with the luminance signal. Also, color sync deflection sync, and blanking are added to produce the complete colorplexed video signal S shown in Fig. 24 · 12. This oscilloscope photo is similar to Fig. 24 · 11h but the color bars are in a different order. The sync

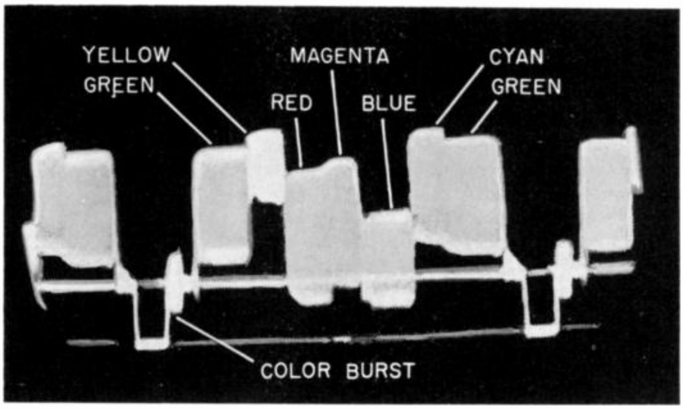


Fig. 24 · 12 Oscillogram of colorplexed total composite video signal S.

pulses are shown downward, as on a studio monitor. Although only horizontal sync and blanking pulses are shown here, corresponding to one line of signal, the total video signal voltage S includes the standard blanking and synchronizing voltage waveform for horizontal and vertical scanning, as in monochrome television.

The colorplexed video signal is transmitted to the receiver as the envelope of the amplitude-modulated picture carrier wave in the broadcast station's assigned channel, along with the associated FM sound signal on a separate carrier separated by 4.5 Mc from the picture carrier frequency, as in monochrome transmission. The chrominance signal is compressed to prevent overmodulation by saturated colors.

At the receiver, the amplitude-modulated picture carrier signal is rectified in the video detector, to recover the colorplexed total composite video signal. In monochrome receivers, the 3.58-Mc color subcarrier in the video signal has practically no effect because there are no color demodulators to detect the chrominance signal. The video-frequency bandwidth is usually less than 3.58 Mc, attenuating the color subcarrier. Also, the chrominance interlace minimizes 3.58-Mc interference by cancellation. However, the required luminance signal is present as the variations in average level of the 3.58-Mc color signal.

In color receivers, the variations in average level provide the luminance signal for the Y video amplifier, while the color demodulators detect the 3.58-Mc chrominance signal to recover its modulation information corresponding to the color video signals. Then the color information is reproduced on the black-and-white picture of the luminance signal.

24·18 Vector addition of color signals

The method of vector addition for two quadrature color video signals is the same as combining two a-c voltages 90° out of phase in a series circuit. For the example of blue C signal, with -0.32 for I and 0.31 for Q,

$$C = \sqrt{(-0.32)^2 + (0.31)^2} = \sqrt{0.102 + 0.096}$$
$$C = \sqrt{0.1980} = 0.44$$

This method can be used to calculate the C values for all the color bars in Fig. 24·11, by vector addition of the I and Q amplitudes.

Just as in an a-c circuit with quadrature voltages, the phase angle of the resultant vector can also be calculated. For the example of blue,

$$\theta = \arctan \frac{Q}{I} = \arctan \frac{0.31}{-0.32} = \arctan - 0.966$$

For a negative tan, $\theta = 180° - 44° = 136°$

The $+Q$ and $-I$ values mean the angle is in the second quadrant, which is why 44° is subtracted from 180°. This angle of 136° is with respect to the I axis, as in Fig. 24·7.

The hue angle is often compared with sync phase. To convert to the reference phase of color sync, add 57°, as this is the angle between I phase and sync phase. When we add 57° to 136°, the sum of 193° shows blue is practically opposite the yellow-green phase of color sync. In general, the complementary colors such as yellow and blue, red and cyan, or green and magenta have the same amplitude of C signal but opposite phase angles differing by 180°. For any color, its combination of I and Q signals determines the resultant phase angle to indicate its hue. The hues of different phase angles are shown in color in Plate X.

SUMMARY

The following alphabetical glossary of terms summarizes the main points of the color television system.

$B - Y$ SIGNAL. Color-difference video signal close to blue. Contains narrow-band information up to 0.5 Mc, approximately.

CHROMINANCE SIGNAL. Is 3.58-Mc color subcarrier with quadrature modulation by I and Q color video signals. Numerically, $C = \sqrt{I^2 + Q^2}$.

COLORPLEXER. Circuit that adds Y video signal to 3.58-Mc modulated chrominance signal. The result is total colorplexed composite video signal that can be transmitted to receiver as amplitude modulation of picture carrier.

COMPATIBILITY. Ability of monochrome receiver to use Y signal for picture in black and white. Also allows color receiver to reproduce monochrome picture. Compatibility results from transmission of Y signal for luminance and use of essentially the same scanning standards for color and monochrome.

DEMODULATOR. Detector circuit for modulated 3.58-Mc chrominance signal. Receiver has two color demodulators to detect two different phases of color video signal.

FREQUENCY INTERLACE. Process of placing harmonic frequencies of chrominance signal midway between harmonics of the horizontal scanning frequency f_H. Is accomplished by making color subcarrier frequency exactly 3.579545 Mc, which is an odd multiple of one-half f_H.

HUE. Dominant wavelength of color, usually in millimicron (10^{-9} meter) units. Different hues of picture information are phase-angle variations of the modulated 3.58-Mc chrominance signal.

I SIGNAL. Color video signal transmitted as amplitude modulation of 3.58-Mc chrominance subcarrier. Contains wide-band information up to 1.5 Mc, for orange and cyan colors.

LUMINANCE. Also brightness, for either color or monochrome information. Different luminance values of the picture information are in the Y video signal, which varies the average-value axis of the 3.58-Mc chrominance signal.

MATRIX. Circuit to combine signals in specific proportions for desired output. Transmitter matrix provides Y, I, and Q video signals in the output with R, G, and B input. In the receiver, usually a three-gun kinescope is the matrix for color video and Y signals in the input to produce red, green, and blue light output.

PRIMARY COLORS. Red, green, and blue. These provide R, G, and B video signal voltages from primary color camera tubes. Opposite voltage polarities are the complementary colors cyan, magenta, and yellow.

Q SIGNAL. Color video signal that modulates 3.58-Mc subcarrier in quadrature with I signal. Q signal contains narrow-band information up to 0.5 Mc for green and magenta colors.

$R - Y$ SIGNAL. Color-difference video signal close to red. Contains narrow-band information up to 0.5 Mc.

SATURATION. A fully saturated color has no dilution by white. Different saturation values in the color picture information are amplitude variations of the modulated 3.58-Mc chrominance signal.

SYNC BURST. Is 8 to 11 cycles of 3.58-Mc color subcarrier, transmitted on back porch of every horizontal pulse. Needed to sync the 3.58-Mc color subcarrier oscillator of receiver to establish exact phase for correct hue in the demodulated color video signals. The hue of color sync phase is yellow-green.

WHITE. Contains red, green, and blue in the proportions $Y = 0.30R + 0.59G + 0.11B$.

SELF-EXAMINATION *(Answers at back of book.)*

Choose (*a*), (*b*), (*c*), or (*d*).

1. Brightness variations of the picture information are in which signal? (*a*) I; (*b*) Q; (*c*) Y; (*d*) $R - Y$.
2. The hue 180° out of phase with red is: (*a*) cyan; (*b*) yellow; (*c*) green; (*d*) blue.
3. Greater peak-to-peak amplitude of the 3.58-Mc chrominance signal indicates more: (*a*) white; (*b*) yellow; (*c*) hue; (*d*) saturation.
4. The interfering beat frequency of 920 kc is between the color subcarrier and: (*a*) associated sound; (*b*) picture carrier; (*c*) lower adjacent sound; (*d*) upper adjacent picture.
5. The hue of color sync phase is: (*a*) red; (*b*) cyan; (*c*) blue; (*d*) yellow-green.
6. Which signal has color information for 1.5 Mc bandwidth? (*a*) I; (*b*) Y; (*c*) $R - Y$; (*d*) $B - Y$.
7. Which of the following is false? (*a*) I video hues are orange or cyan. (*b*) The transmitter matrix output includes Y, I, and Q video. (*c*) A three-gun kinescope can serve as a matrix. (*d*) A fully saturated color is mostly white.
8. Which of the following video signals can reproduce the finest detail? (*a*) $R - Y$; (*b*) Q; (*c*) I; (*d*) Y.
9. What is the hue of a color 90° leading sync burst phase? (*a*) Yellow; (*b*) cyan; (*c*) blue; (*d*) orange.
10. The average voltage value of the 3.58-Mc modulated chrominance signal: (*a*) is zero for most colors; (*b*) is close to black for yellow; (*c*) corresponds to the brightness of the color; (*d*) corresponds to the saturation of the color.

ESSAY QUESTIONS

1. What is meant by color addition? Name the three additive primary colors.
2. What color corresponds to white light minus red? White minus blue? White minus green?
3. Why are the primary color video voltages converted to Y and C signals for broadcasting? What are the two components of the C signal?
4. Define hue, saturation, luminance, and chrominance.
5. State the percentages of R, G, and B primary video signals in the following signals: Y, I, Q, $R - Y$, $B - Y$, $G - Y$.
6. What is the video-frequency bandwidth of the Y signal? I signal? Q signal?
7. What parts of the picture are reproduced in black and white by the Y signal? What parts are reproduced in full color as mixtures of red, green, and blue? What parts can be reproduced only in orange or cyan?
8. What hues correspond to the following: $+I$, $-I$, $+Q$, $-Q$, $R - Y$, $B - Y$, $-(R - Y)$, $-(B - Y)$, color sync burst?
9. How is the 3.58-Mc modulated chrominance signal transmitted to the receiver? Why is the 3.58-Mc signal called a subcarrier?
10. Describe the color sync burst signal and give its purpose.
11. How does the 3.58-Mc modulated chrominance signal indicate hue, saturation, and luminance of the picture information?
12. Why is the chrominance signal transmitted with the subcarrier suppressed?
13. Why is the color subcarrier frequency made exactly 3.579545 Mc?
14. State the horizontal and vertical scanning frequencies for color television broadcasting.
15. Illustrate a matrix circuit for forming the I signal. No values required.
16. Referring to the hue phase diagram in Fig. 24 · 10: (*a*) What is the phase angle with re-

spect to color sync burst for the C vector shown? (*b*) What is the hue of this C signal? (*c*) What is the approximate phase angle for green hues?

17. A scene displays a wide yellow vertical bar of 50 per cent saturation against a white background. How will this picture appear in a monochrome reproduction?
18. What is the effect of the chrominance signal in a monochrome receiver with video-frequency response up to 3 Mc?

PROBLEMS *(Answers to odd-numbered problems at back of book.)*

1. Calculate the relative voltage values of R, G, B, and Y signals for blue 60 per cent saturated.
2. Calculate the voltage values in Fig. 24·4 of Y signal for all colors 60 per cent saturated.
3. Calculate the relative voltage values of saturated blue and saturated yellow for (*a*) $R - Y$; (*b*) $B - Y$; (*c*) $G - Y$ video signals.
4. Prove that if $G - Y$ is $[-0.51\ (R - Y)\ -0.19\ (B - Y)]$ then this equals $-0.30R + 0.41G - 0.11B$.
5. Give the R, G, B, I, Q, and C values for white.
6. Redraw the color vectors in Fig. 24·6*a*, *b*, *c*, and *d*, for full saturation and 20 per cent saturation.
7. Calculate the value of C voltage when $I = -0.4$ and $Q = -0.3$; what is the approximate hue of this color?
8. A bar pattern includes white, blue, and yellow, with 60 per cent saturation. Draw the waveforms with relative voltage values for R, G, B, Y, I, Q, and C signals.
9. For the magenta color bar in Fig. 24·11, calculate the hue phase, with respect to color sync.
10. Do the same as in Prob. 9 for all the color bars in Fig. 24·11.

Chapter 25 Color television receivers

The color reproduction is a monochrome picture on a white raster, with full color added for relatively large areas corresponding to video signal frequencies up to 0.5 Mc. Therefore, the color receiver includes all the circuits of a monochrome receiver, plus the chrominance signal circuits for color. As illustrated in Fig. 25 · 1, the r-f tuner selects the desired station and converts the r-f signal frequencies to the receiver i-f signal frequencies. The i-f amplifier provides enough signal for the video detector. Up to here, the signal waveform contains the colorplexed total video signal voltage S as the amplitude-modulation envelope of the picture carrier. Then the video detector demodulates the AM carrier, as in monochrome receivers. However, the detected output includes two video signals: the luminance signal Y, which is the same as a monochrome signal, plus the 3.58-Mc two-phase modulated chrominance subcarrier signal. This 3.58-Mc subcarrier signal must be detected by the color demodulators to recover the original color video signals.

The main difference in the r-f and i-f circuits is the greater importance of bandwidth for color receivers. Remember that the side frequencies around 3.58 Mc, corresponding to high video frequencies, are essential for the color information in the picture. Without the 3.58-Mc chrominance signal, there is no color—just a monochrome picture. For this reason the fine tuning control on the front end must be tuned exactly for the required response for the chrominance signal to put the color in the picture.

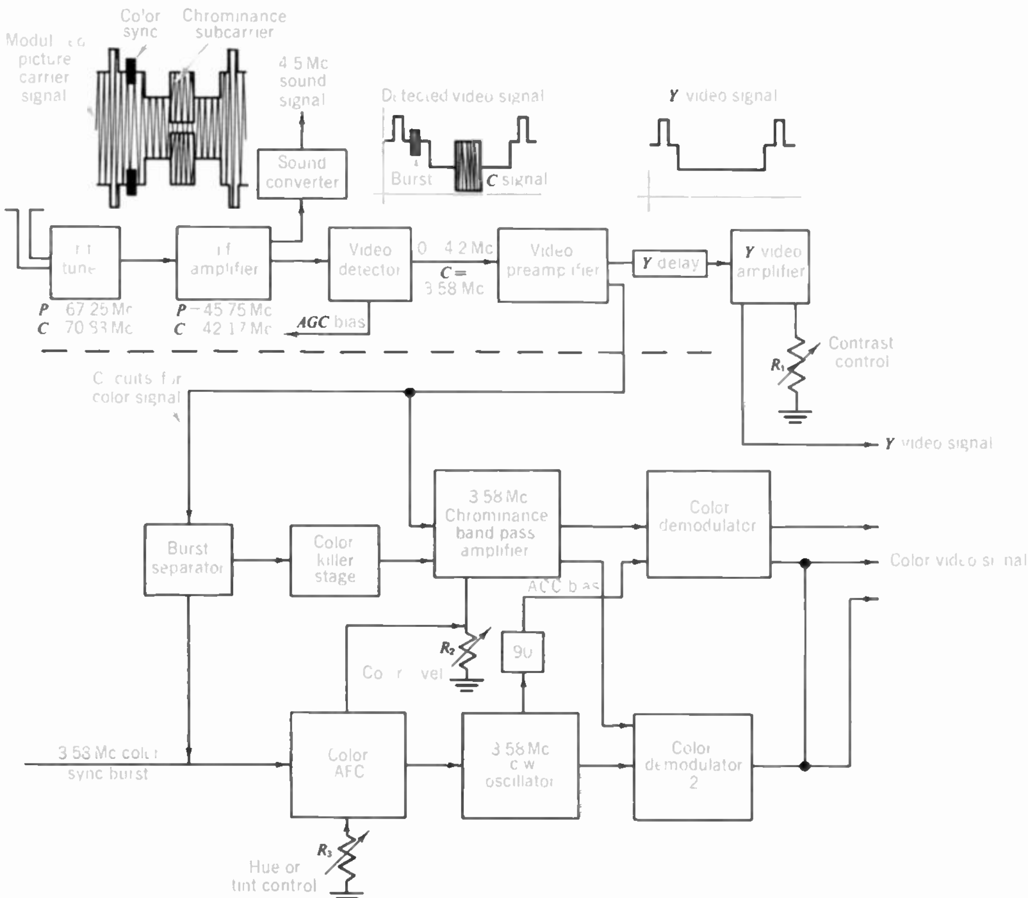
25 · 1 Chrominance section

In Fig. 25 · 1, the video circuits are divided into two parts, one for chrominance and the other for luminance. The Y video amplifier is for the luminance signal only. Its bandwidth is restricted to 3.2 Mc, approximately, to minimize interference from the 3.58-Mc chrominance signal. Note the delay line, which is usually 1 ft of coaxial line for 1 μsec delay of the *Y* video signal. This is necessary to equalize the time delay for the luminance and chrominance circuits before the signals are combined in the matrix.

The *Y* video amplifier determines the amount of luminance signal for the kinescope, which controls the contrast of the reproduced picture. Therefore, the contrast control R_1 is used to vary the amplifier gain. This function is the same as in monochrome receivers.

Chrominance amplifier. This amplifier is fixed-tuned, like an i-f stage, for the 3.58-Mc chrominance signal in all receivers for any station. The

Fig. 25·1 Main sections of color television receiver. Picture carrier frequency indicated for channel 4. Signal waveshapes for vertical color bar down center of gray background. Deflection, sync, sound, and power supply circuits not shown.



input includes total detected video signal but the output is only 3.58 Mc modulated chrominance signal.

The gain of the C amplifier determines how much chrominance signal is coupled to the color demodulators. More chrominance signal means more detected color video signal voltage for the kinescope. More color signal increases the saturation of the color reproduction. Therefore, R_2 varies the C amplifier gain, functioning as a color level control or color saturation control. This stage is often called the *color amplifier* since its gain determines the amount of color in the picture.

ACC bias. Automatic color control (ACC) is generally used to vary the gain of the chrominance amplifier in accordance with the strength of the color burst signal. The idea is the same as automatic gain control, increasing the negative bias on the color amplifier to reduce overload distortion for strong color signals.

Color demodulators. Two color demodulators are necessary to detect two different phases of the modulated 3.58-Mc chrominance signal. Therefore, the output of the chrominance amplifier is coupled in parallel to the two color demodulators. They both have the same C signal input. However, the demodulators have different phases of input voltage from the 3.58-Mc color oscillator, which is the reinserted color subcarrier. The oscillator output is a continuous wave (cw) without any modulation. This voltage beats with the side frequencies in the modulated chrominance signal to recover the color video signals. The phase of the oscillator cw into the demodulator determines the hue of its detected color video signal output. This is why the color demodulator circuits are called *synchronous detectors.* In many receivers one oscillator cw phase is $B - Y$ and the quadrature phase is $R - Y$, to provide $B - Y$ and $R - Y$ color video signals in the output of the demodulators.

Color video signal circuits. With $R - Y$ and $B - Y$ color video signals, these two can be combined to form $G - Y$. The three video signals are matrixed with the Y signal in the kinescope to reproduce the color information in red, green, and blue.

The color video signal output from the demodulators includes the video-frequency range from 0 cps, or direct current, up to the top correction frequency. Therefore, color video signals after the demodulators are amplified in resistance-coupled video amplifier stages. Before the demodulators, tuned amplifiers are necessary for the 3.58-Mc modulated chrominance signal.

Additional video amplifiers may be needed for $R - Y$, $B - Y$, and $G - Y$ signals to provide enough voltage for the kinescope, when low-level demodulators are used. Or, two 3.58-Mc amplifiers can be used to drive the demodulators. In this high-level demodulation, the detected output has enough amplitude for the color video signals to drive the kinescope directly.

