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TELEVISION COURSE

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SECOND EDITION

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PREFACE

Ten years ago we brought to the young television service industry the Photofact Television Course. Through fourteen separate printings, this remarkable book has helped in the training of thousands of today's practicing, professional electronic service technicians. They literally "cut their eye teeth" on this book which has won for itself both high praise and widespread recognition.

It is extremely gratifying to me personally to bring to the electronic industry this entirely new and completely up-to-date edition. I am confident that this revised edition will help further the successful careers of thousands of tomorrow's new technicians.

As in the original edition, we have put our best efforts and the most recent knowledge, skill and techniques into the present book. You are assured of the most accurate, complete and timely information intended to familiarize those with experience and training in radio fundamentals with the basic principles of practical television theory and operation.

Much has happened in the past ten years, but the words we used then are just as true today: "The art of television is destined for phenomenal growth and development."

The information contained in this book is basic and practical and is intended to help you take part in that growth.

Howard W. Sams

May, 1959

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Introduction

Television is a specialized branch of the electronics industry. It requires study of the various types of circuits used in the production of high voltages and special waveshapes, together with a thorough knowledge of the cathode-ray tube.

Modern television receivers are designed around the cathode-ray tube because practically every circuit ultimately contributes to the operation and control of the cathode ray, or electron beam as it is more commonly called.

The cathode-ray tube is essentially a vacuum tube producing a narrow beam of electrons. These electrons are accelerated by a large positive potential and rapidly move toward a chemically prepared screen which fluoresces when bombarded by high-speed electrons. The beam can be deflected and modulated (changed in intensity) so easily that it has many desirable properties for television.

Television requires the use of two distinctly different types of cathode-ray tubes—(1) the camera tube at the television studio, and (2) the picture tube for reproducing the transmitted picture.

The camera tube converts light energy into electrical energy which represents the video signal. This signal from the camera is amplified and used to modulate a carrier wave. After transmission and reception, the picture tube then reconverts the video signal back into light energy for viewing. Transmitted simultaneously with the video signal are blanking and synchronizing pulses. These pulses hold the horizontal and vertical scanning circuits in synchronism with the transmitter to produce 30 frames per second, each frame consisting of 525 horizontal lines, according to present-day U. S. standards.

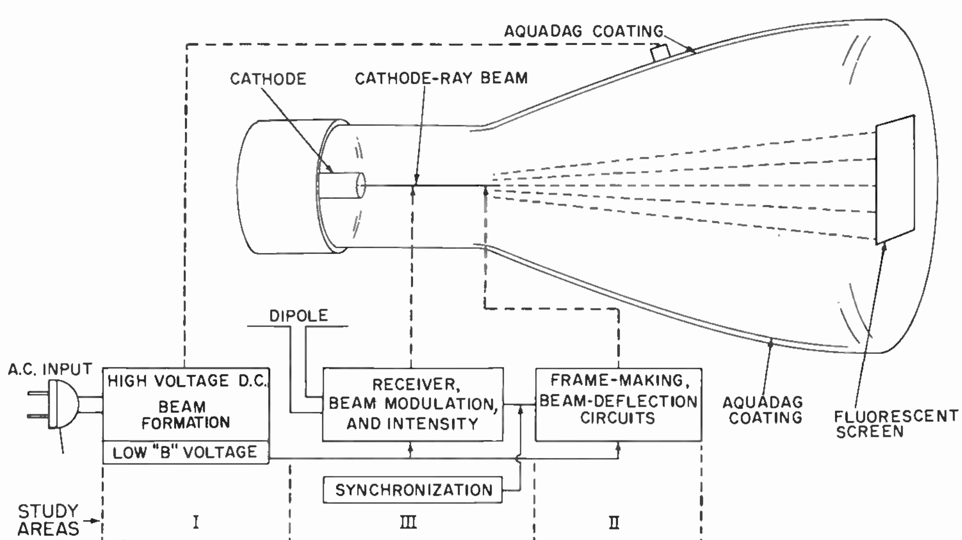
For ease of comprehension in the study of television reception, the receiver is broken down into three major circuit sections, which have independent functions. These sections are shown in the accompanying drawing and are as follows:

I. Beam formation and control — involves the study of high potentials, testing techniques, the cathode-ray tube, and electron optics.

II. Beam-deflection systems—involve the study of blocking oscillators, free-running multivibrators, and waveshaping circuits.

III. Beam modulation and synchronization — involves the study of high-frequency broadband receivers, video detectors, video amplifiers, and DC restorers.

These three master subjects serve as a guide of what and how to study. They are in the correct sequence to give the reader a clear understanding of what takes place in a television receiver. The reader is advised to thoroughly grasp one subject at a time to prevent confusion and disappointment. Studying will not be difficult if this simple rule is followed.



Simplified functional block diagram.

SECTION 1

Beam Formation and Control

Chapter 1

Cathode-Ray Tube — Beam Formation and Electrostatic Control

The cathode-ray tube is also employed in the cathode-ray oscilloscope, one of the most-required instruments used in servicing television receivers. The cathode-ray tubes used in oscillography and television are similar, and the basic principles apply to both; however, the oscilloscope tube is simpler and provides a good introduction to the study of cathode-ray tubes.

BEAM FORMATION

Fig. 1-1 shows a simple cathode-ray tube used in early experiments to develop a narrow beam of electrons. Emitted from a hot filament, the electrons rush forward, attracted by a positively charged plate. Accelerated by this positive charge, the beam reaches a velocity of many thousand miles per second, depending upon the force of this positive attraction. The equal negative charges carried by all the electrons set up a repelling action that causes the emitted beam to widen or scatter. However, the small hole in the center of the anode narrows the beam and permits some of the fast-moving electrons to race to the end of the tube without losing much speed.

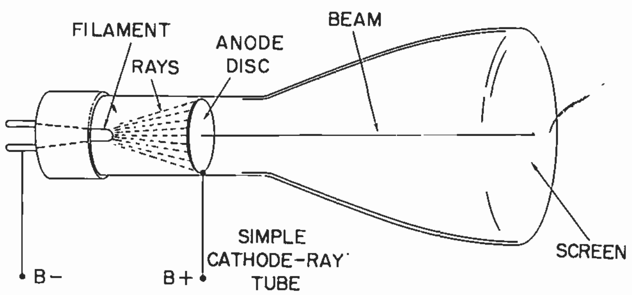


Fig. 1-1. Formation of a narrow beam of electrons. (Early experimental form.)

The electrons making up this narrow beam are traveling too fast for any effective scattering to take place. Thus, a small illuminated spot is produced on the fluorescent screen. The scattered electrons collected by the positively charged anode produce a current through the B supply circuit. This type of cathode-ray tube is quite elementary, but will do as an example in a discussion on the production and formation of a narrow electron beam which strikes a fluorescent screen and develops a spot of light. The color of the spot depends upon the type of chemical in the screen. Oscilloscope tubes generally have a P1 phosphor, which produces a green trace. Television tubes always have a P4 phosphor, which produces a white trace.

Cathode-ray tubes contain the following: (1) An electron beam source. (2) A fluorescent screen for visible indication. (3) A means for varying the intensity, which controls the brilliance of the spot. (4) A method of focusing the beam which controls the size of the spot. (5) Provision for deflecting the beam which controls the position of the spot.

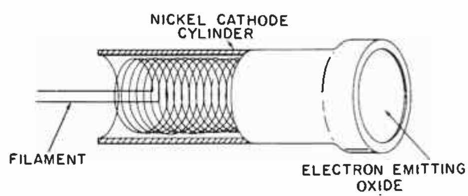


Fig. 1-2. Heater and cathode assembly.

In modern cathode-ray tubes, the electron source is an indirectly-heated cathode, or electron emitter. This cathode is a small cylinder of nickel about one-eighth of an inch in diameter and about one-half inch in length. The nickel sleeve is coated on one end with oxide, which permits a large number of electrons to be emitted toward the fluorescent screen. The heater is a tungsten wire filament wound in the form of a noninductive spiral. The spiral winding tends to cancel any magnetic field that might affect the electron beam. The filament coil is insulated and is inserted into the cathode sleeve. For better heat conduction to the cathode, the insulated filament contacts the nickel sleeve. This important element is illustrated in Fig. 1-2.

BEAM CONTROL

Grid-Control Element

The elementary cathode-ray tube illustrated in Fig. 1-1 has no controlling element to limit the number of electrons emitted from the cathode. The brilliance of the spot on the screen must be controlled. This calls for an additional element, the control grid, between the cathode and the positively-charged anode. In modern tubes, this element is a metal cylinder completely enclosing the cathode element (Fig. 1-3A). The strategic position of the control grid permits controlling the quantity of electrons in the beam. The quantity of electrons limits the spot brilliance because the more negative the control grid is biased with respect to the cathode, the fewer electrons in the beam, and the less the intensity of light produced on the screen. The direction of the electron emission is governed by an aperture in the disc at the end of the grid cylinder.

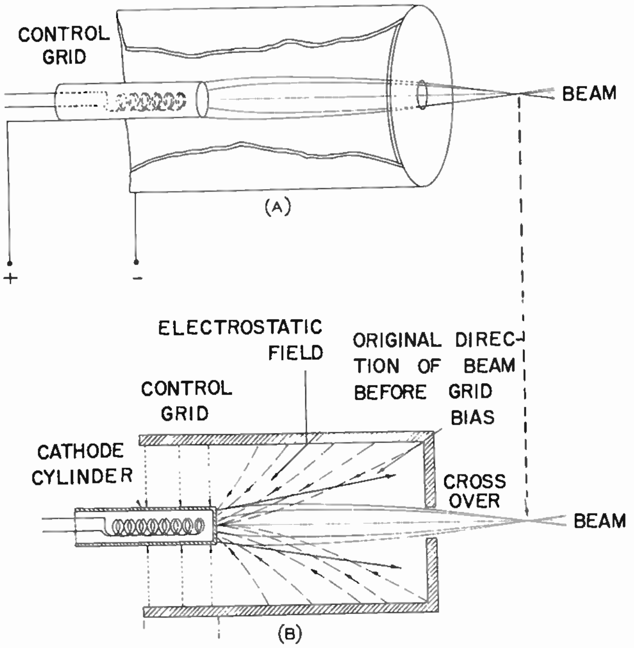


Fig. 1-3. Action of cathode and control-grid assembly.

The lines of force of the electrostatic field developed by the difference of potential between the cathode and control grid and their effect on the beam, are illustrated in Fig. 1-3B.

If the voltage on the control grid is made more negative with respect to the cathode, fewer electrons will be admitted to the beam. If the control grid is sufficiently negative, it will completely shut off, or blank out, the beam.

In earlier cathode-ray guns, the negative field of the control grid also lessened the effect of the positively-charged focus anode located immediately beyond the control grid and further contributed to the reduction of the beam. Many present-day cathode-ray tubes have the anodes placed differently to reduce this interaction between control-grid (brilliance) and focus adjustments. A complete discussion of anode function and structure is given later in this text.

Also, the negative field associated with the control grid causes the beam to cross over itself after it passes through the control-grid aperture (Fig. 1-3B). This crossover is similar to that of a lens and concentrates the beam of electrons in a fine point. Hence, the phrase "electron optics" came into use. The concentration of the beam is shown in Fig. 1-4.

Summing up the foregoing, we recall that the control grid has three functions:

1. Controls the brilliance from zero to maximum at the cathode-ray screen.
2. Lens action concentrates the beam by effecting a crossover.
3. Provides a means for inserting a varying signal to provide intensity modulation of the beam.

A potentiometer varies the control-grid bias and enables the brilliance to be manually adjusted to a comfortable level. In oscillography, this adjustment is provided by the intensity control.

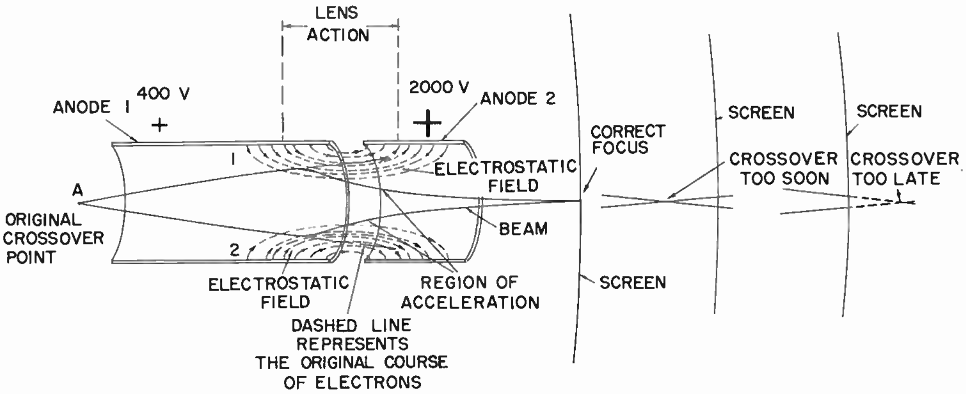


Fig. 1-4. Effect of an electrostatic field on an electron beam.

Focus- and Accelerating-Control Elements

So far, we have been able to control the brilliance of the spot, but yet another control is necessary to bring the spot into sharp focus. The focusing of an electron beam in a cathode-ray tube is similar to the focusing of light (Fig. 1-5). As we understood from the previous discussion, the control grid did focus the beam to a point slightly beyond its aperture, but the beam begins to widen again after the crossover point; therefore, additional focusing is needed. In the earlier picture tubes, focusing was accomplished by two cylindrical anodes (Fig. 1-4).

The two anodes are operated at different potentials, and the lines of force between them establish the lens action in the following manner. The moving electrons are subjected to forces as they enter the electrostatic field set up by anodes 1 and 2. To understand clearly what takes place, refer to Fig. 1-4. Here, the electron beam is seen entering the electrostatic field developed by a difference of potential between the two anodes. Coming from point A, the original crossover, the beam enters a new field. As the electrons cross the first static lines (1 and 2 in Fig. 1-4), they tend to change their course since the forces acting on them will repel negative charges because of the potential gradient between anodes 1 and 2. Although both anodes are positive with respect to the cathode, anode 1, operating at a much lower potential, is negative with respect to anode 2, and an electrostatic field is created which tends to bend the path of the electrons entering that field.

The electrons, traveling at a high speed, are gradually bent into a beam, the greatest repelling force occurring at a point somewhere between the two anodes. When the electrons of the beam converge near the axis of the tube, the lines of force are running almost parallel to the axis and the electrons in the beam begin to accelerate. At the exit end of anode 2, the field is relatively weak, and the electrons keep their direct course, aided by the velocity gained while traveling through anode 2. The two anode cylinders are similar to the grid cylinder, except anode 1, which usually is longer and has two aperture discs spaced for better focus.

After the beam has passed through anode 2, it meets at a second crossover point. This second crossover point is adjusted to take place as the beam arrives at the surface of the screen.

The focus of an electrostatic-type cathode-ray tube, whose construction is shown in Fig. 1-4, is generally adjusted by varying the voltage to anode 1. This controls the amount of force the electrostatic field exerts on the electron beam. By rotating the focus control and observing the screen, the beam can easily be brought into sharp focus.

Before discussing variations in anode structure of the type illustrated in Fig. 1-4, it might be wise to clarify the terminology regarding these elements.

Since the first cathode-ray tubes had two cylindrical anodes (Fig. 1-4), the terms "first anode" and "second anode" were quite logical. However, the newer tubes have more than two anodes or have physically split anodes, and we begin to have difficulty naming them according to their positions in the tube. Therefore, they should be labeled according to their purpose. Consequently, anode 1, which controls the focus, can be called the *focus* anode, and anode 2, the higher potential element, can be called the *accelerating* anode.

Fig. 1-6 shows a more recent form of anode assembly in a typical electrostatic tube. The cylindrical anode adjacent to the control grid is an accelerating anode instead of the focus anode employed in Fig. 1-4. The circular disc with the rather large

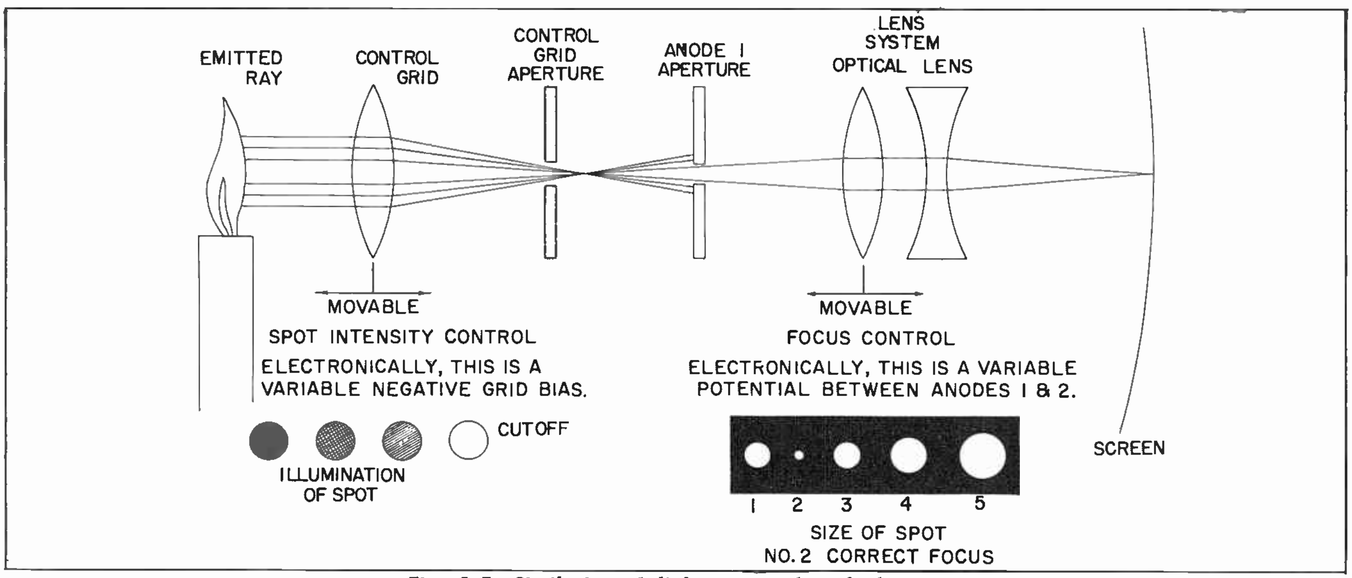


Fig. 1-5. Similarity of light ray and cathode ray.

aperture, following the accelerating anode, is actually the focus anode. The shorter cylindrical anode, combined with the second circular disc, is electrically connected to the first cylindrical anode and is considered part of the accelerating anode structure. The reasons for this construction are as follows:

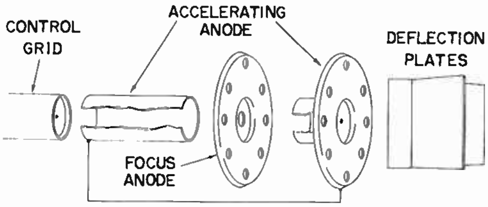


Fig. 1-6. Typical anode assembly.

1. Removal of the focus anode from its position near the control grid lessens any interaction between intensity and focus control adjustments.
2. By proper placement of the focus anode, its aperture can be made larger. The amount of beam current drawn by this anode will be reduced, and its effect on the beam intensity will be lessened.
3. Step 2 is possible because of the separation of portions of the accelerating anode, enabling insertion of the focus anode.
4. In any cathode-ray tubes with appreciable focus-anode current, the focus-control circuits must use sufficient bleed current to insure reliable operation. With the construction outlined in Fig. 1-6, the requirement for bleed currents can be reduced considerably or eliminated altogether.

Remember that construction of the tubes may vary according to the manufacturers' preferences of electrical design and physical support of the various elements.

Summing up, we know that the focus and accelerating anodes have two functions:

1. Focusing the beam for sharpest detail of the image on the screen. The degree of focusing is manually controlled.
2. Acceleration of the beam.

The assembly discussed so far constitutes the electron gun, so called because it shoots bullets (negative particles) to a screen or target. Fig. 1-7 shows a typical electron gun assembly.

Cathode-ray tubes with combined electrostatic focusing and magnetic deflection have been and are being used. The earliest of these tubes were called high-electrostatic focus tubes because they required approximately 3,000 to 5,000 volts for proper focus. These types had to have a high-voltage power supply separate from the regular high-voltage supply. They were soon rendered obsolete by the low-electrostatic focus tubes.

Low-electrostatic focus tubes could be focused by voltages between −100 and +500 volts on the focus anode. These low voltages could be obtained from the low-voltage power supply in the receiver; thus, the extra high-voltage supply could be eliminated. Both the low- and the high-electrostatic tubes re-

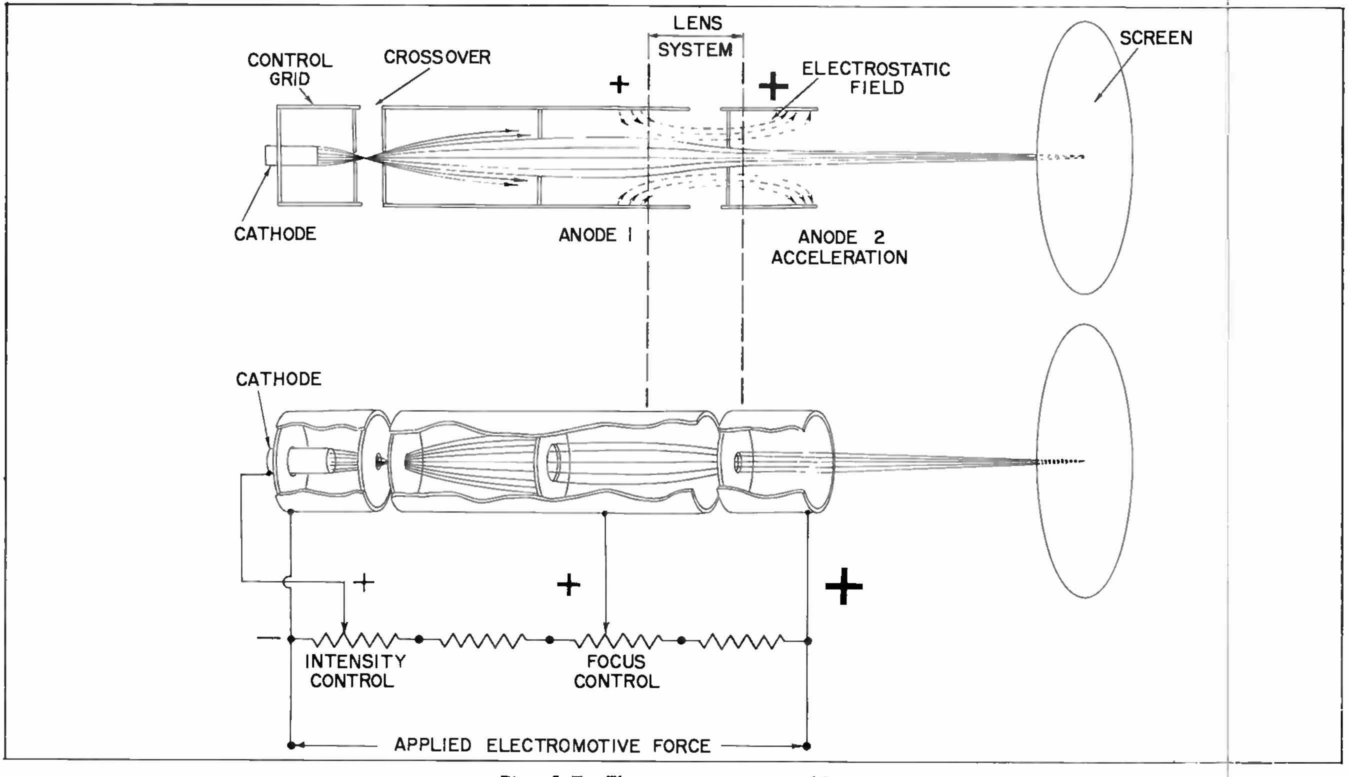


Fig. 1-7. Electron gun assembly.

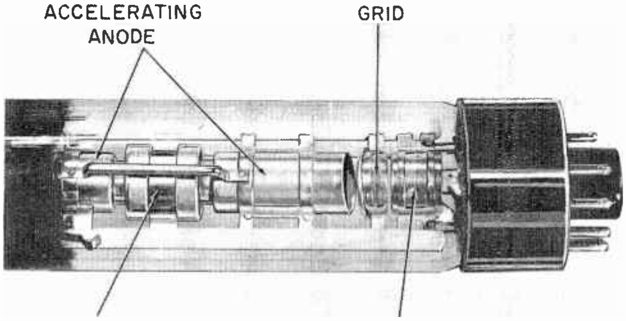


Fig. 1-8. Electron gun of a low-electrostatic focus tube.

quired a potentiometer-type control to vary the focus anode voltage for best focusing. Fig. 1-8 shows the electron gun of a low-electrostatic focus tube.

Several picture tubes having automatic electrostatic focusing were produced. In these tubes, an internal resistance connected the focusing electrode to the cathode of the electron gun. Since the focusing electrode is in or near the electron stream, it will acquire an electrostatic charge from the electrons which hit it. The amount of charge would depend on the beam current, and focusing would be automatic.

Beam Deflection

Now that we have produced and accelerated the beam and can manually control its intensity and focus, the beam must be given a horizontal and vertical movement within the area of the fluorescent screen. Two sets of deflecting plates with horizontal and vertical orientation (Fig. 1-9) are mounted in the neck of the tube. They are so arranged that the electron beam passes between the plates of each pair after it has sped through the anodes toward the screen. The complete assembly

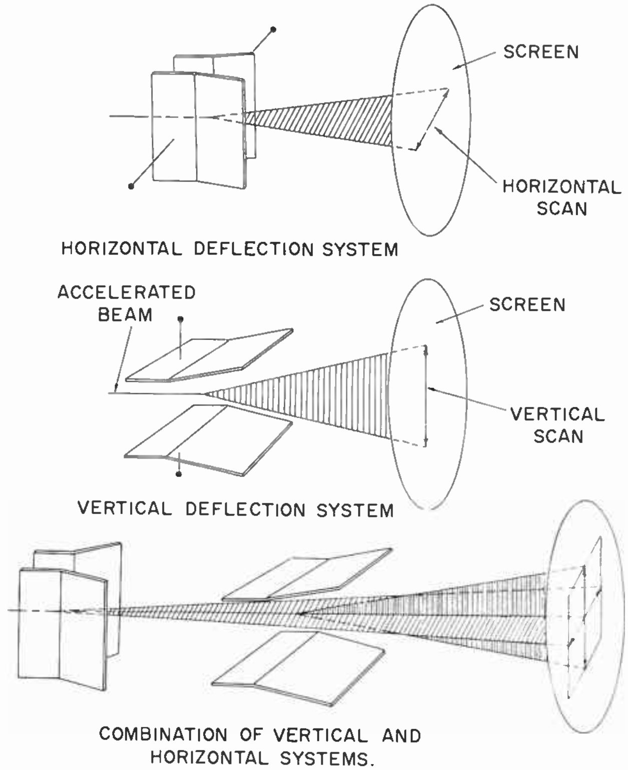


Fig. 1-9. Electrostatic beam deflection systems.

of a cathode-ray tube with electrostatic deflection is illustrated in Fig. 1-10.

Since the electrons in the beam are negatively charged, their movement is governed by the basic law of attraction and repulsion—"like charges repel one another, unlike charges attract one another." Therefore, a positively charged plate will attract electrons, and a negatively charged plate will repel them.

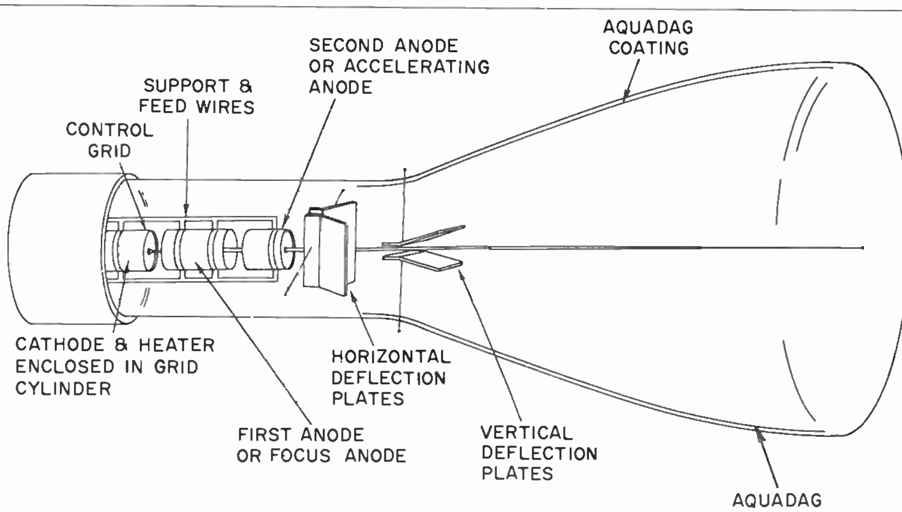


Fig. 1-10. Electrostatic focus and deflection.

An electrostatic field exists between two adjacent plates of opposite polarity. When an electron is shot into an electrostatic field whose lines of force cross its path (Fig. 1-9), the electron tends to drift from its normal course toward the positively charged plate. The electron actually crosses the lines of force because of its own momentum, since the static lines pull the electron in their direction. The high speed at which the electron beam passes through the static field delays its deflection slightly, thus preventing it from hitting the positive plate. Therefore, the amount the beam is deflected off its normal course depends on the velocity of the beam and the strength of the deflecting field. The horizontal and vertical deflections could be increased by increasing the distance between the point of deflection and the screen; this, of course, would increase the length of the cathode-ray tube. The distance the beam or spot is moved across the screen by an applied voltage of one volt across the deflection plates is called the deflection sensitivity. In some specifications, the sensitivity of the horizontal-deflection plates will be greater than that of the vertical plates since the horizontal plates are farther from the screen. However, in a cathode-ray tube, the velocity of the beam and the position of the electrodes are fixed; therefore, to increase the deflection, the deflecting voltage must be increased.

Another way of increasing the deflection sensitivity is by increasing the length of the deflection plates so that the static field is active on the beam for a longer time. In this case, the ends of the plates must be bent to form a flare (Fig. 1-9).

In early cathode-ray tubes, deflection voltage was obtained from single-ended amplifiers. One plate of each set of deflection plates was tied together and connected to the accelerating anode. When a voltage was applied to the deflection plates, a difference of potential existed between the accelerating

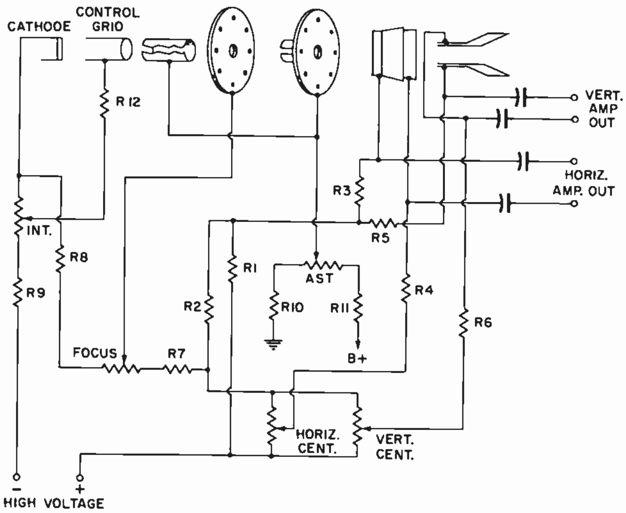


Fig. 1-11. Typical centering-control circuit.

anode and the deflection plates, causing defocus and a change in the beam velocity. This effect is called astigmatism.

In present-day cathode-ray tubes, a separate terminal is provided for each deflection plate, making possible the use of push-pull deflection amplifiers. The average potential remains constant between the plates of either pair because the increase in potential of one plate is equal to the decrease in potential of the other plate. Thus, any defocus or any change in beam velocity is minimized. Some tubes also have a ring or element between the horizontal- and vertical-deflection plates. This element is connected to the accelerating anode and prevents defocus due to any disturbing field between the pairs of deflecting plates.

The accelerating anode in modern cathode-ray equipment is connected to a variable voltage divider, usually called the astigmatism control. This control is adjusted for maximum roundness of the spot on the screen.

Beam Centering

To insure proper operation of the cathode-ray tube, an additional control is required. For correct operation, the cathode-ray electron beam should strike the center of the fluorescent screen when there is no deflection potential. Stray electric and/or magnetic fields, distortion of forces within the cathode-ray tube, and aging or replacement of the tube or its associated supply components may move the beam off center. To correct an off-center condition, two centering controls in most circuits enable the beam to be adjusted for proper horizontal and vertical positioning.

Fig. 1-11 illustrates a typical centering-control system. A high positive potential is applied across the divider network of R1, R2, R7, the focus control, and R8. From the junction of equal resistors R1 and R2, potentials are applied through R3 to one horizontal plate and through R5 to one vertical plate. As shown in Fig. 1-11, each centering control parallels the combination of R1 and R2. The variable arm of the horizontal centering control supplies the remaining horizontal-deflection plate through R4. Similarly, the variable arm of the vertical centering control supplies the remaining vertical-deflection plate through R6. When the variable arms of the centering controls are at their midpoint, no DC potential exists between the individual plates of each pair.

Consider for a moment the horizontal plates. A simplified diagram of their supply circuit is shown in Fig. 1-12. Since R1 equals R2 and XY equals XZ when the centering control is at midpoint, we have, in effect, a bridge circuit; no potential will exist between the arms of this bridge, represented by points A and B in Fig. 1-12. One horizontal plate is supplied by each bridge arm through equal series

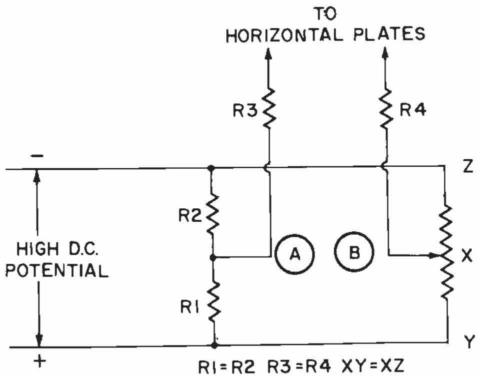


Fig. 1-12. Simplified horizontal-centering circuit.

resistors R3 and R4. Thus, a condition of no DC potential can be maintained between these plates. Should an off-center electron beam be encountered because of any one or a combination of the effects mentioned previously, the beam may be centered by moving the control arm to provide a counteracting DC potential. The vertical-centering circuit operates in a similar fashion.

A cathode-ray tube using electrostatic focus and magnetic deflection must have external control of beam centering. The only method that has been used is the application of external magnetic device. The action of the magnetic field on the electron beam is explained in Chapter 2.

The first magnetic device for centering an electrostatically-focused tube consisted of a small permanent magnet clamped to the neck of the tube. A knurled rod was attached to the magnet; the magnet could be rotated in the clamp, and the clamp could be rotated and slid back and forth on the neck of the tube. All of these motions affected the centering of the electron beam.

A more modern type of centering device consists of two magnetized rings that can be rotated separately or simultaneously around the neck of the tube until the electron beam is centered. This type of centering device is illustrated in Fig. 1-13.

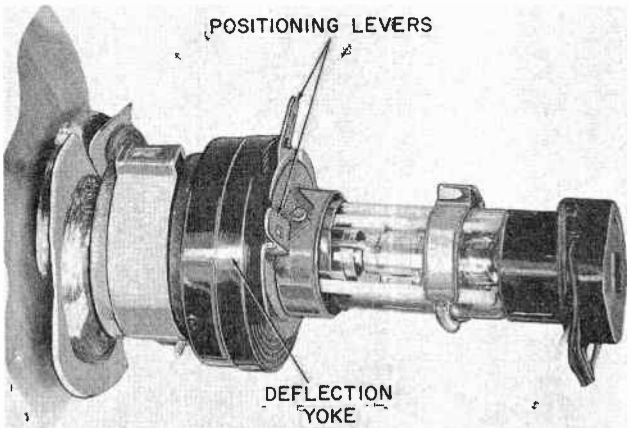


Fig. 1-13. Two magnetized rings used as centering device.

BEAM EFFECT ON THE FLUORESCENT SCREEN

The human eye can retain an image about 1/16 second after the image disappears. This phenomenon is used in motion pictures, where a series of still pictures is projected on the screen so rapidly, the eye does not detect them as separate pictures. In a cathodè-ray tube, the beam is swept so fast, the moving spot looks like a straight line. If the beam is swept over the same line or path at least 16 times a second, the spot looks to the viewer like a continuous line of light without flicker. Therefore, if the combined action of the horizontal- and vertical-deflection voltages sweeps the beam horizontally and vertically at the same time, a frame of light will appear on the screen (Fig. 1-9). In other words, a small spot of light appears at the point where the electron beam strikes the screen and then if the beam is deflected left to right and top to bottom very rapidly, the whole screen is illuminated. This frame of light, whose intensity can be controlled, is called a raster when used in television. Since this raster consists of small spots of light, a signal that will modulate the beam and cause each spot to vary in brilliancy can be inserted into the control-grid circuit of the cathode-ray tube. In this way, a picture is formed.

QUESTIONS

1. Name the main sections of the electron gun in an electrostatic tube.
2. What is the electron source in modern cathode-ray tubes?
3. What determines the intensity of light on the screen of the picture tube?
4. Which anode in the electrostatic tube is used to narrow the beam?
5. What term is used to refer to the distance the beam is moved across the screen by a voltage of one volt across the deflection plates?
6. The amount the beam is deflected off its normal course depends on what two factors?
7. What is the purpose of the centering controls?
8. What is the name of the frame of light that appears on a television cathode-ray tube when the whole screen is illuminated?

EXERCISES

1. Sketch the internal construction of the electrostatic cathode-ray tube and identify each element.
2. Give the function of each element in Exercise 1.

Chapter 2

Cathode-Ray Tube — Electromagnetic Control of the Beam

Up to this point, our discussion of beam formation and control has been primarily about cathode-ray tubes using electrostatic control methods. A second type of cathode-ray electron-beam control is obtained through electromagnetic focusing and deflection by varying the relative force, position, or area of fields adjacent to the beam.

The construction of an electromagnetically-controlled tube is shown in Fig. 2-1. The electron-gun assembly is quite similar to that of the electrostatic unit. Heater and cathode elements are enclosed in the grid cylinder, as they were before. The grid controls the number of electrons admitted to the beam in the same manner. The anode structure is different because there is no provision for internal focusing.

An accelerating anode immediately follows the control-grid cylinder, with its connection terminating in the tube base. Another cylindrical accelerating anode is often connected through contact springs to the Aquadag coating on the inner surface of the flared section of the tube. This Aquadag coating terminates in a button contact on the outside of the flared section.

To more easily understand the over-all operation of this tube, let us review the effect of magnetic fields upon an electron beam.

EFFECT OF MAGNETIC FIELDS UPON AN ELECTRON BEAM

The stream of electrons from the beam source may be considered as being equivalent to a stream of electrons in a solid conductor carrying direct current. The effect of an external magnetic field upon either stream will be the same since any flow of electrons produces its own magnetic field. The direc-

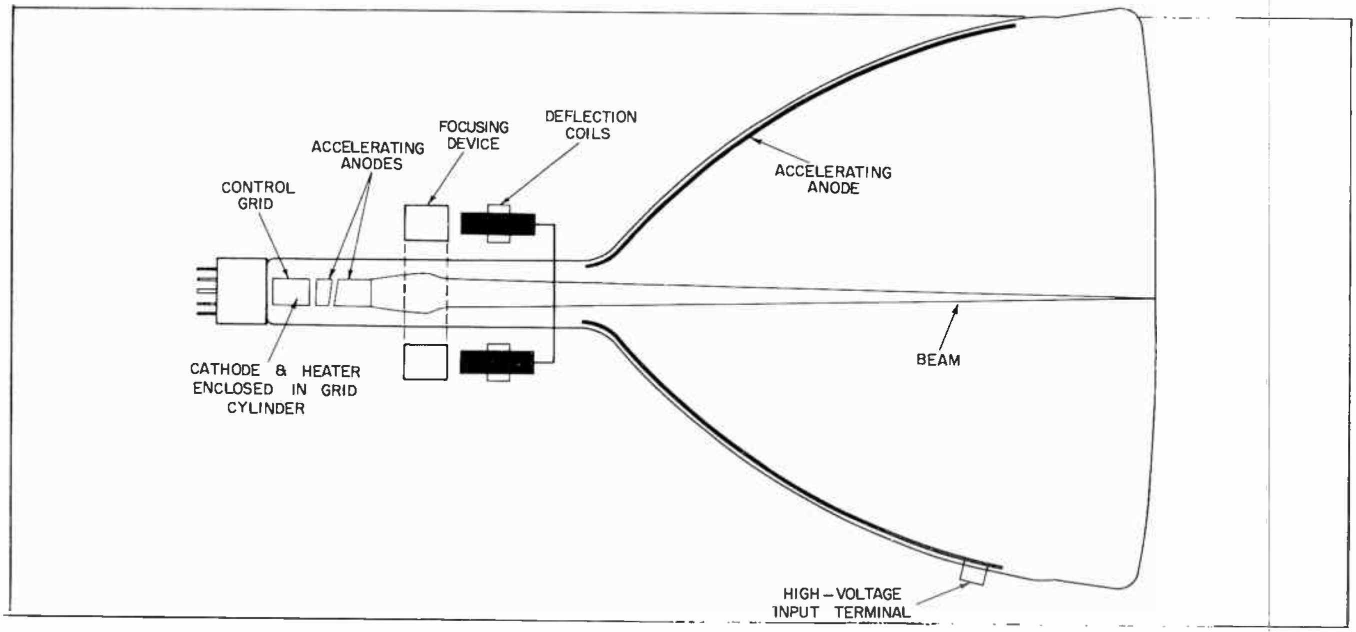


Fig. 2-1. Magnetically-controlled cathode-ray tube.

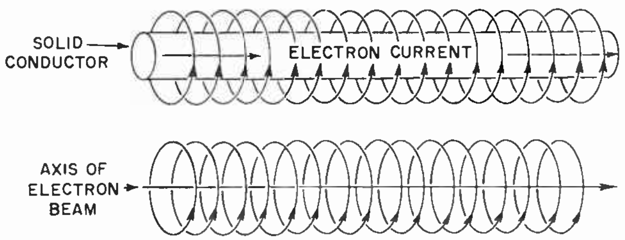


Fig. 2-2. Similarity of solid conductor and electron beam.

tion of the electron flow and the magnetic lines it produces are at right angles (Fig. 2-2).

If this current-carrying conductor is placed in an existing magnetic field with the conductor parallel to the lines of force of this field, no force will be exerted on the electron stream. The magnetic lines from the two sources are at right angles, neither aiding nor opposing one another; therefore, no interaction will result (Fig. 2-3).

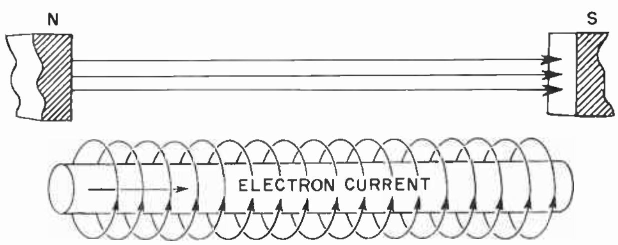


Fig. 2-3. Conductor parallel to magnetic field.

However, if the conductor is at right angles to the existing field (Fig. 2-4), a torque or distortion of the magnetic lines will tend to move the conductor from the field. This is because the lines of the two fields are opposing on one side of the current-carrying conductor or stream of electrons and aiding on the other side.

Fig. 2-4 shows the electron stream or conductor at a right angle to the external magnetic field. However, an electron stream entering the magnetic field at any angle other than parallel will also be affected by the external magnetic field. This effect will be proportionate to the amount of angular variation.

Let's apply the foregoing in terms of electromagnetic control of the beam in a cathode-ray tube.

BEAM FOCUSING

Note in Fig. 2-1 that the focusing device is placed along the neck of the tube. Since its magnetic field controls the size of the electron beam and causes the formation of a small spot of light on the tube face, no internal focusing elements are required.

Provision is made for moving the focus coil along

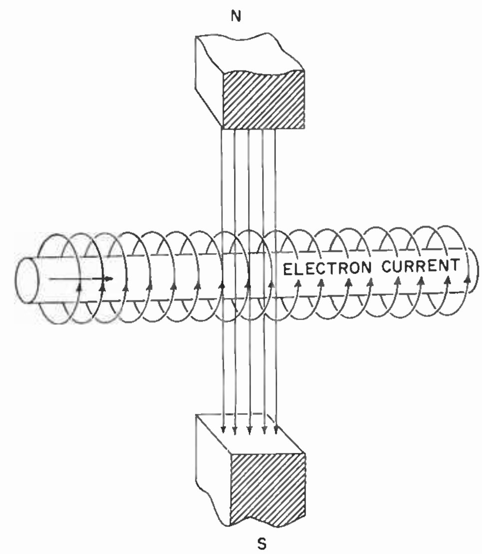


Fig. 2-4. Conductor at right angles to magnetic field.

the neck of the tube. After the coil is in approximately the correct position, control of fine focus is obtained by varying the direct current through the coil. This control (called the focus control) is normally a potentiometer. The construction and application of a typical focus coil is shown in Fig. 2-5.

The first focusing devices were electromagnets. These electromagnets, wound with many turns of wire, were so placed on the neck of the cathode-ray tube that the magnetic field was concentrated about the neck of the tube. Thus, the beam of electrons

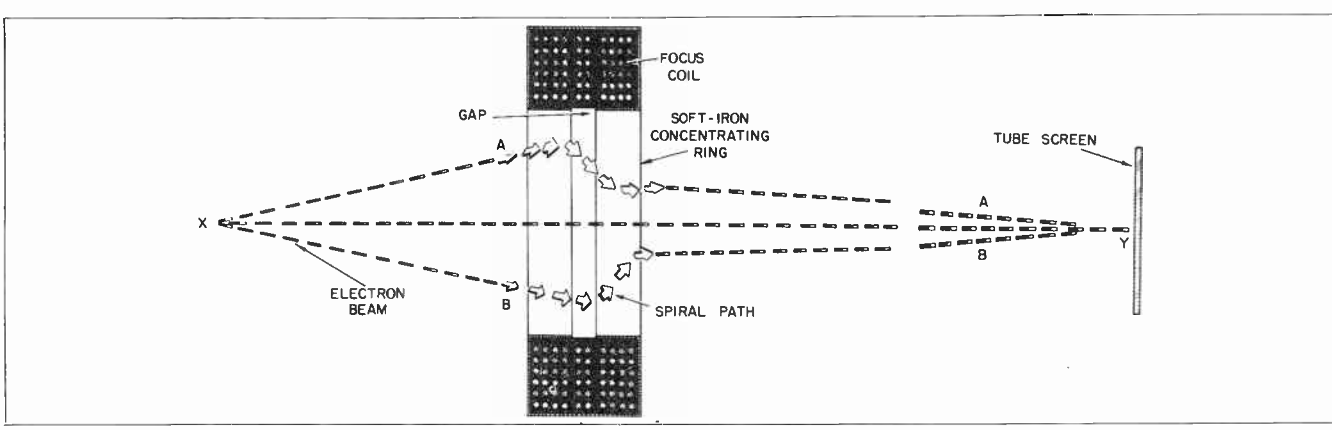


Fig. 2-5. Action of focus coil on electron beam.

inside the tube was surrounded with parallel lines of magnetic force. By concentrating these lines of force, two results are obtained: (1) Less magnetic strength is necessary than with other structures. (2) Stray fields are lessened; therefore, they are less likely to affect the action of the other beam-control elements.

To generate a magnetic field, the focus coil needs a supply of well-filtered direct current. For this reason, the focus coil has generally been replaced by the permanent-magnet (PM) focusing device (Fig. 2-6). Crude focusing is obtained by proper

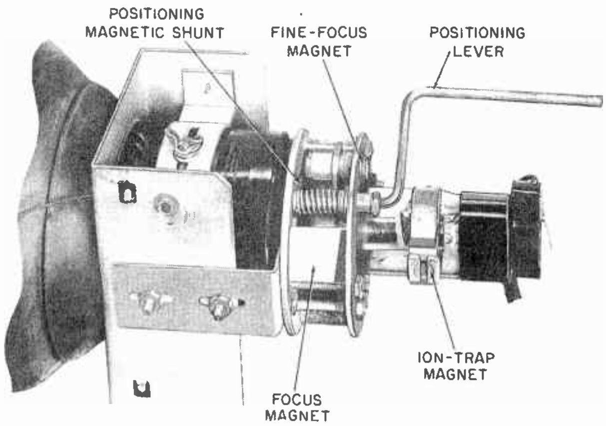


Fig. 2-6. Permanent-magnet focus device.

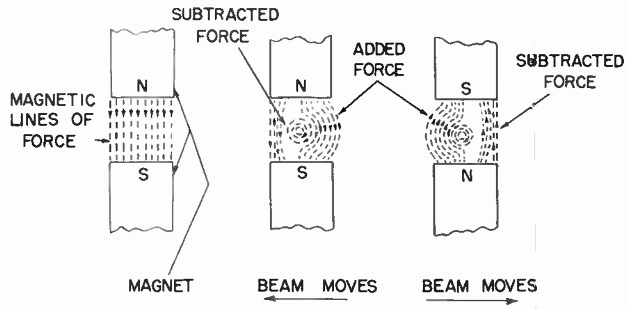
placement of this device along the tube neck. For fine focusing, a fine-focus screw moves a soft-iron magnetic shunt back and forth and thus varies the magnetic strength. The PM focus device acts the same as the focus coil pictured in Fig. 2-5.

Referring to Fig. 2-3, we can see that any electrons traveling along a path parallel to the lines of an external magnetic field will not be affected by that field; they will continue to travel in a straight line. Axis XY of Fig. 2-5 shows this path.

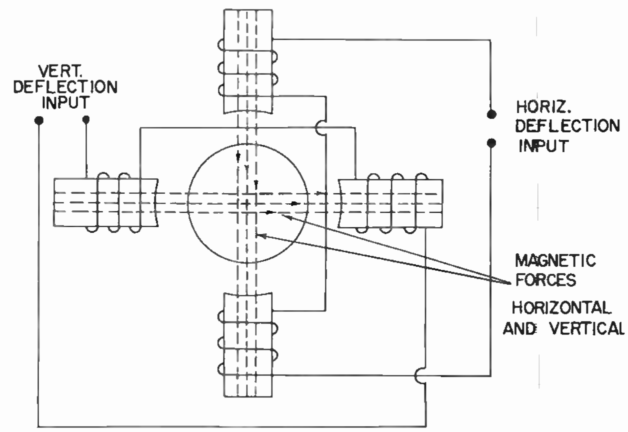
Referring to Fig. 2-4, we can see that a stream of electrons entering an external magnetic field at right angles will be deflected out of the field. Any time this stream of electrons enters the field at an angle other than parallel, the beam will be deflected slightly. Lines A and B of Fig. 2-5 represent the paths of electrons not parallel to axis XY. Therefore, since this beam is entering the magnetic field of the focus coil at an angle, the stream will be pushed sideways. Because the electrons are all traveling very rapidly and each electron has its own magnetic field, the beam within the focus field will follow a path similar to the thread of a wood screw. With a proper balance of the beam velocity, the magnetic field produced by the focusing device, and the potential applied to the accelerating anode, the electron beam will leave the focus field in a converging stream having its focal point at the fluorescent screen.

BEAM DEFLECTION

A magnetic field will deflect a beam of fast-moving electrons at right angles to the direction of the field and the electron motion. (See Fig. 2-7A.)



(A) Movement of electron beam in deflection field.



(B) Arrangement of magnetic coils in deflection yoke.

Fig. 2-7. Action and position of deflection coils.

Therefore, electromagnetic deflection of the beam may be obtained by two sets of coils. These coils are arranged horizontally and vertically over the neck of the tube, near the bulge of the bulb and the focusing device (Fig. 2-1).

The two sets of coils are mounted in what is called a deflection yoke and have four windings, two for horizontal deflection and two for vertical deflection. Fig. 2-8 shows coil construction and assembly of a complete deflection yoke. The two horizontal coils are opposite each other and connected in series for correct polarity. Thus, the magnetic field passes through the neck of the tube at right angles to the path of the beam and will be oriented vertically for horizontal deflection. The vertical-deflection coils are arranged and connected in the same manner, but oriented horizontally for vertical deflection (Fig. 2-7B).

Thus, the spot produced by the beam can be moved anywhere on the screen by passing the correct amount of current through each set of coils. To prevent interaction, good magnetic shielding must be provided between the focus and deflection coils.

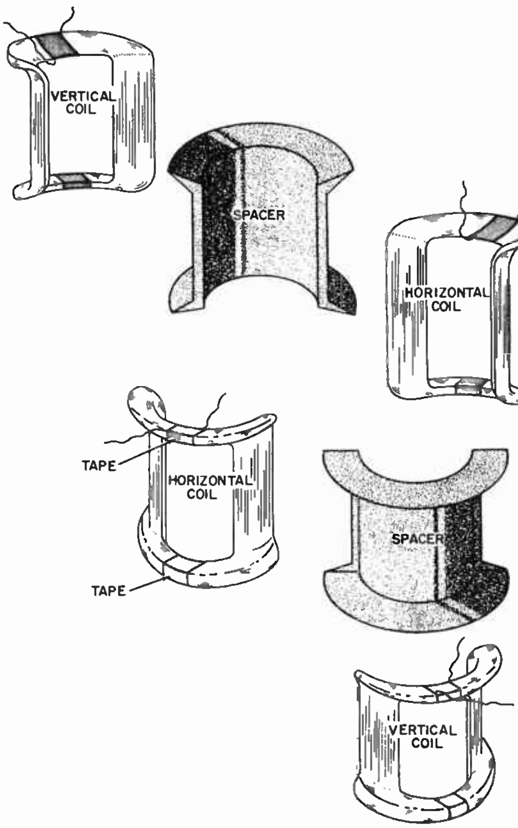
If a sawtooth of current is passed through the horizontal coils, it will cause the spot to move from left to right across the screen and then fly back. Similarly, if a sawtooth of current is passed through the vertical coils, the beam will be made to move from top to bottom and fly back. The combined action of the horizontal and vertical fields will produce a frame of light, or raster, similar to that in electrostatic deflection.

A complete discussion of the development of these sawtooth waveforms will be given in the section covering horizontal and vertical oscillators.

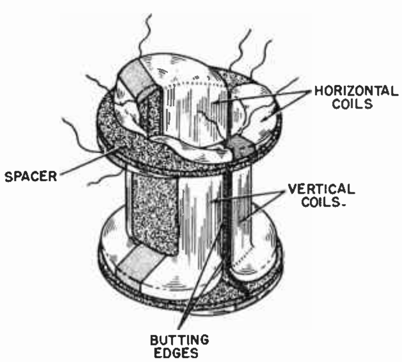
BEAM CENTERING

The beam in an electromagnetically-controlled cathode ray tube can be centered in several ways. One of the early methods was a potentiometer which varied the amount of direct current flowing through the deflection-yoke windings. A direct current provides a fixed magnetic bias for positioning the beam at a point which becomes the center of deflection after deflection voltages are applied to the yoke.

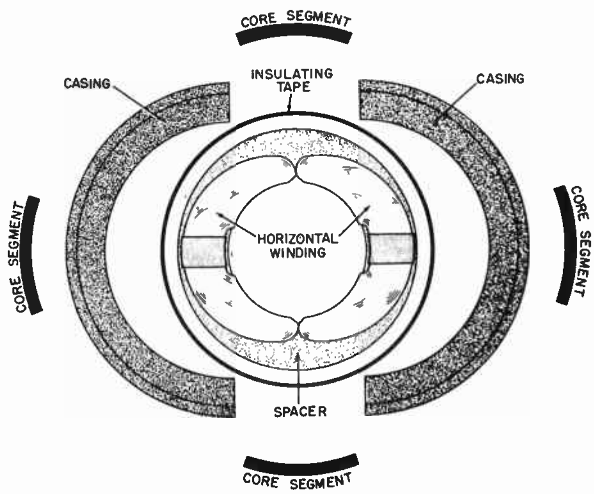
Centering-control potentiometers were soon abandoned in favor of a method that allowed circuit simplification. Because the magnetic field of a focusing device exerts a magnetic bias on the electron beam, the position as well as the focus of the beam can be controlled. The mounting of the focusing device was made variable for variable bias. Eventually, mechanical adjustment of the positioning of the focusing device was improved by adjustment screws at each corner. Each screw has a spring backing to assure that its adjustment cannot change accidentally. Either horizontal or



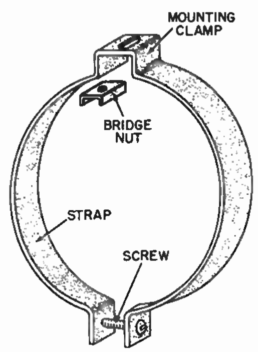
(A) Windings and spacers.



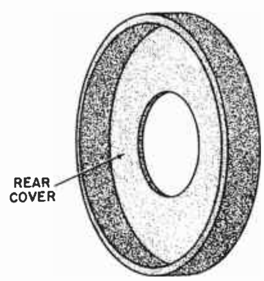
(B) Windings and spacers assembled.



(C) Overhead view of windings and spacers arrangement.

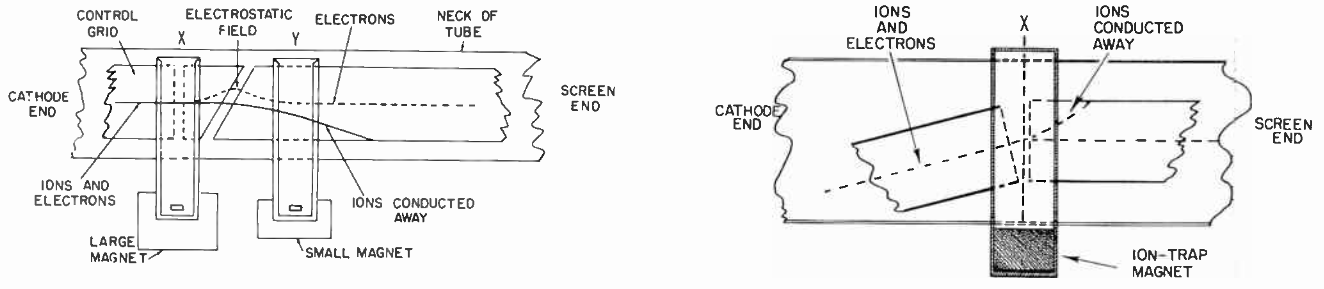


(D) Metal strap which holds core segments in place.



(E) Rear cover.

Fig. 2-8. Deflection yoke assembly.



(A) Double ion trap assembly.

(B) Single ion trap assembly.

Fig. 2-9. Action of ion trap.

vertical centering corrections can be made with these purely mechanical adjustments.

For further simplification, the focusing-device mounting may be stationary and a movable magnetic shunt added. This shunt concentrates the magnetic field, and as the shunt is moved, it centers the electron beam. The arm or handle by which the shunt is moved is visible in Fig. 2-6. The electron beam is always moved at a right angle to the direction of the shunt.

REMOVAL OF IONS FROM ELECTRON BEAM

The emitted electrons from the cathode are mixed with charged atomic particles called ions. These ions are in the tube for two reasons: (1) no matter how well the elements making up the internal tube structure are cleaned, a slight amount of foreign material will be present, and (2) as the cathode is heated, small particles will tend to break loose from it. Each ion is approximately 2,000 times heavier than the electron. If ions strike the screen, a brown spot will gradually appear because they remove the phosphor. The net result is a spot on the screen where no picture image can be produced.

Electrostatically-deflected tubes are not affected by the presence of ions since the electron beam and the ions are deflected simultaneously. However, magnetic fields do not greatly deflect an ion, and tubes using magnetic deflection systems must have some provision for preventing the ions from reaching the center of the phosphor screen; otherwise, a spot will result.

Two methods are used to remove the ions. The first method takes advantage of the fact that ions are not affected by magnetism. The entire beam, containing both ions and electrons, is deflected electrostatically within the gun assembly. Magnets are used to straighten the electrom beam while allowing the ions to continue on their bent path until they hit the accelerating anode and are removed. Fig. 2-9A represents a typical double ion-trap assembly.

The field from the ion-trap magnet bends the electron beam at points X and Y. The ion-trap assembly is adjustable to provide a means of bending the beam at these two points. For correct adjustment, the raster is used as an indicator. The entire assembly is adjusted for maximum brilliance and good horizontal-line focus. Thus, only electrons can emit from the gun structure. These electrons then pass through the neck of the tube and are focused, deflected, and accelerated. Either electromagnets or permanent magnets are used in double ion traps.

The double ion traps were eliminated when picture-tube manufacturers began manufacturing the bent-gun tube. In this type of tube, part of the electron gun is at an angle to the rest of the gun, as shown in Fig. 2-9B. As the ions and electrons emerge from the first part of the gun, they would normally be eliminated by striking the other part of the gun. A single magnet at point X in Fig. 2-9B, however, bends the electrons back to the center of the second part of the gun, and they continue on to strike the screen. A single-magnet ion trap of the type used with a bent-gun tube can be seen in Fig. 2-6.

Another method of removing ions uses an extremely thin film of aluminum on the beam side of the phosphor screen. The aluminum is thin enough that the electrons can pass through and strike the phosphor. Since the ions are larger, they will not penetrate the aluminum and, therefore, will not strike the phosphor. The aluminum film also provides better contrast and more brilliance of the picture. Most picture tubes with the aluminum film employ a straight gun and do not require an ion trap. However, some tubes with an aluminum film do employ a bent gun, and a single-magnet ion trap is required.

The general types of picture tubes and their requirements for deflection have been covered. Before discussing the method of translating received signals into visual patterns on the face of the picture tube, let us see how the televised scene is converted into a video signal at the transmitting studio.

QUESTIONS

1. Is the direction of the magnetic lines produced by electron flow *parallel* or at a *right angle* to the direction of the electron flow?
2. Under what condition will no force be exerted on an electron stream by a surrounding magnetic field?
3. Which one of the following is not used in electromagnetically-focused tubes?
   (1) Accelerating anode.
   (3) Cathode.
   (3) Focusing anode.
   (4) Control grid.
4. How is fine focusing obtained when a PM focus device is employed?
5. What is used for deflecting the beam in an electromagnetic tube?
6. What type of current is passed through the yoke?
7. Which one of the following methods or devices is not used for centering the beam?
   (1) Positioning of the focusing device.
   (2) The yoke.
   (3) A potentiometer.
   (4) A magnetic shunt.
8. What portion of the beam must be removed before the beam strikes the phosphor screen? Why?

EXERCISES

1. Sketch the electromagnetically-controlled tube and show its internal elements and external components. Identify each element and component.
2. Give the function of each element and component shown in Exercise 1.

Chapter 3

The Camera Tube

The following types of camera tubes will be discussed: Iconoscope, Image Orthicon, Image Dissector, Monoscope, and Vidicon. A brief discussion of the internal construction and operation of these camera tubes will assist the reader in associating the scanning technique and transmitted picture with those of the receiving picture tube.

ICONOSCOPE

The Iconoscope (Fig. 3-1) is basically a cathode-ray tube which translates the scene to be transmitted into electrical impulses. It consists of an electron gun similar to the one in the receiving tube; but instead of a fluorescent screen, a large rectangular plate of thin mica is used as a scanning area.

On the front (scene side) of this mica sheet are many microscopic particles of cesium-silver compound—a photosensitive material. Each particle is insulated from the other, giving the mica sheet a mosaic appearance. The back of the mica sheet is covered with a conductive film connected to an output lead. The whole arrangement appears as myriads of small capacitors or cells, with a common lead through which their stored energy is discharged.

To understand the action of the scanning beam, let us assume no scene is being projected optically on the mica sheet, or mosaic. As the beam strikes the small particles of cesium silver, secondary emission takes place. The number of secondary electrons emitted is several times greater than the primary electrons in the beam which strike the particle. Since more electrons, which are of negative potential, are emitted than the number striking the particle, the potential of the particle will change in a positive direction. It will rise to an equilibrium potential of approximately +3 volts. The secondary electrons emitted go either to the collector or to other parts of the mosaic. Since each particle is insulated from all others, this

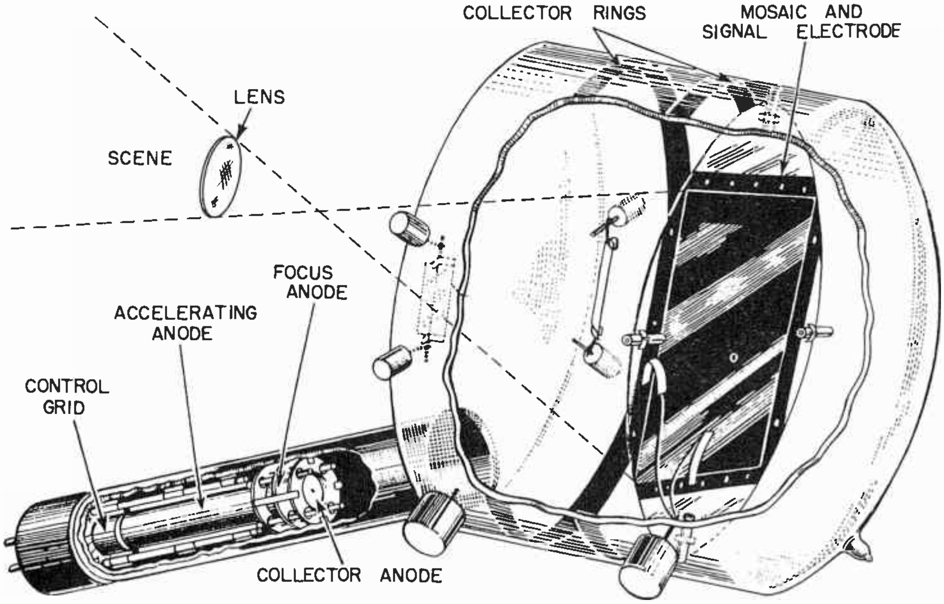
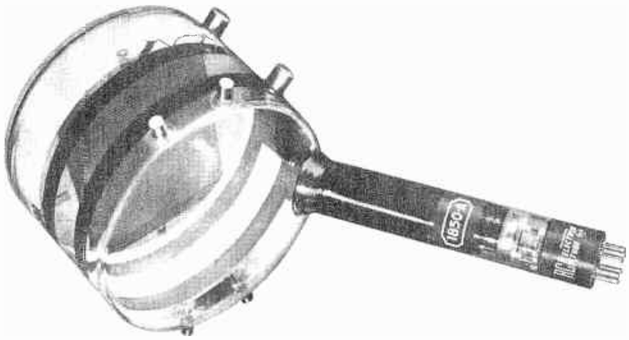
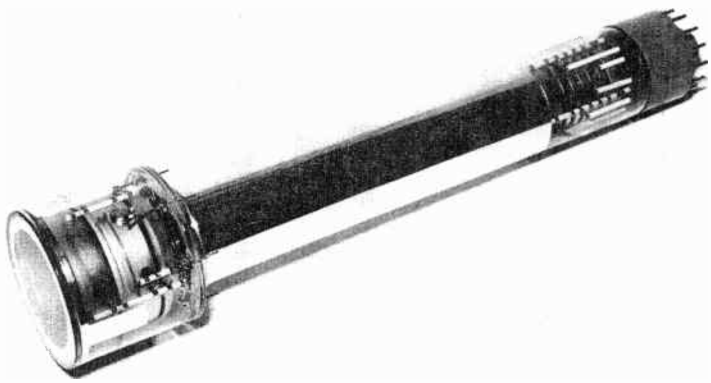


Fig. 3-1. RCA Type 1850-A Iconoscope.



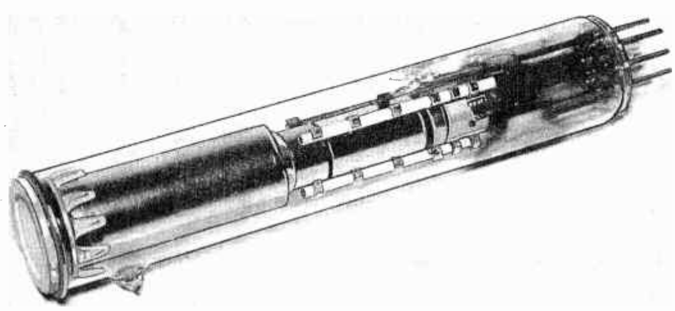
Iconoscope

Image Orthicon



Monoscope

Vidicon



Typical Types of Camera Tubes

charge cannot leak off. However, after the beam continues its sweep, the positively charged particle will attract secondary electrons emitted from other particles in the mosaic and will then go negative. Because so many free electrons are on the face of the mosaic, the particle will actually charge to approximately −1.5 volts. This action closely parallels the familiar contact method of obtaining bias in audio amplifiers. The proximity of the grid to the cathode places it in a cloud of electrons which causes current flow in the grid circuit. However, there can be no current flow in the particle in the mosaic. The particle takes on a negative charge until it is again struck by the beam. Each particle changes from a −1.5 volts to a +3 volts each time the beam strikes it.

The Iconoscope output is obtained from a resistive load connected between the conductive film on the back of the mosaic and ground. A certain capacitance exists between each particle and the conductive film. The instant the beam strikes the particle, the charge on this capacitance cannot change. Hence, the entire voltage appears across the resistive load. A number of electrons equal to the amount lost by the particle will flow from ground to the conductive film to maintain the charge on the existing capacity. This current flow results in a 4.5-volt potential across the load. Because no scene has been projected on the mosaic, the potential on each particle will change equally. As a result, the amount of current flow in the load as the beam scans the mosaic does not change. Since there is no AC component, there is no output from the Iconoscope.

We have discussed the action of the tube with no image projected on the mosaic. Let us now assume that half of the mosaic is illuminated. The cesium-silver particles are photosensitive and will emit electrons when struck by light. When the beam of electrons has passed over a particle being struck by light, the particle will attract free electrons. Since some electrons are being emitted because of the photosensitive properties of the compound, the particle will not take on a −1.5-volt charge. Instead, it will assume some charge in a positive direction from the −1.5 volts. The amount, of course, depends upon the amount of light present. For illustration, let us assume the intensity of light on the illuminated half of the mosaic allows the illuminated particles to charge to a −1 volt. The particles in the nonilluminated area will charge to a −1.5 volts. As the beam of electrons from the electron gun strikes the nonilluminated particles, a change of 4.5 volts takes place, resulting in a 4.5-volt potential across the load. When the beam strikes the illuminated particles, each particle will change only 4 volts while charging to the equilibrium potential of +3 volts, since the original charge was only −1 volt. This results in a potential of 4 volts across the load giving an AC component in the output.

When an image is projected on the mosaic, each particle will charge to a certain potential, depending on the amount of light present. As the beam scans the mosaic from left to right and from top to bottom, each particle will be returned to the equilibrium potential, and current pulses will flow in the load resistor. This train of pulses, varying with the charge on the particles, constitutes the video signal. The output from the Iconoscope is of negative polarity since there is less current flow when an illuminated particle is scanned than when a nonilluminated particle is scanned.

IMAGE ORTHICON

The Image Orthicon is intended for use in black-and-white cameras for outdoor and studio pickup. It is very stable in performance at all incident light levels. For a better understanding of the

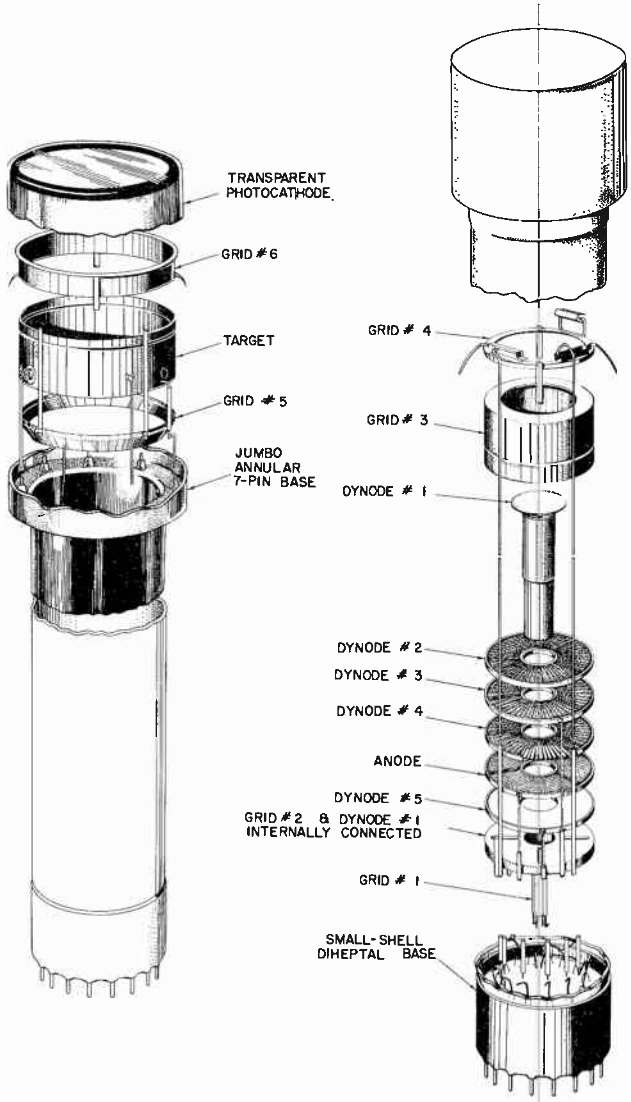


Fig. 3-2. RCA Type 2P23 Image Orthicon.

operation of the Image Orthicon, refer to Fig. 3-2 while studying the following paragraphs.

Light from the scene being televised is focused on the semitransparent photocathode. This photocathode emits electrons proportional to the amount of light striking the area. These electrons are accelerated toward the target by grid 6 and focused by the magnetic field produced by an external coil. The target consists of a special thin glass disc with a fine mesh screen on the photocathode side. Focusing is also accomplished by varying the potential of the photocathode.

When the electrons strike the target, secondary emission from the glass takes place. These secondary electrons are collected by the wire mesh, which is maintained at a constant potential of approximately one volt. This limits the potential of the glass disc and accounts for its stability in changing intensities of light. As electrons are emitted from the photocathode side of the glass disc, positive charges are built up on the other side. These charges vary with the amount of electrons emitted. Thus, a pattern of positive charges corresponding to the intensities of light of the scene being televised is set up. This constitutes the image section of the Image Orthicon. The action described is completely independent from the electron beam and scanning circuits of the tube.

The backside of the target is scanned with a low-velocity beam from the electron gun. The beam is focused by the magnetic field generated by an external coil and by the electrostatic field of grid 4. The potential applied to grid 5 adjusts the decelerating field between grid 4 and the target. As the low-velocity beam strikes the target, it is turned back and focused on dynode 1, the first element of an electron multiplier. When the beam is turned back from the target, however, some electrons are taken from the beam to neutralize the charge on the glass. The greater the charge on the glass, the more electrons are taken from the beam. Thus, when the beam scans a more positively charged area, corresponding to a brighted area in light intensity, fewer electrons are returned to dynode 1. This action leaves the scanned side of the target negatively charged while the opposite side is positively charged. Because the glass-disc target is extremely thin, these charges neutralize themselves by conduction through the glass. This neutralization takes place in less than the time of one frame.

As the amplitude-modulated stream of electrons strikes dynode 1, secondary electrons are emitted. Several secondary electrons are emitted for each primary electron striking the element. These free electrons are then accelerated toward dynode 2. As the electrons strike the element, more secondary emission takes place at dynode 2. This same process continues on through dynodes 3, 4, and 5. The electrons are finally collected by the anode. Thus, the electrons returned to dynode 1 are amplified, or multiplied, many times before the signal reaches the anode. The multiplication per element equals the difference between secondary electrons emitted and electrons striking the element. The approximate gain of the multiplier section of this tube is 500. The load resistor of the Image Orthicon is connected from the anode to the power supply. More current flow in the multiplier, which corresponds to a dark area in the televised scene, causes more current flow in the load, giving a negative output. A brighter area causes less current flow, giving a positive output. Thus, the output of the Image Orthicon will be of positive polarity.

IMAGE DISSECTOR

Both camera tubes previously discussed are known as the storage type since their operation depends upon the neutralization of positive charges by the scanning beam. The Image Dissector (Fig. 3-3), on the other hand, employs instantaneous scanning.

The tube consists of an evacuated glass cylinder closed at both ends. The elements within the tube

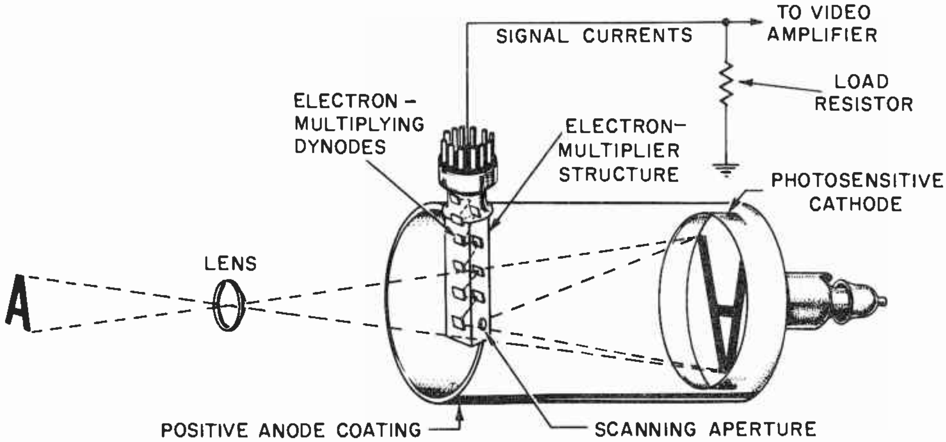


Fig. 3-3. Image Dissector operation.

are a photosensitive cathode, an anode, a shielded target having a small aperture, and an electron multiplier. The cathode, upon which a cesium-silver oxide film has been formed, is placed at one end of the cylinder. The anode, which accelerates the electrons emitted from the photocathode, is a conductive coating on the inner surface of the cylinder. The target is near the other end of the cylinder, which is a plane glass end. The target is at the end of an electron multiplier, which is used for amplification. In front of the target is a small aperture. This aperture lets only a small portion of the electron image fall on the target.

The entire cylinder is placed in a focusing coil. The coil produces an axial magnetic field throughout the entire length of the cylinder. The horizontal- and vertical-deflection coils are placed around the cylinder and act as a supporting framework also. The scene to be televised is focused on the photocathode. Electrons are emitted from this cathode according to the amount of light striking the particular area. It can be said that an "electron image" is emitted from the cathode. This image corresponds to the optical image projected on the cathode. The "electron image" is then accelerated toward the target by the anode, which has a positive potential of several hundred volts. The image is maintained in focus by the axial magnetic field of the focusing coil.

The "electron image" is deflected horizontally and vertically by the magnetic field set up by the sawtooth current flow in the deflection coils. As the "electron image" is deflected past the aperture, only a small portion of the image can strike the target. The image, however, is swept past the aperture in a series of 525 interlaced lines thirty times per second. Instead of a beam scanning the image, the entire image is scanned past the aperture, which "dissects" the image—thus, the name, Image Dissector.

MONOSCOPE

Another cathode-ray tube used in the development of television signals is the Monoscope, which is used for testing and adjusting studio equipment. When transmitted by the station, the signal is useful for proper adjustment of receiving equipment. The primary difference between this tube and the other camera tubes discussed previously is that it has a test pattern in front of the tube envelope. This test pattern is reproduced as the video signal.

The difference in amount of secondary emission of electrons between two materials is used to produce the output. Usually, a sheet of aluminum, which has high emission, is marked with high-carbon content ink. Carbon has fairly low emission, and as the electron beam scans the entire pattern, secondary electrons are emitted from both materials in proportion to their emission ratios. Any pattern with any line shape may be drawn on the aluminum sheet. The Monoscope is a stable video signal source and provides both the television engineer and the service technician with a useful tool.

VIDICON

The Vidicon is a small camera tube suitable for use in black-and-white and color-television cameras for either broadcasting or closed-circuit applications.

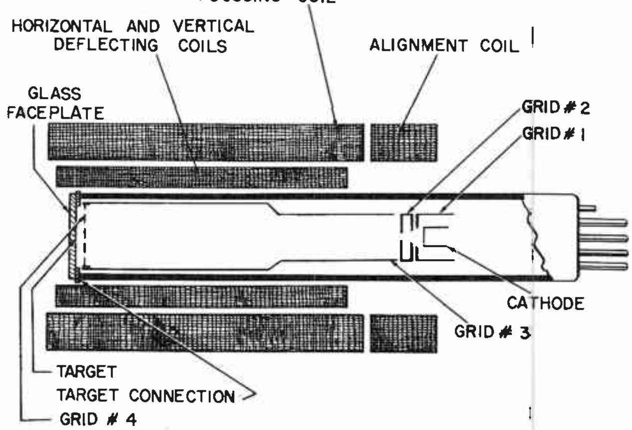


Fig. 3-4. Structural arrangement of the Vidicon.

The structural arrangement of the Vidicon, shown in Fig. 3-4, consists of a target, a fine mesh screen (grid 4), a beam-focusing electrode (grid 3), and an electron gun. The target is composed of a transparent conducting film on the inner surface of the faceplate. Grid 4 is adjacent to the photoconductive layer, and grid 3 is connected to grid 4. Each small portion of the photoconductive layer is an insulator when there is no light on the faceplate, but becomes slightly conductive when illuminated.

The gun side of the photoconductive layer is scanned by a low-velocity beam produced by the electron gun. The gun consists of a cathode, a control grid (grid 1), and an accelerating grid (grid 2). The beam is focused by the magnetic field of an external coil and by the electrostatic field of grid 3. Grid 4 provides a uniform decelerating field between itself and the photoconductive layer so that the beam will approach the layer perpendicular to it, a condition necessary for linear scanning.

When the gun side of the photoconductive layer with its positive-potential pattern is scanned by the beam, electrons are deposited from the beam until the surface potential is reduced to that of the cathode. Thereafter, the electrons are turned back to form a return beam, which is not used. Deposits on the scanned surface of any portion of the layer change the difference of potential between the two surfaces of the portion. When the two surfaces of the portion, which in effect is a charged capacitor, are connected through the external target circuit and the scanning beam, a capacitive current is produced. This capacitive current constitutes the video signal.