Assuming a three-gun tube in an $R - Y$, $B - Y$ receiver, the control grid of each gun has $R - Y$, $B - Y$, or $G - Y$ video signal voltage. All three cathodes have $-Y$ signal voltage. The net result cancels the Y component of the color video signals, as the cathode and grid voltages have opposite effects on the beam current. Effectively, therefore, the three guns have red, green, or blue video signal voltage in each grid-cathode circuit, to reproduce the color information on the kinescope screen. This matrixing applies for color video frequencies up to 0.5 Mc. For higher video frequencies, only the Y signal information is present. Then the luminance of fine details is reproduced in monochrome.

Color synchronization. The 3.58-Mc color sync burst is on the back porch of every horizontal sync pulse. To obtain the color sync alone, the total video signal S is coupled into the burst separator in Fig. 25·1. This stage is a 3.58-Mc amplifier, tuned to the color subcarrier frequency, but keyed on during horizontal flyback time only. During trace time, the burst separator is held cut off. Therefore, plate current flows for 3.58-Mc signal only for the color sync burst during flyback time.

Although the burst separator is a tuned amplifier for 3.58 Mc, like the chrominance amplifier, they have opposite times for conduction. The burst separator is on during flyback time and off during trace time. The chrominance amplifier is on during trace time to supply the chrominance signal needed for the color demodulators.

The amplified output of the burst separator supplies the required color synchronizing voltage for the 3.58-Mc color subcarrier oscillator. Although the oscillator has a 3.58-Mc crystal for frequency stability, its phase must be accurately controlled for correct hues in the picture. Therefore, the color sync voltage goes to an AFC circuit, which in turn controls the oscillator phase. The color AFC circuit holds the hue of the reproduced colors at the correct values, after the manual *hue* or *tint* control has been set.

Color killer stage. This name describes its function, which is to cut off the chrominance amplifier for programs broadcast in monochrome. Then there is no signal into the color demodulators and no output from the color video circuits. The purpose is to prevent tube noise in the chrominance amplifier from being amplified. This noise can produce color snow, called *confetti*, in the picture. For a color program the color killer stage is cut off, however, allowing the color amplifier to supply color signal to drive the demodulators.

Tuning in the color in the picture. For a program that is being broadcast in color, the fine tuning control must be set to provide chrominance signal for the receiver. Without the r-f side bands corresponding to the 3.58-Mc color signal, the picture is reproduced in monochrome by the luminance signal. Adjusting the fine tuning control, the color usually is maximum just off the point where 920-kc sound interference is in the picture. The color or saturation control must be up, as usually there is no chrominance signal at the minimum level. It is preferable to adjust for just a little color. Too much saturation usually makes the colors look artificial. In addition, amplitude distortion of the chrominance signal can distort the color values. Finally, the hue control is set for natural tints. This can be determined by observing flesh tones of people in the scene. Also check any recognized hues, such as green grass and a blue sky.

25·2 Chrominance amplifier circuits

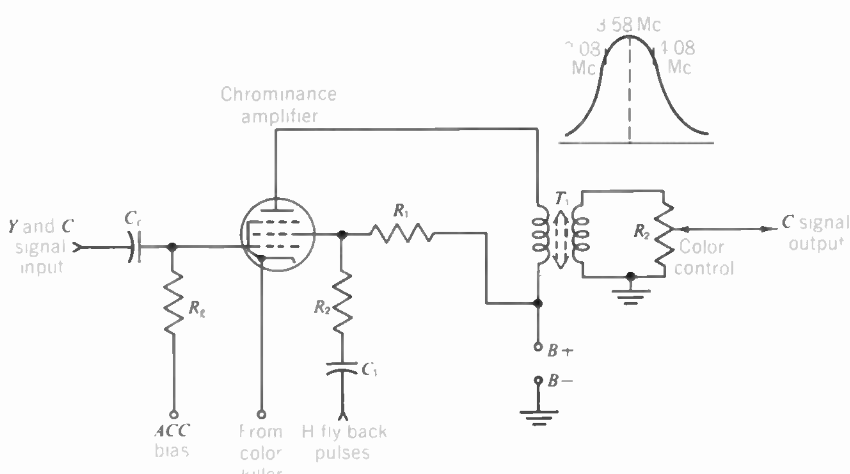
In Fig. 25·2, the input consists of Y and C video signals but because of the tuned plate load, the output is only 3.58 Mc chrominance signal. The color subcarrier frequency of 3.58 Mc to which this section is tuned remains the same for all stations in any receiver. In fact, we can consider 3.58 Mc as the color intermediate frequency for the color circuits.

Input signal for the chrominance amplifier is obtained from the video amplifier or video detector. The chrominance signal output goes to the color demodulators or to an additional color amplifier to drive the demodulators. A second chrominance amplifier is often called the *demodulator driver stage.*

Requirements of the chrominance amplifier. The following features are important for this section of the receiver, which is the main factor in providing the required amount of color signal.

1. The amplifier is tuned to 3.58 Mc with the required bandwidth. For $R - Y$, $B - Y$ demodulators supplying 0.5-Mc color video, the bandwidth of 1 Mc or ± 0.5 Mc is sufficient. This response is provided by T_1 in Fig. 25·2.
2. The output level of the chrominance amplifier determines the amount of saturation in the picture. Either the gain or the amount of signal-voltage output can be varied. R_2 in Fig. 25·2 varies the amount of chrominance signal output. This must be a low-resistance control to minimize variations in phase angle, for the high video frequencies.
3. Automatic color control (ACC) bias is generally used to control the

Fig. 25·2 Basic requirements of chrominance amplifier circuit.



gain of the chrominance amplifier. This arrangement is essentially the same as AGC bias, but only for 3.58-Mc chrominance signal.

4. When no color signal is being received, the chrominance amplifier is cut off by excessive bias from the color killer. In Fig. 25 · 2 the cathode voltage can be made positive enough by the color killer to cut off plate current in the chrominance amplifier.
5. In addition, internal horizontal blanking is necessary to make sure that the burst cannot produce a yellow-green vertical bar at the left side of the picture. In Fig. 25 · 2, negative *H* flyback pulses at the screen grid key off the chrominance amplifier. During horizontal flyback time, then the amplifier cannot conduct. Also, in Fig. 25 · 15, *H* blanking pulses key off the color demodulators; in Fig. 25 · 27, the blanking amplifier keys off the second video amplifier.

Two-stage chrominance amplifier. In Fig. 25 · 3, V_1 and V_2 form a staggered single-tuned pair, with an overall bandpass of 1 Mc centered at 3.58 Mc. The chrominance signal output drives the $R - Y$, $B - Y$ demodulators, which require a bandwidth of ± 0.5 Mc. Input signal for V_1 is taken from the video detector. The input signal includes luminance and chrominance information but the plate load L_{25} is tuned for chrominance signal output. Note the 4.5-Mc trap in the input to reduce 920-kc beat between the sound signal at 4.5 Mc and the chrominance signal at 3.58 Mc.

Fig. 25 · 3 Two-stage chrominance amplifier circuit. C values more than 1 in μμf. M is megohms. (Zenith chassis 29JC20.)

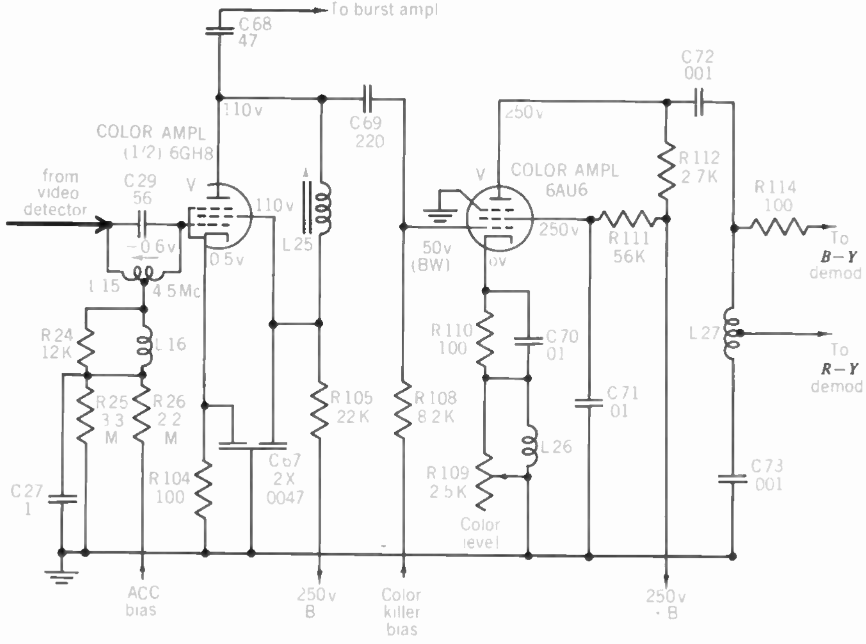
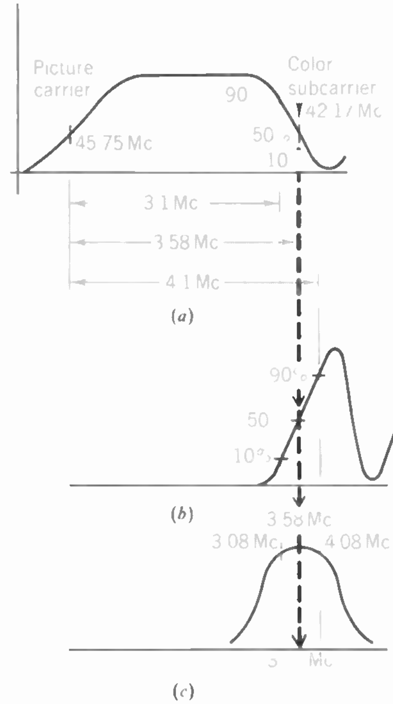


Fig. 25·4 *Sloping response curve used for some chrominance amplifiers. (a) Response of i-f section. (b) Opposite response of chrominance section. (c) Net result for symmetrical chrominance response centered at 3.58 Mc.*



The output signal of V_1 goes in two paths. C_{68} couples chrominance signal to the burst amplifier for color synchronization. In addition, C_{69} supplies input signal to the second chrominance amplifier V_2. The tuned plate load for V_2 is L_{27}, which supplies chrominance signal to the demodulators. Cathode bias is provided by R_{110} with R_{109} varying the stage gain for color control. The small inductance L_{26} minimizes phase-shift variations as R_{109} is varied. Note that the d-c voltages for V_2 are indicated for monochrome reception, with -50 volts at the control grid from the color killer to cut off the chrominance amplifier when no color signal is received.

Chrominance band-pass response. The overall bandwidth required is indicated by a chrominance i-f response curve. As shown for T_1 in Fig. 25·2, its response is centered at 3.58 Mc, with ± 0.5 Mc bandwidth.

Some $R - Y$, $B - Y$ receivers use the sloping response characteristics illustrated in Fig. 25·4. The overall r-f, i-f response curve for the receiver in *a* attenuates chrominance side frequencies near the sound i-f carrier. This feature results in less beat-frequency interference between the sound and chrominance signals. Also, the reduced i-f bandwidth allows more gain. However, the chrominance response curve in *b* has the opposite slope to complement the i-f gain for chrominance signal. The net result for chrominance signal into the demodulators, therefore, is the symmetrical response in *c*, centered at the color subcarrier frequency of 3.58 Mc.

Note the continuity of signal frequencies for color. In the i-f response in Fig. 25·4*a*, the color subcarrier frequency is 42.17 Mc because this is 3.58 Mc below the i-f picture carrier at 45.75 Mc. The chrominance side frequencies are above and below 42.17 Mc. In *b*, the chrominance subcarrier frequency is 3.58 Mc after the modulated picture signal has been detected in the video detector. Here the chrominance side frequencies are above and below 3.58 Mc. Similarly, the subcarrier in *c* is 3.58 Mc in all parts of the chrominance circuits before the color demodulation.

After detection of the chrominance signal, the subcarrier frequency corresponds to 0 cps, or direct current in the demodulator output. Also, the chrominance side frequencies become color video frequencies corresponding to the difference from 3.58 Mc. For example, as the chrominance side frequencies of 3.08 and 4.08 Mc beat with the reinserted subcarrier of 3.58 Mc, the resulting color video frequency is 0.5 Mc.

The effect of narrow bandwidth for the chrominance i-f response curve is reduced color detail in the picture. Insufficient gain for the chrominance signal causes weak color.

25 · 3 Burst amplifier circuits

The reference phase for hue is transmitted as a burst of 8 to 11 cycles of the 3.58-Mc subcarrier on the back porch of each horizontal sync pulse (see Fig. 25 · 5). There is an important difference between the two 3.58-Mc waveforms shown. The burst on the back porch is cw at 3.58 Mc, unmodulated, with fixed phase for a timing reference. The 3.58-Mc *C* signal occurs during trace time, with two-phase modulation to convey chrominance information, although the varying modulation cannot be seen in the drawing.

In order to utilize the phasing information in the color sync burst it must first be separated. This function is accomplished by a stage having the following requirements:

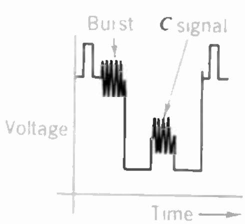
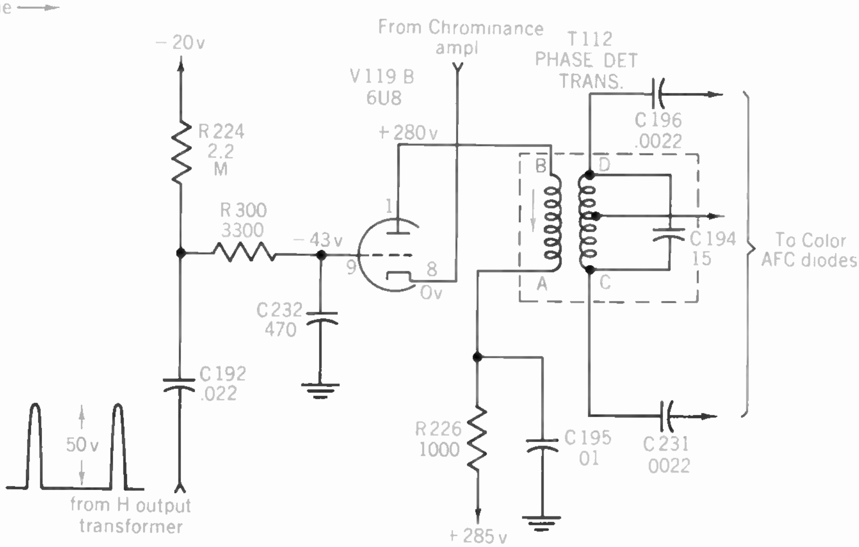


Fig. 25 · 5 The 3.58-Mc color burst on the back porch of horizontal sync during flyback time.

Fig. 25 · 6 Burst separator circuit. C values more than 1 in μμf. M is megohms. (RCA chassis 21CT662.)



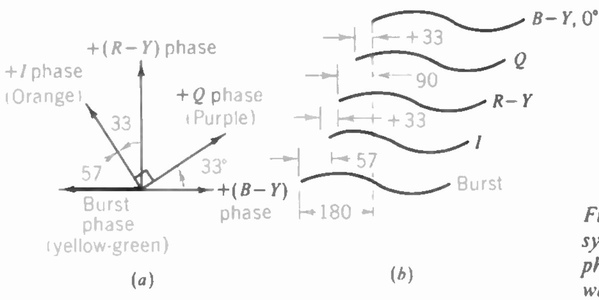


Fig. 25·7 Hue values in terms of color sync phase. (a) Vector diagram showing phase angles. (b) Phase angles for one sine wave.

1. Tuned to 3.58 Mc to amplify the burst signal.
2. Keyed on for conduction during horizontal flyback time only, which is when the burst occurs.
3. Cut off during horizontal trace time to prevent the 3.58-Mc chrominance signal from interfering with color synchronization.

With these functions, the stage is generally called a *burst amplifier, burst separator,* or *burst keyer.* In any case, its function is to separate the burst, amplify this 3.58-Mc reference signal, and couple the output to the color AFC circuit, as shown in Fig. 25·6.

In Fig. 25·6, the chrominance signal input is applied to the cathode of the triode V_{119B}. The control grid is used for positive keying pulses from a winding on the horizontal output transformer. These pulses key the tube into conduction for the burst signal. Grid current produces grid-leak bias of −23 volts, which, added to the fixed bias of −20 volts, results in −43 volts grid bias. Therefore, the tube is cut off except for the time of each flyback pulse, as its 50-volt peak is enough to overcome the negative bias. R_{300} and C_{232} in the grid circuit delay and shape the flyback pulses so that the peak occurs at the time of sync burst. In the plate circuit, T_{112} is tuned to 3.58 Mc. The output is 3.58-Mc signal, therefore, but only during flyback time. The amplified output then is color sync burst, without the chrominance signal.

25·4 Color AFC circuits

The function of the color sync burst is to hold the 3.58-Mc subcarrier oscillator at the desired phase with respect to burst, for correct hues in the picture. The phase angles of different hues are compared with the yellow-green hue of burst phase in Fig. 25·7, to illustrate how phase angle determines hue. It follows that if the 3.58-Mc color reference oscillator is not locked in by the sync burst, the detected hues will change continuously, as the oscillator frequency or phase drifts. Then the picture will have horizontal or diagonal color stripes, as shown in Plate IV.

Most receivers use the AFC circuit illustrated in Fig. 25·8. Since the color oscillator is a sine-wave oscillator, it must be controlled by a reactance tube. The function of each stage is:

1. **Color phase detector.** Detects the difference between color sync

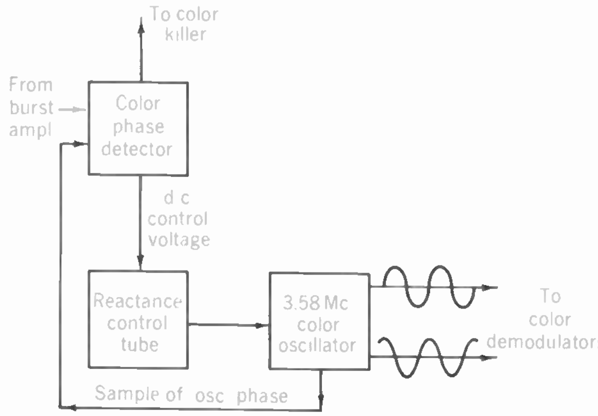


Fig. 25·8 Block diagram of color AFC circuit.

phase and a sample of the color oscillator phase. The required output here is d-c control voltage to vary the bias for the reactance tube.

2. **Reactance tube.** Corrects the phase of the sine-wave oscillator that produces the 3.58-Mc subcarrier reinserted in the synchronous demodulators. This control stage is essentially the same as the reactance tube described for frequency modulation in Sec. 22·4.

3. **Color subcarrier oscillator.** The phase of its cw output determines the hue of the video signal from the synchronous demodulators. The oscillator output circuit provides voltages of different phases for the two color demodulators.

Phase detector circuit. Referring to the complete AFC circuit in Fig. 25·9, the 6AL5 at the top is the color phase detector stage. The input transformer secondary is center-tapped to provide opposite polarities of 3.58-Mc burst signal to the two diodes. Also, input signal from the 3.58-Mc oscillator is applied to the two diodes in the same polarity, through C_{722}. The two input voltages, one burst signal and the other oscillator cw, must be in quadrature. Here, the burst phase transformer T_{702} supplies the required quadrature phase for both polarities of the push-pull input voltages.