SCANNING

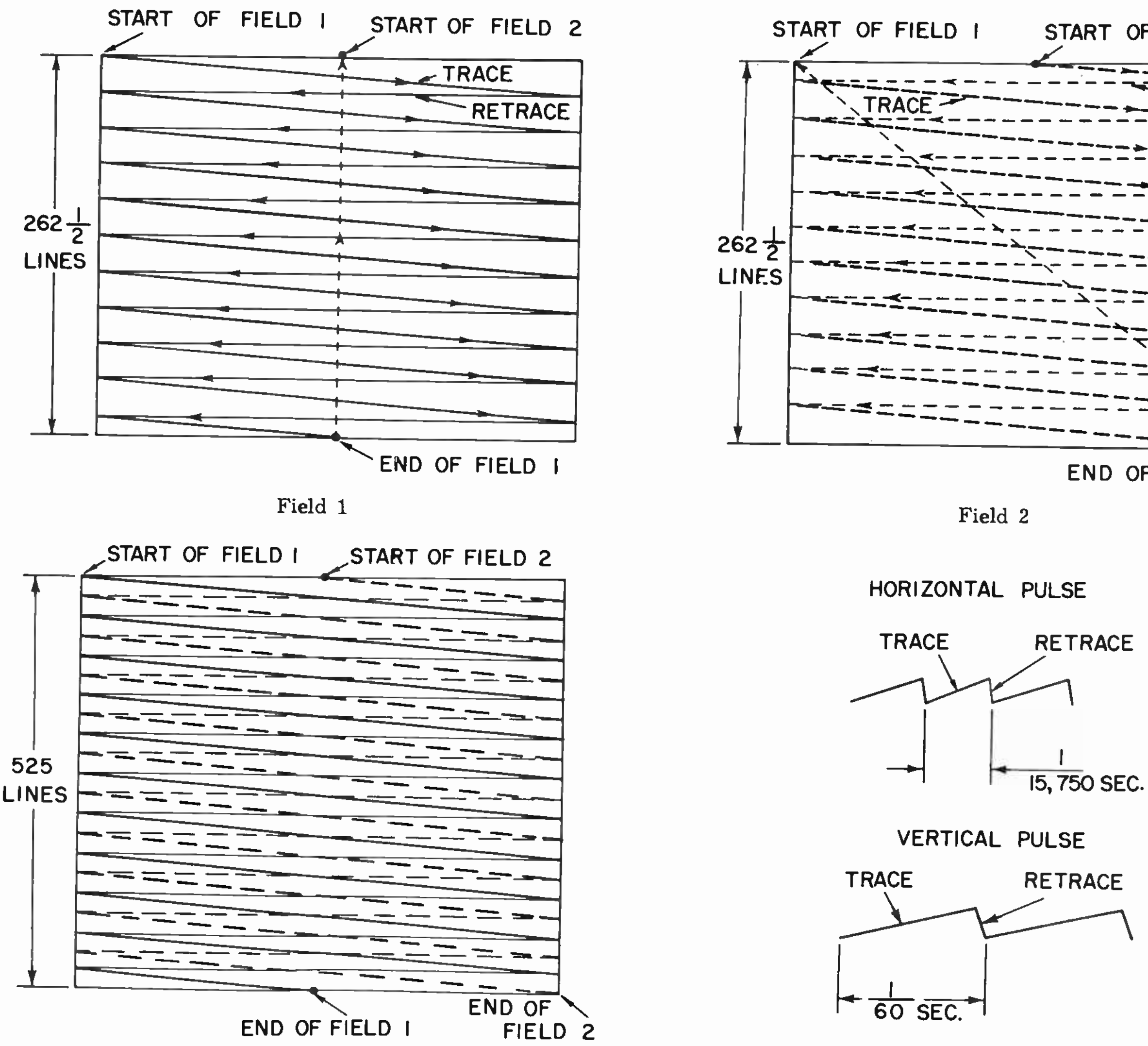
All of the picture-generating tubes discussed have associated external focus and deflection elements. In all but the Image Dissector, these elements cause the electron beam to scan the active picture surface at the front of the tube.

The video signal carries the picture information to be transmitted over the air. Since the timing of the scanning process is important, the video signal must contain other information in the form of electrical pulses. One of these is termed a blanking pulse and blanks out the return trace of the cathode-ray beam in the camera tube during flyback time. Other pulses, the synchronizing pulses, are used by the receiver to synchronize the horizontal and vertical sawtooth generators.

The path traveled by the beam across the screen of the picture tube should be identical to the path traveled by the beam in the camera tube so that the picture may be reconstructed in the correct sequence at the receiver.

For picture resolution, the U.S. standards for television broadcasts are 30 frames per second, each frame being constructed of 525 horizontal lines using interlaced scanning. (If alternate lines are so transmitted that two series of lines are necessary to produce a complete frame, the system is called interlacing.)

Therefore, to produce one frame of 525 lines interlaced, 262½ horizontal lines are scanned on the first down sweep of vertical deflection, and the beam returns to the top and scans 262½ al-



Field 1

Field 2

Complete frame.

Sawtooth deflection currents.

Fig. 3-5. The principles of interlaced scanning.

ternate lines. The horizontal and vertical scanning traces are the result of passing current with a sawtooth form through the respective deflection coils. The rapid return of the electron beam, or retrace, for the start of the succeeding scanning function is a result of the rapidly decreasing current in this sawtooth waveform. To produce interlaced scanning with 525 lines and 30 complete frames per second, the vertical-sweep frequency must be 60 cycles per second, and the horizontal-sweep frequency must be 15,750 cycles per second. Fig. 3-5 further explains the complete scanning operation.

QUESTIONS

1. What is the purpose of the camera tube?
2. When an image is projected on the mosaic of an Iconoscope, the potential to which each particle will charge depends upon what condition?
3. In an Image Orthicon, what is the scene focused on?
4. Which one of the following tubes is used to reproduce a test pattern for testing and adjusting studio and receiving equipment?
   (1) Vidicon.
   (2) Image Dissector.
   (3) Monoscope.
   (4) Image Orthicon.
5. How many frames per second and how many horizontal lines per frame make up the television picture?
6. What are the vertical- and horizontal-sweep frequencies?

Chapter 4

Power Supplies

Power supplies in television systems are more complicated than those normally found in radio and other electronic devices. At least two separate and basically different load conditions must be provided for in the set.

1. A low-voltage, high-current system to power the oscillator, amplifier, and similar stages where applied potential does not exceed 450 volts. In addition, some receivers power the modulation and deflection systems from this source.

2. A high-voltage, low-current system to supply the accelerating anode potential for the cathode-ray or picture tube.

LOW-VOLTAGE, HIGH-CURRENT SUPPLIES

The signal reception portion of the receiver, which includes sound and video amplifier or control tubes, presents a power requirement not greatly different in voltage range from that of other electronic devices. Therefore, this portion of the television power supply is similar to the ones found in large radio receivers.

In general, the voltage requirement is no more than 450 volts. The current required, however, is frequently much greater than that necessary for radio operation. In addition, a good supply regulation is needed to operate the sawtooth oscillators for deflecting the electron beam of the cathode-ray tubes. These oscillators tend to produce currents in the power supply. If not properly filtered, these currents would appear as serious modulation hum in the beam control and sound circuits.

Early receivers required heavy-duty power supplies to provide the current (as much as 300 milli-

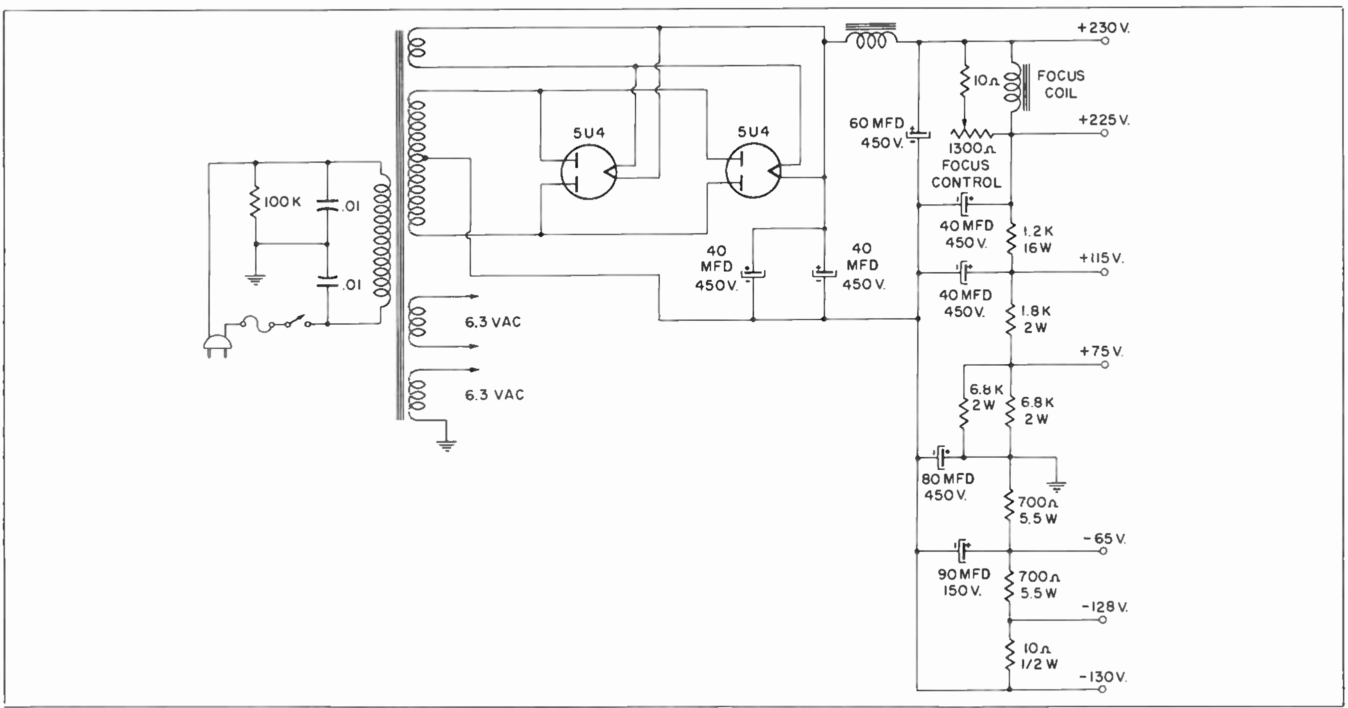


Fig. 4-1A. Schematic of typical low-voltage power supply in early receivers.

amperes in some receivers) drawn by the 30 to 40 tubes in the receiver. These supplies employed from one to three rectifiers; the number, of course, depended upon the current requirements and the designer's preference as to voltage and current distribution.

Fig. 4-1A shows a schematic of a power supply employing two 5U4G rectifier tubes. This power supply is typical of those found in the very earliest receivers. Different voltages—both positive and negative—were needed by the other stages in the receiver, and an elaborate voltage-divider or distribution network was needed to produce these voltages. The current drawn by several of the stages was passed through the focus coil and, thus, made to do double duty.

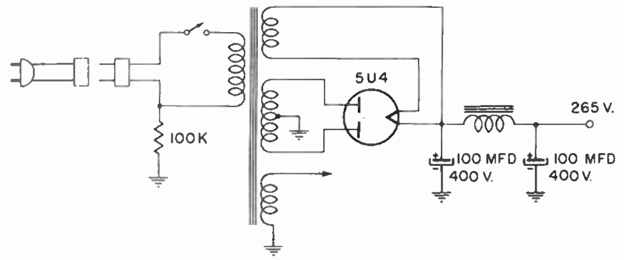


Fig. 4-1B. Schematic of typical low-voltage power supply used in present-day receivers.

As the receivers were simplified, the voltages and currents needed from the power supply decreased; and the power supply was also simplified. Fig. 4-1B shows the schematic of a power supply simplified almost as much as a transformer-type supply can be. Only one output voltage is available; stages requiring less voltage are provided with individual voltage-dropping resistors. The current requirement is low enough that one 5U4G can provide it. This type of supply, with many minor modifications and additions, was used in numerous television receivers.

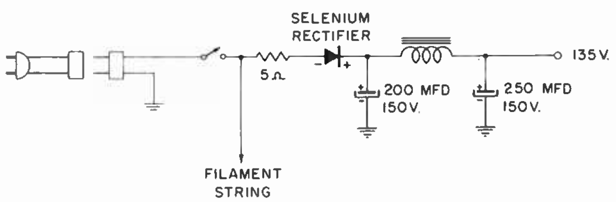


Fig. 4-1C. Schematic of typical low-voltage power supply using a selenium rectifier.

The introduction of semiconductor rectifiers offered an opportunity to simplify and lighten receivers by eliminating the heavy and bulky power transformer. Fig. 4-1C shows a typical half-wave rectifier circuit using a selenium rectifier. Since only 135 volts are available from the supply, the other circuits in the receiver must be designed to operate on this relatively low voltage. Those few circuits requiring more voltage derive it from the boosted B+ voltage from the damper. The development of this extra voltage is discussed in Chapter 8.

The circuit in Fig. 4-1D is used in many portable receivers. The doubler connection produces a volt-

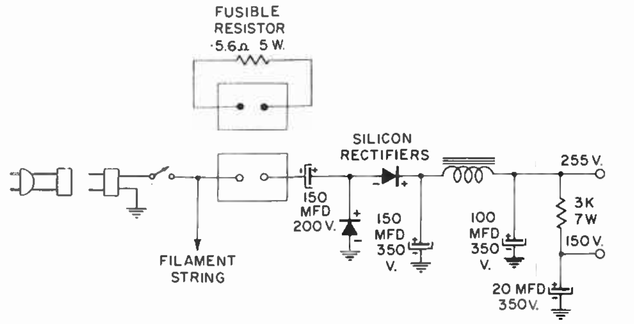


Fig. 4-1D. Schematic of typical low-voltage power supply used in portable television receivers.

age slightly higher than twice the line voltage. The 5.6-ohm fusible resistor in series with the line acts as a fuse. This circuit and the one in Fig. 4-1C have one disadvantage—the chassis is connected to one side of the power line. The user could receive a dan-

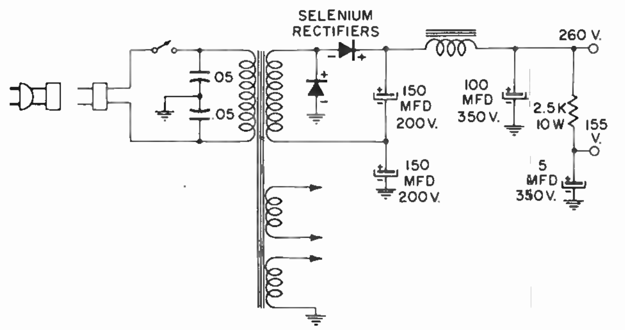


Fig. 4-1E. Schematic of power supply using a transformer and semiconductor rectifiers.

gerous electrical shock between the chassis and any grounded metal if the AC plug is so inserted that it connects the chassis to the high or hot side of the AC line. A power transformer can be used with semiconductor rectifiers, as in the circuit of Fig. 4-1E, to eliminate the shock hazard. This circuit is a full-wave voltage doubler, or full-wave bridge. The bridge configuration is more evident if the rectifier circuit is redrawn, as in Fig. 4-1F.

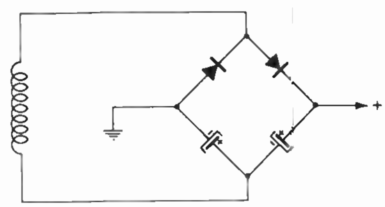


Fig. 4-1F. Schematic of Fig. 4-1E redrawn to show bridge configuration.

When other circuits need various voltages from a power supply, voltage-dropping resistors are necessary. These are usually high-wattage resistors, and they generate considerable heat. To conserve power and to lower the heat in the receiver, manufacturers have used the circuit in Fig. 4-1G. In this circuit, the audio output tube serves as a voltage-dropping resistor, in addition to its role as an output tube.

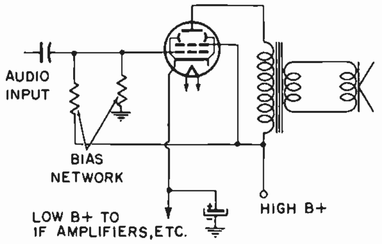


Fig. 4-1G. Schematic of power supply using audio output tube as voltage divider.

The plate resistance of the tube serves as a dropping resistor. The resultant voltage at the cathode is supplied to such stages as the IF, sync, and video amplifiers. Although the output-tube plate current varies at the audio rate, the low B+ is held constant by a large value filter capacitor.

HIGH-VOLTAGE, LOW-CURRENT SUPPLIES

The high-voltage supply differs considerably from the supply just discussed. The current required is quite small, usually around 300 microamperes, while voltages may be extremely high, especially in television receivers using projection-type cathode-ray tubes. Present-day receivers may have accelerating potentials from 3,000 to 30,000 volts.

Since high-voltage power supplies represent extremely dangerous shock hazards, it might be well to consider normal precautions to be followed when working with them.

First, position your bench well away from metal objects or wiring, which might provide an accidental contact to ground or to a voltage source. Don't use metal bench tops. If the floor is concrete, it should be covered with a substantial rubber mat or other good insulating material. Make the mat large enough to avoid any possibility of your stepping off during normal service operations.

Second, don't attempt high-voltage measurements unless they are actually necessary. Resistance measurements will identify many power-supply troubles. If high potentials must be measured, do so in the approved manner. With power switch off and plug disconnected, hook up the test lead to the ground, or low-potential, side of the circuit. If you connect a clip lead to the high-potential side of the circuit, use only one hand. Make sure the power is off and the high-voltage supply is completely discharged before making any connection to the receiver.

If a probe-type instrument is used with the power supply operating, use only one hand to place the probe to the test point. Keep your other hand in your pocket to avoid contact which might establish a circuit through the body. In other service operations, such as alignment, etc., the high-voltage circuit should be disabled. Testing any electronic circuit involves nine mental operations to one manual operation. Think—then act! You owe it to yourself and to your family to take all necessary precautions.

Three different types of high-voltage power supplies have been used in commercial television receivers:

1. A "brute force," 60-cycle, half-wave power-supply system, using a highly-insulated step-up transformer.
2. An RF power-supply system which uses an RF oscillator as a high-frequency voltage source, steps up this voltage through a suitable RF transformer, and then rectifies it for application to the anode.
3. Used almost exclusively in modern TV receivers, the horizontal-flyback or "kickback" type high-voltage supply, employing the pulse voltage generated by the collapsing field in the primary of the horizontal-deflection transformer. This pulse voltage is stepped up, rectified, and supplied to the anode.

"Brute Force" Type

Almost all prewar and a few postwar television receivers used the half-wave "brute force" power supply. Also most oscilloscopes use this type of high-voltage supply. It is essentially the same power supply used in ordinary radio receivers, except that the power transformer must be adequately insulated against high voltages. The half-wave circuit is employed because the high-voltage transformer can be held to the lowest possible number of turns and smallest size. The fact that the filter must be effective for a 60-cycle hum or ripple instead of the 120-cycle ripple of the full-wave supply is not particularly bothersome since the current requirement is relatively small.

Fig. 4-2 shows a "brute force" power supply which provides approximately 12,000 volts DC through the

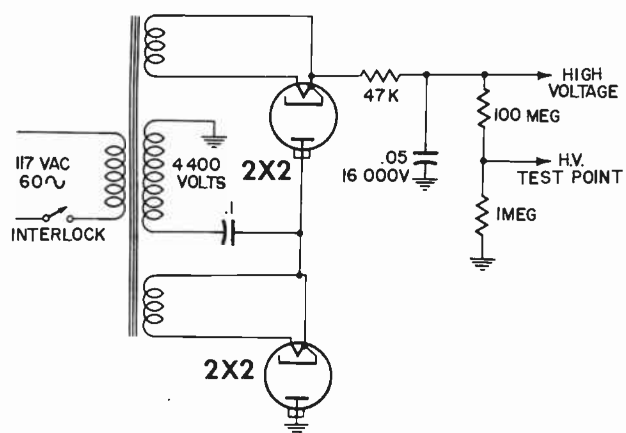


Fig. 4-2. "Brute Force" type high-voltage power supply employing doubler circuit.

use of a half-wave doubler circuit. A pair of 2X2 rectifiers is used in a conventional doubler system fed by a high-voltage winding of the power transformer. This application uses a separate high-voltage supply transformer.

RF Type

The RF oscillator high-voltage power supply was used in some older receivers, especially where electrostatic deflection was employed. It is compact, requiring only two tubes, and is independent from the deflection system of the receiver. Power is generated by an RF oscillator operating at frequencies between 50 and 500 kc. The RF output is stepped up through a transformer to several thousand volts and then rectified. Because of the low current drain on the power supply, the output voltage is nearly equal to the peak voltage applied to the rectifier. A typical circuit is shown in Fig. 4-3.

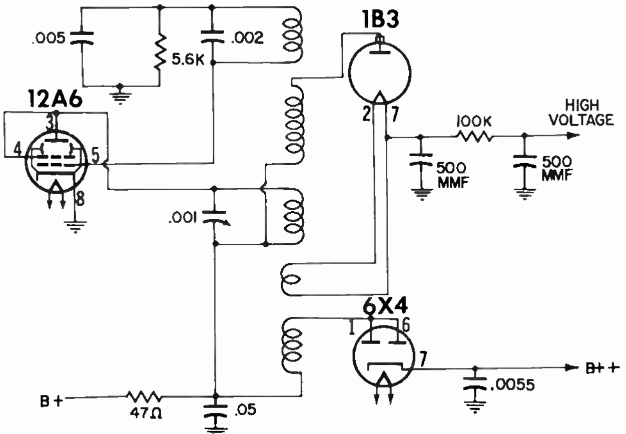


Fig. 4-3. RF oscillator and rectifier high-voltage supply system.

The oscillator usually employs a power-output tube which can generate 10 to 15 watts of RF energy. It normally is connected as a tuned-plate oscillator with tickler feedback. The plate circuit is tuned to the natural resonant frequency of the high-voltage winding, which provides a minimum load on the oscillator circuit. One of the features of this power supply is the fact that any change of capacitance in the circuit, which would result if a hand were inadvertently placed near it, will reduce the oscillator output and lessen the danger of a high-voltage shock. Nevertheless, all aforementioned precautions should be taken when voltage measurements are made; and to avoid severe burns you should not get too close to the cap of the rectifier tube.

Be careful not to change the position of the 1B3 filament winding because any change in coupling will increase or decrease the filament voltage. Since the voltage applied to the filament is RF and there is no practical method of measuring its heating efficiency, a reasonably accurate check can be made by visually comparing the brilliance of the heater on a similar tube with its filament connected across a 1.5-volt dry-cell battery. Since the filament of the 1B3GT is quite easily paralyzed by a momentary overload, another 1B3GT should be tried if trouble is suspected in the rectifier circuit.

The particular receiver in which the circuit of Fig. 4-3 was used had an additional transformer winding connected to the plate of the 6X4 rectifier tube. Thus, additional B+ voltage was provided for the vertical- and horizontal-sweep generators.

Flyback Types

This method of obtaining high voltage is used in nearly all modern TV receivers. It uses the high-voltage pulse created in the plate circuit of the horizontal amplifier during retrace. With this system, few additional components are required since all magnetically-deflected receivers use a matching transformer between the horizontal output amplifier and horizontal-deflection coil. This system also guards against modulation of the video signal by stray energy from the high-voltage supply since the screen is blanked out during retrace. Fig. 4-4 shows an early type of horizontal-flyback system.

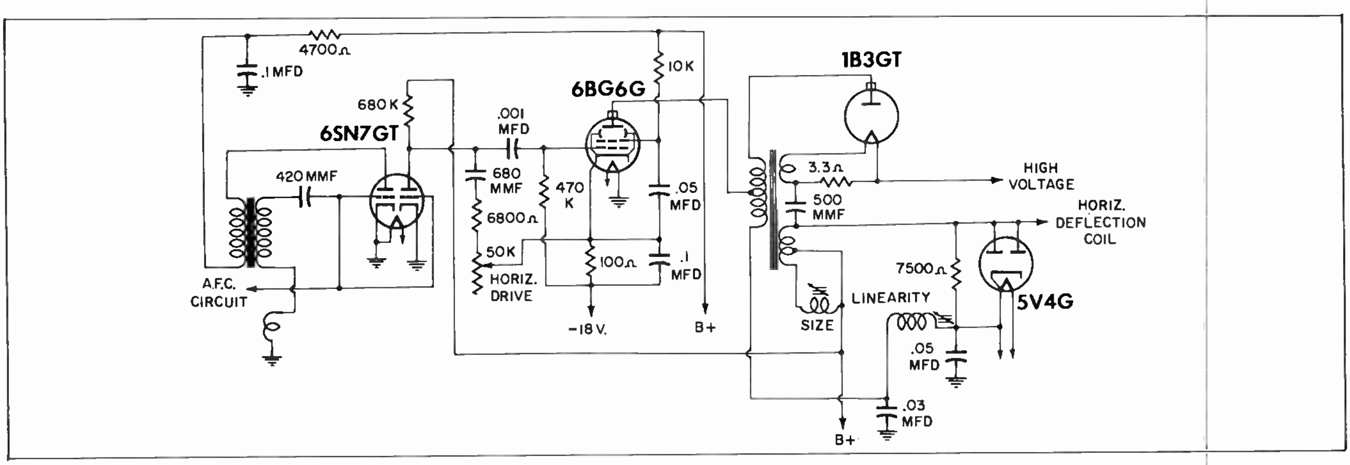


Fig. 4-4. Early type of horizontal-flyback, high-voltage system.

The addition of two windings to the horizontal-output transformer makes it possible to use this system. When the plate current through the horizontal deflection amplifier collapses because of the saw-tooth voltage on its grid, the primary winding (part of the plate circuit of the horizontal-deflection amplifier) will have a relatively high pulse voltage produced across it because of self-induction.

By the addition of an autotransformer winding to the primary of this deflection transformer, the pulse voltage can be stepped up to any desired value. This high voltage is then fed to the plate of a rectifier, where it is rectified and filtered and becomes the high anode potential for the picture tube.

The tertiary winding on the transformer consists of one or two turns which provide the filament power for the rectifier tube. This is possible since a low current is required for this particular tube.

The transformer employs pressed-powdered iron in the form of a shell around the windings. Because of the high horizontal-scanning frequency (15,750 cycles per second), powdered iron can be used in this transformer. The time of one cycle is approximately 63 microseconds. Fifty-three microseconds are used up in the forward scan. The remaining 10 microseconds are employed for flyback and for starting the next horizontal line.

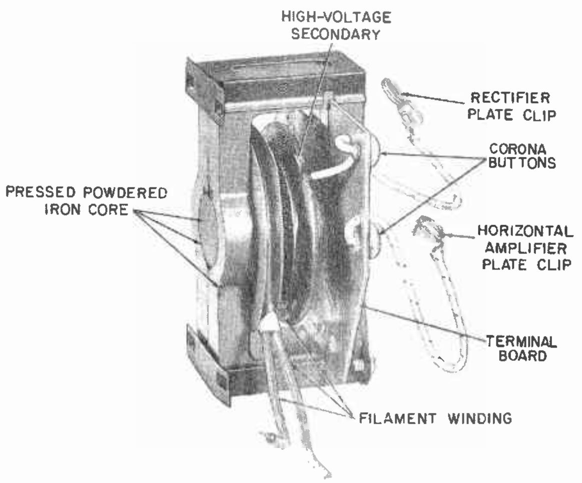


Fig. 4-5. Typical early horizontal-flyback transformer. *(Photo from sample, courtesy of RCA.)*

The windings are of the universal type and are well impregnated. The parts of an early style horizontal-flyback transformer are shown in Fig. 4-5. Since the frequency is high and the current drain is low, the filtering required on this high-voltage supply is small. Actually, the output filter capacity is often realized by using the capacity between the outer and inner layers of Aquadag coating on the picture tube.

The schematic of Fig. 4-6 is typical of a modern flyback-type, high-voltage system. Fig. 4-7 shows a

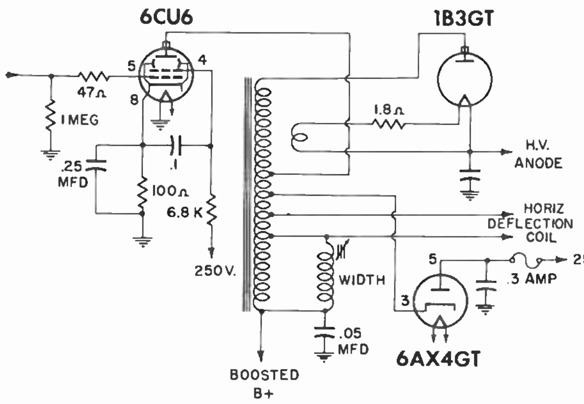


Fig. 4-6. Modern type of horizontal-flyback system.

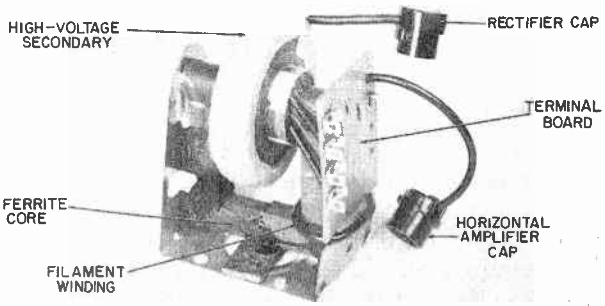


Fig. 4-7. Typical modern horizontal-flyback transformer.

horizontal-flyback transformer of the compact type used in this circuit.

Tripler Type of Flyback System

In the systems previously described, a single rectifier tube was employed to obtain voltages up to 20,000 volts. Projection tubes require potentials between 25,000 and 30,000 volts. Such voltages cannot be readily obtained by single rectifiers because of limitations of the tubes. To overcome this difficulty, power supply systems for projection tubes employ voltage multiplication. A number of capacitors are individually charged to the peak voltage of the system through respective rectifier tubes associated with the capacitors. Fig. 4-8 shows a power supply of this type.

The transformer is similar to the one illustrated in the circuit of Fig. 4-4. The difference is that it has three filament windings for three 1B3GT high-voltage rectifier tubes. Another difference in this system is the use of two horizontal-output 6BG6G tubes connected in parallel to provide the additional energy required.

The diagram shows a "ladder" arrangement of rectifier tubes, capacitors, and resistors for accomplishing this voltage multiplication. This circuit is somewhat different from the familiar "common line" type of voltage doubler or tripler. The individual capacitors need a voltage rating no greater than the

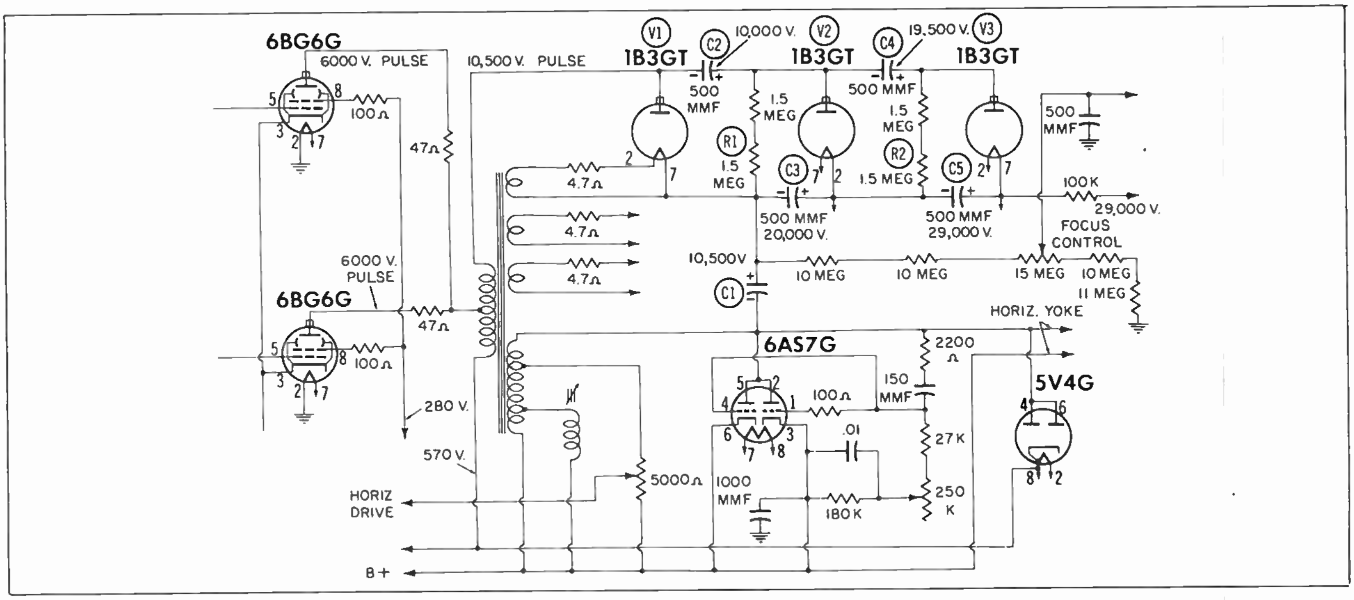


Fig. 4-8. Horizontal-flyback high-voltage tripler.

peak supply voltage. In the familiar multiplier circuits, the voltage ratings increase in each stage. The voltages shown in the diagram of Fig. 4-8 are measured from ground. The circuit operates as follows:

A pulse produced across the transformer primary is applied to tube V1. The rectified current of this tube charges capacitor C1 to approximately the peak value of the pulse. Practically all of this peak voltage charge will appear at the plate of V2 through the low-current path of R1. Then, at the next pulse from the output transformer, the conduction of V2 produces an additive charge across C1 and C3 of 20.000 volts. This voltage is impressed through R2 to the plate of V3. Finally, another pulse from the output transformer is added to the nearly 20,000 volts at the plate of V3 and produces an additive charge across C1, C3, and C5 of 29.000 volts.

This series of events will require a number of cycles of operation by the horizontal oscillator before each capacitor can assume its final charge. When a steady condition is reached. the charge across each capacitor in this group will be approximately the peak voltage of the supply system. Capacitors C1, C3, and C5 in series provide the output voltage for the accelerating voltage of the projection tube.

QUESTIONS

1. What two requirements make the low-voltage power supplies in television different from those in radio?
2. What is the advantage offered by a power supply that employs selenium rectifiers? The disadvantage?
3. When various values of voltages are needed from a low-voltage power supply, what are the methods used?
4. In the RF-type power supply, power is generated by an RF oscillator. What happens to this power before it appears at the output of the supply system?
5. How is the filament voltage for a 1B3GT obtained?
6. In the flyback-type power supply, what is created in the plate circuit of the horizontal amplifier for use in developing high voltage? During what scanning period is it developed?
7. Is the required filtering *small* or *large* in the flyback-type power supply? Why?
8. What is the range of accelerating potentials required by projection tubes? These potentials are produced by what type of voltage supply?

EXERCISES

1. Show the following circuits of present-day low-voltage supplies:
   (a) Using transformer and tube.
   (b) Using selenium rectifiers.
2. Draw the *basic* circuit of a flyback-type power supply using the output stage, the horizontal output transformer, the rectifier, the damper, and the yoke.

SECTION 2

Beam Deflection

Chapter 5

Resistance-Capacitance Circuit Characteristics

The elements in a cathode-ray tube provide an emitter, or source, of electrons; a means of forming an electron beam and accelerating it; and a phosphor-surfaced screen which will fluoresce, or glow, when bombarded by the stream of electrons. Elements are also provided for moving the beam horizontally and vertically to form a frame of light, or raster, on the face of the picture tube.

The voltage or current waveforms required for deflecting the electron beam are obtained from sweep generators followed by special waveshaping circuits. The sweep generators are triggered by the synchronizing pulses derived from the transmitted signal. Thus, the sweep circuits of the receiver can be synchronized with those of the transmitter. The formation of a certain waveshape is required for a linear sweep. This waveshape may be a sawtooth, as is required for electrostatic deflection, or it may be a more complex voltage waveform used to obtain a sawtooth current flow in magnetic-deflection coils. Some circuits will pass a waveform with a minimum of distortion. Others distort greatly when generating, amplifying, or passing a waveform. The behavior of these distortion circuits can best be understood by studying a charging or discharging capacitor in series with a resistor.

Elementary theory states that a voltage, or IR drop, is developed across a resistor when electrons flow through it. The voltage developed by a current flowing through a resistance is found by applying Ohm's law:

$$E = I \times R,$$

where E is in volts, I is in amperes, and R is in ohms.

A further study of fundamentals reveals that a capacitor can store a charge of electrons. When charged, one plate contains more free electrons than the opposite plate. When the capacitor is completely discharged, both plates contain the same number of free electrons. When the accumulation of electrons on one plate exceeds the accumulation on the other plate, a potential difference exists across the terminals of the capacitor. This potential will continue to increase until it equals, for practical purposes, the applied or charging voltage. The value of voltage developed by a charging capacitor is computed with the following equation:

$$E = \frac{Q}{C},$$

where Q is in coulombs, C is in farads, and E is in volts.

One coulomb is the quantity of electrons transferred when one ampere flows for one second.

R-C CIRCUIT CHARGING

A capacitance and a resistance in a voltage-divider circuit (Fig. 5-1) develop a potential across their respective terminals. This circuit is commonly known as an R-C circuit. Both Kirchoff's and Ohm's laws apply to this circuit. Fig. 5-1 shows the voltage divider AB of the circuit diagram in various time

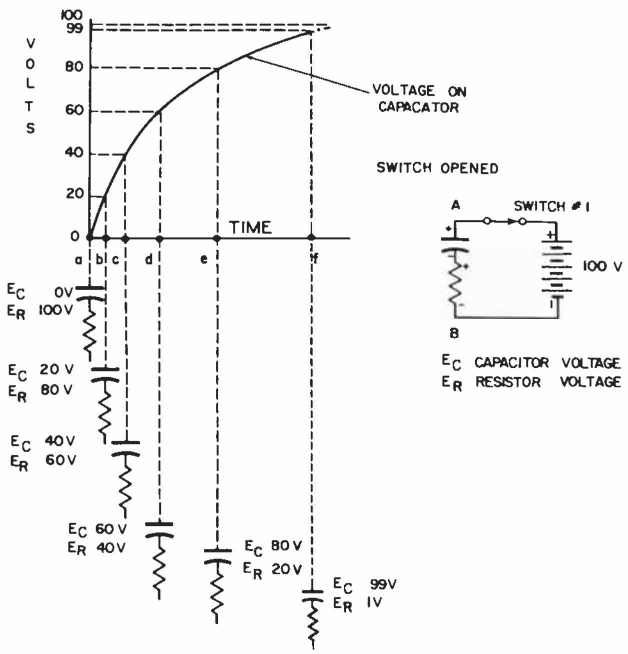


Fig. 5-1. Resistor-capacitor charging curve.

positions on the graph after the switch is closed. As time progresses, the voltage E_C on the capacitor gradually increases, while the voltage E_R across the resistor gradually decreases.

When the switch is closed, electrons are displaced from the upper plate of the capacitor, and a positive charge is developed, causing electrons to be attracted to the lower plate through the resistor. This flow of electrons charges the capacitor. At the instant the current begins to flow, there is no charge on the capacitor, as seen at point *a* on the graph. Therefore, the applied voltage E across the divider must appear as a voltage drop across the resistor, and the initial charge current must equal E/R.

Kirchoff's law states that the sum of the voltages in a closed circuit is equal to zero. Likewise, the sum of the voltage drops in a closed circuit must equal the applied voltage. Therefore, if 100 volts is applied to an R-C circuit, 100 volts appear across the resistor when the switch closes. The graph shows that the instant the switch is closed, the entire applied voltage appears across R, while the voltage across C is zero.

However, the current flowing in the circuit soon charges the capacitor slightly, and a voltage appears across this capacitor. See position *b* of the voltage divider plotted on the graph. E_C is now 20 volts and E_R is 80 volts, the sum of the two being equal to the applied voltage. As time elapses, E_C becomes greater and E_R smaller, as will be noted at the time points *b*, *c*, *d*, *e*, and *f*. Actually, the capacitor voltage becomes a reactive voltage, or back pressure (opposite in polarity and opposed to the applied potential). This reactive voltage causes a decrease in the charging current and in the IR drop across the resistor; therefore, the capacitor charges at a slower rate.

This charging continues until the capacitor is almost fully charged. At this time, the charging current and the voltage across R are practically zero. Theoretically, a capacitor never fully charges, and some minute voltage will always appear across the resistor. However, if the switch is closed long enough, an almost steady state is reached; and the capacitor is considered fully charged for all practical purposes.

Charging a capacitor can be likened to inflating a flat tire. A current of air first rushes into the inner tube and gradually tapers off as the tire becomes inflated. The current of air flowing into the tire rings a bell on the compressor. At first the bell rings in rapid succession, but gradually slows up as the tire builds up a back pressure.

R-C CIRCUIT DISCHARGE

Suppose, just at the time of point *f* on the charging curve (Fig. 5-1), the charging switch (No. 1) is thrown open and a discharging switch (No. 2) is closed, as shown in Fig. 5-2. Note that the capacitor voltage reaches a value of 99 volts. This value would have been slightly higher if the charging circuit had been left closed longer.

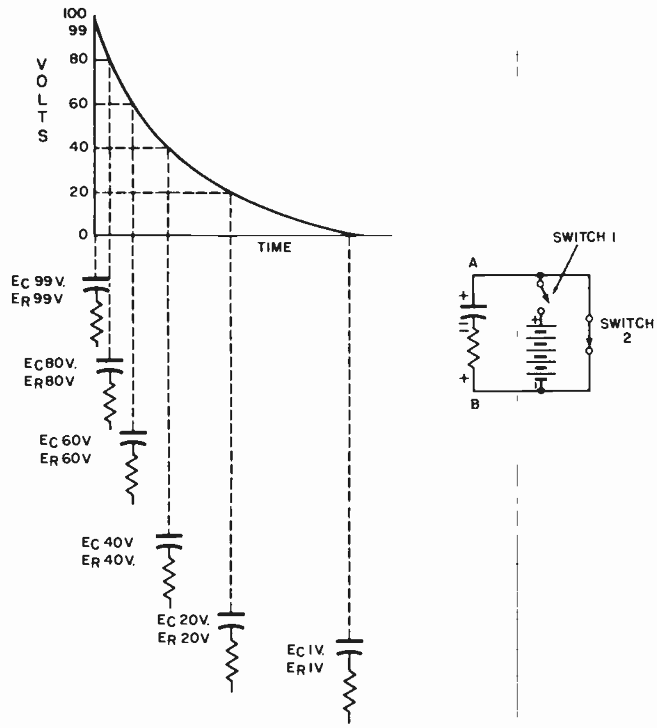
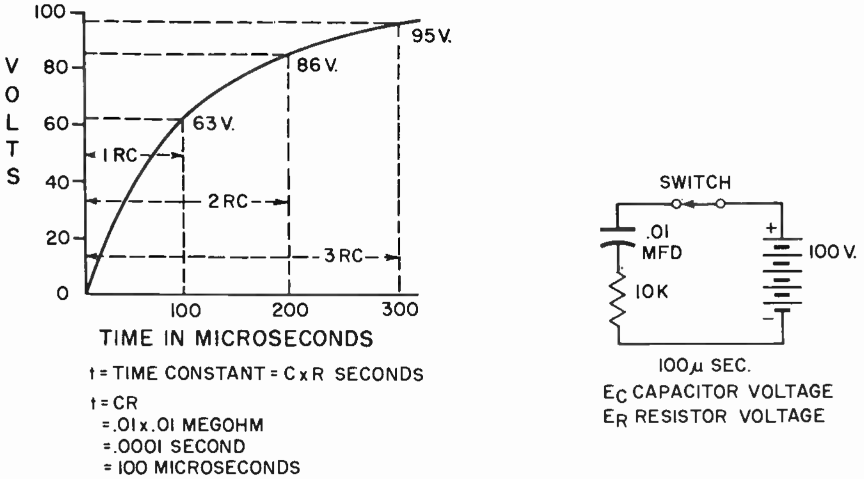


Fig. 5-2. Resistor-capacitor discharging curve.

In Fig. 5-2, the battery switch is open. A short-circuit path is switched across the divider. The 99 volts of potential stored by the capacitor now becomes the applied voltage of the discharge circuit, and a current flows around the circuit. The discharge current will flow in an opposite direction from the charging current, and an IR drop will develop across the resistor. The voltage drop across the resistor due to the discharge current will be opposite in polarity from the voltage drop developed by the charging current.

During discharge, the capacitor voltage drops from its initial value and, representing the applied voltage of the discharge circuit, will equal the voltage drop across the resistor (Kirchoff's law). Since the capacitor voltage now represents the applied voltage of the discharge circuit, E_C and E_R will slowly approach zero together.

The charging curve is not linear throughout. However. the charge portion of the curve (Fig. 5-1) up to 40 volts is practically straight. Because of this linearity, we will be more concerned with this portion of the curve in sweep circuits. This is the most important point of the discussion and should be borne in mind for future reference. Also note that the capacitor voltage does not reverse in polarity during the charge and discharge cycle. This is not true with the resistor voltage because the current through the resistor actually reverses its direction between the charge and discharge period.



1 RC	100μ SEC.
2 RC	200μ SEC.
3 RC	300μ SEC.
4 RC	400μ SEC.
5 RC	500μ SEC.

TIME	E_C	E_R
1 RC	63 V.	37 V.
2 RC	86 V.	14 V.
3 RC	95 V.	5 V.
4 RC	98 V.	2 V.
5 RC	99 V.	1 V.

Fig. 5-3. Time constants of an R-C circuit.

TIME CONSTANTS OF AN R-C CIRCUIT

The diagram in Fig. 5-3 shows an R-C circuit connected across an applied voltage. The time required to charge the capacitor to 63.2% of the applied voltage is known as the time constant of the circuit. The value of this time constant in seconds equals the product of the circuit resistance in ohms and the capacity in farads. It may be found by using any of the following relations:

1. R (in ohms) × C (in farads) = t (in seconds).
2. R (in megohms) × C (in microfarads) = t (in seconds).
3. R (in ohms) × C (in microfarads) = t (in microseconds).
4. R (in megohms) × C (in micromicrofarads) = t (in microseconds).

For example, a 0.1-microfarad capacitor in series with a 100K-ohm resistor will take one-hundredth (.01) of a second or 10,000 microseconds to reach 63.2% of the applied voltage.

In Fig. 5-3, a .01-mfd capacitor is in series with a 10K-ohm (.01 megohm) resistor. Multiplying the capacitance (.01 mfd) by the resistance (.01 megohm) gives a time constant of .0001 seconds, or 100 microseconds, for this circuit. This means that after 100 microseconds have elapsed, 63.2% of the applied voltage is across the capacitor and 36.8% is across the resistor.

Since the applied voltage is 100 volts, the capacitor charge will be approximately 63 volts. Because of the charging current, the IR drop across the resistance will be approximately 37 volts.

In twice the time, or 200 microseconds, 63.2% of the remaining 37 volts is added to the original 63.2% charge. Approximately 86 volts will be across the capacitor, and approximately 14 volts will be across the resistor. Or:

$$200 \text{ microseconds} = 2RC = 63 \text{ volts} + (63.2\% \times 37) = 86.4 \text{ volts}.$$

This value can be found by following the E_C curve in Fig. 5-3.

Theoretically, the capacitor never reaches a fully charged condition. After 5 time constants, approximately 99% displacement of voltage across the circuit has occurred. For all practical purposes, this is sufficient to be considered a full charge. (Refer to the chart in Fig. 5-3.)

The time required to discharge a capacitor through a certain resistance is the same as the time required to charge it through the same resistance. Therefore, the time constant is proportional to the time required to charge or discharge a capacitor.

In 1 time constant of the discharge period, 36.8% of the original charge will remain in the capacitor. The charge and discharge curves are shown in Fig. 5-4. Note the similarity; exponentially, they are the same.

FORMATION OF SQUARE AND SAWTOOTH WAVES

If a source of DC voltage connected to a resistive load is switched on and off in equal alternate periods, the applied electrical pressure across the resistor will be a symmetrical square wave of voltage (Fig. 5-5).

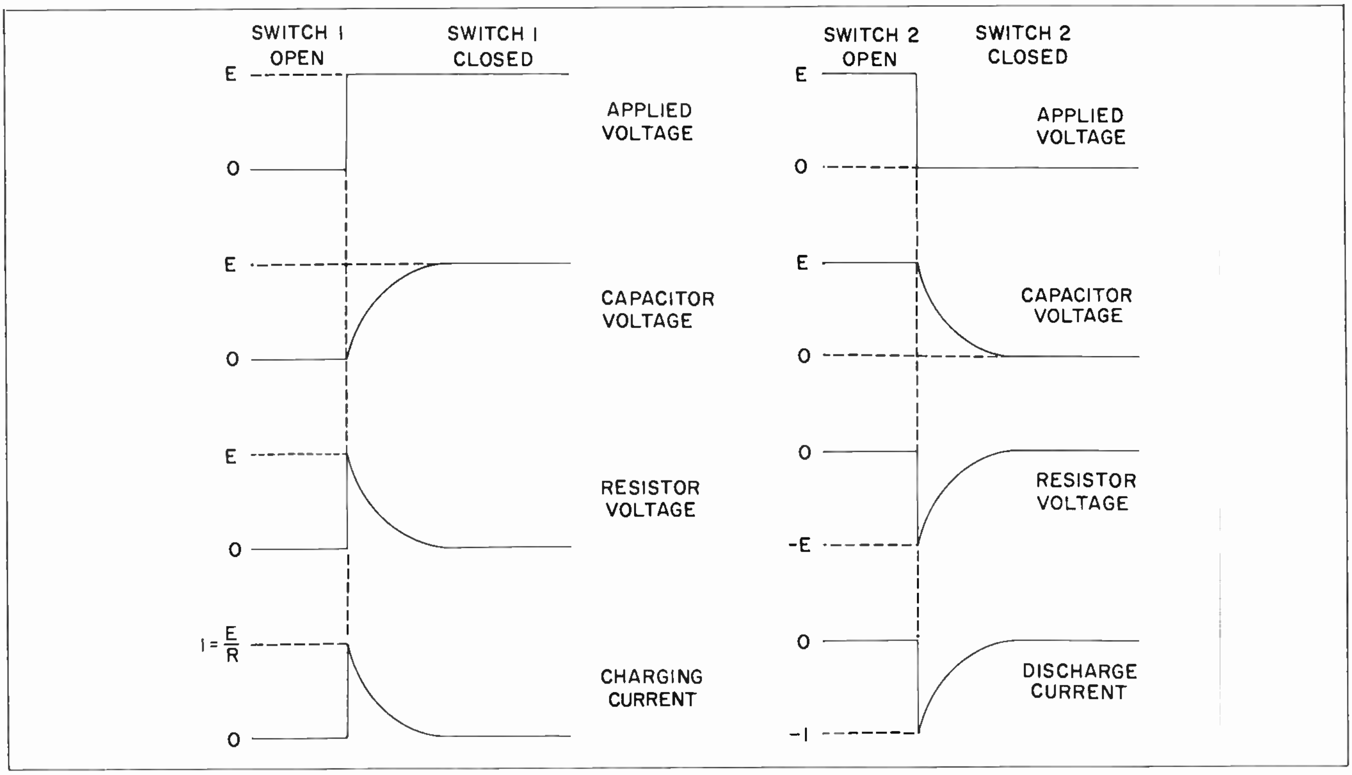


Fig. 5-4. R-C charge and discharge curves.

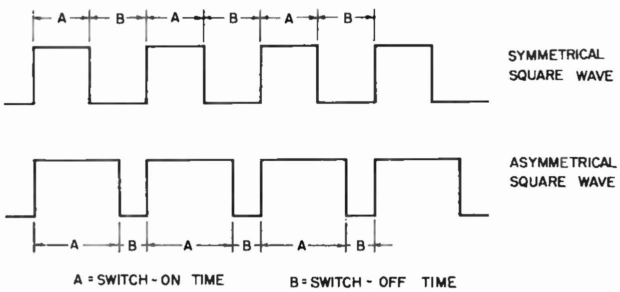


Fig. 5-5. Square waveforms.

On the other hand, if the circuit is switched on and off in unequal alternate periods, the applied voltage to the load will be an asymmetrical square wave (Fig. 5-5). Therefore, by mechanically operating an on-off switch, we can generate two types of voltage waveforms. They are:

1. Symmetrical square wave of voltage.
2. Asymmetrical square wave of voltage.

Now, if a fairly large capacitor is connected in series with the resistor and a DC source of supply is switched on and off in equal time periods to produce an applied square wave of voltage, the resistive and capacitive components of the circuit will produce the following waveshapes (Fig. 5-6).

1. The capacitor voltage, known as the integrator voltage, will appear as a back-to-back sawtooth.
2. The voltage drop across the resistor, known as the differentiator voltage, will appear as a partially distorted square wave.

The polarity of the integrator voltage is unchanged during the charge and discharge period. The differentiator voltage is driven in two directions, positive and negative.

Increasing or decreasing the value of the capacitor in the R-C network will change the integrator and differentiator voltage waveforms (Fig. 5-7). Note that the output waveforms for the 100-microsecond circuit are similar to the ones in Fig. 5-6. When the capacitor is increased to give a time constant of 1,000 microseconds, only a slight voltage is obtained across the capacitor. The voltage across the resistor is distorted very little from the applied waveform. On the other hand, when the value of the capacitor is reduced to give a time constant of 10 microseconds, the voltage waveform across the capacitor is similar to the applied voltage. The waveform across the resistor is greatly differentiated and gives sharp positive and negative peaks. Because the waveforms applied to each of the circuits in Fig. 5-7 have the same frequency and amplitude, a square wave can be differentiated or integrated a variable amount to give the desired waveshape by properly selecting the values in the R-C circuit.

At this point, we are interested in the integrator voltage. So, as we advance still further in the study of waveshapes and circuit analysis, let us refer to Fig. 5-8. Here we have a circuit wherein a capacitor is permitted to charge through a 10K-ohm resistor. By means of a switch, the capacitor is discharged

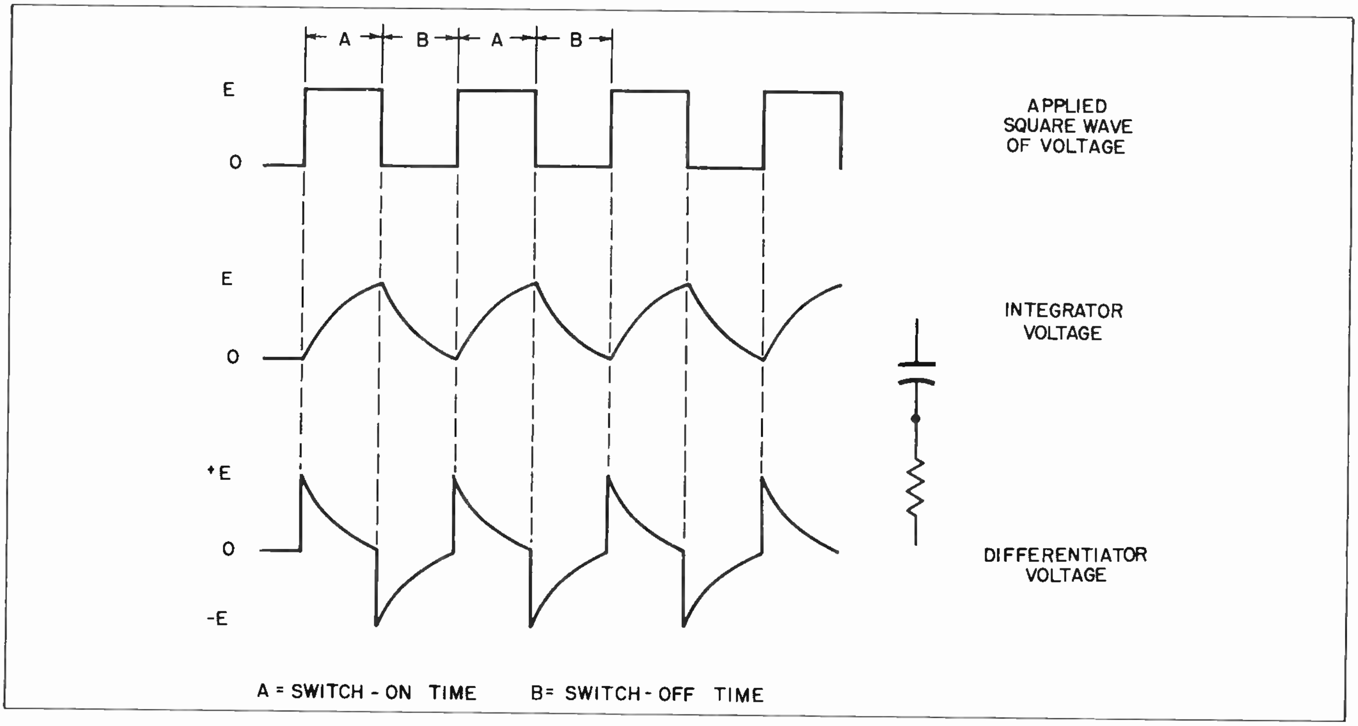


Fig. 5-6. Application of square wave of voltage to R-C circuits.

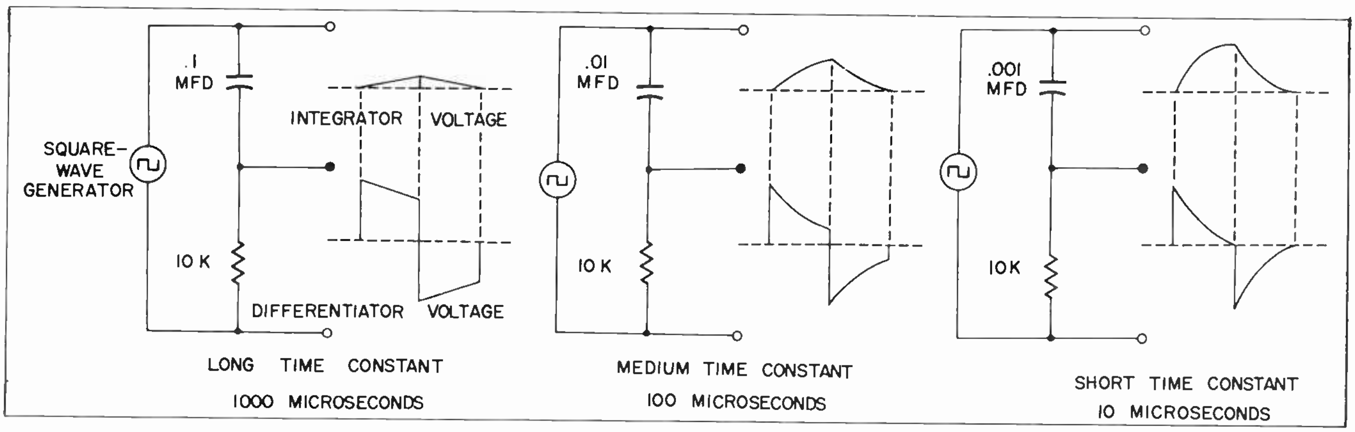


Fig. 5-7. Examples of time constants and their R-C waveforms.

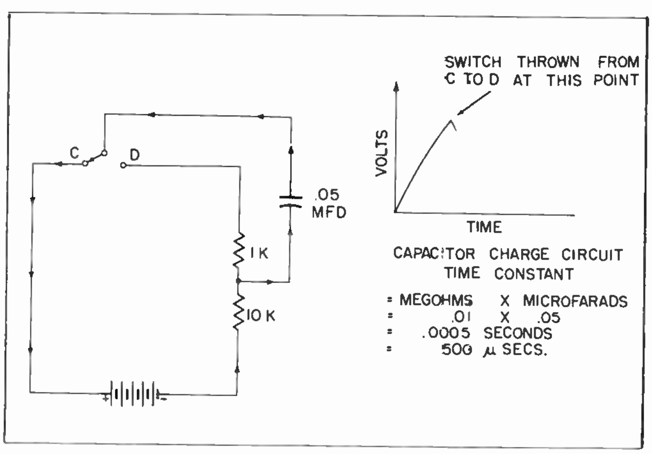


Fig. 5-8. Capacitor charge circuit for sawtooth waveform.

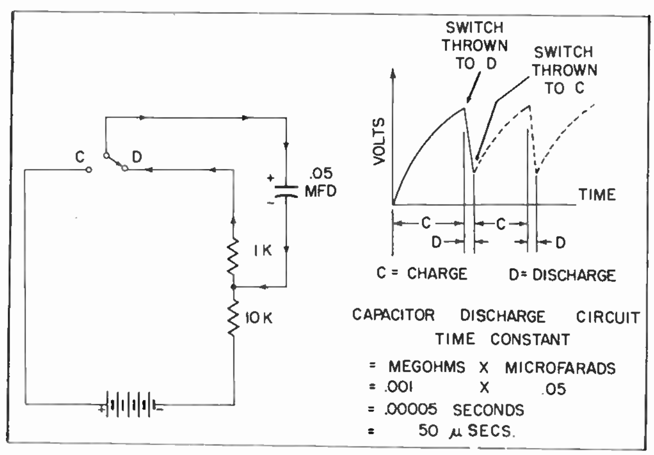


Fig. 5-9. Capacitor discharge circuit for sawtooth waveform.

through a 1K-ohm resistor. If the charge time is longer than the discharge time (for instance, say 10 times longer), the charge and the discharge voltages of the capacitor will be a sawtooth of voltage (Fig. 5-9).

The slow charge and rapid discharge effects can be clearly seen. To obtain linear horizontal and vertical scanning for building a frame or raster, linear sawtooth waves must be generated and applied to the deflection systems of the cathode-ray tube.

QUESTIONS

1. What is the condition of a charged capacitor? What is the condition of a discharged capacitor?
2. When a potential is placed across an R-C circuit, what happens to the voltage across the capacitor and resistor as time progresses?
3. Are the charge and discharge curves of a capacitor linear or nonlinear? What types of curves are these?
4. When the applied voltage is removed from an R-C circuit and the fully-charged capacitor is allowed to discharge, what happens to the voltage across the capacitor and the resistor?
5. Define the time constant of an R-C circuit. What basic equation is used to find the value of the time constant?
6. When a square wave is applied across an R-C network, what is the capacitor voltage called? The resistor voltage?
7. As the charge time constant of an R-C circuit is increased by changing the capacitance, what happens to the voltage across each component when a square wave input voltage is applied?

EXERCISES

1. Show the charge and discharge curves of a capacitor in relation to voltage and time.
2. Find the time constant of an R-C circuit when:
   (a) $R = 100K$ ohms.
   $C = .005$ microfarad.
   (b) $R = 1$ megohm.
   $C = .01$ microfarad.
   (c) $R = 1.5$ megohms.
   $C = 1,000$ micromicrofarads.
3. Show the following voltage waveshapes in an R-C circuit in time sequence.
   (a) Applied square wave of voltage.
   (b) Integrator voltage.
   (c) Differentiator voltage.

Chapter 6

Sawtooth Generators

Production of a sawtooth wave usually involves the charge and discharge of a capacitor through resistors which differ greatly in size between the charge and discharge circuits. An introduction to this concept has been covered in Chapter 5. It has been shown that, to produce a sawtooth waveform, we need a simple circuit consisting of a source of voltage; a single-pole, double-throw switch; resistors; and a capacitor. The capacitor is charged through a high value of series resistance. The voltage across the capacitor at any instant has been shown in Fig. 5-1. Note that the initial portion of this voltage-versus-time curve is essentially a straight line. If we can short-circuit the capacitor before extreme curvature of the charge waveform has occurred and immediately initiate another charging cycle, we have produced a sawtooth wave. This sequence of events could be accomplished with a mechanically operated switch (Fig. 5-9). Since this entire operation must occur in a few millionths of a second, such a mechanically-operated switch is obviously impractical. For this reason, we will resort to some of the properties of electron tubes to accomplish the switching sequence.

Although modern television receivers employ vacuum-tube oscillators and vacuum-tube wave-shaping circuits to produce the ideal sawtooth scanning motion, it will be instructive to examine the earlier forms of circuits. These methods are no longer used in television sets. They are undoubtedly familiar, however, since they are still used in the cathode-ray oscilloscope.

NEON-TUBE OSCILLATOR

The familiar neon gas-filled tube employed in sign lighting is one of the simplest automatic switches for short-circuiting a capacitor at the proper instant to produce a sawtooth voltage wave. A gas-filled tube having a pair of electrodes and connected to a source of electrical potential exhibits interesting properties as the voltage across the electrodes is gradually increased. No electrical current will flow through such a tube until the voltage reaches a value known as the ionization potential. Until this voltage has been reached, the tube acts as an open circuit or as an extremely high resistance. However, when the ionization potential has been reached, the voltage removes electrons from the atoms of the gas and leaves these atoms (ions) with a positive potential. The free negative electrons are rapidly collected by the positive electrode. The positively charged ions are correspondingly attracted to the negative electrode, and current passes through the tube. The resistance of the tube suddenly changes, and it can be considered a voltage-operated switch. Fig. 6-1 shows how such a gas-discharge valve is used as a switch across a capacitor.

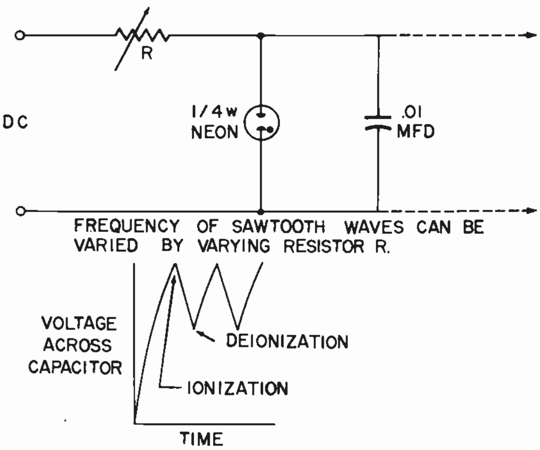


Fig. 6-1. Neon-tube relaxation oscillator.

When the charge in the capacitor has produced a voltage across the tube equal to its ionization potential, the tube will suddenly conduct and start to discharge the capacitor. Once current conduction has started, it will continue even though the voltage has dropped below the original ignition point. Current will continue to flow until a lower voltage level, known as the deionization potential, is reached. At this point, the tube returns to its nonconducting condition.

The charging cycle from the voltage source through the series resistor is resumed, and the cycle continues until ionization again occurs. This sequence of events is diagrammed in Fig. 6-1. Such an automatically-operated switching circuit is known as a relaxation oscillator because the circuit is under normal (or relaxed) charging condition during the periods between its tripping action.

THYRATRON OSCILLATOR

The grid-controlled thyratron is an improved form of gas-discharge tube. Such a tube acts essentially the same as the simple neon lamp previously described, with the following exceptions:

1. A source of electrons from an electrically heated cathode supplies the electrical current for the discharge portion of the cycle.
2. The triggering action is controlled by an additional element similar to the grid of the familiar vacuum tube. This element is normally held at a negative potential and prevents current conduction between the cathode and plate by its repelling action on the electrons emitted by the cathode.
3. The gas normally employed is mercury or argon instead of neon.

A relaxation oscillator can be built with this tube; it is easier to control than a simple neon gas tube. Fig. 6-2 shows the basic circuit of a thyratron sawtooth generator. To assure that the rise of the charging potential is linear, only a small portion of the B+ voltage is allowed to charge the capacitor. The grid is held at a sufficiently negative potential to insure that there is no plate current. In the diagram in Fig. 6-2, this grid potential is provided by a bias battery. Tripping of the circuit can be produced by a positive pulse applied to the grid. Once initiated, plate current will flow until the plate voltage has dropped to a point which corresponds to the deionization potential described in connection with the neon oscillator.

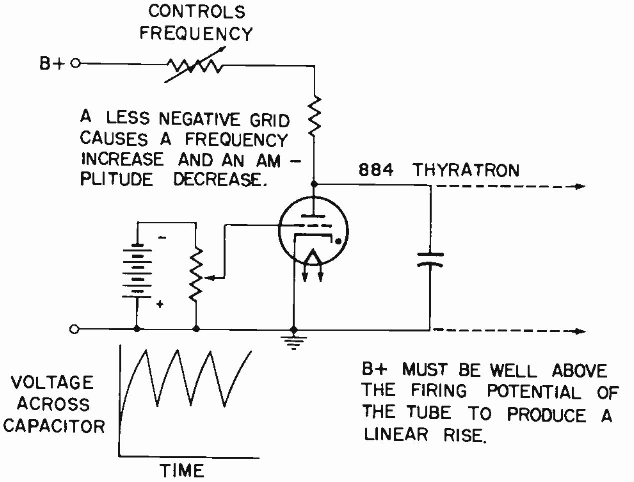


Fig. 6-2. Thyratron relaxation oscillator.

This circuit (Fig. 6-2) is frequently employed in cathode-ray oscilloscopes and was also used in prewar television sets, both in this country and abroad. Television receivers no longer employ gas-tube relaxation oscillators, mainly because such oscillators are not sufficiently reliable in operation with fluctuating power-line voltage, temperature, and time.

VACUUM-TUBE SAWTOOTH GENERATORS

We have seen that, to produce a sawtooth voltage waveform, a voltage is applied to a capacitor through a series resistance. After the capacitor has reached a predetermined charge, the voltage is removed by a short circuit. This action can be more readily accomplished with high-vacuum tubes rather than with gas tubes.

Television sets have generally employed three types of circuit arrangements, or combinations of these circuits, to produce sawtooth waveforms.

1. *The multivibrator.* This circuit arrangement has many variations. The most popular variation is the cathode-coupled version.
2. *The blocking oscillator.* This type of circuit permits the formation of a short pulse of energy. This pulse can be used to produce the sawtooth wave across a capacitor directly associated with the oscillator tube. Or the pulse can trigger a discharge tube which acts as a switch across the capacitor.
3. *The sine-wave oscillator.* An oscillator of the correct frequency supplies the timing voltage for the discharge tube. The sine-wave output of this oscillator is modified into short pulses by wave-shaping circuits. These pulses then operate a discharge tube to produce sawtooth waves.

Multivibrators

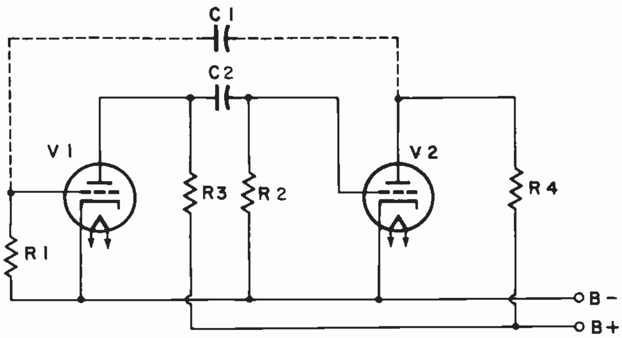
One of the more popular television sawtooth generators is the multivibrator. The multivibrator is another form of relaxation oscillator and employs vacuum tubes, resistors, and capacitors in a feedback arrangement.

The multivibrator is useful because tubes can act as automatic switches to control the charge and discharge of capacitors. This action produces a sustained output of rectangular waveform whose frequency can easily be controlled by the horizontal or vertical synchronizing pulses.

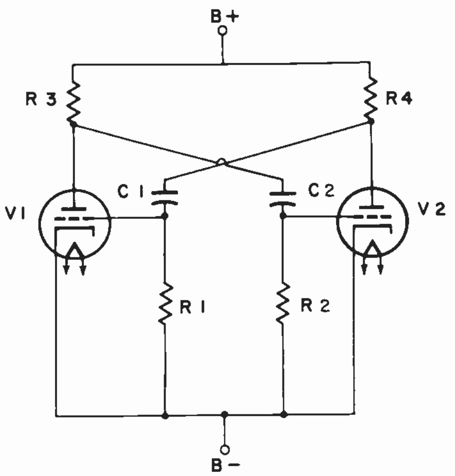
Several versions of the multivibrator circuit are found in modern television sets. Since they are derived from a basic or conventional type, we will first examine the operation of the fundamental circuit.

The Conventional Multivibrator—The basic free-running multivibrator can be considered a two-stage resistance-capacitance coupled amplifier. Over-all feedback is applied by means of a capacitor connected from the output of the second stage to the input of the first stage.

Fig. 6-3A shows a familiar two-stage audio amplifier. The added capacitor C1 (shown in dotted lines) converts this amplifier into a free-running multivibrator. Fig. 6-3B shows a symmetrical rearrangement of the same circuit as it usually appears in textbooks and receiver schematics.



(A) A two-stage amplifier with capacitive feedback.



(B) Same circuit as (A), showing symmetry of circuit.

Fig. 6-3. The conventional multivibrator.

In a single-stage resistance-coupled amplifier, the plate voltage is 180° out of phase with the input voltage. Therefore, the output voltage of the second stage (V2) of the two-stage amplifier will again have been inverted and will be in phase with the input voltage to V1. Capacitor C1 of Fig. 6-3 will impress upon the first grid a voltage of the proper polarity to increase the original input voltage, and oscillation can take place.

To show how oscillation starts and is maintained in this circuit, let us assume that the cathodes are heated and that B+ voltage is applied. Both grid circuits are returned to their respective cathodes through grid resistors. At the instant B+ voltage is applied, the grids will be at cathode potential, or zero bias. Grid and plate currents will start to flow in each tube.

It will be instructive at this point to list the sequence of events which produces the sustained rectangular-shaped output wave of the multivibrator. (Refer to Figs. 6-3 through 6-6.)

1. Since the resistance of the internal cathode-to-grid path under this initial zero bias and high grid current is much lower than the resistance of grid resistors R1 and R2, capacitors C1 and C2 will begin charging from the B+ supply through resistors R4 and R3, respectively. This charging path is shown by the arrows in Figs. 6-4A and 6-4C.

2. If the characteristics of both tubes and the value of the circuit elements were exactly matched, the charging rate of both capacitors would be identical, and the plate currents of both tubes would rise simultaneously. A state of equilibrium would be reached, and the circuit would not oscillate. Such conditions are not met in practice, and a balance is not established.

3. Actually, one of the tubes will start to conduct plate current sooner than the other. This conduction could be caused by lower plate resistance, hotter cathode, or a slightly lower plate-load resistor. Let us assume that the plate current of V1 has started to rise a fraction of a second ahead of the plate current of V2.

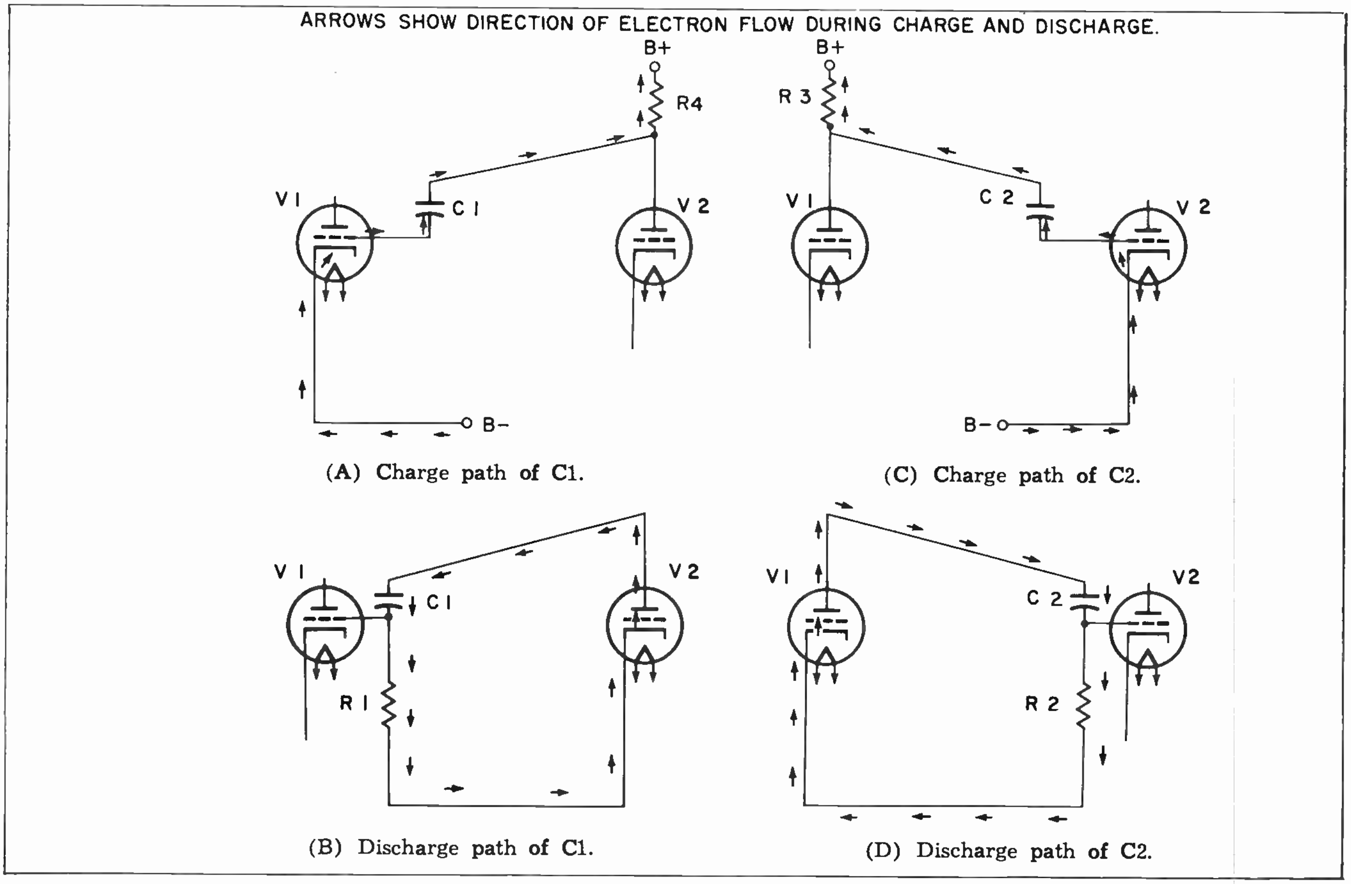
4. This rise of plate current will be accompanied by a drop in plate-to-cathode resistance and by a corresponding drop in plate-to-cathode voltage. Fig. 6-5A shows this set of operating conditions. The low plate resistance of V1 forms a discharge path for C2, as shown in Fig. 6-4D.

5. The discharge current of C2, flowing through the high value of grid resistor R2, develops a high negative grid bias on the grid of V2. This bias drives the tube beyond plate-current cutoff, as shown in Fig. 6-5B. The bias developed by this discharge can be as high as 30 to 50 volts in the example shown.

6. Since the plate current of V2 has been cut off, its plate-to-cathode voltage becomes that of the B+ supply (Fig. 6-5B). The plate-to-cathode voltage will remain at that value until the grid voltage has reached the point where the grid is no longer cut off.

NOTE: Since the foregoing conditions have brought the cycle of operation to one of the two stable, or relaxed, operating points of the circuit, it would be helpful to summarize the changes of circuit voltages which have occurred over the period covered by steps 1-6.

V1 — Plate-to-cathode voltage at its minimum value and steady. Tube conducting. Control-grid voltage zero and steady.



(A) Charge path of C1.

(C) Charge path of C2.

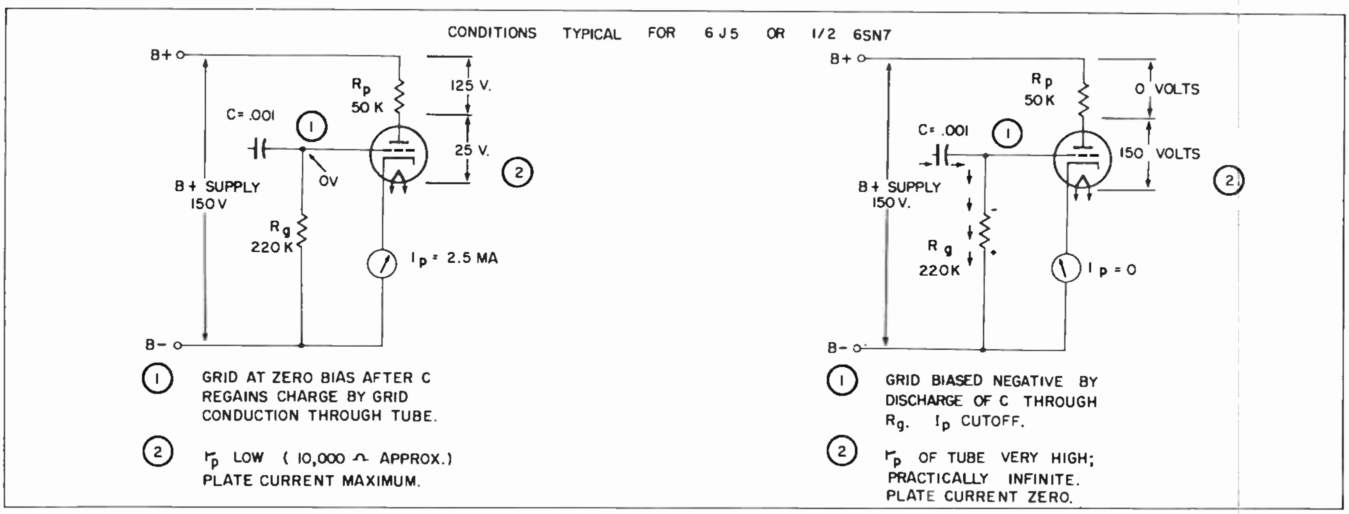
(B) Discharge path of C1.

(D) Discharge path of C2.

Fig. 6-4. Capacitor charge and discharge paths in a multivibrator.

V2 — Plate-to-cathode voltage at its maximum value and steady. Tube not conducting. Control-grid voltage highly negative, but falling exponentially with time as C2 discharges through R2.

7. The time required for C2 to discharge will depend on the time constant of the discharge circuit of C2, R2, and the plate resistance of V1. (See Fig. 6-4D.) The negative voltage across R2, which constitutes the grid bias of V2, finally becomes low enough to allow V2 to conduct heavily. Fig. 6-6 shows the waveforms of the grid and plate voltages of both tubes as a function of time. That part of the waveforms between (a) and (b) in Fig. 6-6 covers the steps outlined up to this point.



(A) Tube conducting.

(B) Tube not conducting.

Fig. 6-5. Tube operating conditions in a multivibrator.