For quadrature phase, the input signals make both diodes conduct the same amount. Then the net voltage at test point 701 at the junction of the diode load resistors is zero. Resistors $R_{716\text{-}A}$ and $R_{716\text{-}B}$ are a matched pair to provide equal and opposite d-c output voltages for zero d-c control voltage. In some circuits the diode load resistances are variable, to be adjusted for exact balance.

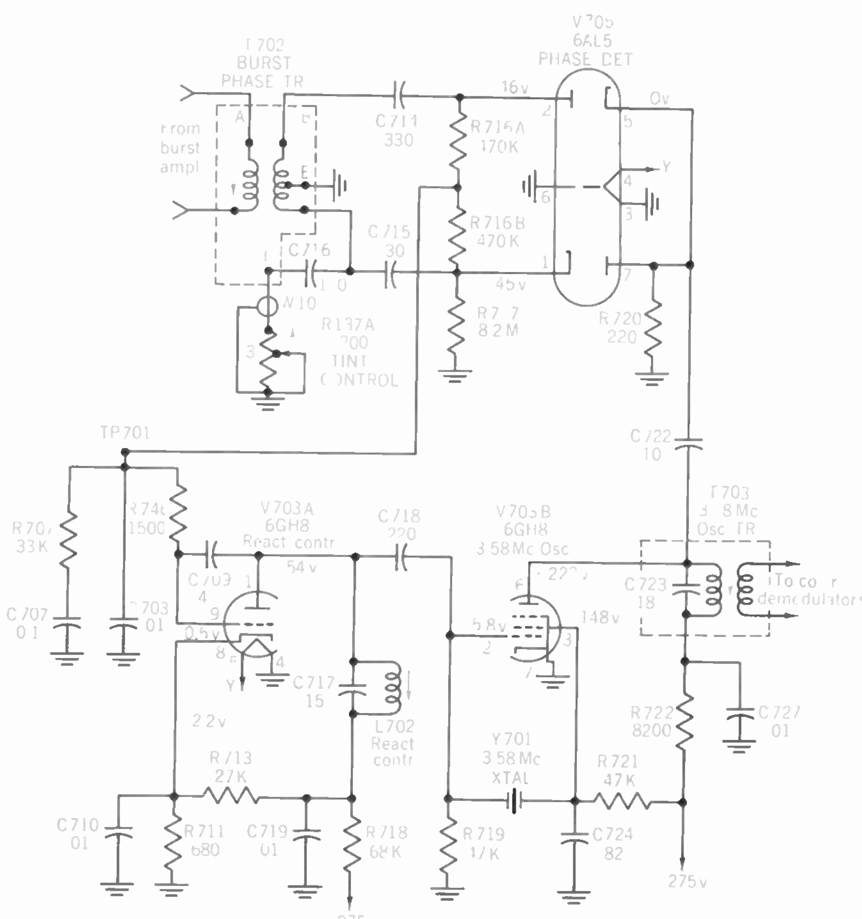
If the phase of the color oscillator drifts, its output voltage for the phase detector will not be exactly in quadature with the burst signal input. Then one diode conducts more than the other, resulting in a net d-c control voltage across C_{703}. This voltage at TP701 is filtered and d-c coupled to the grid of the control tube to vary its reactance. The reactance tube then corrects the color oscillator.

Hue or tint control. In Fig. 25·9, $R_{137\text{-}A}$ in the ground return for T_{702} is the hue or tint control. Varying this resistance makes a slight change in the phase of the entire secondary signal voltage, compared with the 3.58-Mc burst signal in the primary. This phase control enables the viewer to vary the hues. The hue control is adjusted for correct flesh tones in the reproduced picture.

25·5 Automatic color control (ACC) bias

The idea here is the same as an AGC circuit, in that the gain of the 3.58-Mc chrominance band-pass amplifier is automatically controlled by the ACC bias. The voltage is usually taken from one diode of the color phase detector. In Fig. 25·10, the top diode of the 6AL5 supplies negative d-c voltage output proportional to the amount of color sync burst. For no

Fig. 25·9 Typical color AFC circuit. C values more than 1 in μμf. M is megohms. (RCA chassis CT-11.)



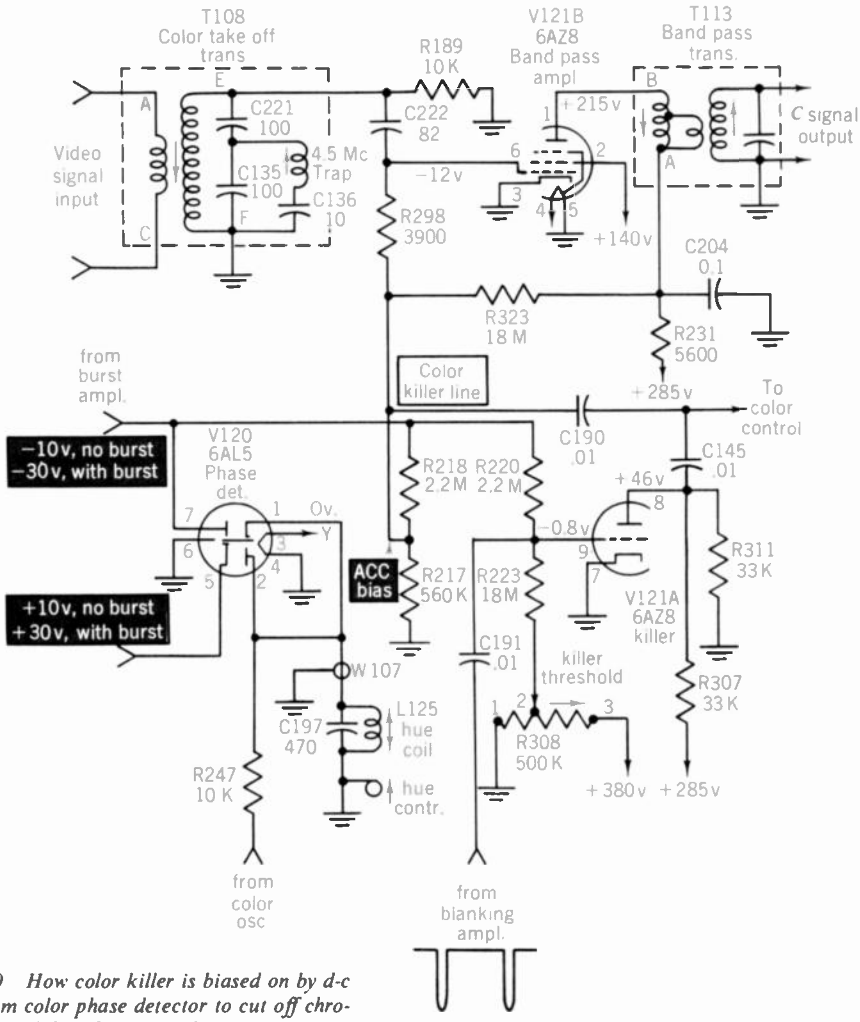


Fig. 25·10 How color killer is biased on by d-c voltage from color phase detector to cut off chrominance amplifier for monochrome reception. C values more than 1 in μμf. M is megohms. (RCA chassis 21CT662.)

signal, or monochrome reception, the d-c voltage at plate pin 7 is −10 volts. With color signal, this d-c voltage can increase up to −30 volts. The voltage divider of R_{217} and R_{218} reduces the d-c output voltage for the ACC bias line to the grid of the color amplifier. Only the voltage across R_{217} is used for the ACC bias.

Note that the negative d-c voltage from the 6AL5 pin 7 is also used for bias on the color killer stage. The entire d-c voltage is d-c coupled through R_{220} to the control grid of the color killer.

These negative d-c voltages for ACC bias and the color killer should not be confused with d-c control voltage for the color AFC circuit. The ACC bias and killer bias are unbalanced output voltages from one diode plate, which is always negative. For color AFC, the phase detector output is a balanced output from the center-tapped input circuit, which is not shown in Fig. 25 · 10. This net d-c control voltage can be zero, positive, or negative.

25 · 6 Color killer circuit

Tube noise in the chrominance signal is reproduced as color streaks, or *confetti*, on the kinescope screen. This is much more noticeable than black-and-white snow because of the colors. Also, the confetti is coarser, as it includes components of lower frequency than snow. The most obvious effect is crosstalk between the color snow and high-frequency components of a monochrome signal, producing sparkling colors at the edges of black-and-white objects in the scene. The confetti is different from steady color fringes, however, which may appear at the edges of all objects in the picture, for both monochrome and color. This edge fringing in colors is a problem of the convergence adjustments in setting up the tricolor kinescope. The color streaks represent noise that should be eliminated by the color killer stage.

Since the color streaks of confetti would be obvious in a monochrome picture, the chrominance signal must be cut off during monochrome reception. For a monochrome picture, the color killer conducts to cut off the color amplifier. For a color picture, the color killer is cut off by rectified burst signal, allowing the color amplifier to operate.

In Fig. 25 · 10 the 6AZ8 stage V_{121A} at the lower right corner is the color killer. Note the following three voltages at control-grid pin 9.

1. Horizontal flyback pulses of negative polarity from a separate blanking amplifier stage. These pulses are coupled by C_{191} to the grid of the killer stage. When V_{121A} conducts these pulses are amplified and inverted to feed positive pulses to the control grid of the color amplifier.
2. Negative d-c voltage from the phase detector circuit. This voltage indicates whether color sync burst is present in the chrominance signal.
3. Positive d-c voltage from the *threshold control.*

The combination of negative and positive d-c voltages at the control grid determines whether the color killer conducts or is cut off. In Fig. 25 · 10, the d-c voltages are shown for no color signal. The -10 volts from the 6AL5 pin 7 is opposed by $+9.2$ volts from the threshold control, resulting in -0.8 volt at the grid. Then the color killer conducts, with almost zero grid bias. The amplified flyback pulses in the plate circuit are coupled by C_{145} and C_{190} to the color amplifier. These pulses cut off the color amplifier during trace time, by means of the grid-leak bias produced by the positive grid pulses. The -12-volt bias shown for V_{121B} is approximately the grid-cutoff voltage, which reduces the gain to practically zero.

The -12-volt bias on the color amplifier is the sum of -2 volts ACC bias, equal to one-fifth the voltage from the 6AL5, added to -10 volts grid-leak bias. This bias voltage results from grid current produced by amplified flyback pulses from the color killer. Although the amplifier is cut off during trace time by -12 volts bias, note that it conducts when the flyback pulses drive the grid positive. Therefore, the color amplifier can supply burst signal during horizontal flyback time, although it is cut off during trace time. This method is necessary here to allow burst signal output when the receiver is switched to a color program. It should be noted, though, that in receivers with two color amplifiers, the second stage can be cut off by the color killer while the first color amplifier can supply burst signal to the phase detector (see Fig. 25 • 3).

With a color program, the burst signal drives the 6AL5 diodes in Fig. 25 • 10 to produce more d-c output voltage. If we take the example of -30 volts at the 6AL5 pin 7, with the same $+9.2$ volts from the threshold control, the net d-c voltage at the killer grid now is -20.8 volts. This voltage is negative enough to cut off the killer stage. Then it cannot conduct, allowing operation of the color amplifier. Without the grid-leak bias of the flyback pulses, the grid bias of the color amplifier is about -6 volts from the ACC line, for class A operation in amplification of the chrominance signal.

Color threshold adjustment. This is R_{308} at the bottom right in Fig. 25 • 10. It must be set to the point where negative bias from the phase detector can cut off the color killer during color reception, but allow the killer to conduct to cut off the color amplifier for monochrome. The procedure is as follows:

1. Tune the receiver to a strong black-and-white station.
2. Adjust the threshold control until color streaks just disappear from the picture.
3. Check on a color program to make sure that color appears.

Incorrect setting of the threshold control can cause the trouble of no color on a color program. When the bias on the color killer is set correctly, though, the receiver automatically reproduces color for a color program as the rectified burst signal cuts off the killer. The operator need not adjust the receiver for color, assuming the fine tuning is set right and the color level control is not at zero.

Color on-off switch. Some receivers have a manual switch for the color killer stage, instead of the threshold adjustment. The switch is at the rear of the color level control. When the switch is on, it cuts off the color killer.

25 • 7 Color demodulator circuits

The chrominance signal output from the band-pass amplifier consists of the side bands produced by modulation of the 3.58-Mc color subcarrier. In order to recover the color video modulation, the chrominance signal must be detected. Remember that the chrominance modulation is an AM signal, but two different color video signals modulate the subcarrier in quadrature phases of the color subcarrier. Therefore, two separate demodulators are necessary. Furthermore, the 3.58-Mc subcarrier must be reinserted so that the chrominance side bands can beat with the subcarrier to detect

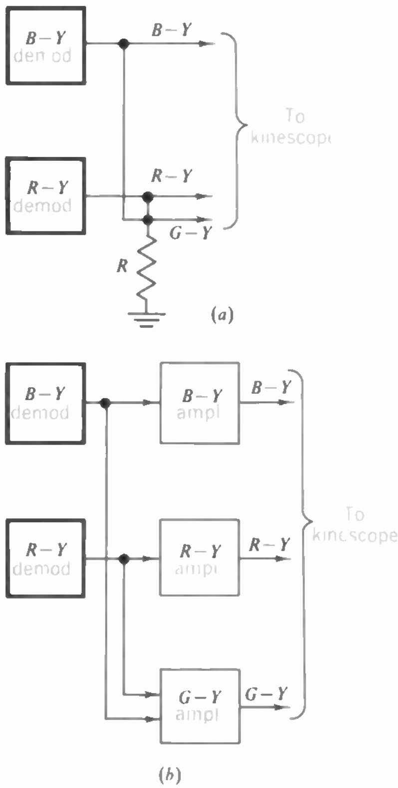


Fig. 25·11 Color video signal output from demodulators, illustrated for R − Y, B − Y receiver. (a) Direct coupling to kinescope. (b) Amplified in color video amplifiers.

the color video as the difference frequencies between the side bands and the subcarrier. Since the detected output depends on the frequency and phase of the reinserted subcarrier, the demodulator is called a *synchronous detector.*

The two color demodulators have the same modulated chrominance signal input from the color amplifier. However, each has a different phase of oscillator cw from the color subcarrier oscillator. The output from the two demodulators consists of the two desired color video signals. Any two color video signals can be detected by using the appropriate phase of oscillator cw in the demodulators. The detected output in some receivers is amplified by one video stage for each color to provide enough color video signal to drive the kinescope grid-cathode circuit. The required amplitude is about 80 volts peak to peak. In other receivers, the demodulator output voltage has enough amplitude to drive the kinescope directly. These two methods are illustrated in Fig. 25·11.

Basic circuit. The circuit in Fig. 25·12 illustrates a $B-Y$ demodulator, as a specific example. Note that the phase of oscillator cw into the demodulator is shown the same as the phase of the C signal for blue. This phase is opposite from yellow. There is enough oscillator voltage input to make the stage operate as a nonlinear amplifier, providing rectified output in the plate circuit. A tube with two control grids is used, so that both signal input voltages can control the plate current. However, a triode can be used with one signal at the control grid and the other at the cathode.

The output of the demodulator is no longer 3.58-Mc sine-wave voltage but is a color video signal. Its waveshape depends on the chrominance modulation. This voltage corresponds to the envelope of the 3.58-Mc modulated chrominance signal when its phase is the same as or opposite from the oscillator cw phase. Quadrature phase has no effect on the output. In Fig. 25·12 the color video modulation represents blue, white, and yellow bars. Maximum plate current flows for blue video, as the C signal and oscillator cw then have the same phase. The maximum IR drop across

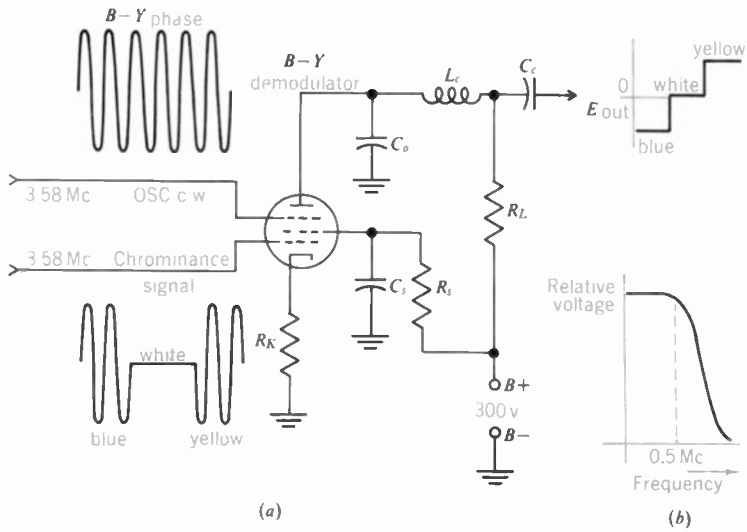


Fig. 25·12 Synchronous detector for B − Y phase. Voltage waveshapes shown for blue, white, and yellow bars. (a) Circuit with series peaking for video signal output. (b) Frequency response curve.

the plate load resistor R_L results in minimum plate voltage. For yellow hue with opposite phase, the plate current is minimum and the output voltage maximum. Therefore, the demodulator output has its peak values of video voltage for the hues on the axis of its oscillator cw input.

Note the series peaking coil L_c for high-frequency compensation, to provide the desired frequency response for the color video signal voltage. In this example, the top correction frequency is 0.5 Mc. A typical value of R_L for this frequency response is about 20,000 ohms. The cutoff at 0.5 Mc filters out the 3.58-Mc subcarrier component of the rectified signal.

Synchronous waveshapes. Figure 25·13 illustrates how the amount of output from the synchronous detector depends on the phase between the *C* signal and oscillator cw. Assume first that there is oscillator voltage but no *C* signal, as in *a*. Half cycles of 3.58-Mc sine-wave current flow in the plate circuit. The negative half cycles are beyond cutoff, as in a class B amplifier. However, the half cycles of direct plate current have an average value, taken here at 10 ma. It is this average value of I_b that will vary with the modulation information of the *C* signal, to provide the desired color video signal output.

In *b*, the *C* signal of the same phase as the oscillator cw aids the flow of plate current for one-half cycle. Then the average I_b increases to 12 ma. In *c*, the *C* signal of opposite phase decreases the average I_b to 8 ma. This decrease is still a desired signal variation, as it just indicates an opposite change with respect to the no-signal axis of 10 ma.

The example in *d* with quadrature *C* signal indicates no detected output because the average I_b is the same 10 ma as when there is no *C* signal. The

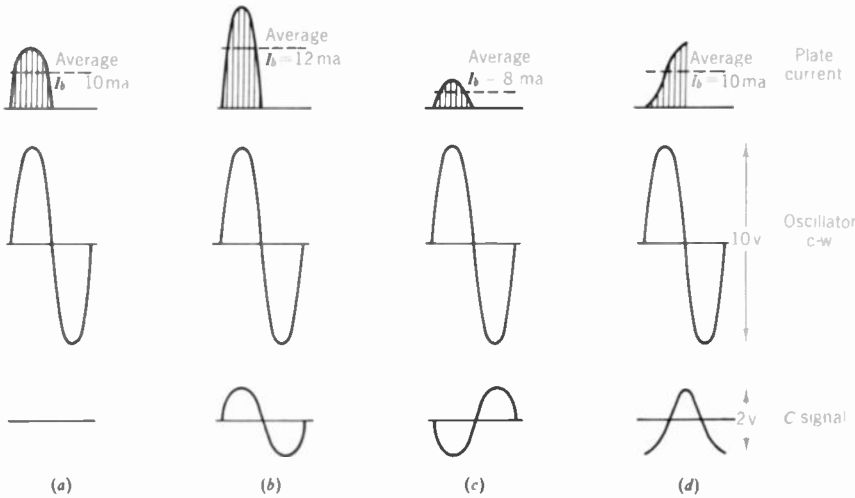


Fig. 25 · 13 How the phase between input voltages of synchronous detector determines average plate current I_b. (a) Oscillator cw input but no C signal. (b) C signal in phase with oscillator cw to increase I_b. (c) C signal of opposite phase decreases I_b. (d) Quadrature phase has same I_b as no C signal.

reason for this static condition with 90° phase between the two input voltages is that each quarter cycle has an opposite effect, increasing and decreasing plate current by the same amount around the 10-ma axis. The same result applies to a phase angle of 270°. This case of quadrature phase shows why a demodulator circuit has zero signal output for hues of quadrature phase, eliminating interference between the two modulation phases of the chrominance signal. Therefore, separate demodulator circuits are necessary to recover two different color video voltages that are not on the same axis. An axis refers to a straight line for any given phase angle and its opposite angle 180° away.

Demodulator phase angle. The two synchronous demodulators can detect the chrominance signal on any two axes to provide different color video voltages. Typical phase angles for the oscillator cw are shown in Fig. 25 · 14. The axes that have been used for most color receivers are:

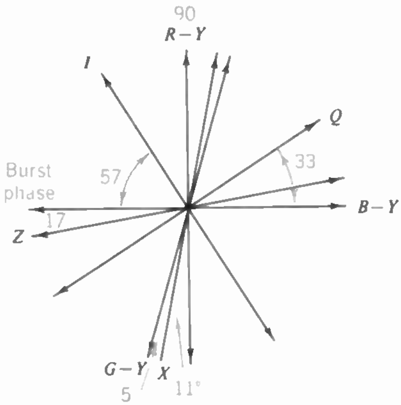


Fig. 25 · 14 Hues for different phases of oscillator cw into demodulators.

1. Any pair of the color-difference signals $R-Y$, $G-Y$, and $B-Y$. As examples, one demodulator can detect $R-Y$ phase and the other $B-Y$, or one for $R-Y$ and the other for $G-Y$. Only two axes are necessary to detect two color-difference signals, as the third can be obtained by matrixing the two detector output voltages in the correct proportions.
2. I and Q axes. These are the same as the transmitted chrominance signal.
3. X and Z axes. The Z phase is close to $-Q$, which is green-yellow, while the X phase is 11° from the hue of $-(R-Y)$, which is magenta.

For any axis, its two opposite polarities provide maximum detector output for that specific phase of oscillator cw. Furthermore, two axes in quadrature or close to 90° out of phase are often used in order to minimize crosstalk between the two detected signals. Remember that the synchronous demodulator produces maximum signal output for the axis of oscillator cw phase injected, but minimum output for the quadrature phase.

Gated-beam tube demodulators. The circuit in Fig. 25 · 15 uses gated-beam

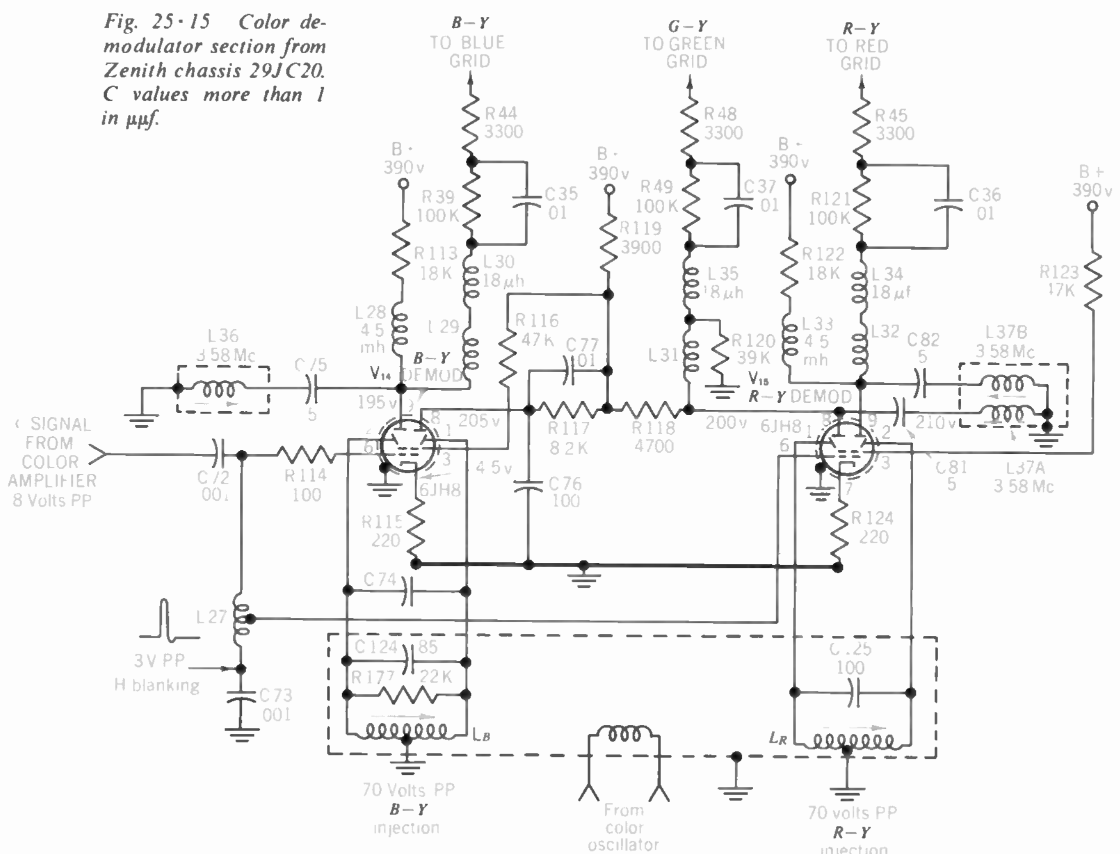


Fig. 25 · 15 Color demodulator section from Zenith chassis 29JC20. C values more than 1 in μμf.

tubes for the synchronous detectors. Each is equivalent to a pentode with two control grids. Note the dual plates (pins 8 and 9) and dual deflectors (pins 1 and 2) for each demodulator. The tubes are designed to conduct plate current only for positive deflector voltage. Furthermore, the dual structure enables one tube to produce two output voltages of opposite polarities. This feature simplifies the addition of negative $B - Y$ and $R - Y$ voltages to provide $G - Y$ signal. The tubes can handle the large grid voltage required for high-level modulation, without excessive amplitude distortion. Note that the $B - Y$, $R - Y$, and $G - Y$ signals are coupled directly to the kinescope grids for this high-level demodulator circuit.

Analyzing the demodulator at the left for $B - Y$, the two deflectors of the 6JH8 have push-pull oscillator cw input. These voltages are taken from the secondary winding L_B of the oscillator transformer T_{12}, which is center-tapped to provide equal and opposite voltages at the two ends. Therefore, the phase of oscillator cw is $(B - Y)$ for one deflector but $-(B - Y)$ phase for the other. The single control grid has chrominance signal input, which controls the current for both plates. C signal is taken from the second color amplifier in Fig. 25 · 3.

Because of the push-pull oscillator voltage on the deflectors, one plate in the $B - Y$ demodulator conducts maximum for $B - Y$ phase in the C signal, the other for $-(B - Y)$ phase. As a result, the rectified output includes $B - Y$ and $-(B - Y)$ video signals. The $B - Y$ signal from plate pin 9 goes to the kinescope for matrixing. The $-(B - Y)$ signal from plate pin 8 is combined with $-(R - Y)$ signal to provide $G - Y$ video signal.

The $R - Y$ demodulator V_{15} operates the same way but the oscillator phases for the deflectors are $(R - Y)$ and $-(R - Y)$, in quadrature with the $B - Y$ axis. The quadrature phase results from resonance at 3.58 Mc in the secondary winding L_R of T_{12}. Detected output from V_{15} includes $R - Y$ video signal from plate pin 9 to the kinescope for matrixing, and $-(R - Y)$ signal from plate pin 8. This signal and the signal from V_{14} are combined across R_{120}. The resulting $G - Y$ signal for the kinescope is formed with the proportions: $G - Y = -0.19(B - Y) - 0.51(R - Y)$. All three color video signals are d-c coupled to the kinescope to preserve the d-c axis for the correct brightness level without the need for d-c restorers.

Note the high-frequency compensation for the video signal output from the demodulators. For instance, L_{28} and L_{29} correspond to L_o and L_c as shunt and series peaking coils for the $B - Y$ video signal. The plate load resistor R_{113} is 18,000 ohms for the top correction frequency of approximately 0.5 Mc. The 3.58-Mc traps make sure there is no subcarrier voltage in the detected output signal.

25 · 8 Color picture tubes

As illustrated in Figs. 25 · 16 and 25 · 17, the kinescope used in practically all color receivers is a 21-in. round glass tube that has three electron guns to

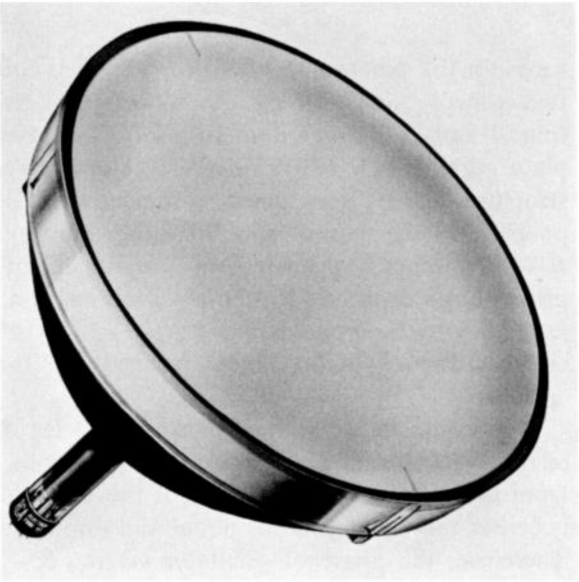


Fig. 25·16 Color kinescope type number 21FJP22. Internal construction illustrated in Fig. 25·17. (RCA.)

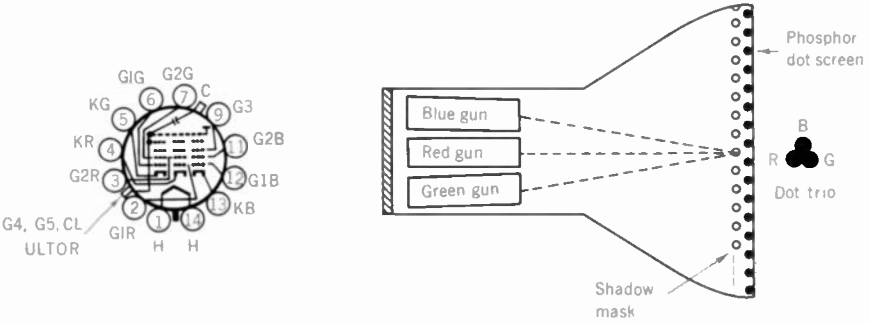


Fig. 25·17 Three-gun kinescope with shadow mask for color dot trios. Schematic of three guns with base connections shown at left for type number 21FJP22.

excite the phosphor screen coated with trios of red, green, and blue dots. The three electron guns are spaced 120° apart, as shown in Fig. 25·18. The gun with the lateral converging pole pieces produces beam current to excite the blue phosphor dots; one gun is for red and the third is for green. The blue gun is located near base pin 12, at the keyway. When the kinescope is installed, the blue gun must be at the top. Typical color kinescopes are listed in Table 25·1. All have an aluminized screen, requiring no ion-trap magnet. Video input signal is applied to the three guns at the same time for either a color picture or monochrome reproduction. Early color receivers used 15- and 19-in. kinescopes but these are not listed. The later types 25AP22 and 25BP22 have a larger rectangular screen and 90° deflection angle to fit in a more compact receiver cabinet.

Shadow mask. This is a thin metal sheet with tiny holes, each in line with a dot trio. See Fig. 25·17. For a 21-in. screen, there are about 400,000 trios of dots and the same number of apertures. The mask enables each beam to excite its respective color dots, when the beam converges at the

Table 25 · 1 Color picture tubes

Type number	*Envelope*	*Deflection angle*	*Maximum ultor voltage*	*Ultor terminal*	*Notes*
21AXP22	Metal	70°	25,000	Metal shell	21AXP22-A has new red phosphor*
21CYP22A	Glass	70°	25,000	Two cavity caps	Graded-hole shadow mask*
21FB22	Glass	70°	27,500	Cavity cap	
21FJP22	Glass	70°	26,500	Cavity cap	Integral safety window
21FKP22	Glass	70°	27,500	Cavity cap	Integral safety window
25BP22	Glass	90°	27,500	Cavity cap	Rectangular screen

* Also applies to later tubes listed.

correct angle, without exciting the other two colors. This action occurs over the entire screen, as the yoke deflects all three beams to produce the scanning raster. Any electrons that are not converging at the required angle are blocked by the metal shadow mask. In later tubes, the shadow mask has graded holes, which increase in diameter from the outer edge inward to the center, for increased brightness and improved convergence.

Electron guns. Each uses high-voltage electrostatic focusing with magnetic deflection. The focusing voltage required is about 5 kv, equal to one-fifth the ultor voltage of 25 kv. All three ultors are connected internally to a cavity cap connector for the high voltage. Similarly, the three focus grids are connected to base pin 9, for focusing voltage from a separate high-voltage supply. There are three heaters connected in parallel, for 6.3 volts at 1.8 amp at pins 1 and 14. The only electrodes with three separate connections are the screen grid, control grid, and cathode. The screen-grid voltage is usually made adjustable for each gun, between +130 and +370 volts, to produce a balanced white raster. The grid-cathode circuit of each gun needs separate input conections for red, green, and blue video signals.

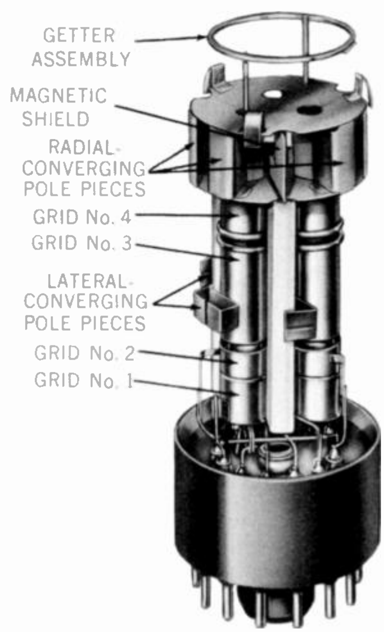


Fig. 25 · 18 Assembly of three electron guns in 21FJP22 color kinescope. (RCA.)

Color matrixing. In practically all receivers the kinescope is the matrix for red, green, and blue output from the color-difference video signals. Typically, $-Y$ video signal with positive sync polarity is applied to all three cathodes. Then, without color video for a monochrome program, the three guns produce a black-and-white picture. For a color picture, however, $R - Y$, $G - Y$, and $B - Y$ video signals are applied to the respective guns. For each color signal at the grid, though, its $-Y$ component is canceled by the $-Y$ signal at the cathode.

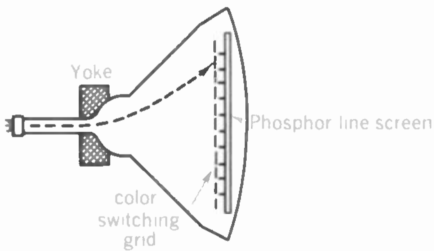
With positive sync polarity the $-Y$ video at the cathodes becomes more negative for white information. The $-Y$ component in the color-difference voltages also is more negative for white. However, one signal is at the control grid, while the other is at the cathode, for opposite effects on beam current. The net result for color, therefore, is as though R, G, and B video were applied to the three guns. D-c coupling is generally used for the color video signals, to eliminate d-c restorer circuits.

Color balance. A monochrome picture is reproduced as the sum of the contributions of red, green, and blue. To reproduce standard white the 21FJP22 color kinescope, as a specific example, requires the following percentages of beam current: $0.42R + 0.30G + 0.28B$. These proportions result from the unequal phosphor efficiencies of the three colors. To balance the colors for a white raster, the d-c voltages are adjusted for each gun. Included are screen-grid adjustments for each color and individual grid-cathode bias controls. These are also called *kinescope temperature adjustments*, as the idea is to produce white corresponding to 9300° Kelvin.

Black-and-white tracking. This term refers to maintaining the color balance within the normal range of the contrast and brightness controls. The main problem is obtaining a black-and-white picture, without color in the raster, for different video signal levels. Usually there are provisions for adjusting the amount of video signal drive to the three guns for good tracking, in addition to the screen-grid controls. A typical procedure is as follows:

1. Disable vertical scanning to have just a horizontal white line. There is usually a *normal-service* switch on the rear apron of the chassis for this purpose.
2. Set the three controls for red, green, and blue screen-grid voltage to the point where there is just enough light for a white line, barely visible. The bias controls need adjustment also. All these controls vary the kinescope d-c voltages, without a picture.
3. Return the switch to normal for a monochrome picture on a white raster.
4. Adjust the color video drive for each gun to produce good gray and white in the dark and light parts of the picture. Proper tracking should be obtained at all normal brightness levels.

When the kinescope is replaced, it may be necessary to readjust the



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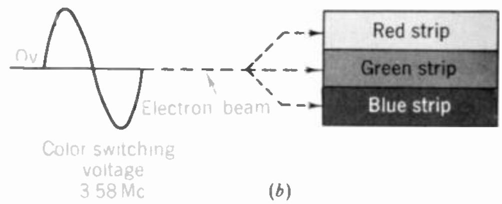


Fig. 25·19 Single-gun line-screen color kinescope. (a) Construction. (b) Effect of color-switching voltage on phosphor strips.

video drive for each color to compensate for different phosphor efficiencies. Also check the color-temperature adjustments.

One-gun line-screen color kinescope. The color picture tube in Fig. 25·19 has only one electron gun. It produces three colors by exciting strips of red, green, and blue phosphors on the kinescope screen. The electron beam must pass through the wire mesh serving as the color-switching grid. Zero voltage on this grid results in no displacement of the beam, which then strikes the center of the phosphor-strip trios. Positive voltage moves the beam to a different color strip and negative voltage enables the beam to excite the third color. The required amplitude of color-switching voltage is about 1,000 volts peak to peak. The switching frequency of 3.58 Mc is used in order to minimize dot pattern in the picture.

Note that only one color can be produced at any one time with a single gun. Therefore, this arrangement is sequential, compared with the simultaneous use of three colors in a three-gun tube. In order to use a single-gun tube, therefore the receiver requires switching voltage for the kinescope, timed correctly with respect to gated video signal for each color. The net result must allow video signal at the control grid just for red information, as an example, only at the time when the red phosphor is excited. Similarly, the green and blue video information must be gated on when the kinescope is reproducing these colors.

25·9 Kinescope set up adjustments

In Fig. 25·20, the magnets on the neck of the color kinescope are the deflection yoke, radial convergence magnet assembly, purity ring magnet, and blue lateral bar magnet. The yoke deflects all three beams. The convergence magnet has three parts, one for each of the beams. The purity magnet compensates for alignment of all three guns. The blue lateral magnet, which looks like an ion-trap magnet mounted on its spring clamp, has the function of moving only the blue beam left or right.