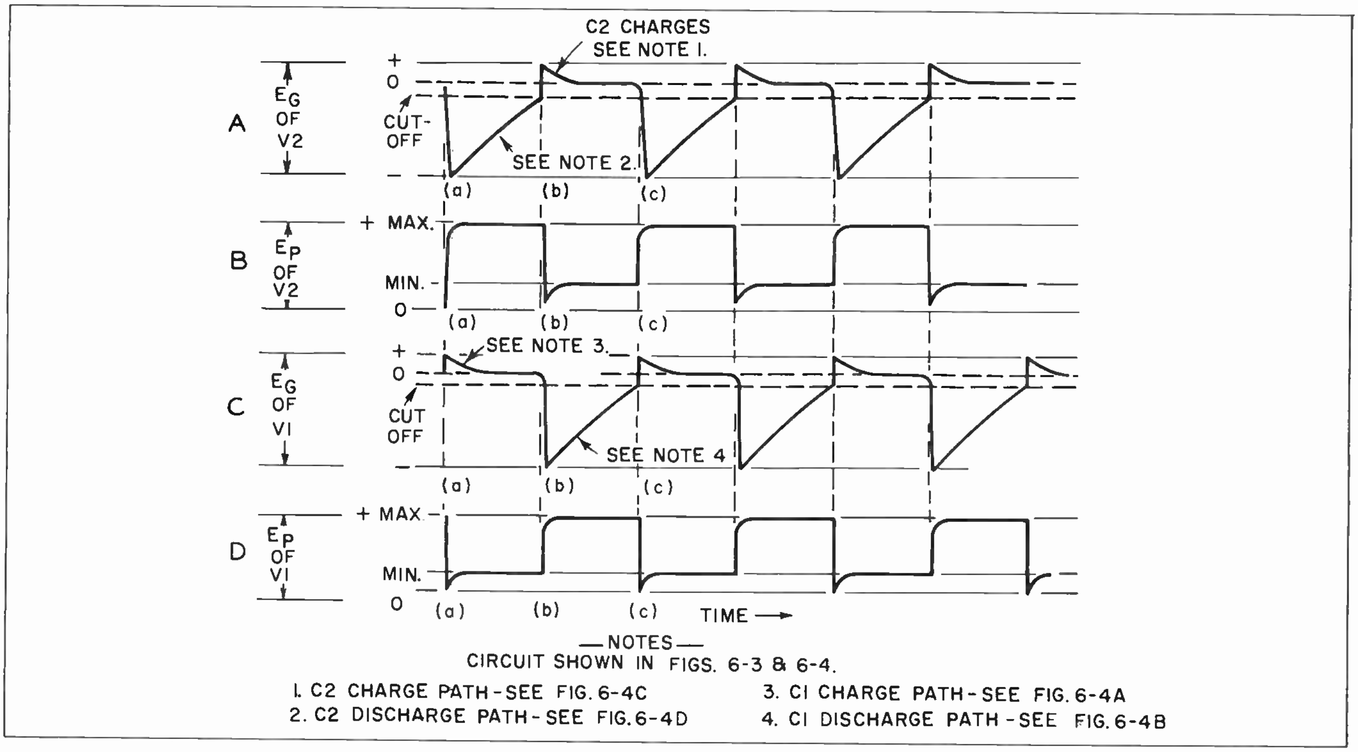


Fig. 6-6. Typical waveforms of a symmetrical multivibrator, showing square-wave switch action.

8. As V2 starts to conduct, conditions in this tube become identical to those in step 4 for V1 except that the tubes and capacitors have exchanged functions, and the discharge path of capacitor C1 is now as shown in Fig. 6-4B.

9. The discharge current of C1 flowing through R1 now biases V1 beyond cutoff, as described in step 5.

10. Since the plate current of V1 has been cut off, the plate-to-cathode voltage assumes the value of the B+ supply, similar to V2 in step 6.

11. The rise in plate voltage of V1 is impressed on capacitor C2, starting the charging cycle shown in Fig. 6-4C.

12. Since the internal grid-to-cathode path of V2 is conductive because of the zero grid-bias, the charging resistance is small, and C2 is rapidly charged. This action is shown at time (b) in Fig. 6-6.

We can now summarize the conditions of circuit voltage and compare them with those found at the end of step 6.

V1 — Plate-to-cathode voltage at its maximum value and steady. Tube not conducting. Control-grid voltage highly negative, but falling exponentially with time as C1 discharges through R1.

V2 — Plate-to-cathode voltage at its minimum value and steady. Tube conducting. Control-grid voltage zero and steady.

Note that the new conditions, which represent the other stable, or relaxed, operating point are the same as before except that the tubes and grid circuits have changed places. This cycle of events is shown in the waveform diagrams of Fig. 6-6 between times (b) and (c). This square-wave generation will continue at a frequency determined by the charge and discharge time constants of coupling networks R1-C1 and R2-C2.

In this symmetrical circuit it has been assumed, but not stated, that the corresponding grid resistors, plate resistors, and coupling capacitors are equal. When this is true, the time constants are equal, and the output waveforms from the plates are identical. The frequency of this multivibrator can be changed by altering either resistors R1 and R2 or capacitors C1 and C2. A lower time constant

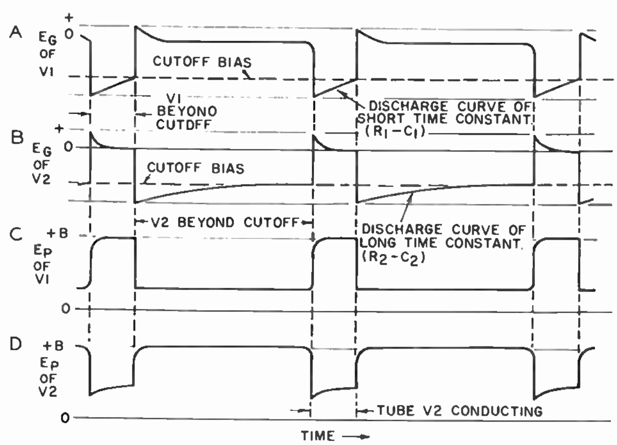


Fig. 6-7. Waveforms of an asymmetrical, or unbalanced, multivibrator.

will increase the frequency. If the values of R or C are changed equally, the output wave remains symmetrical.

The Asymmetrical, or Unbalanced, Multivibrator—To produce the type of sawtooth wave required for television scanning with a multivibrator, succeeding square waves must be unequal in length or spacing. For this reason, the time constant of the R-C circuit of one tube is made much greater than the time constant of the other. Such a multivibrator is called asymmetrical. Fig. 6-7 shows the waveforms obtained when the circuit constants of the symmetrical multivibrator just described are so changed that the product (R1 × C1) in the grid circuit of V1 is much smaller than the product (R2 × C2) of V2.

Waveform D of Fig. 6-7 shows a short pulse of plate current occurring in V2 once each cycle. We will employ this pulse to produce the scanning sawtooth in proper time relationship to the scanning of the camera tube at the transmitter.

Use of the Multivibrator to Produce Sawtooth Scanning—Fig. 6-8 shows a circuit similar to those discussed for symmetrical and asymmetrical multivibrators. By adding two new circuit elements, we can generate sawtooth voltage waves to control the electron beam with either the horizontal or the vertical sweep circuits. These new circuit elements are C3 and C4 (Fig. 6-8). Coupling capacitor C3 connects the multivibrator circuit to a source of synchronizing pulses, which are part of the transmitted television signal. The function of the synchronizing pulses and how they control the frequency of the multivibrator will be covered in a later chapter.

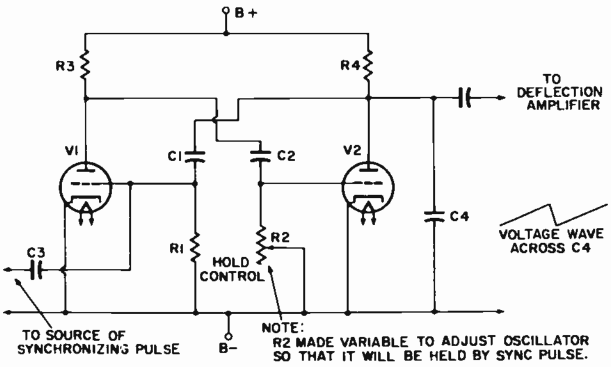


Fig. 6-8. Asymmetrical multivibrator, showing input coupling and sawtooth-forming capacitor.

At this time, the additional circuit element which concerns us is capacitor C4 between the plate and cathode of V2. For a horizontal-line scanning frequency of 15,750 cycles per second, the circuit is so arranged that the time constant of R1 and C1 is approximately one-ninth the time constant of R2 and C2. The plate current of V2 will consist of short pulses (Fig. 6-7) which represent low resistance. During the time shown as the conducting period, V2 will act as a short circuit across capacitor C4. The multivibrator thus acts as a periodic switch and fulfills the requirements covered previously for producing a sawtooth wave.

A significant difference between the circuit in Fig. 6-8 and the one in Fig. 6-3 is that R2 has been made variable. This variable resistor is one of the major controls of a television receiver. From the previous discussion of multivibrator theory, we know that varying R2 alters the length of the portion of the operating cycle controlled by R2 and C2 in Fig. 6-7. This represents the active portion of the sawtooth wave when the face of the cathode-ray tube is scanned during video modulation. This variable adjustment permits the multivibrator to be locked-in with the synchronizing pulse and is known as a hold control.

The voltage of the sawtooth wave across capacitor C4 (Fig. 6-8) is too small to produce the required deflection of the electron beam; therefore, amplifiers are needed.

Cathode-Coupled Multivibrator — A variation of the multivibrator is the cathode-coupled circuit. This circuit is shown in Fig. 6-9. A significant difference between this circuit and the conventional multivibrator is that feedback is accomplished in two ways. Coupling capacitor C2 transfers charges from the plate of V1 to the grid of V2. In addition, this circuit employs a cathode-bias resistor common to V1 and V2. This common-cathode resistor is responsible for the unique action of the circuit. The second tube (V2) functions as a switch or discharge tube for capacitor C4, which produces

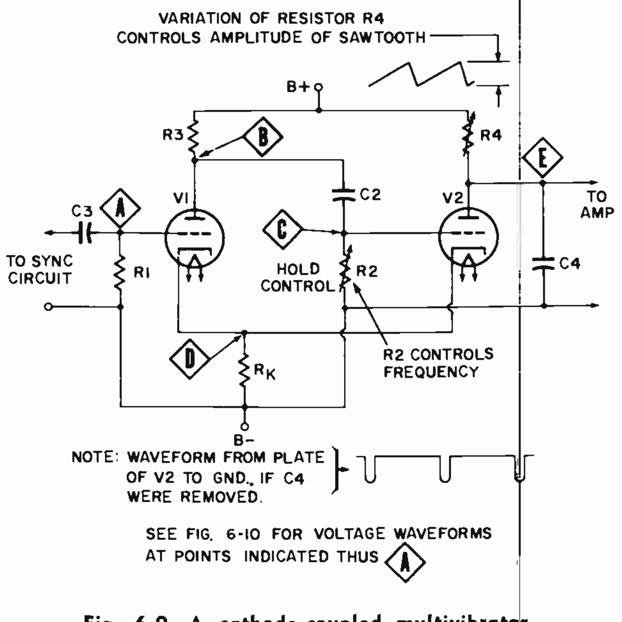


Fig. 6-9. A cathode-coupled multivibrator.

the sawtooth waveform. From the theory of the conventional and asymmetrical multivibrators as previously discussed, the action of this cathode-coupled version can be readily understood. We will again assume that the cathodes of the tubes are heated and that B+ potential is applied. Let us follow the sequence which allows this circuit to generate asymmetrical pulses.

1. Capacitor C2 will charge through R3 and the grid to ground (or B−) circuit of V2. This action occurs quite rapidly since the grid of V2 is initially at zero potential.

2. Plate current will start to flow in both V1 and and V2. Bias voltage for both of these tubes will be developed across cathode resistor R_K.

3. This bias voltage will immediately start to decrease the plate current of both tubes, which were initially in a conductive condition since the control grids were at zero potential.

4. The flow of plate current through V1 causes a lower plate-to-cathode voltage drop and a corresponding lower plate resistance in this tube.

5. The low-resistance path of V1 initiates the discharge of coupling capacitor C2 through R2 and R_K. As in the conventional multivibrator previously discussed, this current flow through R2 produces a high negative bias on the control grid of V2 which immediately drives the tube beyond its plate-current cutoff point. Note that this circuit differs from the conventional multivibrator; there is no coupling capacitor between the plate of V2 and the grid of V1. For this reason, V2 is immediately thrown into plate-current cutoff.

As was done in our discussion of the conventional multivibrator, it will be constructive to summarize the voltage conditions which have occurred up to this point:

V1 — Plate-to-cathode voltage at its minimum value. Tube conducting. Control-grid voltage negative and steady. The tube is self-biased by its own plate current through common cathode resistor R_K.

V2 — Plate-to-cathode voltage rising along the linear portion of the charging curve. This rise of plate voltage charges capacitor C4 and initiates the first part of what will eventually become a sawtooth wave of voltage. Control-grid potential highly negative and exponentially diminishing in value.

6. V2 has been cut off during this period, and no plate current has been flowing. C2 has been discharging through R2, R_K, and the cathode-to-plate circuit of V1. It is interesting to note that, as the plate current of V2 was cut off by the high negative bias produced across R2 by the discharge of C2, the plate-to-cathode voltage did not immediately assume the value of the B+ supply. This is because the plate voltage of V2 was maintained by the charge of C4, which started simultaneously with the closing of the B+ circuit. For this reason, the plate voltage wave of V2 will be sawtooth instead of rectangular because of the charge flowing into C4. If C4 is removed from the circuit, the plate-to-cathode voltage of V2 would rise immediately to the B+ value since the grid of this tube is cut off by the high negative voltage resulting from the discharge of C2.

7. As in the other types of multivibrators already discussed, when the bias of V2 falls to a value equal to the grid cutoff potential, V2 will start to conduct.

8. When conduction occurs in V2, C4 will be rapidly discharged through the plate-to-cathode circuit of this tube, and its voltage will drop to a minimum value (Fig. 6-10E). The sequence of

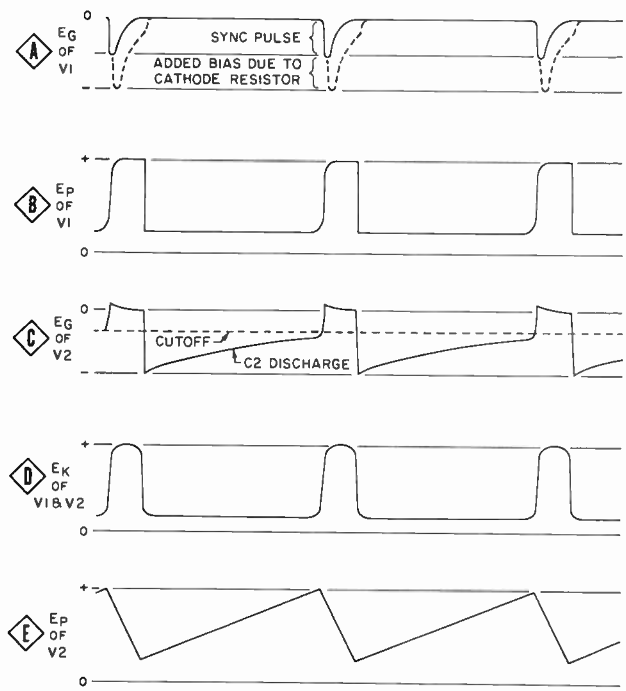


Fig. 6-10. Voltage waveforms in cathode-coupled multivibrator.

events to this point has resulted in a sawtooth wave of voltage across capacitor C4. Thus far, the circuit is like the conventional and asymmetrical multivibrators previously discussed.

9. A different action now takes place. The sudden pulse of plate current, which occurs when V2 conducts, flows through cathode resistor R_K. Since this resistor is common to the cathode circuits of both V1 and V2, the voltage produced by this plate-current pulse immediately drives the grid of V1 negative with respect to its cathode.

10. This negative bias suddenly increases the plate-to-cathode resistance and the plate-to-cathode drop of V1. The sudden increase in V1 plate voltage charges C2. A positive voltage is instantane-

ously impressed on the grid of V2; and the plate current pulse of V2, which started the cycle, is momentarily increased.

11. The cumulative increase of plate current through common-cathode resistor R_K finally biases the grid of V1 enough to completely cut off the plate current, and the plate-to-cathode voltage of V1 rises to its maximum value.

12. Capacitor C2 has become charged, and the plate current of V2 relaxes.

13. This decreased current flow in R_K reduces the bias of V1. As V1 conducts, C2 is discharged through R2, R_K, and the plate circuit of V1. The current flow in R2 drives the grid of V2 to cutoff, and the cycle is repeated.

Because a sudden and cumulative action was produced in the circuit by the coupling of tubes V1 and V2 through a common-cathode resistor, this circuit is called a cathode-coupled multivibrator. The cathode-coupled multivibrator is preferred in television over the asymmetrical multivibrator because:

1. It can be triggered and controlled by a negative pulse of voltage. Thus, the control circuits can often be simplified.

2. Its sudden and cumulative pulsing action in V2 permits a higher ratio of linear-sweep time to return time.

3. Variable resistors R2 and R4 of V2 (Fig. 6-9) permit control of its scanning frequency and amplitude.

The waveshapes of the voltages at various points in this circuit as a function of time are shown in Fig. 6-10. These waveshapes are identified in Fig. 6-9 by letters enclosed in diamonds. Voltages are always measured to B– or ground.

Blocking Oscillators

Another type of vacuum-tube circuit for producing controlled sawtooth voltage waves is the blocking oscillator. The blocking oscillator was originally quite popular, especially as a vertical-sweep oscillator, but has been discontinued in favor of the multivibrator.

Fig. 6-11 shows a simple blocking oscillator. Upon casual inspection, it looks like a Hartley

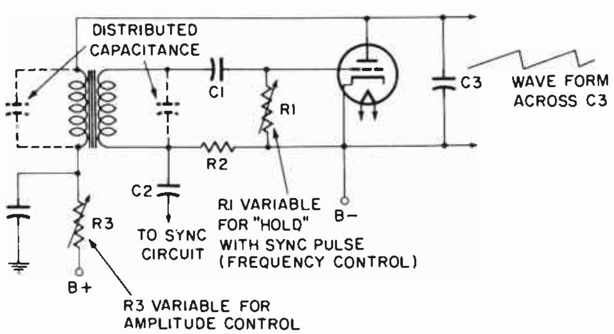


Fig. 6-11. Blocking oscillator.

oscillator with an iron-core transformer. Basically, it is such an oscillator. However, instead of sustained sine-wave oscillations, it produces short pulses of energy, with correspondingly long intervals of relaxed action. For this reason, it is classified as another form of relaxation oscillator. Two significant differences distinguish this circuit from the common Hartley oscillator:

1. The time constant of grid resistor R1 and grid capacitor C1 is such that long periods of blocked plate current occur between short periods of plate-current conduction. During these short conductive periods, oscillation takes place.

2. The natural period of oscillation of the transformer, with its associated distributed and lumped circuit capacitances, is such that the desired pulse time approximates one-half cycle of the frequency at which the circuit would oscillate if it were the continuous sinusoidal type.

As was done with the other types of sawtooth oscillators, it will be instructive to follow through, in sequence, the various actions which take place in this circuit.

1. We will again assume that the cathode of the tube is heated and that the plate circuit is suddenly closed to provide B+ potential. Since the grid is initially at cathode or zero potential, plate current will start to flow through the transformer primary. This sudden rush of current will set up a magnetic field in the core of the transformer, and a secondary voltage will be induced across the grid winding. The direction of these windings is such that the primary current will cause a positive potential to appear at the grid with respect to the cathode or ground.

2. The positive voltage applied to coupling capacitor C1 makes the grid more positive than the cathode. The grid then attracts electrons from the emitted cathode current, and grid current flows through resistor R1.

3. Simultaneously, the increasing positive grid potential makes the plate draw still more current, until plate-current saturation is reached. When the plate current reaches a steady maximum value, no further change of current occurs in the primary winding of the transformer.

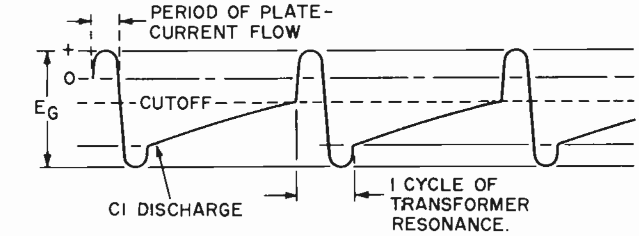
4. Since the voltage induced into the secondary depends upon the change of magnetic flux, the secondary voltage of the transformer will cease to rise.

5. As the grid becomes less positive (C1 discharging through R1), the plate current through the primary falls, and the magnetic field linking the secondary coil collapses. The time taken for this sudden rise and fall of grid voltage is governed by the natural resonant frequency of the transformer and its associated circuit capacitances.

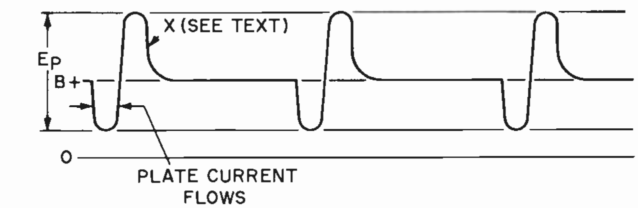
6. The collapsing field in the transformer due to the dropping plate current induces a second-

ary voltage. This secondary voltage is opposite to the original plate-current pulse. Capacitor C1 discharges through resistor R1. The grid is driven more and more negative. The plate current quickly falls until it finally reaches a cutoff point. Although the reversal of grid voltage and the cutoff of plate current has taken considerable time to describe, the action is practically instantaneous.

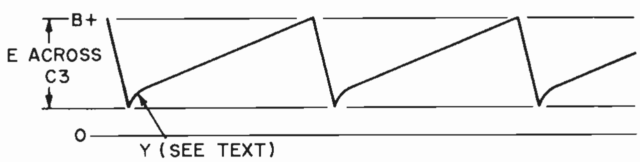
7. From this point, the action in the tube follows the action in the multivibrator. The grid potential follows an exponential curve of R-C discharge until plate conduction is again reached. The waveform of grid voltage is shown in Fig. 6-12A.



(A) Grid-voltage waveform.



(B) Plate-voltage waveform which would occur if C3 (Fig. 6-11) were disconnected.



(C) Sawtooth waveform across C3.

Fig. 6-12. Voltage waveforms in a blocking oscillator.

8. The time taken for the discharge of C1 depends upon the time constant (R1 + R2) × C1.

9. As the tube starts to conduct again, oscillation begins and the cycle is repeated. From the curve of Fig. 6-12B, we see that the plate voltage of the tube is nearly steady and is at the B+ value between these oscillatory pulses. We have fulfilled the conditions of sawtooth charge and discharge of capacitor C3 in Fig. 6-11 and have, therefore, produced a sawtooth scanning wave. As in previous circuits, grid resistor R1 can be made variable for use as a frequency control (hold control), and a plate-circuit resistor can be made variable for use as a width or height control.

In the circuit in Fig. 6-11, note that the sawtooth-generating capacitor C3, connected from plate to cathode of the tube, has modified the shape of the plate-voltage wave. A similar action occurred in the cathode-coupled multivibrator circuit i Fig. 6-9. Without capacitor C3, the voltage wav between the plate and cathode of the tube woul look like the wave in Fig. 6-12B. The amplitud of the wave above the B+ axis at point X caused by the energy stored in the transforme primary.

When capacitor C3 is connected, the voltag wave from plate to cathode assumes the shap shown in Fig. 6-12C. This is the desired sawtoot deflection wave except for the distorted sectio at point Y. This sudden rise of the curve at th point is due to the additional charging voltag when the magnetic flux of the transformer co lapsed, as explained in the foregoing. The plat voltage could not follow the curve in Fig. 6-12 because of the terminal voltage of capacitor C The distorted section of the wave at point Y actually of no consequence because it is blanke out, as will be explained later.

Discharge or Trigger Tube — In some televisio receivers, pulse generators do not directly contr the sawtooth charge and discharge of the capac tor; they are used to trigger an additional tu known as a discharge tube. This tube short-circui (discharges) the capacitor. Fig. 6-13 shows a bloc ing oscillator of the kind previously described. I grid is conected directly to a second tube, whos only function is to conduct plate current and di charge capacitor C2 at the proper instant to pr duce the sawtooth wave. This extra tube is mo independent than the simpler form in Fig. 6-1 which might suffer from some interaction of co trols. The discharge tube can be used with a relaxation oscillator where the plate current o curs over short intervals of the operating cycle.

Sine-Wave Generators

The familar sine-wave oscillator can produ pulses which will trigger a discharge tube asso ated with a sawtooth-waveforming capacitor. On a short portion of the sine wave is used. It passed through clipping stages to "bite off" a sma section of the wave. The output of the clipper a pulse. The usefulness of this type of circuit w be discussed later when we consider how sca ning is controlled by synchronizing pulses.

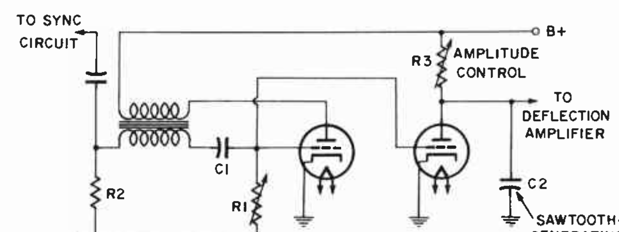


Fig. 6-13. Blocking oscillator and discharge tube as sawtooth generator.

These oscillators are biased to run in a fashion similar to Class-C transmitter technique, where plate current is cut off for part of the cycle. Part of the necessary clipping of the sine wave to produce a pulse is already done in the oscillator. This will be covered in greater detail when we discuss actual television deflection circuits.

A SUMMARY OF MULTIVIBRATORS AND BLOCKING OSCILLATORS

1. Multivibrator circuits employ vacuum tubes as electronic switches to control the charge and discharge cycles of a capacitor for the production of sawtooth waves.

2. The switching action by the tubes is practically instantaneous and is limited in timing by the circuit elements rather than by the tube itself. The velocity of electrons in a tube suddenly driven to saturation after plate-current cutoff can approach one-tenth the speed of light, or 18,600 miles per second.

3. When an electron tube acts as a voltage-operated switch, it suddenly changes from no plate current to where plate current is limited only by the ability of the cathode to supply electrons (plate-current saturation). These two conditions are illustrated in Fig. 6-5. Since this action is also found in other sections of a television receiver, you should become thoroughly familiar with the steps involved.

4. Multivibrator circuits may either be symmetrical or asymmetrical. Symmetrical circuits produce square voltage waves. The symmetrical multivibrator is used in the television transmitter to produce the complex television signal. It is not used in the receiver; for that reason, we are interested only in the asymmetrical type.

5. The asymmetrical multivibrator produces voltage output waves in which a short rectangular pulse is followed by a long "gap." This short pulse of voltage can be used to switch a vacuum tube from plate-current cutoff to plate-current saturation, and back again to cutoff.

6. If a capacitor of the correct value is connected across the plate of the output stage, the long periods between pulses will be occupied by the gradual charging cycle, and the voltage across the capacitor will rise linearly.

7. When the short rectangular pulse of voltage is suddenly applied to the grid of the tube, the tube becomes conductive and short-circuits the capacitor, and a new sawtooth wave is started.

8. An important reason for using this means of generating sawtooth waves is that dual-triode tubes, such as the 6SN7GT or 12AU7, can be employed instead of the two tubes shown in the illustrations.

9. Checking the values of resistors and capacitors, making sure that there are no open or short circuits, and substituting tubes constitute the normal service procedure in multivibrator scanning generators.

10. The blocking oscillator requires only one vacuum tube, contrasted with the multivibrator which depends upon a phase rotation of 360° through two tubes. The phase of feedback in the blocking oscillator is provided by the relationship of transformer windings.

11. The characteristics of the transformer determine the length of the conduction time or "closed switch" part of the cycle. This time is approximately one-half cycle of the transformer resonant frequency as tuned by its associated capacitances.

12. The "relaxed" time between "switch on" or conduction periods is determined by the time constant of the grid-circuit capacitor and its discharge resistors. The grid resistor can be made variable to control the frequency of oscillation and allow synchronization with the transmitted signal.

13. In the blocking oscillator, plate-to-grid feedback through the transformer sets up a grid current. This grid current charges the grid capacitor, and the grid is momentarily driven positive. The plate current rises to its saturation value and then falls. As the induced grid voltage reverses, the grid capacitor discharges through the grid resistor. The grid is driven very negative, and plate current is cut off. The cycle repeats as soon as the voltage across the resistor reaches a value which will allow the grid to again initiate plate current.

14. Discharge, or trigger tubes are frequently used with pulse generators, such as the blocking oscillator, to provide separate control of scanning-wave size and shape. Discharge tubes also provide a means of producing special waveshapes for particular scanning requirements. This subject will be covered in detail later in the text.

QUESTIONS

1. Name two gas-filled tubes that can be used in oscillator circuits.
2. How is feedback accomplished in a conventional multivibrator circuit?
3. What controls the bias of a conventional multivibrator circuit?
4. What are the distinguishing characteristics of an asymmetrical, or unbalanced, multivibrator?
5. By what two ways does the circuit of a cathode-coupled multivibrator differ from that of a conventional multivibrator?
6. When the second stage (V2) of the cathode-coupled multivibrator conducts, what action does the sawtooth-forming capacitor across the output perform?
7. In a blocking oscillator, what happens in the transformer when plate current is increasing?

8. What happens in the transformer when the plate current decreases?
9. What is the purpose of the trigger tube sometimes used with a blocking oscillator?
10. What type of waveform is produced by each of the following circuits in the absence of a sawtooth-forming capacitor?
    (a) Symmetrical multivibrator.
    (b) Asymmetrical multivibrator.
    (c) Blocking oscillator.

EXERCISES

1. Draw the circuit of a conventional multivibrator.
2. Break down the circuit in Exercise 1 and show the charge and discharge paths of the coupling capacitors.
3. Draw the circuit of a cathode-coupled multivibrator. Include the input capacitor and the sawtooth-forming capacitor.
4. Draw the basic circuit of a blocking oscillator.

Chapter 7

Sawtooth Generator Control and Production of Scanning Waveforms

At this time, let us examine in greater detail the sawtooth scanning waves which produce the raster. We have mentioned before that the horizontal and vertical sawtooth motion of the electron beam must keep in step with similar sawtooth scanning movements occurring at practically the same instant in the camera tube at the transmitter.

To accomplish this synchronization, pulses which control the horizontal and vertical scanning are transmitted in the television signal. These pulses occur between each horizontal frame. During the scanning of the frame, the receiver is "on its own." However, during the short interval between successive horizontal frames, the deflection circuits of the receiver are the absolute "slave" of the transmitter if the set is well designed, operating properly, and being used in an area of adequate field strength.

How the synchronized pulses are separated from the complex signal will be taken up later in this course. At this point, we will examine the relationship between the timing of these pulses and the control of the sawtooth scanning of the receiver. Fig. 7-1 shows the sequence of events during the scanning of one horizontal line and during the return of the electron beam to start the scanning of the next line.

The sawtooth line in Fig. 7-1 shows the desired linear trace and retrace motions of the electron beam in the TV picture tube. It does not necessarily depict the exact wave of current which must be passed through the deflection coils of the picture tube. As a matter of fact, we will find later that the deflection current must be distorted somewhat to accomplish the linear sweep and rapid flyback of the beam of electrons tracing the picture.

Fig. 7-1 shows, therefore, the *ideal* sawtooth for controlling the horizontal scanning motion in a television receiver. At point A, the electron beam starts to cross the face of the picture tube hori-

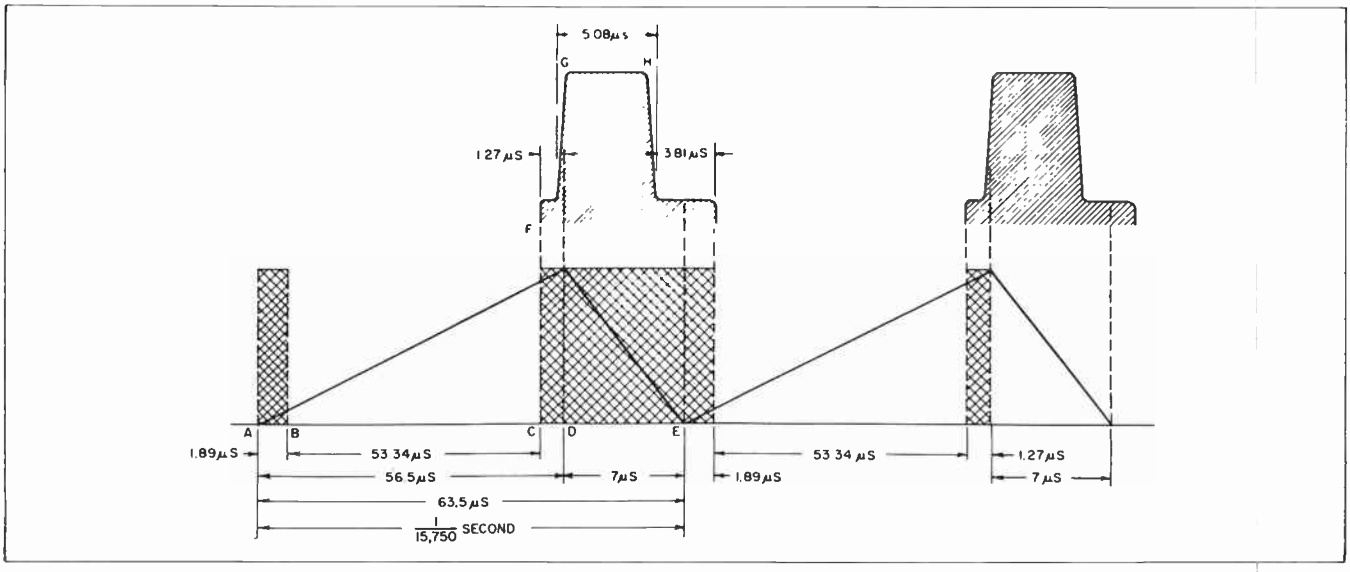


Fig. 7-1. Horizontal-scanning wave and synchronizing signal.

zontally from left to right. We will assume the picture tube is a 17-inch type and has an active picture width of 14 inches. The beam has blanked out from A to B, and the picture starts at point B. Between points B and C as the uniform motion progresses, the video modulation produces the picture.

As we have previously stated, the picture frame consists of 525 horizontal lines reproduced each 1/30th of a second (30 frames per second times 525 lines per frame equals 15,750 horizontal lines per second). This means the time for the trace of a line and its return to start another line is 1/15,750th of a second.

We should introduce the idea of talking about these extremely short time intervals in multiples of one millionth of a second. The unit of measurement is a microsecond. It is the length of time required for the completion of one cycle of carrier wave at the middle of the broadcast band, or 1,000 kc. The entire horizontal action, including the tracing of the picture line and the return to start a new line, occurs in 63.5 microseconds.

Let us divide the picture width (14 inches) by the time of active scanning (53.34 microseconds). We obtain a velocity of 4.1 miles per second. The retrace time (between points D and E in Fig. 7-1) is 7 microseconds. Since this retrace is over the same 14 inches or horizontal motion, the speed of the spot (blanked out to produce no light) must obviously be much faster. Actually, this retrace can reach a speed of 31.5 miles per second.

As we previously stated, the sequence of events must occur exactly in step with a similar sequence occurring at the same instant in the camera tube at the transmitter. To accomplish this action, pulses are sent out from the transmitter between each horizontal trace. The shape of these pulses is shown above the sawtooth wave in Fig. 7-1. At the instant shown as F, enough voltage appears at the grid of the picture tube to blank out all light. The region from F to G is known in television slang as the "front porch." This region is slightly more than one-millionth of a second in duration. At point G, the carrier wave of the transmitter abruptly increases by approximately 25% of its average value. This sharp rise in the carrier triggers the scanning generators in the receiver. The scanning generators produce the required sawtooth motion of the electron beam. Exactly how the pulse accomplishes this triggering will be described later.

The horizontal beam does not trace a line parallel with the top of the picture, but has a slight downward slope. This vertical motion is controlled by a scanning sawtooth which moves the scanning spot to the bottom of the image and then rapidly returns it to the top. The electron beam moves from the top to the bottom of the picture and back to the top in 1/60 of a second. It is easy to see that this vertical scanning is much slower than the horizontal line tracing action and requires 16,666 microseconds. Pulses are sent out between successive fields to lock in, or control as a slave, the vertical-scanning oscillator of the receiver. A cycle of the vertical-deflection sawtooth wave, together with an enlarged section of that part of the wave which occurs during blanking and retrace, is shown in Fig. 7-2.

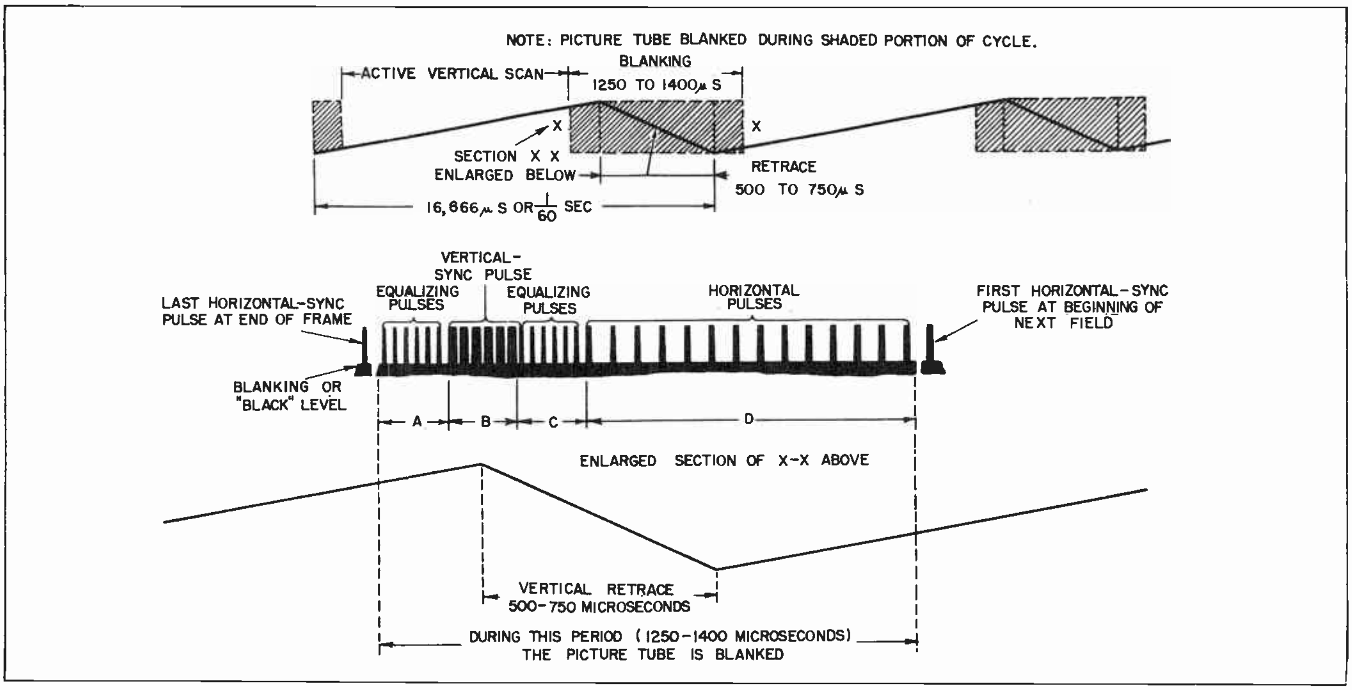


Fig. 7-2. Vertical-scanning wave and synchronizing signal.

The portion of the television signal which controls vertical retrace and synchronization is much more complicated than the single horizontal pulses which occur between successive horizontal lines. The vertical synchronizing signal resembles a comb with uneven teeth. If its only function were to trigger the vertical oscillator and to blank out the picture-tube screen during retrace, it could be made in the form of a single long rectangular pulse whose time duration would be from 20 to 22 horizontal lines (1,250-1,400 microseconds). However, the vertical synchronizing signal must perform two other functions. It must continue to keep the horizontal-scanning oscillator in step during vertical retrace and also assure that alternate fields have proper interlace of the horizontal lines.

Horizontal synchronization is kept in step by notches B and pulses A, C, and D (Fig. 7-2). Interlace is controlled by equalizing pulses A and C (Fig. 7-2) preceding and following the vertical sync pulse.

We are not concerned at this time with the exact composition of the complex wave making up the television signal because this subject is covered in Chapter 9.

CONTROL OF SCANNING GENERATORS BY SYNC PULSES

We have seen that the scanning systems of the receiver must keep in accurate step with the scanning raster of the camera tube at the transmitter. We have also described the type of synchronization pulses which are made a part of the television signal to satisfy this requirement.

For a satisfactory reproduced picture, the picture elements of adjacent horizontal traces must line up accurately, and the lines of alternate fields must interlace or space accurately between each other.

To avoid a displacement of more than one picture element in successive horizontal lines, the frequency stability of the horizontal oscillator must be 0.2% or better. Fig. 7-3A illustrates horizontal displacement.

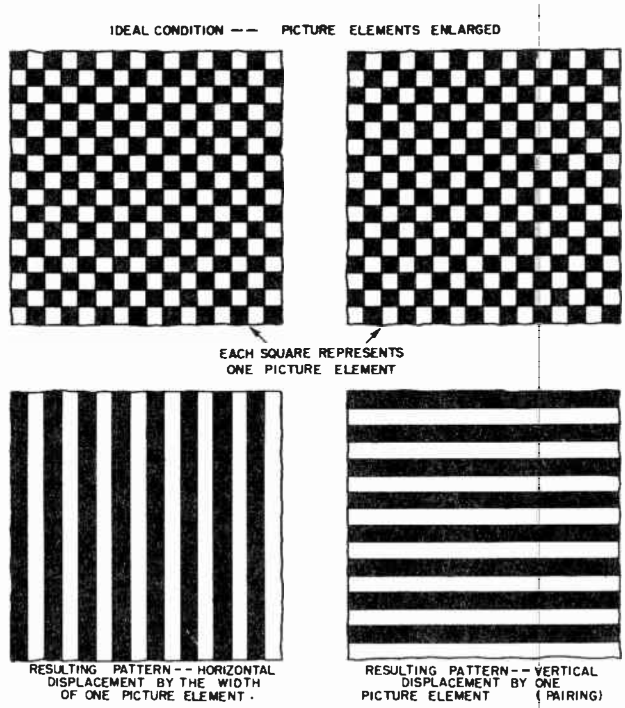
To avoid "pairing" of the lines of successive fields (the line lying on top of those of the preceding field instead of being properly interlaced), the stability of the vertical oscillator must be better than 0.05%. Fig. 7-3B illustrates this displacement.

In each of the impulse-generating circuits (cathode-coupled multivibrator and blocking oscillator) suitable for television scanning, coupling means have been indicated in the grid circuits for the introduction of synchronizing pulse controls (see Figs. 6-9, 6-11, and 6-13).

The horizontal and vertical pulses are clipped from the signal, amplified, and passed through circuits which classify the pulses so that each will control its own scanning oscillator only. How these operations are accomplished will be described later. The end result is a short, sharp "pip" for the horizontal control and a long triangularly-shaped pulse for the vertical control.

In considering how the pulse controls the oscillator frequency, three items are important:

1. The free-running frequency of the sweep generator. — This frequency is that which would be generated at any particular setting of the hold control if the sync pulses were not present. The frequency can be slower or faster than the pulse repetition rate, or in exact step. We will show later that, for proper stable operation, the slow rate is required.



(A) Horizontal displacement.

(B) Vertical displacement.

Fig. 7-3. Picture-element displacement which might result from scanning-oscillator instability.

2. The firing point of the sweep generator. — This is the grid-bias voltage required for the controlled tube in order to initiate conduction in the discharge tube, and to start capacitor discharge and scanning-wave retrace. At this point in the cycle, the oscillator is most sensitive to control by the sync pulse.

3. The synchronizing frequency. — This is the rate at which the pulses are applied to the control-input terminal of the oscillator—60 cps for vertical synchronization and 15,750 cps for horizontal synchronization.

Since the control action of the blocking oscillator can be more readily diagrammed, we will consider it first.

Pulse Control of the Vertical Blocking Oscillator

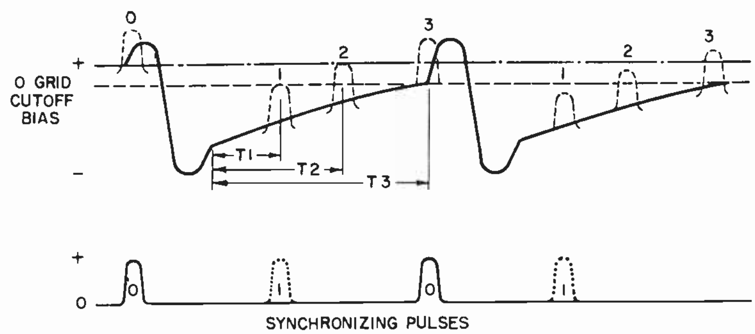
As we mentioned in the preceding chapter, the triggering, or firing, of the blocking oscillator occurs when the grid voltage passes the cutoff point. The free-running frequency of the oscillator (if no sync pulses are present) is determined solely by the time constant of the grid capacitance and resistance. Once the oscillator is fired, it functions on its own until it is blocked again by the grid cutoff voltage.

If a pulse of positive potential from an external source is fed to the grid while the grid capacitor is discharging through the resistor, the grid voltage will pass the cutoff point, and the tube will begin to conduct. The sawtooth-forming capacitor will discharge, and retrace will occur. A new scanning cycle then begins. Therefore, the repetition of the positive sync pulses can control the firing of the blocking oscillator and lock the picture into synchronization.

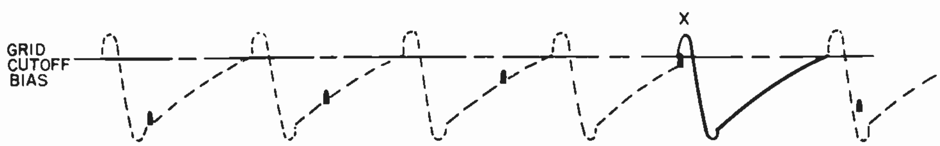
Fig. 7-4 illustrates this action in detail. Fig. 7-4A is an enlarged portion of the blocking-oscillator grid voltage. The synchronizing pulses below the grid waveform show a series of pulses marked "O" whose leading edges are exactly in step with the wave. These pulses do not affect the free-running frequency of the oscillator. They merely add to the grid voltage at the same instant it is being driven positive by the plate-current pulse. On the other hand, if the sync pulses occur at the points indicated as "1", the pulse voltage added to the discharge voltage of C1 through R1 (Fig. 6-11) is still short of the cutoff bias point and will not fire the tube. However, if the pulse occurs at points 2 or 3, the critical bias will be exceeded, the tube will immediately conduct, and retrace will begin.

Since the free-running frequency of the blocking oscillator can be changed by varying the time constant of the capacitance and resistance in the grid circuit, let's examine this action under the following conditions: (1) oscillator running faster than the sync-pulse rate, and (2) oscillator running slower than the sync-pulse rate.

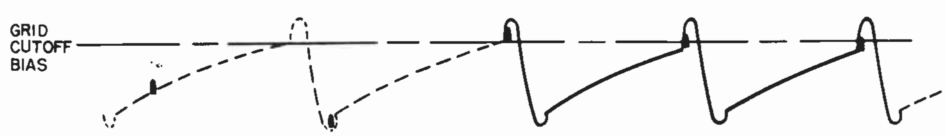
Fig. 7-4B shows what happens when the oscillator is running faster than the sync-pulse rate. The dotted portion of the waveform indicates lack of synchronization at that point. Notice that several cycles occur before the pulse reaches point "X". At this point the grid cutoff voltage is exceeded, and the tube fires. Normally, you would expect the picture to lock in satisfactorily. However, this is not true. Lock-in occurs only momentarily during the field initiated at point "X". Succeeding fields do not lock in. With the oscillator running in this fast condition, the sync pulses are occurring during the scanning interval. Consequently, the picture is divided by the blanking bar. Also, the oscillator, running faster than the sync-pulse rate, can easily be triggered into erratic operation by automobile ignition and static interference. Therefore, the picture will be unstable. In modern receivers, however, improved



(A) Grid-voltage waveform with a series of sync pulses.



(B) Oscillator frequency fast.



(C) Oscillator frequency slow.

Fig. 7-4. Pulse control of a free-running blocking oscillator.

circuit designs have resulted in more stable pictures, even under heavy interference conditions.

Fig. 7-4C shows what happens when the free-running frequency of the blocking oscillator is slower than the sync-pulse rate. Notice that lock-in occurs much faster and a good stable picture is obtained. This is obviously the desired condition because the oscillator should run slightly slower than the sync-pulse rate. The sync pulses will then take over and force the blocking oscillator to lock in with each succeeding sync pulse.

The height and width of the sync pulses are not important as long as they are high enough to drive the grid above the cutoff point. With the sync-pulse amplitude reasonably high, the hold control can be varied over a fairly wide range without loss of synchronization.

For control of the blocking oscillator just described, the sync pulses are positive. When we study pulse clipping and amplification, we will find that a sync pulse can be made either positive or negative with respect to ground (or chassis), depending on receiver design. Negative pulses are ideal for the cathode-coupled multivibrator.

Pulse Control of the Cathode-Coupled Multivibrator

From Fig. 6-10 and the text, we know that "tripping," or discharge, of the cathode-coupled multivibrator is initiated by a negative-voltage pulse on the grid of the first tube. Once "tripping" begins, it immediately receives an additional negative voltage from the cathode-bias resistor, which is common to both tubes. Although the control actions and principles just described for the blocking oscillator hold true, it is not feasible to show them in diagram form. As a matter of fact, the small step in the leading edge of the grid-voltage curve of Fig. 6-10A (indicating the sync-pulse contribution to the grid voltage) is really a matter of "poetic license" since the action is so rapid and so cumulative that it is impossible to tell where pulse control stops and the circuit takes over.

Summary of Scanning-Generator Pulse Control

1. A positive synchronizing pulse controls the frequency of a blocking oscillator.
2. A negative pulse controls the frequency of a cathode-coupled multivibrator.
3. The free-running frequency of the scanning oscillator should always be slightly less than the synchronizing-pulse repetition rate. The hold, or frequency, control of the oscillator controls this action.
4. As the grid voltage of a pulse generator approaches the "trigger" point, the oscillator becomes increasingly sensitive to control by additional grid voltage. At this point, the scanning can be "tripped" by interference. Special circuit combinations have been devised which are controlled by the pattern of the pulses, rather than by the individual pulses. Such a system is relatively insensitive to interference, which seldom has a regular pattern. Its operation will be described later in the course when specific circuits are covered.

REQUIREMENTS FOR ELECTROSTATIC PICTURE TUBES

In an electrostatic tube, the desired raster can be produced by applying sawtooth voltage waves of the shape and time requirements shown in Figs. 7-1 and 7-2 to the horizontal- and vertical-deflecting plates.

The once popular 7-inch Type 7JP4 required deflection voltages of about 200 volts per inch to move the beam. The picture size on this tube was 4 × 5½ inches. The horizontal-deflecting plates required a peak-to-peak voltage of the sawtooth wave of 5.5 times 200, or 1,100 volts.

The sawtooth generators described usually cannot directly produce such voltages. For this reason, amplifiers were needed between the scanning-generator circuit and the picture-tube deflecting plates. As we learned in Chapter 1, the deflection and centering circuits of the tube were the balanced or push-pull type. Therefore, the deflection amplifier feeding them was a push-pull amplifier, usually a phase inverter. Typical electrostatic-deflection systems will be covered in detail later.

SCANNING REQUIREMENTS FOR ELECTROMAGNETIC PICTURE TUBES

We have seen that electrostatically-deflected tubes required only an amplified sawtooth wave of volt-

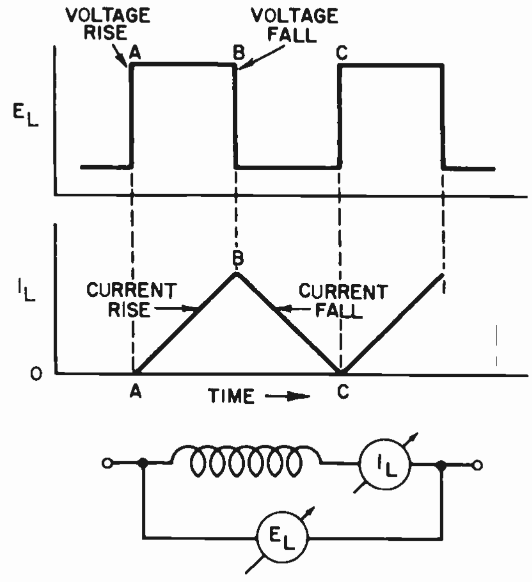


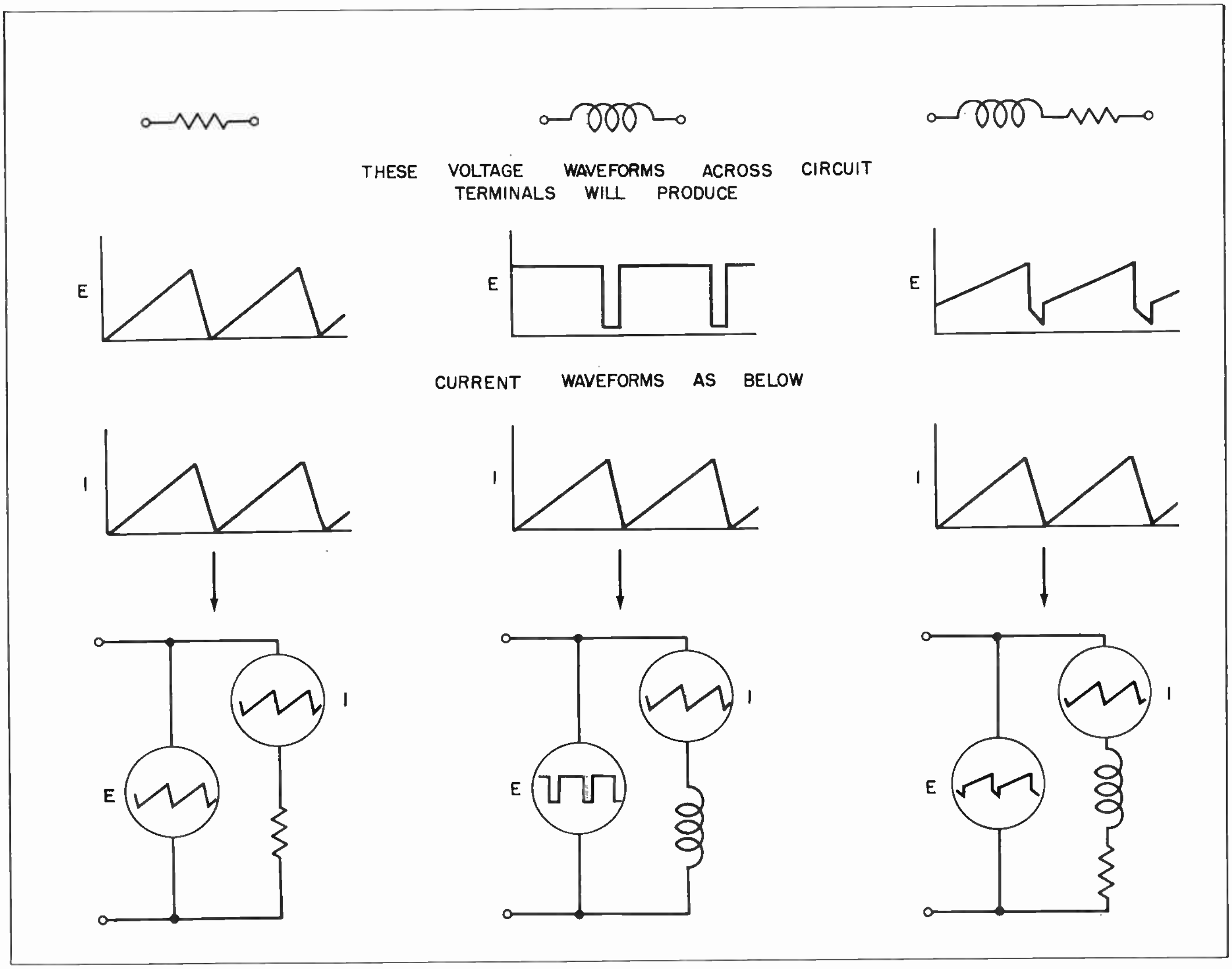
Fig. 7-5. Rise and fall of current through a pure inductance when a square voltage wave is applied.

age to produce the desired pattern or raster. Magnetically deflected tubes, however, impose a different set of requirements because of the deflecting coils.

Chapter 2 explained the theory and mechanical arrangement of the horizontal- and vertical-deflecting coils. It stated that a sawtooth wave of current through the coils can be made to produce the desired raster. In other words, the amount an electron beam in an electromagnetically-deflected cathode-ray tube is deflected depends upon the strength of the magnetic field produced by the deflecting coils. The magnetic field is proportional to the amount of current passing through the coils, and these fields cross the path of the electron beam within the neck of the tube.

We must supply a linear sawtooth of current through the coil so that the electron beam will trace the proper raster under the combined influence of the horizontal- and vertical-deflecting coils. In Fig. 7-5 we see the resultant shape of a current wave which would flow through a pure inductance if a symmetrical square wave of voltage were applied across its terminals. This type of wave, as we have seen, can be developed by a conventional or symmetrical multivibrator. At point A, the voltage has suddenly been applied to the coil in much the same fashion as if a switch had been closed to connect the coil to a DC source of potential, for example, a battery.

Notice that the current through the coil did not immediately rise to maximum. The self-induced voltage of the coil opposed the sudden change. The current, therefore, increased linearly over that portion of the cycle when the applied voltage was steady. (Theoretically, the current rises exponentially; but for practical purposes, we can consider it to be a linear change.) At point B, the impressed



(A) Pure resistance. (B) Pure inductance. (C) Series inductance and resistance.

Fig. 7-6. Voltage and current waveforms in inductance and resistance circuits.

voltage was suddenly removed (the switch was opened). At this point, the current did not immediately fall to zero since it was maintained by the energy stored in the magnetic field. The self-induced voltage of the coil served as the driving potential to produce the linear fall of current from point B to point C.

We have now produced a triangular wave of current through the coil. If we can make the rise portion of the curve longer than the decay portion, we can produce the desired sawtooth-scanning current wave. This wave can be produced by making the impressed voltage wave asymmetrical, as shown in Fig. 7-6B.

Since a deflection coil cannot be built as a pure inductance, we must now consider what effect the resistance of the windings will have on the voltage waveform producing a sawtooth of current.

Fig. 7-6 illustrates three types of circuits and the voltage waveform necessary to produce a sawtooth wave of current through each circuit. Fig. 7-6A shows a pure resistance. The current is in phase with the voltage, and a sawtooth wave of voltage impressed across the resistor will cause a sawtooth wave of current through it. Energy losses occur only in the form of heat. The voltage required to produce a certain current is equal to the IR drop, as determined by Ohm's law.

Fig. 7-6C shows the circuit represented by a deflection coil. The voltage waveform will be seen as a combination of the sawtooth of A and the rectangular wave of B (often called a trapezoidal waveform). In reality, this shape is the sum of an instantaneous pulse and a sawtooth. We might think of its function as follows:

1. The sawtooth or linear rise portion of the wave tends to produce a sawtooth wave of current through the resistive part of the circuit.

2. The instantaneous pulse portion of the wave forces a sawtooth wave of current through the inductive part of the circuit.

To produce this combination waveshape, additional circuit elements are added to the sawtooth-capacitor charging circuit. The circuit is then known as a peaking type of waveshaping circuit. By proper choice of capacitor and resistor values, either the sawtooth portion or the impulse portion of the wave can be made to predominate. The circuit action will be described later.

It is interesting to note that one part of the wave must predominate over the other because of the differences between the horizontal- and vertical-deflection coils. In the vertical-deflecting coil, the resistive component predominates over the inductive component. Thus, the sawtooth portion of the wave predominates over the impulse portion. For example, this coil might have a resistance of 68 ohms and an inductance of 50 millihenries. When the retrace rate is 60 cycles, a predominantly resistive circuit is presented.

In the horizontal-deflecting coil of the same receiver, the conditions are reversed; the inductive component predominates. The impulse portion is more important, and the required waveshape approaches that of Fig. 7-6B. For example, we would find a resistance of only 14 ohms and an inductance of 8 millihenries. Since this coil operates at the much higher frequency of 15,750 cycles per second, the circuit is essentially inductive.

Fig. 7-7 represents a comparison, in block-diagram form, between the basic elements of an electrostatic and an electromagnetic scanning circuit.

PEAKING CIRCUITS FOR ELECTROMAGNETIC DEFLECTION

By a simple change in the discharge-tube circuit (Fig. 6-13), the combination sawtooth and impulse wave required for electromagnetically-deflected scanning can be generated. The modified circuit is shown in Fig. 7-8. Resistor R2 has been added in series with the discharge capacitor C2. The circuit action will be described in sequence:

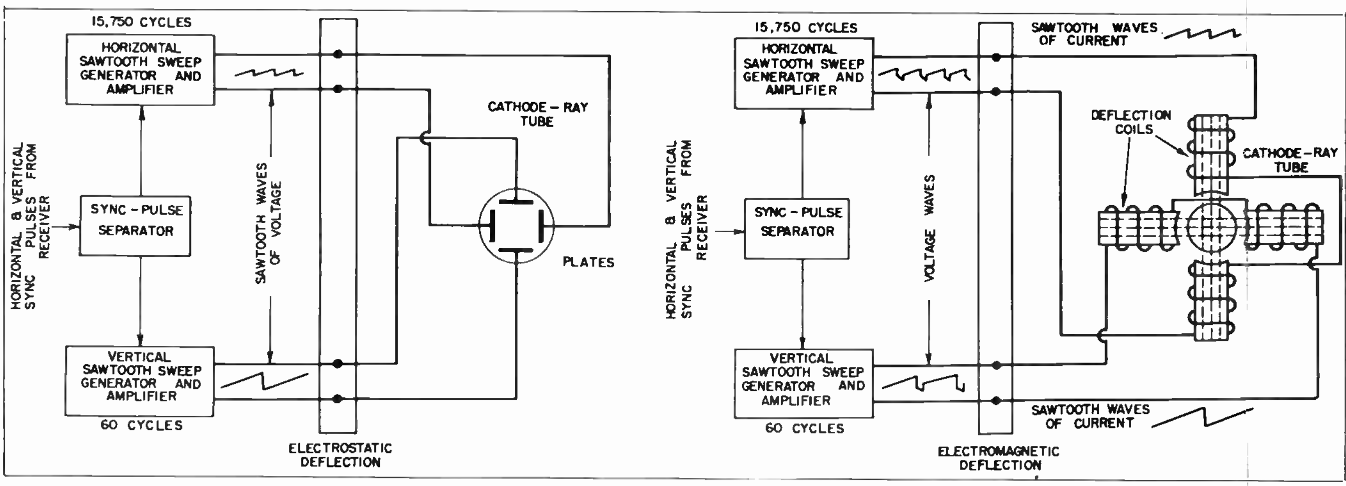


Fig. 7-7. Electrostatic and electromagnetic scanning systems.

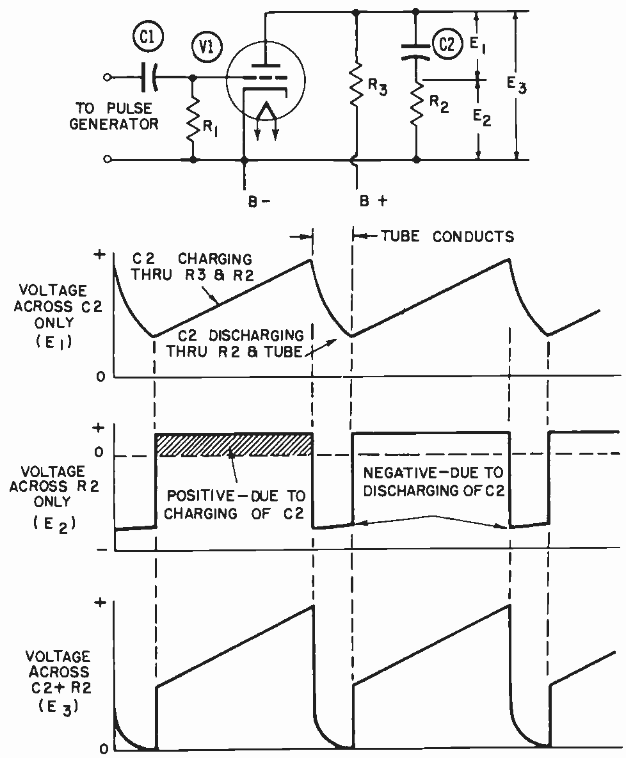


Fig. 7-8. Typical peaking circuit and associated waveforms.

1. The sawtooth-forming capacitor C2 is charged from the B+ source through resistors R3 and R2. This charging action takes place when the tube is not conducting.

2. The output-voltage waveform of the circuit is taken across the series combination of R2 and C2. R2 is known as the peaking resistor.

3. During the charging portion of the cycle, the voltage across the capacitor is a sawtooth wave.

4. When the tube conducts because of a positive pulse on its grid, the voltage across C2 and R2 is shunted by the low plate resistance of the tube.

5. The voltage across the capacitor cannot change instantly because its discharge path through R2 and through the plate resistance of the tube is not zero. Therefore, the difference in voltage must suddenly appear across peaking resistor R2. After this initial sudden change of voltage, the capacitor discharges exponentially through R2 and the tube until the tube again becomes nonconductive.

6. As the tube is cut off, the B+ potential is applied to the capacitor through R2 and R3 in series. Again, the capacitor voltage cannot rise instantaneously. The voltage across R2 must once more change abruptly, after which the capacitor charges through R2 and R3 in its normal sawtooth fashion.

By changing the values of R2 and C2, the ratio of the amplitude of the peaking impulse to the sawtooth can be adjusted to match the inductive and resistive requirements of the particular deflecting coil.

QUESTIONS

1. How long does the beam take to move from the top to the bottom of the picture and back to the top again?
2. How many horizontal lines are produced per second?
3. What are the functions of the vertical-synchronizing signal?
4. Which one of the following pulses controls interlace?
   (a) Vertical-sync pulses.
   (b) Horizontal-sync pulses.
   (c) Equalizing pulses.
   (d) Horizontal pulses.
5. What is the polarity of the synchronizing pulse that controls the frequency of a blocking oscillator? Of a cathode-coupled multivibrator?
6. What is the current waveform passed through the deflection coils? What is the voltage waveform applied to the coils?
7. How is the output circuit of a sawtooth generator changed in order to obtain the voltage waveform needed for electromagnetic deflection? What is the circuit called?

EXERCISES

1. Draw the current and voltage waveforms for the following.
   (a) Pure inductive circuit.
   (b) A series resistive and inductive circuit.
2. Draw the output stage of a sawtooth generator that is used to provide sweep voltage. Sketch the voltage waveform across each component of the peaking circuit and the combined waveform across the circuit.

Chapter 8

Deflection Systems – Commercial Applications

FOCUSING AND CENTERING CIRCUITS FOR ELECTROMAGNETIC DEFLECTION

Chapter 1 described the circuits for beam centering in the electrostatic tube; Chapter 2, focusing and deflection of the beam in an electromagnetic tube. The circuits that control the focusing and centering of electromagnetic tubes were not shown, since they are related to the action of sweep and deflection circuits and to the voltage waveform requirements of electromagnetic deflection.

Fig. 8-1 shows a schematic of the picture-tube control circuits in a once popular type of receiver. This type of circuit is no longer used, but will be discussed in order to illustrate some of its functions.

Focus Control

In Chapter 2, focus coil L1 (Fig. 8-1) was described and illustrated. The amount of current through the coil is determined by the setting of variable resistor R1, and the electron-beam paths are brought together to form a small concentrated spot at the surface of the fluorescent screen. In the circuit illustrated, a high-current coil of relatively low resistance (250 ohms) has been used. This coil and the ion-trap electromagnet coils are in the negative-return lead of the power-supply system and carry a large portion of the total B+ current. In this example, the average current through the coil is approximately 115 milliamps. Other receivers may use focus coils having more turns of finer wire, hence higher resistance, in other parts of the supply circuit. Although the circuit shows R1 as a single variable resistor to achieve smoother control action, this control is often part of a network of series and parallel resistors.

The focus coil is so mounted that its axis can be rotated relative to the axis of the picture tube. We will see later that this action either centered the picture on the fluorescent screen or (in some receivers) provided a fixed control that governed the range of action of the back-mounted centering controls. The electro-magnetic type of focus control has been replaced in modern receivers by the permanent-magnet type discussed in Chapter 2.

Ion Trap

Ion-trap action was described and illustrated in Chapter 2. In the circuit of Fig. 8-1, ion-trap magnet coils L2 are shunted by resistor R2. This shunt is used because the ion-coil current (approximately 105 milliamps) for proper operation is less than the focus-coil current. The electromagnetic type of ion trap, as we stated in Chapter 2, has been replaced by the permanent-magnet type, or has been eliminated altogether.

Vertical Centering

The vertical-centering circuit (shown in simplified form in inset A, Fig. 8-1) consists of a series arrangement of vertical-deflection coils L3, secondary L4 of the vertical-output transformer, and a source of DC voltage from potentiometer R5. R5 has a center tap; as the contact arm passes this center point, the DC voltage introduced into the series circuit changes polarity. In this way, a DC polarizing current can be produced through the deflection coils in either direction, and a steady magnetic-field bias will exist in the deflection system. The sawtooth scanning current operates around this bias as a center point; and by adjustment of R5, the picture can be moved up or down.

Horizontal Centering

Horizontal-centering circuits in prewar receivers were usually the same type just described for vertical centering (a center-tapped wirewound control carrying high current in the B+ return).

Inset B, Fig. 8-1, shows in simplified form the portion of the large diagram which governs horizontal centering. As before, we find a series circuit consisting of horizontal-deflection coils L5, secondary L6 of the horizontal-output transformer with its shunt width-control coil L7, and centering control R6. In this circuit, however, the DC bias current through the horizontal-deflection coils cannot be made to reverse direction, as was possible with the vertical-centering system.

To obtain the equivalent of control in both directions, the focus coil, which exerts a magnetic-field bias on the electron beam, must be positioned properly. During initial receiver installation, the following steps are taken:

1. The horizontal-centering control, potentiometer R6, is first set at its midrange.
2. The position of the focus coil is adjusted in its mounting to center the raster.
3. Adjustment of R6 will now slightly correct the picture horizontally (right or left). The focus-coil position has provided a magnetic-field bias equivalent to the introduction of an opposing current to that flowing in the series-deflection circuit because of the initial setting of potentiometer R6.

The latest receivers dispense entirely with the potentiometer-type centering controls, as explained in Chapter 2.

DAMPING CIRCUITS

The horizontal-output circuit of Fig. 8-1 can be recognized as the high-voltage supply system described in Chapter 4, which utilizes the collapse of magnetic energy in the horizontal-output transformer and horizontal-deflection yoke. No explanation was given at that point of the action of the 5V4G damping or reaction scanning tube. Its function, as well as the action of horizontal-linearity control L8 and horizontal-width control L7, will now be considered. This tube and its associated circuit components stop, or damp out, oscillations in the system and help produce the required linear current sawtooth through the deflection coils.

We pointed out in Chapter 7 that horizontal retrace must be accomplished in the extremely short time of 7 microseconds. Because the horizontal-

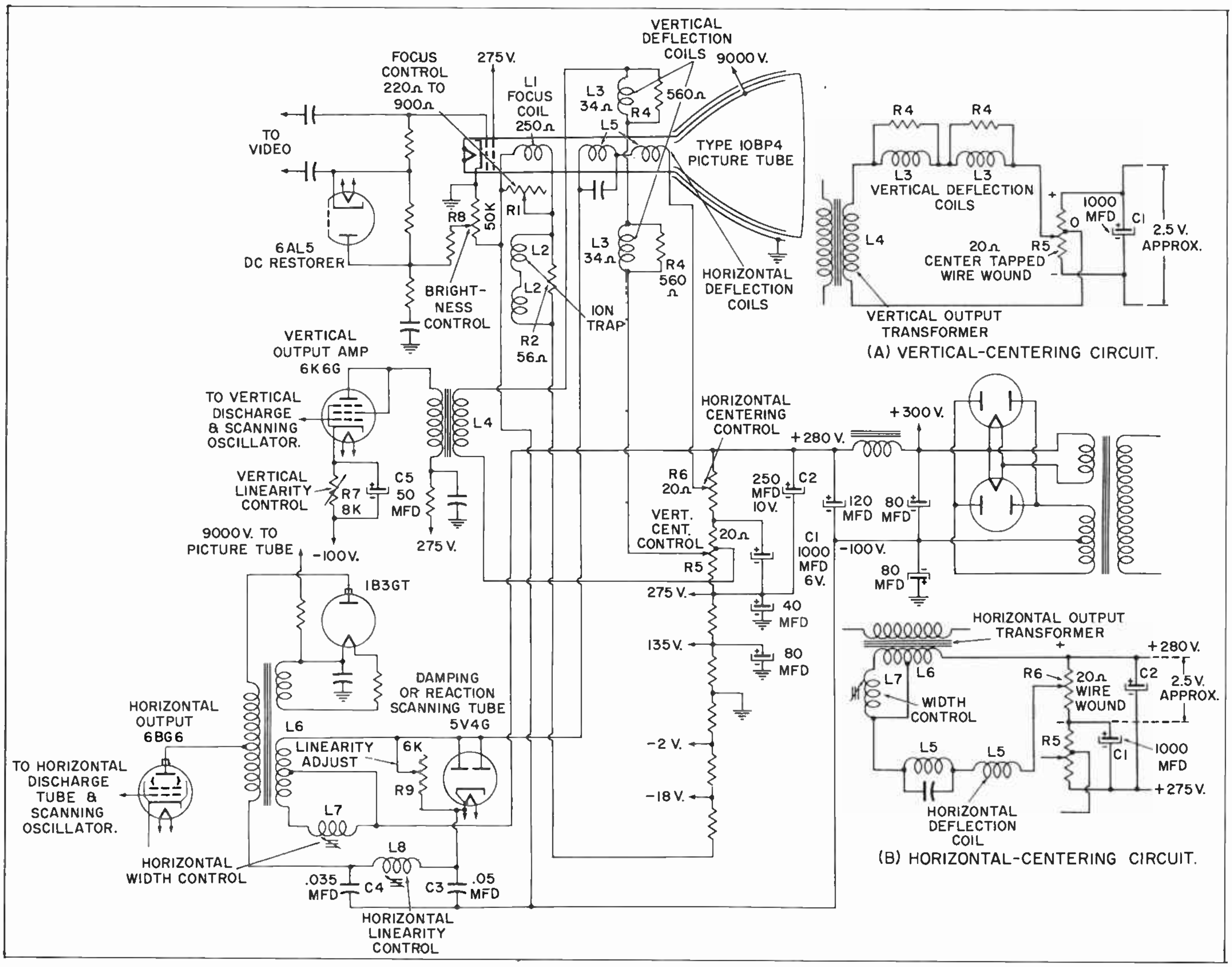


Fig. 8-1. An early type of electromagnetic-deflection system.

deflection coil system is predominately inductive, a different method of operation from that of the lower-frequency vertical system must be employed. To obtain the rapid reversal of current through the horizontal-deflection coils, the output transformer and deflection coil circuit are tuned to a frequency of approximately 71 kilocycles by the associated circuit capacitances. This frequency is used because one-half cycle of oscillation is equal to the required retrace time of 7 microseconds.

The current through the deflection coils is maximum at either the extreme left or extreme right of the picture frame, with the axis (zero point) at the center. When the right-hand end of the trace is reached, the horizontal-output tube is conducting high plate current, and a maximum of magnetic energy is stored in the deflection coils. At this instant, a negative pulse arrives at the grid of the horizontal-output tube from the plate circuit of the horizontal-discharge tube, and the output-tube plate current is suddenly cut off. The magnetic field in the transformer and in the deflection coils starts to collapse at a rate determined by the resonant frequency of the system (71 kc). This collapse will shock-excite the circuit into damped oscillation which, if allowed to continue, would produce the waveform of current shown in Fig. 8-2.

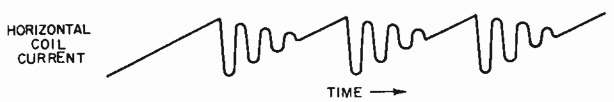


Fig. 8-2. Oscillation of horizontal-deflection coil current if damping tube were not used.

Continuation of this oscillation would seriously distort the left-hand side of the picture. An absorption, or damping, device is used to kill the oscillation during the 10 microseconds the picture tube is blanked. This function is provided in the circuit of Fig. 8-1 by the 5V4G damping or reaction scanning tube.

As the output-tube plate current is cut off by the negative scanning pulse, the induced voltage caused by the collapsing magnetic field immediately becomes negative at the plate of the damping tube. The tube will not conduct, and no load will be imposed upon the circuit. Therefore, it oscillates for one-half cycle at its resonant frequency of 71 kc (approximately 7 microseconds). This oscillation causes the current through the deflection coils to reverse to a maximum in the other direction and accomplishes the rapid return trace. The circuit would continue to oscillate for many cycles if it were not for the damping tube, which acts during the second half-cycle.

As the current reverses through the deflection coils to start the second half-cycle of oscillation, the self-induced voltage also reverses; and the polarity becomes positive at the plate of the damping tube. This tube starts conducting and acts as a load across the terminals of the deflection-coil system, and any further tendency toward oscillation is prevented. The current through the tube decays at a linear rate determined by the circuit constants. This linear current starts the next active scanning wave for the visible, or unblanked trace.

If no additional current is supplied to the circuit from the horizontal-output tube, the electron beam stops at the center of the picture tube as the current through the damping tube decreases to zero. The latter portion of the decay current wave departs from the desired linear sawtooth form. To overcome this nonlinearity, conduction of the horizontal-output tube is so timed by the scanning generator and discharge tube that it starts contributing current to the deflection coils before the original current has completely decayed.

This current contribution from the horizontal-output tube, which deflects the beam from the center to the right-hand side of the screen, is so shaped at its start that it corrects the nonlinearity of the original decay-current wave. As shown in Fig. 8-3, the two current waves overlap at the

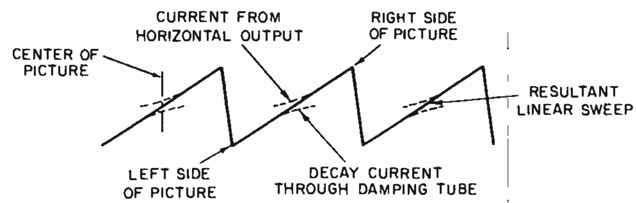


Fig. 8-3. Current waveforms in horizontal scanning.

center of the scanning, and the combination produces a linear coil current.

To summarize the circuit action to this point:

1. The damping tube allows one-half cycle of natural resonant oscillation to occur in the deflection circuit, after which it loads the circuit and prevents further oscillation.

2. The first half-cycle of oscillation accomplishes beam retrace in the required 7 microseconds.

3. Decay of current through the damping tube produces the first half of the 53-microsecond active trace.

4. The horizontal-output tube starts to contribute power to the system before final decay of the damping-tube current and produces the final half of the active scanning wave.

5. At the end of the active scanning cycle, a negative pulse causes plate-current cutoff in the output tube and starts a new oscillation; and the cycle is repeated.

Three adjustable circuit components are associated with the damping tube in Fig. 8-1. These are the horizontal-linearity control L8, the horizontal width control L7, and the linearity-adjustment resistor R9.

Horizontal-Linearity Control

The network comprised of L8, C3, and C4, performs two functions in this circuit. It provides a means of operating the horizontal-output tube at a higher plate voltage than that of the power supply. At the same time, it is an adjustable means of making small corrections in the shape of the active sweep section of the sawtooth current wave.

Since these two functions are interrelated, an explanation of the voltage-addition action will precede the description of the control operation. Notice that the plate voltage of the horizontal-output tube is supplied by a series circuit consisting of the primary of the output transformer, inductor L8, and the cathode-to-plate conducting path of the 5V4G damping tube. The damping tube is conducting over most of the sawtooth, as shown in Fig. 8-3. Capacitors C3 and C4 are charged during this conduction period. They then discharge while the damping tube is not conducting. Thus, the plate current of the horizontal-output tube is maintained over this portion of the cycle when the damping tube is not conducting.

The charge on these capacitors is greater than the B+ supply voltage because the rectified surge, or "kickback", of the deflection coils is added to the charge placed upon them by the B+ supply. In a typical receiver employing this circuit, the output-tube plate voltage is approximately 50 volts higher than the B+ voltage.

Capacitors C3 and C4 do not seem to be high enough in value to act as storage capacitors. This fact gives a clue to the action of the circuit as a linearity control. During the first half of the trace period, the voltage across C3 rises because of the rectified deflection-coil "kickback". During the second half of the period, the voltage falls because of the current demand of the horizontal-output tube (which is conducting at that time). Thus, a ripple voltage with the same frequency as the sawtooth is impressed on the plate supply of the horizontal-output tube. If the phase of this ripple is shifted with respect to the plate-current requirements of the output tube, the shape of the current wave from the output tube can be modified. Changing the value of inductor L8, by adjustment of a powdered iron core, controls the phase of the ripple voltage and permits minor adjustments of horizontal linearity. Other controls which affect horizontal linearity are the horizontal-drive control (to be described later), linearity adjustment R9, and horizontal-width control L7. These controls are interdependent; and when one is adjusted, the others must often be readjusted.

Linearity-Adjustment Resistor

Resistor R9 is known as a damping resistor. Rather than being made continuously variable, it has a series of taps. This is primarily a factory adjustment to compensate for manufacturing variations in the deflection yoke and output transformer. Resistor R9 controls the linearity of the trace on the left-hand side of the picture only. It is no longer used in present-day receivers.