Details of the convergence magnet are shown in Fig. 25·21. The entire assembly is mounted directly over the internal pole pieces of each gun. Then the magnetic flux provided by each section is coupled through the glass tube neck to the associated internal pole pieces. These poles, with the internal shield, shape and confine the flux to affect only the one beam.

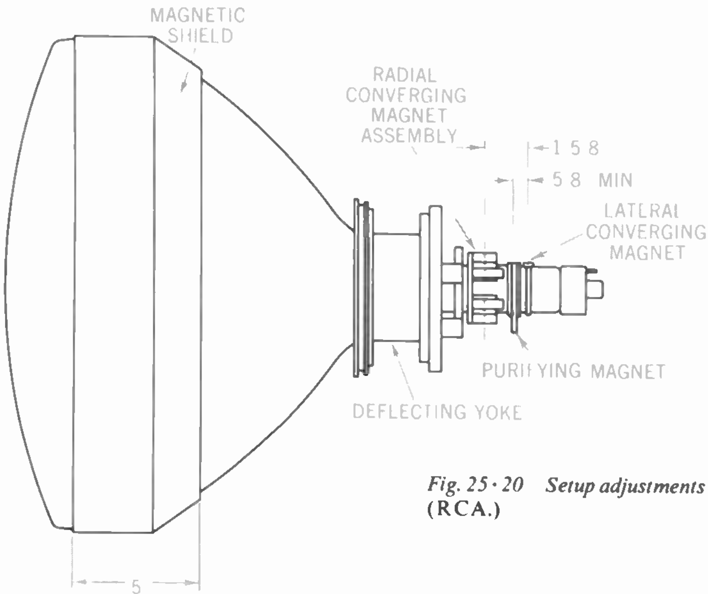
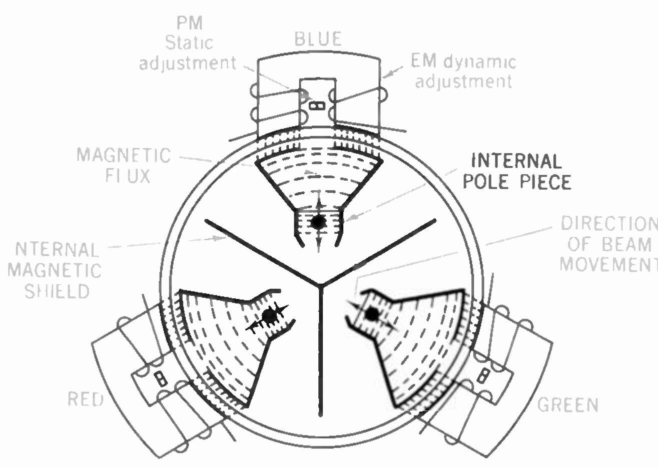


Fig. 25·20 Setup adjustments on neck of color kinescope. (RCA.)

Adjusting the strength of the flux from the external magnets deflects each beam slightly. Because of the position of the kinescope with the blue gun at top, with red and blue spaced 120° apart, the blue magnet moves its beam vertically, while the red and green beam moves diagonally. To provide horizontal movement of the blue beam, the lateral convergence magnet is used.

The net result of the four possible displacements allows convergence of the three beams to produce white. This idea is illustrated in Fig. 25·22. It should be noted that these are not phosphor dots. They are square dots produced by video signal input from a dot generator. Only the center of the screen is shown here, out of 15 to 25 horizontal and vertical rows of dots. Assume there is no convergence to start. Adjusting the red magnet

Fig. 25·21 Details of radial convergence magnet. (RCA.)



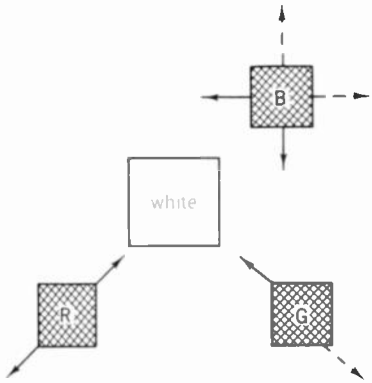


Fig. 25·22 Effect of convergence magnets on moving dots produced on screen by signal generator.

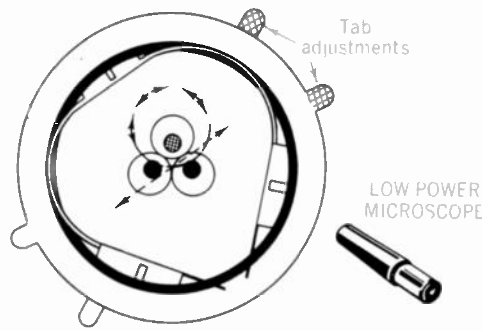


Fig. 25·23 Details of purity magnet. (RCA.)

moves the dot diagonally down to the left or up to the right, as you face the screen. The green dot can be moved in the opposite diagonal. If you move the red dot up a little to the right and the green dot up a little to the left, they will overlap to form a yellow dot. Then move the blue dot with the lateral magnet a little to the left, in line above the yellow dot. Finally, the blue convergence magnet can move the blue dot straight down to converge all three beams for a white dot.

Convergence adjustments. Making the three beams pass through the same opening in the aperture mask at the same time is called *convergence.* Doing this for the center area of the screen, around one-half the area, is *static convergence.* Each section of the radial magnet has a small permanent magnet that is adjusted for static convergence at the center of the screen. The ability to converge the three beams at the edges of the picture, as the beam is deflected to fill the height and width, is *dynamic convergence.* The electromagnets in the radial assembly have correction current from the deflection circuits to provide dynamic convergence at the top, bottom, left, and right edges.

It is important to realize that the test of convergence is a black-and-white picture. The evidence of good convergence with a dot-bar generator is white dots and bars. In a monochrome picture. there should be no color fringing at the edges of objects (see Plate II).

Purity adjustments. The center axis of the three beams should strike the centers of their respective color phosphors in a dot trio. Then the light from each dot is a pure color, without contamination from the other two colors. This result is obtained by the color purity adjustments, which are also called *register* or *beam-landing* adjustments. The correction is necessary to compensate for slight misalignment of the three guns. Also, the earth's magnetic field can displace the electron beams enough to cause poor purity. It is important to note that the test of good purity is a white raster, without any areas of color. The purity adjustments are made without any signal.

In Fig. 25·23 the purity magnet is used to adjust all three beams. The two red tabs on the ring magnet can be spread to increase field strength, which moves the three beams diagonally. Rotating the entire magnet moves the three beams in a circular path. The microscope shown can be used for viewing the illuminated phosphor dots to check purity. Note that the static convergence must be approximately correct to obtain good purity.

The technique for adjusting purity can be as follows:

1. Turn down the green and blue screen controls to view just a red raster.
2. Adjust the purity magnet for saturated red. An example of incorrect adjustment of the purity magnet is shown in Plate III.
3. Check the green raster alone and the blue raster alone for good purity.

After adjusting for good purity, set the screen controls for a white raster. It should be a uniform gray, without color in any area.

The purity magnet affects most of the screen area, but more adjustments may be needed for the edges. Moving the deflection yoke slightly forward or back affects purity at the edges of the raster. The magnetic shield around the front of the kinescope in Fig. 25 · 20*b* minimizes the effects of external magnetic fields that can affect purity. In addition, some receivers have six *field-equalizing* magnets mounted around the screen to correct purity at the edges.

Degaussing. This term refers to demagnetizing metal parts of the color kinescope, which can become slightly magnetized by the earth's steady magnetic field. The main effect is on purity. Different amounts of magnetization are the reason why the purity can change when the receiver is moved to a different location. When perfect purity is a problem, degaussing may be necessary.

A degaussing coil generally consists of 425 turns of No. 20 wire on a 12-in.-diameter form. See Fig. 25 · 24. This coil is connected to the 115-volt a-c power line to supply current for a varying magnetic field. Move the coil around to cover the entire screen area; then move a few feet back to reduce the effect of its magnetic field and disconnect the coil. The entire operation takes about 30 sec. The receiver can be either on or off.

Purity and convergence affect each other. Problems in obtaining good convergence may be due to purity that is not exactly right. The purity adjustments must be made with good static convergence. Although the adjustments interact, remember that purity is for a white raster while convergence eliminates color fringes in the picture.

25 · 10 Convergence procedure

The test of correct convergence is the ability of the color kinescope to reproduce black-and-white areas without color fringing. Therefore, the convergence adjustments are made with a black-and-white picture. The best method is with a dot-bar generator that produces dots, vertical bars, horizontal bars, or a crosshatch pattern, all in black and white.

The steps listed in numerical order here give a specific example of how

Fig. 25·24 Degaussing coil with switch for power-line cord. Diameter about 12 in. (Walsco Electronics Co.)

Fig. 25·25 High-voltage cheater being inserted at back of color television receiver. (Walsco Electronics Co.)

this important job can be done in a logical sequence. Once the convergence is done, the adjustments are stable. First, check all the conventional monochrome adjustments, including height, width, linearity, centering, AFC, AGC, and focus at the required anode voltage. Color receivers usually have a high-voltage adjustment. A separate cheater plug may be needed to operate the high-voltage supply with the back cover off. The plastic "bottle" in Fig. 25·25 opens a metal spring that shorts the high-voltage filter capacitor when the back cover is removed.

1. Adjust purity. No signal is required. Use degaussing coil at start. Adjust purity magnet for pure red raster at center. Check that static convergence at center is good (see step 4). Move deflection yoke back or forward slightly to improve purity at edges of raster. Adjust field equalizing magnets, if provided, around edge of kinescope faceplate. Try for perfect purity, as this makes the convergence easier.
2. Adjust screen-grid voltages and bias voltages of the three guns for white raster.
3. Convergence is checked with a dot-bar signal generator. Obtain a steady black-and-white pattern on the screen. Horizontal bars, vertical bars, crosshatch, bars or dots can be used for different parts of the job. The general requirements are illustrated in Fig. 25·26 with a crosshatch pattern.
4. Static convergence at the center is usually easy. Adjust the red, green, and blue permanent magnets on the convergence assembly, plus the blue lateral magnet, to make the three colors overlap and produce white. Refer back to Fig. 25·22. The usual range of movement of the color dots is $\pm\frac{1}{2}$ in. It may be necessary to remove one or more of the magnets, invert it 180°, and reinsert. For good static convergence, the bars or dots at the center are white without any color fringing.

The dynamic convergence takes the most time because it requires 12 or

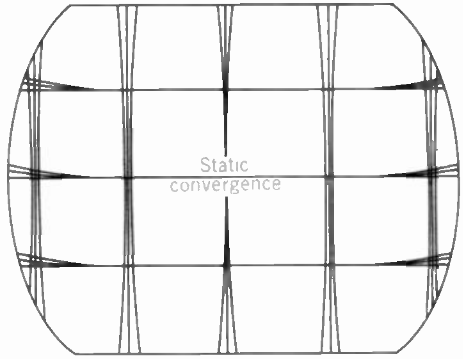


Fig. 25 · 26 Crosshatch pattern with static convergence at center but fringing at edges.

more adjustments. These are usually on a separate subchassis. There are separate adjustments for red, green, and blue, at the top, bottom, left, and right of the picture. When service notes are available, follow the manufacturer's order of instructions for each control. If not, observe the effect of each control at each area around the edges. Then use each control for the corresponding correction.

If there is confusion about parallel lines or dots of colors, move the blue color out with the static magnets and use blue as the reference. Since blue can be moved both vertically and horizontally with separate magnets, this is usually the best color to converge into the other two. Do not be afraid to separate the colors with the static adjustments if this helps in making the dynamic convergence. Normally, it is easy to merge all the colors again with the static magnets.

It is helpful to watch for types of fringing at the edges that can be corrected with the static magnets. For instance, suppose that all the bars or dots across the screen have blue fringing, always to the right. This can easily be corrected by a slight readjustment of the blue lateral magnet. In some receivers, the blue dynamic adjustment for vertical convergence moves the blue dot in opposite directions at the top and bottom. Once this is adjusted for equal fringing in the same direction from top to bottom, the blue vertical static magnet can then produce perfect convergence.

25 · 11 Color controls and adjustments

See Table 25 · 2. In general, the main requirement is setting up the color kinescope to produce a monochrome picture on a white raster. Small areas of color in the raster indicate poor purity. Overall color in the entire raster and picture is a problem of d-c voltages for the three guns. Color fringing at the edges of picture information indicates poor convergence. When the color kinescope reproduces a good black-and-white picture, the color picture will also be good, assuming normal receiver operation.

The usual controls and setup adjustments for a monochrome receiver are omitted here but these must be set correctly before any color adjustments are made. This is especially true of the fine tuning control, which

Table 25·2 Color controls and adjustments

Control	*Circuit*	*Function*	*Adjustment*
Color level, chroma, or saturation	Chrominance amplifier	Vary amount of 3.58-Mc chrominance signal	Strong or weak color
Hue, tint, or phasing	3.58-Mc reference oscillator or color AFC circuit	Vary phase between color oscillator and sync burst	Correct hues in flesh tones or known object
Deflection yoke	Deflection circuits	Produce scanning raster	Back and forth for best purity at edges of raster
Purity magnet	Permanent magnet	Align three electron beams	Pure red in center area of raster
Rim magnets	Permanent magnet	Corrects beam alignment at edges	Pure red at edges of raster
Static convergence magnets	Permanent magnet	Align each beam with respect to the other two	Converge three beams to produce white in center area of picture
Horizontal dynamic convergence	Electromagnets with current from horizontal output circuit	Correct for misconvergence caused by horizontal deflection	Good convergence at left and right sides of picture
Vertical dynamic convergence	Electromagnets with current from vertical output circuit	Correct for misconvergence caused by vertical deflection	Good convergence at top and bottom of picture
Focus	Focus-voltage rectifier in high-voltage supply	Vary focusing-grid voltage of kinescope	Sharp scanning lines in raster; affects convergence
High-voltage	Shunt regulator in kinescope anode supply	Provide correct amount of anode voltage	Affects focus and convergence
Red screen, green screen, and blue screen	Low-voltage power supply	Vary screen-grid voltage for each gun	White
Background or bias	Kinescope cathode-grid circuit	Vary d-c grid bias for each gun	White raster
Blue gain and green gain	Input signal to kinescope	Balance color video signals for three guns	Black-and-white picture
Color killer	Chrominance amplifier	Cut off color for monochrome reception	No color streaking in black-and-white picture

must be set exactly to convert the chrominance signal side bands to the i-f amplifier frequencies. Otherwise, there will be no color. For a color program, the fine tuning control must be set first in order to have color signal. With the color level control turned up, adjust the fine tuning for maximum color in the picture and minimum 920-kc beat, which is a coarse herringbone pattern. When switching between stations broadcasting in color, readjustment of the fine tuning control may be necessary.

With color tuned in, set the hue control for the correct tints and the color control for the desired saturation. It is usually best to set the hue control by correct flesh tones of people in the scene. The range of phase-angle variations for the hue control is about 90°. Do not use too much color, as a small amount of saturation makes the colors more natural. The hue and color controls are generally on the front panel of the receiver as operating controls. The other color controls and adjustments are on the chassis, often behind a hinged cover, and on the kinescope neck.

Dot-bar generator. Although approximate adjustments of static convergence can be made on any steady monochrome picture, like a test pattern, best results are obtained with a dot-bar generator. This is a signal generator to produce a stable pattern of dots, lines, or crosshatch. All the patterns are in black and white, as convergence is a problem of obtaining white from the three-color beam. Some generators provide modulated r-f output, which is connected to the receiver antenna input terminals. Others provide video signal output, to be injected directly in the kinescope grid-cathode circuit. This type usually provides sharper edges in the dots and lines. In some generators, the output can be switched for either r-f or video. For either type, an extra cable may be provided for horizontal sync from the receiver. This lead is generally clipped on the insulation of the wire to the horizontal deflection coils in the yoke. The receiver must be tuned to a station, with the picture in sync, to lock in the generator. Usually, the generator can produce 6 to 20 rows of dots or bars in the vertical and horizontal directions. The more rows you use the more exact is the convergence adjustment.

Color-bar generator. This is used to check the color circuits. As an example, alignment can be checked by noting the strength of the appropriate colors in the picture. A typical generator produces a rainbow of 10 vertical color bars advancing in hue by 30° for each bar. Then, every third bar differs in hue by 90°. Every sixth bar is a complementary color, 180° away.

Stripe test signal. Some stations broadcast this 3.58-Mc color test signal during daytime monochrome programs, to be used for checking color receiver installations when there is no color program (see Plate VI). In black and white, the stripe just shows 920-kc beat. In color, its hue is the greenish-yellow of color sync phase, locked in, or a barber pole of varying colors. The stripe signal occurs just before and after horizontal blanking. To see color, the horizontal sync must be delayed to delay the flyback pulses. Then the stripe signal can operate the burst separator to provide color sync and cut off the color killer. In some receivers, a stripe-test terminal is available which shunts a 0.001-μf capacitor across the coupling capacitor for the sync separator output, to provide the required delay.

25 · 12 Schematic of color section of receiver

Figure 25 · 27 shows how the color circuits fit together in a typical receiver. Note the sections marked for different functions. Starting with the colorplexed composite video signal at the top left in the diagram, this output voltage from the video detector is coupled to the grid-cathode circuit of V_{110}, which is the first video amplifier. Its plate current produces signal across R_{127} in the cathode circuit. The circuit supplies signal to the luminance amplifier V_{111}, through the delay line. Then the amplified signal $(-Y)$ is coupled to the three cathodes of the kinescope. H blanking pulses are applied to the screen grid of V_{111} from the blanking amplifier V_{122A}.

In the plate circuit of V_{110}, the transformer T_{108} supplies 3.58-Mc modulated chrominance signal to the color amplifier V_{121B}. Its output is amplified chrominance signal for two paths: the demodulator driver V_{123} and the burst separator V_{119B}.

Following the path for color signal first, V_{123} is a second chrominance amplifier stage to provide enough signal to drive the high-level demodulators. The transformer T_{114} supplies C signal to the plate of each demodulator V_{125A} and V_{125B}. In addition, each grid has oscillator cw input from the two secondary windings on the oscillator transformer T_{115}. The triode demodulators operate as grid-controlled rectifiers, producing maximum detected signal output for the axis of oscillator cw phase in the input. The $R - Y$ and $G - Y$ video signals from the two synchronous demodulators are combined in the $B - Y$ amplifier V_{124B}. Then each of the three color-difference signals is applied to the control grid of the appropriate electron gun. With $-Y$ signal at the three cathodes, the kinescope serves as the matrix to provide red, green, and blue. Direct coupling is used so that d-c restorers are not necessary.

Referring back to the chrominance amplifier, its output transformer also couples C signal to the cathode of the burst separator V_{119B}. In this stage, the phase detector transformer T_{112} couples the separated color sync burst to the color AFC diodes in V_{120}. The rectified output voltage of the top diode, across R_{218} and R_{217}, is used to bias the color killer. Only the voltage across R_{217} provides ACC bias. This circuit is the same as Fig. 25 · 10.

The d-c control voltage from the phase detector, at the center arm of R_{228}, controls the reactance tube V_{124A}. Note that the oscillator cw for the $R - Y$ demodulator is also coupled to the phase detector diodes as a sample of oscillator output.

In the high-voltage supply, at the lower left, the 3B2 rectifier can supply the required beam current for three electron guns, at 25 kv. Note the separate rectifier V_{114} for focusing voltage at about 5 kv and the regulator tube V_{113} to maintain the high voltage and good convergence with changes in load current. The high-voltage filter capacitor of 2500 $\mu\mu$f is shorted by the safety interlock S_{104} when the back cover is removed. This switch is opened by the cheater in Fig. 25 · 25.

25·13 Color troubles

Typical problems are listed in Table 25·3, separated according to raster, monochrome picture, and color picture. The first requirement of the color receiver is that it produce a monochrome picture on a white raster. Note that the d-c voltages for the three electron guns determine the color of the raster. These voltages include grid-cathode bias and screen-grid voltage for each gun. With d-c coupling of the signal to the kinescope, however, troubles in the signal circuits can affect the raster. As an example, if the plate voltage rises to $B+$ in an $R - Y$ demodulator directly coupled to the control grid of the red gun, the raster will be red because of excessive red beam current.

Assuming the receiver produces a good monochrome picture, the main color troubles are no color, weak color, wrong hues, or no color synchronization. For the case of no color, remember that the chrominance signal must be picked up by the antenna for the r-f circuits and converted by the fine tuning control to the correct i-f signal frequencies. Similarly, weak color can be caused by insufficient chrominance signal in the r-f and i-f circuits. You can check the 3.58-Mc chrominance signal in the output of the video detector with an oscilloscope. The peak-to-peak amplitude of color burst should be the same as for the horizontal sync pulse.

When the chrominance signal output of the video detector is normal but the picture has no color or weak color, the trouble is in the chrominance section. The chrominance amplifier may be weak or not operating. Check the color killer circuit and its threshold adjustment. Remember that the burst amplifier must operate to cut off the killer to allow operation of the chrominance amplifier.

No output from the color reference oscillator means no demodulation and therefore no color. Note that, if only one demodulator operates, the result is wrong colors. However, no oscillator output means both demodulators cannot produce color video signal output.

Table 25·3 Color troubles

Indicator	*Troubles*	*Cause*
Raster	Entire raster in color	Kinescope d-c voltages
	Color areas	Poor purity
Monochrome picture	Color fringing	Poor convergence
	Color sparkling at edges	Chrominance amplifier not cut off
Color picture	No color	No color signal
	Weak color	Weak color signal
	Wrong hues	Color oscillator phase; color video balance
	Varying color stripes	No color AFC

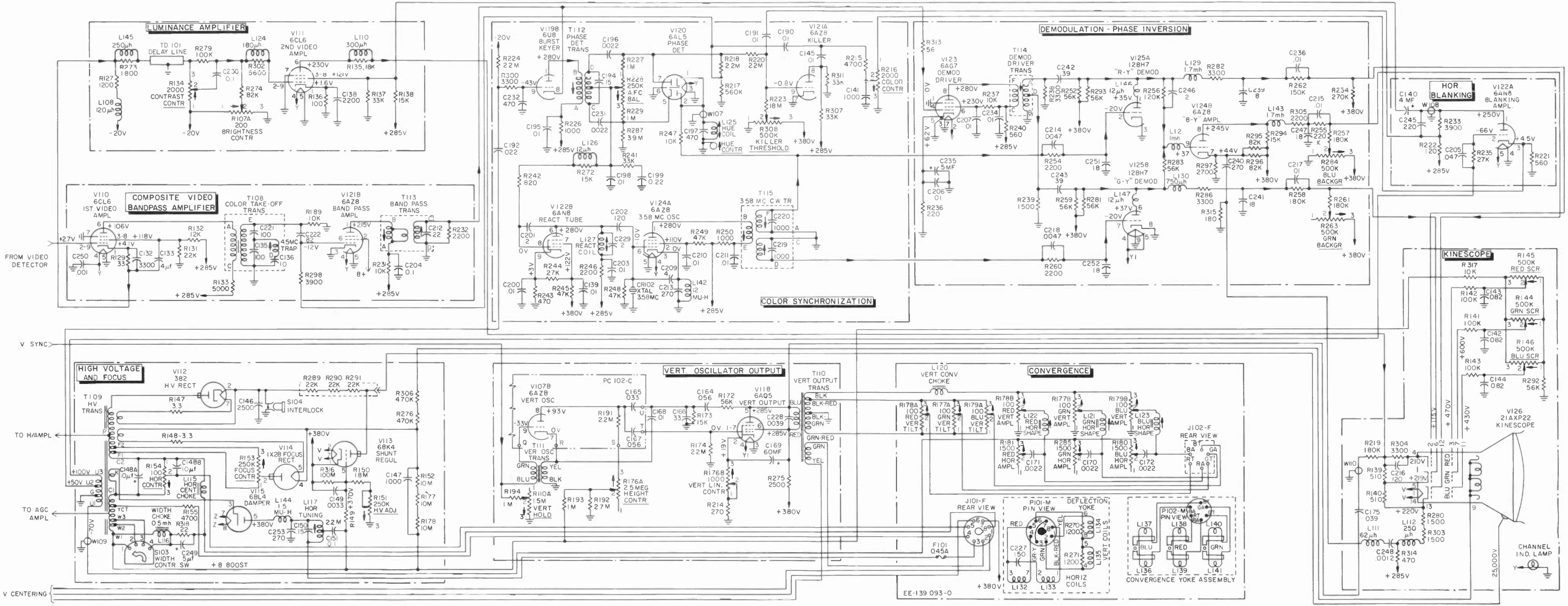


Fig. 25·27 Schematic of color section of receiver. C values more than 1 in μμf. M is megohms. (RCA chassis 21CT662.)

Incorrect phase of the oscillator output for the demodulators produces wrong hues. However, this is mainly an alignment problem, which seldom changes by itself. See if the hue control is effective in shifting colors through approximately 90° of hues. It is helpful to note which colors are either missing or too strong, as this can indicate which color circuit has the trouble. For instance, blue and yellow horizontal hum bars drifting up or down in the picture indicate hum in the blue video circuits. Weak output for $R - Y$ video will cause insufficient red and cyan in the color picture. In summary, the color video signals must be balanced to have a picture with correct colors.

Without color synchronization, the picture has horizontal or diagonal color stripes (Plate IV). The continuously changing stripe pattern shows the color AFC circuit is not locking in the color reference oscillator. Check the reactance tube generally used in the oscillator control stage, the phase detector producing d-c control voltage, and the burst separator. If the color stripe pattern remains stationary this shows the oscillator is locked in at the wrong frequency. This trouble can be caused by the oscillator crystal.

SUMMARY

1. The color section of the receiver includes the 3.58-Mc chrominance band-pass amplifier, two demodulators for color video signal to the kinescope, the color reference oscillator and the burst separator supplying sync to the color AFC circuit. In addition, the color killer stage cuts off the color amplifier during monochrome reception. The remaining circuits are like a monochrome receiver, but the band pass and tuning must be exactly right to tune in the color signal.
2. The chrominance amplifier is fixed-tuned to 3.58 Mc, like an i-f amplifier just for color, to supply enough color signal to drive the demodulators.
3. The color level control adjusts saturation in the picture by varying the amount of signal output from the chrominance amplifier.
4. Automatic color control (ACC) bias is generally used to control the gain of the chrominance amplifier according to the strength of color signal.
5. When the color killer stage conducts, it cuts off the color amplifier for a monochrome signal. The purpose is to eliminate color noise, or confetti, from the black-and-white picture. With color signal, however, the killer stage is cut off, allowing the color amplifier to operate.
6. The burst separator is tuned to 3.58 Mc but cut off during trace time and gated on by horizontal flyback pulses to recover the color sync burst. Its input includes chrominance signal but the output is only the 3.58-Mc burst on the back porch of horizontal sync.
7. The color AFC circuit holds the color reference oscillator at the correct frequency or phase. No color hold results in stripes of varying hue in the color picture.
8. The 3.579545-Mc reference oscillator generates the color subcarrier for the demodulators.
9. The hue or tint control varies the phase of either the sync burst or oscillator input to the color AFC circuit.
10. There are two color demodulators to detect two color signals, which are combined to produce the third. The synchronous detector circuit used requires modulated chrominance signal input and cw input from the color reference oscillator. The phase of the oscillator input to each demodulator determines the hue axis of detected color video signal output.
11. In the three-gun tricolor kinescope, all three beams must pass through each aperture in the shadow mask to excite the trios of red, green, and blue phosphor dots. Proper con-

vergence makes the three beams pass through one aperture for each dot trio. Poor convergence causes color fringes, especially evident in a monochrome picture. Purity refers to having each beam excite its correct phosphor dots, for saturated colors. Poor purity causes color areas in a white raster.

SELF-EXAMINATION *(Answers at back of book.)*

Choose (*a*), (*b*), (*c*), or (*d*).

1. Which of the following does *not* affect the chrominance signal? (*a*) Antenna; (*b*) fine tuning control; (*c*) synchronous demodulator; (*d*) 4.5-Mc sound i-f amplifier.
2. The output of the chrominance band-pass amplifier: (*a*) drives the color reference oscillator; (*b*) is varied by the hue control; (*c*) drives the synchronous demodulators; (*d*) operates the color killer.
3. Which of the following applies for a monochrome signal? (*a*) Color amplifier conducting; (*b*) color killer conducting; (*c*) maximum output from color phase detector; (*d*) maximum $G - Y$ video signal.
4. Which of the following is *not* tuned to 3.58 Mc? (*a*) Chrominance band-pass amplifier; (*b*) $G - Y$ amplifier; (*c*) sync burst separator; (*d*) demodulator input circuit.
5. To reduce color fringes at the bottom of a monochrome picture, adjust the: (*a*) purity magnet; (*b*) color video drive signals; (*c*) dynamic convergence; (*d*) color level control.
6. To eliminate color in the center area of a white raster, adjust the: (*a*) purity magnet; (*b*) hue control; (*c*) color level control; (*d*) red screen-grid control.
7. If you turn down the blue screen-grid control but allow the red and green guns to operate, without a picture you will see: (*a*) yellow raster; (*b*) blue raster; (*c*) color fringes; (*d*) diagonal color stripes.
8. Which of the following is cut off during horizontal trace time? (*a*) Color amplifier; (*b*) sync burst separator; (*c*) color reference oscillator; (*d*) Y video amplifier.
9. With oscillator cw phase at 254° for $G - Y$ phase, the hue of detected output is: (*a*) red or blue; (*b*) orange or cyan; (*c*) yellow or blue; (*d*) green or magenta.
10. The hue of sync burst phase is: (*a*) orange; (*b*) cyan; (*c*) blue; (*d*) yellow-green.

ESSAY QUESTIONS

1. Give the specific functions in terms of the color picture reproduction for the following: r-f local oscillator in front end, i-f amplifier, video detector, chrominance band-pass amplifier, burst amplifier, color reference oscillator, color killer, $R - Y$ demodulator, $B - Y$ demodulator, $G - Y$ amplifier, color phase detector, and horizontal output transformer.
2. Describe briefly operation of the color level control. Why is this similar to a contrast control?
3. Describe briefly the operation of the hue control. How is it set for correct hues?
4. Describe how to set the fine tuning control for color in the picture.
5. How is it possible for a narrow-band antenna to prevent color in the picture but still allow a black-and-white picture?
6. Give one advantage and one disadvantage of a receiver demodulating $R - Y$ and $B - Y$ color video, compared with I and Q signals.
7. A color program is being received on channel 7 (174 to 180 Mc). List the numerical values for the color subcarrier frequency in the r-f, i-f, and chrominance circuits.
8. Define the following: ACC bias, color temperature balance, kinescope matrix, synchronous demodulator, confetti, green-stripe test signal.
9. What controls are adjusted for a white raster?
10. Give the function of all components in the color amplifier circuit of Fig. 25·2.
11. Draw schematic diagrams for the following circuits: chrominance band-pass amplifier, Y video amplifier, $G - Y$ video amplifier. Give one difference between the chrominance amplifier and the other two circuits. Also between the Y and $G - Y$ amplifiers.
12. Indicate which of the following stages are conducting during trace time for a color pro-

gram and for a monochrome program: chrominance amplifier, burst separator, color AFC diodes, and color killer.

13. Why are two color demodulators used, instead of one or three?
14. List two pairs of quadrature axis for the color demodulators, in addition to the I and Q phases. What is the advantage of 90° phase between the two demodulators?
15. List the input and output signals for the following stages: video detector, chrominance amplifier, burst separator, color phase detector, oscillator control tube, $B - Y$ demodulator, three-gun kinescope used as matrix.
16. How does the picture look without color hold? What stages affect this?
17. How does the picture look if the color oscillator does not operate?
18. What two controls can remove color from the picture to see it only in monochrome? How is it possible to have a monochrome on a color kinescope?
19. Define the following: static convergence, dynamic convergence, degaussing, purity, blue lateral magnet, shadow mask, color-switching grid.
20. Give three adjustments that affect color balance.
21. What is the difference in adjustments for a raster that is too blue all over, compared with just a blue area in one corner?
22. Name three adjustments that affect purity.
23. How can you tell the difference between poor purity and poor convergence?
24. Explain briefly how the video signal input for a single-gun color kinescope differs from the requirements of a three-gun tube.
25. What will be the effect in the picture if each of the following stages does not operate: chrominance amplifier, oscillator control tube, color killer, color reference oscillator, $R - Y$ demodulator. Assume just the one trouble at a time.
26. List the input and output signals of a $B - Y$ demodulator, as an example of a synchronous detector.
27. Why does a synchronous demodulator produce maximum signal output voltage for the axis of its cw input phase?
28. How is a color-bar generator used?
29. How is a dot-bar generator used? Make a drawing showing the directions red, green, and blue dots can be moved to produce white.
30. List briefly the main steps in setting up the tricolor kinescope for purity and convergence.
31. How do you adjust the color killer threshold level control?
32. In Fig. 25 · 3, give the function for R_{104}, L_{25}, C_{69}, and R_{111}. How will the picture look if R_{104} opens? R_{105} opens?
33. In Fig. 25 · 10, to what frequency is T_{108} tuned? T_{113}?
34. In Fig. 25 · 12, give the function of each component.
35. In Fig. 25 · 15, find three wave traps for the color subcarrier frequency. Why are they used? Calculate the inductance of L_R to resonate with C_{125} at 3.58 Mc.
36. In Fig. 25 · 27, list each stage with all controls and give their functions.
37. In Fig. 25 · 27, calculate the stored energy in joules in the high-voltage filter capacitor C_{146}.

Appendix A

Television channel frequencies

Channel number	Frequency band, Mc	Picture carrier frequency, Mc	Sound carrier frequency, Mc	Channel number	Frequency band, Mc	Picture carrier frequency, Mc	Sound carrier frequency, Mc
2	54–60	55.25	59.75	43	644–650	645.25	649.75
3	60–66	61.25	65.75	44	650–656	651.25	655.75
4	66–72	67.25	71.75	45	656–662	657.25	661.75
5	76–82	77.25	81.75	46	662–668	663.25	667.75
6	82–88	83.25	87.75	47	668–674	669.25	673.75
7	174–180	175.25	179.75	48	674–680	675.25	679.75
8	180–186	181.25	185.75	49	680–686	681.25	685.75
9	186–192	187.25	191.75	50	686–692	687.25	691.75
10	192–198	193.25	197.75	51	692–698	693.25	697.75
11	198–204	199.25	203.75	52	698–704	699.25	703.75
12	204–210	205.25	209.75	53	704–710	705.25	709.75
13	210–216	211.25	215.75	54	710–716	711.25	715.75
14	470–476	471.25	475.75	55	716–722	717.25	721.75
15	476–482	477.25	481.75	56	722–728	723.25	727.75
16	482–488	483.25	487.75	57	728–734	729.25	733.75
17	488–494	489.25	493.75	58	734–740	735.25	739.75
18	494–500	495.25	499.75	59	740–746	741.25	745.75
19	500–506	501.25	505.75	60	746–752	747.25	751.75
20	506–512	507.25	511.75	61	752–758	753.25	757.75
21	512–518	513.25	517.75	62	758–764	759.25	763.75
22	518–524	519.25	523.75	63	764–770	765.25	769.75
23	524–530	525.25	529.75	64	770–776	771.25	775.75
24	530–536	531.25	535.75	65	776–782	777.25	781.75
25	536–542	537.25	541.75	66	782–788	783.25	787.75
26	542–548	543.25	547.75	67	788–794	789.25	793.75
27	548–554	549.25	553.75	68	794–800	795.25	799.75
28	554–560	555.25	559.75	69	800–806	801.25	805.75
29	560–566	561.25	565.75	70	806–812	807.25	811.75
30	566–572	567.25	571.75	71	812–818	813.25	817.75
31	572–578	573.25	577.75	72	818–824	819.25	823.75
32	578–584	579.25	583.75	73	824–830	825.25	829.75
33	584–590	585.25	589.75	74	830–836	831.25	835.75
34	590–596	591.25	595.75	75	836–842	837.25	841.75
35	596–602	597.25	601.75	76	842–848	843.25	847.75
36	602–608	603.25	607.75	77	848–854	849.25	853.75
37	608–614	609.25	613.75	78	854–860	855.25	859.75
38	614–620	615.25	619.75	79	860–866	861.25	865.75
39	620–626	621.25	625.75	80	866–872	867.25	871.75
40	626–632	627.25	631.75	81	872–878	873.25	877.75
41	632–638	633.25	637.75	82	878–884	879.25	883.75
42	638–644	639.25	643.75	83	884–890	885.25	889.75

Appendix B

FCC frequency allocations from 30 kc to 300,000 Mc

Band	Allocation	Remarks
30–535 kc	Includes maritime communications and navigation, international fixed public band, aeronautical radio navigation	Very low, low, and medium radio frequencies
535–1,605 kc	Standard radio broadcast band	AM broadcasting
1,605 kc–30 Mc	Includes amateur radio, loran, government radio, international short-wave broadcast, fixed and mobile communications, radio navigation, industrial, scientific, and medical equipment	Amateur bands 3.5–4.0 Mc and 28–29.7 Mc; industrial, scientific, and medical band 26.95–27.54 Mc
30–50 Mc	Government and nongovernment, fixed and mobile	Includes police, fire, forestry, highway, and railroad services; VHF band starts at 30 Mc
50–54 Mc	Amateur	6-meter band
54–72 Mc	Television broadcast channels 2–4	Also fixed and mobile services
72–76 Mc	Government and nongovernment services	Aeronautical marker beacon on 75 Mc
76–88 Mc	Television broadcast channels 5 and 6	Also fixed and mobile services
88–108 Mc	FM broadcast	Also available for facsimile broadcast; 88–92 Mc educational FM broadcast
108–122 Mc	Aeronautical navigation	Localizers, radio range, and airport control
122–174 Mc	Government and nongovernment, fixed and mobile, amateur broadcast	144–148 Mc amateur band
174–216 Mc	Television broadcast channels 7–13	Also fixed and mobile services

Band	*Allocation*	*Remarks*
216–470 Mc	Amateur broadcast, government and nongovernment, fixed and mobile, aeronautical navigation, citizens' radio	Radio altimeter, glide path and meteorological equipment; citizens' radio band 460–470 Mc; civil aviation 225–400 Mc; UHF band starts at 300 Mc
470–890 Mc	Television broadcasting	UHF television broadcast channels 14–83
890–3,000 Mc	Aeronautical radionavigation, amateur broadcast, studio-transmitter relay, government and nongovernment, fixed and mobile	Radar bands 1,300–1,600 Mc
3,000–30,000 Mc	Government and nongovernment, fixed and mobile, amateur broadcast, radio navigation	Superhigh frequencies (SHF) Used for radio relay; theater television
30,000–300,000 Mc	Experimental, government, amateur	Extremely high frequencies (EHF)

Appendix C

PRINCIPAL TELEVISION SYSTEMS OF THE WORLD

Many countries throughout the world use television standards that are not the same as in the United States. These are listed, with our own FCC standards, in Table C · 1. The field rate of 50 cps is used where this is the a-c power-line frequency, instead of the 60 cps used in this country. Note that the combination of 625 lines per frame and 25 frames per second, as used in Western Europe, results in the line-scanning frequency of 15,625 cps, which is very close to the 15,750-cps frequency in our standards. In all cases, odd-line interlaced scanning is used with two fields per frame, the aspect ratio is 4:3, and amplitude modulation is used for the picture carrier.