Horizontal-Width Control

Variable inductor L7 shunts a portion of the secondary of the horizontal-output transformer L6. This action controls the output voltage and, hence, the width of the sweep. Since the horizontal-width control also has a minor effect on the phase of the

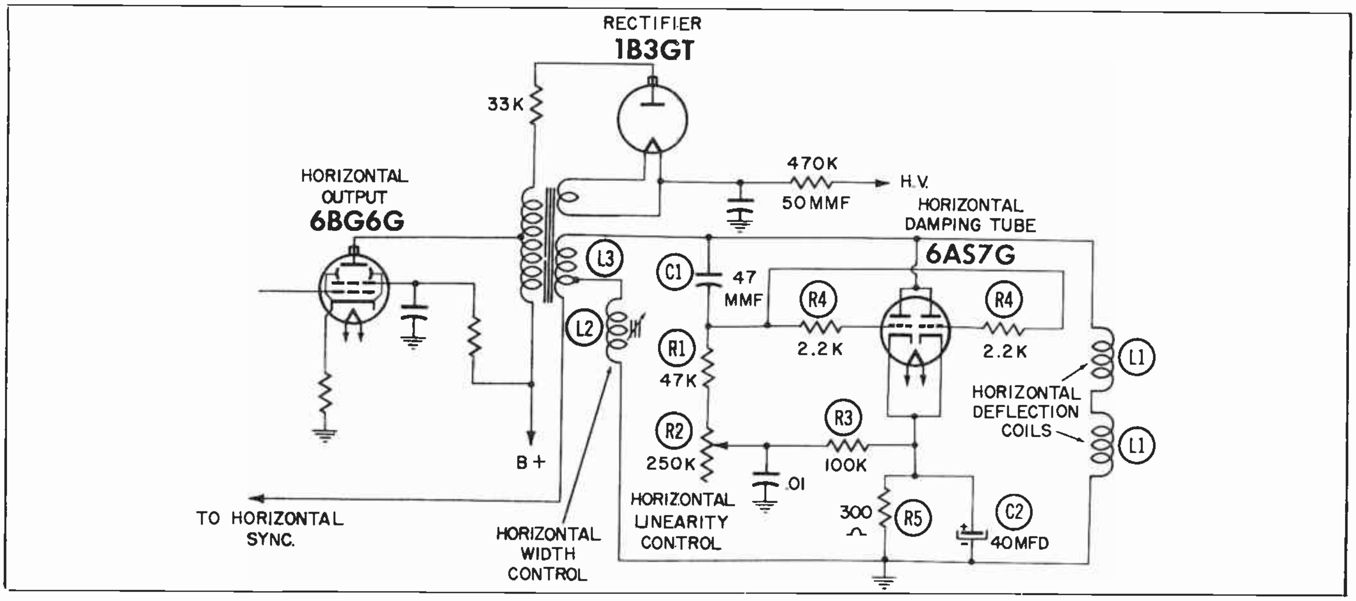


Fig. 8-4. Use of triode as horizontal-damping tube.

plate voltage of the output tube, it causes slight changes in the linearity of the right-hand side of the picture as the width is being changed.

Horizontal-Drive Control

The horizontal-drive control is a variable resistor which is the pulse-forming element in the plate circuit of the horizontal-discharge tube. An explanation of the horizontal-drive control appears in Chapter 7. Fig. 4-4 shows its application to the circuit under discussion. The value of this resistor determines the ratio of the negative pulse amplitude to the sawtooth amplitude impressed on the grid of the horizontal-output tube. The resistor can, therefore, control the point on the trace at which the output tube conducts. Increasing its value widens the picture, crowds the right side, and stretches the left side. If the horizontal-drive control is adjusted, the horizontal-width control may have to be reset.

Using the Triode for Horizontal Damping

In the circuit just discussed, a power rectifier type of diode (5V4G) was used to damp the oscillation produced in the horizontal-deflection coil system by cutoff of output-tube plate current. In some receivers, a triode was used as the damping tube, since the presence of a control grid gave the designer another opportunity to modify circuit action for better sweep linearity.

Fig. 8-4 shows the 6AS7G, a low-mu dual-power triode, used as a damping tube. For a tube to act effectively as a damping tube, its plate resistance must be extremely low during conduction. The 6AS7G, designed especially for voltage regulators and television, met this requirement.

At the instant of plate-current cutoff, the circuit conditions in the horizontal-output tube are similar to those described for the 5V4G. The natural resonant frequency of the deflection circuit is approximately 71 kc. The first half-cycle of oscillation finds the plate of the damping tube negative, and no plate conduction can occur. During this time (7 microseconds), horizontal retrace takes place.

On the next half cycle, the plate of the tube becomes positive and causes the following series of events:

1. Since no charge exists on C1 and the grid is at cathode, or zero bias, plate current immediately starts to flow, and the tube damps the oscillation.
2. The energy stored in the horizontal-deflection coils discharges through the tube, and the next active retrace is initiated.
3. As conduction starts capacitor C1 has become charged through the internal grid-to-cathode path of the tube. Resistors R4 in the grid circuit limit the grid current to a safe value.
4. At the start of conduction, C1 begins to discharge through the grid resistance network. The time constant of the discharge path, resistors R1 and R2, can be varied from 2 to 15 microseconds by adjusting R2.
5. The changing bias on the tube grid makes its plate resistance vary as scanning progresses and allows the shape of the decay part of the curve in Fig. 8-3 to be altered by adjusting R2 (linearity control).

The action of this type of horizontal-linearity control is independent from that of the horizontal-output tube plate circuit. Some interaction with the horizontal-width control (L2 in Fig. 8-4) does occur, however, and setting one may require readjusting the other.

Horizontal-width control L2 varies the proportion of the horizontal-transformer output reaching the deflection coils and controls the sawtooth current strength and the width of the picture.

VERTICAL-OUTPUT AMPLIFIER AND DEFLECTION CIRCUITS

The vertical-deflection system in Fig. 8-1 is much less involved than the horizontal system shown. A 6K6G, triode-connected, is used as the output amplifier. Transformer L4 matches the plate impedance of the tube to the impedance of the vertical-deflecting coil.

Any tendency for shock-excited oscillations to occur following vertical retrace is nullified by damping resistors R4, which are connected across vertical-deflection coils L3.

The proper shape of the combination sawtooth and pulse wave of voltage is supplied to the vertical-output tube grid from a blocking oscillator having a resistance-capacitance peaking circuit in its output.

Vertical-Linearity Control

The grid-voltage versus plate-current characteristic of the triode-connected 6K6G is not a straight line over its entire range. Cathode bias resistor R7 is made variable to shift the point of operation of the tube. As the operating point is shifted along the curve, the shape of the output scanning wave changes. These variations are sufficient to correct any lack of vertical linearity.

Since the tube gain varies as the bias is changed, this linearity adjustment will also affect height. Therefore, the vertical-height control may have to be readjusted.

TYPICAL MODERN HORIZONTAL-DEFLECTION SYSTEM

A typical modern horizontal-deflection system (Fig. 8-5) has some points of similarity to the older circuit of Fig. 8-1, but it also has many differences. The horizontal-drive adjustment consists of a variable capacitor at the grid of the horizontal-output

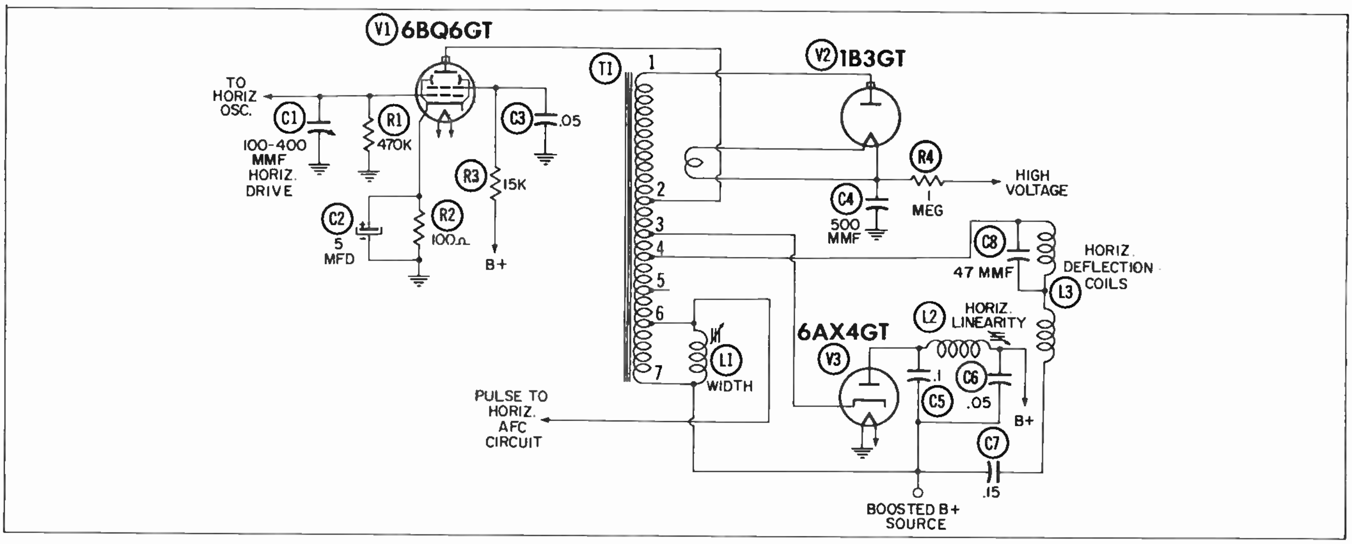


Fig. 8-5. A modern horizontal-deflection circuit.

tube. This capacitor and the coupling capacitor (not shown) form a voltage divider. A change in the setting of the drive adjustment changes its capacitance; thus, the voltage across it changes. Resistor R2 is not put in the cathode circuit to develop a bias voltage, but to protect the output tube if the oscillator fails.

The output transformer is an autotransformer. Plate current from the output tube flows to the B+ supply through the damper tube and through that portion of the winding between terminals 2 and 3. Autotransformer action builds up the higher voltage for rectifier V2, and the lower voltage to the deflection yoke is obtained from terminals 4 and 7 of the transformer. Capacitor C7 is needed to eliminate direct current in the deflection yoke. Width coil L1 functions by absorbing more or less power from the transformer.

Damper tube V3 damps the oscillations in the circuit when the voltage between terminals 7 and 3 reverses because of the inductive kickback produced by the sudden cutoff of output-tube plate current. The voltage provided by the damper conduction is stored in capacitors C5 and C6 and is used as the boosted B+ source. Linearity coil L2 is in series with the damper tube and changes linearity by changing the phase of the damper conduction current.

In a receiver utilizing a deflection circuit like this, horizontal and vertical centering and an ion trap, if one is used, are provided by permanent magnets on the picture-tube neck.

The vertical-deflection circuit in this type of receiver will almost always use a multivibrator driving an output stage, or a multivibrator in which the output tube is part of the multivibrator.

EARLY COMMERCIAL ELECTROSTATIC-DEFLECTION SYSTEMS

Electrostatic scanning (discussed in Chapter 7) consists of producing a sawtooth wave of voltage and applying it to the deflection plates by a push-pull output circuit.

Two methods of meeting these requirements were used. The first, and most widely used, consisted of a pulse-controlled, cathode-coupled multivibrator

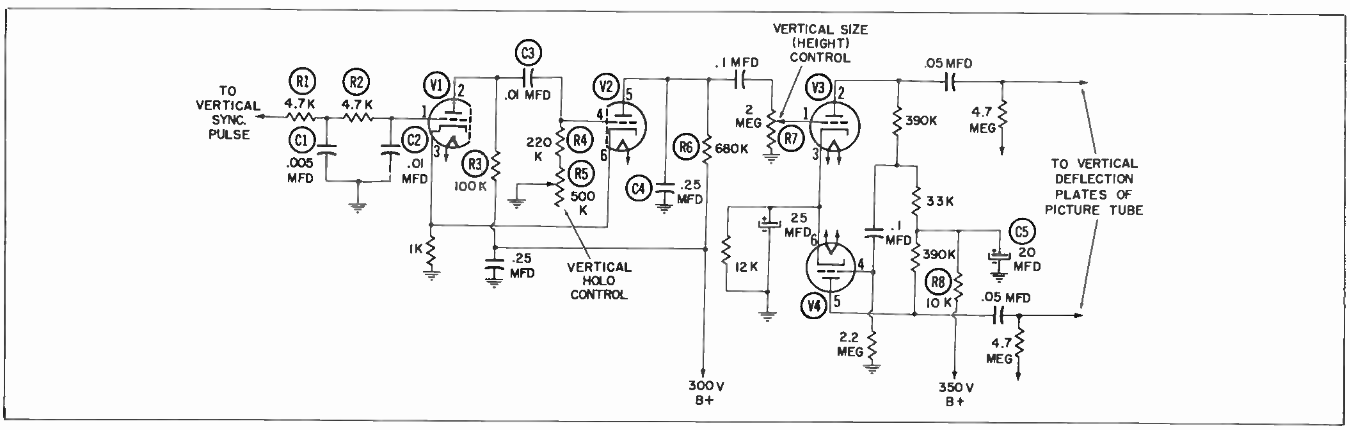


Fig. 8-6. Vertical-deflection circuit for an electrostatic picture tube.

feeding a phase-inverted push-pull amplifier. The second method was an unusual adaptation of the blocking oscillator, in which a single tube and its associated circuits fulfilled all scanning requirements.

Circuits Which Used the Cathode-Coupled Multivibrator

Most electrostatic-type television sets employed the cathode-coupled multivibrator with a phase-inverted amplifier for both horizontal and vertical systems. Since a number of differences in circuit constants and arrangements existed between the vertical and the horizontal systems, typical circuits of each will be discussed separately.

Vertical-Deflection Circuit—Fig. 8-6 shows a vertical-deflection circuit. Its action is as follows:

1. The vertical synchronizing signal, a series of negative pulses with respect to ground and spaced as shown in Fig. 7-2, passes through the network of R1, C1, R2, and C2. This is known as an integrating circuit. It adds all the small serrations of the vertical signal as a voltage across capacitor C2 until the negative potential on the grid of V1 reaches a value corresponding to plate-current cutoff. This action trips the multivibrator circuit.
2. The time constant, determined by the values of capacitor C3 and the sum of resistors R4 and R5, controls the free-running frequency of the multivibrator. R5 can readily adjust the frequency to lock in with the vertical-sync pulse at 60 cycles per second. Hence, it is called the vertical-hold control.
3. During the interval between the triggered conduction pulses of V2, sawtooth-forming capacitor C4 is linearly charged from the B+ source through resistor R6. Firing of V2 by the sync pulse discharges C4 and causes retrace.
4. Control R7 adjusts the input voltage impressed on the phase-inverted push-pull amplifier consisting of tubes V3 and V4. Because this control varies the output, it is known as the vertical size or height control.
5. The circuit associated with these tubes is the same type found in the audio system of many broadcast receivers and requires no further mention except to note:

   a. The coupling capacitors and load resistors are of such values that the circuit gain remains flat to the low frequency of vertical scanning (60 cycles).

   b. Plate-decoupling capacitor (C5) has a higher value than would ordinarily be required for an audio system. This high value is necessary to keep the vertical-sawtooth ripple out of the video- and audio-supply circuits.

In this circuit, it is not necessary to have a linearity control because the sawtooth produced across C4 is linear enough to provide good pictures.

Horizontal-Deflection Circuit—Fig. 8-7 illustrates the horizontal system employed in the same receiver as the vertical system of Fig. 8-6. It is similar to the vertical system with the following exceptions:

1. The horizontal-synchronizing signal, which consists of short pulses (5 microseconds) at the end of each horizontal picture line (Fig. 7-1), are impressed upon the grid of V1 through the network of R1, C1, and R2. This is known as a differentiating circuit. Here, the grid triggering voltage is taken across the resistor rather than across the capacitor, as is done in the vertical circuit. The circuit sharpens the pulse and delivers a negative stab of voltage on the grid to lock in the oscillator at the proper time.
2. The time constant of free-running oscillation, determined by the value of C2 times the sum of R4 plus R5, is much shorter (1/262.5 of the vertical time) than the vertical time-constant just discussed. This accounts for the lower values of capacitance and resistance, as compared with those in Fig. 8-5. As in the vertical circuit, R5 can be adjusted to lock in with the repetition rate of the sync pulses. Therefore, it is known as the horizontal-hold control.
3. The size or width of the picture is adjusted by a voltage control in the plate circuit of V2 (R7 in series with B+). This type of size or width control

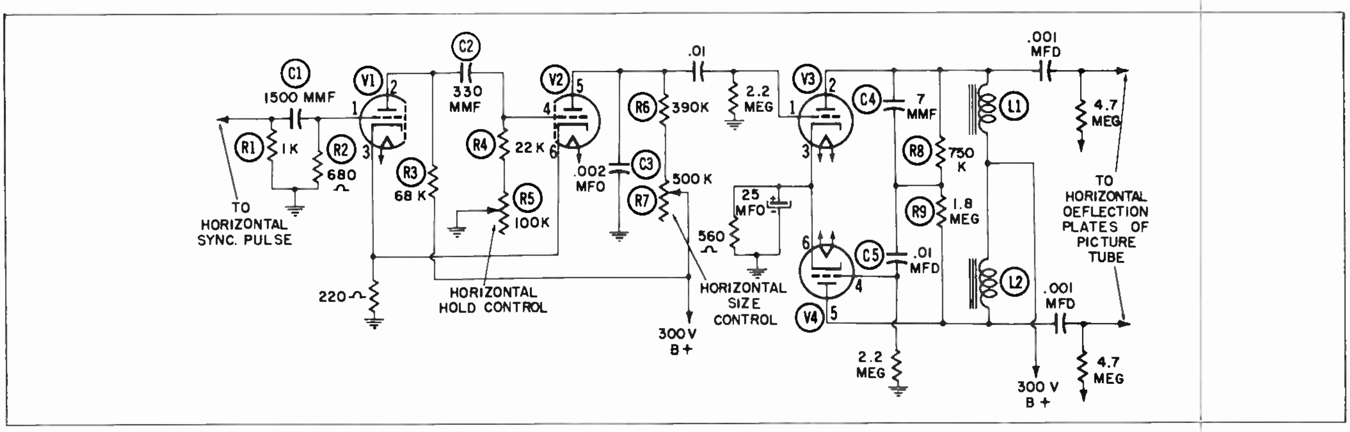


Fig. 8-7. Horizontal-deflection circuit for an electrostatic picture tube.

is sometimes used in both the vertical and horizontal circuits. It controls the charging voltage impressed on sawtooth-waveforming capacitor C3 and, hence, the height of the wave.

4. Compared with familiar audio practice, the phase-inversion circuit is a bit unusual. Since the fundamental horizontal-scanning frequency is 15,750 cycles, the capacitance as well as the resistance balance must be considered. This accounts for capacitor C4.

5. A word or two should be said about the choke supply to the plates of V3 and V4. Many receivers employed resistance coupling in this circuit, as shown in the vertical circuit in Fig. 8-6. However, chokes were sometimes employed because the high frequency (15,750 cycles) permitted an economical design in which the impedance of the chokes was high enough to have negligible shunting action across R8 and R9 as far as the horizontal-line frequency was concerned. The advantages of using chokes were (1) the plate voltage and, consequently, the output were higher, and (2) their series impedance acted as an excellent isolation of the system from the audio and video supplies.

Circuits Which Used the Blocking Oscillator

Vertical-Deflection Circuit—Fig. 8-8 combines a number of the circuit arrangements covered separately in our discussion of sawtooth generators and deflection means. The circuit was straightforward with regard to blocking oscillator V1 and its associated circuit elements. The unusual features were found in the phase-inverted push-pull amplifier.

As in the cathode-coupled multivibrator just described, an integrating network (R1, C1, R2, C2) sorts the vertical-sync pulse from the combined sync signal. However, the polarity of this pulse is positive with respect to ground. Control of the oscillator frequency by the pulse occurs as described in Chapter 6.

The oscillator free-running frequency is adjusted by controlling the time constant of the R-C network in the grid circuit of blocking oscillator V1. R5 in this circuit is the vertical-hold control. Vertical size or height of the picture is controlled by adjusting the charging voltage impressed on sawtooth-waveforming capacitor C6 by means of vertical-size control R7.

The phase-inverted push-pull amplifier V2 and and V3 is a 6SL7GT high-mu twin-triode. To supply the grid voltage of V3, the voltage is divided by a capacitance voltage divider C7 and C8, rather than the usual resistance divider found in the audio systems of broadcast receivers. Contact bias for both tubes is derived by grid current through the 10-megohm grid resistors R8 and R9. The plate-supply voltage of approximately 900 volts (from a bleeder across the high-voltage supply) feeds the tubes through 4.7-megohm plate resistors R10 and R11. The actual voltage at the plate is approximately 450 volts. With this supply, a plate-voltage swing of approximately 700 volts can be derived.

Like the cathode-coupled multivibrator circuit, the linearity of the sawtooth-voltage wave produced by this circuit is good enough to assure satisfactory pictures. Therefore, no linearity control is required.

Horizontal-Deflection Circuit — In the horizontal system that used the blocking oscillator (Fig. 8-9), the deflection-sawtooth voltages (balanced to ground) were developed in the oscillator itself. No amplifier stage was needed. With a DC plate supply of only 250 volts, this circuit delivered a balanced linear sweep of 1.200 volts peak-to-peak to the deflection plates of the picture tube. Note that this circuit differs from the customary blocking oscillator in that the cathode is not grounded. The plate-to-ground and cathode-to-ground circuits are symmetrically arranged with respect to ground or B−. This accounts for its balanced or push-pull output.

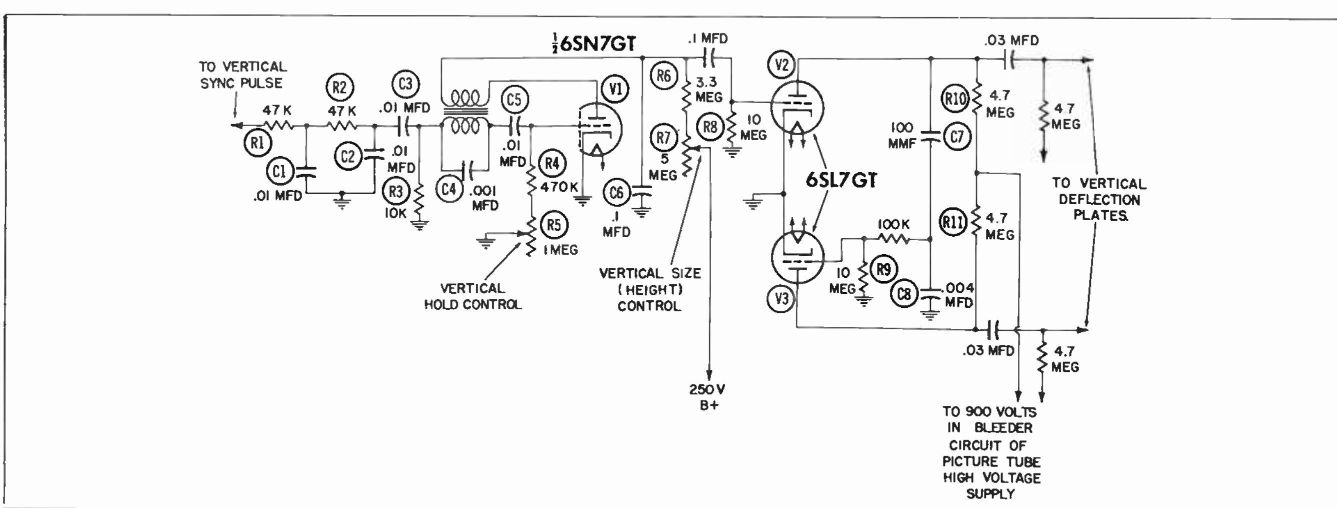


Fig. 8-8. Blocking oscillator—vertical electrostatic-deflection circuit.

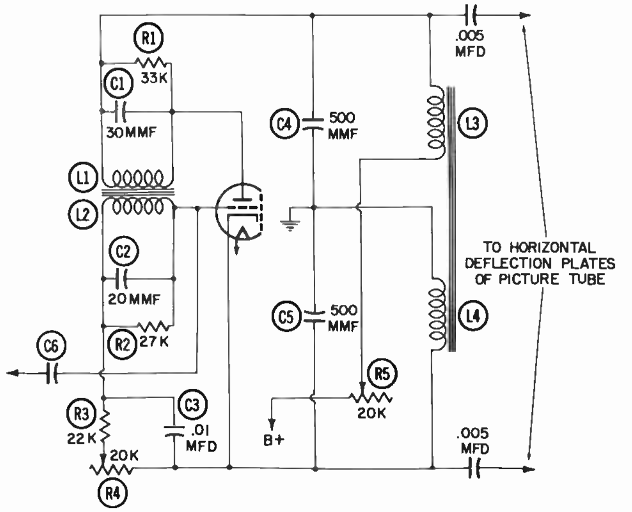


Fig. 8-9. Blocking oscillator—horizontal electrostatic-deflection circuit.

TYPICAL COMMERCIAL ELECTROMAGNETIC-DEFLECTION SYSTEMS

When we analyze the circuits of electromagnetically-deflected television receivers, which have become universally used in recent years, we find much greater diversity of design than in electrostatic systems. It will be instructive to examine the typical circuits and determine the operating principles of their various parts.

Vertical-Deflection Systems

Fig. 8-10 shows the details of an early vertical-deflection system. The system combines a number of the circuit elements previously discussed, such as the vertical pulse-integrating circuit, the blocking oscillator, the discharge tube, and the series R-C circuit which forms the combination sawtooth and pulse wave required for electromagnetic deflection.

The network of resistors and capacitors R1-C1, R2-C2, and R3-C3 comprises the integrating circuit which shapes the vertical-sync signal and applies it in series with the grid circuit of blocking oscillator V1. The pulse arrives at the grid at the proper time to trigger the plate-current pulse, as described and illustrated previously. This insures that retrace of the vertical scanning from bottom to top of the picture occurs at the correct instant (Fig. 7-2).

The action of the blocking-oscillator circuit of V1 has been covered. In the present circuit, its application is conventional. The free-running frequency is controlled by the time constant of capacitor C4 and of its discharge resistors R4 and R5 in series. R5 is variable and acts as the hold control to adjust the oscillator frequency to lock

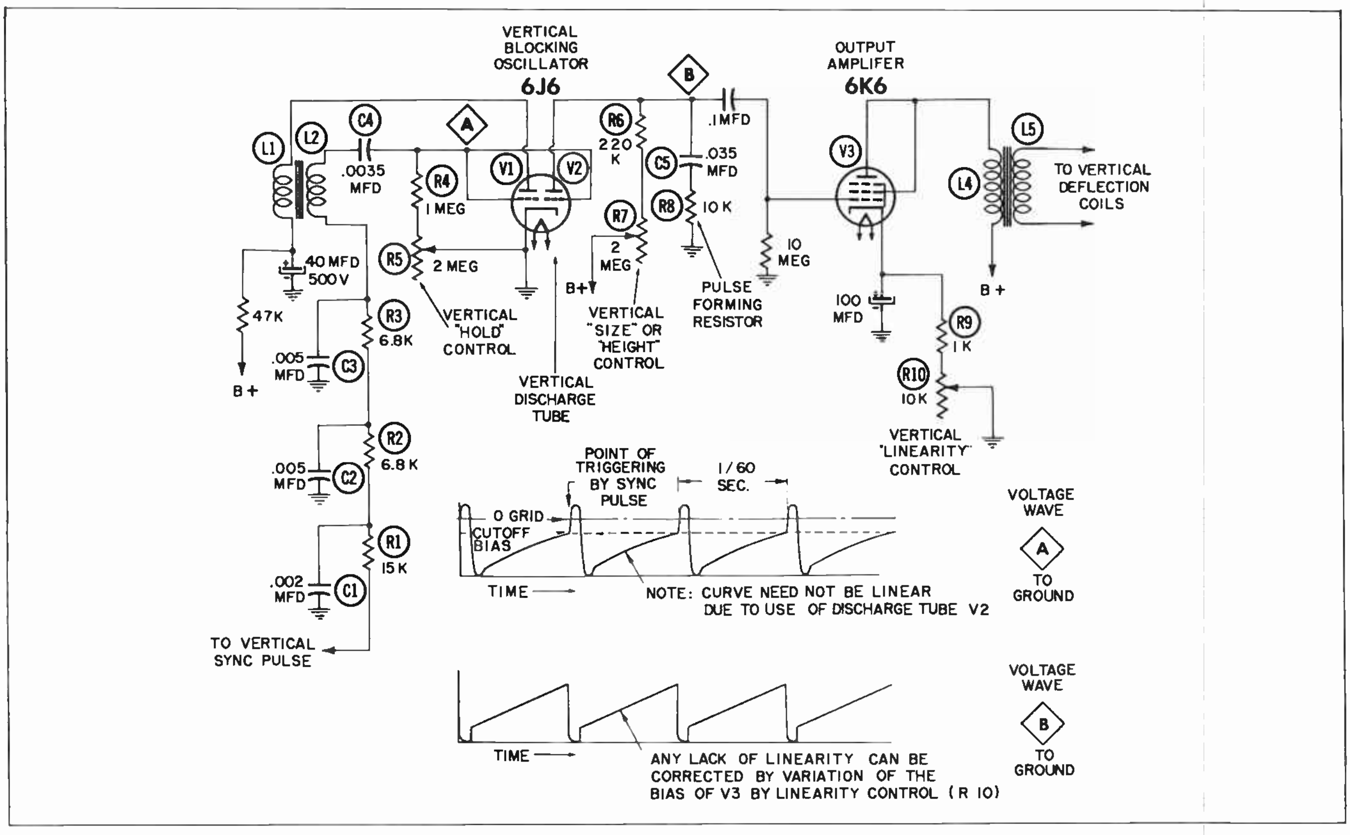


Fig. 8-10. Use of the blocking oscillator in a vertical-sweep circuit.

with the repetition rate of the vertical-sync pulses.

The second section of the dual triode acts as a discharge tube to short the network of C5 and R8, which forms a voltage wave consisting of a linear sawtooth followed by a pulse, as described in Chapter 7.

The charging voltage supplied to network C5-R8 is controlled by the series combination R6 and R7 connected to the B+ supply. R7 is made adjustable to act as the height or size control.

Tube V3 is a triode-connected output amplifier. It increases the amplitude of the sawtooth pulse to the proper level. The only feature of special interest in this part of the circuit is the variable cathode-bias resistor (R9 and R10 in series). Adjustment of the operating point on the grid-voltage versus plate-current curve by the setting of R10 introduces the proper amount of distortion to correct any departure from linearity of the sawtooth scanning wave. This control is known as the vertical-linearity control.

In practice, adjustment of R7 (vertical height) and R10 (vertical linearity) are somewhat interdependent. In this type of circuit, the vertical size or height control primarily affects the lower half of the picture, and the vertical-linearity control has its major effect upon the upper half of the picture. Waveforms A and B in Fig. 8-10 show the voltage waveforms appearing between ground and points A and B, respectively, on the schematic.

The vertical-deflection circuit of Fig. 8-11 employs the asymmetrical multivibrator and the vertical-output tube, which have been covered previously. A circuit not previously discussed, known as a sync clipper, is included since its plate-circuit components are an essential part of the deflection circuit under consideration. The sync clipper operates at a very low plate voltage (approximately 15 volts in this circuit) and derives its bias by grid rectification of the video signal. Plate-current conduction occurs on the peaks of the video signal only. These peaks are actually the sync pulses. Thus, V1 clips the sync pulses and passes waveform A (Fig. 8-11) on to sync amplifier V2.

Although we have labeled V2 a sync amplifier (its common commercial name), it should really be considered a coupling tube and isolation stage. Since resistor R7, the load of the tube, is in the cathode circuit, the voltage gain of the stage is less than unity. The output is taken from the

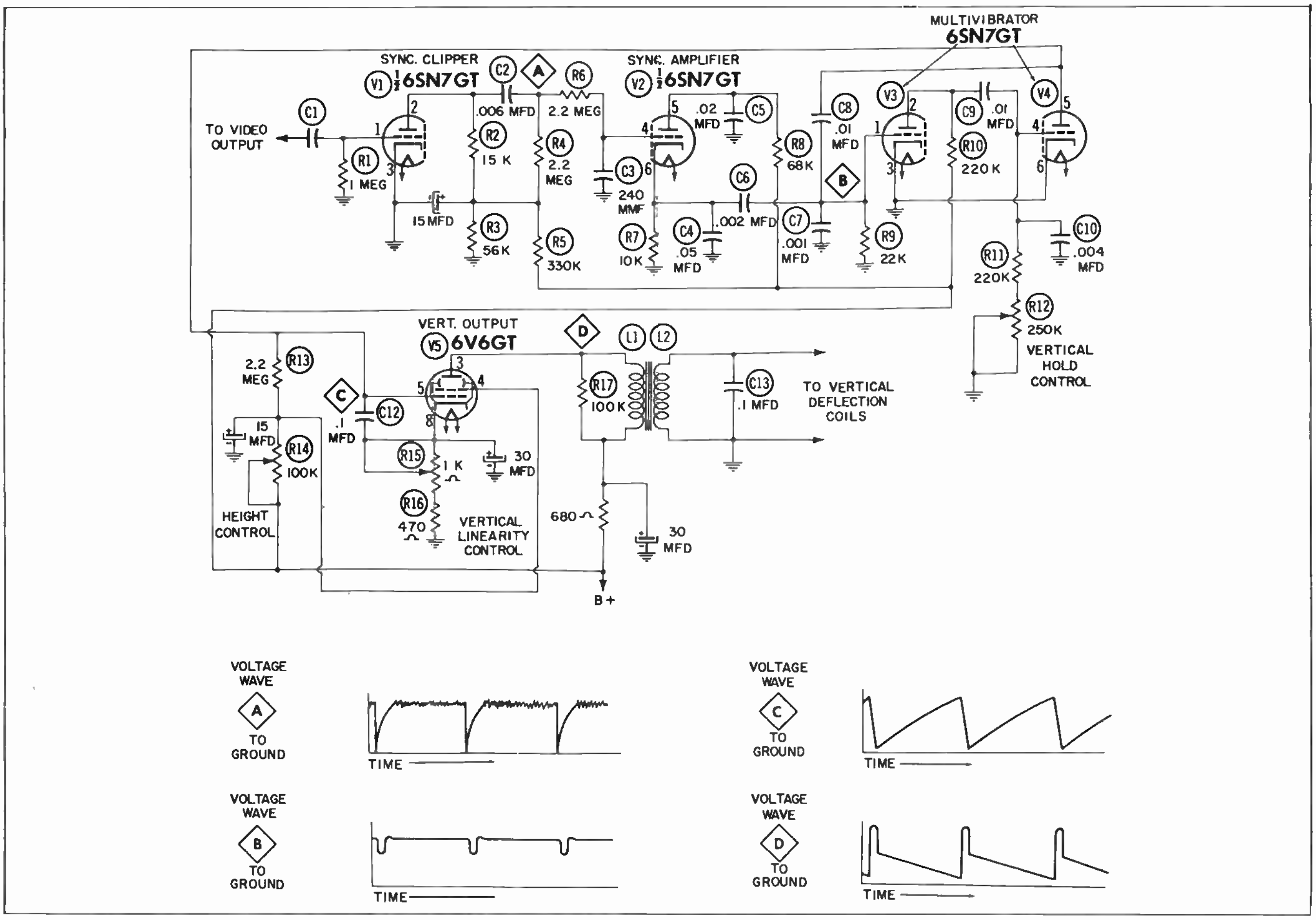


Fig. 8-11. Vertical-deflection circuit using multivibrator.

cathode. The voltage pulse this tube delivers to the grid of multivibrator V3 is negative with respect to ground. This is the requirement for tripping the sweep action. The shape of the negative pulses which appear at the grid of V3 is shown in waveform B of Fig. 8-11. The coupling network between sync clipper V1 and sync amplifier V2 has several unusual features:

1. Resistor R6 and capacitor C3 constitute an integrating circuit to accept the vertical-sync signals and reject the horizontal-sync signals. To make use of this circuit, an unconventional grid-biasing method becomes necessary.

2. Note that the grid of V2 is connected to B+ through resistors R6, R4, and R5 and to ground through resistor R3. This group of resistors acts as a voltage divider across B+. The positive voltage thus impressed on the grid opposes the negative bias from the cathode-bias resistor R7. As a result, the actual grid bias with respect to the cathode is approximately +3 volts. The positive bias assures a condition of plate-current conduction until cutoff is initiated by the negative pulse from the vertical-sync signal.

The asymmetrical multivibrator, consisting of tubes V3 and V4, is actually a dual-triode 6SN7 and is the same circuit covered in Fig. 6-8. The time constant of the grid circuit of V4 is approximately ten times that of tube V3. The short time-constant of V3 corresponds to the retrace period. The longer time-constant of V4 controls the active trace portion of the sawtooth wave.

As in the multivibrator circuits previously discussed, the free-running frequency is controlled by the time constant of capacitor C9 and the series combination of R11 and R12. R12 is made variable to act as the vertical-hold control.

Adjustable cathode-bias resistor R15 in the circuit of V5 serves as a linearity control. A slight amount of inverse, or negative, feedback occurs in this stage because C12 is connected to the cathode of the tube rather than to ground. This feedback helps improve the linearity of the sweep applied to the deflection coils.

Two new circuit elements, capacitor C13 and resistor R17, appear in connection with the output transformer. Like the circuit in Fig. 8-1 and its explanation, the secondary L2 of the vertical-

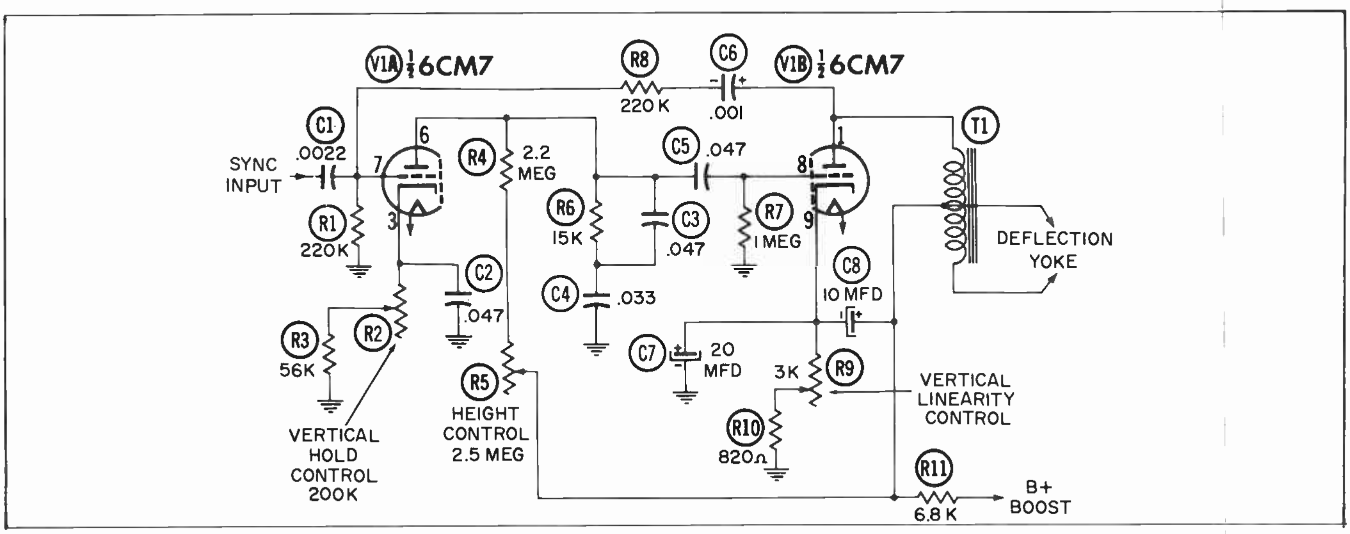
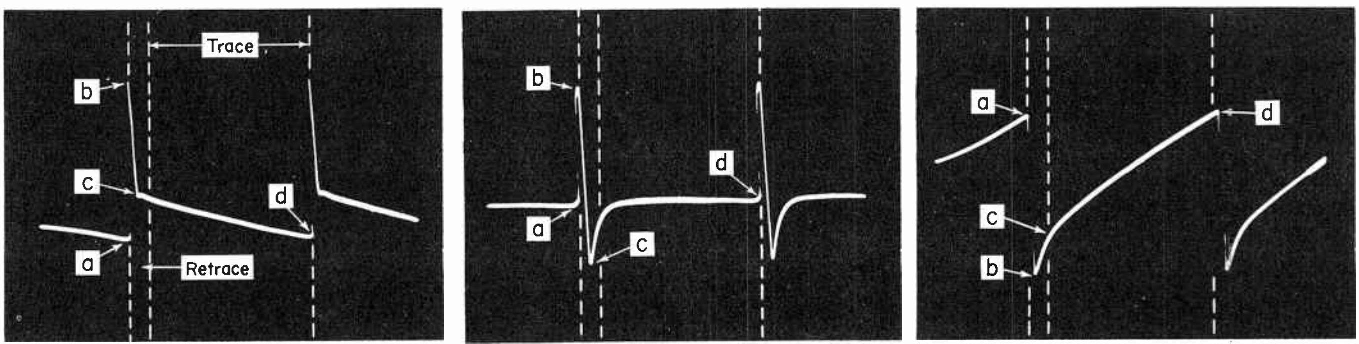


Fig. 8-12A. Vertical oscillator-output system employed in a modern TV receiver.



Waveform A—Plate of V1B. Waveform B—Grid of V1A. Waveform C—Plate of V1A.

Fig. 8-12B. Associated waveforms of circuit in Fig. 8-12A.

output transformer, capacitor C13 and the vertical-deflection coils are tuned to resonate at such a frequency that the first half-cycle corresponds to the vertical retrace time (500-750 microseconds). When the plate current of the vertical-output tube is suddenly cut off by the negative swing of voltage on its grid, the ringing, or oscillation, of the output circuit causes a rapid retrace of scanning. Any tendency of this circuit to continue in a state of oscillation is suppressed by the damping resistor R17.

The vertical oscillator and output circuit in Fig. 8-12 is found in one of the latest portable receivers. This circuit is also typical of many other modern receivers, and tube designers have provided a special tube for it. The 6CM7 tube consists of two dissimilar triode sections—the output section will handle more current and provide more output power than the oscillator section.

Like all multivibrators, oscillation takes place due to alternate conduction of the two sections of the tube. The circuit is free-running and will oscillate at its natural frequency without the necessity for external triggering. To synchronize the oscillator action with a signal from the transmitter, however, positive-going sync pulses are coupled to the system through capacitor C1.

Tube V1A acts somewhat as a switch which automatically charges and discharges the sawtooth-forming capacitor C4. During vertical trace time, V1A is nonconductive and capacitor C4 charges. The positive-going signal voltage developed across the combination of C4, C3, and R6 is then coupled to the grid of V1B, where it causes a sawtooth-deflection current in the plate circuit and through the yoke.

When retrace starts, V1A conducts and capacitor C4 discharges, so that a sharp negative voltage is applied to the grid of V1B and cuts it off. The circuit is so designed that V1A conducts during the retrace period only, or approximately 1/15th of the vertical cycle.

We can begin a detailed circuit description with the moment the set is turned on. With cathodes heated and plate voltage applied, both tubes will start to conduct. For the circuit to oscillate, however, one section must conduct while the other is cut off. Let us assume that this is initially accomplished by the following action:

As B+ is applied to the circuit, capacitors C5 and C6 attempt to charge to their associated plate potentials. Because of the relatively lower resistance in the charge path of C6 and the higher plate voltage on V1B, the initial capacitor-charging current through R1 and R8 will exceed that through R7, R4, and R5. The current through grid resistor R1 causes the grid voltage of V1A to rise slightly in a positive direction before the grid voltage of V1B rises. The positive-going voltage on the grid of V1A increases plate current through the tube. More voltage is dropped across resistors R4 and R5, and the plate potential is lowered. Because this negative-going signal is coupled through C5 to the grid of V1B, its conduction diminishes and its plate voltage increases. As the plate voltage of this tube rises, capacitor C6 couples the positive voltage back to the grid of V1A, and it conducts still more. The variation in plate-current flow between the two triode tube sections is amplified by this continued action until V1A becomes highly conductive and V1B is completely cut off. This is the action which occurs between points a and b on waveforms A, B, and C of Fig. 8-12.

Vertical-output transformer T1 presents a relatively large value of inductance in the plate circuit of V1B. When the output triode is cut off, current ceases to flow through T1, and the magnetic field produced by this current collapses. The voltage induced across the inductance by this collapsing field is very high and accounts for the high value (approximately 1,400 volts) of the pulse at point b in waveform A of Fig. 8-12. Since V1B is cut off, its plate voltage drops rapidly toward point c in waveform A after the magnetic field has collapsed.

A portion of the high pulse developed across the top section of T1 is divided between R8 and R1. Note the resulting positive grid swing in waveform B of Fig. 8-12. During this pulse, grid current flows in V1A and charges C6 in the polarity shown in Fig. 8-12. When the plate voltage on V1B drops to point c in waveform A, the charge on C6 adds to this negative voltage swing and causes the grid voltage on V1A to drop all the way to point c in waveform B. This negative grid voltage cuts off V1A.

To explain the shape of waveform C in Fig. 8-12 between points b and c, we must say that the charge on C4 and C3 is immediately dissipated when V1A conducts. Shortly after time b, when the grid starts in a negative direction and the plate current begins to fall off, the plate voltage rises once again. However, by the time the tube cuts off, the voltage reaches point c only. From point c to point d, capacitors C4 and C3 slowly charge through the path consisting of the power supply, R11, R5, and R4. This action accounts for the sloping rise of V1A plate voltage (waveform C) during this period.

The almost linear shape between points c and d of waveform C produces a steady increase in V1B grid voltage and, consequently, in V1B tube current. This is the reason for the steady decrease in V1B plate voltage between points c and d in waveform A.

Tube V1A is now cut off and V1B is conducting; this is the opposite of the initial circuit action. Conduction in V1A does not start immediately again after time c because its grid voltage is held just a few volts more negative than cutoff. The level portion of the trace between points c and d in waveform B of Fig. 8-12 represents this bias level. The exponential rise beginning at point c represents the decrease in the discharge current of C6. During the remainder of the trace period, C6 continues to discharge through R1 at a lesser but steady rate because of the linear decrease in V1B plate voltage. The steady rate of discharge through R1 holds V1A cut off during trace time. Another factor contributing to the maintenance of cutoff in V1A is the low plate voltage of V1A during the early part of the trace period.

All inductors have a saturation point, that is, a point at which additional current will not cause further expansion of the magnetic field. This is what happens within a blocking-oscillator transformer, and in this case, with T1. Current through V1B increases linearly until T1 becomes saturated. When this saturation occurs, V1B plate voltage will no longer decrease, and C6 will cease to discharge through R1. With bias removed, V1A starts to conduct, and the whole cycle is repeated.

The frequency of oscillation in this particular circuit is primarily determined by the charge and discharge of capacitors C4 and C3, as well as the inductance of T1. The vertical-synchronizing pulses coupled to the grid of V1A by way of C1 tend to force the oscillator into sync with any incoming signal.

Horizontal-Deflection Systems

Using Sine-Wave Oscillators—Fig. 8-13 shows the horizontal-deflection system of a receiver which is typical of early types using sine-wave oscillators. It consists of a combination of circuit elements which we have treated separately up to this point: the sine-wave oscillator controlled by the pulse through an AFC (automatic frequency control) system and reactance tube, pulse-shaping circuits, a horizontal-output tube, and a triode damping tube.

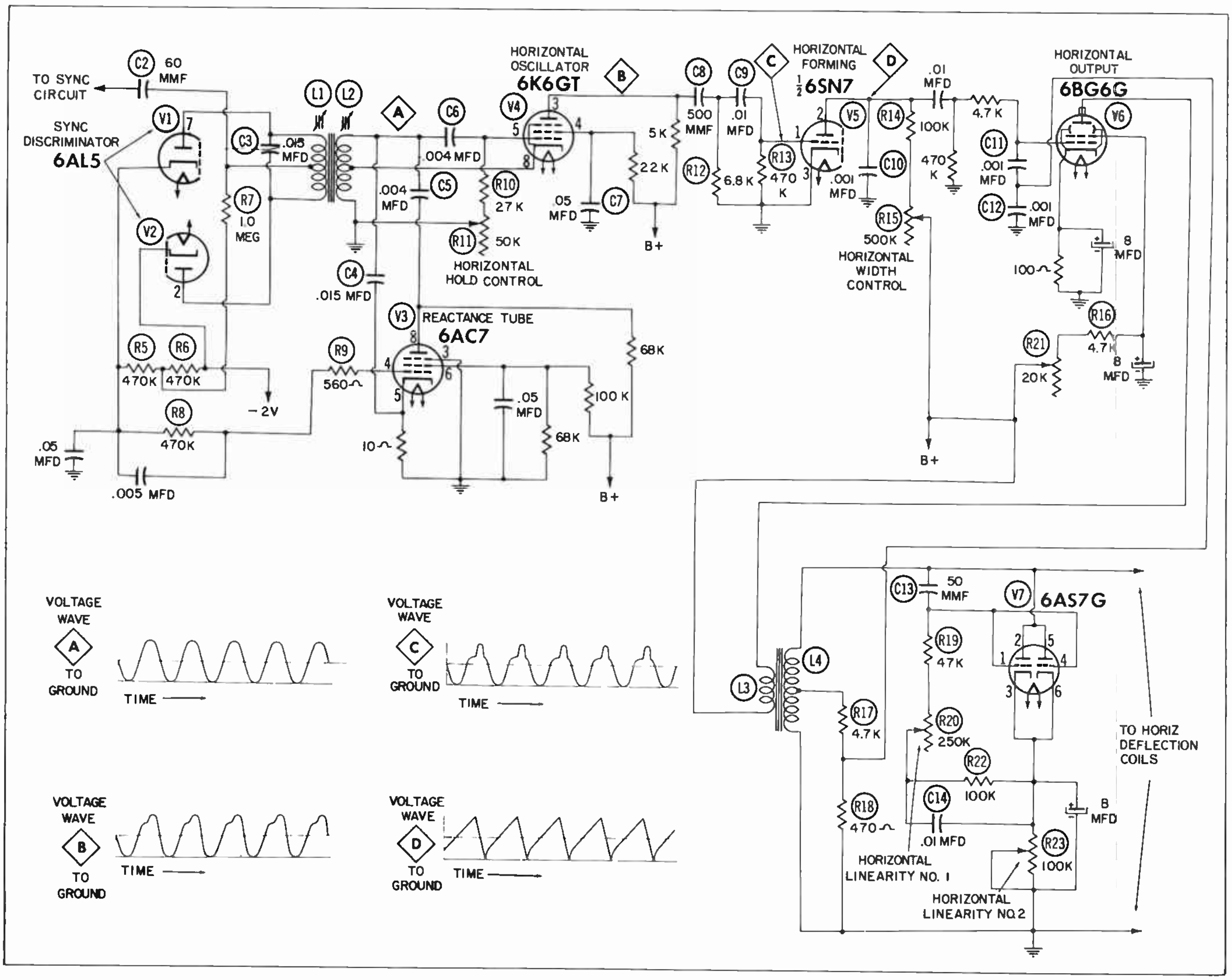


Fig. 8-13. Horizontal-deflection system using sine-wave oscillator and flywheel AFC.

The sync pulses are applied through C2 to the AFC sync discriminator, or phase detector. This circuit, employing dual-diode 6AL5, will be recognized as being similar to the Foster-Seeley discriminator used for FM detection. It is used in this circuit to compare the repetition rate of the sync pulses with the frequency of the horizontal oscillator and to produce a DC output voltage for control of reactance tube V3. The reactance tube automatically adjusts the horizontal-oscillator frequency and keeps it in step with the horizontal scanning. The action of this circuit will be described in more detail later.

Tube V4 is employed as an electron-coupled oscillator. The screen is at ground potential for the oscillator frequency (15,750 cycles) because of bypass capacitor C7. Feedback to sustain oscillation is furnished by the "hot" cathode tap on transformer secondary L2. In this circuit, there are two adjustable frequency controls: the iron-core tuning slugs of transformer windings L1 and L2, and the grid-resistor combination R10 and R11. The transformer adjustments are preset or semifixed to assure that the action of horizontal-hold control R11 will cover the proper frequency range on either side of the center of its rotation.

Waveform A of Fig. 8-13 shows the sine-wave voltage appearing across the oscillator-grid circuit. This voltage is electron-coupled to the plate circuit of V4. In this circuit, the sine wave is changed to that shown in waveform B of Fig. 8-13. This action takes place when the plate swing reaches the region of plate-current cutoff during part of the cycle.

Coupling network C8-R12 and C9-R13 acts as a differentiating circuit and further sharpens the waveshape, as shown in waveform C of Fig. 8-13. The voltage represented by waveform C is impressed on the grid of horizontal-forming tube V5. This tube acts much like the discharge tube described earlier. The positive voltage peaks on the grid cause plate conduction to discharge sawtooth capacitor C10 in the plate circuit. This discharge produces waveform D which drives horizontal-output tube V6. R14 and R15, in the plate circuit

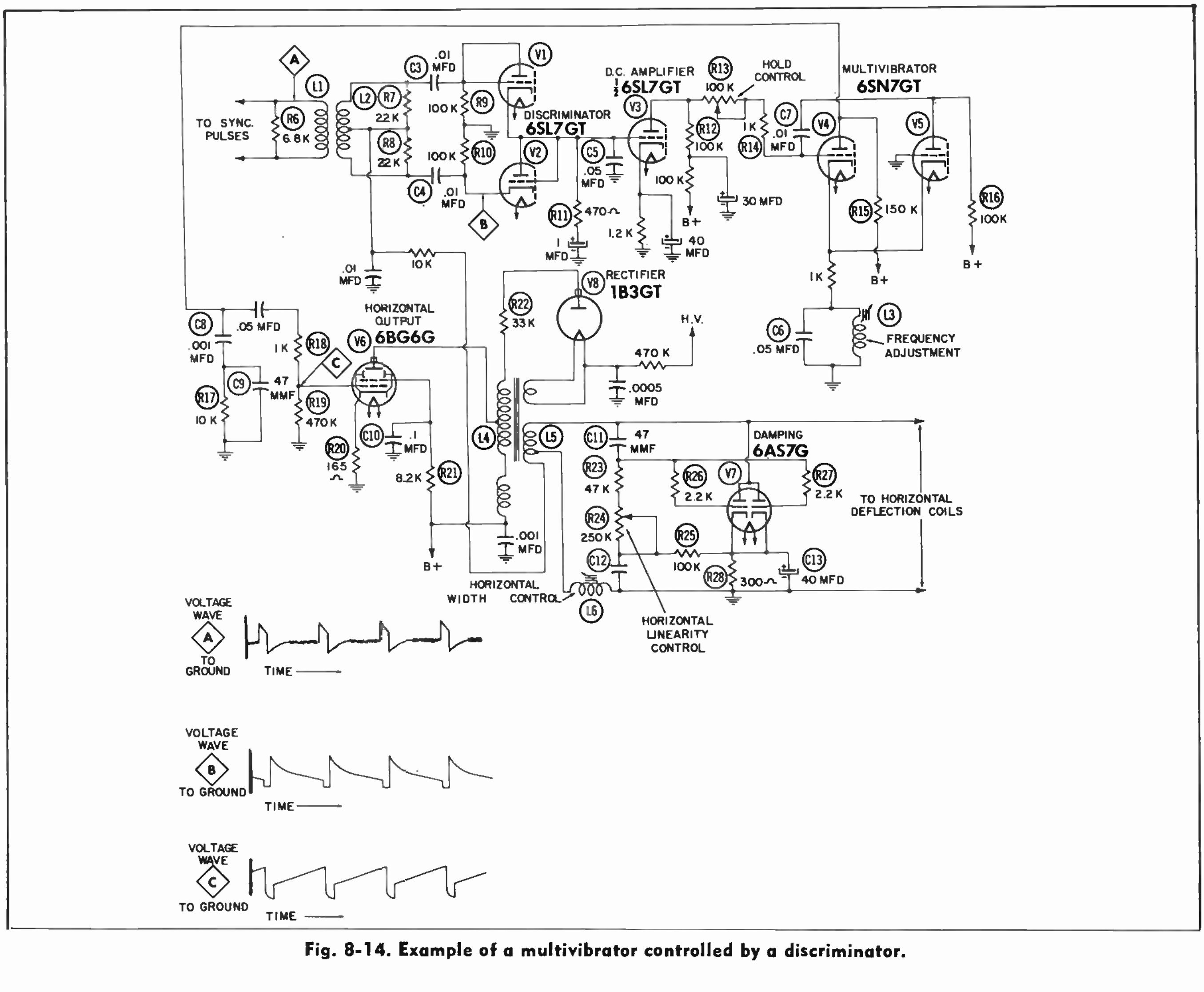


Fig. 8-14. Example of a multivibrator controlled by a discriminator.

of V5, control the charging voltage of C10. Variable resistor R15 acts as a horizontal size or width control.

Using Cathode-Coupled Multivibrators—The early type of horizontal-deflection system in Fig. 8-14 used a cathode-coupled multivibrator. A sample of the scanning-wave voltage is taken from output-transformer secondary L5 and injected into discriminator V1 and V2, where it is compared with the repetition rate of the horizontal-sync pulses. The high-mu twin triode (6SL7GT) in the discriminator circuit is employed as a dual diode by connecting the grids and plates of the respective sections to each other. The direct-current output of the discriminator is applied to the grid of DC amplifier V3. The plate resistance of the amplifier is part of the discharge-resistance network which determines the operating frequency of the cathode-coupled multivibrator.

The function of DC-amplifier tube V3 is similar to that of reactance tube V3 in Fig. 8-13. The resistor discharge network comprised of R12, R13, R14, and the plate resistance of V3 controls the time constant of the multivibrator grid circuit. One of these resistors, R13, is made variable to adjust the free-running frequency and, therefore, acts as the horizontal-hold control. The tuned circuit in the cathode return of tubes V4 and V5, consisting of capacitor C6 and inductor L3, is an additional control of the free-running frequency of the multivibrator. This is a service or semifixed adjustment and is used to set the range of the hold control so that it operates symmetrically about its midposition.

Waveforms A, B, and C in Fig. 8-14 show the waveforms found at various points in this circuit. Waveform A of Fig. 8-14 requires explanation, since it differs from the series of single horizontal pulses normally delivered by a sync amplifier. It consists of a positive pip followed immediately by a negative pip. The reason for this lies in the action of the primary of discriminator transformer L1. When a square-wave pulse of current passes through the inductance, magnetic flux is generated during the rapid rise of current at the leading edge of the pulse and the rapid fall of current at the trailing edge. Since these current changes are in opposite directions, they produce waveform A of Fig. 8-14.

Fig. 8-15 is the schematic of a horizontal phase detector and multivibrator system that is typical of a great number of present-day receivers. The phase detector compares the horizontal-sync pulses to pulses fed back from the horizontal output and produces a DC voltage which represents the phase difference between the two signals. This voltage is applied to the grid of the multivibrator and controls the firing point.

The coil in the plate circuit of the first triode of the multivibrator is known as a ringing coil. It and its associated capacitor are tuned to the horizontal frequency and develop a sine wave of voltage. This sine wave steepens the slope of the grid waveform on the second triode of the multivibrator when this triode is approaching its conduction point. Thus, the sine wave lessens the chance of accidental triggering of the oscillator by a noise pulse.

Using Blocking Oscillators—The circuit shown in Fig. 8-16 is a blocking-oscillator type of horizontal-scanning generator. It is very similar to the vertical generator of Fig. 8-10 except for the resistor

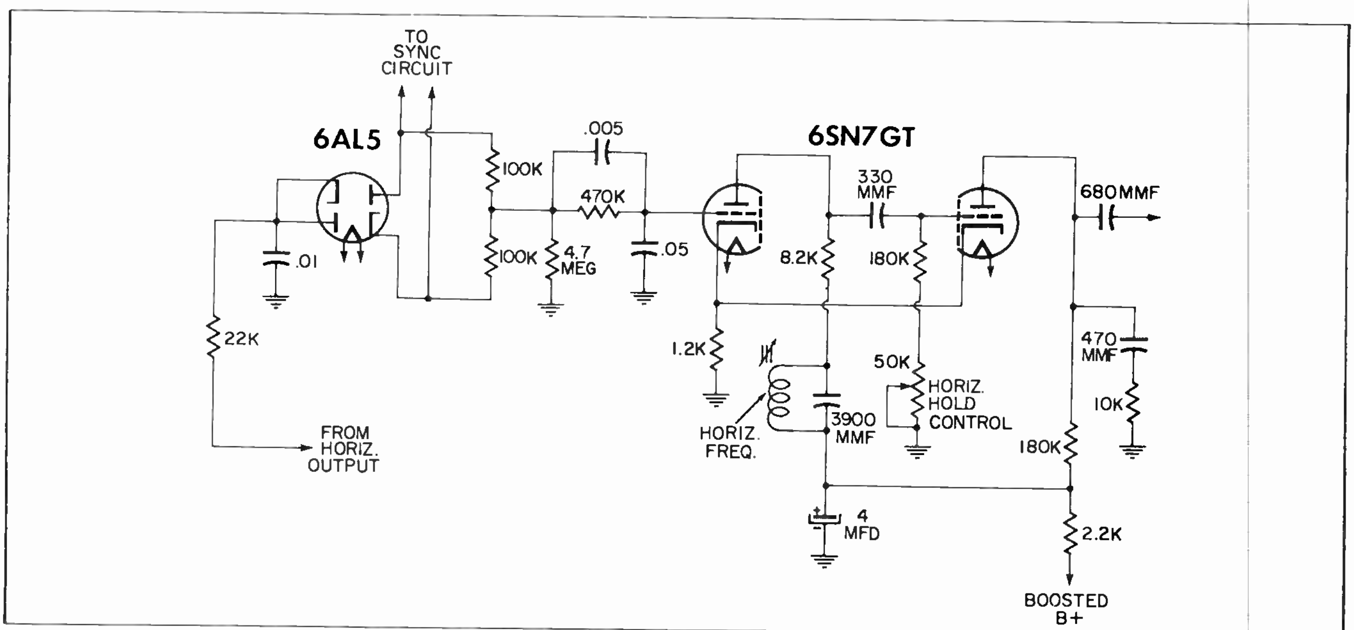


Fig. 8-15. Horizontal-phase detector and multivibrator circuit in many modern receivers.

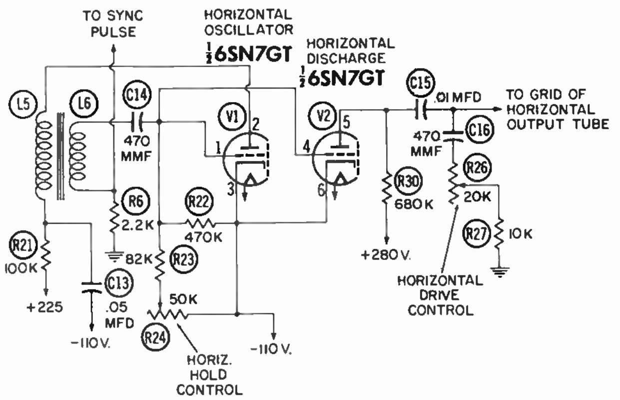


Fig. 8-16. Blocking-oscillator type of horizontal scanning generator.

and capacitor values, which differ because of the different frequencies being generated.

Horizontal-sync pulses are fed to the grid of blocking oscillator V1. The circuit consisting of resistor R6 and the secondary L6, with its distributed capacitance, accepts the horizontal pulses and rejects the vertical pulses. Blocking oscillator V1 is thus controlled by the horizontal-sync pulses, and proper scanning occurs.

The grid of horizontal-discharge tube V2 is connected directly to the grid of blocking oscillator V1 and functions as described previously. The sawtooth-generating and peaking circuit of C16, R26, and R27 is discharged, or triggered, by plate conduction of V2 which is controlled by blocking oscillator V1. Resistor R26 is made variable as a horizontal-drive control.

Using Pulse-Width Systems—Fig. 8-17 shows the schematic of a type of horizontal-scanning generator which originated in some of the earliest postwar receivers and is still used in modified form today. This is known as the Synchroguide or pulse-width system. It consists of a blocking oscillator and an AFC control tube.

The pulse output of the sync system must be positive with respect to ground. This pulse is applied to the grid of control tube V1 through the differentiating network of C1, C2, and C3. The action of the remainder of the circuit involves simultaneous events which can best be described in sequence.

1. Control tube V1 fulfills a number of functions:

   a. Since a portion of its cathode circuit, resistor R6, is common to the grid circuit of blocking oscillator V2, it can affect the oscillation frequency by influencing the time constant of the circuit of C10, R11, and R6 (paralleled by C8 and C9).

   b. The voltage drop across R6, due to the plate current of V1, can be controlled by potentiometer R8, which adjusts the plate voltage. R8 is the horizontal-hold control.

   c. Tube V1 serves as an automatic frequency control of scanning by mixing the original sync pulse, a signal fed back from the output of blocking oscillator V2, and a pulse from the horizontal-deflection coils through R15 and C12.

2. The pulse voltage derived from the deflection coils is the voltage which occurs at the instant of retrace. It is sharpened by R15 and C12. This pulse is negative with respect to the original sync pulse.

3. A small portion of the sawtooth voltage in the plate circuit of blocking oscillator V2 is also fed

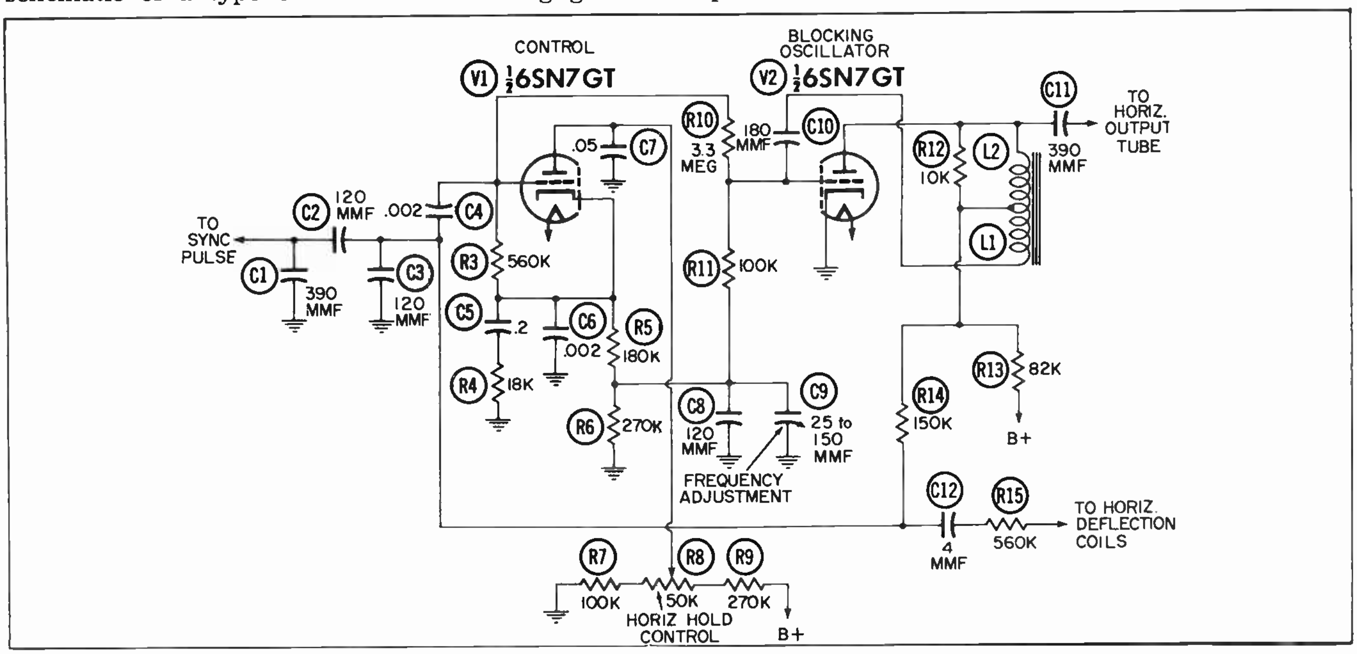


Fig. 8-17. Circuit using automatic pulse-width control of blocking oscillator.

back through resistor R14 to the grid of control tube V1. This voltage is positive (the same polarity as the original sync pulse) and is combined with the sharp negative pulse from the deflection coil. This combination produces a sawtooth which has a very sharp trailing edge.

4. The combination voltage just described is applied to the control-tube grid. It will coincide with the original sync pulse only if the oscillator is in exact step with the sync pulses.

5. As exact synchronization is reached, the control-tube grid pulse, which consists of the original sync pulse added to the feedback pulse, will be narrow and of high amplitude. If the feedback pulse is slightly fast or slow, it will not add to the original sync pulse, but instead will widen the original pulse.

6. From this combination of variable width and height of the voltage pulse on the grid of tube V1, very precise timing is achieved.

7. The plate-current pulse of tube V1, flowing through cathode resistor R6, adjusts the grid-circuit time constant of blocking oscillator V2 and produces the required exact synchronization.

Fig. 8-18 shows a modernization of the pulse-width circuit. A variable capacitor C3 has been

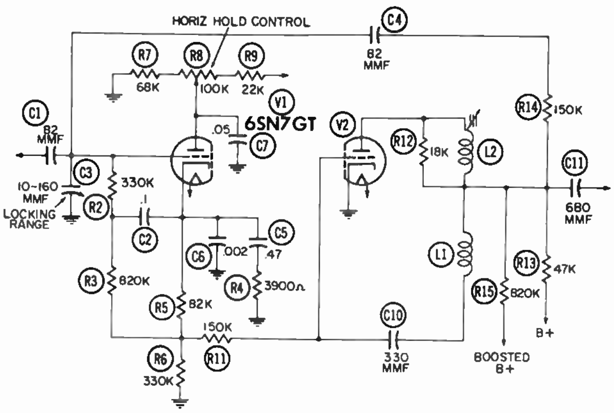
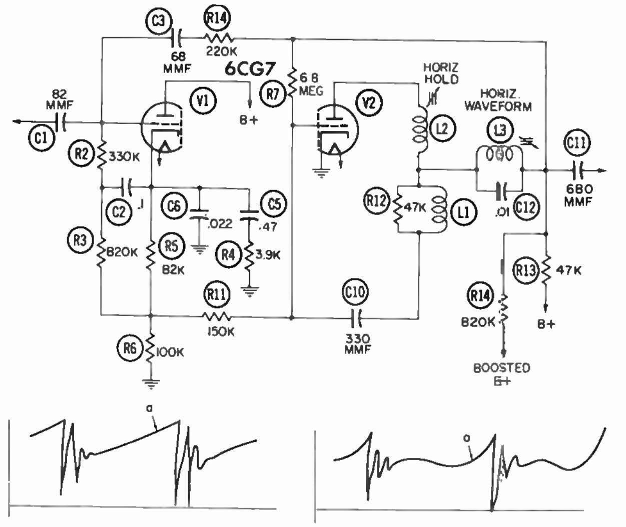


Fig. 8-18. A later version of the pulse-width circuit.

added at the grid of V1 as a preset or service control. It functions as part of a capacitance voltage divider. Adjusting it will vary the input voltage to the grid of V1. The feedback pulse from the horizontal output does not come from the deflection coils; instead, it comes from the boosted B+ source. The signal voltage at this point is due to the conduction of the damper tube and accomplishes the same function as a signal from the deflection yoke.

One of the latest versions of the pulse-width system is shown in Fig. 8-19. All the adjustments of the control tube have been eliminated, including the horizontal-hold control potentiometer. A hold adjustment has been provided by extending



Waveform A — Grid of V2 without stabilizing coil.

Waveform B — Grid of V2 with stabilizing coil.

Fig. 8-19. The most modern pulse-width circuit, and waveforms showing effect of stabilizing coil.

the screw on the slug in coil L2 and putting a knob on it.

A further refinement, which has been used with various pulse-width systems in the past, is coil L3 labeled horizontal waveform and commonly called a stabilizing coil. This inductance and the capacitance across it are tuned to the horizontal frequency, and they will produce a sine wave of voltage in series with the plate circuit. The sine-wave voltage steepens the slope of the grid-voltage waveform immediately before conduction time. This action improves the frequency stability of the oscillator by reducing the possibility of random noise triggering the oscillator. Waveforms A and B in Fig. 8-19 show the V2 grid voltage, and point (a) in each waveform shows the change of slope.

The pulse-width system previously discussed and the cathode-coupled multivibrator of Fig. 8-15 are being used in more than 90 per cent of present-day television receivers.

Flyweel or AFC Sync Control

The first postwar electromagnetic-deflection circuits used either the pulse-width or what is known as the flywheel, or AFC, system. The intricate flywheel system was eliminated in favor of simpler and more easily adjusted circuits.

Fig. 8-20 shows a typical circuit adaptation of the automatic-frequency control of horizontal scanning by the sync pulses. More properly, it should be called automatic-phase control since it locks the phase as well as the frequency of the horizontal oscillator.

The operation of the circuit is as follows:

1. A very stable oscillator tube V4 with its associated circuits generates a sine wave of 15,750

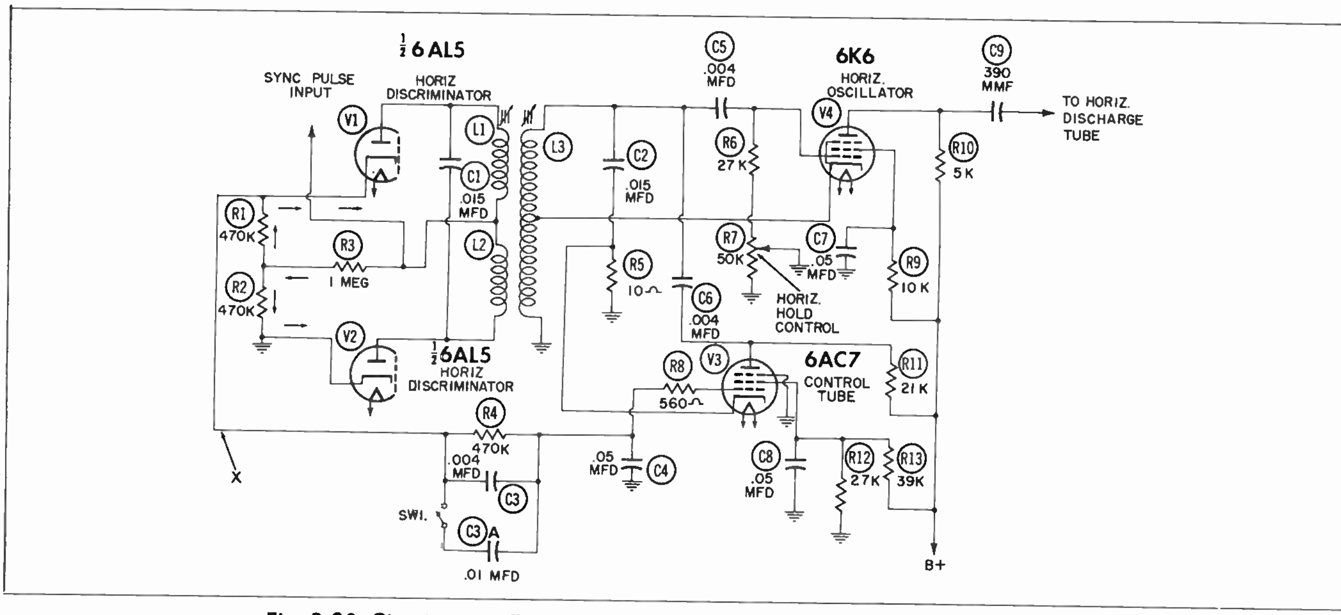


Fig. 8-20. Circuit using flywheel or AFC control of horizontal-scanning oscillator.

cycles per second. This oscillator circuit is the "hot cathode" Hartley type. Its frequency-determining components are inductor L3 and capacitor C2. The free-running frequency is adjusted by the powdered-iron core of inductor L3 with hold control R7 in midposition.

2. A tapped-coil circuit, L1-L2, is tightly coupled to L3 and tuned slightly off resonance. The voltage across these coils is applied to a discriminator circuit similar to the type used for frequency-modulation detection. The voltages from the horizontal oscillator, which are applied to the plates of discriminator tubes V1 and V2, are equal in amplitude and opposite in phase (180° out of phase with each other). See 1 and 2 in Fig. 8-21.

3. The clipped and amplified sync pulses are applied to the discriminator across resistor R3. They appear in phase and of the same amplitude at each diode of the discriminator, as shown in Fig. 8-21.

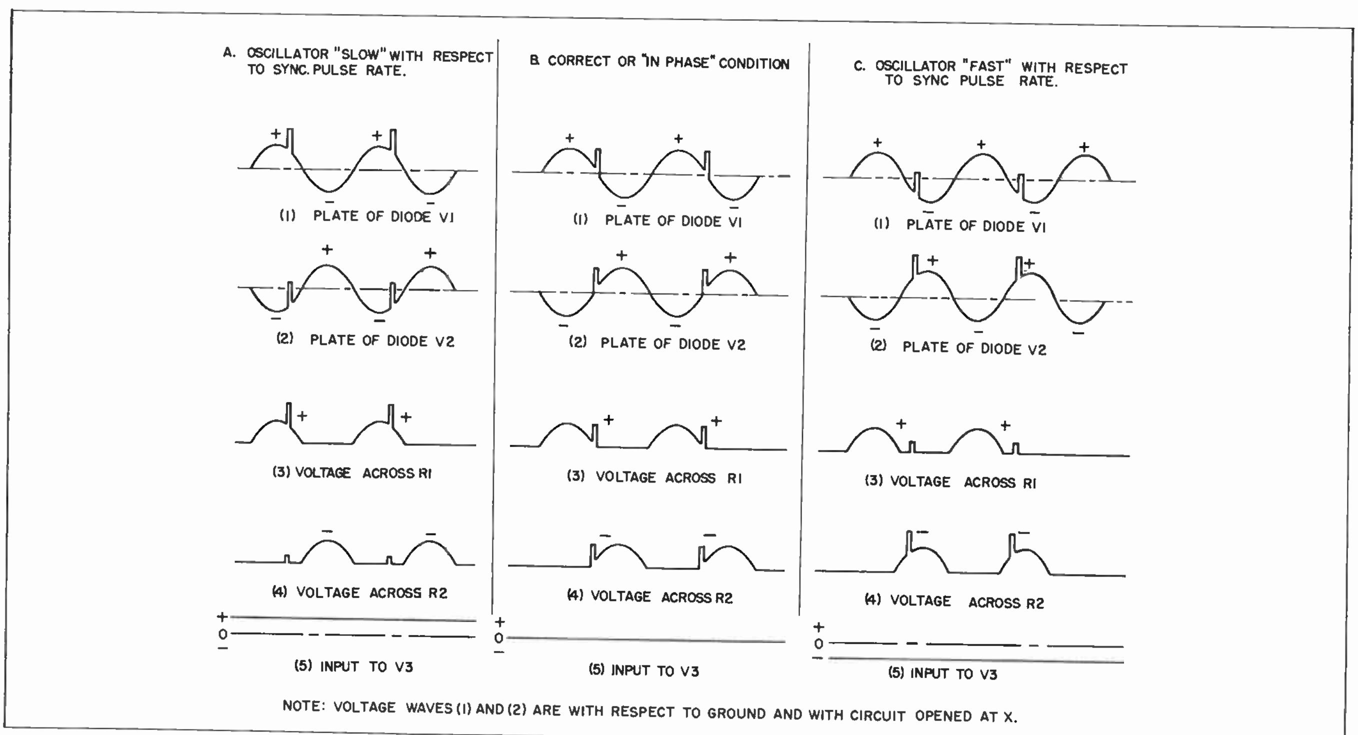


Fig. 8-21. Discriminator action in horizontal AFC sync control circuit of Fig. 8-20.

4. Three conditions of the system in which the relation of the oscillator frequency to the sync-pulse repetition rate is slow, correct, and fast are shown in Figs. 8-21A, 8-21B, and 8-21C, respectively. Note that diode load resistors R1 and R2 are so connected that the voltage from ground to point X is the difference between the rectified output of the diodes V1 and V2. Arrows show the direction of electron flow due to the plate current of these tubes.

5. If only tube V1 is conducting, point X will be positive with respect to ground. If tube V2 is conducting and no current flows through tube V1, point X will be negative with respect to ground. In the absence of sync pulses, equal and opposite voltages appear across the diodes. The total rectified voltage between point X and ground is zero over the cycle.

6. Stable operation occurs with the sync pulses riding at zero phase, or on the axis of the wave, as shown in Fig. 8-21B. Here, the rectified output of diode V1 would produce a voltage across R1, as shown at (3) in Fig. 8-21B. The lower diode V2 would produce the voltage wave shown at (4) in Fig. 8-21B. These voltages oppose each other because of the method of connecting the diodes to series load resistors R1 and R2. The net charge on filter network C3, R4, and C4 is zero, as shown at (5) in Fig. 8-21B.

7. If the horizontal-scanning oscillator is running slower than the repetition rate of the sync pulses (Fig. 8-21A), the upper diode output voltage will exceed that of the lower diode, as shown at (3) and (4) in Fig. 8-21A. After smoothing through the filter network, a positive voltage will appear at the grid of reactance-control tube V3.

8. Similarly, if the oscillator is running faster than the repetition rate of the sync pulses, the conditions shown in Fig. 8-21C will apply. Here, a negative control voltage will appear at the grid of V3.

9. The 6AC7 reactance-control tube is connected across the oscillator-tank inductance L3. This tube controls the instantaneous frequency of the oscillator in the following manner:

a. The plate-current verses plate-voltage curves of a high-gain, sharp-cutoff pentode such as the 6AC7, exhibit a long range of plate voltage over which there is substantially no change of plate current. This region is shown in the curves of Fig. 8-22. The plate current is controlled by the grid voltage, as indicated by the individual curves of the plate current vs. plate voltage family.

b. Referring to the circuit of Fig. 8-22, we find that the plate is connected to the high side of the oscillator tuned circuit through capacitor C2. The alternating voltage from the oscillator, appearing across the tube, is swinging over the

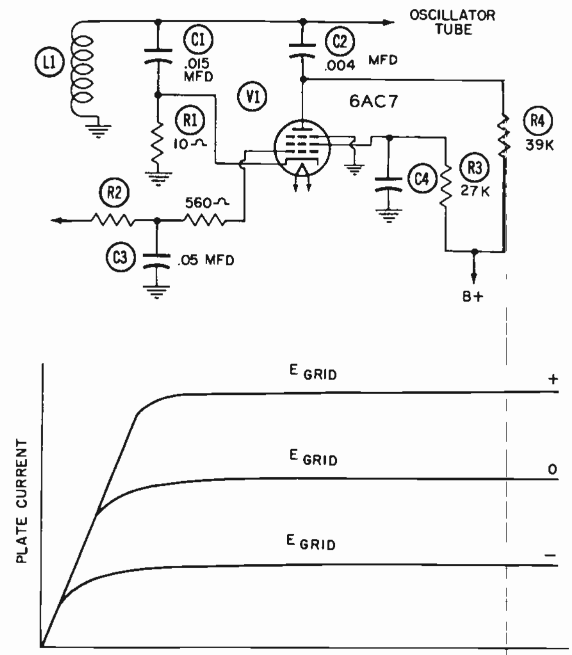


Fig. 8-22. Reactance-control operation.

flat portion of the curve. Therefore, the plate current does not change. On the other hand, if an alternating voltage is applied between grid and cathode, the plate current will be changed in amplitude. This change will be in phase with the alternating grid voltage.

c. If the alternating voltages applied to the grid and plate are made 90° out of phase, the plate current will then have an alternating component 90° out of phase with the plate voltage, i.e., in phase with the grid voltage.

d. Any circuit where the current is not in phase with the applied voltage is reactive. The circuit then appears to be capacitive or inductive. Thus, V1 of Fig. 8-22 can be made to appear as a reactance across oscillator inductance L1.

e. The phase shift of applied alternating grid voltage with respect to the alternating plate voltage is provided by the network C1-R1. The voltage across R1 is leading the voltage across the tank circuit by almost 90°. Since the reactance of coupling capacitor C1 is low, the voltage applied between cathode and grid is approximately 90° out of phase with the voltage applied between plate and ground.

f. The control grid is returned to ground or to the end of R1 opposite the cathode through capacitor C3. The reactance of C3 is very low at the oscillator frequency of 15,750 cycles. The DC return path of the control grid is through the output load resistors of the discriminator stage (R1 and R2 of Fig. 8-20). Since the alternating grid voltage is applied between cathode and the grounded grid, rather than between grid and ground, the plate differs by 180° from the phase it would have if the cathode had been grounded and the grid voltage made variable. The plate current will lag the plate voltage, and the tube will appear as an inductance to the tuned circuit.

g. The bias voltage on the grid of the reactance tube controls the value of its transconductance (ratio of change in plate current to the change in grid voltage which caused it). The alternating plate-current amplitude can thus be changed by varying the bias on the control grid. A low or positive bias will increase the AC plate current, and a more negative bias will reduce the AC plate current. Since the AC plate voltage is fixed by the oscillator output, a bias change will change the ratio of voltage to current. Thus, the apparent value of inductance across the oscillator circuit will be changed. For example, a more positive bias on the reactance tube grid would cause an increase in inductive plate current. This would simulate a decrease in the inductance across the tuned circuit and cause the frequency of the oscillator to speed up.