*Table C · 1 Principal television systems of the world**

Where used	*Western Hemisphere and Far East; includes U.S., Canada, Japan, and most South American countries*	*Western Europe; includes Germany, Italy, and Holland but not France*	*Soviet countries*	*England*	*France*
Lines per frame	525	625	625	405	819
Frame rate	30	25	25	25	25
Field frequency, cps	60	50	50	50	50
Line frequency, cps	15,750	15,625	15,625	10,125	20,475
Video bandwidth, Mc	4	5	6	3	10.4
Channel width, Mc	6	7	8	5	14
Video modulation polarity	Negative	Negative	Negative	Positive	Positive
Sound modulation	FM	FM	FM	AM	AM

* Adapted from C. H. Owen, Television Broadcasting, *Proc. IRE*, May, 1962.

Appendix D

DEVELOPMENT OF TELEVISION BROADCASTING

1941 FCC assigned present VHF channels 2 to 13 for commercial television broadcasting, with standards used now. Channel 1, 44–50 Mc, was assigned to other services because of interference problems.

1946 Marketing of first popular television receiver, RCA model 630TS. This was a 30-tube receiver with a 10-in. screen. Modern circuit innovations included horizontal AFC (Synchrolock), flyback high voltage, reaction scanning, stagger-tuned i-f amplifiers, and a tuner for Channels 1 to 13.

1949 Experimental color television systems demonstrated by Radio Corporation of America and Columbia Broadcasting System. The CBS method used a color wheel rotating in front of a black and white kinescope, in a dot sequential system that required incompatible scanning frequencies for the standard 6-Mc channel. The RCA method was all electronic and compatible with monochrome broadcasting. The FCC approved the CBS color system for commercial broadcasting, because of excellent color reproduction, but this was later withdrawn in favor of the NTSC compatible system used now.

1952 UHF channels 14 to 83 assigned by the FCC to provide for more television broadcast stations. Some channels are reserved for educational television operated by schools.

1953 Color television standards of National Television Systems Committee authorized by FCC, for a system compatible with black and white (monochrome) broadcasting. This system is similar to the RCA compatible electronic system. All commercial color television broadcasting, however, follows the NTSC standards exactly, as described in Chapter 24.

1962 World wide television transmission using satellites circling the earth started with Telstar project of American Telephone and Telegraph Company.

1964 All receivers shipped in interstate commerce must have VHF and UHF tuners. This federal law was passed to increase the use of UHF television broadcast channels.

*Table D · 1 Number of television stations and receivers**

	Commercial television stations authorized	*Television receivers in use*
1946	30	5,000
1947	66	150,000
1948	109	1,010,000
1950	109	9,785,000
1952	108	21,460,000
1954	573	32,750,000
1956	609	42,810,000
1958	665	49,900,000
1960	653	56,210,000

* C. H. Owen, Television Broadcasting, *Proc. IRE*, May, 1962.

Appendix E

TIME CONSTANT AND TRANSIENT RESPONSE

In capacitive and inductive circuits, the voltage or current that results with a sudden change of applied voltage is called the *transient response.* This application is important for the nonsinusoidal waveshapes used in sawtooth deflection, sync pulse circuits, and square-wave testing.

RC charge

Figure E · 1*a* shows a circuit for charging C through the series R by means of the 100-volt battery. Note that the battery and switch are used here to provide a square-wave pulse of applied voltage. The leading edge of such a pulse means that maximum voltage is applied. Furthermore, the voltage is maintained for the time of the pulse width. This input voltage corresponds to closing the switch in the simplified circuit. The trailing edge of the pulse is equivalent to removing the applied voltage by opening the switch.

When the switch is closed, current flows in the direction shown to charge C. Since the voltage across C is proportional to its charge Q, a small voltage e_c appears across the capacitor. Remember that $Q = CE$ for a capacitor; or the charge develops a potential difference E equal to Q/C. If the charging continues until C is fully charged, e_c will equal the battery voltage. When the voltage across C equals the applied voltage, the charging current drops to zero.

Initially, when the switch is closed the amount of current flowing is maximum, since at that time there is no voltage across C to oppose the applied voltage. Therefore, C charges most rapidly at the beginning of the charging period. As C becomes charged, the capacitor voltage opposing the battery becomes greater and the net applied voltage for driving current through the circuit becomes smaller. Plotting the capacitor voltage e_c against time as it charges, the RC charge curve of Fig. E · 1*a* is obtained. The capacitor voltage increases at a decreasing rate, as shown by the exponential curve, because the charging current decreases as the amount of charge on the capacitor increases.

The voltage across the resistor e_R is always equal to $i \times R$. This voltage has the same waveshape as the current because the resistance is constant, making the voltage vary directly with the current. When the capacitor is completely charged, the current is zero and the voltage across the resistor is also zero. At any instant the voltage drops around the circuit must equal the applied voltage. Or, e_R must be equal to $E - e_c$. As an example, with 100 volts applied, at the time when e_c is charged to 63 volts, e_r then is 37 volts.

RC discharge

Figure E · 1*b* shows a circuit for discharging C through the same R. Let the voltage across C be 100 volts at the instant the switch is closed for discharge. As electrons from the negative plate of the capacitor flow around the circuit to the positive plate, C loses its charge and e_c decreases exponentially as shown. The voltage decreases most rapidly at the beginning of the discharge because e_c, now acting as the applied voltage, then has its highest value and can drive maximum discharge current around the circuit.

The discharge current has its peak value of E_c/R at the first instant of discharge. Its direction is opposite to the current flow on charge because the capacitor is now acting as the generator and is connected on the opposite side of the series resistance R. As C discharges,

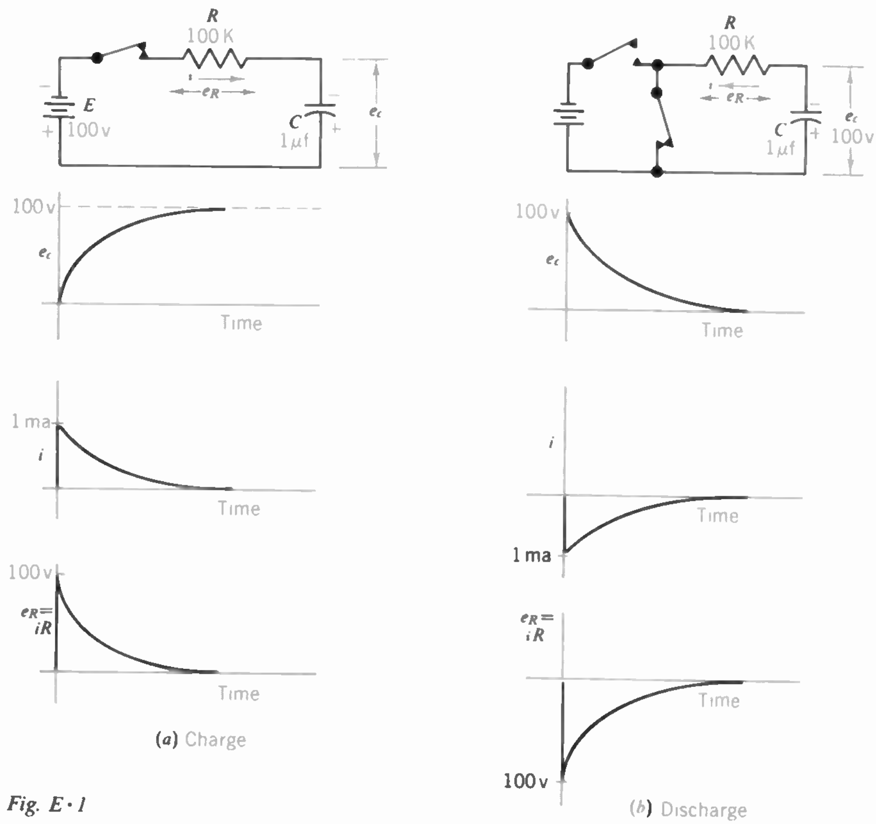


Fig. E·1

its current decreases because of the declining value of e_c. After C is completely discharged the current is zero. The voltage across R is equal to iR and has the same waveshape as the current.

RC time constant

The series R limits the amount of current. Therefore, higher resistance results in a longer time for charge or discharge. Larger values of C also require a longer time. A convenient measure of the charge or discharge time of the circuit, therefore, is the RC product. The time constant is defined as

$$R \times C = t$$

where C = capacitance in farads, R = resistance in ohms in series with the charge or discharge current, and t is the time constant in seconds. As an example, with 1 μf C in series with 0.1 megohm R, the time constant equals 0.1 sec.

The time constant states the time required for C to charge to 63 per cent of the applied voltage. For discharge, the time constant states the time required for C to discharge 63 per cent, or down to 37 per cent of its original voltage. In our example, C charges 63 per cent in 0.1 sec. If the applied voltage is 100 volts, after 0.1 sec of charge e_c will be 63 volts. If the capacitor discharges from 100 volts, e_c will be 37 volts after 0.1 sec. The capacitor is practically completely charged to the applied voltage after a time equal to five time constants. On discharge, the capacitor voltage is practically down to zero after five time constants. See the universal charge and discharge curves in Fig. E·3.

Note the following facts about charge or discharge of the capacitor. When the applied voltage is more than e_c the capacitor will charge. The capacitor will continue to charge as long as the applied voltage is maintained and is greater than e_c. The rate of charging is determined by the RC time constant. When e_c equals the applied voltage, however, the capacitor cannot take on any more charge, regardless of the time constant. Furthermore, if the applied voltage decreases, the capacitor will discharge. The capacitor will discharge as long as its voltage is greater than the applied voltage. When the capacitor voltage discharges down to zero it cannot discharge any more.

In summary, the capacitor charges as long as the applied voltage is greater than e_c. Similarly the capacitor discharges as long as e_c is able to produce discharge current. Note that if C is discharging and the applied voltage changes to become greater than e_c, the capacitor will stop discharging and start charging.

Also, it should be noted that the 63 per cent change of voltage in RC time refers to 63 per cent of the net voltage available for producing charging current on charge, or discharging current on discharge. As an example, if 100 volts is applied to charge a capacitor that already has 20 volts across it, the capacitor voltage will increase by 63 per cent of 80 volts, in one RC time, adding 50.4 volts. The 50.4 volts added to the original 20 volts produces 70.4 volts across the capacitor.

L/R time constant

When voltage is applied to an inductive circuit, the current cannot attain its steady-state value instantaneously but must build up exponentially because of the self-induced voltage across the inductance. This rise of inductive current corresponds to the rise of voltage across a capacitor (see Fig. E·2). The transient waveshapes are exactly the same, with current in the inductive circuit substituted for capacitor voltage. Since the voltage across a resistor is $i \times R$, the e_R voltage with inductive current corresponds to the e_c voltage. This parallel between

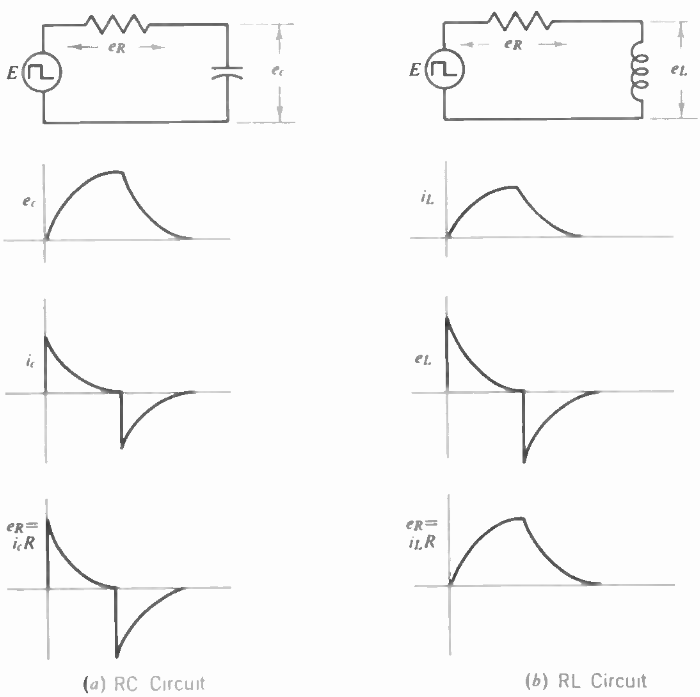


Fig. E·2

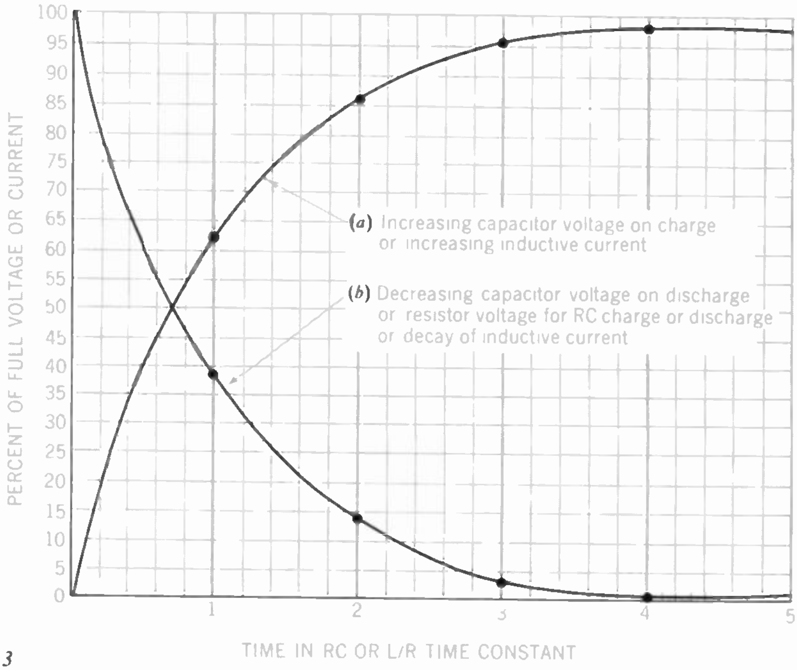


Fig. E · 3

inductive current and capacitor voltage applies to either a current rise or decay, corresponding to the capacitor charge or discharge.

The time constant for an inductive circuit is

$$\frac{L}{R} = t$$

where L is in henrys, R in ohms, and t in seconds. This time constant indicates the same 63 per cent charge as for the RC time constant, but for current in the RL circuit. Also, the transient response is completed in five time constants. However, note that higher resistance provides a shorter time constant for the current rise or decay in an inductive circuit.

Universal time-constant graph

With the curves in Fig. E · 3 you can determine voltage and current values for any amount of time. The rising curve *a* shows how the voltage builds up across *C* as it charges in an RC circuit; the same curve applies to the current increasing in the inductance for an RL circuit. The decreasing curve *b* shows how the capacitor voltage declines as *C* discharges in an RC circuit or the decay of current in an inductance.

Note that the horizontal axis is in units of time constants, rather than absolute time. The time constant is the RC product for capacitive circuits but equals L/R for inductive circuits. For example, suppose that the time constant of a capacitive circuit is 5 μsec. Therefore, one RC unit corresponds to 5 μsec, two RC units equals 10 μsec, three RC units equals 15 μsec, etc. To find how much voltage is across the capacitor after 10 μsec of charging, take the value on graph *a* corresponding to two time constants, or approximately 86 per cent of the applied charging voltage. The point where curves *a* and *b* intersect shows that a 50 per cent change is accomplished in 0.7 time constant. These values are listed in Table E · 1. The curves can be considered linear within the first 40 per cent of change.

In Fig. E · 3 the entire RC charge curve actually adds 63 per cent of the net charging voltage for each increment of RC time, although it may not appear so. In the second interval of RC

Table E · 1 Charge or discharge percentages

FACTOR	CHANGE, PER CENT
0.5 time constant	40
0.7 time constant	50
1 time constant	63
2 time constants	86
3 time constants	96
4 time constants	98
5 time constants	99

time, for instance, e_c adds 63 per cent of the net charging voltage, which is $0.37E$. Then 0.63×0.37 equals 0.23, which is added to 0.63 to give 0.86 or 86 per cent.

The decay curve b in Fig. E · 3 is plotted from the following equation:

On discharge, $$e_R = e_c = E_c\,(\epsilon^{-t/RC})$$

where E_c is capacitor voltage at the start of discharge and e_c is its voltage at time t, while the RC product is in ohms and farads for time in seconds. The constant factor ϵ is the base in the Napierian or natural system of logarithms, equal to 2.718. We can say $e_R = e_c$ on discharge because the magnitudes must be the same then, with R across C and no other voltage applied.

Converting to common logarithms, log ϵ to base 10 equals 0.434. Therefore,

On discharge, $$e_R = \text{antilog}\,(\log_{10} E_c - 0.434\,t/RC) \qquad \text{(E · 1)}$$

With this formula, you can calculate any value on curve b.

Note that this decay curve applies to e_c on discharge and to e_R for either charge or discharge. The only other possibility is e_c on charge for curve a. For these values, calculate e_R and subtract from the applied voltage E_a to obtain the e_c voltage on charge.

As an example for discharge, we can calculate e_c after a discharge time of one time constant, where $t = RC$, starting from E_c equal to 100 volts. Then

$$\begin{aligned} e_c &= \text{antilog}\,(\log 100 - 0.434) \\ &= \text{antilog}\,(2 - 0.434) = \text{antilog}\ 1.566 \\ e_c &= 37 \text{ volts} \end{aligned}$$

If the example were to find e_c after one RC of charge, from a 100-volt E_a, we could calculate e_R first. The e_R value is the same 37 volts because the formula applies to e_R for either charge or discharge. Then subtract 37 from 100 to obtain e_c as equal to 63 volts. This is the voltage across C after one RC of charge.

Since the formula for the RC decay curve is convenient for calculations, we can transpose terms to have t alone for calculating time. To generalize the terms, e_s (for start) will be used instead of E_c, and e_f (for finish) instead of e_c. Then the equation can be used to calculate time for (1) capacitor discharge voltage and (2) resistor voltage for either charge or discharge. Using e_f and e_s,

$$e_f = \text{antilog}\,(\log e_s - 0.434\,t/RC)$$

Taking the log of both sides,

$$\log e_f = \log e_s - 0.434\,t/RC$$

Transposing terms,

$$0.434\ t/RC = \log e_s - \log e_f = \log\left(\frac{e_s}{e_f}\right)$$

$$t = \frac{RC}{0.434}\log\left(\frac{e_s}{e_f}\right)$$

$$t = 2.3\ RC\log\left(\frac{e_s}{e_f}\right) \qquad \text{(E}\cdot\text{2)}$$

As an example, we can calculate the time for the resistor voltage to drop from 100 to 50 volts, in a circuit with an RC time constant of 50×10^{-6} sec. Then

$$t = 2.3 \times 50 \times 10^{-6} \times \log\left(\frac{100}{50}\right)$$

$$= 2.3 \times 50 \times 10^{-6} \times 0.3$$

$$t = 34.5 \times 10^{-6}\ \text{sec}$$

This is the time for e_R to drop from 100 to 50 volts while e_c is discharging between the same values. The time is the same when e_c is charging from 0 to 50 volts with 100 volts applied. Note that this time is practically equal to 70 per cent of the 50 μsec time constant for a 50 per cent change, as in Table E · 1 and Fig. E · 3.