10. If the horizontal-oscillator frequency should shift with respect to the sync-pulse rate because of variation of supply voltage or other causes, the DC output of the discriminator would change. The transconductance of the control tube will be changed, and the instantaneous oscillator frequency will be shifted to restore the equilibrium shown in Fig. 8-21B.

The network of resistors and capacitors (R4, C3 or C3A, C4, and R8) which connect the output of the discriminator to the grid of control tube V3 (Fig. 8-20) requires explanation. Resistor R4 provides a DC return path for the grid of control tube V3. Resistor R8 acts as a parasitic suppressor to prevent any possible high-frequency oscillation in the high-mutual conductance control tube 6AC7.

Capacitors C3 and C4 constitute a voltage divider and a filter. The ripple on the rectified output of the discriminator diodes is smoothed to furnish a DC voltage to the control tube. Since the reactance of the smaller capacitor C3 (.004 mfd) is on top of the divider and the larger capacitor C4 (.05 mfd) is on the bottom, extremely rapid voltage changes are prevented from affecting the grid voltage of the control tube. Rapid changes can be caused by bursts of noise or by the serrations of the vertical-sync signal. These changes are rejected by this voltage divider.

The ratio of the voltage divider C3 and C4 can be changed by switching capacitor C3A (.01 mfd) in parallel with C3. This link connection for the addition of C3A was provided to take care of a condition which existed in some television transmitters. If any phase modulation of the exact position of the horizontal-sync pulse occurs, the AFC control tube must follow this modulation, and a faster response circuit is required. This faster response is provided by increasing the capacitance of C3 by adding C3A. Phase modulation of the horizontal pulses, if not suppressed by the original capacitance ratio, would produce a horizontal displacement of part of the picture with respect to the raster.

QUESTIONS

1. What are the main purposes of the damper circuit?
2. Is the horizontal-deflection coil system predominantly resistive or inductive?
3. Does the damper tube conduct during the first or second half of the trace period?
4. Name the three controls that affect the horizontal linearity.
5. What is the purpose of the vertical-output transformer?
6. What type of network is used to reject the horizontal-sync pulses and to accept the vertical-sync pulses? What type of network accepts the horizontal-sync pulses?
7. What signals are applied to the input of the control tube in a Synchroguide circuit?
8. What is the purpose of the stabilizing, or ringing, coil in a multivibrator?
9. What is the purpose of the flywheel sync-control circuit?

EXERCISES

1. Show, by using a block diagram, the stages of a vertical-deflection system by starting at the output of the sync amplifier. Describe the function of each stage.
2. Show, by using a block diagram, the stages of a horizontal-deflection system by starting at the output of the sync amplifier. Describe the function of each stage.

SECTION 3

Beam Modulation and Synchronization

Chapter 9

The Composite Television Signal

Before going further in our discussion, let us take a look at a block diagram of a television receiver and check the sections that have been studied up to this point and those that still need investigation.

A block diagram of a television receiver is shown in Fig. 9-1. Shown in this diagram are those sections that have been discussed up to this point, those that will be discussed in the following chapters, and those that will not be discussed. The chapter in which each section is covered is indicated by number under each block. Arrows indicate the signal paths.

The audio amplifier and the speaker are shown as sections that will not be discussed. It is suffi-

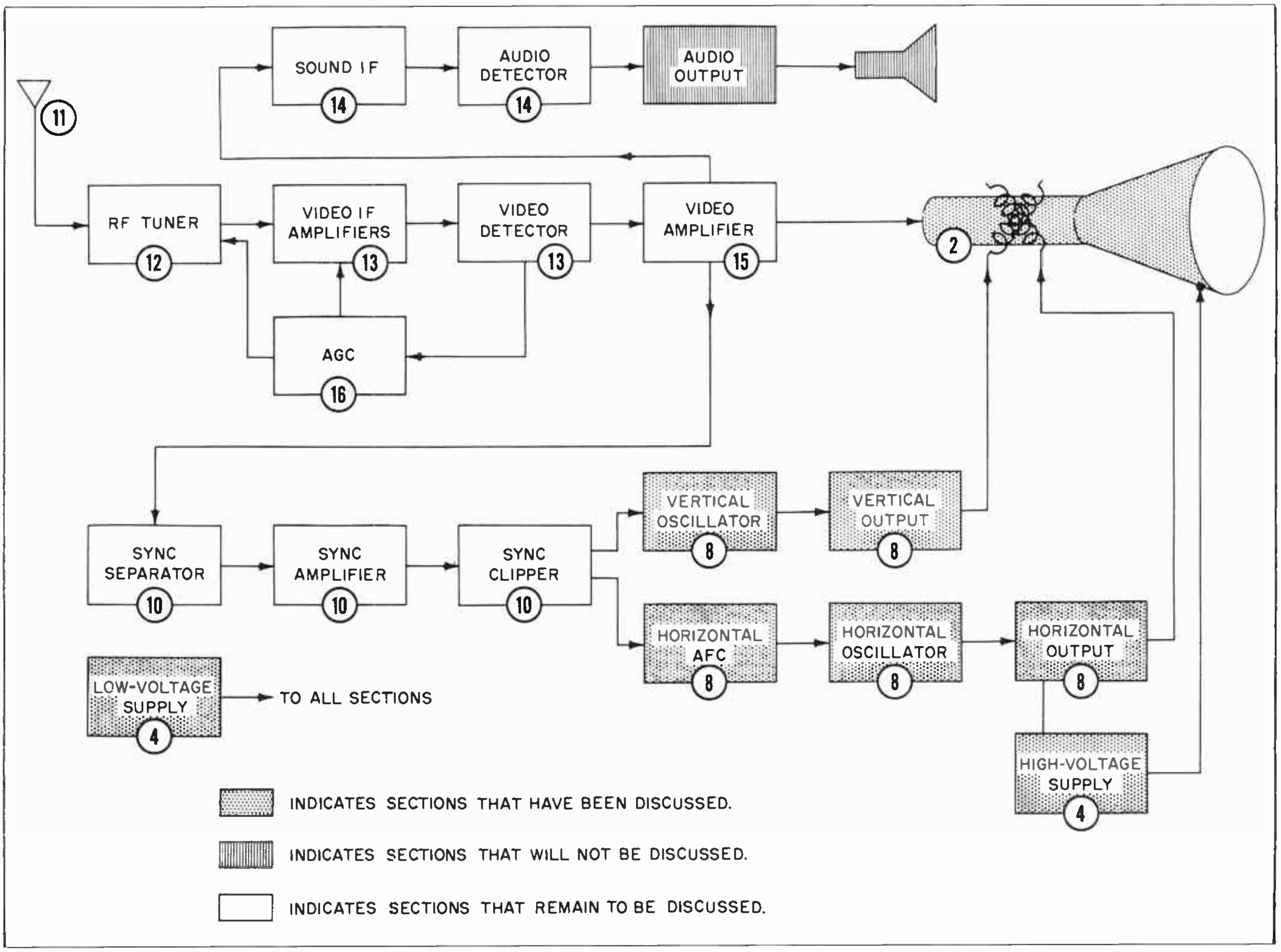


Fig. 9-1. Block diagram of a television receiver.

cient to say that they serve the same purpose in television receivers as they do in radio and audio equipment.

In order to understand the operation of the remaining sections, we must study the nature of the composite television signal at this point.

The television signal is more complex than broadcast and short-wave signals, which carry only speech or telegraphic information. To better present the reasons for the complexity of the television signal, we will review the composition of the familiar broadcast-band, amplitude-modulated carrier before describing the television carrier.

During the amplitude modulation of a carrier wave by speech or music, frequencies higher and lower than the carrier are produced. These frequencies are known as sidebands. These sidebands occur as a result of the beat, or heterodyne, between the carrier frequency and the modulating frequency. In the broadcast band, for example, the highest modulating frequency for speech or music transmission is approximately 5,000 cycles per second. Thus, if a broadcast station operates at 1,000 kc (one million cycles), the 5,000-cycle modulation will produce sidebands that extend from 1,000,000 cycles minus 5,000 cycles to 1,000,000 cycles plus 5,000 cycles (995 kc to 1,005 kc). A double-sideband modulated broadcast carrier is shown in Fig. 9-2.

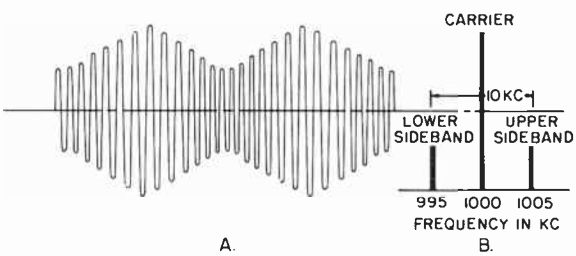


Fig. 9-2. Double-sideband modulated carrier wave; 50 per cent modulation, 5,000-cycle tone.

The modulating frequencies encountered in the television video signal extend from less than 30 cycles per second to over 4,000,000 cycles per second. It will be interesting to examine the reasons for this tremendous frequency range in the output of the transmitter camera tube.

As the signal spot in the camera tube sweeps across the scene, approximately 480 lines are scanned in 1/30th of a second. (Although there are 525 lines per frame, only 480 lines contain video modulation because of those lost during vertical retrace.)

If we assume that the picture element is square and that the resolution, or definition, is the same both horizontally and vertically, we can calculate the maximum frequency produced. The ratio of the width of the picture to its height has been set as 4:3. This is known as the aspect ratio. If the vertical resolution is set by the 480 horizontal lines, the number of dots in one horizontal line for equal resolution horizontally would be 480 × 4/3, or 640. Multiplying the number of horizontal lines (480) by the number of dots in one horizontal line (640), we obtain a possible maximum of 307,200 picture elements for a single frame. Thirty frames per second (two 60-cycle fields) multiplied by this figure of 307,200 produces an upper frequency of 9,216,000 picture elements per second.

Fortunately, this figure does not represent the top frequency, as you can see from the video signal produced by scanning a checker board (Fig. 9-3) consisting of white and black squares, 480 squares

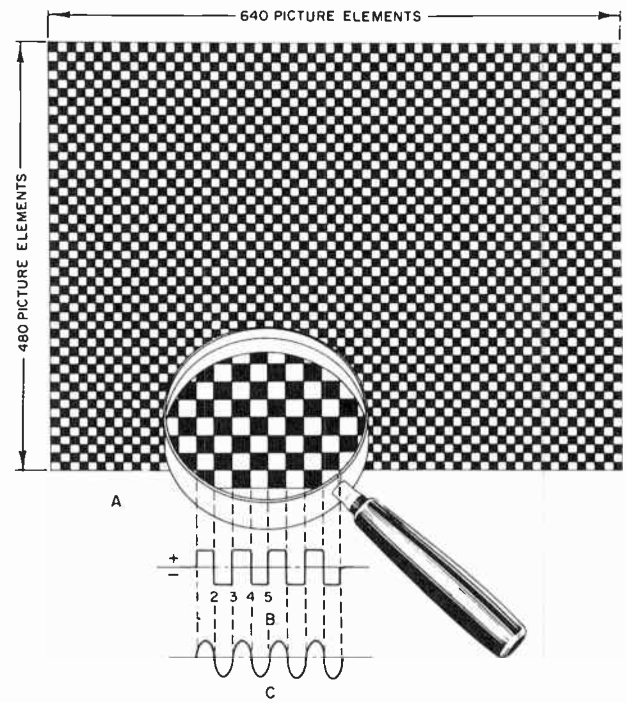
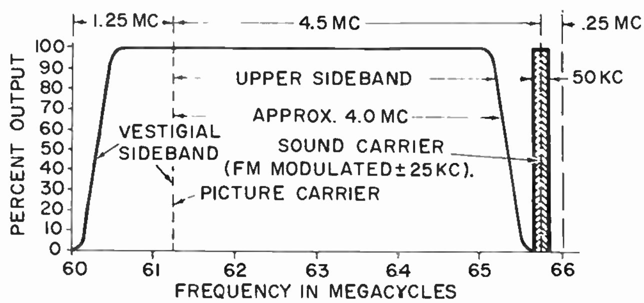


Fig. 9-3. Video modulation produced by scanning a checker board in which the picture elements are the limit of system resolution.

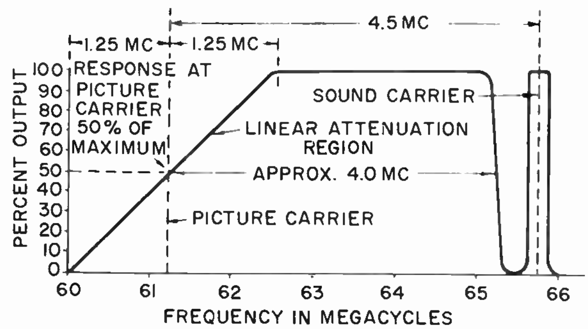
vertically and 640 horizontally. In the television transmission system that is standard in the United States, an increase in carrier strength caused by an increase in video-modulating voltage produces a darker spot on the cathode-ray tube. The brightest spot in the picture corresponds to the lowest modulating voltage. Thus, when the scanning beam in the camera tube reaches the edge of a black square at point 1 in Fig. 9-3, the video-modulating voltage of the transmitter assumes its maximum positive value and remains at this value as the scanning action crosses the black portion of the image. At point 2, the scanning reaches a white area, and the modulating voltage changes to its maximum negative value. The video voltage then remains negative until point 3 is reached, at which point the cycle begins to repeat. Since two picture elements have been scanned to produce one cycle of video voltage, the top modulating frequency required will be one-half the number of picture elements, or in this example, 4,608,000 cycles (4.6 megacycles) for 9,216,000 picture elements. Since we used as an example the maximum resolution of the system, in which a picture element has been made the same size as the scanning spot, the video output would be the sine wave at C rather than the square wave at B in Fig. 9-3.

If double-sideband modulation of the type employed in radio broadcasting were used in television, sidebands extending 4.5 megacycles on both sides of the carrier would require a bandwidth of 9.0 megacycles for the video signal. In the early days of television research, when the pictures were of much lower definition (90 lines instead of 525), double-sideband modulation modulation was employed. In addition to the bandwidth required for the video carrier, the accompanying sound carrier and a guard space must be provided. The Federal Communications Commission has assigned a channel space of only 6 megacycles for each television channel. Therefore, it is impossible to use the double-sideband type of modulation and still retain the definition that is possible in a 525-line picture. Other types of modulation which require less channel space to accommodate the wide band of modulation frequencies encountered in television are (1) single-sideband modulation and (2) vestigial-sideband modulation (one sideband plus a vestige of the other).

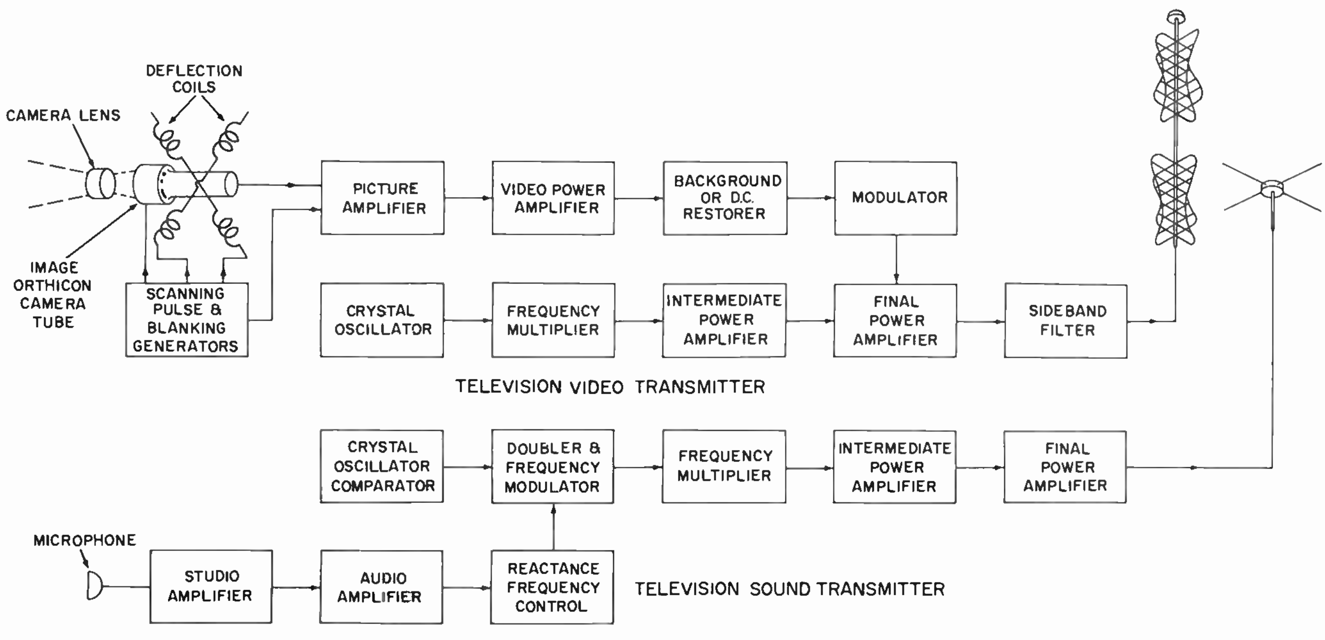
In the process of detecting (demodulating) an amplitude-modulated carrier, one sideband is always eliminated, and the information in the other is amplified and used to produce the picture. The single-sideband method requires only half the space in the radio spectrum for a given maximum modu-



(A) Channel-3 output.



(B) Ideal receiver response characteristic.



(C) Essential elements of a transmitter.

Fig. 9-4. Output characteristic of a television transmitter, the required response characteristic, and the block diagram of the essential elements of the transmitter.

P = PICTURE CARRIER FREQ. (MC)
S = SOUND CARRIER FREQ. (MC)

CHANNEL NUMBER	P	S	FREQ. LIMITS OF CHANNEL (MC)
2	55.25	59.75	54–60
3	61.25	65.75	60–66
4	67.25	71.75	66–72
5	77.25	81.75	76–82
6	83.25	87.75	82–88
7	175.25	179.75	174–180
8	181.25	185.75	180–186
9	187.25	191.75	186–192
10	193.25	197.75	192–198
11	199.25	203.75	198–204
12	205.25	209.75	204–210
13	211.25	215.75	210–216
14	471.25	475.75	470–476
15	477.25	481.75	476–482
16	483.25	487.75	482–488
17	489.25	493.75	488–494
18	495.25	499.75	494–500
19	501.25	505.75	500–506
20	507.25	511.75	506–512
21	513.25	517.75	512–518
22	519.25	523.75	518–524
23	525.25	529.75	524–530
24	531.25	535.75	530–536
25	537.25	541.75	536–542
26	543.25	547.75	542–548
27	549.25	553.75	548–554
28	555.25	559.75	554–560
29	561.25	565.75	560–566
30	567.25	571.75	566–572
31	573.25	577.75	572–578
32	579.25	583.75	578–584
33	585.25	589.75	584–590
34	591.25	595.75	590–596
35	597.25	601.75	596–602
36	603.25	607.75	602–608
37	609.25	613.75	608–614
38	615.25	619.75	614–620
39	621.25	625.75	620–626
40	627.25	631.75	626–632
41	633.25	637.75	632–638
42	639.25	643.75	638–644
43	645.25	649.75	644–650
44	651.25	655.75	650–656
45	657.25	661.75	656–662
46	663.25	667.75	662–668
47	669.25	673.75	668–674
48	675.25	679.75	674–680
49	681.25	685.75	680–686
50	687.25	691.75	686–692
51	693.25	697.75	692–698
52	699.25	703.75	698–704
53	705.25	709.75	704–710
54	711.25	715.75	710–716
55	717.25	721.75	716–722
56	723.25	727.75	722–728
57	729.25	733.75	728–734
58	735.25	739.75	734–740
59	741.25	745.75	740–746
60	747.25	751.75	746–752
61	753.25	757.75	752–758
62	759.25	763.75	758–764
63	765.25	769.75	764–770
64	771.25	775.75	770–776
65	777.25	781.75	776–782
66	783.25	787.75	782–788
67	789.25	793.75	788–794
68	795.25	799.75	794–800
69	801.25	805.75	800–806
70	807.25	811.75	806–812
71	813.25	817.75	812–818
72	819.25	823.75	818–824
73	825.25	829.75	824–830
74	831.25	835.75	830–836
75	837.25	841.75	836–842
76	843.25	847.75	842–848
77	849.25	853.75	848–854
78	855.25	859.75	854–860
79	861.25	865.75	860–866
80	867.25	871.75	866–872
81	873.25	877.75	872–878
82	879.25	883.75	878–884
83	885.25	889.75	884–890

Fig. 9-5. Channel allocations for commercial television.

lating frequency, compared with the double-sideband method. Single-sideband modulation is produced by passing a double-sideband modulated carrier through a radio-frequency filter network, which suppresses the undesired sideband. Although the single-sideband method allows the greatest use of channel space, it is impractical for transmission of television signals and has been replaced by the vestigial-sideband method because:

1. The type of filter which must be employed to completely suppress one sideband introduces serious phase distortion of the low-frequency components of the video signal and results in a blurred picture.

2. Even though means of correcting the phase shift may be developed, the single-sideband method would require extremely accurate tuning of the receiver and a high degree of freedom from oscillator-frequency drift. Otherwise, the low-frequency portion of the video modulation would be lost.

VESTIGIAL-SIDEBAND VIDEO MODULATION

To overcome the difficulties encountered with single-sideband modulation, the system of modulation employing all of one sideband and a vestige of the other has been adopted as standard in the United States. The fact that a part, or vestige, of one sideband is transmitted gives it the name vestigial-sideband modulation.

Fig. 9-4A shows a typical television channel employing vestigial-sideband modulation. Channel 3 has been chosen for this example. A block diagram of the essential parts of the transmitter required to produce this type of carrier and modulation is shown in Fig. 9-4C.

In this channel, which extends from 60 to 66 mc, the picture-carrier frequency is placed 1.25 mc above the lower limit of the channel, or at 61.25 mc. Video modulation of this carrier produces an upper sideband with frequencies extending to approximately 4 mc. The accompanying sound has a carrier frequency 4.5 mc above the video carrier, or at 65.75 mc. A guard channel or space of 0.5 mc separates the video modulation from the sound channel. The sound produces a frequency modulation of its carrier with a maximum deviation of 25 kc (a total carrier swing of 50 kc). Between the sound carrier and the upper end of the channel (66 mc) is another guard band or space of 0.25 mc.

Note that the lower-sideband frequencies are extended to approximately 1 mc below the video-carrier frequency. For video modulation frequencies extending to 1 mc, the transmitting system is essentially double sideband. The presence of the lower sideband would overemphasize the low frequencies if the receiver response characteristic were flat throughout the entire band. To compensate for this overemphasis, the receiver characteristic is such that the carrier suffers a 50% loss (Fig. 9-4B). The receiver response is made to drop off linearly from a frequency 1.25 mc above the video-carrier frequency to the lower limit of the band. When the receiver is correctly tuned to the channel, this loss of the lower sideband produces a flat output from approximately 25 cycles to 4.5 mc.

When the Federal Communications Commission first announced channel allocations, 13 channels were assigned for television broadcasting. These channels were divided into a low-band and a high-band group, with each channel occupying a 6-mc band. Channels 1 through 6 originally covered a band of 44 to 88 mc (72 mc to 76 mc being reserved for other services). Channels 7 through 13 covered a band of 174 to 216 mc. Channel 1 (44 to 50 mc) is no longer used for television. In 1952, the UHF band was opened by the FCC. This band includes 70 channels (14 thru 83) between 470 and 890 mc. The VHF and UHF television channels with their respective picture and sound carriers are listed in the table in Fig. 9-5.

Fig. 9-6 illustrates the consecutive arrangement of channel spacing. Note the guard space of 0.25 mc between the sound carrier of the adjacent television channel and the low end of the desired channel.

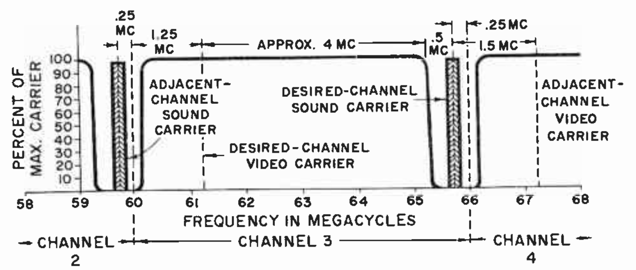
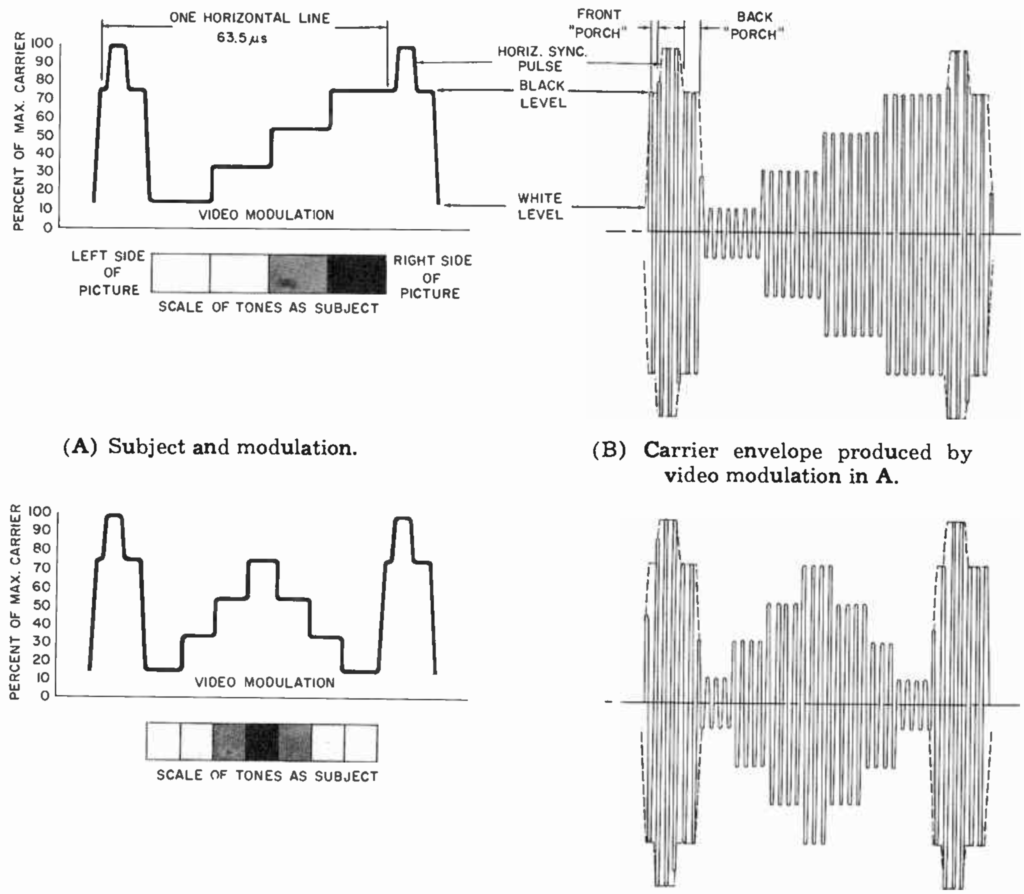


Fig. 9-6. Arrangement of consecutive channel spacing for Channels 2, 3, and 4.

Similarly, there is a space of 0.25 mc between the sound carrier of the desired channel and the low end of the video modulation of the next higher channel. Television receivers must be able to attenuate or reject these undesired transmissions occurring in adjacent channels.

THE VIDEO SIGNAL

Fig. 9-7 shows the video signal which would be produced if a card carrying a series of tones ranging from pure white through a number of greys to black were placed in front of the television camera. Fig. 9-7A represents one horizontal line scanned across the series of tones. The line is preceded and followed by pedestals upon which the horizontal-synchronizing pulses are mounted. A scale at the left of Fig. 9-7A indicates the percentage modulation of the carrier corresponding to the various light values. The region from 75 to 100 per cent is reserved for scanning impulses; therefore, the 75 per cent point is the blackest



(A) Subject and modulation.

(B) Carrier envelope produced by video modulation in A.

(C) Subject and modulation.

(D) Carrier envelope produced by modulation in C.

Fig. 9-7. Video signal and carrier envelope produced by scanning a range of tones.

part of the picture. The region beyond this black point is known in television slang as blacker than black and in engineering circles as the infrablack region.

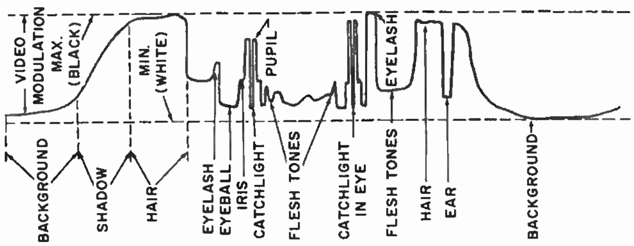
At the other end of the modulation scale (from 10 to 15 per cent of maximum amplitude) is found the brightest highlight, or white region. FCC regulations specify that this white level shall not exceed 15 per cent of maximum carrier amplitude. One basic method of receiver design, known as intercarrier sound, requires a white level of at least 10 per cent of maximum carrier. For this reason, television broadcast stations must now hold the white level within quite narrow limits.

The intermediate grey tones fall between these two extremes (Figs. 9-7A and 9-7C). Fig. 9-7A shows a series of tones in which the white portion of the picture is at the left and the black portion is at the extreme right. Fig. 9-7C shows a succession of white through black to white again; the black level or pedestal is reached at the center of the scan. The television video-carrier envelopes that would be produced by the video-modulating waves of Fig. 9-7A and 9-7C are shown in Figs. 9-7B and 9-7D.

An actual television subject will be represented by continuously varying tone values along each of the 470 to 480 scanning lines. Fig. 9-8A illustrates a typical television image. In this figure, however, the line structure has been produced by approximately 75 lines rather than by the 480 lines of an actual transmission. One of these horizontal lines, which crosses all the tone values of the subject, has been analyzed for change of video signal due to the major differences in reflected light intensity from the subject. The variation of camera signal due to scanning of line X-X causes the video-modulating voltage in Fig. 9-8B. The sudden changes caused by crossing the eyeball, iris, and pupil represent rapid changes of the video voltage. For faithful reproduction of such sudden changes, the system must have extremely wide-band response characteristics. Gradual variations, like those represented by the slight variations in the flesh tones of the face and by gradual transitions of the shadow in the background, can be accommodated by middle frequencies of the video band.

A number of times in our discussion of video modulation, we have indicated that an increase of carrier strength produces a darker picture until

(A) Television subject.



(B) Video voltage.

Fig. 9-8. Variation of video voltage produced by scanning line X-X.

the spot on the picture tube is finally extinguished and a black dot is produced. This is known as negative modulation. Negative modulation was chosen by the National Television System Committee (a radio industry group which acted as an advisory committee to the FCC) after a trial of positive and negative systems in the laboratory and experimental field demonstrations. The reasons for the choice of negative transmission as the United States standard are:

1. Static and noise in the receiver and man-made noise from automobiles, oil-burner ignition, electric razors, etc., produce black spots on the picture in negative transmission, but would produce bright flashes of light if the video modulation were positive. The former condition is the least annoying.

2. In the negative system, both horizontal- and vertical-synchronizing pulses are at maximum carrier ampltude. Thus, receiver synchronization with the transmitter is assured even during low signal strength at the limits of the service area.

Unlike radio broadcasting, which can be received by many types of systems from a simple crystal detector to a complicated superheterodyne, television is absolutely controlled by the choice of type of transmission. For this reason, the final decision of the FCC on television standards was delayed for many years pending industry accord.

THE DIRECT-CURRENT COMPONENT OF THE VIDEO SIGNAL

Video modulation differs from the audio modulation employed in sound broadcasting because it is a varying direct current rather than the familiar alternating current of speech and music modulation.

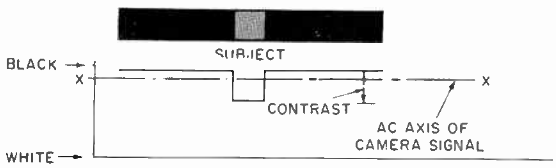
The direct-current component, or bias, corresponds to the average illumination of the scene being televised. This is equivalent to an average of the camera output for all the lines comprising a frame-scanning interval. The camera tube produces an alternating voltage output which is proportional to the variation in brightness of the parts of the picture being scanned. Since the output of the camera tube (such as the mosaic of the Iconoscope) is capacitively-coupled to the input grid of the camera amplifier, any direct-current components of the camera-tube output cannot by passed on to the succeeding stages. The remaining video-amplifying stages of the transmitter are also capacitively-coupled and, therefore, cannot amplify the direct-current component.

Direct-current components in the video signal are due to those portions of the scene being televised which have no change in brilliance over part of the horizontal line. A uniform grey background, for instance, produces a uniform single value charge on that part of the mosaic representing its image. As the scanning beam of the camera tube crosses this section, the output voltage does not change; consequently, no alternating voltage is passed on to the camera amplifier input.

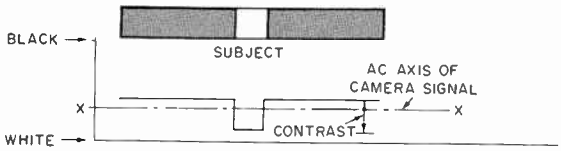
If the average lighting of the scene or background were not taken into account by adding the correct DC bias to the video modulator, the contrast between the various parts of the picture would be correct, but the background illumination or shading would not.

Several instances of typical subjects will illustrate the effect on the received image if the average or DC component were not transmitted.

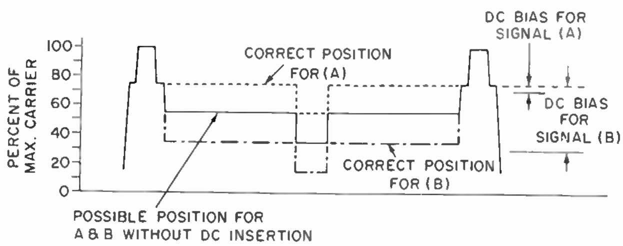
1. A dancer dressed in a white costume and with a black curtain as a background is chosen as the subject. Assume that the television system, from camera tube to the receiving picture tube, has been so adjusted that the dancer's costume is rendered on the picture tube as a satisfactory highlight and the black curtain appears as black. What will happen if several more dancers, similarly garbed in white, enter the scene? If the DC component representing the new average light value of the scene is in the transmited carrier and is restored at the picture tube, the scene will be reproduced correctly. If the new DC bias were not provided at the transmitter, the highlights of the dancers would be grey, and any grey areas in the background would disappear into the black area beyond cutoff.



(A) Scanning a grey on a black background.



(B) Scanning a white on a grey background.



(C) Video modulation.

Fig. 9-9. Examples of why DC component of video signal is needed.

2. As an opposite extreme, consider a hockey arena as the subject. The system has been set up for proper rendition of the ice as the highlight value. If the opposing teams now skate into the field of view, this large area of dark figures will degrade the highlight tones of the ice and will not be as dark as the actual scene contrasts would require.

To understand the reasons for the effects described in the two examples, we chose two types of subject material which, in the absence of the DC component, would produce the same video signal and cause confusion in the reproduced image.

Fig. 9-9A shows the video signal from the camera amplifier which is caused by a grey line being scanned on a black background. Since no DC component is present, the alternating current is averaged about the line X-X. Fig. 9-9B shows the camera amplifier output when a white line across a grey background is scanned. The contrast between the line and the background is the same in each instance, and the AC video modulation would be the same. When the video signal in each is referred to the black level (Fig. 9-9C), the light values are placed at their proper points on the scale, and there is no confusion. The direct-current component caused by the average value of the background now differentiates between the grey line on the black background and the white line on the grey background.

The proper value of DC bias can be produced at the transmitter in several ways:

1. It can be added directly to the modulation circuit by a manually operated control. This control is monitored by an observer, who compares the image on a television screen with the actual scene being televised and continually adjusts the background or average lighting.

2. An auxiliary camera operated with the pickup camera uses a photoelectric cell to integrate the light of the scene and automatically provide a DC bias for the background level.

3. Certain camera tubes like the Image Dissector are nonstorage devices employing a direct-coupled output and can produce a direct-current component representing the brightness of the scene. This type of tube is ideally suited for transmitting directly from motion-picture films.

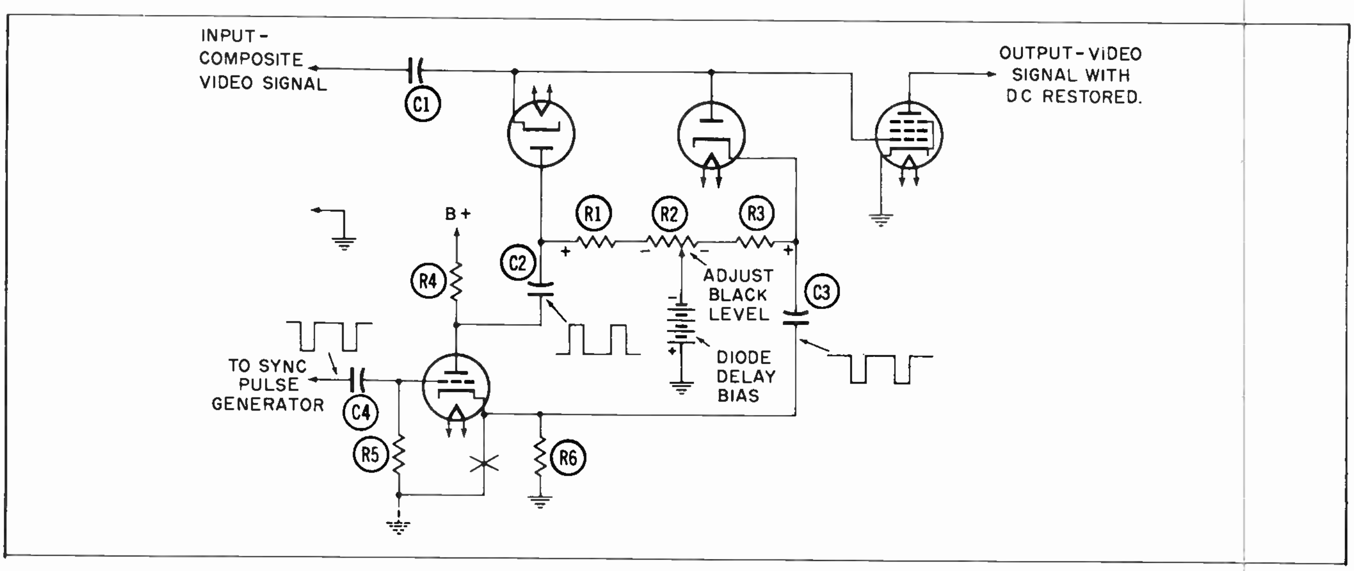


Fig. 9-10. Typical double-diode clamp circuit for reinserting DC at transmitter.

4. Orthicon camera tubes, as well as "flying-spot" scanners, can also accomplish DC insertion automatically.

In all these methods of inserting the DC component or bias level, the end result is the establishment of a fixed DC value for the black level of each frame. In television transmitters, the DC component is often added to the video signal in the camera amplifiers, but is subsequently lost by passing the signal through higher level capacitively-coupled video amplifiers. The DC component is then reinserted by rectifying the peak value of the video signal and adding the rectified DC voltage to the signal.

Fig. 9-10 shows a typical transmitter DC reinsertion circuit. The circuit employs a double-diode tube in a bridge network and is known as a double-diode clamp circuit, since it clamps the video signal at the black level. The sync pulses of the video signal establish the level at which this circuit operates. The DC furnished by the circuit for reinsertion is thus referred to a level previously established by the DC component inserted in the camera amplifier.

QUESTIONS

1. What is the range of modulating frequencies encountered in the television video signal?
2. What is the aspect ratio of a television picture?
3. When a change in video-modulating voltage causes a momentary increase in carrier strength at the transmitter, what happens to the brightness of the spot on the picture tube screen at the receiver?
4. How many megacycles wide is each television channel?
5. Is double-sideband or vestigial-sideband modulation used in transmitting the video signal?
6. Where is the picture carrier placed in the allotted 6-mc channel? What is the limit of the upper sideband?
7. What is the maximum percentage of video modulation? What is the specified white-level limit?
8. Are the horizontal-sync pulses transmitted at maximum or at minimum carrier amplitude?

Chapter 10

Sync-Pulse Separation, Amplification, and Use

Sync-pulse separation has been mentioned in connection with deflection systems. At this time we will consider in greater detail the separation, amplification, and use of sync pulses. Before starting such a study, let us review some of the characteristics of vacuum tubes used as amplifiers and rectifiers.

REVIEW OF VACUUM TUBES APPLIED TO THE SEPARATION AND USE OF SYNC PULSES

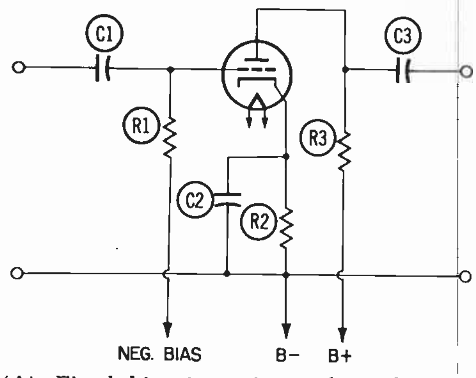
In commercial receivers, we find circuits in which the sync pulses are separated from the composite video signal by diodes, triodes, and pentodes. The following review of vacuum tubes, together with a study of the characteristic curves and data in receiving-tube manuals, will be helpful. These manuals are available from the tube manufacturers.

1. The limits of the plate-current range of a vacuum tube are determined by cutoff and saturation. In triodes and pentodes, plate current can be cut off by a fixed negative grid bias or by negative bias from grid-circuit rectification. A series delay bias can produce cutoff in diodes. Plate-current saturation is produced by operation at a low plate voltage with triodes and at low plate and screen voltages with pentodes.

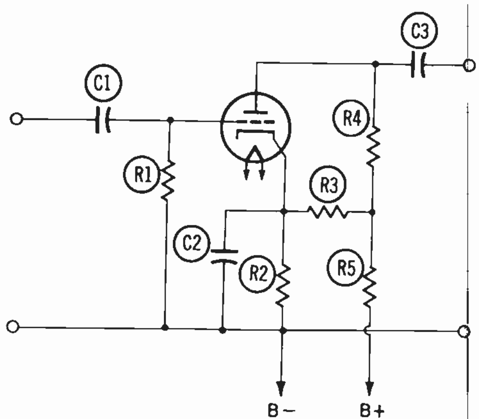
2. The voltage across a cathode resistor due to plate-current flow provides a negative bias if the grid circuit is returned to the end of the resistor opposite the cathode. A tube biased by its own plate current cannot cut itself off. Self-bias from the cathode resistor is often augmented by additional negative bias in series with the grid return or by bleeding current through the cathode resistor from the plate supply to assure that the grid bias will be cut off (Fig. 10-1).

3. A change in the control-grid voltage will change the plate current in the same direction or polarity. Stated another way, we can say that grid voltage and the plate current are in phase.

4. The plate-voltage change caused by a change of grid voltage is in the opposite direction. Hence, the plate voltage is 180° out of phase with the grid voltage. For this reason, a vacuum-tube amplifier with grid input and plate output is essentially a phase inverter.



(A) Fixed bias in series with grid-return resistor.



(B) Additional cathode bias from power supply.

Fig. 10-1. Cutoff by combination bias.

5. The drop in plate voltage caused by a positive pulse at the grid is known as a negative-going pulse. As we have seen, such a pulse is required to control a cathode-coupled multivibrator.

6. The increase in plate voltage caused by a pulse which drives the grid more negative than its initial bias condition is known as a positive-going pulse. This action is required to trip a blocking oscillator.

7. A plate-load resistor decreases the plate-to-cathode voltage as plate current increases. Conversely, with such a load, the plate-to-cathode voltage will increase as plate current drops.

8. The amount of change in plate-to-cathode voltage caused by a change in plate current can be transferred by a coupling capacitor to the grid circuit of a following tube with no change in polarity.

9. When the load resistor of a vacuum tube is in the cathode circuit of the tube and the output is taken from the cathode (Fig. 10-2), the circuit is

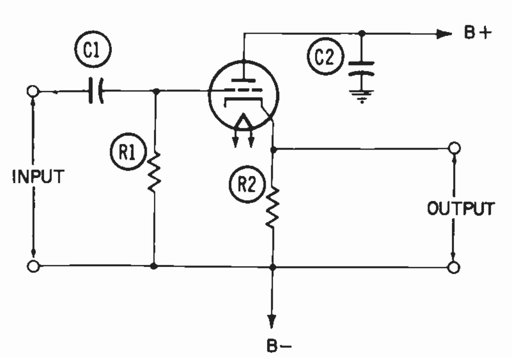


Fig. 10-2. The basic cathode-follower circuit.

known as a cathode follower. This type of circuit was employed in some early television receivers.

SYNC-PULSE SEPARATION

The horizontal- and vertical-synchronizing pulses and their time relationship to the scanning sawtooth were described and illustrated in Chapter 7. These pulses occur when the electron stream in the picture tube is cut off. How the sync pulses are separated from the video signal and used to control the horizontal and vertical-scanning systems of the receiver will be discussed. Specifically, we shall consider the separation of the horizontal pulses occurring at the end of each of the 525 lines in the picture. The vertical pulses will be covered when we consider the methods of segregating the vertical pulses from the horizontal pulses.

The pulses can be clipped from the signal at three places in the circuit:

1. At the video-detector input.
2. At any of the video-amplifying stages.
3. At the point of restoration of the average background light of the picture.

Fig. 10-3B shows the video signal with its picture and synchronizing information as it appears at the input of the video detector (Fig. 10-3A). This signal consists of horizontal pulses (as detailed in Fig. 7-1) mounted on a shelf or pedestal. These are identified in Fig. 10-3B as (1) the pulse and (2) the pedestal.

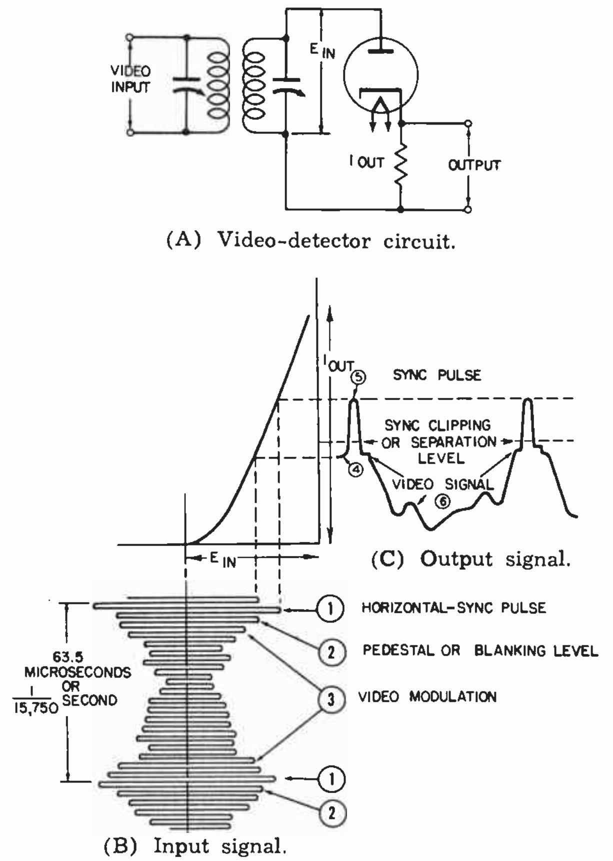


Fig. 10-3. Demodulation (detection) of the video (picture) carrier.

Between the edges of the pedestals, the video-modulated envelope (3) of the carrier is found. This envelope represents the variations in light and dark of the video signal that modulates the cathode-ray beam. Fig. 10-3C shows the form of the detected or demodulated wave; the sync pulse and pedestal are at (5) and (4), respectively. The picture information or video signal is shown in the part of the wave at (6).

Note that the sync pulses are at the top of the signal. Since these pulses occur when the picture tube is black, the darker tones of the picture are just below the pedestal, or point (4), of Fig. 10-3C. The signal, after it passes through the video amplifier, must reach the grid of the picture tube in such a phase that the sync pulses are the most negative part of the wave. This action accomplishes blanking during the return trace. We have seen that the polarity of the sync pulse, as it arrives at the grid of the scanning oscillator, must be of the proper polarity to assure control. For this reason, the method of sync-pulse separation selected in any particular receiver design will depend upon (1) the number of stages of video amplification, (2) the point in the circuit at which sync separation is accomplished, and (3) the number of sync-amplifier or clipping stages employed.

All the methods of separating the synchronizing pulse from the rest of the video signal involve the fact that, in transmission, the pedestal or blanking level is always maintained at a definite point on the carrier wave (75% of maximum carrier). Therefore, the sync pulses occupy the top 25% of the wave. The problem of sync clipping thus resolves itself into one of amplitude separation or of removing the top 25% of the wave without passing the lower values containing the video signal. The methods commonly employed will be covered under the types of tubes used.

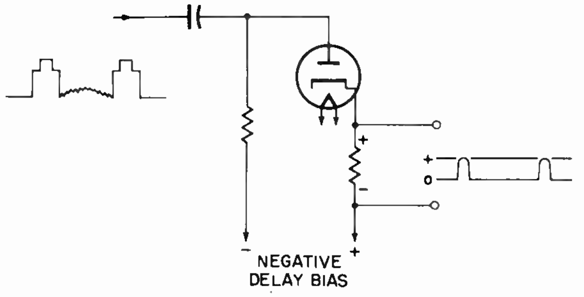
Diode Sync-Separation Circuits

Fig. 10-4 shows three diode sync-separation circuits. The input signal of each circuit is the composite video signal, and the output is a pulse signal that represents the sync pulses.

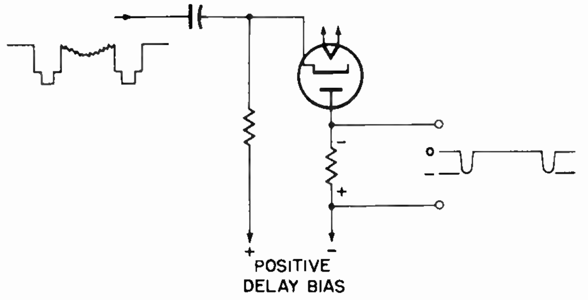
In Fig. 10-4A the video signal is coupled to the plate of the diode. The output signal is developed across a resistor in the cathode circuit. In order that the diode can be kept at cutoff until sync-pulse time, the plate is connected to a negative delay-bias source. When the highly positive sync pulses come along, the delay bias is overridden and the diode is allowed to conduct. During this time, current flows through the cathode resistor and produces a voltage in the polarity shown. The output taken across the cathode resistor is in the form of positive pulses.

Fig. 10-4B is an inverted version of the same circuit. In this circuit the video signal is coupled to the cathode and the output signal is developed at the plate. Cutoff is accomplished by a positive delay bias applied to the cathode. The input signal is of opposite polarity to that applied to the diode in Fig. 10-4A. The sync pulses are highly negative. When the sync pulse comes along, the positive delay bias is overridden and the diode conducts. Current flows through the plate resistor and produces a voltage in the polarity shown. The output taken across the plate resistor is in the form of negative pulses.

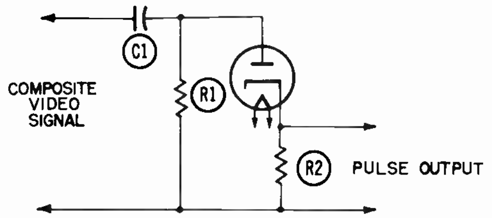
The circuit in Fig. 10-4C is similar in operation to the one in Fig. 10-4A, except that the bias required for delay of the diode action to the correct point is obtained from the charge placed upon C1 during diode conduction. The time-constant used in bias circuit R1-C1 is long compared to the horizontal and vertical scanning times. Thus, the diode can be biased automatically by the video signal to the proper point, so that the sync pulses are clipped from the signal.



(A) Positive sync pulses with delay bias on diode.

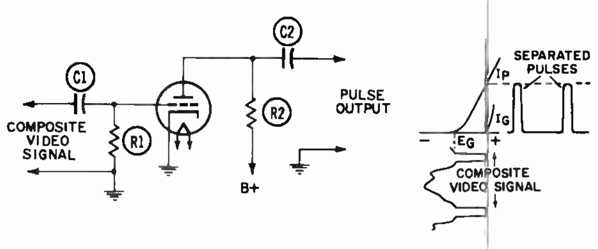


(B) Negative sync pulses.

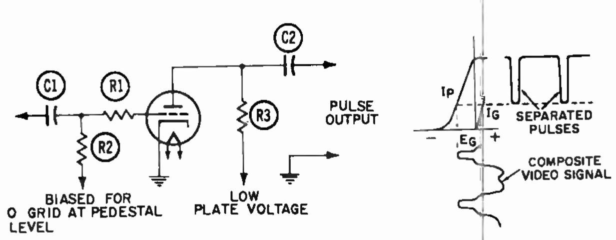


(C) Positive sync pulses with self-biased diode.

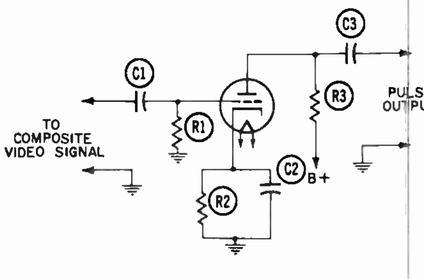
Fig. 10-4. Diode sync-separation circuits.



(A) Grid cut off by signal rectification.



(B) Separation by positive grid and plate-current saturation.



(C) Cathode-bias circuit.

Fig. 10-5. Triode sync-separation circuits.

Triode Sync-Separation Circuits

In many receivers, triodes or pentodes are employed, rather than the diode circuits just discussed. The choice of tube type depends upon the sync-pulse amplitude requirement, polarity requirement, and point in the circuit at which separation is accomplished. Generally, these multi-element tubes allow some voltage gain to be realized in the separation process. In certain circuits, leveling or limiting can also be realized. Three basic types of triode sync-separating circuits are shown in Fig. 10-5.

The circuit of Fig. 10-5A uses grid rectification of the video signal to bias the control grid, so that plate-current cutoff occurs at the desired pedestal level. This action is similar to the one just discussed for the diode circuit in Fig. 10-4C. Two additional actions are found in this circuit:

1. The sync pulses are large enough to drive the grid positive, and the lowered grid resistance limits the input signal by loading.

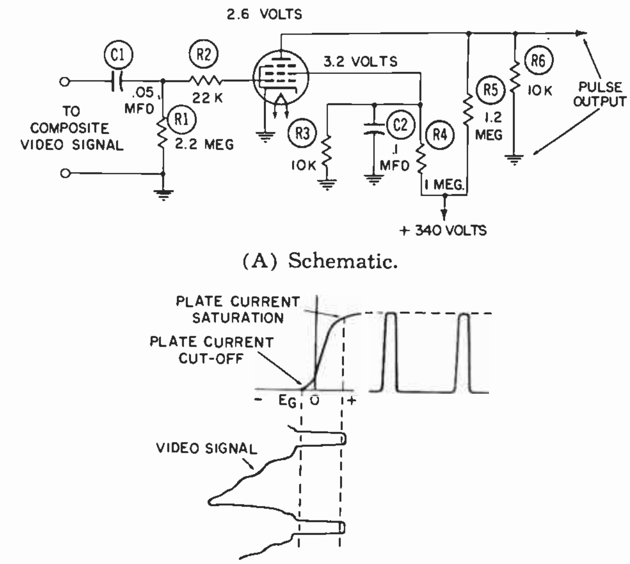
2. Some amplification of the sync pulse occurs because of the amplifying properties of the triode.

In Fig. 10-5B the operating conditions are quite different from those of the circuit in Fig. 10-5A. The tube is biased from an external voltage source, through resistors R1 and R2, until it just starts to draw grid current at the black or pedestal level. The input signal is inverted in polarity from that of the preceding example. The sync-pulse portion of the video-input signal is the most negative. The plate voltage is made so low that the plate current is saturated at a near zero grid voltage. The portion of the video-input signal which is more positive than the desired clipping level lies in the saturation region and produces no further rise in plate current. For this reason, limiting or leveling occurs at this saturation point. The negative grid voltage from the sync-pulse portion of the input signal causes a drop in plate current, as shown in the waveform drawings of Fig. 10-5B. So that the circuit action can be better illustrated, the amplitude of the sync input pulses has been limited in the drawing. If their amplitude were extended to beyond the grid cutoff point, limiting would also occur because of plate-current cutoff, and the output pulses would still be uniform but larger in size. Series resistor R1 in the grid circuit limits the amount of grid current that can flow over any one frame. This limiting prevents a long-time blocking condition from developing because of an excessive charge on capacitor C1.

The circuit in Fig. 10-5C employs cathode bias to establish the correct operating point for plate-circuit separation of the sync pulses. The values chosen for resistor R2 and capacitor C2 are determined by the following considerations:

1. The resistor must have such a value that the plate-current pulses above the clipping level will produce a voltage drop equal to the required operating grid bias. For high-mu tubes, this resistor will be around 10,000 ohms.

2. The value of capacitor C2 must be high enough to maintain constant bias voltage throughout at least one vertical blanking period and yet low enough that it can change its charge as the average background lighting of the scene changes.



(B) Input and output waveforms.

Fig. 10-6. Pentode sync-separating circuit that incorporates leveling or limiting.

Pentode Sync-Separation Circuits

Fig. 10-6 illustrates the use of a sharply cutoff pentode as a combination sync separator and limiting amplifier. Plate-current saturation is assured by operating both the screen and plate at extremely low voltages. This low voltage is accomplished with dividing networks. The screen network consists of resistors R3 and R4. With a B+ voltage of 340 volts, the screen is held at the extremely low voltage of 3.2 volts. The plate-supply network of resistors R5 and R6 maintains the plate at 2.6 volts. Plate-current saturation occurs just after the grid voltage goes positive, as shown in the diagram in Fig. 10-6B. The grid circuit network, resistors R2 and R1 with capacitor C1, establishes the operating point of the circuit by grid-circuit rectification of the video signal. This rectification assures plate-current cutoff just above the pedestal or black region. You can see from the drawings that the synchronizing pulses are clipped at both ends, and limiting occurs. Because a pentode under these conditions exhibits very low voltage gain, a circuit of this type must either be operated at a high signal level or be followed with sync amplifiers.

Series grid resistor R2 has a function similar to that of the resistor R1 in the positive grid triode of Fig. 10-5B. However, R2 in Fig. 10-6A has an additional function. In the absence of resistor R2,

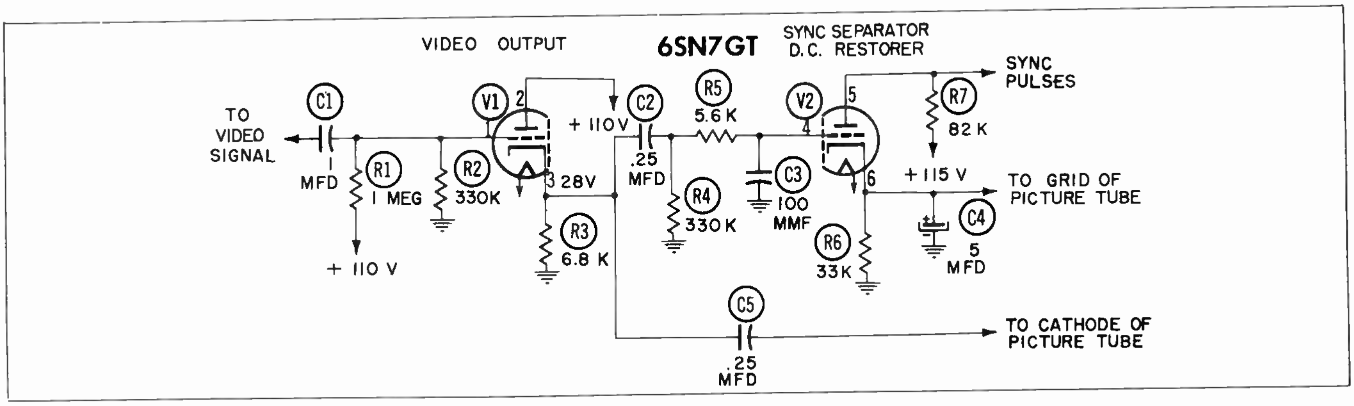


Fig. 10-7. Cathode-follower used with sync-separator.

sharp noise pulses (such as those due to motor-car ignition, whose amplitude might be higher than that of the sync pulses) would cause an excessive bias voltage on C1 and result in blocking. Loss of synchronization would occur while the tube was blocked.

Cathode-Follower Used with Sync-Separation Circuits

Fig. 10-7 shows a circuit which employs a cathode-follower ahead of a sync separator. The cathode-follower tube V1 operates on a linear portion of the curve. This is due to the balance of the bias across resistor R3 and the voltage from the plate supply produced by network R1 and R2. A net negative voltage of −2 volts appears between grid and cathode.

Cathode-follower tube V1 functions both as video output and sync takeoff. The video output for the picture-tube cathode is taken directly from cathode load resistor R3 through capacitor C5. This same load resistor feeds a triode (sync separator and DC restorer) tube V2 through the network of C2, R4, R5, and C3. The combination of R5 and C3 filters the high-frequency components of the video signal.

Sync pulses are taken from the plate of triode V2, the operation of which is similar to the circuit in Fig. 10-5C. The DC bias voltage across network R6-C4 is used to restore the average component of the picture signal responsible for the average light of the televised scene. This DC component has been lost in passing through video-amplifier stages.

Gated Sync-Separation Circuits

The sync separators which have been described have one feature in common—the output signal is low in amplitude. Additional stages of amplification (and sometimes clipping) are needed to produce a sync signal of sufficient amplitude to trigger the sweep oscillators. A simpler type of separator circuit, called the gated sync separator, has been developed. This circuit eliminates the extra amplifier stages because its tube is usually a pentagrid, which supplies many times more gain than a triode.

The schematic of one type of gated circuit is presented in Fig. 10-8. This circuit is primarily designed to stabilize sync operation in weak and noisy signal areas, and operates as follows:

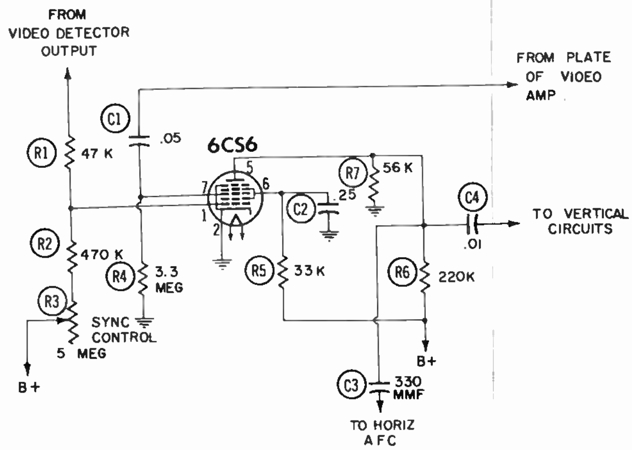


Fig. 10-8. Schematic of a gated-sync circuit.

The composite video signal is sampled from the plate circuit of the video amplifier and fed to the grid (pin 7) of the 6CS6 tube through capacitor C1. The positive-going signal on pin 7 is of high amplitude and causes grid current to flow; therefore, capacitor C1 charges to approximately the blanking level of the composite video signal. The bias developed on pin 7 permits the tube to conduct only on signal peaks which are greater in amplitude than the blanking level. This condition allows only the horizontal- and vertical-sync pulses to appear in the plate circuit.

A noise-cancellation action is provided along with this sync-separation action. A fixed bias is developed on the grid (pin 1) of the 6CS6 tube by resistors R1, R2, and sync control R3. The bias is set by adjusting the sync control, so that the tips of the sync pulses will fall near the cutoff point of the tube.

When part of the composite video signal containing negative-going sync pulses from the video-detector output is impressed on the grid (pin 1) through resistor R1, any noise pulse of sufficient amplitude will cancel the effect of a simultaneous positive pulse on pin 7 by cutting off the tube and will not be present in the plate circuit of the stage. Noise pulses on pin 1 during the sync-pulse interval will also cut off the tube; but since the noise pulses usually have a shorter time duration than the sync pulses, the stability of the vertical and horizontal oscillators will not be affected.

A different circuit is included in some television receivers. It is built around one tube which does triple duty as a sync separator, AGC amplifier, and noise inverter. The most unusual feature of this circuit is that the inverter removes noise pulses from the AGC signal as well as from the sync signal. The AGC amplifier circuit will be described in a later chapter.

A unique tube, the 6BU8, has been used in this circuit (Fig. 10-9). The tube might be called a "Siamese-twin" pentode. The cathode, the first control grid, and the screen grid are common to both sections of the tube. Outside the screen grid,

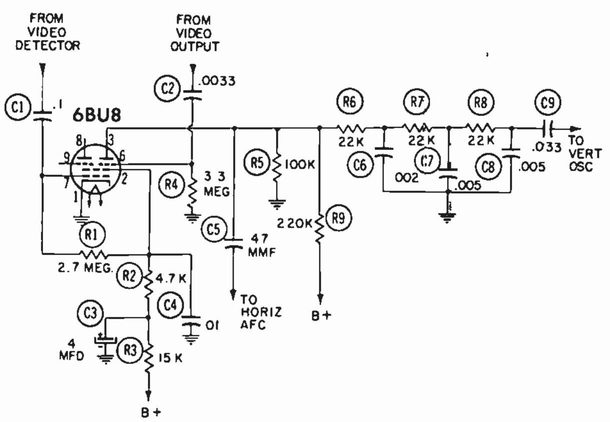


Fig. 10-9. The 6BU8 gated-separator circuit.

there is a second control grid, which is split into two distinct sections; and there are two separate plates in line with these two control grids. The currents in the two plate circuits can be controlled simultaneously by the first control grid or separately by the pair of second control grids.

The cathode is grounded, so that variations in the plate current of one section of the tube will not be coupled to the other section. The sync-separator output is taken from the plate connected to pin 3, and the AGC output is obtained from the other plate. The first control grid (pin 7) is employed as a noise inverter.

A composite video signal containing positive-going sync pulses is coupled from the video-amplifier plate through C2 to one of the second control grids (pin 6) of the 6BU8 tube. Grid-leak bias is developed by C2 and R4. Plate current in the right-hand section of the 6BU8 is cut off, unless a sync pulse is on the grid. The waveform of plate voltage measured at pin 3 of the 6BU8 is typical of the output of a sync separator. Negative-going sync pulses are developed at this plate. They are applied to the horizontal AFC tube and vertical oscillator without further amplification. The vertical oscillator used with this circuit is a multivibrator and requires a negative sync pulse for proper triggering.

The first control grid of the 6BU8 tube receives a composite video signal from the video-detector output. This signal has the same waveform as the sync input signal on pin 6, but is opposite in polarity. One might assume that the two signals would cancel each other. Actually, the signal on the first grid has such a small amplitude that it has only a slight degenerative effect upon the output of the tube. The action of the first grid becomes important only when noise is in the input signals.

The first control grid is biased so that the tips of the negative-going sync pulses will almost drive the grid into cutoff. Noise pulses in the composite video signal sometimes have a greater amplitude than the sync pulses. When one of these bursts of noise reaches the first control grid, the grid voltage is driven below cutoff, and all conduction within the 6BU8 tube ceases momentarily. At the same instant, a positive noise pulse appears at the second control grid, but the temporary interruption of plate current prevents a corresponding pulse from developing at the plate.

If the noise inverter were not in the circuit, strong noise pulses would get into the output signal of the sync separator. They would tend to cause random triggering of the sweep oscillators, and unstable synchronization would result.

SYNC-PULSE AMPLIFICATION, CLIPPING, AND SHAPING

Many television receivers employ more than a one-tube circuit to separate the sync pulses from the video signal. Additional stages are introduced to (1) invert the phase of the pulses (when not of the proper polarity to control the scanning oscillator), (2) clip the pulse width (for more reliable scanning control), (3) amplify the pulse (if it is not strong enough for control), and (4) level the pulse (to take care of variations in the video signal and minimize the effect of interfering noise pulses).

There has been little standardization in naming these stages, and we find the following descriptive titles in the service literature of various manufacturers: "sync clipper," "pulse stripper," "sync amplifier," "sync inverter," "sync leveler," "sync limiter," "pulse limiter," and "clamper." The various actions are self-explanatory; but, it should be noted that, even though a stage is labeled as a sync amplifier, it is usually so biased that either cutoff or saturation contributes leveling or clipping as well as the desired voltage amplification.

SORTING OF THE INDIVIDUAL HORIZONTAL AND VERTICAL PULSES

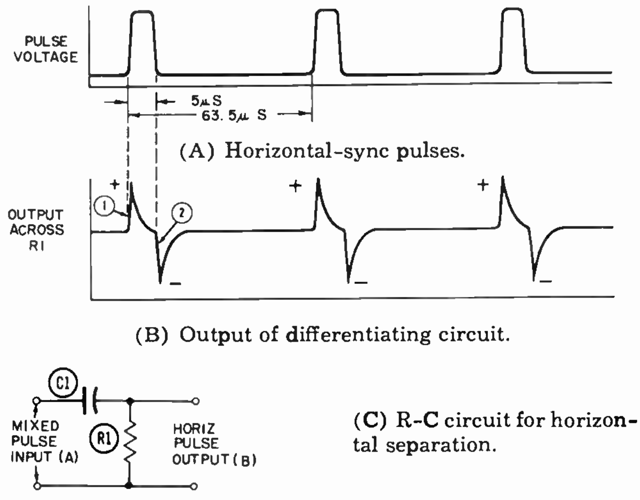
In the foregoing description of the various methods of separating the synchronizing pulses from the composite video signal, only the narrow horizontal pulses were mentioned. The longer vertical pulses are clipped from the signal in the same separation process.

After the sync pulses have been removed from the video signal, the vertical pulses must be sorted from the horizontal pulses; and each one must be fed to its respective deflection-scanning system. Since the horizontal and vertical pulses are equal in amplitude, the methods of separation for clipping them from the video signal cannot be used to distinguish between them. They do, however, differ in time duration. It is on this basis that sorting is accomplished.

Several times, we have mentioned differentiating networks for removing horizontal pulses and integrating networks for the vertical pulse acceptance. We will now consider the action of such systems in greater detail.

Horizontal-Pulse Separation

The horizontal pulses of the transmitted signal are approximately five microseconds in duration, as shown in Fig. 7-1. These pulses are impressed on a circuit of the type shown in Fig. 10-10C, which is known as an R-C differentiating circuit.



(A) Horizontal-sync pulses.

(B) Output of differentiating circuit.

(C) R-C circuit for horizontal separation.

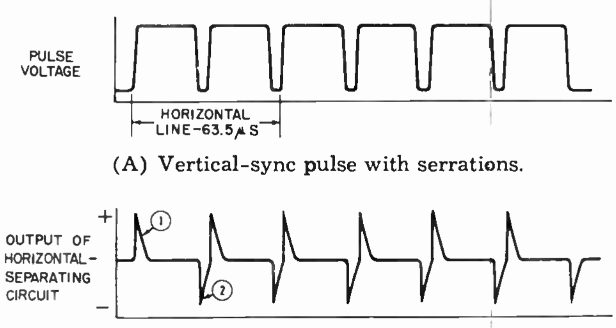
Fig. 10-10. Horizontal-pulse separation, or differentiation.

Differentiation means the breaking down of a quantity into a number of small parts. The pulses in Fig. 10-10A are made into smaller parts, as shown in Fig. 10-10B, by the action of the circuit in Fig. 10-10C. The circuit consists of a capacitive and resistive combination in which the capacitor is in series with the separated pulse input and the resistor is shunted across the output. The time constant of this circuit is made short compared with the duration of a horizontal-sync pulse. The sync pulse is held between 4 and 5 microseconds, and the time constant of the horizontal differentiating circuit is made between 1 and 2 microseconds. As described and illustrated in Fig. 5-7 (an R-C circuit in which the time constant is short compared with the duration of the applied square-wave pulse), the capacitor is completely discharged. A sharp pip of voltage occurs across the resistor at both the leading and trailing edges of the applied square-wave pulse.

The amplitude of the pip is determined not only by the amplitude of the square wave but by the steepness of the edge of the square wave. For this reason, the FCC limits the allowable slope of the leading and trailing edges. These slopes must not occupy more than 0.4 per cent of the horizontal-line scanning interval of 63.5 microseconds.

The voltage pip due to the leading edge of the horizontal-synchronizing square wave is shown as a positive pip at (1) in Fig. 10-10B. The dip due to the trailing edge of the horizontal pulse is shown as a negative voltage at (2). The leading-edge pulses are the ones which control the horizontal-scanning oscillator. The negative pulses are rejected by cutoff or saturation of one or more stages of the sync system.

When the longer duration vertical synchronizing pulses arrive, the differentiating circuit acts as shown in Fig. 10-11. Here again, a positive pip



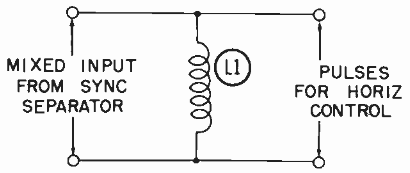
(A) Vertical-sync pulse with serrations.

(B) Output of horizontal-differentiating circuit.

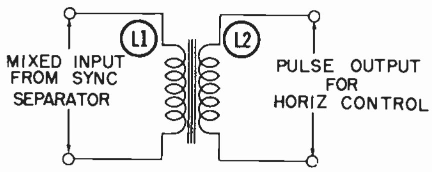
Fig. 10-11. Action of horizontal-differentiating circuit on vertical-sync signal.

occurs at the leading edge of each vertical pulse, and a negative pip occurs at the trailing edge. The leading-edge pulses continue to control the horizontal oscillator during vertical retrace. In this instance, however, two pulses occur during a horizontal line-scanning interval. Only the first of these pulses is used to control the horizontal oscillator. The second pulse cannot cause lock-in, since it occurs while the oscillator is insensitive to tripping.

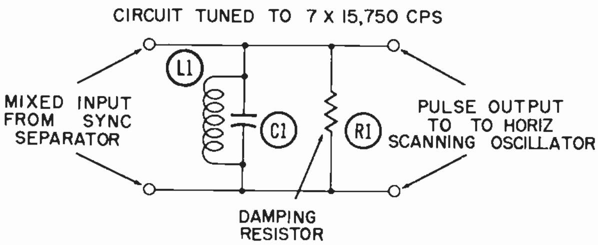
The horizontal pulses can be separated by other means than the R-C differentiating circuit just described. Fig. 10-12 shows two types of differentiating circuits which employ inductance and a third type which uses the properties of a resonant circuit. The inductance of the circuit in Fig. 10-12A is connected in series with the plate circuit of a tube which has been biased to clip the sync pulses from the video signal. The waveform of the sync-pulse plate current consists of steep slopes which



(A) Differentiation by self-inductance.



(B) Differentiation by mutual inductance (transformer action).



(C) Horizontal separation by tuned circuit action (one-half cycle of ringing employed).

Fig. 10-12. Other methods of horizontal sync-pulse separation.

correspond to very rapid changes of current. The voltage across the inductance is proportional to the rate of change of the current through it. Thus, at the leading and trailing edges of each plate-current pulse, a high voltage is produced across the inductor. This voltage is the same form shown for the R-C type of differentiator (Fig. 10-10B). If the pulses are of proper polarity and sufficient amplitude, they can be applied directly to the scanning generator by a capacitor connected to the plate end of the inductor. If the polarity is incorrect, phase reversal can be accomplished by an amplifier stage or by a transformer, as shown in Fig. 10-12B. When a transformer is used, secondary L2 may be connected so that the output voltage pulses have opposite polarity to those across primary L1. The secondary can be connected directly to the grid circuit of the horizontal-scanning generator.

The circuit shown in Fig. 10-12C operates quite differently from the two just described. The resonant circuit, consisting of L1 and C1, is tuned to approximately seven times the horizontal-line frequency of 15,750 cycles, or 110 kilocycles. The separated sync pulses are impressed across the circuit and shock-excite it into oscillation at its resonant frequency. The oscillation is quickly damped out by parallel resistor R1. Only the first half-cycle of voltage across the circuit is used to control the horizontal-scanning oscillator. This corresponds to a pulse duration of approximately 5 microseconds. Several advantages can be cited for this method of horizontal-sync discrimination:

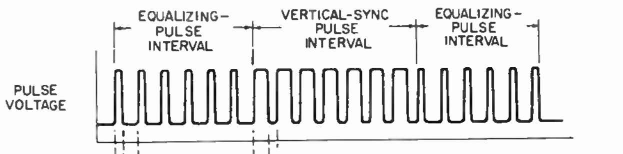
1. An extremely simple pulse-separation and oscillator-control system can be used. The circuit can be connected directly in the plate return of the sync-separator tube and coupled directly to the scanning oscillator because pulse shaping is performed by the resonant action.

2. This method is relatively immune to excitation by static or ignition noise because such pulses would have to be of the proper time duration (5 microseconds) and repetition rate (15,750 cycles) to produce ringing. The probability of such coincidence is slight.

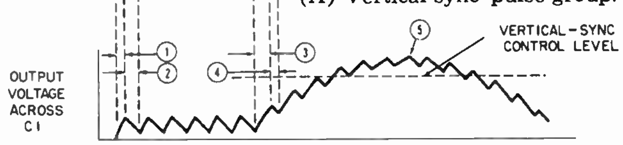
Vertical-Pulse Separation

In the description of vertical-scanning systems, we mentioned integrating networks for segregating the long-time vertical field pulses from the sharp horizontal line pulses. We will now consider the means of sorting these vertical field-scanning pulses from the composite scanning pulses and of using them to control the vertical-oscillator timing.

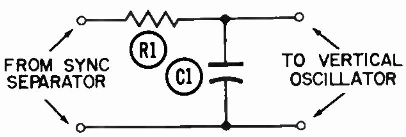
The integrating action which sorts the vertical pulses from the complex video signal is exactly opposite from the differentiation process for separating the horizontal pulses. Integration means the addition of a number of small elements to form a whole. Fig. 10-13C shows an integrating circuit. It is the opposite of the differentiation circuit in Fig. 10-10. The resistor is in series with the input, and the capacitor is connected across the output. The time constant of the combination is much longer than that employed for sorting the horizontal pulses. This time constant is made approximately equal to the duration of a horizontal pulse. Consequently, the charge accumulated by the capacitor because of a horizontal pulse is small and will decay rapidly. This action is shown in Fig. 10-13B. During the time shown at (1), the equalizing pulses produce only a small voltage across the capacitor. This voltage decays to near zero in the interval between pulses, as shown at (2). The much longer vertical-synchronizing pulses produce a greater charge in the capacitor during period (3). This charge does not completely decay during the short serration interval (4). Consequently, each



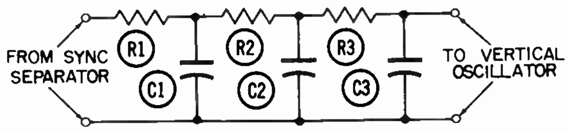
(A) Vertical sync-pulse group.



(B) Output of integrating circuit.



(C) Integrating circuit.



(D) Cascade integrating circuit.

Fig. 10-13. Vertical-pulse separation by integration.

vertical pulse adds an element of charge to the capacitor, and the voltage continues to build up during the interval of vertical pulses. The dotted line in Fig. 10-13B indicates the level at which this voltage becomes large enough to trigger the vertical-scanning oscillator. This point usually occurs after two or three vertical pulses have charged the capacitor.

The vertical-integrating network is seldom the two-element type shown in Fig. 10-13C, but is usually a cascade network, as shown in Fig. 10-13D. The resultant time constant of this network is smaller than that of any of the individual branches (R1-C1, R2-C2, or R3-C3). The over-all time constant calculation is the same as for resistors in parallel. For the three-branch circuit in Fig. 10-13D with T1 for the time constant R1 × C1, T2 for R2 × C2, and T3 for R3 × C3, the effective circuit time constant (T) will be

$$\frac{1}{T} = \frac{1}{T1} + \frac{1}{T2} + \frac{1}{T3}$$

Individual time constants for a three-branch circuit in a modern receiver are 30 to 60 microseconds. The effective over-all circuit time constants are, therefore, between 10 and 20 microseconds.

The reasons for using cascaded integrating circuits are:

1. To prevent erratic control of vertical retrace by random noise or static pulses. Before such pulses could control the vertical oscillator, they would have to be comparable in duration and spacing to the vertical-sync pulses.

2. To smooth out the contour of the rising voltage wave (shown in the interval 3 to 5 of Fig. 10-13B) across the output capacitor. The action is similar to that of the familiar resistance-capacitance, power-supply filter system in which the ripple is reduced by successive stages.