Long and short time constants

A long time constant is at least five times longer than the period of the applied voltage. Then the capacitor cannot take on any appreciable charge before the applied voltage drops to start the discharge. Also, there is little discharge before charging voltage is applied again. In an RC circuit with a long time constant, therefore, the applied voltage is mainly across R, with little voltage across C. An RC coupling circuit is an example. Note that a long time constant corresponds to little capacitive reactance, compared with R, as both require a large value of capacitance.

A short time constant is one-fifth, or less, of the period of applied voltage. This combination permits C to become completely charged before it discharges. Then the applied voltage is on long enough to charge the capacitor for five time constants or more. Also, the capacitor can discharge completely before it is charged again by the applied voltage. In an RC circuit with a short time constant, therefore, the applied voltage is mainly across C (see Fig. E · 4).

Differentiated output

In an RC circuit with a short time constant, the voltage across R is called differentiated output because its amplitude can change instantaneously, in either polarity. In Fig. E · 4, the voltage across R consists of positive and negative spikes corresponding to the leading and trailing edges of the square-wave input. Such sharp pulses are useful for timing applications. In general, differentiated output for pulses corresponds to a high-pass filter for sine waves. In both cases, the high-frequency components of the input are in the output circuit.

In any circuit with nonsinusoidal input, the differentiated output has sharp changes in amplitude. This applies to i and e_R in an RC circuit but not to e_c as the voltage across the capacitor cannot change instantaneously because of its stored charge. In an RL circuit, e_L can provide differentiated output. The current cannot change instantaneously in an inductance because of its magnetic flux.

The instantaneous values of e_L in an RL circuit can be calculated as $L\ di/dt$, where di/dt is the time rate of change of the current. Similarly, in an RC circuit the instantaneous value of i_c is equal to $C\ dv/dt$, where dv/dt is the time rate of change of the capacitor voltage.

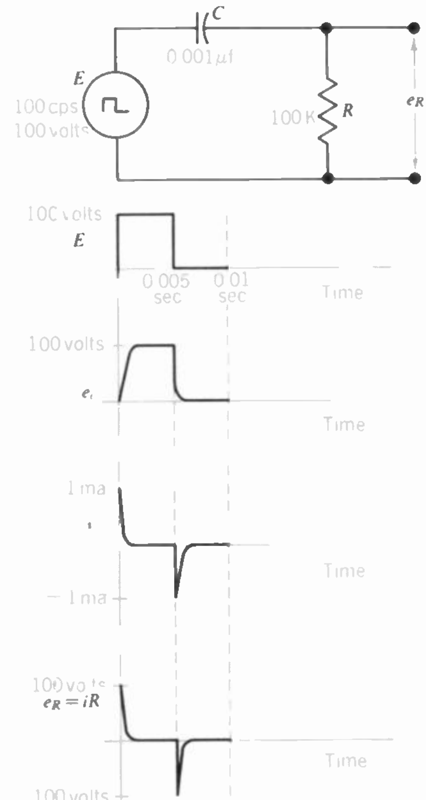


Fig. E·4

Integrated output

In general, an integrated waveform does not have instantaneous changes in amplitude and cannot reverse in polarity. Therefore, e_c in an *RC* circuit can provide integrated output. In comparison with *RC* circuits for sine waves, the integrated output corresponds to the action of a low-pass filter. In this case, a shunt capacitor can be considered a bypass for the high-frequency components of the input. For *RL* circuits, i_L and e_R can provide integrated output, since the current cannot change instantaneously.

Nonsinusoidal waveshapes

For sine-wave circuits, the capacitive or inductive reactance determines the phase angle. With applied voltage that is not a sine wave, there can be changes of waveshape, in either voltage or current, instead of the changes in phase angle that apply only for sine waves. Instead of reactance, the time constant of the *RC* or *RL* circuit must be considered.

Some examples of changes in waveshapes by *RC* and *RL* circuits are illustrated in Fig. E·5. The sawtooth capacitor voltage in *a* results from charging *C* slowly through a high *R* in a long time constant circuit, and discharging *C* fast through a small *R* in a short time constant circuit. The capacitor charge and discharge current has the rectangular waveshape because of the uniform rate of change for the sawtooth voltage. The values of i_c are calculated as $C\, dv/dt$ for a 2-μf capacitor. Similarly, in *b* the sawtooth current through an inductance corresponds to rectangular voltage. Here, the values of e_L are calculated as $L\, di/dt$ for a 2-henry inductance.

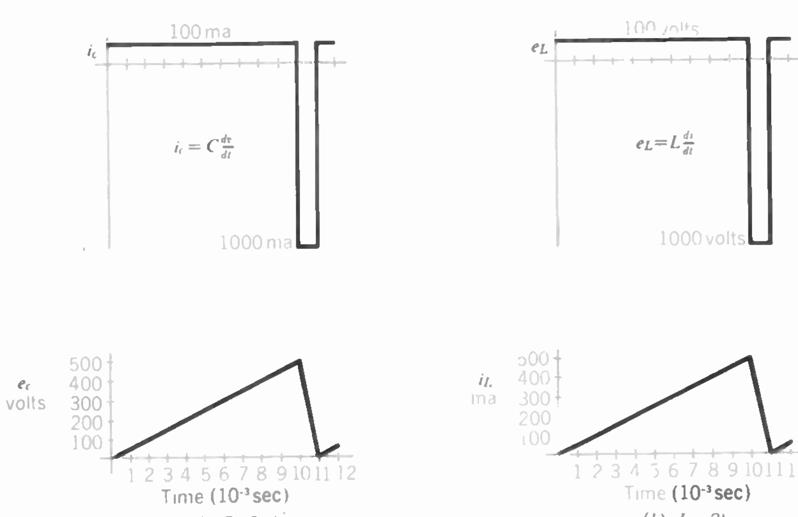


Fig. E·5

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Answers to self-examinations

Chapter 1

1. 30	2. 525	3. 262½
4. 2	5. 15,750	6. 63.5
7. 15,750	8. 60	9. Contrast
10. Detail, resolution, or definition	11. Camera	12. Picture
13. 6	14. AM	15. FM
16. 174 to 180	17. Synchronizing	18. Blanking
19. Cutoff	20. 15,750	

Chapter 2

1. T	2. T	3. T
4. T	5. T	6. T
7. T	8. T	9. T
10. T	11. T	12. T
13. F	14. T	15. T
16. T	17. T	18. T
19. T	20. T	

Chapter 3

1. (*b*)	2. (*a*)	3. (*d*)
4. (*a*)	5. (*b*)	6. (*c*)
7. (*d*)	8. (*a*)	9. (*b*)
10. (*d*)		

Chapter 4

1. T	2. T	3. F
4. T	5. F	6. F
7. F	8. T	9. T
10. T	11. T	12. T
13. T	14. T	15. T
16. T	17. F	18. T
19. T	20. T	

Chapter 5

1. (*b*)	2. (*b*)	3. (*d*)
4. (*a*)	5. (*c*)	6. (*c*)
7. (*d*)	8. (*b*)	9. (*b*)
10. (*a*)		

Chapter 6

Part A

1. (*e*)	2. (*a*)	3. (*b*)
4. (*c*)	5. (*h*)	6. (*d*)
7. (*f*)	8. (*g*)	

Part B

1. (*g*)	2. (*d*)	3. (*a*)
4. (*b*)	5. (*c*)	6. (*e*)
7. (*f*)	8. (*j*)	9. (*i*)
10. (*h*)		

Part C

1. (*b*)	2. (*c*)	3. (*d*)
4. (*e*)	5. (*f*)	6. (*a*)

Chapter 7

1. (*b*)	2. (*c*)	3. (*d*)
4. (*c*)	5. (*d*)	6. (*b*)
7. (*d*)	8. (*b*)	9. (*b*)
10. (*b*)		

Chapter 8

1. (*a*)	2. (*d*)	3. (*c*)
4. (*c*)	5. (*b*)	6. (*d*)
7. (*a*)	8. (*d*)	9. (*a*)
10. (*b*)		

Chapter 9

1. (*a*)	2. (*b*)	3. (*c*)
4. (*b*)	5. (*b*)	6. (*c*)
7. (*d*)	8. (*c*)	9. (*d*)
10. (*c*)		

Chapter 10

1. 35	2. 16	3. 707
4. 1,414	5. 40	6. 25
7. (*a*) 1,600 (*b*) 32 (*c*) 5.6	8. (*a*) 2,400 (*b*) 56	9. 2,880
10. 10.1	11. 36.4	12. (*a*) 10 (*b*) 3,000
13. 6.6	14. 5	15. 1
16. (*a*) 0.7 (*b*) 58.7 (*c*) 126 (*d*) 0.6	17. 2.4	18. 80

Chapter 11

1. (*c*)	2. (*b*)	3. (*c*)
4. (*d*)	5. (*a*)	6. (*b*)
7. (*a*)	8. (*b*)	9. (*b*)
10. (*c*)		

Chapter 12

1. T	2. T	3. F
4. T	5. T	6. F
7. F	8. F	9. F
10. T	11. F	12. F
13. T	14. F	15. T

Chapter 13

1. (*b*)	2. (*d*)	3. (*b*)
4. (*d*)	5. (*b*)	6. (*d*)
7. (*b*)	8. (*d*)	9. (*c*)
10. (*b*)		

Chapter 14

1. T	2. T	3. T
4. T	5. T	6. T
7. F	8. T	9. T
10. T	11. F	12. T
13. T	14. T	15. T
16. T	17. T	18. T
19. F	20. T	

Chapter 15

Part A

1. F	2. F	3. T
4. T	5. F	6. T
7. T	8. T	9. T
10. T	11. T	12. T
13. T	14. T	15. T
16. F	17. T	18. T
19. T	20. T	

Part B

1. 0.15	2. 0.00045	3. 0.003
4. 250	5. 1,000	6. 15,750
7. 0.033	8. 16.4	9. 30
10. 300		

Chapter 16

1. (*b*)	2. (*b*)	3. (*a*)
4. (*d*)	5. (*c*)	6. (*a*)
7. (*c*)	8. (*c*)	9. (*a*)
10. (*b*)		

Chapter 17

1. (*d*)	2. (*a*)	3. (*c*)
4. (*b*)	5. (*b*)	6. (*a*)
7. (*c*)	8. (*b*)	9. (*a*)
10. (*b*)		

Chapter 18

1. (*d*)	2. (*c*)	3. (*b*)
4. (*a*)	5. (*b*)	6. (*c*)
7. (*c*)	8. (*b*)	9. (*d*)
10. (*d*)		

Chapter 19

Part A

1. 41.25	2. 4.5	3. 4,000
4. 5	5. 6,750	6. 27.25
7. 42.75	8. 39.75	9. 2

10. 6,000	11. 24	12. 6,000
13. 10	14. 4	15. 0.05

Part B

1. T	2. T	3. T
4. T	5. T	6. T
7. T	8. T	9. T
10. T	11. T	12. T
13. T	14. T	15. T

Chapter 20

1. (*d*)	2. (*c*)	3. (*c*)
4. (*a*)	5. (*d*)	6. (*c*)
7. (*b*)	8. (*b*)	9. (*b*)
10. (*b*)		

Chapter 21

1. (*b*)	2. (*c*)	3. (*c*)
4. (*c*)	5. (*d*)	6. (*c*)
7. (*d*)	8. (*d*)	9. (*a*)
10. (*c*)		

Chapter 22

Part A

1. 96.3	2. 10	3. 20
4. 50	5. 4	6. 4
7. 90°	8. 75	9. Zero
10. 60		

Part B

1. T	2. T	3. T
4. F	5. T	6. T
7. T	8. T	9. T
10. F	11. T	12. T
13. T	14. T	15. F
16. T	17. T	18. T
19. T	20. T	

Chapter 23

1. (*j*)	2. (*d*)	3. (*k*)
4. (*m*)	5. (*t*)	6. (*n*)
7. (*s*)	8. (*l*)	9. (*e*)
10. (*f*)	11. (*g*)	12. (*a*)
13. (*h*)	14. (*b*)	15. (*i*)
16. (*c*)	17. (*o*)	18. (*r*)
19. (*p*)	20. (*q*)	

Chapter 24

1. (*c*)	2. (*a*)	3. (*d*)
4. (*a*)	5. (*d*)	6. (*a*)
7. (*d*)	8. (*d*)	9. (*b*)
10. (*c*)		

Chapter 25

1. (*d*)
2. (*c*)
3. (*b*)
4. (*b*)
5. (*c*)
6. (*a*)
7. (*a*)
8. (*b*)
9. (*d*)
10. (*d*)

Answers to odd-numbered problems

Chapter 1

1. (*a*) 31.75 μsec
   (*b*) 64 μsec
3. 120,000
5. 0.25 μsec

Chapter 2

1. 0.2 volt
3. 1,024
5. 2 μa, 4 μa, and 8 μa

Chapter 3

1. 60 cps and 31,500 cps
3. 5 μsec and 667 μsec
5. (*a*) 63.5 μsec
   (*b*) 1/60 sec

Chapter 4

1. 2.82 Mc
3. 170
5. (*a*) 1,000 μsec
   (*b*) 960 μsec
   (*c*) 5.2 kc
7. (*a*) 359
   (*b*) 272
   (*c*) 423
   (*d*) 531

Chapter 5

1. (*a*) 55.0 Mc, 55.5 Mc, and 55.25 Mc
   (*b*) 80.25 Mc and 77.25 Mc
   (*c*) 470.75 Mc, 471.75 Mc, and 471.25 Mc
   (*d*) 475.25 Mc and 471.25 Mc
3. (*a*) 55.24 Mc
   (*b*) 77.26 Mc
   (*c*) 187.26 Mc
   (*d*) 193.24 Mc
5. 61.25 Mc and 65.75 Mc.

Chapter 6

No problems

Chapter 7

1. Push yoke forward
3. Turn yoke
5. Pincushion magnets
7. (*a*) −20 to −40 volts
   (*b*) 300 to 500 volts
   (*c*) 10 to 20 kv
   (*d*) 0 volts
   (*e*) 2 to 4 kv
   (*f*) −30 to +300 volts
9. (*a*) 0.2 ma, 0.3 ma, 0.35 ma, and 0.55 ma
   (*b*) Cathode drive and 500 volts on G2

Chapter 8

1. (*a*) 40 volts
   (*b*) 200 volts
   (*c*) 10 ma
3. (*a*) 267 ohms
   (*b*) 14 ohms
5. 42 ohms, 0.945 watt

7. (*a*) 0
   (*b*) 120 volts
   (*c*) 12.6 volts
   (*d*) 86.9 volts

Chapter 9

1. (*a*) 68 volts
   (*b*) 0.2 ma
   (*c*) 1.7 ma

3. 15.9 kc

5. (*a*) 0.025 μsec
   (*b*) 2,500 μsec

7. (*a*) 56.3 kc
   (*b*) 1,312 cps

Chapter 10

1. (*a*) 16.5 μμf
   (*b*) 2.1 Mc

3. 0.8 Mc

5. 2,400 ohms

7. 5,000 ohms

9. E_b is 170 volts
   E_s is 140 volts
   E_k is 2.76 volts

11. (*a*) 4.48 Mc
    (*b*) 2.56 Mc
    (*c*) 3.2 Mc

13. (*a*) 53 ohms
    (*b*) 0.74

15. 8.6

Chapter 11

1. 10 volts

3. (*a*) +150 volts
   (*b*) 150 volts
   (*c*) −27 volts
   (*d*) 177 volts
   (*e*) +27 volts
   (*f*) 123 volts

5. (*a*) −0.9 volt
   (*b*) −3.6 volts
   (*c*) −27 volts

Chapter 12

1. R_L = 4,400 ohms
   L_c = 155 μh

3. 3 ma, 8 ma, 22 ma, 40 ma

Chapter 13

1. (*a*) 0.1
   (*b*) 0.00015
   (*c*) 0.0003
   (*d*) 0.141

3. (*a*) 20
   (*b*) 80
   (*c*) 6
   (*d*) 3
   (*e*) 14
   (*f*) 23
   (*g*) 28

5. (*a*) 80
   (*b*) 70
   (*c*) 3,160

7. 10,000 or 80 db

Chapter 14

1. (*a*) 40 μsec
   (*b*) 28 μsec
   (*c*) 40 μsec
   (*d*) 200 μsec

3. 44 μsec, 41 μsec, and 39 μsec

5. (*a*) 120 volts
   (*b*) 90,000 ohms

Chapter 15

1. (*a*) 10 volts
   (*b*) 40 volts
   (*c*) 50 volts
   (*d*) 63 volts
   (*e*) 86 volts
   (*f*) 99 volts

3. (*a*) 73 volts
   (*b*) 27 volts

5. (*a*) 100 μa
   (*b*) 100 volts
   (*c*) 50 μa
   (*d*) 50 volts

7. (*a*) 100 volts
   (*b*) 90 volts

9. (*a*) 200 volts
   (*b*) 250 volts
   (*c*) 100 volts

Chapter 16

1. +1.5 volts
3. (*a*) 120 μsec
   (*b*) 8 μsec
5. 360 cps
7. 0.028 henry

Chapter 17

1. 32 volts
3. (*a*) 4,900 ohms
   (*b*) 13 henrys
5. 33⅓ volts

Chapter 18

1. (*a*) 5,000 μsec
   (*b*) 79
   (*c*) Yes
3. (*a*) 75 ma
   (*b*) 705 watts
5. (*a*) 0.96 watt
   (*b*) 24 volts
7. 100 μμf

Chapter 19

1. $Q = 15$
   $f = 2$ Mc
   $Z_L = 4{,}500$ ohms
3. (*a*) 146.75 Mc
   (*b*) 106.75
5. 800, 8,000, and 4,000
7. (*a*) 4 μh
   (*b*) 10 μμf
   (*c*) 8 μh
9. 30 Mc with Q of 10
   31.29 Mc with Q of 20
   28.71 Mc with Q of 20

Chapter 20

1. 101 Mc and 257 Mc
3. 2.09 ft
5. 47 Mc
7. 106.75 Mc
9. 0.028 ft
11. 0.54 volt

Chapter 21

1. 1,000 μv, 600 μv, 200 μv, and 0
3. Dipole = 8.6 ft
   Reflector = 9.0 ft
   Spacing = 3.44 ft
5. 3.76 ft
7. 2.31 ft
9. 4 ft
11. −20 db
13. −12 db
15. 69 ohms

Chapter 22

1. (*a*) 5 kc
   (*b*) 15 kc
   (*c*) 30 kc
   (*d*) 45 kc
3. 3 db down at 2.133 kc
5. 8 Mc
7. (*a*) 15.9 μμf
   (*b*) 1,590 μμf
   (*c*) 1,590 μh
   (*d*) 15.9 μh
9. (*a*) V.A. = 36
   (*b*) 10.8

Chapter 23

1. 32 ma
3. 15 ma
5. 130 volts
7. 30°
9. 2,000 cps, 3,000 cps, 1,500 cps

Chapter 24

1. $R = 0.4$, $G = 0.4$, $B = 1.0$, $Y = 0.466$

3. (*a*) -0.11 and $+0.11$
   (*b*) $+0.89$ and -0.89
   (*c*) -0.11 and $+0.11$

5. $R = 1.0$, $G = 1.0$, $B = 1.0$, $I = 0$, $Q = 0$, $C = 0$

7. $C = 0.5$; green hue

9. 118.7°

Chapter 25

No problems

Index