Because of this smoothing action, an individual horizontal pulse cannot cause pairing of lines during retrace. The sections of the cascade network are usually not made with equal time constants. This unbalance prevents accidental triggering by noise pulses.

THE FUNCTION OF VERTICAL-EQUALIZING PULSES

In Chapter 3 we briefly discussed interlaced scanning, which prevents flicker of the image. For simplicity, the retrace from bottom to top of the picture was shown as a straight line or single jump. Actually, the horizontal oscillator must be kept in

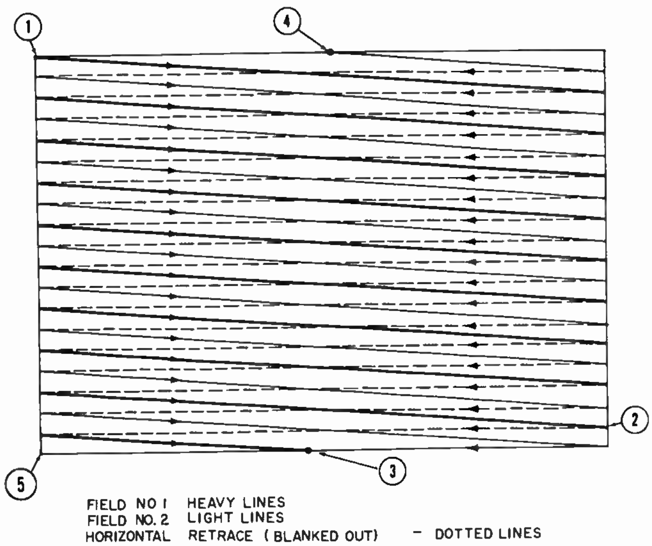


Fig. 10-14A. Active downward fields.

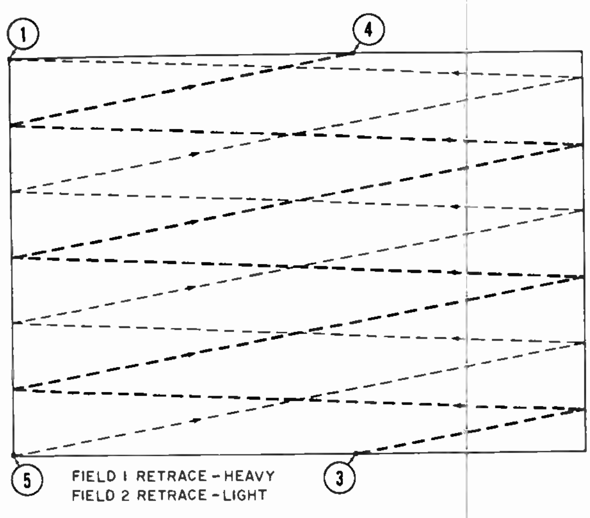


Fig. 10-14B. Inactive upward fields (vertical retrace).

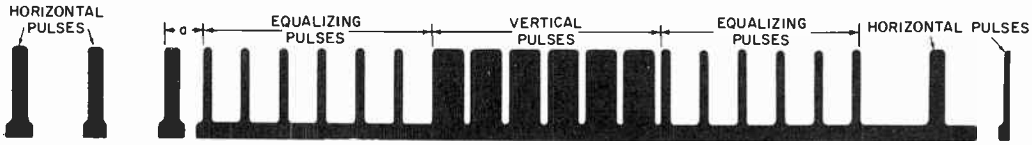


Fig. 10-14C. Vertical-sync signal for retrace after field 1.

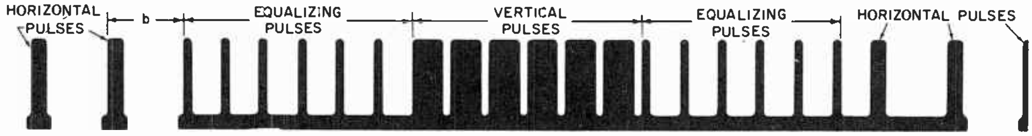


Fig. 10-14D. Vertical-sync signal for retrace after field 2.

step with the transmitter during vertical retrace, which lasts from 1250 to 1400 microseconds (20 to 22 horizontal lines). Fig. 10-14A shows a simplified version of the downward scanning, in which nine and one-half lines have been drawn to represent each field. Actually, a field consists of 262½ lines, less the lines lost during retrace. The first field, which starts at the upper left-hand corner (Point 1) and ends at the bottom center of the picture (Point 3), is shown by heavy lines. The second, or interlaced field, starts at the top center (Point 4) and ends at the lower left-hand corner (Point 5), and is shown by light lines. During vertical retrace, when the picture is blanked out, the beam moves upward under the combined action of both the vertical- and the horizontal-deflection systems. This is represented in simplified form by the diagram in Fig. 10-14B. Here, three lines represent the twenty to twenty-two lines actually required during vertical retrace. Again, a heavy dotted line represents the retrace of field No. 1, and a light dotted line the retrace of field No. 2.

The dual functions of producing vertical retrace at the proper instant and of keeping the horizontal oscillator in synchronism are controlled by the equalizing and vertical pulses shown in Figs. 10-14C and 10-14D. The vertical sync signal for the retrace of field No. 1 differs from that of field No. 2 by the spacing between the last horizontal pulse and the first equalizing pulse. In Fig. 10-14C for field No. 1, this space (a) consists of only one-half of a horizontal line, since field No. 1 ends at the middle of the last line, as shown at Point 3 of Fig. 10-14A. In Fig. 10-14D for field No. 2, the space (b) between the last horizontal pulse and the first equalizing pulse consists of an entire horizontal line. Vertical blanking starts at the leading edge of the equalizing pulses. Thus, the successive field blanking time is accurately set up by the signal.

Even though retrace blanking is accurately established, vertical retrace may not take place at the proper instant unless the critical charge on the integrating capacitor occurs at exactly the same point for each successive vertical-sync signal. How the equalizing pulses assure this condition is shown in Fig. 10-14E. At (1) is shown the composition of a vertical-sync signal which would follow field No. 1 if the equalizing pulses were not present. This signal input to the integrating circuit would charge the capacitor as shown by dotted line (1) on the charge curves of Fig. 10-14E. This curve crosses the sync-control level at time X.

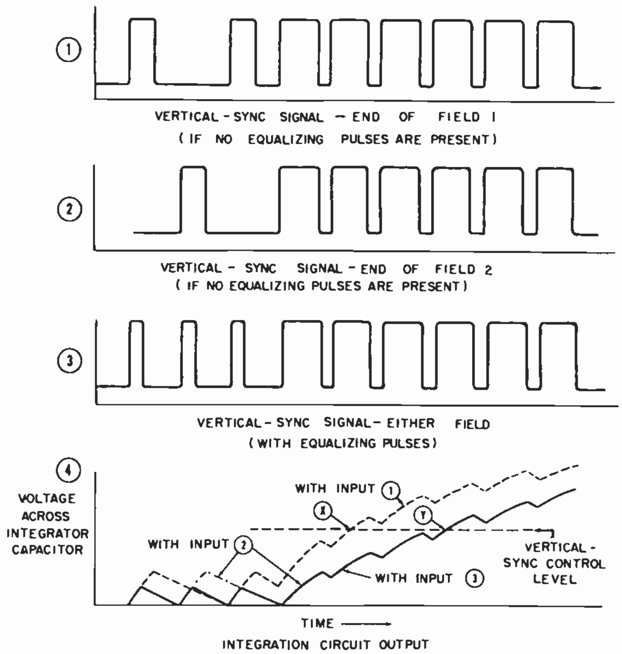
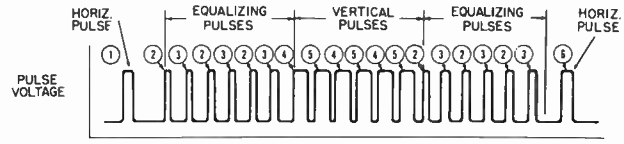


Fig. 10-14E. Action of vertical-integrating circuit for successive fields (with and without equalizing pulses).

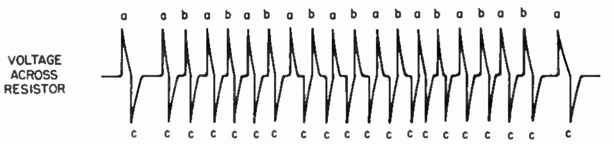
The vertical signal, without equalizing pulses, for retrace at the end of field No. 2, would be as shown at (2) in Fig. 10-14E. On the charge curves, the critical sync-control level would be reached at time Y, which is so much later than time X that proper interlace would not occur. When equalizing pulses are employed as shown at (3), the critical firing point for the vertical oscillator is at time Y for both fields. Successive fields preceded by equalizing pulses will therefore accurately control the oscillator and assure proper interlace.

ACTION OF THE HORIZONTAL-DIFFERENTIATING CIRCUIT DURING THE VERTICAL PULSE

The formation of positive and negative pips at the leading and trailing edges, respectively, of the vertical sync pulses was described briefly. We will now consider in detail the action of the horizontal-differentiating circuit during the entire vertical pulse. Fig. 10-15A shows the pattern of the vertical signal following field No. 2.



(A) Vertical pulse group following field 2.



(B) Output of horizontal-differentiating circuit.

Fig. 10-15. Action of horizontal-differentiation circuits during vertical pulse period.

The horizontal pulse which starts retrace of the bottom line of the picture is shown at (1) in Fig. 10-15A. The positive output pip produced by its leading edge is shown at (a) in Fig. 10-15B. The pips produced by the trailing edge of this horizontal pulse and of all the other pulses of the period (labeled c) are rejected by the sync system, as previously explained.

Each equalizing pulse (2 and 3) before and after the vertical pulse also produces a pair of positive and negative pips. Only the pips marked (a) are used for oscillator control. Those labeled (b) are rejected, since they occur in the scanning cycle, while the horizontal oscillator is not sensitive to pulse control.

Each pulse of the vertical group (4 and 5) also produces a pair of positive and negative pips. However, only the positive pips (a) of Fig. 10-15B are used. The horizontal pulse shown at (6) is one of a group occurring during the blanking period. The pips produced by the pulse at 6 are the same as those produced by horizontal pulse (1).

It is evident that the vertical-pulse group, because of the individual pulses and their different lengths, can assure vertical retrace at the proper time and keep the horizontal oscillator in step with the scanning in the camera tube at the transmitter.

QUESTIONS

1. In what three places in the circuit can the synchronizing pulses be removed from the composite signal?
2. What is the function of a sync separator?
3. In a diode sync-separation circuit, what is the polarity of the output pulses when the input signal is applied to the plate of the diode?
4. What is the advantage of using a gated sync-separator circuit?
5. In a differentiator circuit, are the output pulses formed across a resistor or across a capacitor? Where are they formed in an integrator circuit?
6. What factor about the horizontal- and vertical-sync pulses makes it possible to distinguish between the two?
7. For a three-branch integrating circuit, is the total time constant smaller or larger than an individual time constant? How is the total time constant calculated?

EXERCISES

1. Draw a gated sync-separation circuit that uses a 6CS6 tube and explain the basic operation.
2. Show an integrator and a differentiator circuit, and draw the output waveform of each during the vertical sync pulse period.

Chapter 11

The Receiving Antenna

Television receiving antennas are more critical in performance and play a more important role in the production of satisfactory television reception than the antennas employed for AM or FM reception. AM or FM receivers are sensitive enough and the broadcast signal is strong enough that outside antennas are unnecessary in most instances. The antenna normally is inside the cabinet. A loop is employed for the broadcast band, and a simple half dipole or doublet is employed for the FM band. Reception is usually satisfactory except when the receiver is inside a steel building.

Many unique factors in television wave propagation are not encountered to as great a degree in other forms of broadcasting. A discussion of these factors will help the television service technician understand the major problems which will be experienced and arrive at the best compromise solution.

The elements of the antenna problem which must be considered are:

1. The nature of radiation and propagation of radio waves in the television bands.
2. Horizontal versus vertical polarization of television waves.
3. The wide-band nature of the television channel, and its relation to the susceptibility of the system to noise.
4. Ghost images due to multiple signal paths caused by the reflection of the signal from buildings, mountains, hills, or other obstructions.
5. Ghost images due to reflections in the transmission line (lead-in), produced by a mismatch between the impedance of the line and that of the antenna or receiver input circuit.
6. The necessity for special types of antennas to achieve a desired directional reception pattern.
7. Wide-band antennas to allow reception of stations widely separated in the television frequency spectrum.
8. Problems peculiar to fringe reception (reception of stations beyond the limits of the primary service area).

In our study of the television channel with its vestigial amplitude-modulated video and frequency-modulated audio components, we have seen that a channel width of 6 megacycles is required for high definition television. The modulation process requires that the carrier frequency be made at least ten times the highest modulating frequency. For this reason, the carrier frequencies of the television transmitters have been assigned in the portion of the radio spectrum above fifty megacycles. The standard classification of the part of the spectrum between 30 and 300 megacycles is VHF (very high frequencies). The propagation characteristics of VHF waves are considerably different from those of the lower frequencies employed for commercial radio broadcasting. A review of how waves travel from the transmitter to the receiver will be helpful to the reader in understanding some of the transmission phenomena which occur in the television bands.

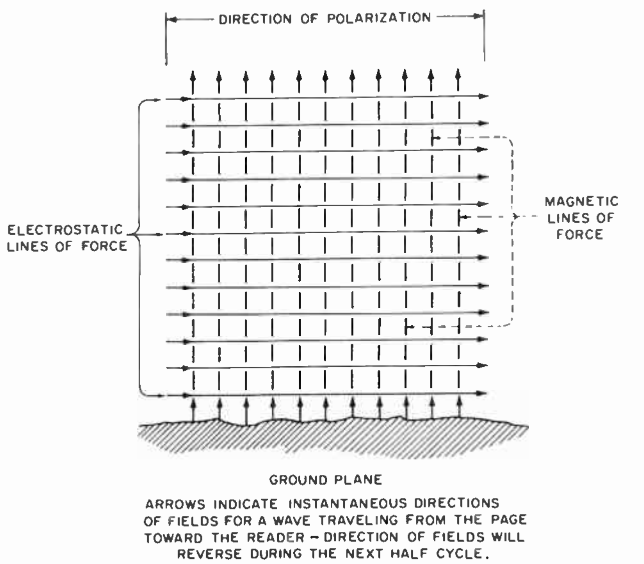


Fig. 11-1. Graphic representation of a horizontally-polarized electromagnetic wavefront.

After a wave has traveled several wavelengths from the transmitting antenna, it consists of two components, an electrostatic field and a magnetic field. These two fields are made up of lines of force at right angles to each other. The energy in the wave is divided equally between these two traveling alternating fields. Fig. 11-1 shows part of the wavefront of an electromagnetic wave in space. Since the wave is traveling in all directions from the transmitting antenna, the lines of Fig. 11-1 represent a small portion of a spherical surface.

POLARIZATION OF THE TRANSMITTED WAVE

The direction of the lines of force of the electrostatic component define the direction of polarization of the wave. In the graph of Fig. 11-1, the solid horizontal lines are the electrostatic lines of force. Such a wave is said to be horizontally polarized with respect to the earth's surface. If the electrostatic lines are perpendicular to the earth's surface, the wave is said to be vertically polarized.

In the VHF portion of the television spectrum, the plane of polarization of the transmitted wave is the same as the position of the transmitting antenna with respect to the earth's surface; that is, a vertical antenna produces vertically-polarized waves and a horizontal antenna produces horizontally-polarized waves. The United States has adopted horizontal polarization. The reasons advanced for choosing horizontal polarization are:

1. Many types of man-made interference, as well as interference from other radio communication transmitters, is vertically polarized; horizontal polarization helps reduce interference from these sources.
2. Horizontal-polarized waves suffer less loss when reflected from the earth or when passing through the atmosphere.

TYPES OF WAVE PATHS BETWEEN THE TRANSMITTER AND RECEIVER

Waves of different carrier frequencies follow different paths between the transmitter and the receiver. The waves are basically the same and would act alike in free space (beyond the earth's atmosphere). The earth's surface, the earth's atmosphere, and the presence of objects comparable in size to the length of the wave modify the transmission path as the frequency of the wave is changed.

Based on the paths they follow, waves can be classified into two main groups, ground waves and sky waves. The ground-wave group, which is of greater interest to us, can be further subdivided into a direct wave, a surface wave, and a space wave.

Sky Wave

The sky wave, which accounts for long-distance reception on the low frequencies such as the broadcast and short-wave bands, is a wave that is bent back to the earth by ionized layers (ionosphere) in the upper atmosphere. Fig. 11-2 shows the effect of the ionosphere upon transmitted waves. Some waves penetrate the ionized layer and others are refracted to the earth. Whether the waves are refracted or not dpends upon a number of principles—the frequency of the wave, the density of the ionized layer, and the angle at which the transmitted wave enters the ionosphere.

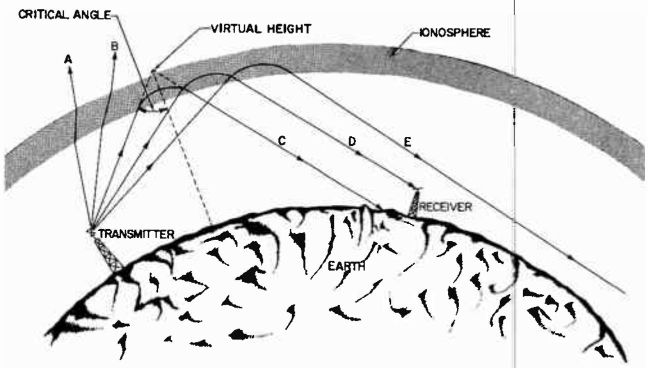


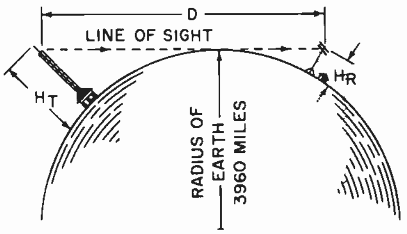
Fig. 11-2. The bending of radio waves back to the earth from the ionosphere. (Occurs at frequencies below 40 megacycles.)

The waves in Fig. 11-2 are all of the same frequency, and the ionosphere presents the same density for each wave. Because of their angle of entry, waves A and B in Fig. 11-2 can penetrate the ionosphere; therefore, they pass into space and are lost. The smaller the entry angle, the easier the wave is bent. A wave entering the layer at the angle of wave C cannot penetrate the ionosphere and is refracted to the earth. The wave is steadily bent as it passes into the ionized layer and emerges toward the earth as though it had been reflected. The amount of bending that a wave will encounter is a function of the frequency or wavelength. At each frequency there is a critical angle at which the required density for refraction of the wave is obtained (Fig. 11-2). All waves steeper than wave C will pass through the ionosphere. Those with more gradual slopes (to the right) will be reflected to the earth.

Frequencies of more than 30 to 40 megacycles are not returned by the ionosphere except under unusual sporadic conditions because the higher the frequency, the easier the wave will penetrate the ionized layer. Therefore, this type of wavepath cannot be depended upon for television transmission.

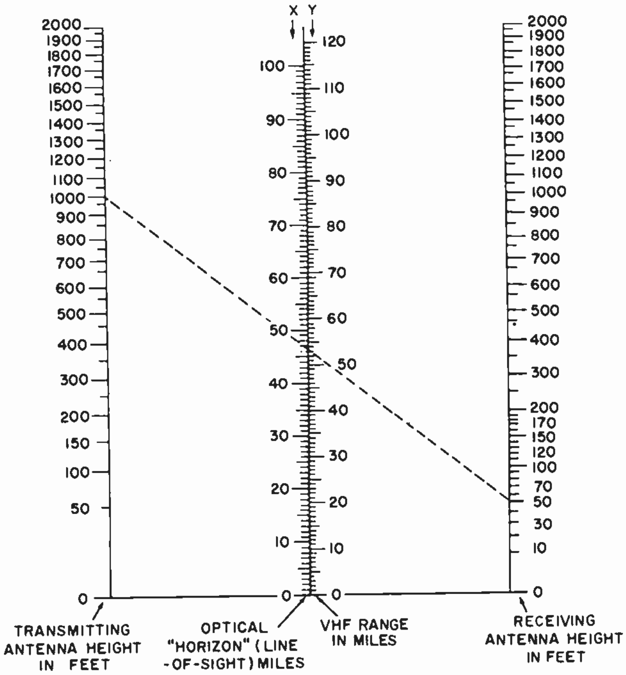
Ground Wave

The action of frequencies in the range above approximately 40 megacycles is often termed quasi-optical, since the waves act in a manner similar to light rays. (Quasi in Latin means "as if" and optical means "light.") Transmitters operating in this frequency range, such as television stations, employ as high an antenna as is practical and economical. The carrier waves leaving the antenna act like the rays of light which would be produced if a powerful electric light were mounted on top of the antenna mast. The earth's curvature would cut off these light waves at the horizon. The horizon distance can be extended by erecting a tower for the observer of the light.



$$D\,(\text{MILES}) = 1.23\left[\sqrt{H_T(\text{FT.})} + \sqrt{H_R(\text{FT.})}\right]$$

(A) Line of sight.



VHF RANGE = OPTICAL HORIZON DISTANCE X 1.156

(B) VHF range chart.

Fig. 11-3. Line-of-sight distance and VHF range as related to height of transmitting and receiving antennas.

Waves emitted by television transmitters exhibit this line-of-sight action, as shown in Fig. 11-3A; however, the cutoff at the horizon is not sharply defined. In traveling through the air, the wave is bent slightly toward the earth, and the VHF range is considered to extend some 15% beyond the line-of-sight or optical horizon. Beyond this range, the strength of the signal decreases very rapidly.

Fig. 11-3B shows, by means of a nomograph, the effect of the height of the transmitting and receiving antennas on the optical horizon (shown at X) and on the VHF range (shown at Y). This chart is calculated for a smooth, spherical earth. If the transmitting and receiving antennas are at different heights above sea level, this difference should be taken into account when the VHF range is calculated. Large intervening objects such as buildings or hills can reduce seriously the signal level and thus the range of satisfactory reception. Fig. 11-3B indicates the advisability of locating the receiving antenna as high as possible.

WAVELENGTHS OF THE TELEVISION CHANNELS

In the ranges of frequencies assigned to television transmission (54 to 88 megacycles, 174 to 216 megacycles, and 470 to 890 megacycles), the length of the electromagnetic wave is short when compared with the height of obstructions, such as buildings. This accounts for some of the peculiar transmission phenomena encountered. These will be described when we examine the effect of reflections. On the credit side of the ledger is the fact that highly efficient antennas, which are resonant to the wavelength and possess directional characteristics, can readily be constructed.

For these reasons, a review is needed of some of the relationships between the frequency of an electromagnetic wave and its wavelength. The television service technician should become familiar with the actual wavelengths (in feet) of the television carrier frequencies in his locality, since the lengths of antenna elements are directly related to the wavelength.

All electromagnetic waves (cosmic waves, X rays, ultraviolet waves, light, infrared, and radio waves) travel through space at the same speed—approximately 300,000,000 meters, or 186,000 miles, per second.

The distance an electromagnetic wave travels in one cycle (360° of its sine-wave oscillation) is called one wavelength. Fig. 11-4 shows the relationship of wavelength, velocity, and frequency. When the velocity is in feet per second and the carrier frequency is in cycles, both the numerator and the denominator of the fraction in Fig. 11-4 can be divided by one million. The simple expression, 984 divided by the frequency in megacycles, is the result. The wavelengths encountered in the VHF television bands are between four and twenty feet.

A handy rule-of-thumb from Fig. 11-4 is that the television wave travels approximately 1,000 feet in one microsecond (actually 984 feet). This information is valuable for determining the effect of wave reflections from objects.

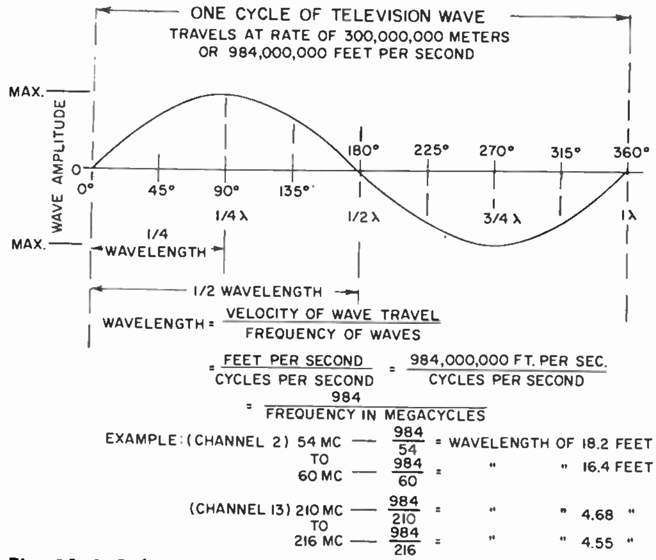


Fig. 11-4. Relation of carrier frequency to wavelength of television signal.

THE NOISE PROBLEM

Television requires an extremely wide band to accommodate both the video modulation and the sound. One of the axioms or principles of electrical communication states that when amplitude modulation is employed, the noise susceptibility of a system is proportional to its bandwidth. For example, automobile ignition systems produce interference. Because of the length of the spark-plug wiring, this interference is quite often tuned and radiated within the wide television bands. Fortunately, the strength of a radiated interfering field drops rapidly as the distance from the source of interference increases. If the TV antenna system is relatively high, is not sensitive to vertically polarized waves, and has a balanced lead-in system, the effects of this and other types of man-made interference can be reduced. These are added reasons for the installation of an efficient antenna.

GHOSTS DUE TO MULTIPLE-PATH TRANSMISSION

Before the television antenna is considered in detail, let us examine some of the common picture distortions caused by indirect transmission paths. How these defects are overcome or alleviated will then be discussed.

Fig. 11-5A shows a condition which might occur in an urban residential section. The direct signal arrives at the receiver over path A (assumed to be 10,000 feet). When the signal is 12,800 feet from the transmitter, the office building reflects the signal to the receiver over path C (9,000 feet). The combined path B plus C is 21,800 feet. This reflected wave path is 11,800 feet longer than the direct path. We find from the wavelength-frequency relationship in Fig. 11-4 that the wave travels 984 feet per microsecond. In the foregoing instance, the reflected wave will arrive 11,800/984, or 12 microseconds later than the direct wave.

Fig. 11-5B shows the effect of the delayed, or ghost, signal on the picture. The horizontal-scanning time relationships discussed in Chapter 7 will explain the image displacement. Approximately 54 microseconds elapse while one line is scanned. On a twenty-one inch tube, the picture is about ninteen inches wide. The signal arriving over path B plus C requires twelve microseconds more than the direct path; during this time, scanning will have advanced four and one-quarter inches. Thus, a second image due to the reflected wave will appear displaced from the main image as shown. This ghost image will generally suffer some loss of signal strength due to imperfect reflection and to energy absorption and will therefore be less intense than the image caused by the direct signal. Fig. 11-5C shows the appearance of a ghost of the type just discussed, photographed from a television receiver with the test pattern as the subject.

In the example just discussed, the ghost was caused by a single reflection. Often, a series of overlapping ghost images due to multiple reflections from a number of buildings or from other obstructions will appear. When such multiple ghost images arrive over paths only slightly longer than the direct signal path, vertical lines on the picture are widened and the fine detail is obscured. Since the picture appears smeared, this phenomenon is known in television slang as a "smear ghost."

Ghost images are eliminated or reduced until they are no longer objectionable by using the directional properties of certain antennas. The theory and application of such antennas will be covered later in this section.

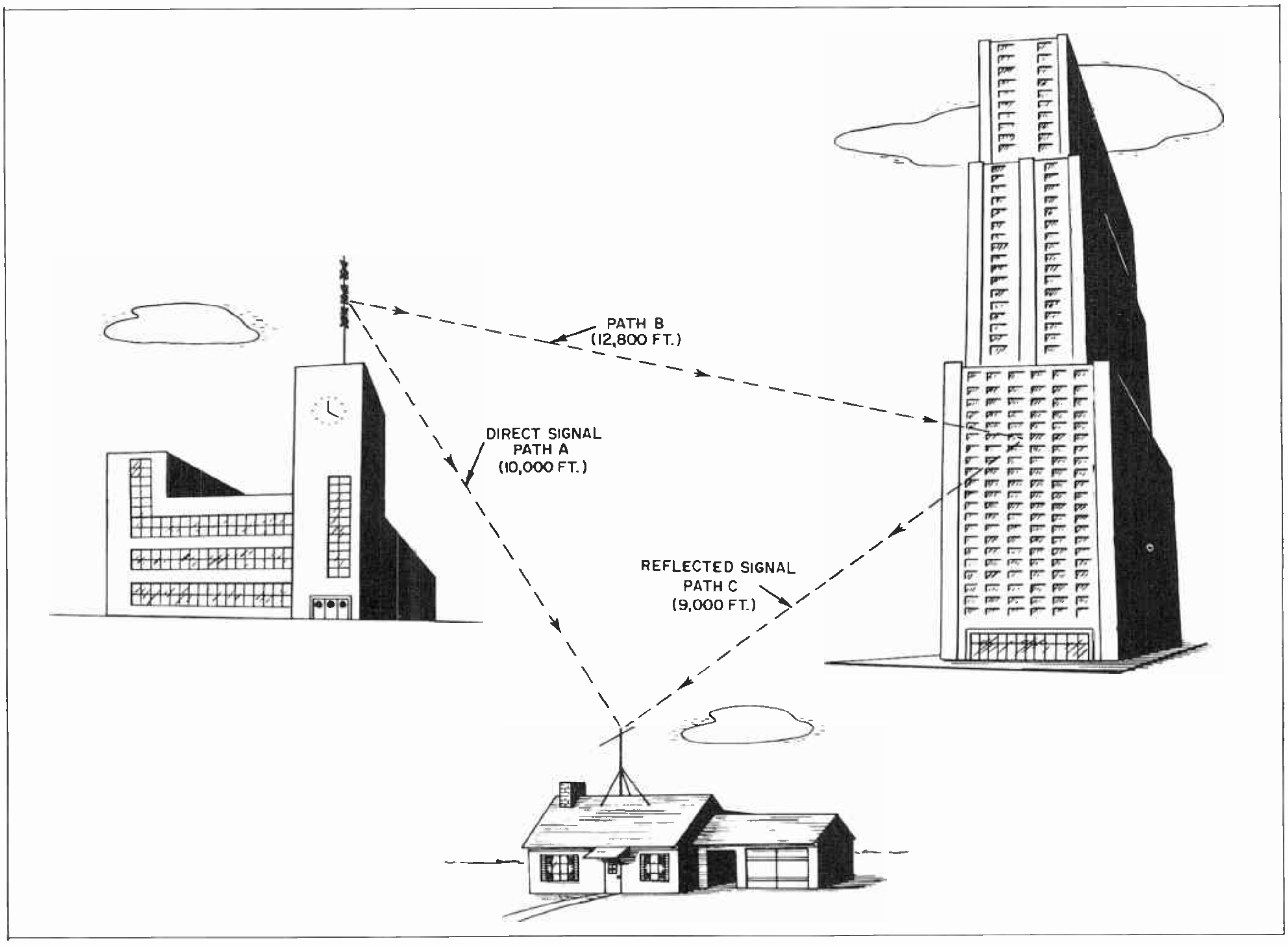
In certain urban locations, the receiving antenna is shielded by buildings from a line-of-sight signal from the transmitter. One of the reflection paths will often produce a stronger signal than the direct path. The antenna is beamed to accept the stronger reflected signal. The dominant reflection is then considered a substitute primary signal, and the direct path produces a ghost which must be suppressed.

GHOSTS DUE TO REFLECTIONS IN THE LEAD-IN

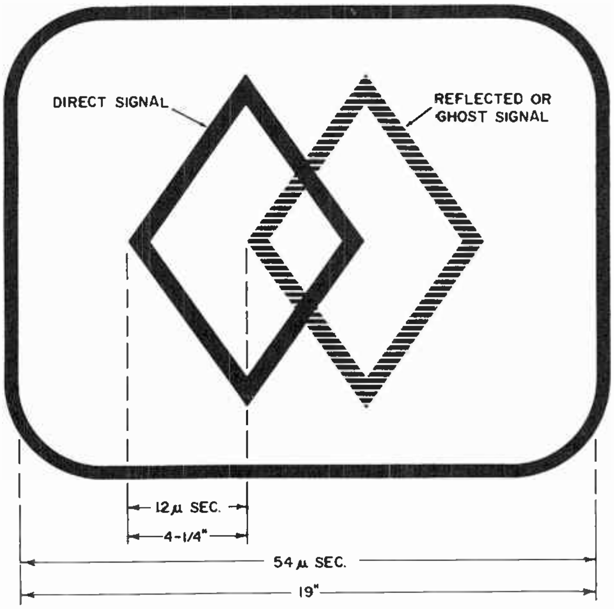
The resonant antennas employed for television require a special lead-in. This lead-in (or transmission line) consists of either a parallel-wire line, a twisted-pair line, a shielded twisted-pair line, or a coaxial (concentric) cable.

The reasons for these special types of lead-ins are: (1) to transfer the maximum amount of energy from the low-impedance antenna to the first tube input in the receiver, and (2) to restrict signal pickup to the antenna only.

Maximum energy transfer of the intercepted signal from the antenna to the receiver input amplifier



(A) Illustration of multiple-path transmission.



(B) Ghost caused by reflection over path B plus C.

(C) Ghost due to single-reflection path.

Fig. 11-5. Production of a ghost image by multiple-path transmission.

requires that the characteristic impedance of the line be matched to the impedances of the antenna and the receiver input circuit. The various types of lead-ins and their application will be covered in greater detail later. At present we will consider the effect of line mismatch in the production of ghost images.

When the impedance of the transmission line differs considerably from that of the antenna or the receiver input, energy will be reflected from the receiver to the antenna and back to the receiver again. A ghost image is produced when the reflection reaches the receiver. Unless the mismatch is considerable, only one secondary or ghost image will be produced.

When ghost images are caused by line mismatch, only a small displacement of the image occurs because of the relatively short transmission line. If the lead-in is less than 70 feet long, the ghost will be so close to the main image that the eye cannot distinguish it as a separate image. The horizontal resolution will be impaired, and the image will appear improperly focused.

It is possible to distinguish between a ghost image due to line reflection and one due to reflection by some object. If the antenna is rotated while the service technician watches the image, a ghost due to line mismatch will not change. On the other hand, if the ghost is due to a reflected path, its intensity will change with respect to the main image due to the directional characteristic of the antenna.

THE HALF-WAVE DIPOLE

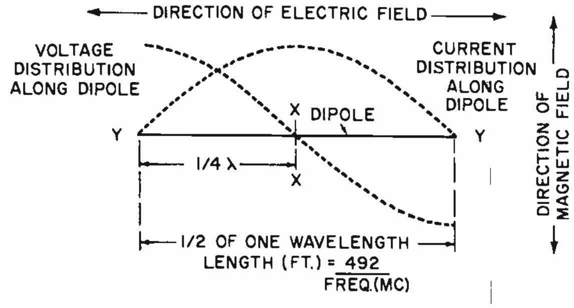
Having reviewed the nature of wave propagation and some of the basic phenomena of the transmission of television signals, we are now prepared to examine the types of antennas which have been developed to cope with the problems encountered.

Since the advent of commercial television in the United States, there has been much activity in the search for types of antennas to meet requirements. Among the desirable characteristics are more uniform response or higher pickup and sharper directional pattern.

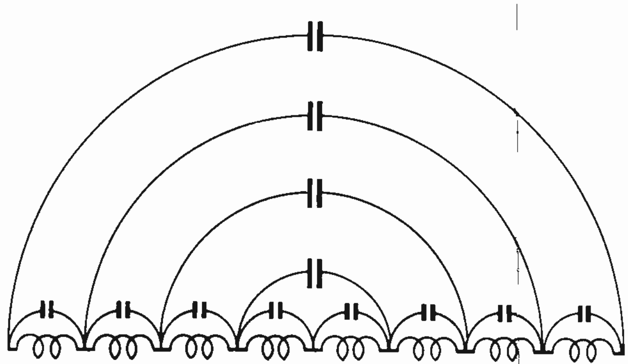
Practically all television antennas are variations of the basic half-wave dipole or Hertzian doublet. An understanding of this resonant type will help in the study of the more complicated arrays. The half-wave doublet is the industry standard against which the efficiency of other types is compared.

Fig. 11-6A shows the concept of a half-wave dipole in space. If such an antenna were of very small wire and could be removed from the earth's influence, its resonant electrical length would equal its physical length. This length, in turn, would be one-half the wavelength of the electromagnetic wave in space.

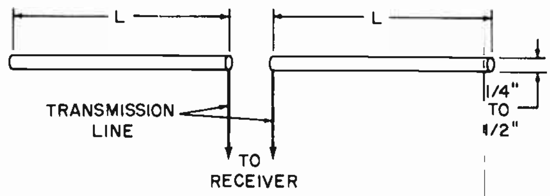
Although Fig. 11-1 shows the electric and magnetic fields as fixed uniform lines, both of these fields actually are varying at a sine-wave rate. Thus, a varying magnetic field will induce an alternating flow of electrons in the dipole wire, and a wave of current will pass down the dipole to its end. At the end, reflection will occur, and a standing wave having the voltage and current distributions shown in Fig. 11-6A will build up along the wire. Since the length of the theoretical dipole in Fig. 11-6A is equal to one-half wave in space, the voltage along the wire



(A) A theoretical dipole (a very thin wire isolated in space). Length equals half of carrier wavelength.



(B) Electrical circuit equivalent of the dipole.



2L = 1/2 WAVELENGTH X .95

$L\,(FT.) = \frac{234}{FREQ.(MC)}$

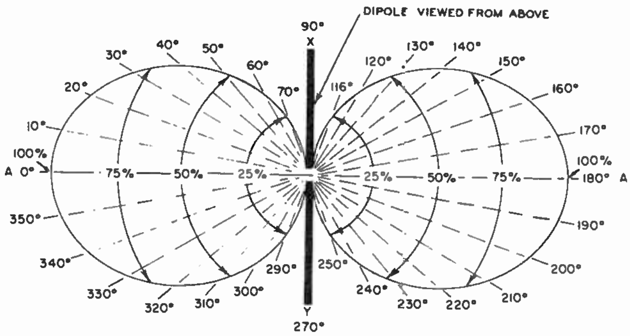
(C) A practical receiving dipole.

Fig. 11-6. The half-wave dipole antenna.

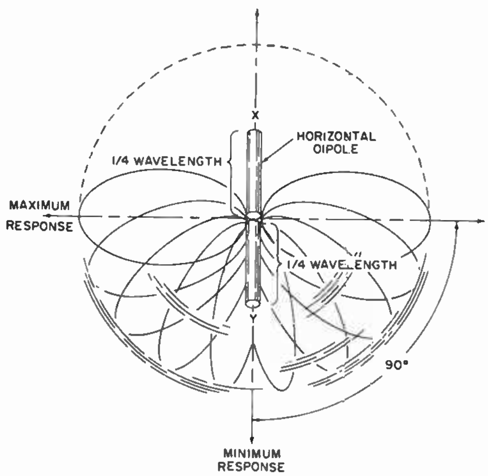
and the current through it will reach a resonant condition similar to that found in a parallel-tuned circuit. Actually, the dipole may be considered such a tuned circuit, but here the inductance is distributed along the length of the wire; and the capacitance consists of many elements across the individual small inductances. This tuned circuit equivalent is shown in Fig. 11-6B.

The foregoing discussion has been concerned with hypothetical dipoles. They could never exist because the dipole wire must be infinitely small in diameter compared with its length and the dipole must be so far from surrounding objects, including the earth itself, that there is no influence from such objects. In practice, the dipole must be erected within a few wavelengths of ground and must be large enough in diameter to be self-supporting. Both requirements cause the velocity of wave travel in the practical dipole to be less than the speed of electromagnetic waves in free space.

Although the length of the dipole for resonance will vary slight, depending upon the proximity of other objects, an average figure for the reduced length due to the proximity effect is five per cent.



(A) Reception pattern of a half-wave dipole, showing relative response versus direction in a horizontal plane.



(B) Solid pattern in space of half-wave dipole.

Fig. 11-7. The reception pattern in space of a half-wave dipole.

In other words, the half-wavelength value obtained from the formulas in Fig. 11-4 should be multiplied by .95 to find the length of a dipole for any given frequency. A practical dipole formula is found in Fig. 11-6C.

The dipole produces the greatest current when parallel to the electric field of the wave front. In other words, a line drawn from the receiving antenna to the transmitter should be at right angles to the length of the dipole. As the dipole is rotated, its response varies as shown in Fig. 11-7A. A curve of this type is known as a polar response curve. The response at different angles away from the normal or maximum response position is determined by taking the ratio of the length of the radius at any point to the maximum radius, or line A-A.

The response curve of Fig. 11-7A represents a cross section taken in the plane of the antenna. The response to waves arriving from the sky or reflected from the ground is determined by the solid figure obtained by rotating the curve in Fig. 11-7A around axis X-Y as shown in Fig. 11-7B; this solid figure is doughnut-shaped.

From the reception pattern in Fig. 11-7A, we see that the simple dipole is bidirectional and receives equally well from front or rear, but shows practically no response in line with the length of the dipole. This directional characteristic is valuable for discriminating against spurious reflections, which would produce ghosts in the picture. The simple dipole is satisfactory when the signal strength is high, when only one station (or stations whose frequencies are close to each other) is to be received, and when the ghost problem is not too severe.

If greater directional selectivity or more signal pickup is required, the pattern and response of the simple dipole can be improved by additional elements known as reflectors and directors. The use of these elements in more complicated antenna structures or arrays will be covered in detail later.

The impedance of the simple half-wave dipole at its center is approximately 72 ohms. For maximum energy transfer, the lead-in also should have a characteristic impedance of 72 ohms. This impedance value is rather low and is best suited to the coaxial-type lead-in. The mismatch of the antenna to the lead-in can vary as much as two to one without serious signal loss and without ghosts, if the line is appropriately matched to the receiver input. Therefore, lead-ins having impedances between 36 ohms and 144 ohms can be used with the half-wave dipole.

The simple half-wave dipole is most efficient when its length is correct for a particular carrier frequency. If several stations in adjacent television channels are to be received with a simple antenna, its length should be made correct for the geometric means of the lowest and highest channel frequencies required. For example, if a station on Channel 3 (video carrier frequency, 61.25 mc) and a station on Channel 6 (video carrier frequency, 83.25 mc) are to be received with approximately equal antenna response, the dipole should be made the correct length for the geometric mean of these frequencies. Therefore, if the proper length of a dipole to receive Channel 3 is 7 feet, 7¾ inches and the proper length for Channel 6 is 5 feet, 7½ inches, then the length required to give an approximately even response over both channels with a single dipole would be the average length or the sum of both divided by two. For example.

$$7'\ 7\tfrac{3}{4}'' + 5'\ 7\tfrac{1}{2}'' = 13'\ 3\tfrac{1}{4}'' = 159\tfrac{1}{4}''$$

Then,

$$159\tfrac{1}{4}'' \div 2 = 79\tfrac{5}{8}''.$$

This would be the over-all length of the dipole for a fairly equal response between the two channels. The length of each dipole section would thus be

$$79\tfrac{5}{8}'' \div 2 = 39\tfrac{13}{16}''.$$

THE FOLDED DIPOLE

If two half-wave dipoles are placed parallel to each other with their ends connected, the received currents will be in phase. The reaction of one dipole upon the other will increase the impedance at the center of either dipole from 72 ohms to approximately 300 ohms. Fig. 11-8 shows such a folded dipole and the formula for calculating the required dimensions. The folded dipole possesses several advantages over the single half-wave dipole:

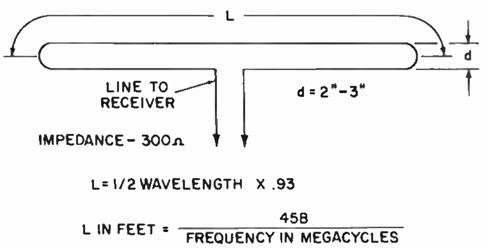
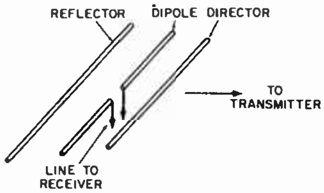


Fig. 11-8. Dimensions of a folded dipole.

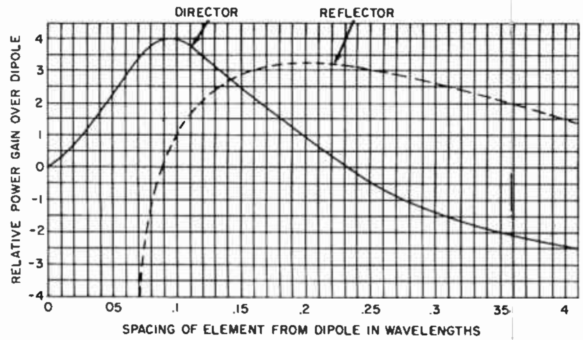
1. Its higher impedance provides an ideal match for the often-used polyethylene-insulated parallel, or twin-lead, transmission line. The most popular type of this lead-in has a characteristic impedance of 300 ohms, and most television receivers can match a 300-ohm lead-in.
2. It is receptive to a wider frequency band than the simple half-wave dipole, but has the same directional pattern.
3. Because of its design, it has a more rigid structure and will withstand greater wind pressure.

ANTENNA STRUCTURES EMPLOYING THE DIPOLE WITH REFLECTORS AND/OR DIRECTORS—YAGI ARRAYS

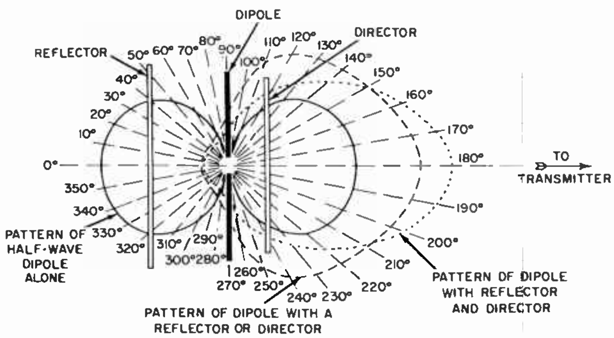
If a second half-wave dipole is placed parallel to and closer than a half-wave length from the receiving dipole, it is called a "parasitic" element and its magnetic and electrostatic fields will modify the directional pattern and increase the gain of the receiving dipole. When a parasitic element is on the side away from the transmitter, the element is known as a reflector. An element on the side toward the transmitter is known as a director. A reflector is approximately 5% longer than the receiving antenna, and a director is about 4% shorter. This is equivalent to saying that the reflector is tuned to a somewhat lower frequency than the operating frequency and that the director is tuned to a slightly higher frequency. Fig. 11-9A shows the arrangement of the dipole with a director and a reflector. The effect of the spacing of these elements on the power gain of the dipole antenna is shown in Fig. 11-9B.



(A) Dipole with director and reflector.



(B) Effect of director and reflector spacing on gain of parasitic array.



(C) Typical reception patterns of a dipole with director and reflector elements singly or in combination.

Fig. 11-9. The dipole antenna with parasitic elements (directors and reflectors).

Fig. 11-9C compares the directional pattern and gain of the dipole alone with a dipole using either a reflector or a director, or both. The gain is increased by these additional elements and the directional pattern is made sharper, so that stations at other points of the compass are not received with the same strength. For this reason, arrays of this type will increase the pickup from the desired station and, at the same time, suppress reflection paths which would produce ghosts. These multiple-element arrays are sometimes called Yagis, since a Japanese by that name first proposed using them as directional antennas.

Arrays with directors and reflectors, although an improvement over the dipole from the standpoint of pattern and gain, will not accept as wide a frequency band as the simple dipole or folded dipole. It is often necessary to erect a number of these antennas tuned to various stations in the band. Fig. 11-10 shows two sets of arrays on a single pole. The lower array is for the low band of television frequencies, i.e., Channels 2 through 6; and the upper array of smaller size is for Channels 7 through 13.

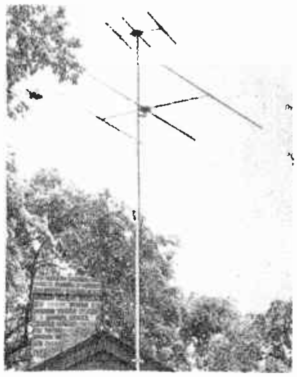


Fig. 11-10. Array with separate dipoles for high and low VHF bands.

If increased gain and directivity are required, additional director elements can be added. Arrays with as many as ten elements are often employed to receive a single station when the receiving location is in a fringe area.

THE BROADBAND PROBLEM

The half-wave dipole and the arrays employing half-wave elements are most efficient at the frequency for which they have been cut. Satisfactory performance is obtained when only one station serves the area or when relatively high signal strength exists for several stations in the same television band. When stations in both the low band (54 mc to 88 mc) and the high band (174 mc to 216 mc) must be received, antennas with wider frequency responses are necessary.

The folded dipole exhibits the broadest frequency response of any of the television antennas discussed to this point. Fig. 11-11 shows a combination of folded dipoles which provide directional discrimination and reception in both bands. This arrangement is called an "in-line" antenna. In the low-frequency band, the larger folded dipole backed by a reflector is the receiving antenna, and the smaller folded dipole acts as a director. In the high-frequency band, the small folded dipole functions as the antenna, and the large folded dipole behind it acts as a reflector. The two dipoles are connected to each other at their center (high-current) points by the proper length of twin-lead transmission line, so that a transition occurs between the two bands. An additional feature of this array, determined by the size of the elements and by their spacing, is a low efficiency in the gap between the two television bands. This gap contains the FM broadcast stations (88 mc to 108 mc), which may be a source of television interference. In the following text, means for increasing the bandwidth, other than by use of the folded dipole, are described and illustrated.

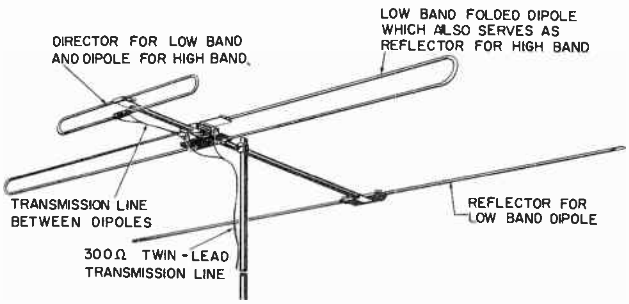


Fig. 11-11. "In-line" antenna array of folded dipoles for reception on both bands.

A Large-Diameter Dipole

A larger-diameter dipole will have a wider frequency response than a smaller-diameter dipole of the same length. If the dipole is a large cylinder, i.e., three to six inches in diameter, its Q will be decreased and its response will be broadened. Such a cylinder would be awkward to install and could be easily damaged by wind. An equivalent of the cylinder can be obtained by constructing a cage of wires with their ends connected to rings.

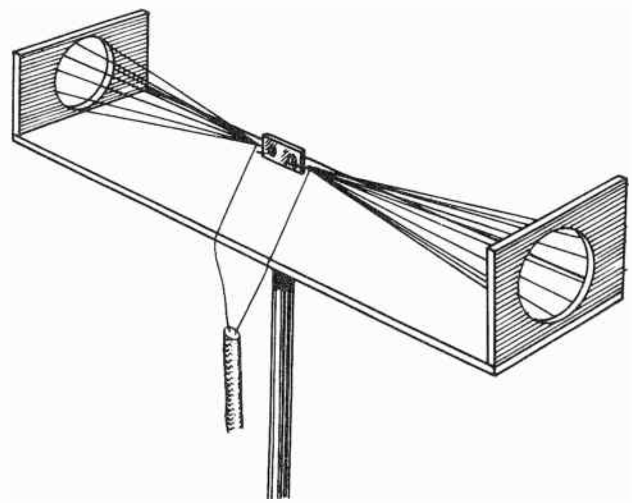


Fig. 11-12. The cage antenna—a variation of the dipole for wide-band reception.

A Cone-Shaped Dipole

The cage construction can be further modified to the form of two cones whose apexes meet at the lead-in or feed point. Fig. 11-12 shows such an antenna, although this is still not a practical design because of its awkward construction.

A Modified Cone

By taking a cross-section of the antenna of Fig. 11-12, antenna designers arrived at the fan-type antenna of Fig. 11-13. This antenna belongs in the "conical" family of antennas and has a broad frequency response.

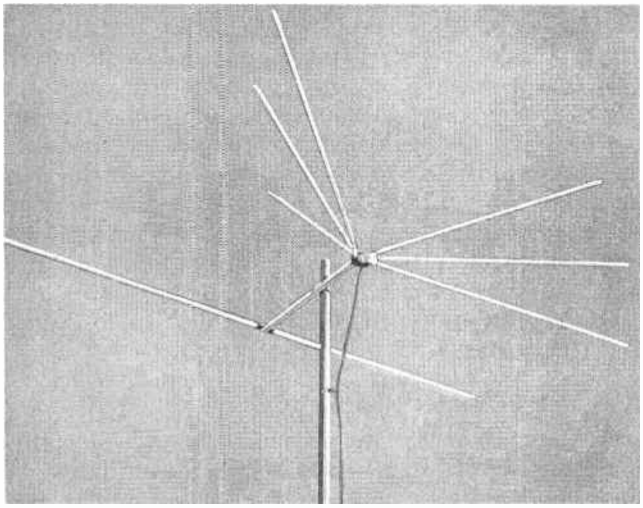


Fig. 11-13. A fan, or "conical," antenna having a broadband response.

A Nonparallel Dipole

The V-type antenna consists of a dipole with V-shaped elements. Moving the dipole elements from a straight line broadens the band, yet still retains the directional pattern. Fig. 11-14 shows a simple V antenna. Note the similarity to an indoor "rabbit-ear" antenna.

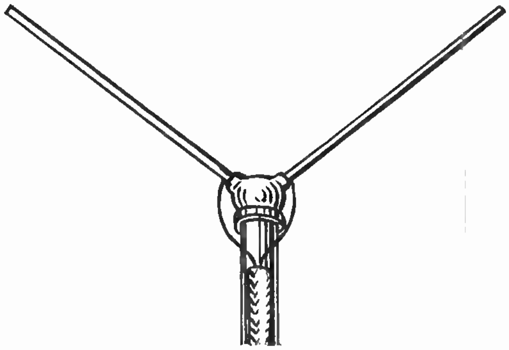


Fig. 11-14. The V antenna.

STACKED ARRAYS

Any type of antenna can be stacked by erecting another identical antenna above the first and in the

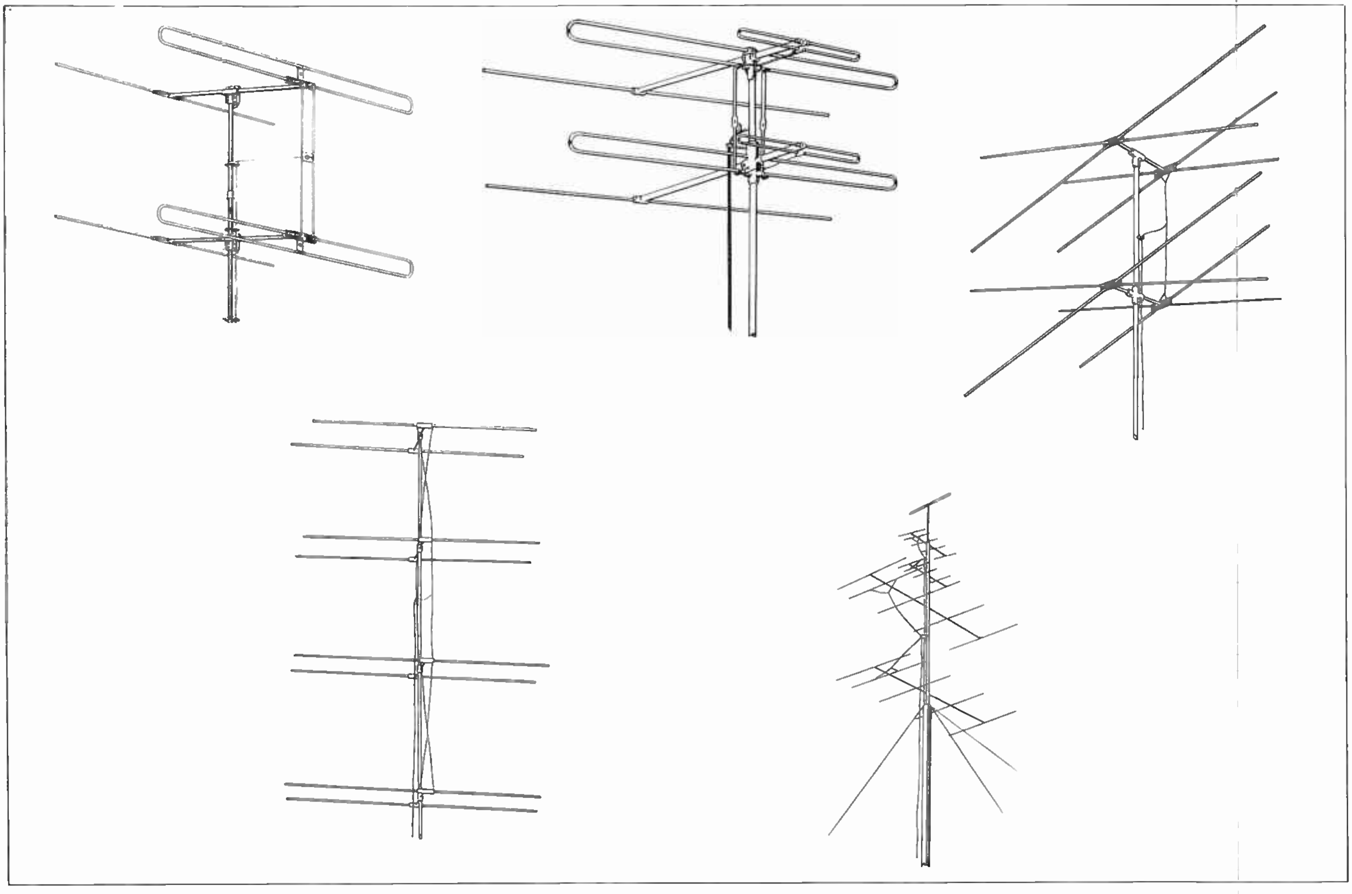


Fig. 11-15. Stacked arrays for higher gain and improved directional pattern.

same vertical plane. The antennas are critically spaced (usually one-half wavelength) to provide in-phase operation to a common lead-in. The advantages of vertical stacking are twofold:

1. Additional gain is obtained because of the added antenna.

2. Some vertical directivity is contributed by the mutual interaction of the antennas. This interaction discriminates against a reflected wave from the ground and confines the reception to the direct or sky wave. Fig. 11-15 illustrates several versions of vertically-stacked arrays.

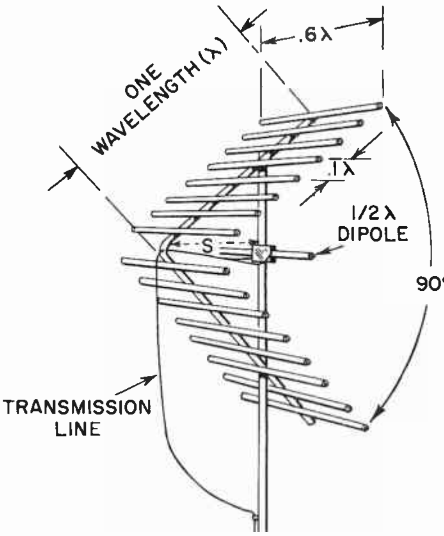
THE CORNER-REFLECTOR ANTENNA

Fig. 11-16 shows an antenna structure which, because of its very high front-to-back ratio, greatly increases the pickup of the dipole. This design is used for receiving UHF television channels; the reflector becomes too large in the VHF range.

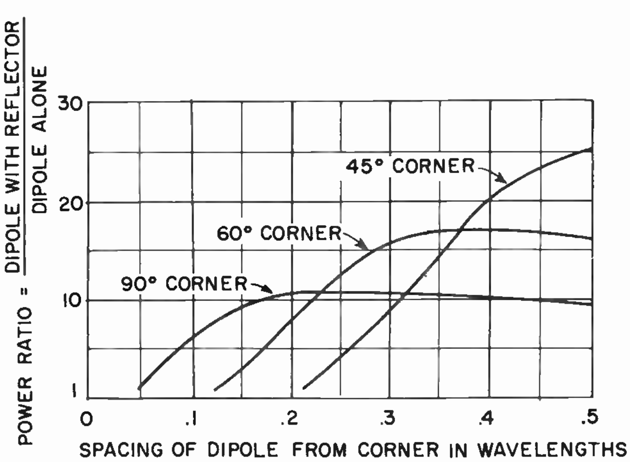
Fig. 11-16B shows the ratio of power received with the reflector to power received with the dipole alone for various spacings of the dipole from the corner and for corner angles of 90 degrees, 60 degrees, and 45 degrees. Fig. 11-16C illustrates the influence of the corner reflector on the antenna impedance. This has been expressed in percentage change of dipole impedance, since the corner reflector can be used with any type of dipole.

ROTATABLE ANTENNAS

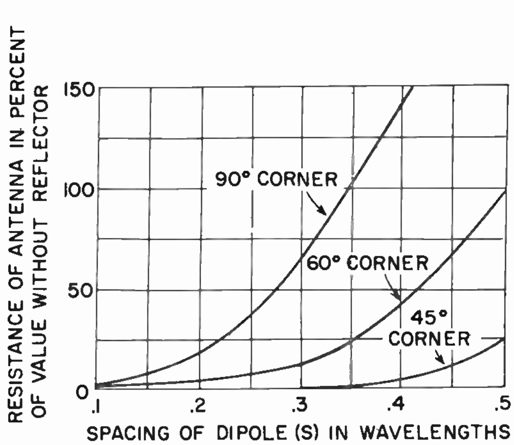
Most television antennas are directional; for best reception, the receiving antenna should be pointed toward the transmitting antenna. The foregoing creates a problem in some areas where there are more than one television signal and the transmitting



(A) 90° corner-reflector antenna.



(B) Effect of corner angle and dipole position on gain of corner reflector.



(C) Effect of corner angle and spacing on dipole resistance.

Fig. 11-16. The corner-reflector antenna.

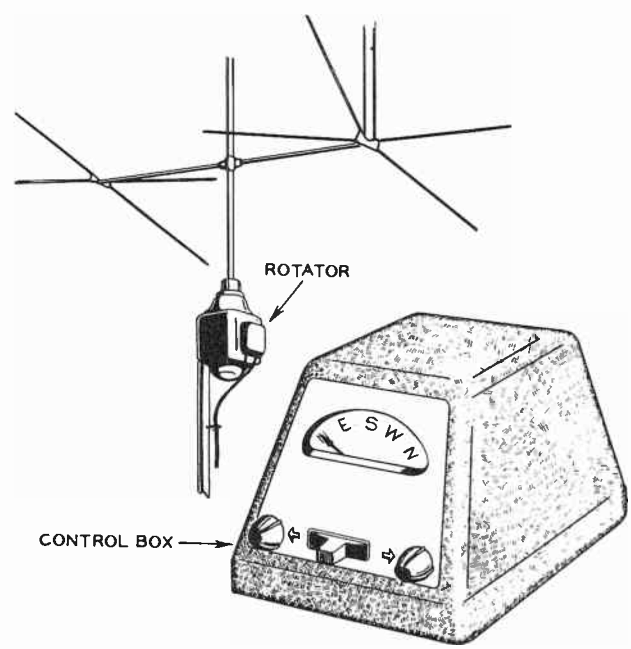


Fig. 11-17. Typical antenna rotator installation.

towers are located in different directions. Also, in areas where there are multiple-path reception difficulties, the antenna must be rotated in order that the best signal can be located. For best reception on each channel, antenna rotators are employed to rotate the antenna.

Fig. 11-17 shows a remote-controlled, motor-driven antenna. The motor which rotates the antenna is mounted on the antenna mast and the antenna is then mounted to the motor. The motor is remote-controlled by a unit at the receiver. Power to the motor is obtained by a cable connected from the motor to the control unit. Some control units indicate the direction in which the antenna is pointing. With these devices, the user can record the best position of the antenna for any given station and can subsequently return the antenna to this position. The motor is geared for slow rotation of the antenna, so that the point of best reception is not overlooked. The rotator turns the antenna through a complete revolution of 360°.

TYPES OF LEAD-INS

In our discussion of ghosts produced by reflections in lead-ins (transmission lines), we indicated that maximum power transfer and freedom from reflections occur when the characteristic impedance of the lead-in matches the antenna resistance and the receiver input resistance. The power developed in the antenna must be transferred to the grid of the first RF amplifier tube with as little loss as possible in order to override noise and produce steady, high-contrast pictures. Three types of lead-in lines have been used extensively. They are the two-wire parallel line, the coaxial or concentric cable, and the shielded two-wire parallel line. Each of these types exhibits its own particular characteristics of impedance and of loss, or attenuation, per foot.

The Two-Wire Parallel Line

The most popular lead-in consists of two uniformly spaced stranded wires molded parallel to each other in a flat strip of polyethylene insulating material. This low-loss insulating material has excellent VHF properties, including high resistivity, low water absorption, and a low dielectric constant. The lead-in is available in impedances of 75, 150, and 300 ohms and provides a match for any of the antennas discussed. Fig. 11-18A shows this type of parallel lead-in cable. The only disadvantage of parallel-lead cable is that it is not shielded and, therefore, will pick-up noise from electrical equipment and automobile ignition.

A variation of this type of parallel line consists of two wires molded opposite one another in the side walls of a plastic tube. The advantages of this construction are: (1) the leakage path has been increased, (2) the dielectric between the wires is mostly air, and (3) the spacing between the wires

(A) Flat twin-lead.

(B) Tubular twin-lead.

(C) Coaxial cable.

(D) Universal standoff insulator.

Fig. 11-18. Types of lead-ins and a standoff.

can be smaller for a given characteristic impedance than the width of comparable ribbon type. Thus, the loss per foot is smaller; and the performance, especially in wet weather, is improved. Fig. 11-18B shows such a tubular two-wire line.

Coaxial or Concentric Cable

Coaxial cable for television lead-ins consists of a flexible conductor molded in the center of a solid polyethlene cylinder. This cylinder is surrounded by a braided-copper outer conductor, and the entire cable is covered by a weatherproof vinyl sheath. The outer conductor is grounded at the receiver and acts as a shield for the inner conductor. Coaxial cable has a very low loss per foot, is free from noise pickup, and is available with characteristic impedances from 50 to 150 ohms. The impedance is determined by the ratio of the diameter of the outer conductor to the diameter of the inner conductor. For this reason, impedances higher than 150 ohms would require either a large outer diameter or an extremely small and weak inner conductor. Fig. 11-18C shows a sample of coaxial cable.

Shielded Two-Wire Parallel Line

This lead-in is similar to the coaxial cable except that two conductors, equally spaced from the center, are molded in the solid dielectric. The line is balanced because the outer conductor is merely a shield and is connected to ground; the two inner conductors are connected to the balanced input of the receiver and to the antenna. The losses per foot are higher than losses in coaxial cable of the same impedance. Impedances from 40 to 100 ohms are available. This lead-in is especially valuable for short transmission lines in unusually noisy locations.

Fig. 11-18D shows a universal standoff insulator that can be used with each of the lead-ins shown. The purpose of the standoff is to secure the cable and to hold it away from the building.

TELEVISION RECEPTION IN FRINGE AREAS

The service area of a television transmitter for rural and residential areas is normally defined by a contour line beyond which the signal strength falls below about 500 microvolts per meter. This contour depends upon the power of the transmitter, the height of the transmitting antenna, the topography of the land, and the effect of shielding by buildings and other structures.

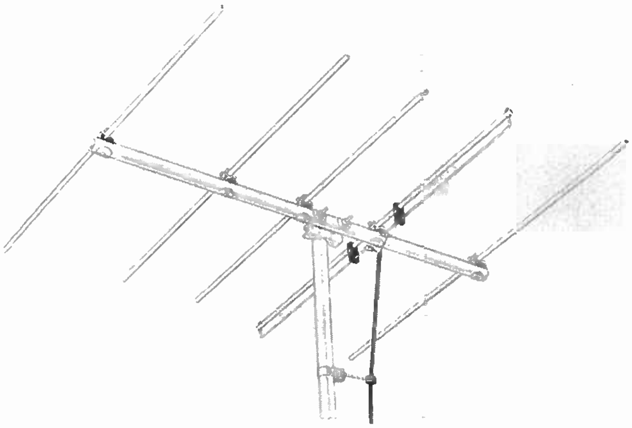


Fig. 11-19. Multielement Yagi antenna.

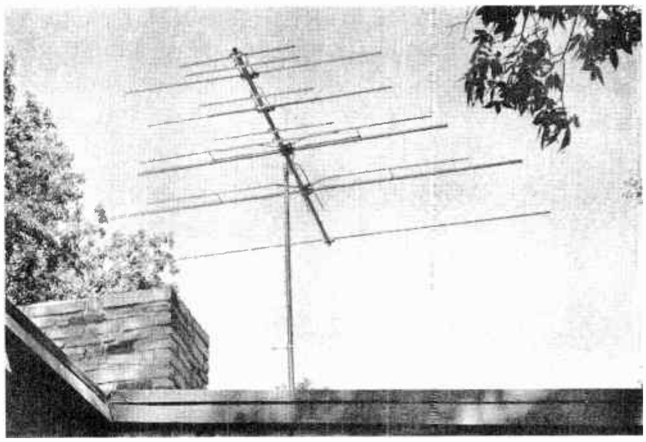


Fig. 11-20. A popular wide-band antenna now being used.

Fringe reception at points beyond the service area requires highly efficient antennas. Yagi antennas like the one in Fig. 11-19 are suitable for the reception of individual channels, and modified Yagis can be used for multichannel reception. Fig. 11-20 shows a popular wide-band antenna for all-channel reception. As shown in Fig. 11-3 the VHF range is increased by additional height at the receiving location. Towers for this purpose are commercially available. Television boosters can provide additional RF gain; and many types of these are also used in weak signal areas.

QUESTIONS

1. When the electrostatic lines of force of a television signal are horizontal to the earth's surface, is the signal horizontally or vertically polarized? What type of polarization is standard in the United States?
2. What determines the transfer of maximum energy of the signal from the antenna to the receiver input?
3. How are ghost images due to line reflections differentiated from those due to signal reflections?
4. What are three desirable characteristics of television receiving antennas?
5. At what position with respect to the transmitted signal will a receiving dipole produce the greatest current?
6. What are the director and reflector elements of an antenna assembly? How are they identified?
7. What is meant by stacked arrays? What are their advantages?

EXERCISES

1. Sketch the following types of antennas:
   (a) A single dipole.
   (b) A folded dipole.
   (c) A dipole with a director and a reflector.
2. Calculate the length of a dipole element to be used for Channel 4. Also, one to be used for Channel 8.

Chapter 12

RF Tuners

The RF section (or tuner) consists of an RF amplifier stage, a converter or first-detector stage, and a local oscillator. This combination of circuit elements performs the same function as its counterpart in a conventional broadcast or short-wave superheterodyne receiver.

A number of complications not encountered in the reception of ordinary broadcast signals require a more complicated design than is found in standard broadcast sets because:

1. The broadband nature of the television channel requires the acceptance and amplification of a band of frequencies six megacycles wide.
2. The frequency allocation of television channels (54 mc to 88 mc and 174 mc to 216 mc for the VHF band; and 470 to 890 mc for the UHF band) necessitates special types of coupling circuits to maintain uniform gain at the extremes of each band.
3. Balanced types of input circuits must be matched in impedance to the characteristic of available parallel-lead or coaxial transmission lines.
4. Undesired or spurious responses to signals outside the desired television band must be avoided. These responses may be caused by:

   a. The adjacent-channel sound carrier.
   b. Cross modulation due to other television channels.
   c. Direct transmission, through the RF system, of signals at the intermediate frequency.
   d. Interferences due to other television channels and to FM stations.
   e. Overloading of the input tube due to strong broadcast stations.

5. Local-oscillator energy has a tendency to be radiated by the antenna. Such radiation must be suppresed to prevent interference with neighboring television receivers.

Television tuners can be classified according to the type of RF amplifier circuit employed or the mechanical means of channel selection. From the standpoint of the type of RF amplifier circuit, tuners can be classified as pentode, triode, tetrode, and cascode.

From the standpoint of mechanical means of channel selection, tuners can be classified as continuous (variable inductance and variable capacitance), push-button, switch, and turret types.

PENTODE RF AMPLIFIERS

The pentode RF circuit like the one in Fig. 12-1 was the first to be used; it gets its name from the type of tube employed, a sharp-cutoff pentode. Its chief advantage is that it performs satisfactorily, without any neutralization, from a B+ supply volt-

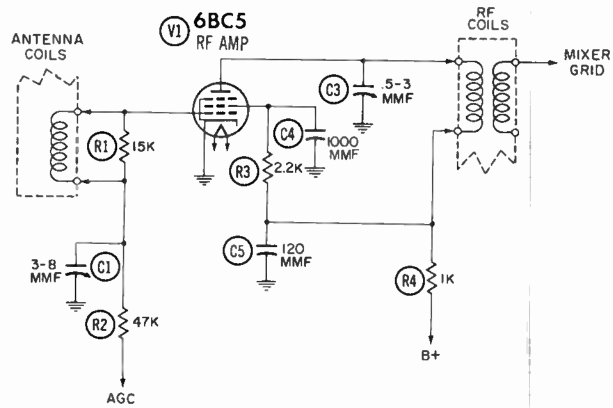


Fig. 12-1. Pentode RF amplifier circuit.

age as low as 100 volts. The fact that this circuit generates more noise than a triode circuit and is, therefore, not too suitable for fringe-area operation does not lessen its usefulness in strong-signal areas. For the most part, the circuit is quite stable, and tube replacement or minor component aging has only a small effect on its frequency-response characteristic.

The RF amplifier stage must govern most of the selectivity and sensitivity of the tuner. The grid circuit is very broadly tuned, and the plate circuit presents the RF signal with an impedance as high as possible to maximize the gain and selectivity. The necessary bandwidth is further governed by the degree of coupling between the RF-plate and mixer-grid coils.

TRIODE RF AMPLIFIERS

When a triode is used with a grounded cathode as an RF amplifier, a relatively high capacitance exists between the plate and the grid. A feedback signal will be impressed on the grid because of this capacitance, and the stage will oscillate. One way this oscillation can be eliminated is by neutralizing the stage, which is done by providing a second feedback path. An out-of-phase signal can be applied through this path to the grid to cancel the plate-to-grid signal.

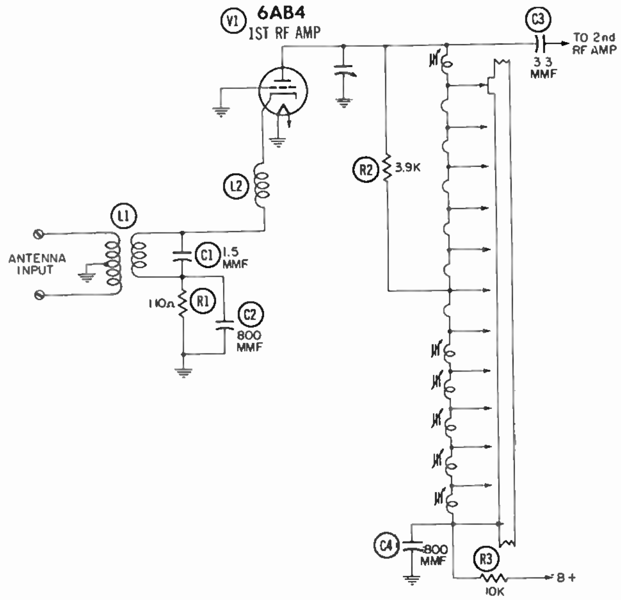


Fig. 12-2. RF amplifier using a grounded-grid triode.

The original solution to the oscillation problem was a grounded-grid circuit like the one in Fig. 12-2. If the grid is grounded and the signal is applied to the cathode, stable amplification can be attained because the grounded grid acts as a shield and eliminates all interelement coupling within the tube.

NEUTRODE RF AMPLIFIERS

Some of the latest receivers use a new version of the triode RF amplifier called the neutrode circuit. The schematic for this circuit is shown in Fig. 12-3. As the name implies, this circuit is a neutralized triode RF amplifier. The neutrode circuit is employed in a newly designed turret tuner that also features a printed-wiring board for the RF and oscillator-mixed circuits.

The RF amplifier is neutralized by a capacitor connected from the low side of the plate coil to the control grid. A 5-mmf fixed capacitor is paralleled with the trimmer to increase the minimum capaci-

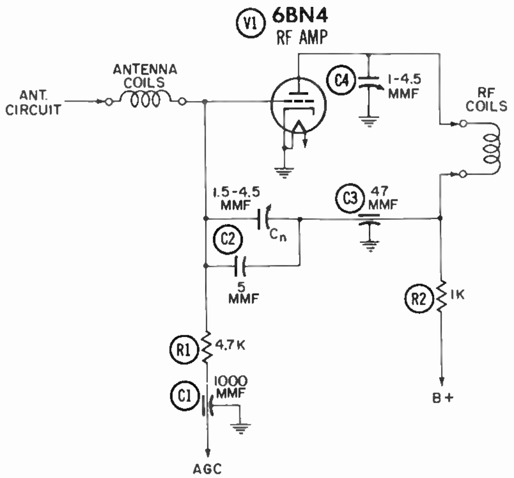


Fig. 12-3. The neutrode RF amplifier circuit.

tance. The required 180° phase inversion is obtained through the plate coil. The amount of feedback is controlled by adjustment of C_n to obtain the exact amount of signal required for most effective neutralization.

The performance of the neutrode tuner compares favorably with that of cascode tuners. Fields tests have shown that about 32 db of gain with less than 8 db of noise can be achieved.

TETRODE RF AMPLIFIERS

A further development in the continuing search for simpler circuits was the tetrode RF amplifier circuit in Fig. 12-4. This circuit is designed around a tetrode series which includes the 2CY5, 6CY5, etc. These tetrodes feature a high transconductance (8,000 micromhos) with a noise figure approaching that of some of the RF triodes.

Circuitwise, the tetrode RF amplifier is almost identical to the pentode RF amplifier. Of course, there is no suppressor grid in the tube; and there is a small inductance in series with the screen-grid bypass capacitor. This inductance is actually in the lead of the capacitor itself, and it has a stabilizing effect on the stage. The tetrode tuner will provide about 35 db of signal gain at a 6- to 8-db noise figure, which is comparable to the performance of cascode RF amplifiers.

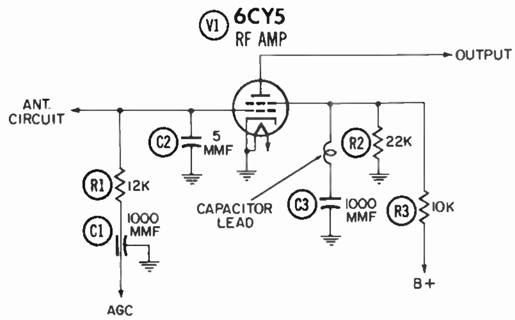


Fig. 12-4. The tetrode RF amplifier circuit.

CASCODE RF AMPLIFIERS

The cascode circuit was developed as a result of efforts to improve the signal-to-noise ratios of TV tuners. High signal-to-noise ratios are needed for good fringe reception. The cascode circuit in Fig. 12-5 consists of a grounded-cathode triode driving a grounded-grid triode; both triodes are connected in series as far as plate current is concerned. The cathode input impedance of the grounded-grid stage is the plate load for the first stage and is such that maximum power gain is attained for both stages. The cascode tuner has the high gain normally associated with a pentode amplifier and the low noise of a triode amplifier. The average cascode tuner will have about 35 db of gain with about 6 db of noise.

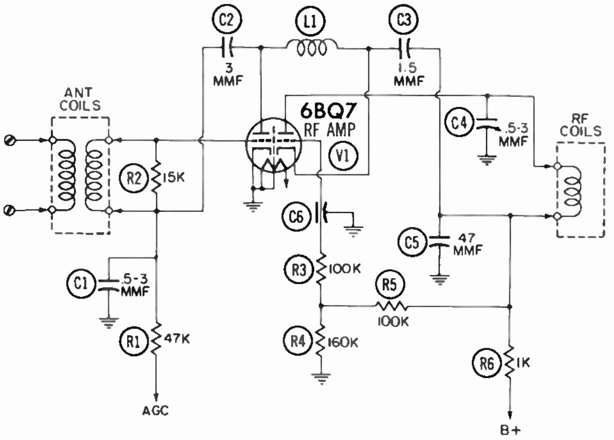


Fig. 12-5. The cascode RF amplifier circuit.

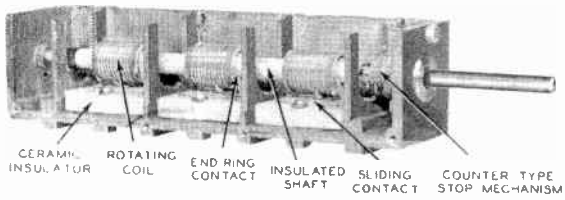
Like any other grounded-cathode triode amplifier operating at radio frequencies, the input triode must be neutralized. This is done by coil L1, which develops a small signal voltage that is then fed back to the bottom of the grid coil through capacitor C2.

TUNING SYSTEMS

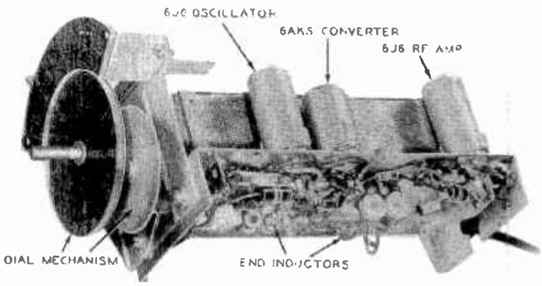
Variable-inductance tuning is done by the use of sliding electrical contactors or sliding iron cores to vary the inductance.

Tuning by Continuously-Variable, Contactor-Type Inductor

Fig. 12-6 illustrates an older type of tuner in which the inductance of the circuit is continuously variable. Such a method of tuning is particularly adapted to VHF because a high ratio of inductance to capacitance and, consequently, higher circuit impedance at the high-frequency end of the range is possible. A sliding contactor made of a high silver-content alloy rides in trolley fashion on a silver-wire inductor. End rings on the inductor permit connection to both ends of the inductance. The sliding contactor shorts



(A) Top view of tuning mechanism (shield removed).



(B) View of associated circuit elements.

Fig. 12-6. A continuously variable-inductance tuner for television reception.

the unused portion of the inductance. The tuner in Fig. 12-6 is a three-gang inductor equivalent to a three-gang capacitor. In the input system shown in Fig. 12-6A and schematically in Fig. 12-7, the three units are used in a coupled bandpass selector stage and in the local oscillator. In Fig. 12-7, coupling elements C5 and C7, and the shunt circuit L6-C6, are so proportioned that the acceptance band remains constant at six megacycles over the entire tuning range. The tuning range extends continuously from 44 to 216 megacycles and includes not only the television channels, but also FM stations, two amateur bands, aviation channels, VHF radio telephone, and in commercial services. A unique dial arrangement (Fig. 12-6B) makes it possible to use the television receiver for reception of FM also.

Adjustable end inductors L5, L7, and L9 (Fig. 12-7) perform the same function circuitwise as the high-frequency trimmers of a variable tuning capacitor. The oscillator circuit shunt inductor L8 is equivalent to the series padder of the conventional superheterodyne oscillator circuit.

The service technician will recognize other features of the schematic in Fig. 12-7 as being similar to the input systems of lower-frequency entertainment receivers. The injection of local-oscillator voltage by capacitance coupling between the oscillator and the control grid of the converter tube is a conventional method at VHF frequencies.

Continuously-Variable Tuning by Powdered-Iron Cores

Fig. 12-8 shows an early tuning system in which a special type of high-frequency, powdered-iron core can be moved to change the value of the tuned-circuit inductances. This permits selection of television

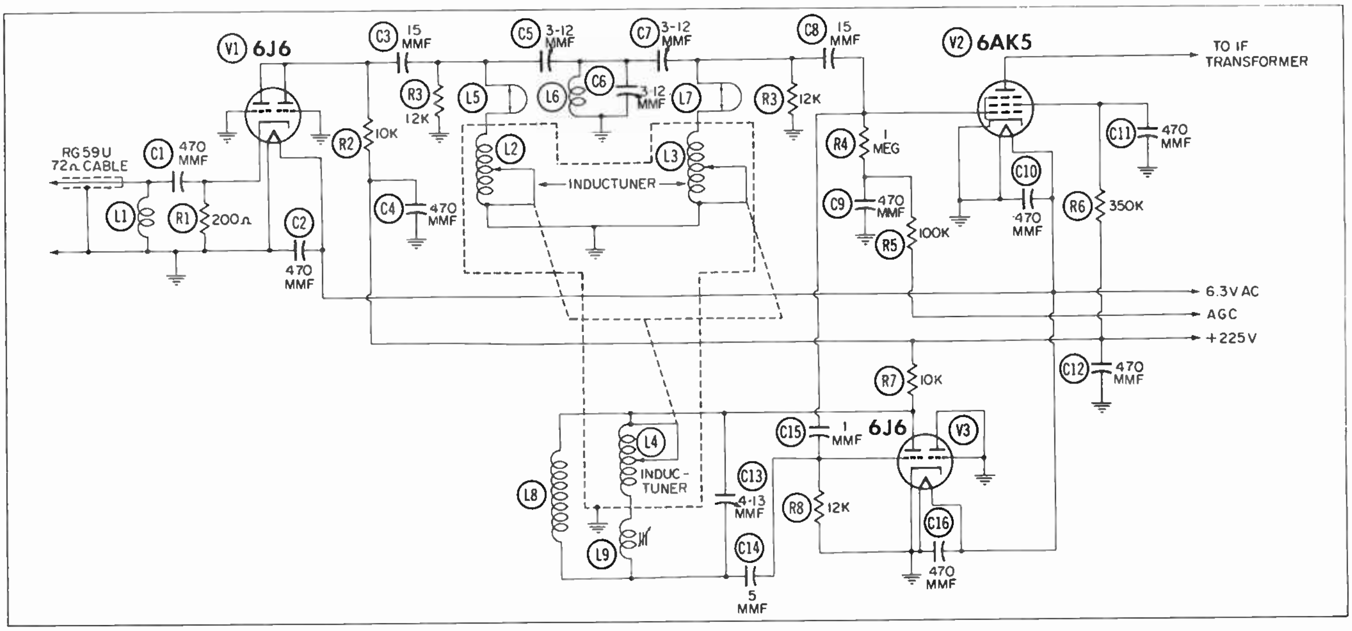


Fig. 12-7. Schematic of inductively-tuned input system illustrated in Fig. 12-6.

channels on both low- and high-frequency bands. A separate set of inductors and sliding cores are provided for each band, and electrical switching transfers from one band to the other. Fig. 12-9 shows the electrical circuit for this input system. Since tuning is continuous in this circuit and in that of Fig. 12-7, a separate fine-tuning control is not needed. The tuning adjustment is set for the best performance of the sound channel, and this automatically assures a proper setting for the video carrier.

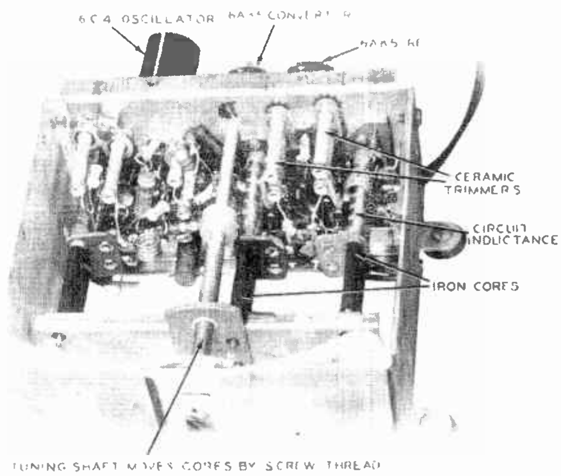


Fig. 12-8. A variable-inductance tuner employing movable powdered-iron cores.

Antenna primary coils L1 and L3 in Fig. 12-9 are balanced to ground, and a center tap is connected to the chassis ground. Thus, a balanced parallel-lead line can be used for the transmission line. Both the RF and converter stages employ miniature 6AK5's in pentode connection. Broadband response is obtained by loading the tuned circuits with low-value parallel resistors. In the high-frequency band, parallel resistors are not required because of higher circuit losses.

Capacitively-Tuned Input Systems

A tuner which uses a variable capacitor is shown in Fig. 12-10. This tuner appeared in some of the early television receivers, and is no longer used. The schematic in Fig. 12-11 shows that the inductors for each tuned circuit were switched for changing from low to high television channels.

What might be considered as being another type of capacitively-tuned tuner is shown in Fig. 12-12, and its schematic is shown in Fig. 12-13. This tuner is more properly called the push-button type because push buttons inserted a set of pretuned capacitors into the circuit for each channel. This tuner was also introduced in early receivers, and has since been replaced by newer types.

Artificial-Line Input Tuner

Fig. 12-14 shows a type of input system found in many early receivers. A 6J6 twin-triode provides push-pull operation in each stage. The artificial equivalent of a quarter-wave transmission line, consisting of inductors with their associated distributed capacitors, is employed in the RF, converter, and oscillator circuits to perform the functions usually associated with lumped-tuned circuits. These lines are balanced with respect to the chassis or ground. The various television channels are tuned by switching a short circuit along the line.

RF amplifier tube V1 (6J6) of Fig. 12-15 is push-pull connected and is cross-neutralized by capacitors C4 and C5. This cross-neutralization increases the

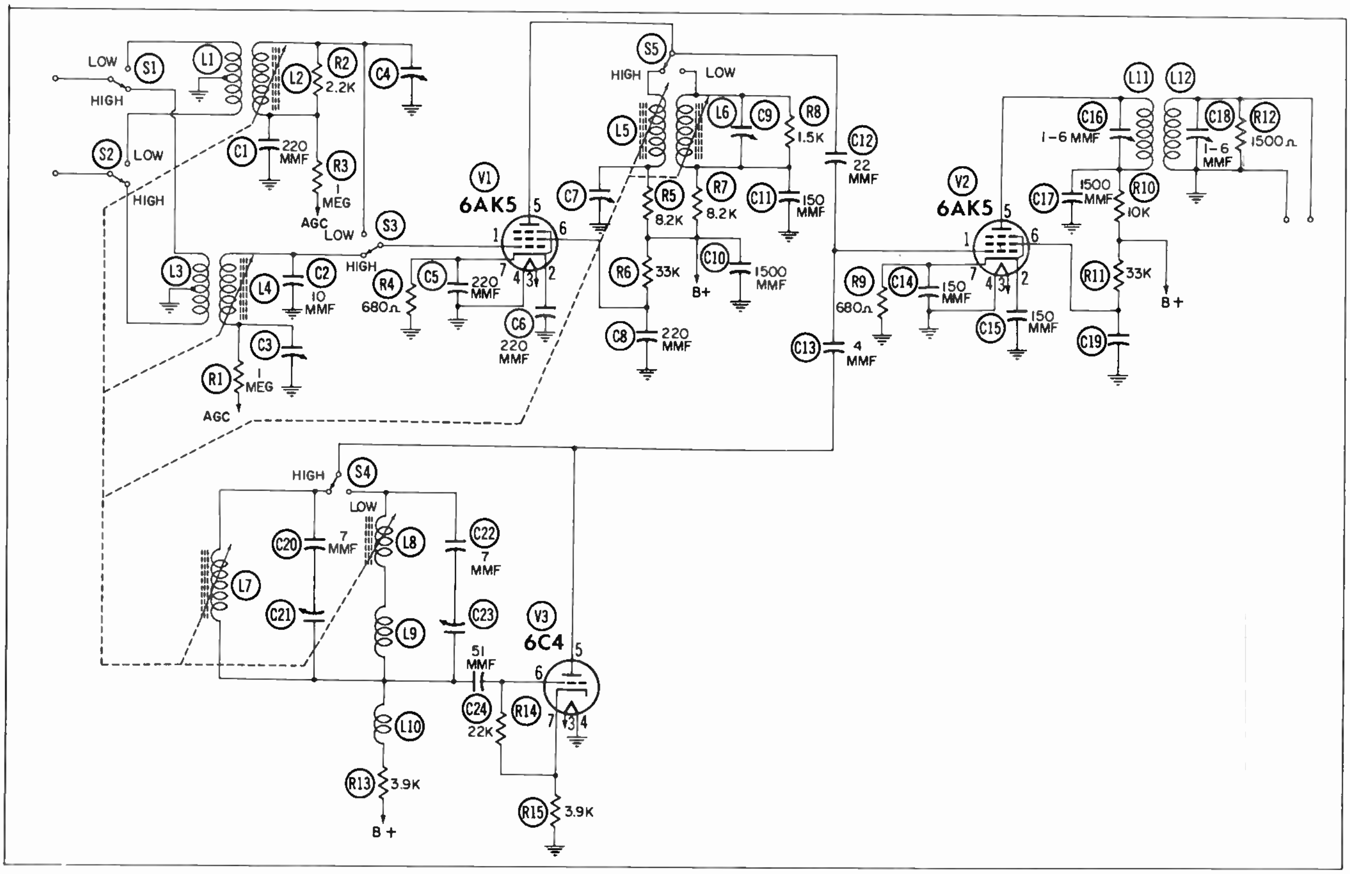


Fig. 12-9. Schematic of the iron-core tuning system illustrated in Fig. 12-8.

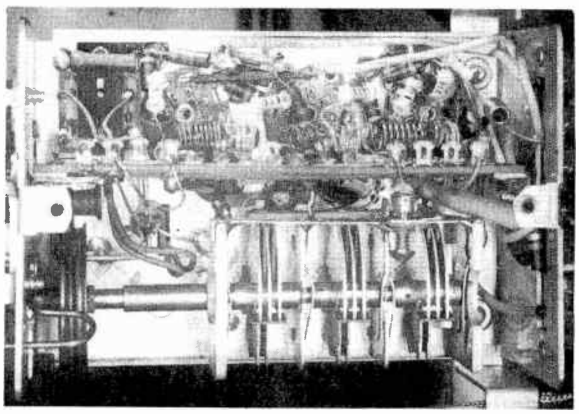


Fig. 12-10. A tuner which uses a variable capacitor.

stability of the stage by preventing regeneration or oscillation and also reduces the transmission of oscillator energy through amplifier tube V1 to the antenna.

The plate load of V1 consists of the equivalent of a quarter-wave transmission line and is made up of inductors L1 through L26. A rotary switch shorts out sections of the line so that it can be tuned to any one of the twelve VHF television channels.

Capacitors C2 and C3 and inductive link L36 act as coupling elements betwen the line in the RF plate circuit and in the corresponding line comprised of inductors L31 through L57 in the converter-grid circuit. Because of the value of these elements, the channel width is kept constant over both VHF television bands.

The oscillator circuit consists of a transmission line similar to the ones just described, but with the additional feature that each channel inductance is separately adjustable (see inductors L76 through L87). The oscillator tube (a push-pull connected 6J6) derives its plate-to-grid feedback for sustained oscillation from a crossed pair of capacitors (C25 and C27), the values of which are greater than those of the grid-to-plate capacitance of the 6J6.

Oscillator voltage is injected into the converter-grid circuit by magnetic coupling between the tuned transmission lines, augmented by coupling link L74.

Input Systems with Switch-Selected Variable Inductors

Figs. 12-16 and 12-17 show an early tuner in which a rotary wafer switch selects the proper inductances for tuning to the desired television channel. Several features not previously mentioned are evident in this circuit. The input circuit and the method by which tube V1 is connected constitute a means of coupling between a balanced-to-ground transmission line and a single-ended tube circuit. Both the grid and the cathode act as input elements and are con-

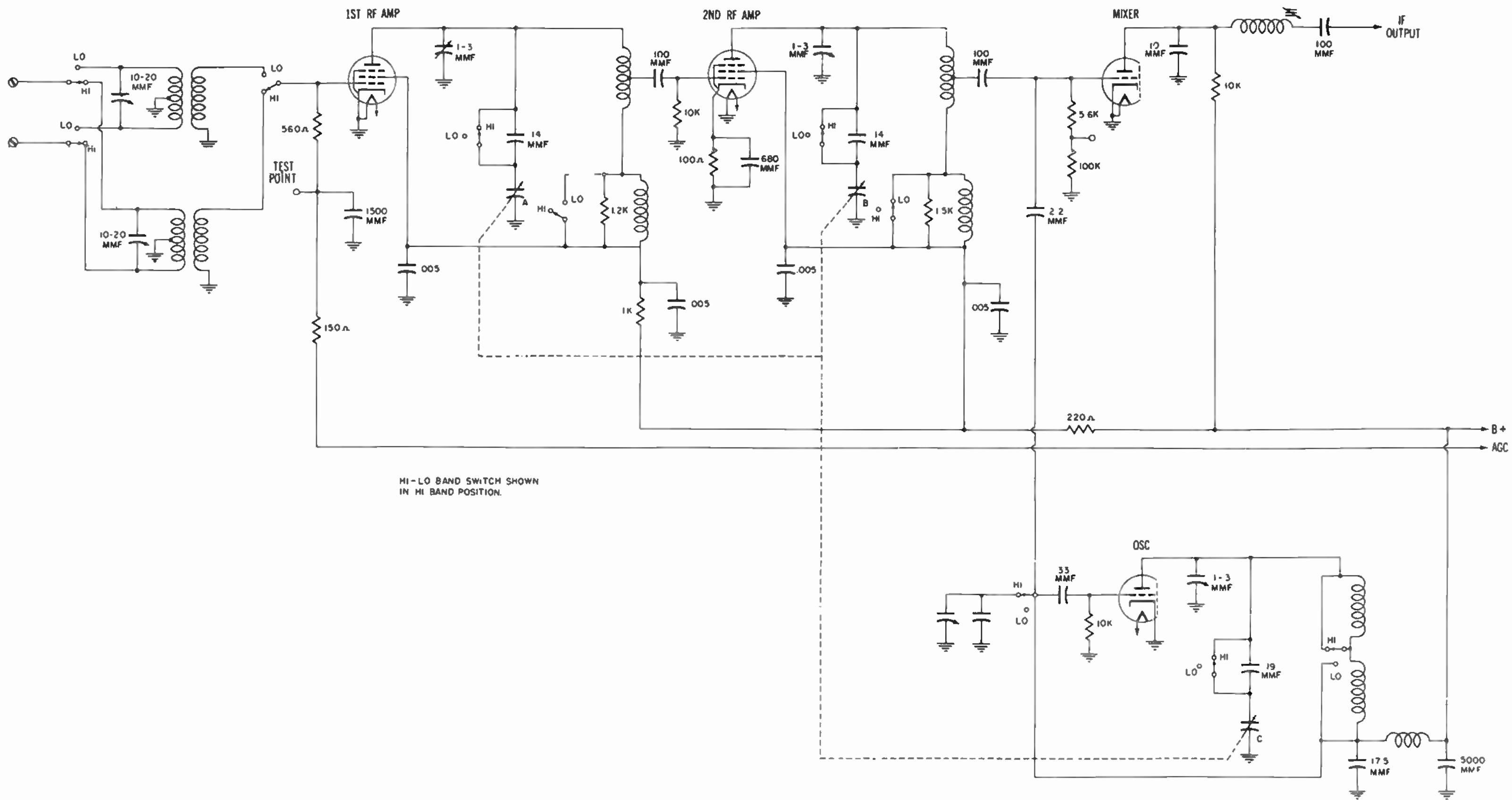


Fig. 12-11. Schematic of the tuner shown in Fig. 12-10.

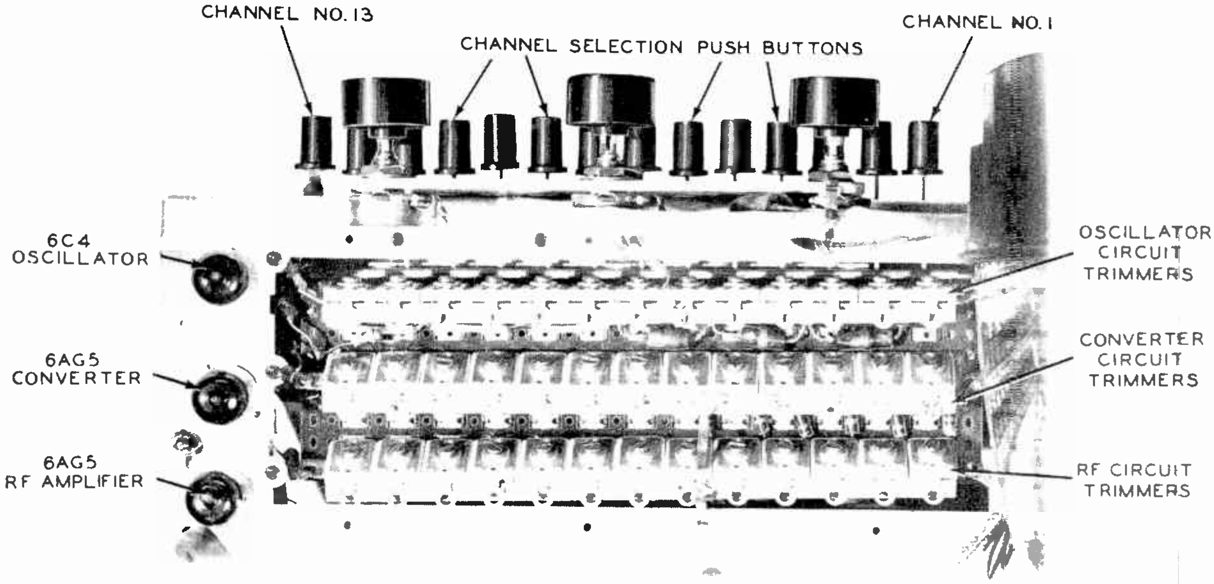


Fig. 12-12. Tuner employing push-button, trimmer-tuned circuits.

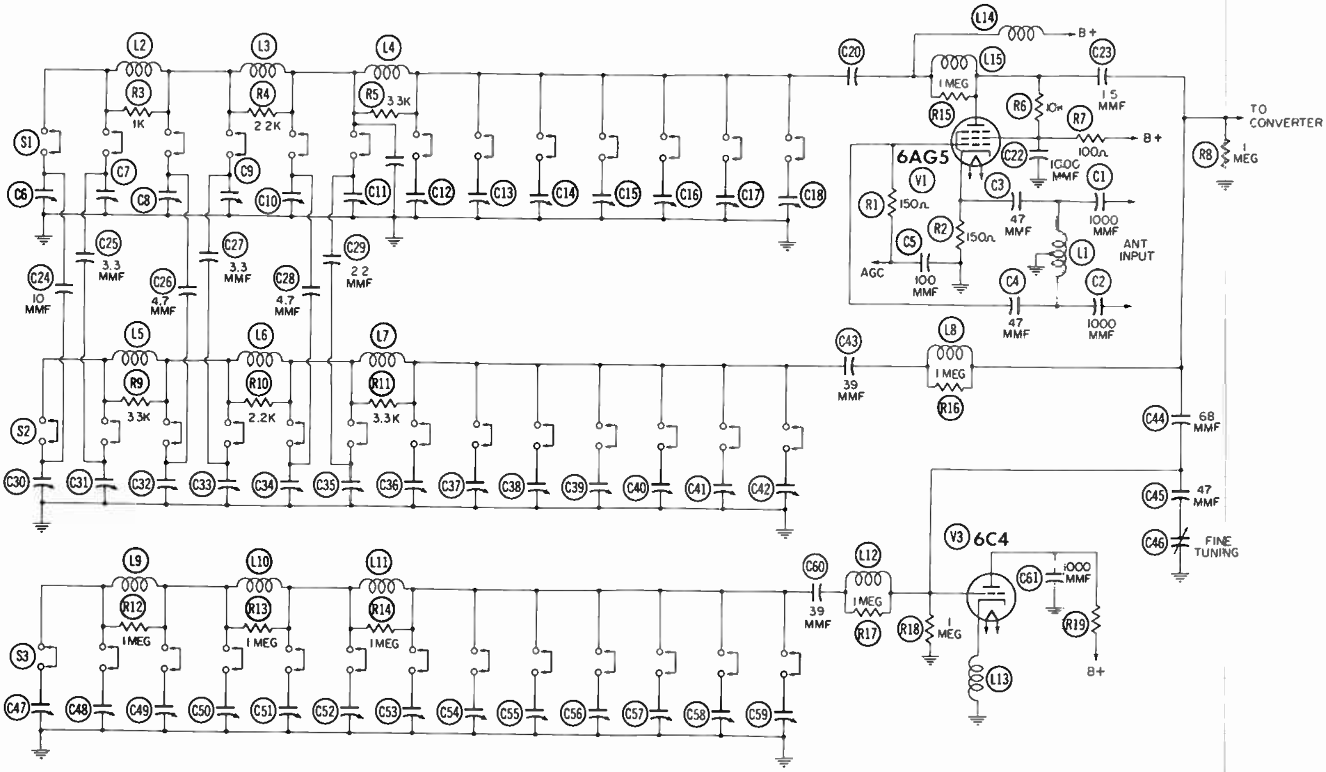


Fig. 12-13. Schematic of tuner shown in Fig. 12-12.

nected to opposite ends of center-tapped inductor L3. On the high-frequency band (channels 7 through 13), inductors L1 and L2 in series are connected across L3 and resonate broadly at the higher frequencies.

The interstage coupling system consists of two tuned circuits with coupling adjusted to obtain the proper bandwidth for all channels. The plate- and grid-circuit inductances (L4 through L15 and L16 through L27) are selected by the rotary switch. The tuning capcitances consist of the capacitances of the tubes and of the wiring to ground. Both the circuit shunt loading (R6) and the coupling (C5, C7, C8, and C9) are switch-controlled to provide uniform (6-mc) bandwidth for all channels. The oscillator-circuit switching involves the selection of individually-adjusted inductors (L30 through L40). Oscillator-tuning capacitor C13 is made variable for fine tuning.

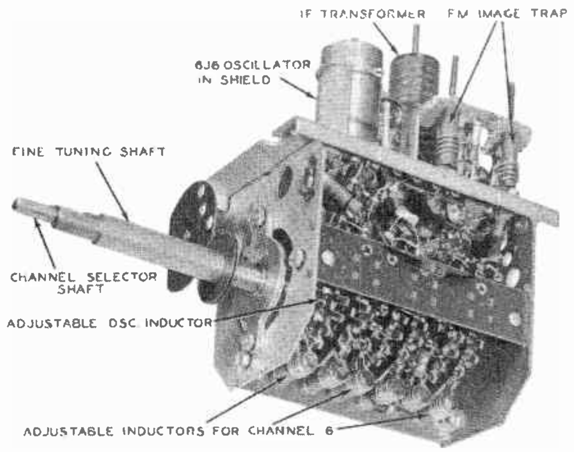


Fig. 12-14. Input tuning system employing lumped equivalents of quarter-wave transmission lines.

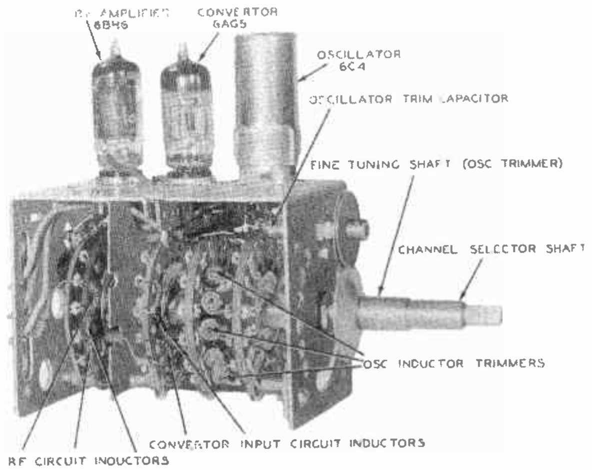


Fig. 12-16. Input tuning system employing rotary-switch inductors.

Some of the present-day receivers still use the switch-type tuners, but in their modern form. There is very little difference between the old and the new tuners of this type. Modern RF amplifier circuits include the pentode, cascode, tetrode, and the new tube types; the mixer and oscillator tubes are contained within one tube envelope. The very latest switch type tuner is pictured in Fig. 12-18, and its schematic, in Fig. 12-19.

Turret Tuners

Figs. 12-20 and 12-21 show a widely-used tuning system in which the input, converter, and oscillator tuned circuits are mounted on a rotating turret. Only the circuit elements associated with a single television channel are connected in the circuit at any one time. Spring contacts associated with the various tube circuits provide a means for connection to the terminals of the tuned circuits. As the turret is rotated by the channel-selector knob, studs at the end of each set of coils are positively indexed into contacting position by wiping action which assures low-circuit resistance.

The coils are mounted on Bakelite moldings which are clipped into the turret drum. Two coil strips are used for each channel. One strip (L1) contains the input coils, and the other strip (L4) has the RF plate, mixer grid, and oscillator coils. A spring-

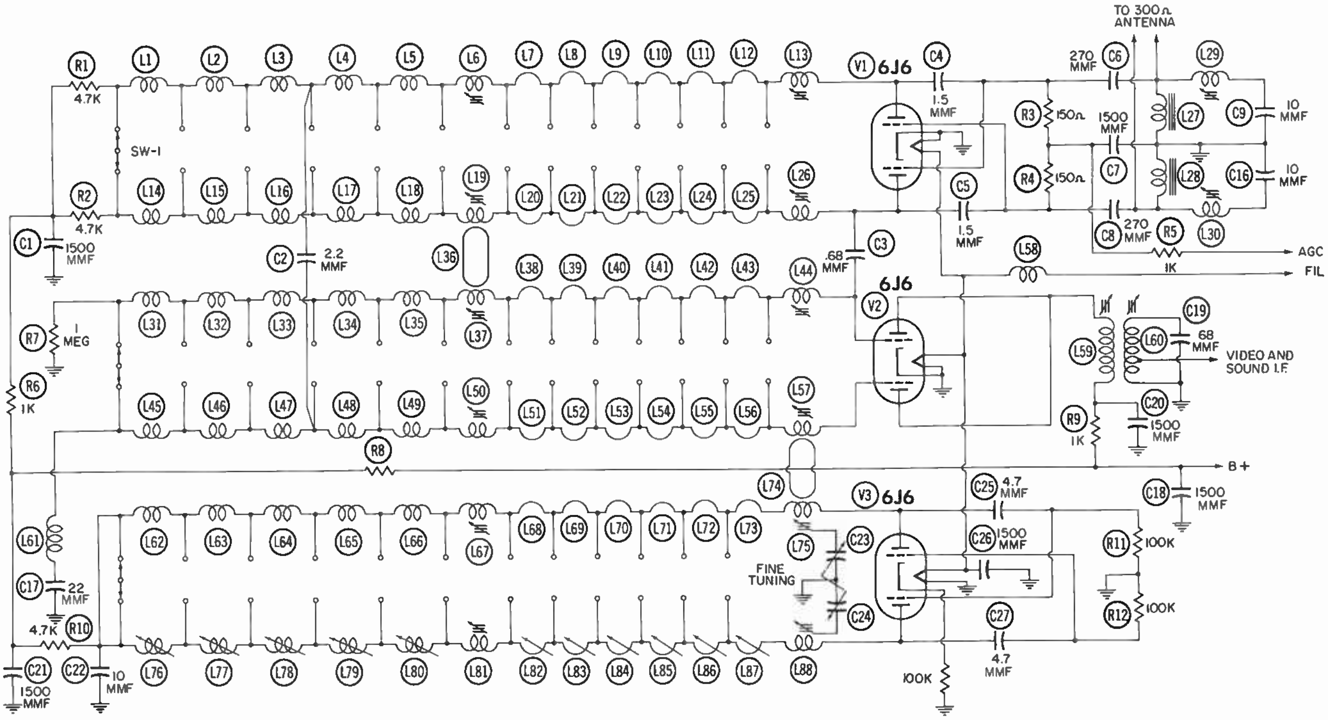


Fig. 12-15. Schematic diagram of tuning system illustrated in Fig. 12-14.

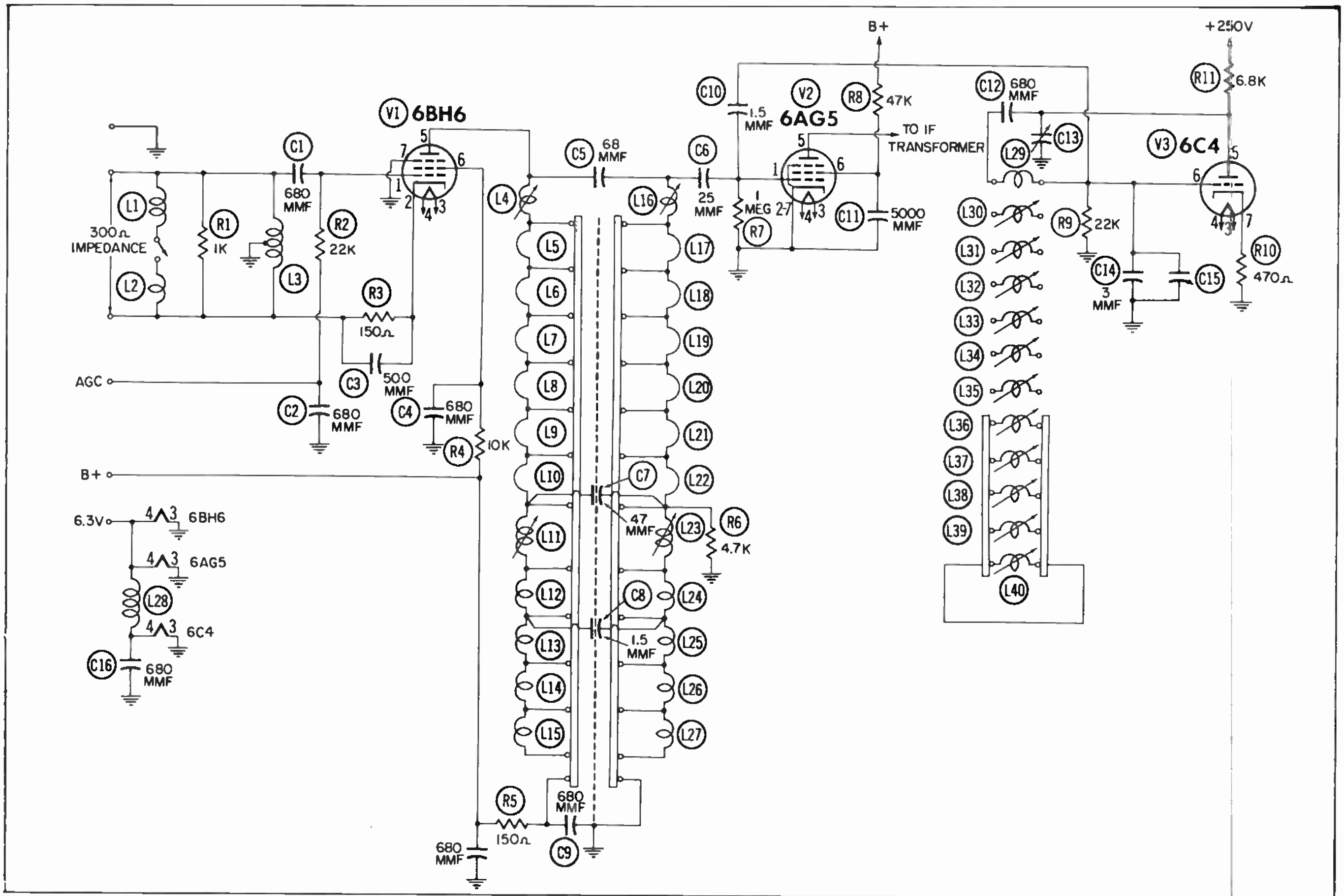


Fig. 12-17. Schematic diagram of tuner illustrated in Fig. 12-16.

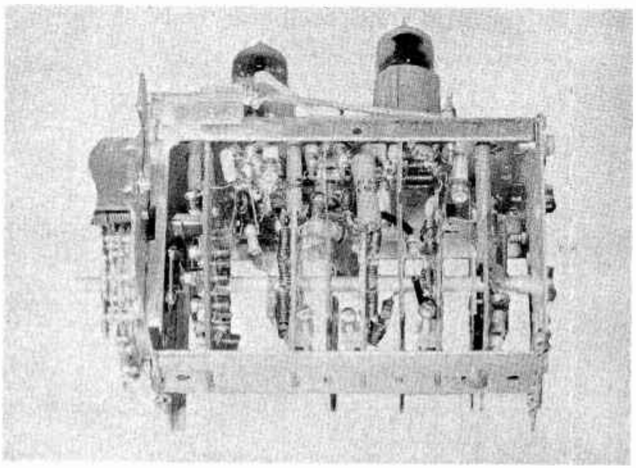


Fig. 12-18. A modern switch-type tuner.

loaded arm with a roller engages the detent plate to lock the turret into position at the desired channel. The detent plate also acts as a shield between L1 and L4. The spring contacts are mounted on an insulated strip and contact the coil contact buttons. The fine-tuning mechanism is controlled by a shaft which is concentric with the channel-selector shaft. Each oscillator coil has a tuning slug which can be reached through a hole at the front of the tuner.

Disc Tuners

A further development of the turret tuner is the disc tuner. Fig. 12-22 shows a disc tuner which combines certain principles of both the turret- and switch-type tuners. Attached to the shaft inside the tuner are two large discs upon which the tuning inductors are arranged. The antenna coils are on the upper disc. A detent mechanism resembling that of the turret tuner is mounted on the lower disc. Contact buttons and fingers, like those of the turret tuner, connect the disc to the external circuitry.

An important difference between the turret and disc tuners is that the disc tuner does not have a completely separate set of coils for each channel. There are several sets of basic coils plus incremental inductances which are switched in series with the basic coils to tune certain channels. The incremental design is commonly used in wafer-switch tuners. The coils are broken into more separate groups in the disc tuner than in the typical switch tuner.

All of the contact buttons on the under side of the upper disc are arranged in six concentric circles. The buttons in the intermost two circles are connected to the plate coils for the RF amplifier, the middle two are connected to the mixer input circuit, and the outer two are connected to the oscillator

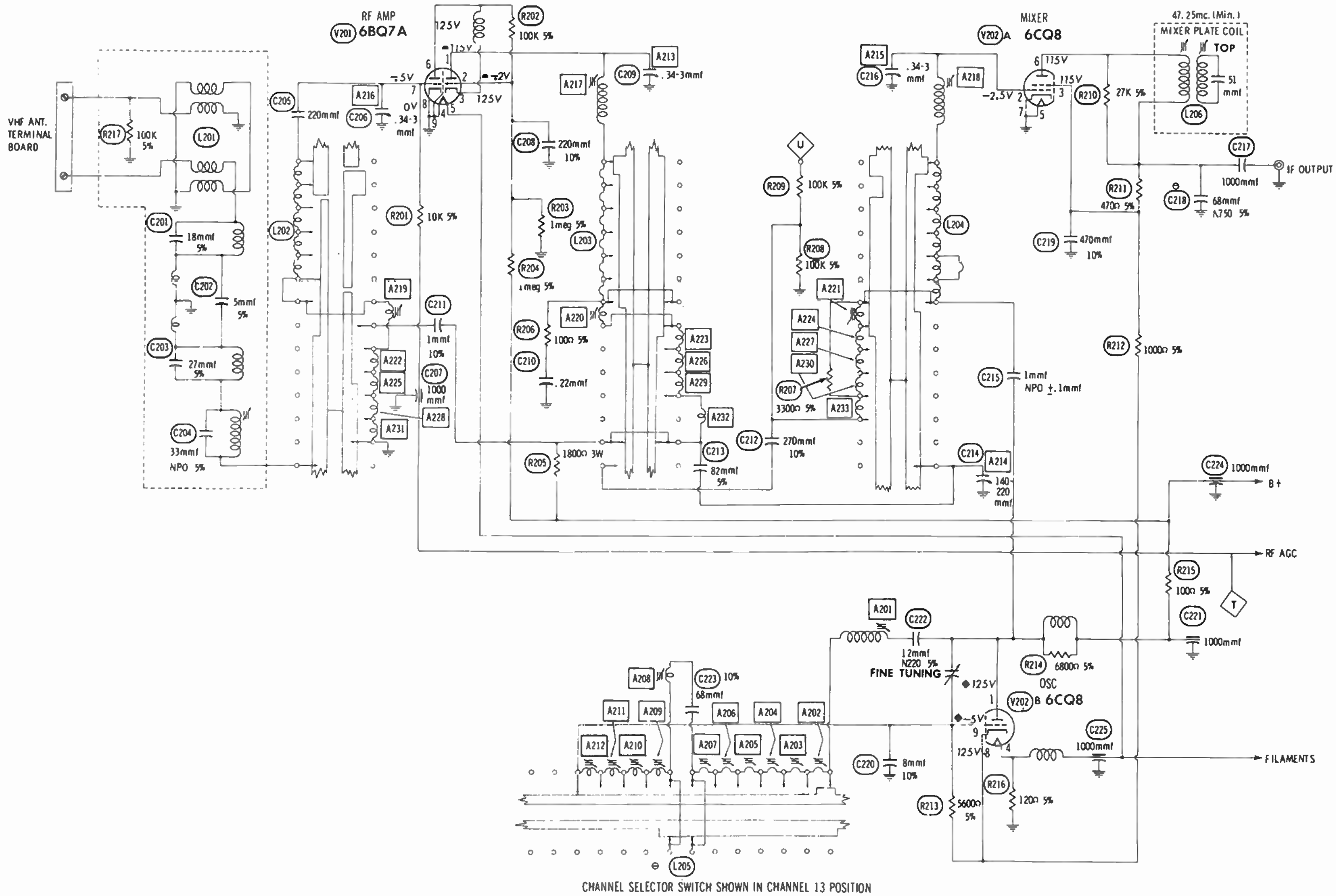


Fig. 12-19. Schematic of the switch-type tuner shown in Fig. 12-18.

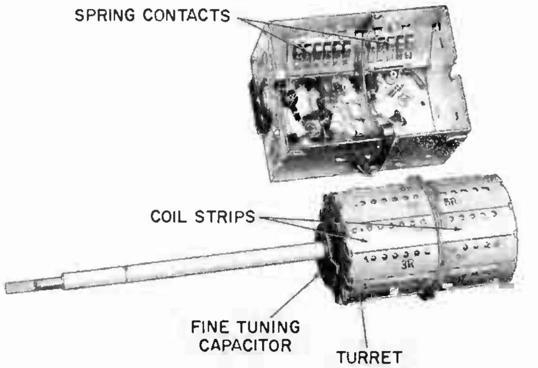
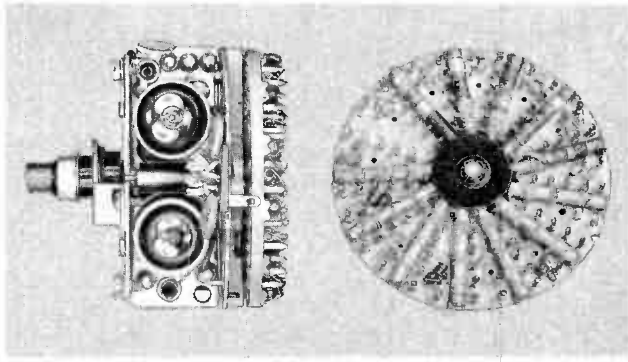


Fig. 12-20. A turret-type tuner.



(A) Top view. (B) End view.

Fig. 12-23. A smaller disc-type tuner.

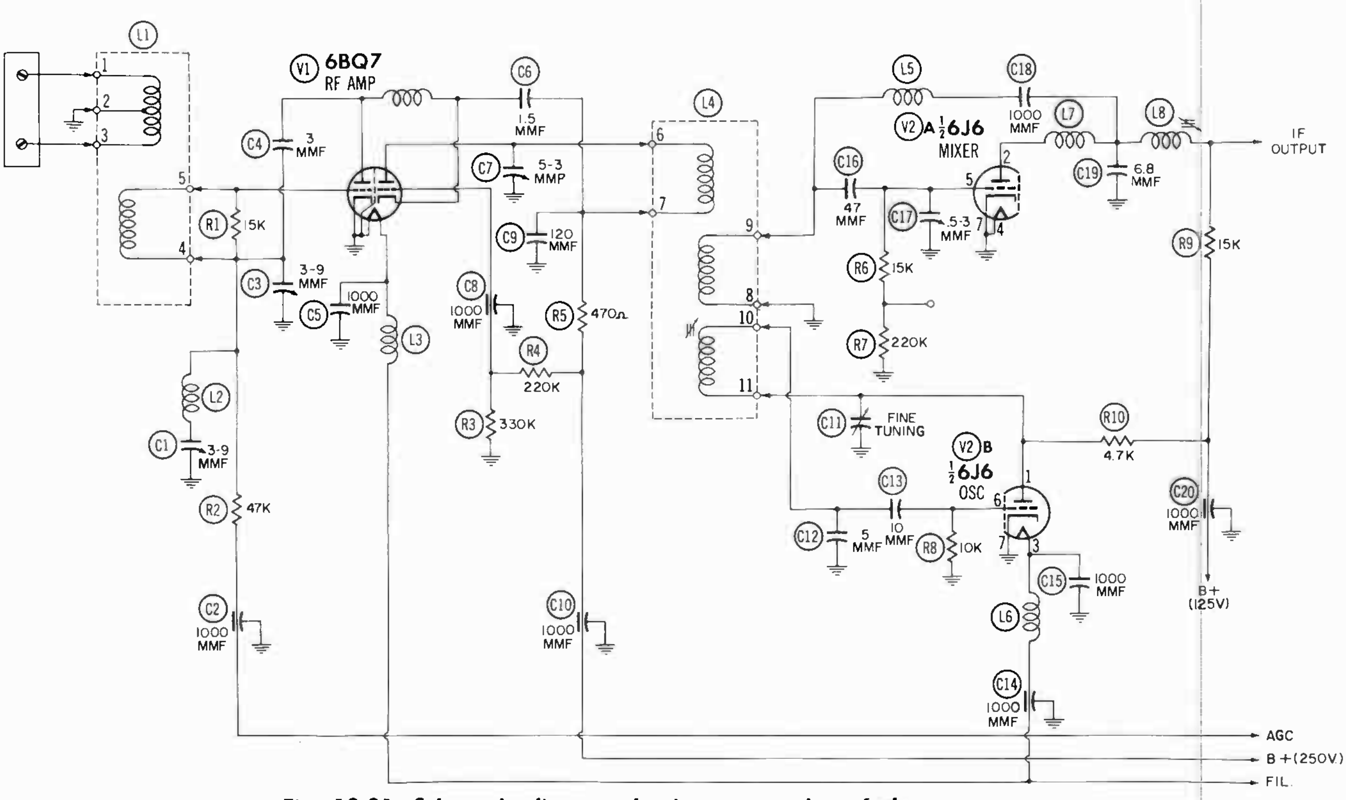


Fig. 12-21. Schematic diagram showing one version of the turret tuner.

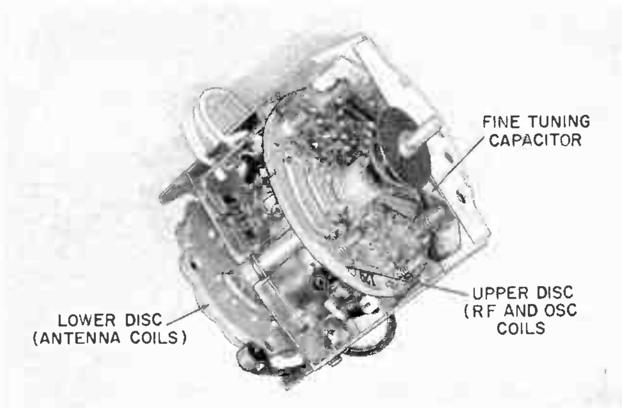


Fig. 12-22. A disc tuner.

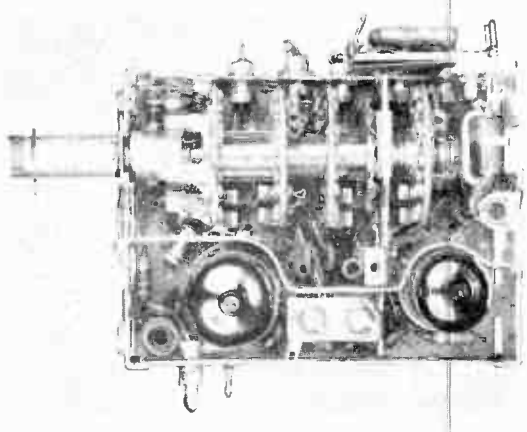


Fig. 12-24. Another version of the disc-type tuner.

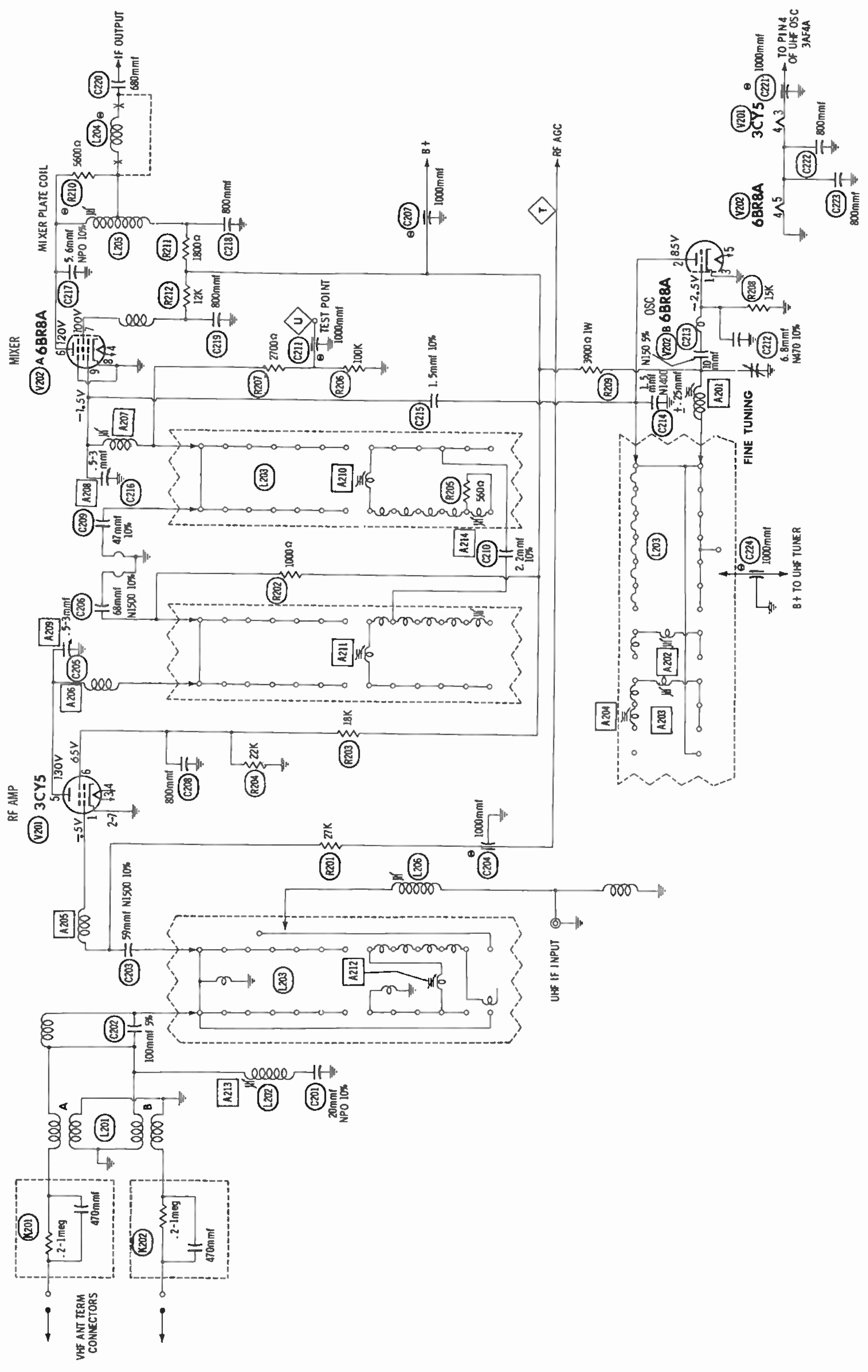


Fig. 12-25. Schematic diagram of tuner shown in Fig. 12-24.

coils. A 1.5-mmf capacitor connects the low-band RF coils to the corresponding mixer coils.

The fine-tuning capacitor is located at the top of the tuner (Fig. 12-22). The movable plate is a strip of metal attached to the tuning shaft, and the fixed plate is printed on a small wiring board.

A top view of a still smaller disc tuner is shown in Fig. 12-23. The antenna, RF, and oscillator coils for all 12 TV channels have been mounted on one disc. Connections to these coils are made by spring contacts like those in a turret tuner. Proper positioning of the wheel for each channel is assured by a detent mechanism like the one in a turret tuner.

Fig. 12-24 shows one of the latest tuners. It resembles the switch type, yet should be listed as a disc tuner because the tuning inductors are mounted on discs which rotate against stationary contacts. A schematic of this tuner is shown in Fig. 12-25.

UHF TUNERS

The preceding tuners in this chapter are intended for receiving the 12 VHF channels. Since the release in 1952 of the 70 UHF channels, tuners for receiving these channels have appeared.

The reception of frequencies between 470 and 890 megacycles presented many problems to the design engineer. The tubes available in 1952 were very inefficient at these frequencies. Also, the frequencies are too high for conventional lumped coil and capacitor tuning methods and too low for the waveguides and resonators employed in receivers operating at 1,000 mc and beyond. The methods which have been developed tend to strike a compromise between the lumped inductance and capacitance of low-frequency circuits and the distributed inductance and capacitance of high-frequency circuits. It is well to remember that a straight length of wire can possess inductance and that capacitance can exist wherever there is a difference of potential between two surfaces.

When the UHF channels were released, there were many VHF receivers already in the field. To enable these receivers to pick up the UHF broadcasts, UHF converters were designed. A UHF converter is simply a UHF tuner and mixer-oscillator circuit in its own cabinet and with an output signal on one of the VHF channel frequencies. This output frequency was usually adjustable, so that it could be placed on one of the unused VHF channels.

Those early receivers having turret tuners could be internally modified for UHF reception by removing unused channel strips from the turret and placing UHF channel strips in their place. Receivers without turret tuners had to use converters. Some present-day receivers have built-in UHF tuners.

Many methods of tuning have been used for UHF reception. One of the most common has been the shorted quarter-wave or half-wave transmission line with a movable short. A transmission line possesses both inductance and capacitance, and these are distributed along the line. Any change in the length of the line will change the frequency to which it is tuned. To change the frequency, the movable shorting bar is ganged with the tuning control and dial indicator.

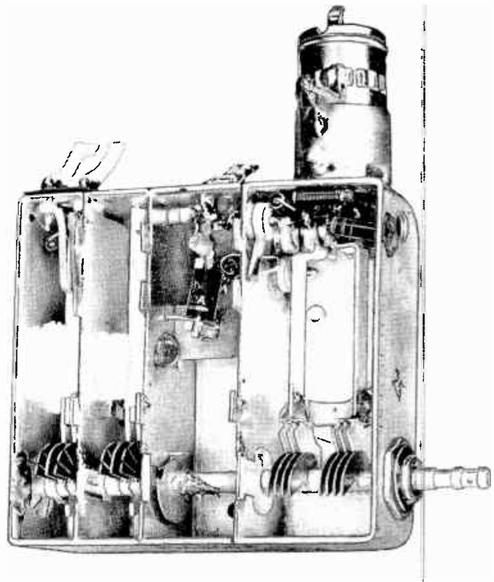


Fig. 12-26. A UHF tuner with capacitance tuning.

Another popular form of tuning is a transmission line with a variable capacitor across the open end of the line. This method is as good as the sliding shorting-bar method; an added feature is that no moving contacts are needed. Fig. 12-26 shows a modern UHF tuner that tunes a transmission line by a variable capacitor.

QUESTIONS

1. Name the stages in an RF tuner.
2. Most of the selectivity and sensitivity of the tuner is governed by what stage?
3. In the neutrode RF amplifier, how is feedback accomplished?
4. What are the identifying features of a cascode RF amplifier?
5. What circuit components are mounted on the clipped-in strips of a turret tuner?
6. The principles of what two types of tuners are combined in the disc-type tuner?

Chapter 13

Video IF Amplifiers and Detectors

INTRODUCTION

In our discussion of the composite television signal carrying amplitude-modulated video and frequency-modulated sound, we said that a six-megacycle channel is required. Since the television receiver is a high-frequency superheterodyne, signal amplification and selectivity are obtained mostly in the intermediate-frequency amplifier system. The television intermediate-frequency amplifier system differs in one major respect from the familiar broadcast or short-wave type. The signal at the converter grid consists of two carriers separated by 4.5 megacycles. These two carriers are the video (picture) carrier and the audio (sound) carrier. The information in the modulation of these carriers must be separated and converted to a video voltage at the picture-tube grid and to an audio voltage at the grid of the output tube feeding the loudspeaker, respectively.

Early television receivers had two intermediate-frequency amplifier systems, the first for video IF amplification and the second for audio IF amplification (Fig. 13-1). Since the two carrier frequencies are accurately separated at the transmitter by crystal control, the receiver local oscillator, by beating with the composite television signal, automatically produces two intermediate frequencies in the plate circuit of the converter or mixer. These intermediate frequencies are separated in the so-called "split-sound" system, and each is given separate amplification and detection as shown in Fig. 13-1. The audio IF passband, however, is quite narrow; and any drifting of local-oscillator frequency produces a loss of audio output.

The fact that the video carrier and audio carrier are always separated by a fixed difference of 4.5 megacycles makes possible the present-day method of separating the video and audio signals. This system, known as intercarrier sound, was proposed originally as a means of simplifying military television equipment. A block diagram of the intercarrier sound system is shown in Fig. 13-2. The IF amplifier response is made broad enough to accept both the video and the audio IF carriers. At the output of the second detector, a new beat note of 4.5 mc appears between the two signals. This beat is in effect a new intermediate frequency and is frequency-modulated in accordance with the audio signal. Since the video-

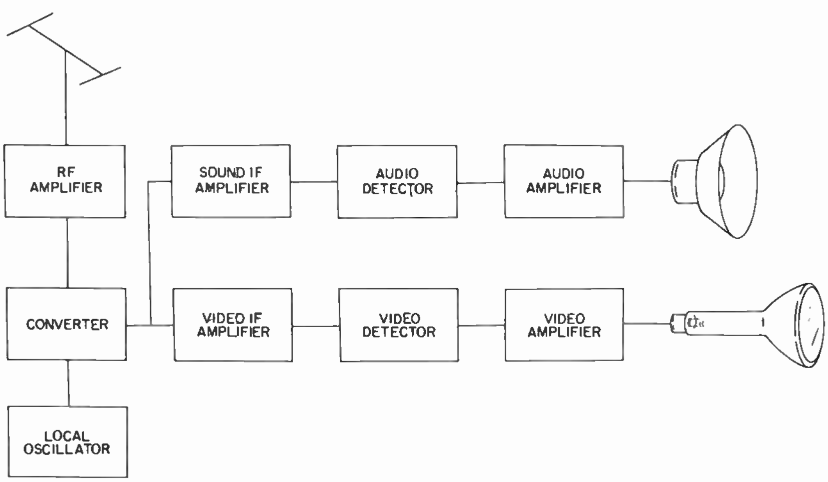


Fig. 13-1. Block diagram of a dual-IF channel television receiver.

amplifier response is wide enough, this frequency-modulated, 4.5-megacycle signal appears in amplified form at the picture-tube grid. A trap circuit prevents it from modulating the picture tube. The 4.5-megacycle signal at the video-amplifier output is passed to an FM detector and to an audio amplifier. This intercarrier sound system will be considered in greater detail later.

Requirements of the television IF amplifier system are more complex than IF amplifiers in broadcast receivers. In a superheterodyne receiver, the IF amplifier must pass a band of frequencies extending only 10 kilocycles on either side of the center frequency. In FM receivers, the bandwidth need not be more than ±200 kilocycles. In a television receiver, the intermediate-frequency amplifier must pass a band approximately 5 megacycles wide. In addition to this bandpass requirement, interfering signals from adjacent television channels must be rejected.

The factors influencing the choice of intermediate frequency for television are:

1. Bandwidth.
2. Selectivity.
3. Harmonics of the intermediate frequency (which might fall within a television band).
4. Direct IF interference (externally-generated signals having frequencies within the IF band and passing through the RF amplifier).
5. Images due to FM or television stations on the image frequency.
6. Cross modulation due to television and FM stations.
7. Oscillator radiation which will interfere with the operation of other television or FM receivers in the vicinity.

The choice of any particular intermediate frequency must be a compromise in view of the various factors just listed, but the higher intermediate frequencies seem to be the better choice.

VIDEO IF SYSTEMS

The requirements of the video IF system are:

1. The bandwidth must accept a total band approximately five megacycles wide.
2. Frequencies beyond the edges of the passband must be rejected or attenuated so that they cannot interfere with the picture. These frequencies are:

   a. The associated or cochannel sound carrier, which must be reduced to a level 23 to 26 db below the picture carrier so that sound bars do not appear in the picture.

   b. The sound carrier of the adjacent lower-frequency channel.

   c. The video carrier with its modulation in the next higher adjacent channel.

3. The response characteristic of the video-IF amplifier must be so shaped that the lower frequencies corresponding to the double-sideband part of the transmission are properly attenuated to prevent overemphasis.

Required Response Characteristics of the Over-all Video System

An analysis of the transmitter output in Fig. 9-4A reveals that the amplitude is constant from approximately 0.75 megacycles below the video carrier to approximately 4 megacycles above the video carrier. If such a carrier and its sidebands are impressed on a linear detector, the double-sideband nature of the region below the picture carrier would cause high output from the detector for modulating frequencies from 0 to 0.75 megacycles, and about half output for frequencies higher than 0.75 megacycles. Fig. 13-3 shows the output of such an ideal detector. The output over the region from 0 to 0.75 megacycles is twice that of the region from 1.25 megacycles to 4 megacycles; the output drops linearly from 0.75 megacycles to 1.25 megacycles.

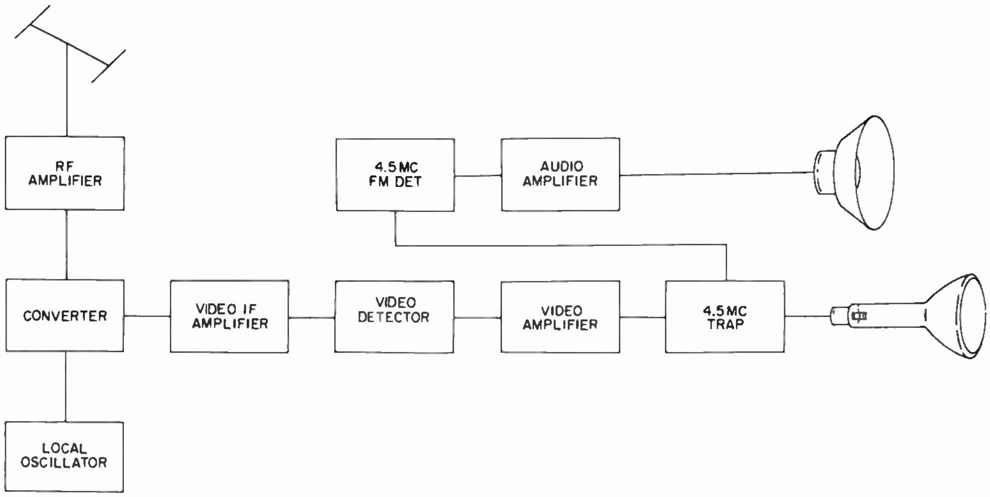


Fig. 13-2. Block diagram of an intercarrier-sound television receiver.

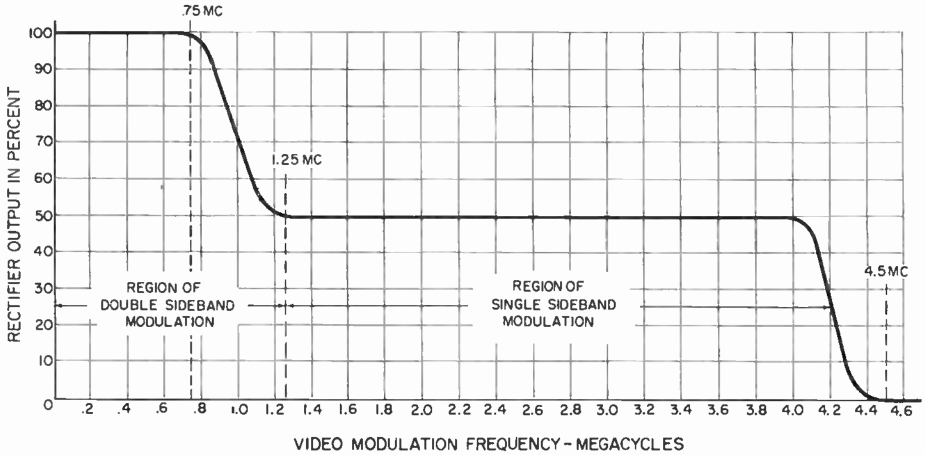


Fig. 13-3. Output of an ideal (linear) detector rectifying the vestigial-sideband modulated video carrier.

To compensate for this increased low-frequency output due to the lower sideband, the over-all response curve of the receiver should follow the linear slope of curve B in Fig. 13-4. This curve passes the 50% response point at the video-carrier frequency. In actual television receivers, the IF curve is usually shaped like C in Fig. 13-4.

As long as the area under curve B to the left of line X-X is equal to the area above curve B to the right of line X-X, these increments of the video-

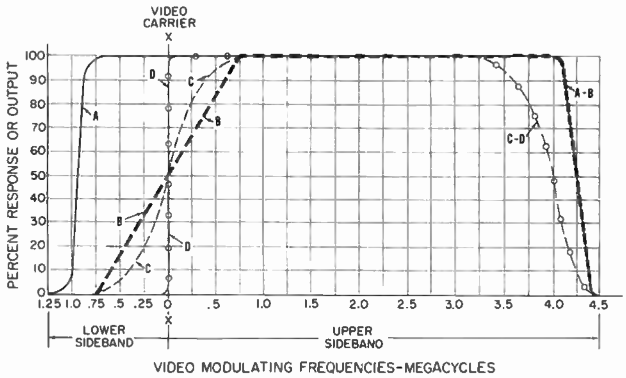


Fig. 13-4. Over-all receiver characteristic required to compensate for vestigial-sideband modulation.

detector output will add and will produce the desired curve D. The ideal response characteristic for the high-frequency end of the band is a sharp cutoff at 4 megacycles (curves A and B in Fig. 13-4). Although some receivers do approach this type of curve, most designs provide a more sloping cutoff (C and D in Fig. 13-4).

Although the reduction of high frequencies causes some loss of fine detail in a test pattern, this loss is not noticeable when a moving scene is being televised. For economy, the passband can be reduced to 3 megacycles without noticeably degrading the picture.

Up to this point our discussion of the video and sound carriers and of the adjacent-channel frequency has been concerned with their position in the spectrum as transmitted RF signals. Since the local oscillator is at a frequency higher than that of the signal, all these frequencies will be inverted in order when they appear as intermediate frequencies in the detector output.

Fig. 13-5 shows the over-all video IF response curve for a typical modern receiver. This curve

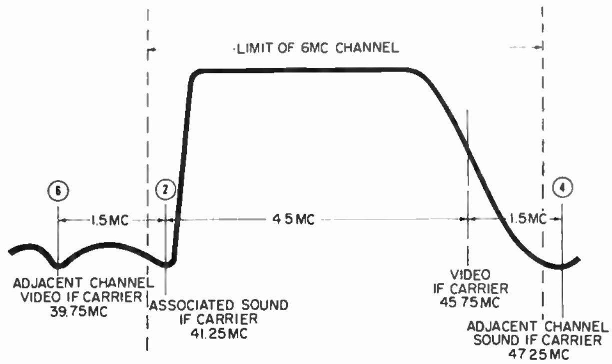


Fig. 13-5. Over-all video-IF response curve, showing location of IF trap frequencies.

shows the significant intermediate frequencies of the desired channel and the position of adjacent-channel IF carriers which might interfere with the operation of the receiver. The curve is depressed at these points; this reduction in response is done with trap circuits.

The choice of an intermediate frequency is limited by the television system and the frequency assignments. The intermediate frequency chosen must be between the highest video-frequency component on the one side and the lowest television frequency-channel assignment on the other. In other words, the IF cannot be lower than approximately 8 megacycles or higher than 50 megacycles. Here is a list of the video and sound IF carriers that have been employed in commercial television receivers:

Video IF Carrier (megacycles)	*Sound IF Carrier (megacycles)*
15.2	10.7
22.9	27.4
25.75	21.25
26.1	21.6
26.2	21.7
26.4	21.9
26.25	21.75
26.6	22.1
26.75	22.25
37.3	32.8
45.75	41.25

Methods of Obtaining Wide-Band Response Required of Video IF Amplifiers

Many types of coupling networks can be used between the video-IF amplifier stages to accomplish the wide-band response required. Two major methods are currently found in receiver design. These are:

1. Overcoupled transformers with shunt resistance loading.
2. Staggered-tuned circuits.

Video-IF Amplifiers with Overcoupled Transformers—If an interstage transformer whose primary and secondary are tuned to the same frequency has the coupling between the circuits progressively increased, a double-humped resonance curve will occur after the critical coupling point has been passed. These double humps can be smoothed to a flat-topped response curve by loading the primary and secondary circuits with the correct resistance. This, of course, results in lower Q and consequently lower gain than would be obtained with a narrower passband. Fig. 13-6 shows an intermediate-frequency amplifier system using overcoupled transformers.

The amplifier in Fig. 13-6 uses three bandpass-coupled stages employing 6CB6 miniature tubes. Transformers L2, L3, and L4 are overcoupled and are then loaded with resistance in the secondary circuits to produce the flat-topped bandpass characteristic. No primary shunt resistance is shown because the plate-to-cathode resistance of the tubes provides the required primary loading.

Decoupling, or isolation, of each stage from all others except the signal path is important in television IF amplifiers. If the second amplifier tube V2 in Fig. 13-6 were to receive the signal in the output of tube V3, the entire IF strip would oscillate. The feedback signal could be returned through the B+ supply line, through the filament supply line, or through capacitive coupling between components in different stages. No signal amplification is possible when the amplifiers are oscillating.

The B+ supply line in Fig. 13-6 is decoupled, therefore, at each stage in the amplifier. A 1,000-ohm series resistor and a bypass capacitor isolate the stages. The filament supply line is decoupled by a bypass capacitor at each tube socket. Many receivers also include RF chokes between each tube filament in the amplifier (Fig. 13-8).

Video IF Amplifiers with Staggered-Tuned Circuits—Another solution for designing a video amplifier to meet wide-band requirements is to couple the tubes by single-tuned circuits tuned to different frequencies within the desired passband.

If a number of single-tuned circuits (between amplifier tubes) are all tuned to the same frequency, the bandwidth decreases as stages are added, and the over-all response grows more peaked or selective. However, if the individual circuits are staggered-tuned about the center frequency and the Q values of the circuits are properly adjusted, the desired bandwidth can be obtained and a satisfactory over-all curve will result.

Curves in Figs. 13-7B and 13-7C illustrate the effects of the tuning frequency on the bandwidth of

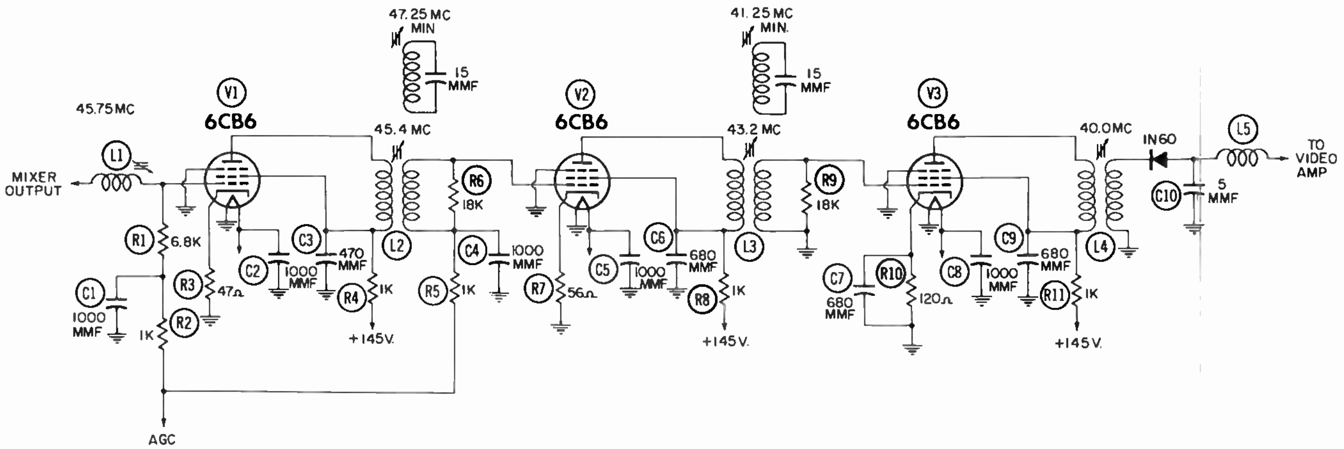
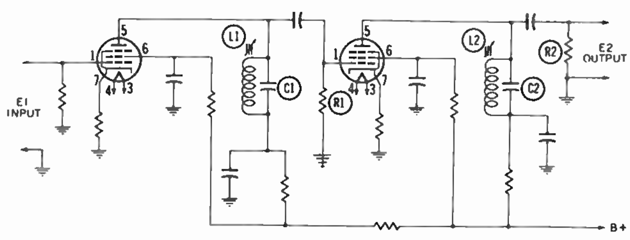
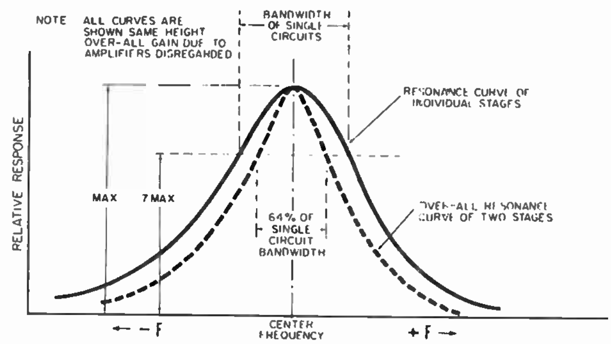


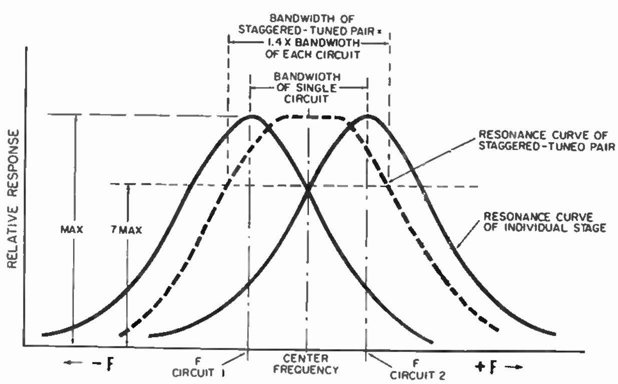
Fig. 13-6. Video IF amplifier employing overcoupled transformers.



(A) Single-tuned coupling circuits between video IF stages.



(B) Effect of cascading circuits on the same frequency.



(C) Effect of staggered tuning.

Fig. 13-7. The effect of tuned circuits on bandwidth.

a pair of identical circuits (Fig. 13-7A), first to the same frequency and then to frequencies separated from the center frequency by half the bandwidth of the individual circuits. The bandwidth of a circuit is defined as the frequency spread between the points on the resonance curve at which the response is 0.7 of that at resonance. When both circuits are tuned to the same frequency (Fig. 13.7B), the over-all bandwidth of the amplifier (Fig. 13-7A) will be 64 per cent of the bandwidth of the individual circuits.

Although we have used two circuits to explain the effects of staggered-tuning, the practical television video-IF amplifier involves the staggering of more circuits to achieve the required 4- to 6-megacycle bandwidth. In the two circuits just illustrated, the individual bandwidths and circuit Q values have been made identical for simplicity. In television receiver circuits, however, the Q values are varied to produce resonance curves and stage gains which will, in turn, cascade to produce approximately the over-all curve in Fig. 13-5. Figs. 13-8 and 13-9 illustrate the details of a typical staggered-tuned video IF amplifier system.

Fig. 13-9 shows the individual resonance curves of the various coupling circuits. Four stages of video-IF amplification are required to obtain sufficient IF gain and the necesesary wide-band characteristic. The effective Q of each circuit is determined by either the shunt plate resistance or a grid circuit resistance. The response curves produce the desired over-all response shown by the broken line in Fig. 13-9.

In a staggered-tuned system, variations of the individual amplifier stage gains do not affect the shape of the response curve; thus, the amplifier is not unduly sensitive to tube interchange. Another important feature of staggered tuning is the fact that alignment can be accomplished by a signal generator set to each of the individual circuit frequencies and with a VTVM at the video detector for output indication.

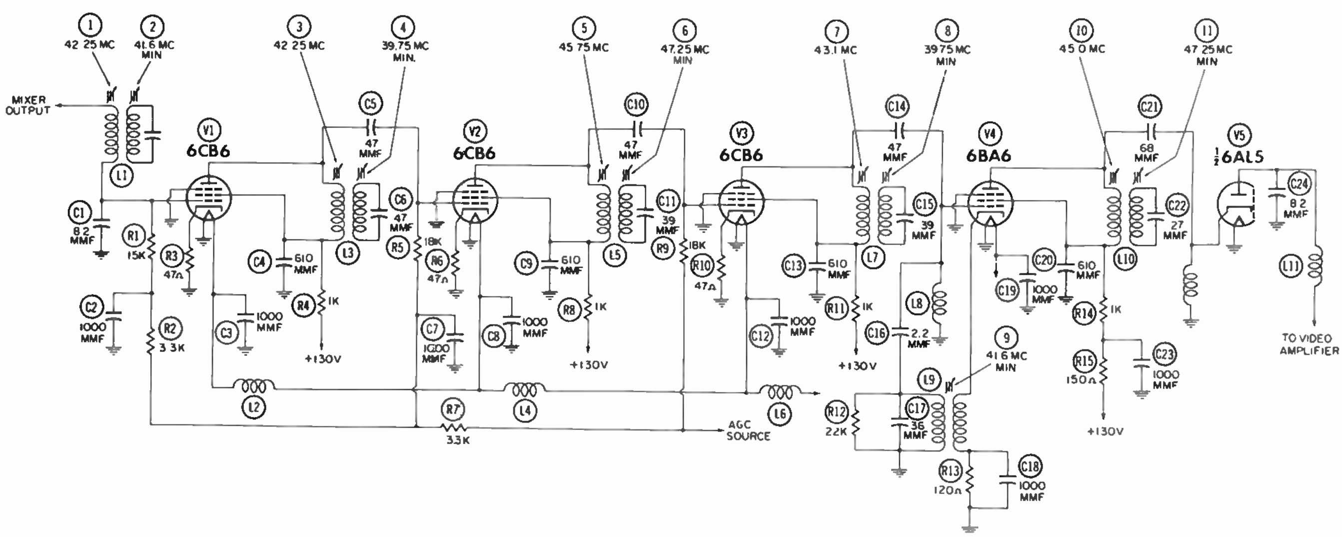


Fig. 13-8. An example of a staggered-tuned video-IF system.

A sweep generator and an oscilloscope are required for observation of the over-all curve only.

Rejection of Undesired Adjacent-Channel and Co-channel Carriers

In our discussion of the required response curve of the video IF amplifier (Fig. 13-5), we saw that the response must be reduced at the position which would be occupied by the sound carrier of the same channel being received, the video carrier of the next higher adjacent channel, and the sound carrier of the next lower adjacent channel. The low-response points or "notches" at the intermediate frequencies corresponding to these carriers can be produced in the video-IF amplifiers by one or a combination of the following circuits:

1. Series-tuned circuit traps.
2. Parallel-tuned circuit absorption traps.
3. Cathode-circuit or degenerative traps.
4. Bridged-T networks as coupling elements and rejection circuits.

The trap circuits (items 1 and 2) function like the interference-elimination traps used in radio to eliminate image interference or interference by strong local stations. The degenerative trap principle is often used in audio circuits for compensation. The fourth, or bridge-T type of trap is like the null bridge used for measurement purposes.

Series-Tuned Traps—The acceptance-type series-tuned trap acts as a short circuit across the system at its resonant frequency. The impedance at this frequency is very low and equals the AC resistance of the circuit components. In the amplifier in Fig. 13-6, the circuit comprised of coil L1 and the input capacitance of V1 is used to adjust the slope of the high-frequency end of the IF bandpass curve. The point at which the video carrier drops to 50% of maximum response (for compensation of vestigial-sideband effect) can be controlled. When this adjustment is made, the curve drops fast enough to reject the next lower-channel sound carrier, which falls at 47.25 megacycles.

Shunt-Tuned or Absorption-Type Traps—The absorption-type trap is the most frequently used rejection circuit. Six traps of this type are shown in the schematic in Fig. 13-8 (circuits 2, 4, 6, 8, 9, and 11). They consist of parallel-tuned circuits coupled to the plate or cathode coils of the video IF stages. At the resonant frequency, a high circulating current is developed in the trap. Because the trap reacts through its inducitve coupling, the load impedance of the stage and the amplification at the tuning point are reduced. The effects of the trap circuits in Fig. 13-8 on the amplifier response curve are shown in the correspondingly numbered points on the over-all amplifier curve in Fig. 13-9.

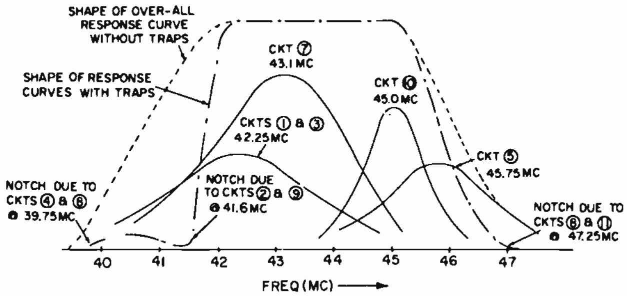


Fig. 13-9. The effect of cascading the staggered-tuned circuits of Fig. 13-8.

Cathode-Circuit or Degenerative Traps—Another method of reducing the amplifier gain at a particular point is by selective degeneration. This is done by placing a parallel-resonant circuit in the cathode-to-ground circuit of one of the amplifier stages. At the resonant frequency of the trap, a high impedance appears in the circuit. The trap acts like a large unbypassed resistor; it causes degeneration and reduces the amplification of the stage to a very low value. The rejection figure of such a degenerative trap can never be greater than the gain of the stage.

Bridged-T Networks—Fig. 13-10 shows a network of circuit elements known as the bridged-T. These circuits use a T-shaped branch consisting of two capacitors and a resistor (C1, C2, and R1) bridged by inductor L2 to reject the unwanted associated sound and adjacent-sound IF carriers.

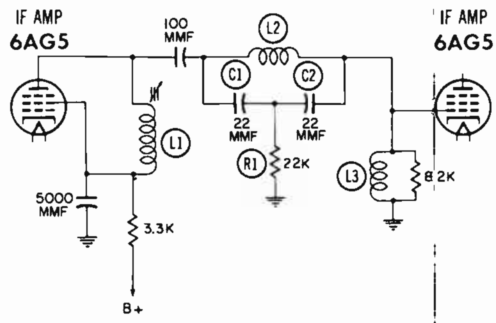
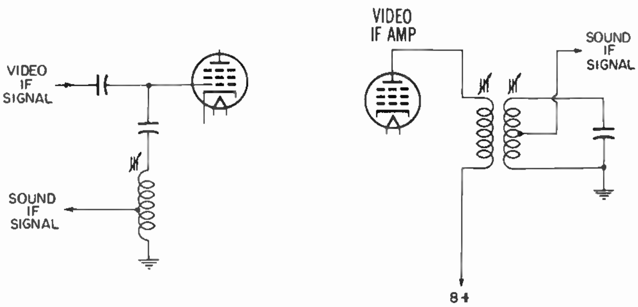


Fig. 13-10. The bridged-T network used as a trap.

This arrangement acts like a bridge circuit, but has the advantage of the input and output circuits having a common terminal at ground potential. Balance (for a null or low output) occurs when the reactance of the variable inductance equals the reactance of the capacitors in series and when the resistor in the center leg of the T is approximately one-fourth the parallel resistance of the tuned circuit. The circuit also acts as an anti-resonant trap in the line with a secondary balance for the resistance losses of the circuit. Much sharper null notches and greater rejection can be obtained with the bridge-T connection than with the trap circuit alone. Another variation of the bridged-T circuit (not illustrated) employs a center-tapped coil for the resistance branch and a single trimmer for the capacitance branch.

The intercarrier sound system is used in almost all the receivers on the market. Those IF circuits just discussed illustrate only intercarrier receivers. In nonintercarrier, or separate-sound, receivers, the take-off point for the sound IF signal can follow the converter or the video IF amplifier. Usually, this take-off point will be a tap on a trap coil (Fig. 13-11). This method allows a co-channel sound trap to serve a dual purpose.



(A) Series-tuned trap. (B) Parallel-tuned trap.

Fig. 13-11. Take-off points for sound-IF signal in nonintercarrier receivers.

Reflex Amplifier

A new use of an old circuit is illustrated in Fig. 13-12. In the early days of radio, a circuit which allowed one- and two-tube receivers to have more than normal gain was developed. A high-frequency amplifier was used to amplify a low-frequency signal at the same time. Thus, an RF amplifier could serve as an audio amplifier without either signal being adversely affected. This type of circuit was called the reflex amplifier.

The reflex circuit has been revived for television receivers. In modern intercarrier receivers, the video IF signal is at 45.75 mc and the sound IF signal is at 4.5 mc. These two frequencies are separated enough that they can both be amplified by the same tube. This dual amplification is accomplished by the circuit in Fig. 13-12.

The signal at the grid of V1 is the video-IF signal from the first video-IF amplifier. This signal is amplified by V1 and detected by M1. The output of M1 is the video modulation between 30 cycles and 4 megacycles and the sound IF signal at 4.5 megacycles. This 4.5-mc signal is coupled back to the grid of V1 through C7, L5, and C1. Coil L5 is adjustable and tunes the feedback path to 4.5 mc.

The 4.5-mc signal is amplified by V1 and appears in the plate circuit. The primary of L2 (like the secondary of L1) is tuned to the video-IF frequency and presents almost no impedance to the 4.5-mc signal. Coil L3, however, is tuned to 4.5 mc; and the 4.5-mc output signal of V1 is developed across L3. The 4.5-mc signal is coupled through C5 to the remainder of the sound IF system.

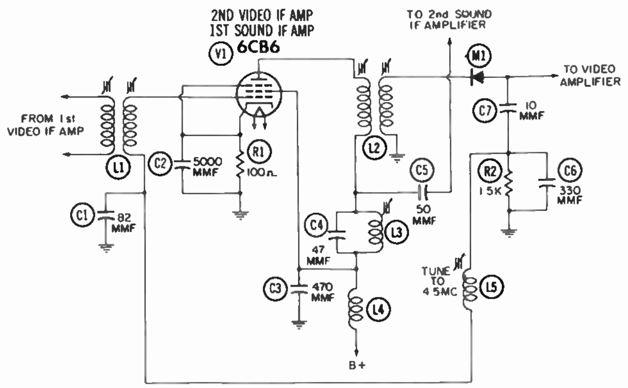


Fig. 13-12. Circuit of the reflex IF amplifier.

Cascode Amplifier

The two video stages in Fig. 13-13 are similar to the cascode RF stage discussed in Chapter 12. The one important difference is that the second stage is not a grounded-grid stage. As far as the B+ supply is concerned, the two tubes, just like the cascode tubes, are in series. Since the cathode of V2 is above ground by approximately one-half the supply voltage, a voltage divider (R6 and R7) biases the grid of V2 to the approximate cathode potential.

VIDEO DETECTORS

The video-IF amplifier is followed by the video detector, which is essentially the same as the second detector in AM broadcast or short-wave radio receivers. However, two significant circuit differences

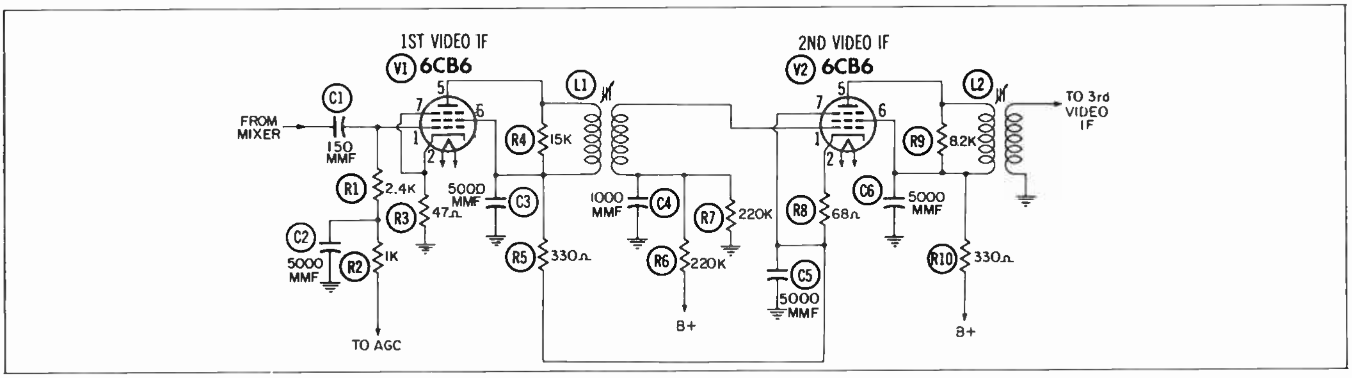
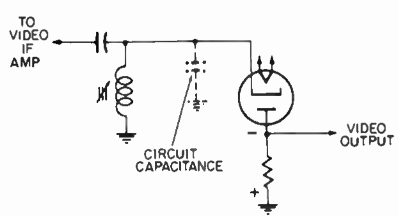


Fig. 13-13. Video-IF amplifier with two tubes in series.

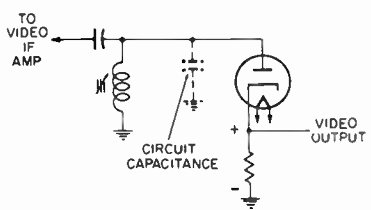
in the TV video detector must be taken into consideration: (1) a means of compensation must be used to prevent the loss of the higher video frequencies, and (2) the polarity of the detector output must be considered.

The video signal may be applied to the grid or to the cathode of the picture tube. To what element the signal is applied, and the number of amplifiers between the detector and the picture tube determine the polarity of the detector-output signal.

If there are an even number of video amplifying stages between the detector and the picture-tube grid, the detector output must be negative-going. In other words, an increase in IF carrier strength at the detector results in a more negative video signal with respect to ground. Fig. 13-14A shows a detector which supplies a negative picture polarity.



(A) Video output increases negatively with respect to ground as carrier increases.



(B) Video output increases positively with respect to ground as carrier increases.

Fig. 13-14. Diode video-detector output polarity.

If an odd number of video amplifying stages are employed (in most instances, this will be a single stage) and the video signal is applied to the grid of the picture tube, the detector must be connected as shown in Fig. 13-14B. This circuit, with the plate of the diode connected to the high side of the video coupling circuit, produces a video output which becomes more positive as the video carrier strength is increased.

A typical video detection and amplifying system employing a video signal of negative polarity from the diode is shown in Fig. 13-15. Load resistor R2 has a value of 3,900 ohms and has associated with it a group of circuit elements, C3, L3, L4, and R1. These elements assist in producing a flat video response from 30 cycles to more than 4 megacycles.

A significant difference between a radio detector and a video detector is in the value of the load resistor. In the second detector of a radio, a typical diode load varies from 0.5 to 2 megohms and maintains this high load resistance over the range of frequencies required for sound reproduction. In most instances, no frequencies higher than 5,000 cycles are involved. In the video detector, however, the capacitance of the diode and the capacitance to ground of its associated circuit prevent the use of a high diode load resistance because flat response to at least 4 megacycles must be provided. At these high frequencies, the circuit reactance and tube capacitance would become lower than the load resistance and thus bypass the high frequencies. For this reason, the load resistance is made the low value; and the compensating elements just discussed produce a resonant rise of circuit impedance at the high end of the video band. L3 is a series-peaking choke and L4 is a shunt-peaking choke. These compensating elements are employed in each video stage. (See L5-L6 and L7-L8 in Fig. 13-15.) A more complete discussion of their function will be presented when video amplification is considered.

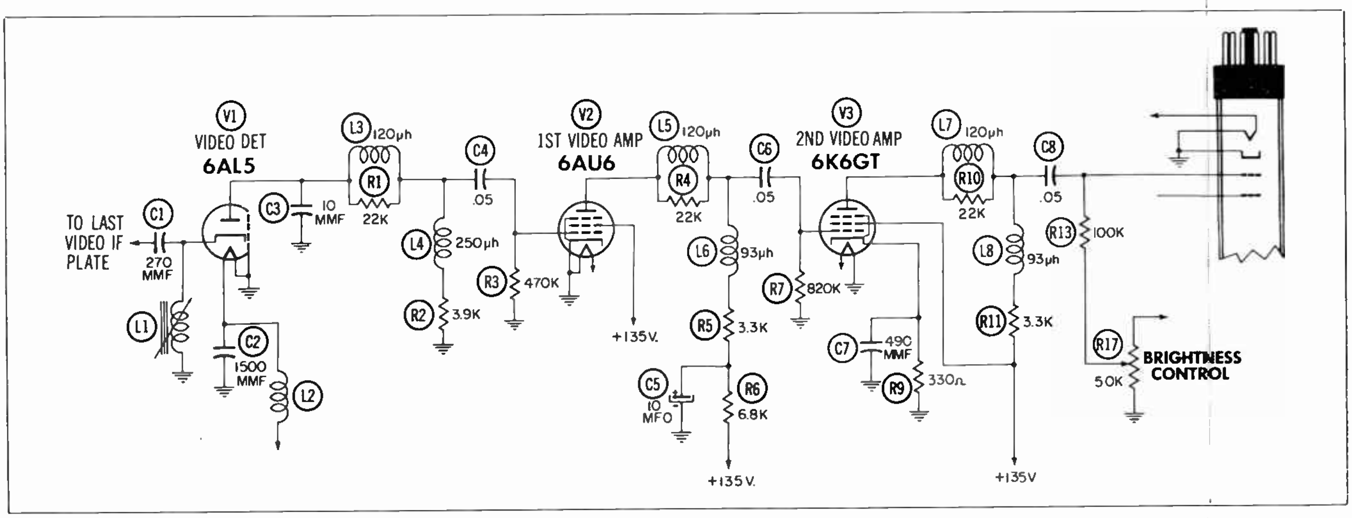


Fig. 13-15. Video detector of negative polarity feeding picture-tube grid through two video-amplifier stages.

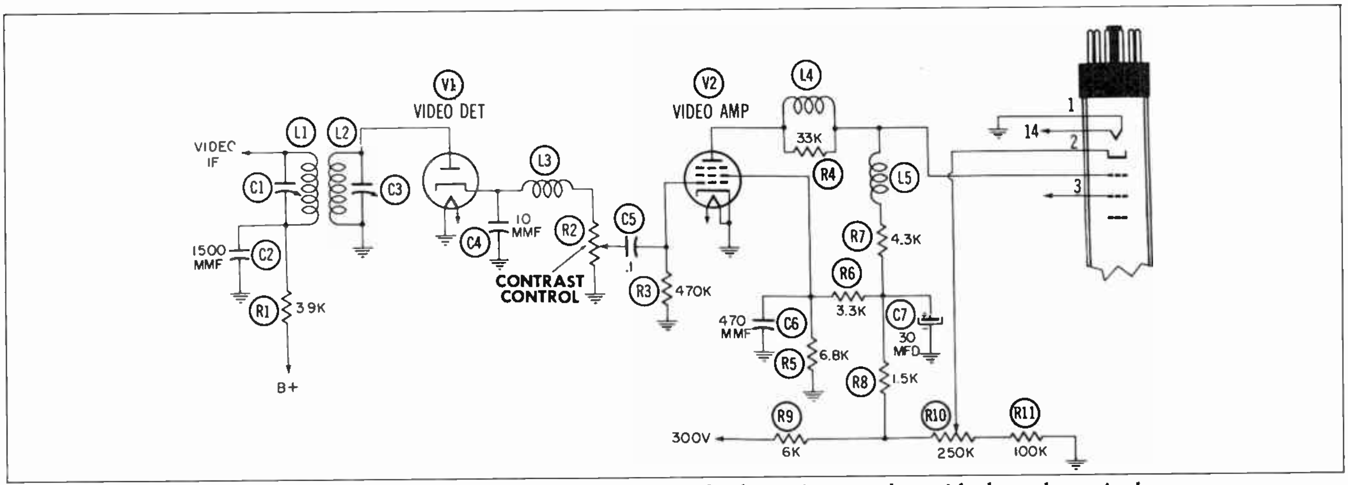


Fig. 13-16. Video detector of positive polarity feeding picture-tube grid through a single video-amplifier stage.

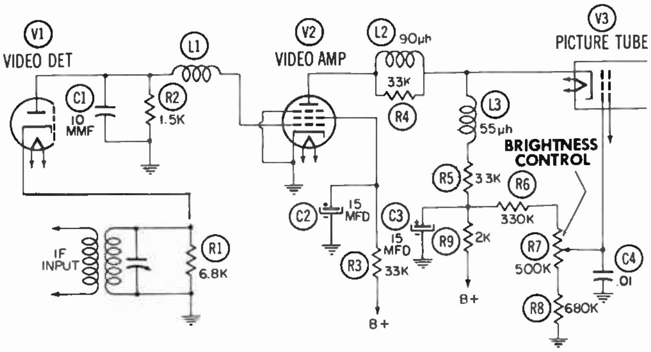


Fig. 13-17. A negative-output detector feeding the cathode of the picture tube through a single DC amplifier.

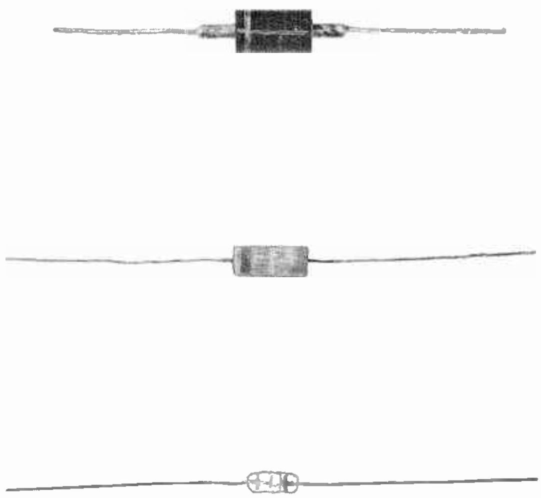
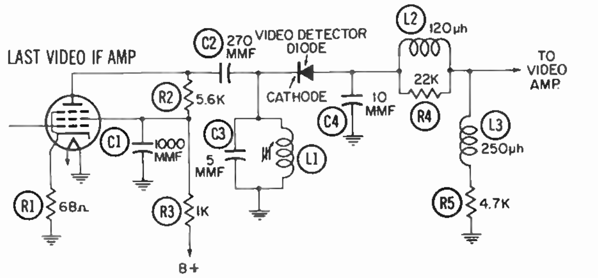


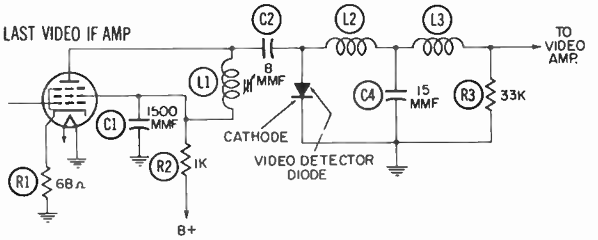
Fig. 13-18. Three types of fixed crystal detectors. (Shown actual size.)

Fig. 13-16 shows a video detection and amplification system employing the diode with its plate connected to the high side of the IF input circuit in order to produce a positive-going video signal. The polarity of the signal is inverted once by the single video-amplifier stage; thus, a more negative voltage is produced at the picture-tube control grid as the carrier strength increases. A combination of L3 and C4, together with the circuit capacitances, resonates at the high end of the video band aind maintains a flat response from the detector.

In Figs. 13-15 and 13-16, the video-output tube is coupled to the control grid of the picture tube. For circuit simplification in some receivers, the picture-tube input is inverted, and the video-output tube is coupled to the cathode rather than to the control grid. A positive-going rather than a negative-going signal must be fed to the picture-tube cathode. Fig. 13-17 shows an example of a circuit employing a negative output detector V1 with a single-stage amplifier V2 feeding the cathode of picture tube V3.



(A) Crystal diode with series load.



(B) Crystal diode with shunt load.

Fig. 13-19. The semiconductor crystal as a video detector.

The Semiconductor Crystal Diode As a Video Detector

Small fixed crystal rectifiers (Fig. 13-18) are being used extensively as a substitute for the thermionic diode. Fig. 13-19 shows two typical applications of crystal detectors used as video demodulators. Fig. 13-19A shows a load circuit consisting of resistor R5 and high frequency-compensation network C4, L2, L3, and R4 connected in series with the diode. Fig. 13-19B shows a shunt conection of similar elements.

The crystal diode is recommended over the vacuum-tube type for several reasons. These are: (1) lower dynamic resistance, (2) reduction of power consumption (no heater required), and (3) ease of mounting (no socket required). The lower capacitance and lower dynamic resistance of the crystal diode improve its performance as a video detector by providing higher rectification efficiency for a given bandwidth. Its improved linearity at low signal levels helps preserve highlights in the picture.

QUESTIONS

1. When the high frequencies of the television signal are lost, what happens in the reproduced picture?
2. What two coupling methods are used to obtain the wide-band response required from the video IF amplifiers?
3. What results when the primary and secondary of a coupling transformer are loaded with a shunt resistance?
4. What is meant by staggered tuning?
5. Why are traps necessary in the video-IF circuits?
6. What is the purpose of the reflex amplifier?
7. When the video signal is applied to the grid of the picture tube and there are two video-amplifier stages, what must be the polarity of the video-detector output? In this case, is the signal taken off the plate or off the cathode of the video detector?
8. What other type of component is being used for video detectors in place of the vacuum-tube diode?

EXERCISES

1. Draw the over-all IF response curve and show:
   (a) The frequency limits.
   (b) The location of the video carrier frequency.
   (c) The location of the traps.

Chapter 14

Sound IF Amplifiers and Audio Detectors

The sound IF system of the television receiver is similar to the IF system of an FM radio receiver. Two major differences from the standard FM receiver will be noted:

1. The intermediate frequency for FM receivers has been standardized at 10.7 megacycles, whereas the sound IF for intercarrier television receivers has been standardized at 4.5 megacycles. (This is the frequency difference between the video and sound carriers and is held constant at the transmitter.) For nonintercarrier receivers, the sound intermediate frequency will always be 4.5 megacycles lower than the video intermediate frequency.

2. The deviation of the television sound carrier for maximum modulation has been established at 25 kc (a total sweep of 50 kc), whereas the maximum deviation for standard FM broadcasting has been set at 75 kc (a maximum sweep of 150 kc). Because a lower deviation is employed for television sound, a narrower passband can be used for the television sound IF system, and a shorter linear region can be used for the detector. The passband of a typical intercarrier sound IF amplifier is about 150 kc, and the linear region of the detector usually does not exceed 100 kc. A high-quality FM receiver might have an IF passband of 300 kc and a linear range in the detector of as much as 2 megacycles.

A nonintercarrier receiver requires a passband and a linearity region of several hundred kilocycles to render the receiver immune to drifting of the local-oscillator frequency. An extremely small percentage of frequency drift of the local oscillator would be sufficient to move the audio IF out of the passband if it were made only wide enough to accommodate the 50-kc deviation. This small drift would not noticeably affect the picture because of the vestigial-sideband nature of the video signal.

SOUND IF TAKE-OFF

The take-off methods for nonintercarrier receivers were discussed in Chapter 13. Take-off methods for intercarrier sound systems are different because the take-off point must follow the video detector. A sound detector requires a definite amount of signal to provide a noise-free signal. The amplification this signal receives is provided by the sound IF amplifier, the video amplifier, or combinations of both.

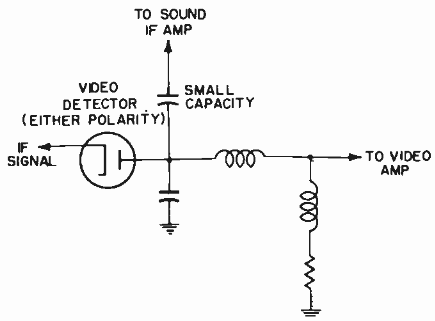


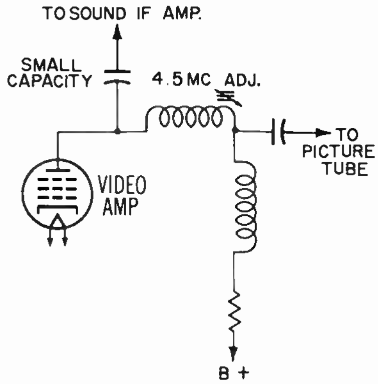
Fig. 14-1. Sound IF take-off point in video detector.

The take-off point for most modern receivers is at the video-detector output (Fig. 14-1). The 4.5-mc signal is coupled through a small capacitor from the video-detector output to the sound-IF amplifier input. In other receivers, sound-IF take-off point is in the plate circuit of the video amplifier to take advantage of the gain supplied by the video amplifier. Typical circuits are shown in Fig. 14-2.

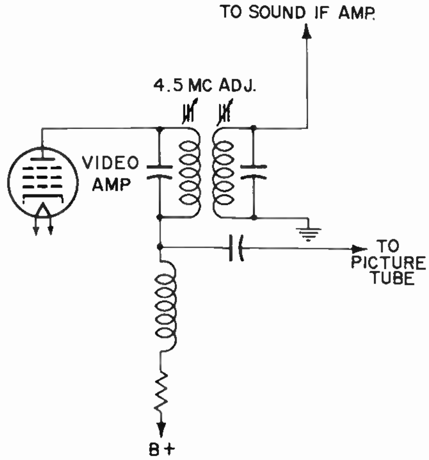
TYPICAL SOUND IF SYSTEMS

A sound IF circuit that is typical of those used in intercarrier receivers is shown in Fig. 14-3. The 4.5-mc signal from the video detector is tuned by the grid coil and then amplified by the pentode tube. The output is applied through the IF transformer to the audio detector, which can be any of the commonly used detectors except a discriminator. A discriminator requires more amplification stages and at least one limiting stage.

A second type of sound IF amplifier is shown in Fig. 14-4. In this circuit, the triode portion of a triode-pentode tube is used as the sound IF ampli-



(A) Capacitor coupling.



(B) Transformer coupling.

Fig. 14-2. Sound IF take-off points in video amplifier.

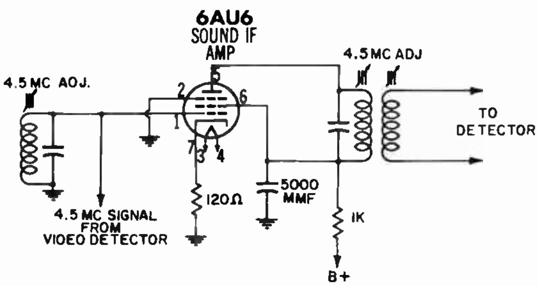


Fig. 14-3. A typical sound-IF amplifier circuit.

fier. Since this tube is a triode operating at a relatively high frequency, it must be neutralized. Neutralization is accomplished by the tapped plate coil and the 4.5-mmf capacitor.

The reflex amplifier discussed in Chapter 13 should be considered a sound IF amplifier. Another type of reflex amplifier is shown in Fig. 14-5. In this circuit, the sound output from the audio detector is applied to the input of the sound IF amplifier, so that the tube can amplify both signals simultaneously. The plate load for the 4.5-mc signal is the transformer, and the plate load for the sound signal is the 27K resistor.

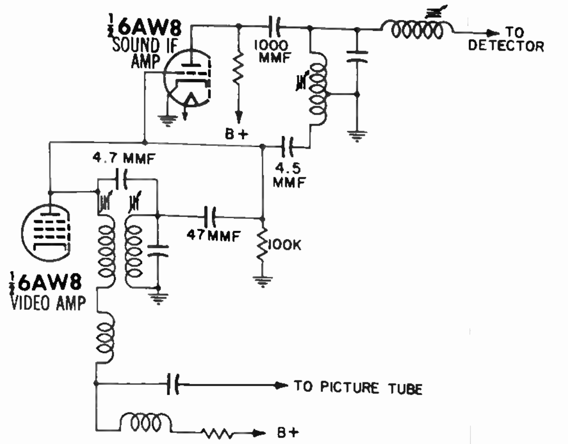


Fig. 14-4. A neutralized triode used as a sound IF amplifier.

AUDIO DETECTORS

Television receivers employ at least five types of FM demodulation circuits: the Foster-Seeley discriminator, the ratio detector, the gated-beam detector, the locked-oscillator detector, and the Delta sound system.

The Foster-Seeley FM Discriminator

The Foster-Seeley or phase-shift discriminator (Fig. 14-6) is familiar to the television service technician, not only as an FM detector, but also as a method of obtaining a control voltage for AFC systems. The discriminator provides an audio output which, at any instant, is equal to the difference between the rectified output of diode V2 (across R4) and diode V3 (across R5). These two resistors and RF bypass capacitors C7 and C8 are equal in value.

In this circuit, the detectors interpret a change in frequency as a change in amplitude because of the method by which the signal is coupled to the plates of the diodes. A phase-shift principle is used, and the output voltage is obtained in the following manner: The transformer, consisting of primary L1 and secondary L2, couples the output of the limiter to the diode plates (V2 and V3). One end of the secondary is connected to the plate of V2, and the opposite end is connected to the plate of V3. Both windings of the transformer are tuned to the center frequency by capacitors C3 and C6.

The primary voltage is also coupled to the center tap of the secondary by capacitor C5. Coil L3 provides a DC return for the diode plate current. The reactance of C5 is negligible at the signal frequency. When the secondary is resonant, as it is here, the primary voltage at the center tap of the secondary is in quadrature (90°) with the voltage across each half of the secondary. This 90° phase relationship varies as the signal frequency changes. The difference in phase between the primary voltage at the center tap and the voltage across each half of the secondary determines the audio output. Let us see

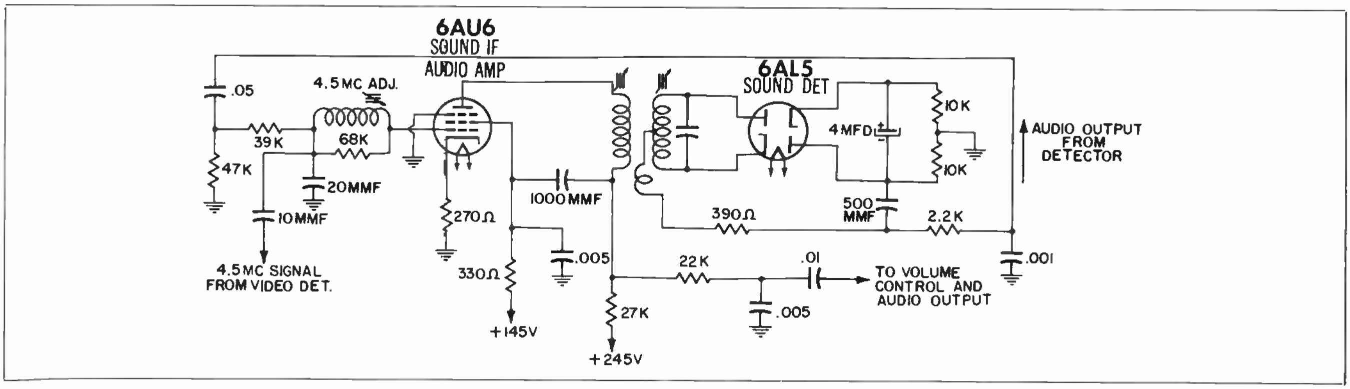


Fig. 14-5. A reflex amplifier which handles the sound IF and audio signals.

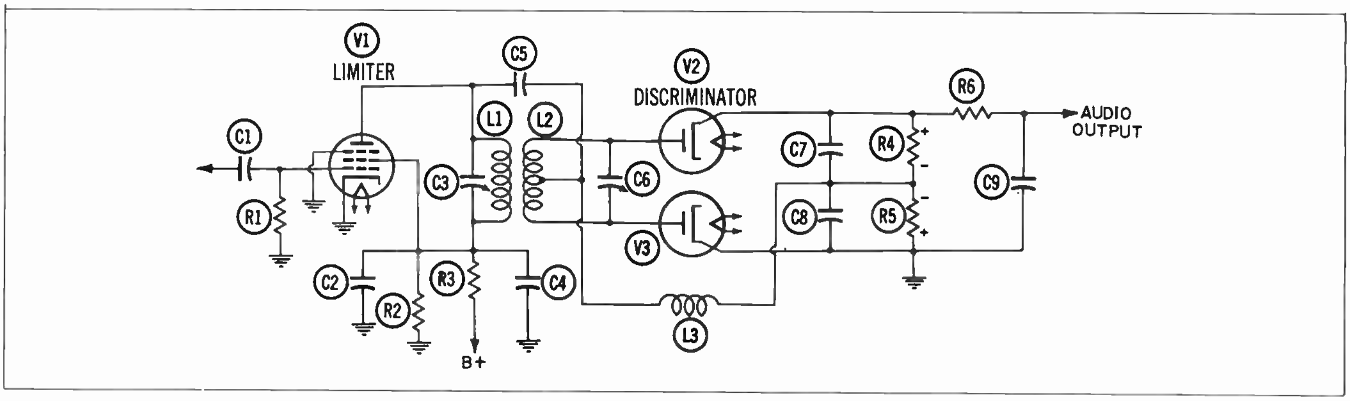


Fig. 14-6. Foster-Seeley discriminator circuit.

what takes place during different operating conditions.

When a frequency equal to the center frequency appears at the limiter output, the diodes conduct equally. During this time, the voltage across the top half of the secondary is 90° out of phase with the primary voltage at the center tap of the secondary. The resultant of these two voltages is applied to the plate of V2. At the same time, the voltage across the bottom half of the secondary is also 90° out of phase with the primary voltage at the center tap. The resultant of these two voltages is applied to the plate of V3. Because the phase angles are equal, the voltages applied to the diode plates are equal; therefore, the diodes conduct equally. During this equal conduction, the voltages across load resistors R4 and R5 are equal, but opposite in polarity. Therefore, the voltages cancel and the output is zero.

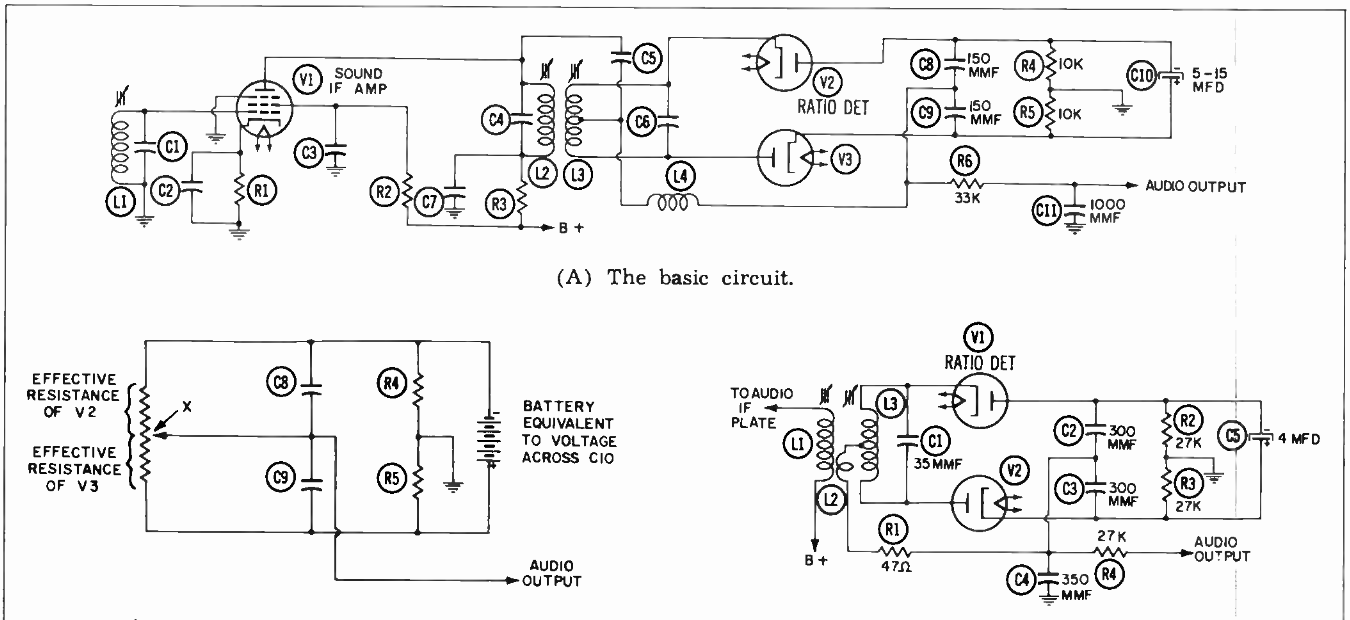
When the signal frequency deviates from center frequency, this deviation is reflected in the output circuit. The following happens when the input frequency is higher than the center frequency. The phase difference between the primary voltage and the voltage across the top half of the secondary is less than 90°. At the same time, the phase difference between the primary voltage and the bottom half of the secondary becomes more than 90°. As a result, the voltage on the plate of V2 is higher than the voltage on the plate of V3. Diode V2 then conducts more than diode V3 does. The current through R4 increases, and a larger positive voltage is developed. The current through R5 decreases, and the negative potential is reduced. The difference of potential across R4 and R5 results in a positive voltage in the output.

When the input frequency is lower than the center frequency just the opposite happens. The plate voltage on V2 becomes less than the plate voltage on V3. V3 then conducts more than V2. The voltage becomes more negative across R5 and less positive across R4. This change in potential across the load resistors results in a negative voltage in the output.

In summary, the IF signal varies above and below the center frequency. The phase angle of the secondary voltage, with respect to the primary voltage, varies above and below 90°. Because of this changing phase angle, the inductively- and capacitively-coupled voltages combine and provide the balanced detectors with a signal voltage that varies in amplitude as the IF signal varies in frequency. The signal is detected, and appears as an audio signal in the output.

This type of FM detector will discriminate against amplitude change at its zero-modulation frequency only. For this reason, a limiting stage must precede the detector.

For a better signal-to-noise ratio, the high-frequency portion of the audio signal is pre-emphasized



(A) The basic circuit.

(B) The equivalent circuit of the ratio detector.

(C) The most common variation of the basic circuit.

Fig. 14-7. Ratio-detector circuits.

in the modulation of the television transmitter. A de-emphasis circuit consisting of a series resistor and a shunt capacitor (R6 and C9 in Fig. 14-6) corrects this pre-emphasis in the receiver. The time constant of this combination is fixed at 75 microseconds, which conforms to the FCC standard.

Ratio Detectors

The ratio detector is a popular type of FM detector in television receivers. Besides the function of detection, it features an inherent rejection of amplitude modulation.

Fig. 14-7A shows a ratio-detector circuit. The input transformer connections are identical to those in the basic Foster-Seeley discriminator in Fig. 14-6. The resultant voltages due to the changes in frequency are applied to the diodes. The resemblance stops at this point. The diodes are arranged in series or additive fashion, rather than in opposition to one another as they are in the discriminator. The output is taken from the tap on the secondary of the input transformer, instead of across load resistors R4 and R5. An electrolytic capacitor C10 is used across the load resistors.

We have seen that the output of a discriminator is proportional to the difference of the absolute values of the voltages impressed on the diodes. The ratio-detector output, however, is proportional to the ratio between the amplitudes of the impressed voltages. Since the ratio of the voltages instead of the difference is used, amplitude modulation of the incoming signal has little effect on the output.

The conduction of the diodes in series produces a stable charge across capacitor C10. It is convenient to think of the stabilizing capacitor C10 as being a form of bias battery across the two diodes in series. See Fig. 14-7B. The diodes may be considered as being variable resistances across this battery. Each resistance is determined by the amount of IF voltage impressed on that diode. With frequency modulation, the IF voltage impressed on one diode increases as the voltage impressed on the other decreases, and vice versa. Therefore, the variable resistances of the diodes have been represented by a potentiometer in Fig. 14-7B.

When an unmodulated carrier is received, equal voltages are impressed across the diodes. The diodes conduct equally, and their effective resistances are equal. The potential at their junction X in Fig. 14-7B does not change. Consequently, there is no audio output.

When the frequency of the incoming signal increases because of modulation, the voltage impressed on diode V2 becomes higher than the voltage impressed on V3; and the effective resistance of V2 decreases while the effective resistance of V3 increases. This has the same effect as if the arm of the potentiometer in Fig. 14-7B were moved upward. The output voltage swings in a negative direction.

When the frequency of the incoming signal decreases, the arm of the potentiometer is effectively moved downward and the output voltage swings in a positive direction. In this way, FM detection takes place.

When the incoming signal is amplitude-modulated, the voltages on the diodes change in the same direction. This is equivalent to an identical change in the effective resistances of *both* halves of the potentiometer in Fig. 14-7B; the arm does not move, and there is no output voltage.

Fig. 14-7C shows the most common variation of the basic type of ratio detector. The major difference is that a tertiary winding L2 instead of a capacitor supplies the quadrature voltage. Winding L2 is closely coupled to the primary winding; therefore, the voltage phases on the two windings are substantially the same. The phase relationship of the coupled voltage to the voltage of the secondary is still 90° at the center frequency.

The Gated-Beam Detector

The circuit in Fig. 14-8 presents a unique method of FM detection. A special tube, the 6BN6, functions

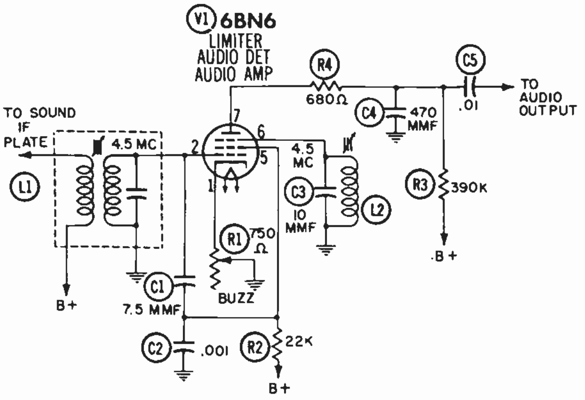


Fig. 14-8. The gated-beam 6BN6 detector.

as a limiter, audio detector, and 1st audio amplifier. The construction of the 6BN6 is shown in Fig. 14-9. Its cathode provides a thin beam of electrons that are guided and accelerated toward the plate. In the path of the electrons, a limiter grid and a quadrature grid act as gates: either grid can cut off the electrons. Both grids have a sharp cutoff feature—a small signal voltage can drive the tube from cutoff to saturation. Amplitude modulation is rejected or limited because large signal variations have no greater effect on plate current than small signal variations.

As the plate current of the tube flows past the quadrature grid, a voltage is induced on this grid by space-charge coupling. The 4.5-mc tuned circuit (L2

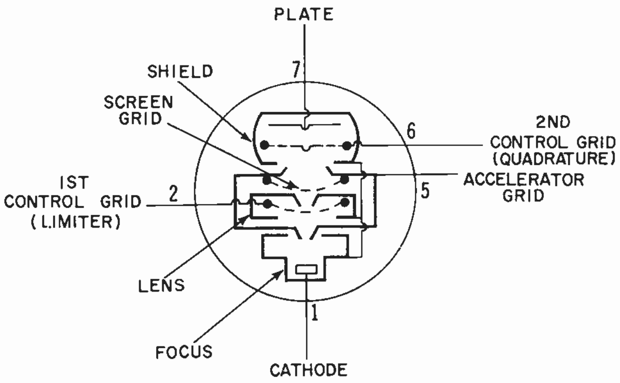


Fig. 14-9. Internal construction of the 6BN6 tube.

and C3) is connected to the quadrature grid. This circuit causes the alternating voltage on the quadrature grid to be 90° out of phase with the center frequency of the incoming IF signal (connected to the limiter grid) when the IF signal is unmodulated.

When modulation is impressed on the IF signal, the phase difference between the limiter and quadrature grids varies above and below 90°. When the input signal goes positive, the limiter grid allows current to pass. A short time later, depending upon the instantaneous phase difference, the quadrature grid goes positive and allows current to pass. When the input signal goes negative, the plate current stops. The average plate current will depend upon the length of time the plate current flows on each positive cycle. When the audio modulation varies, the phase difference, the conduction time, and the average plate current will all vary proportionately. The plate current variations then contain the same modulation as an amplified audio signal large enough in amplitude to drive the output stage directly.

The gated-beam detector usually has an adjustable resistor in the cathode circuit. The resistor so adjusts the tube bias that the incoming signal will drive the limiter grid into both saturation and cutoff. This resistor is called a buzz control because it eliminates the buzzing sound caused by sync signals getting through to the audio section.

The gated-beam detector becomes very inefficient on weak signals. When the incoming signal at the limiter grid drops below a certain voltage, limiting stops, and sync buzz and noise are reproduced.

The Locked-Oscillator Detector

A circuit which operates much like the gated-beam detector is shown in Fig. 14-10. The similarity between the gated-beam and the locked-oscillator circuits can be clearly seen by comparing Figs. 14-8 and 14-10. The most obvious difference is that the locked-oscillator detector lacks a buzz control.

The 6DT6 is not as complex as the 6BN6 and is much like an ordinary pentode except that the control and suppressor grids can both sharply cut off plate current. During reception of moderate or strong signals, quadrature-grid detection takes place much the same as it does in the 6BN6 circuit except for a few differences in limiting.

The locked-oscillator mode of operation of the 6DT6 circuit does not come into play until the input signal becomes fairly weak. Below a certain signal amplitude, the detector will break into 4.5-mc oscillation. This oscillation tends to keep the signal amplitude constant in the detector despite amplitude variations which may occur in the input signal because of noise or fading.

The circuit can oscillate because of positive feedback from the quadrature-grid circuit to the control-grid circuit through the interelectrode capacitance of the tube. A similar arrangement using a 6BN6 can-

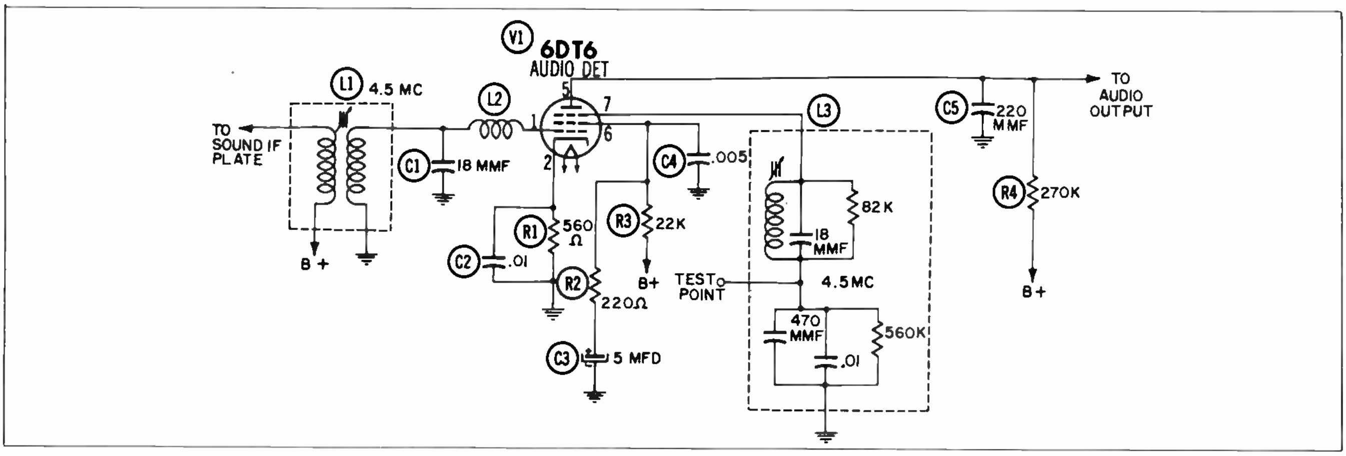


Fig. 14-10. The locked-oscillator 6DT6 detector.

not be set up because the interelectrode capacitance of this tube is too small to provide the required amount of feedback.

The input signal voltage to the 6DT6 can drop to as low as ⅓ volt rms at the secondary winding of the detector input transformer without a loss of oscillation. Normally, the signal generated in the control grid circuit by the oscillation is approximately one volt rms. The space-charge coupling to the quadrature grid is accompanied by a voltage gain. The oscillations in the quadrature tank have about three times the amplitude of the oscillations in the input circuit. This amplitude is sufficient to develop the required bias voltage across the 560K-ohm resistor in the quadrature circuit.

The locked feature of the 6DT6 oscillator refers to the fact that the phase of the oscillations in the control-grid circuit will follow the phase of the incoming sound-IF signal. During the locked-oscillator mode of operation, the input signal serves as not much more than a type of sync signal having little amplitude but yielding frequency information. Normal quadrature-grid detection takes place in the oscillating detector, and the process is kept under control by the input signal at all times. The oscillation boosts the weak-signal sensitivity of the quadrature-grid circuit, so that its performance becomes comparable to that of a ratio detector. Clear sound can be received even when the station signal is so weak, the picture is not fit to watch.

Limiting of strong signals is done somewhat differently in the 6DT6 than in the 6BN6. The characteristic curve of control-grid voltage of the 6DT6 does not show a rapid leveling off or saturation of plate current at small positive values of grid voltage as the corresponding curve for the 6BN6 does. Limiting in the 6DT6 depends on the damping of strong signals in the grid circuit. A strong input signal causes the control grid to draw considerable current, which loads down the tuned circuit connected to the grid. The oscillation is suppressed by the grid loading, the tuning of the input circuit is broadened, and the peak voltage swing at the grid is held to only a few volts.

Degeneration of audio-frequency signals in the cathode circuit of the 6DT6 also contributes to limiting. The value of cathode resistance required in the 6DT6 circuit for best AM rejection is not critical, and no control is needed in the cathode circuit.

The Delta Sound Detector

Fig. 14-11 shows the delta sound detector circuit, which uses a dual-diode/triode tube. The FM sound signal is demodulated in this circuit by a simple, well-known process called slope detection. This process has seldom been applied to commercial circuits because it cannot reject AM noise interference. The delta design makes the slope detector practical by coupling the detector with an efficient noise-suppression circuit.

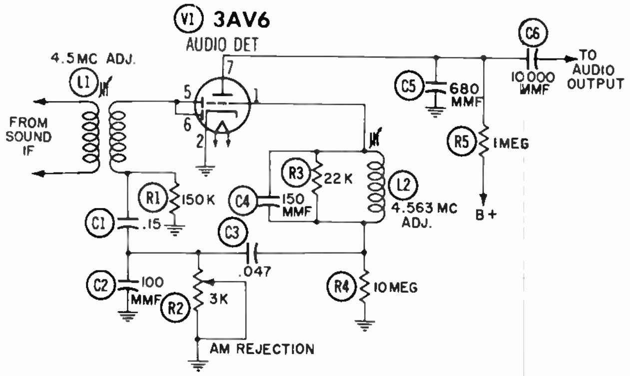


Fig. 14-11. The delta sound detector.

A frequency-modulated, 4.5-mc signal fed into the secondary of L1 from the sound-IF stage is developed across C2, which is a part of the tuned circuit associated with the secondary of L1. From C2, the signal is coupled through C3 to the tuned circuit composed of L2, C4, and R3. This tuned circuit, pulsed by the signal fed to it through C3, develops the actual driving signal for the grid of the triode. If the pulsing signal is negative with respect to ground at a certain instant, the voltage applied to the triode grid (connected to the other end of the tuned circuit) is positive at the same instant.

The circuit of L2, C4, and R3 is not tuned to exactly 4.5 mc as in many sound detectors, but to 4.563 mc. The result is shown in Fig. 14-12. The tuned-circuit response in other sound detectors falls off symmetrically on each side of 4.5 mc, whereas the entire band-

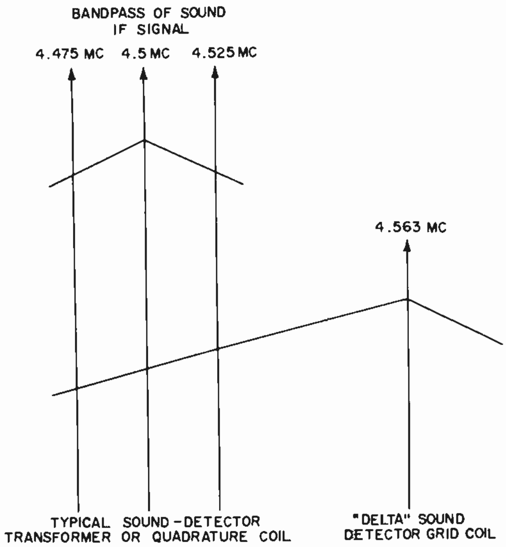


Fig. 14-12. The frequency-response curve of the delta sound detector.

pass of the sound-IF signal will fall on one slope of the response curve of the delta circuit—hence the name, slope detector. The amplitude of the signal pulses developed by the tuned circuit and impressed on the triode grid will change linearly as the incoming signal frequency varies. Fig. 14-12 shows this change. Thus, any frequency modulation in the input signal will develop a corresponding amplitude modulation in the signal at the grid of the triode. The action of the triode is such that it rectifies and amplifies this signal; and in the plate circuit, capacitor C5 filters the 4.5-mc component of the signal to ground, allowing the audio signal to pass on to the audio output stage.

No amplitude modulation should be in the signal output from the triode except the amplitude modulation due to slope detection. The diode circuit and AM rejection control act to reduce any AM interference signals. The average conduction of the diodes changes whenever an amplitude change occurs in the input signal from the preceding stage. The change in average current through the secondary of transformer L1 is reflected back to the plate circuit of the preceding stage as a changing load impedance. This causes a compressing action on any AM peaks which might appear in the signal. The residual AM peaks appear across R2 as negative peaks only. When these are applied to the grid of the triode through C3 and the tuned circuit, they are prevented from reaching the plate circuit because the tube is biased near cut-off by the proper setting of control R2.

QUESTIONS

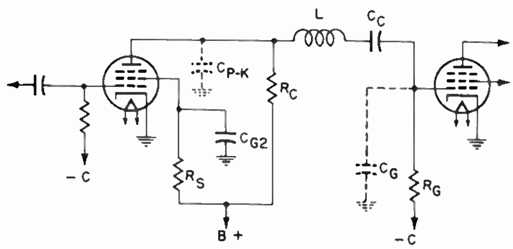
1. What is the standard sound IF for television?
2. What is the deviation of the television sound carrier for maximum modulation?
3. In intercarrier receivers, the sound signal may be taken off at the outputs of what two stages?
4. In the ratio detector, are the diodes arranged in opposition to one another or in series?
5. Why isn't a limiter necessary with a ratio detector?
6. Name the three functions of the 6BN6 used in a gated-beam detector?
7. How is limiting accomplished in the 6DT6 used in a locked-oscillator detector?

Chapter 15

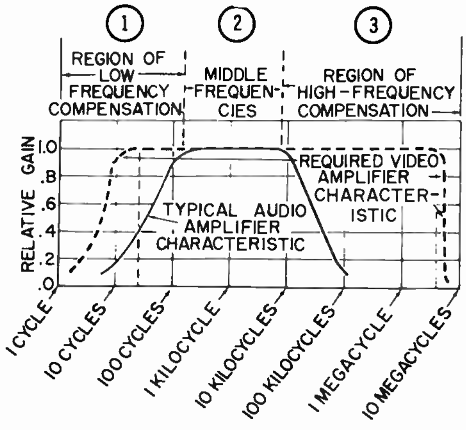
Video Amplifiers

The output of the video detector seldom exceeds a few volts. Since the picture tube requires a grid swing of approximately 40 volts for its range of black to white, the signal from the video detector must be amplified through one or more stages of video amplification.

In our study of the nature of the video modulating signal, we have seen that the range of frequencies extends from 30 cycles to over four million cycles per second. For an amplifier to provide uniform gain over this extended band, compensating circuits and



(A) Typical resistance-coupled audio-amplifier stage.



(B) Required video bandwidth compared with typical audio bandwidth.

Fig. 15-1. Response requirements of video-versus-audio amplifiers.

pentode amplifying tubes of high transconductance must be used. The basic circuit, to which correction networks are applied, is the familiar resistance- and capacitance-coupled audio amplifier.

Fig. 15-1A shows a typical pentode amplifier with its coupling elements. The interelectrode capacitances of the tubes are indicated by the dotted lines. When the amplifier is employed in a conventional broadcast receiver, the solid-line curve of Fig. 15-1B would represent an entirely adequate gain-versus-frequency characteristic for sound reproduction. The dotted line in Fig. 15-1B shows the required gain characteristic of an amplifier for the video band of frequencies.

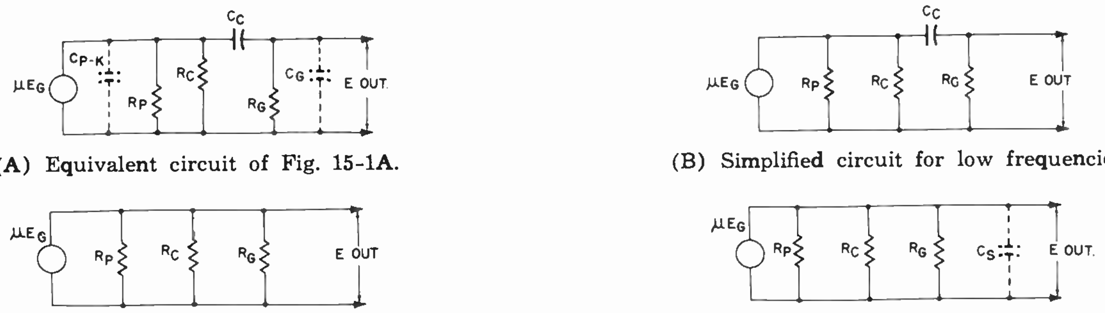
The effect of the circuit elements on the frequency characteristic is shown in Fig. 15-2. The equivalent circuit of the amplifier at all frequencies is shown by the network in Fig. 15-2A. This network has been redrawn in Figs. 15-2B, 15-2C, and 15-2D to represent the equivalent circuits at low, medium, and high frequencies. These ranges correspond respectively to regions 1, 2, and 3 of Fig. 15-1B.

Three separate methods of extending the range of a resistance- and capacitance-coupled amplifier for video use are:

1. Low values of plate load or coupling resistance R_C with high-transconductance tubes.
2. Low-frequency compensation to overcome effects of coupling network C_C-R_G.
3. High-frequency compensation to overcome effects of total circuit capacitance C_S.

EFFECT OF PLATE-LOAD RESISTOR ON BANDWIDTH AND GAIN

Fig. 15-3 shows the effect of changing the value of plate-load resistor R_C (see Figs. 15-1 and 15-2) in a resistance-coupled amplifier stage employing a high-transconductance pentode, such as the 6CL6, 6AC7, or 6AU6. The band of video frequencies over which the output is flat is greatly extended as the coupling resistance is decreased. The choice of load resistor is a compromise between bandwidth and gain. The



(A) Equivalent circuit of Fig. 15-1A.

(B) Simplified circuit for low frequencies.

(C) Simplified circuit for middle frequencies.

(D) Simplified circuit for high frequencies.

R_C = Plate coupling resistor; R_P = Plate resistance; R_G = Grid resistor; C_C = Coupling capacitor;
C_G = Grid-to-cathode capacitance; $C_{P\text{-}K}$ = Plate-to-cathode capacitance; $C_S = C_{P\text{-}K} + C_G$ = Total shunt capacitance.

Fig. 15-2. Equivalent circuits of resistance-coupled stage at low, middle, and high frequencies.

voltage gain of a video stage is seldom more than 20, whereas in resistance-coupled audio stages gains of as high as 150 are possible. Load resistors of 2,000-4,000 ohms are common in video amplifiers. After the value of load resistance is determined, the stage is compensated to raise the gain at frequencies below approximately one-hundred cycles and above several hundred kilocycles.

LOW-FREQUENCY COMPENSATION

At the low-frequency end of the video range (region 1 in Fig. 15-1B), the equivalent circuit of the amplifier is the same as the circuit in Fig. 15-2B. Coupling capacitor C_C in series with grid resistor R_G acts as a voltage divider. As the frequency decreases, the reactance of capacitor C_C increases. Since the voltage across grid resistor R_G constitutes the output voltage of the amplifier, the output drops as the capacitive reactance increases.

To compensate for this voltage drop, a "bass-boost" type of circuit, R4-C4 in Fig. 15-4A, is connected in series with coupling resistor R3 in the plate load circuit. The bass-boost circuit increases the plate load as the frequency decreases. The size of capacitor C4, usually an electrolytic, is such that at all frequencies above approximately 100 cycles its reactance is low compared with that of resistor R4. Capacitor C4 virtually short circuits the resistor, leaving load resistor R3 effective for the middle- and high-frequency range. As the frequency decreases, the reactance of C4 increases. This increases the total plate load. The stage gain follows this increase in plate load, as shown in Fig. 15-3.

Fig. 15-4C shows the effect on the low-frequency gain when the product of R4 times C4 (the time constant of the correction network) is changed. In the circuit in Fig. 15-4A, the time constant of R4 times C4 is one second.

The low-frequency components of the video signal are generated by scanning large objects or areas of uniform tone. Poor low-frequency response produces improper contrast of large areas to the smaller objects or fine detail of the picture.

Other sources of attenuation or loss of low-frequency gain in the resistance-coupled amplifier of Fig. 15-1A are the screen and cathode circuits. The effect of the screen circuit is minimized by the use of a large electrolytic bypass capacitor.

When negative grid-bias voltage for the video-amplifying stage is obtained from a resistor in the cathode circuit, the resistor can introduce degeneration and cut the stage gain unless it is bypassed by a sufficiently large capacitance. The reactance of this capacitor increases as the frequency decreases. Thus, there is a loss of low-frequency gain with respect to the middle- and high-frequency ranges. Low-frequency loss from this source can be compensated in the same manner as described for the coupling network. To avoid this loss, many receivers operate with the cathode grounded and the bias voltage

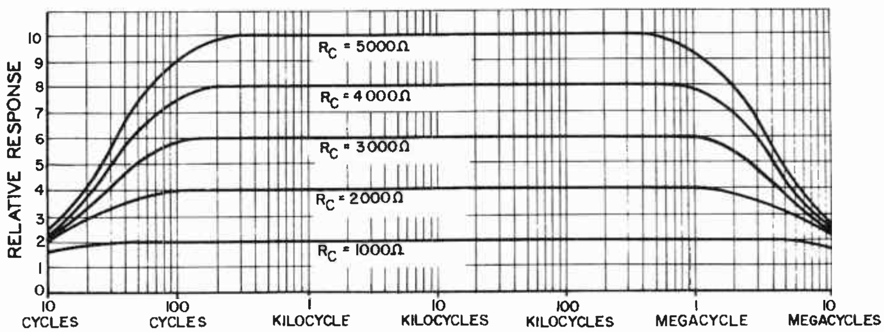
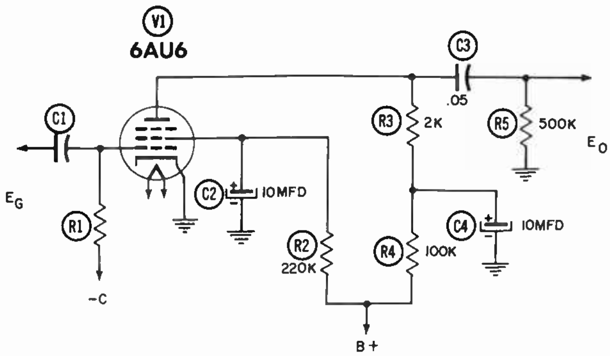
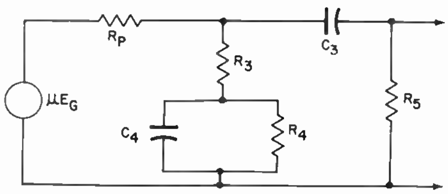
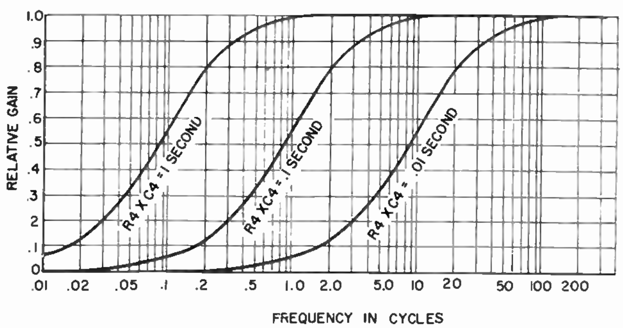


Fig. 15-3. Effect of plate-coupling resistor on gain and bandwidth.



(A) Low-frequency compensation for effect of coupling network C3-R5.


(B) Equivalent low-frequency circuit.



(C) Effect of correction network R4-C4 on gain.

Fig. 15-4. Low-frequency compensation for loss of gain due to coupling circuits.

developed by grid current through a high value of grid leak.

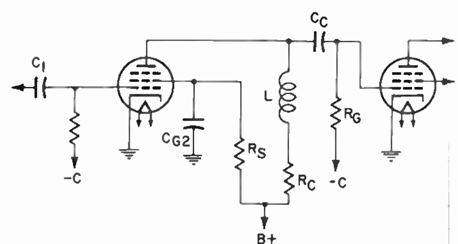
HIGH-FREQUENCY COMPENSATION

The high-frequency region of video amplification (region 3 in Fig. 15-1B) is responsible for the fine detail of the reproduced picture. If the video-amplifier gain is deficient at high frequencies, fine lines or small picture elements will be missing from the picture or will be blurred. Lack of high-frequency gain can be detected by examining the center or narrow portion of the vertical wedges of a test pattern.

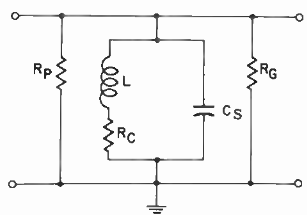
Loss of high-frequency gain in a resistance-coupled amplifier is caused by the shunting effect of the stray capacitance of the various circuit elements to ground. These stray capacitances include the plate-to-cathode capacitance of the amplifying tube, the capacitance of the various coupling components (resistors and capacitors), and the input or grid-to-cathode capacitance of the next tube. In the equivalent circuit in Fig. 15-2D, the shunt capacitance (the sum of these elements) is shown by the dotted capacitor C_S.

We have seen in Fig. 15-3 that the high-frequency response can be extended by decreasing the value of the plate-load resistor. This is accomplished at the expense of stage gain. The gain of the stage at high frequencies is proportional to the impedance of the parallel combination of plate-coupling resistor R_C, grid resistor R_G, and total shunt capacitance C_S. At the frequency at which the capacitive reactance of C_S is equal to the value of the resistors in parallel, the gain is 70.7% of its flat, middle-frequency value. See point A on the uncompensated curve of Fig. 15-5C. This frequency is usually considered the limit at which the gain can be brought up to the mid-frequency level by corrective networks. These networks employ inductors (known as peaking coils) whose reactances increase with increasing frequency and thus compensate for the loss of impedance due to the shunt capacitance.

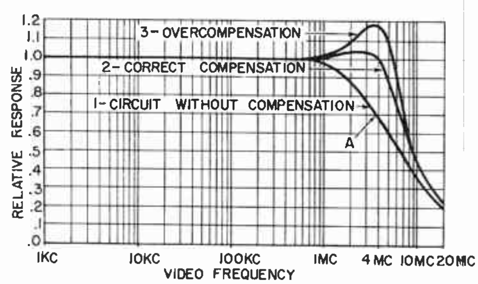
A great variety of high-frequency compensating circuits have been proposed, but commercial receivers employ only two basic circuits, shunt and series compensation circuits. A combination of shunt and series compensation is usually employed.



(A) Shunt-peaking, high-frequency compensation.



(B) High-frequency equivalent circuit.



(C) Effect of shunt-peaking choke.

Fig. 15-5. High-frequency compensation by shunt peaking.

Shunt Peaking

Fig. 15-5 shows the basic shunt-peaking type of high-frequency compensation. Although peaking coil L is in series with plate-load resistor R_C, peaking coil L raises the impedance of the shunt circuit of R_C and C_S. This circuit is in parallel, or shunt, with the plate resistance and the following grid resistor (Fig. 15-5B).

Curve 1 of Fig. 15-5C shows the frequency characteristic of the circuit in Fig. 15-5A before correction is applied. Notice that point A of 70% response occurs at approximately 3.5 megacycles. At this point, the capacitive reactance of C_S equals the effective shunt resistance of R_C and R_G in parallel. At this frequency, the value of L is chosen so that the parallel impedance of the circuit of R_C, R_G, C_S and L is increased to yield the same gain obtained in the middle frequency range. This condition is shown as curve 2 in Fig. 15-5C.

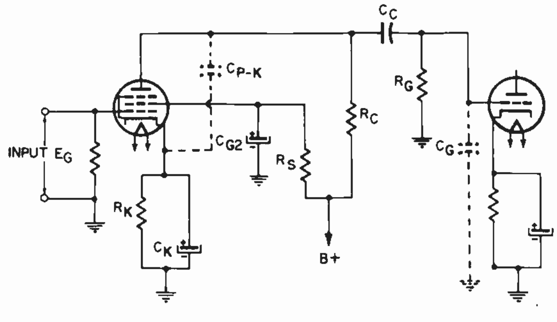
The network in Fig. 15-5B will be recognized as a parallel-tuned circuit with series and shunt resistors. Under the condition for correct compensation (curve 2 in Fig. 15-5C), the resonant frequency of the circuit is 1.41 times the frequency discussed in the previous paragraph. If the inductance is increased to reduce the resonant frequency, overcorrection will be obtained (curve 3 in Fig. 15-5C). Such overcompensation will make fine details in the picture too dark with respect to large areas. The peaking coils of a television receiver are designed to operate with particular types of tube and circuit capacitances. Should the peaking coils need replacing, exact duplicates must be used.

Series Peaking

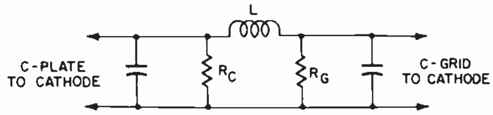
Fig. 15-6A shows another method of raising the response curve at the high-frequency end. A series-peaking coil in series with the circuit produces an equivalent network, as shown in Fig. 15-6B. This will be recognized as a pi-type filter network. The inductance isolates the plate-to-cathode capacitance of the amplifier tube from the input capacitance of the following tube. A higher value of coupling resistor with consequently higher stage gain can be used. The gain of the stage with series peaking can be made 50% greater than that of one with shunt peaking alone for the same bandwidth. With series peaking, the resonant frequency of the circuit L-C_G is made higher than the upper limit of the desired video range and tends to counteract the loss due to capacitance $C_{P\text{-}K}$ across coupling resistor R_C.

Combination Shunt and Series Peaking

From the foregoing discussion, it is evident that shunt peaking and series peaking operate independently and can complement each other. Most modern television receivers employ a combination of both methods. A typical example is shown in Fig. 15-7.



(A) Series-peaking, high-frequency compensation.



(B) Equivalent circuit at high-frequencies.

Fig. 15-6. High-frequency compensation by series peaking.

Here, inductor L3 operates as a shunt-peaking inductor, and L2 as a series-peaking inductor. Note that both of these coils have a shunting resistor connected across them. The shunting resistor reduces the Q of the circuit to smooth out the gain-versus-frequency curve. This resistor also damps the circuit and thus prevents it from being shock-excited into transient oscillation by sharp video impulses or noise peaks. Such oscillation produces smearing and a negative image following fine-detail portions of the picture, as shown in Fig. 15-12.

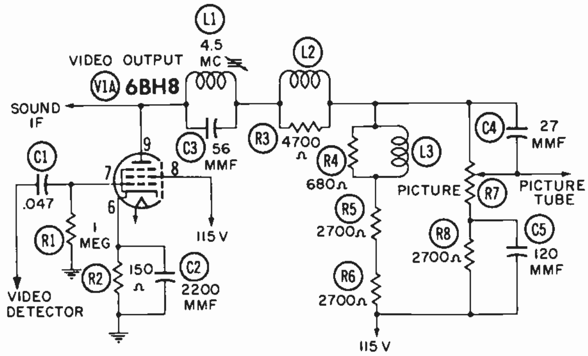


Fig. 15-7. Video stage with both high- and low-frequency compensation.

In the discussion of video detectors, mention was made of series and shunt peaking for correction of high-frequency losses in the detector-output circuit. This peaking is the same as the type just discussed.

PHASE SHIFT IN THE VIDEO AMPLIFIER

In our discussion of the video amplifier up to this point, we have considered the requirement of gain versus frequency only. Another equally important consideration is the possible phase shift, or time delay, of the signal as it passes through the amplifier.

In the amplification of sound voltages in a broadcast receiver, phonograph amplifier, or sound system, phase shift at one end of the audio-frequency spectrum with respect to the other is seldom important. The ear is not sensitive to phase shift, and for this reason, the service technician need not give it any consideration. In the television video amplifier, however, phase shift is extremely important. If not corrected, phase shift can cause badly distorted and smeared pictures.

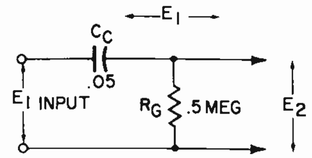
Almost all sound waveforms are sinusoidal. An understanding of the action of sine waves of voltage passing through an amplifier is sufficient background for understanding the operation of sound amplifiers. The video signal, however, is often a square or rectangular waveform. This can readily be understood by considering the output from the camera tube as it scans a black bar in front of a white background. During the scanning process, the video signal is near its zero amplitude as is crosses the background. Suddenly the video signal rises to maximum amplitude and remains there as the bar is crossed; then it drops to zero amplitude again when the background is reached. The width of the bar determines the half-wave of a low frequency. This frequency is the fundamental frequency for that particular element of the picture. The square wave is composed of this fundamental frequency plus a great number of harmonics of different amplitudes. Together, this fundamental frequency and the harmonics will produce the rapid rise and fall at the ends of the bar. Thus, to reproduce a black bar, no matter what its length, all of the harmonic frequencies must be amplified. As we have seen, these harmonic frequencies can extend to four megacycles.

In our study of the action of amplifier stages, we learned that the wave applied to the grid is shifted 180° in phase when it appears across the plate-load resistor. As far as the tube itself is concerned, a wave of any frequency in the video range is shifted 180° in phase during voltage amplification through a single tube. The network of resistors and capacitors which constitutes the coupling elements between the tubes can, however, cause a phase shift which differs both in amount (number of degrees) and in direction for different frequencies in the video range.

For the middle range (region 2 in Fig. 15-1B), the coupling network is resistive (see Fig. 15-2C), and a constant 180° phase shift occurs because of the tube action.

Low-Frequency Phase Shift

At low frequencies (region 1 in Fig. 15-1B), coupling network C_C-R_G (whose effect on low-frequency gain has already been shown) produces a leading phase shift which increases as the frequency de-



$$\text{TANGENT OF PHASE SHIFT ANGLE} = \frac{X_C}{R_G} = \frac{1}{2\pi F C_C R_G}$$

Frequency (cps)	*Phase Shift of E_2 (Degrees)*	*Time Delay (μs)*	*Horizontal Displacement of Image (inches) (21-in. Tube)*
500	0.7	4	1.4
200	1.8	25	9.0
100	3.6	100	1 Line + 13
70	5.2	206	3 Lines + 5.5
50	7.2	400	6 Lines + 6.8
30	12.0	1,110	17 Lines + 10.8

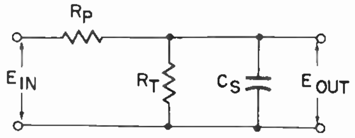
Fig. 15-8. Chart showing low-frequency phase shift due to coupling elements.

creases. This phase shift is proportional to the ratio of the reactance of the coupling capacitor to the resistance of the grid resistor. If no correction were applied, this phase shift would cause the effects shown in the table in Fig. 15-8. A very small phase shift can cause a large time difference at very low frequencies. As a result, large areas will show smearing of the edges and an uneven tonal reproduction in the picture. Excessive phase shift at very low frequencies (30 to 70 cycles) will also cause a gradual shading of the picture from top to bottom because the effect of a single picture element lasts for more than one horizontal line.

Since the phase shift is a leading effect, it can be corrected by a shunt circuit consisting of capacitance and resistance in parallel. This is the same type of network required to compensate the low-frequency loss of gain (R4-C4 in Fig. 15-4). Thus, both phase shift and gain can be corrected by the same network of added circuit elements.

High-Frequency Phase Shift

At high frequencies (region 3 in Fig. 15-1B), shunt capacitance C_S can cause a lagging phase shift of the high frequencies with respect to the middle-frequency portion of the video signal. Fig. 15-9 shows the parts of the equivalent amplifier circuit (see Figs. 15-1 and 15-2) responsible for high-frequency increases, the time delay drops, and the image displacement becomes less. Again, as with low-frequency phase shift, the same corrective networks which compensate for loss in gain are used to produce a corrective phase shift. The shunt- and series-peaking coils and their combinations produce an over-all circuit phase shift which is proportional to frequency.



$R_T = 4\,000\,\Omega$
$C_S = 50$ MMF

TANGENT OF PHASE SHIFT ANGLE $= -2\pi F C_S R_T$

Frequency	*Phase Shift (Degrees)*	*Time Delay (μs)*	*Horizontal Displacement of Image (inches) (21-in. Tube)*
10 KC	.72	2.0	.71
100 KC	7.2	2.0	.71
1 MC	64.4	1.8	.64
2 MC	103.8	1.4	.5
3 MC	124.0	1.266	.45
4 MC	136.6	0.95	.34

Fig. 15-9. Chart showing high-frequency phase shift due to shunt capacitances.

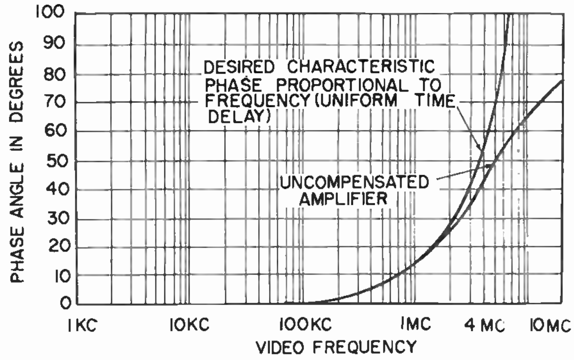
The Ideal-Phase Shift Proportional to Frequency or Uniform Time Delay

At the horizontal scanning speed (15,750 cycles per second), the spot moves across a picture 19 inches wide at approximately 356,000 inches per second. Thus, a time delay of one microsecond will produce a difference in position of about three-eighths of an inch. If all frequencies in the video signal suffered this time delay, the picture would be satisfactory, but would be displaced three-eighths of an inch to the right and could be moved back by adjusting the horizontal centering. A uniform time delay means a different phase shift at each frequency (a phase shift proportional to the frequency).

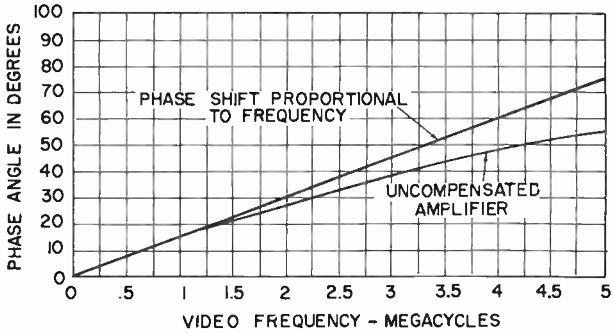
The relation between time delay and frequency is expressed by the equation:

$$\text{Time Delay} = \frac{\text{Phase Shift in Degrees}}{360° \times \text{Frequency in Cycles}}$$

Fig. 15-10 shows the desired phase shift relationship proportional to frequency, or uniform time delay, as applied to the video amplifier of a television receiver. At high frequencies, the phase shift of the uncompensated amplifier drops below the desired proportional curve which will produce a picture without blurred or shifted elements.



(A) Ideal phase characteristic — plotted in logarithmic coördinates.



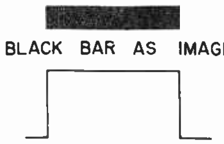
(B) Same curve as in A but on linear basis.

Fig. 15-10. Phase shift requirements of a video amplifier for perfect reproduction.

SMEARING OF THE TELEVISION IMAGE BY AMPLITUDE OR PHASE DISTORTION

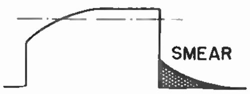
Fig. 15-11 illustrates the effects on a square wave when various deficiencies exist in the video amplifier. Fig. 15-11A shows the correct video signal that should be produced when the black bar is scanned.

Fig. 15-11B shows the square waveform after it has passed through an amplifier having insufficient high-frequency gain. The leading edge of the amplified wave has a gradual slope rather than an abrupt rise. This slope will cause shading from gray to black in the reproduction. At the trailing edge, the exponential decay of the wave will cause a gray-to-white smear.

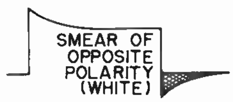


(A) Correct video signal.

(B) Insufficient high-frequency response.



(C) Excessive low-frequency response.



(D) Insufficient low-frequency response.

Fig. 15-11. Square-wave video signals, showing the effect of amplifier deficiencies.

Fig. 15-11C shows the effect of overcompensation, or excessive low-frequency response. This effect is the same as the effect caused by insufficient high-frequency response, although not as pronounced.

Fig. 15-11D illustrates insufficient low-frequency response accompanied by attendant phase shift. The front edge of the bar is black, shading to gray; and the smear following the bar is white, shading to gray. This type of smear is known as a "trailer."

Fig. 15-12 shows the effect of insufficient damping of the peaking circuits used for high-frequency compensation. The oscillation due to shock excitation of these circuits by the short-duration square-wave picture elements can be seen as alternate white and gray ghosts adjacent to the black circular rings of the test pattern.

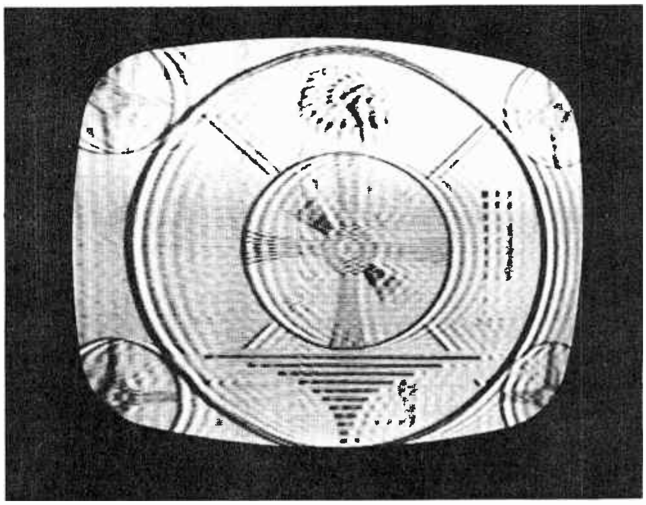


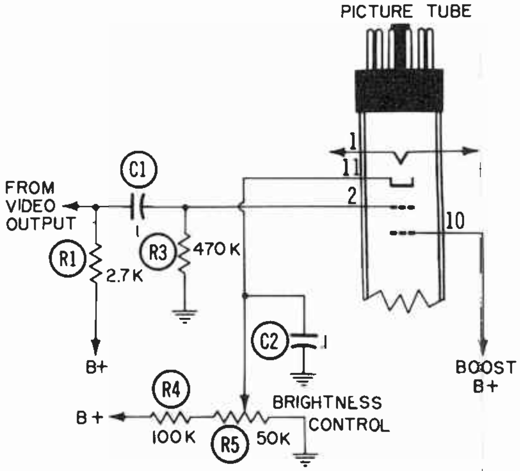
Fig. 15-12. A test pattern, showing the effect of transient oscillation.

BRIGHTNESS CONTROL

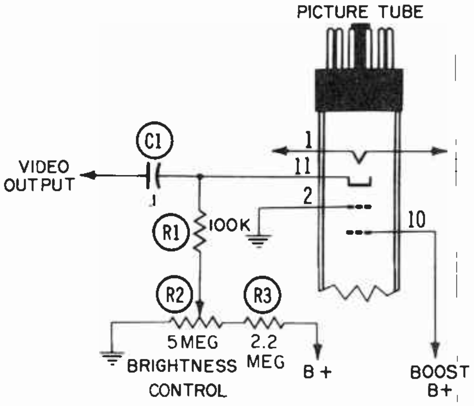
Our previous study of the television receiver system concerned itself with the motion of the electron beam in the picture tube as it scanned the picture area. We showed in Chapter 1 that the intensity of illumination of the cathode-ray screen could be controlled by a grid cyclinder in the picture-tube gun. At several points in the text, we indicated that the variation of video or picture signal on this control grid is responsible for the instantaneous changes of spot illumination which makes up the elements of the picture. When the signal voltage on this control element changes in a negative direction, a darker spot is produced on the screen. Finally, at some critical negative voltage, the spot of light is entirely extinguished.

One of the essential controls of the television receiver is a bias adjustment on the grid. This adjustment assures that the blanking level, or pedestal, of the signal occurs at the black point. Fig. 15-13 shows two bias systems in which the voltage established by adjustment of the brightness control biases the control grid with respect to the cathode and determines the correct picture brightness. Because of the polarity of the video signal in Fig. 15-13A, the signal from the plate of the video-output tube is connected to the control grid of the picture tube.

Fig. 15-13B shows the video signal applied to the cathode of the picture tube. In either case, the brightness control is a voltage adjustment of the bias between the control grid and cathode, and establishes the correct blanking or black level.



(A) Video signal applied to the grid.



(B) Video signal applied to the cathode.

Fig. 15-13. Brightness-control circuits.

In older television receivers, the brightness control is adjusted with no picture signal present. A point will be reached at which the vertical retrace lines caused by the motion of the spot from the bottom to the top of the picture will become visible. The control should be turned counterclockwise until these lines just disappear. In modern receivers incorporating vertical-retrace blanking circuits, the brightness control can be adjusted to suit the viewer.

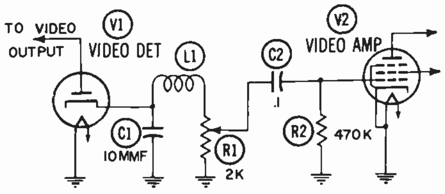
PICTURE OR CONTRAST CONTROL

The contrast control of a television receiver has much the same function as the volume control of a broadcast or communication receiver. "Contrast" means the ratio of light intensity of the brightest highlight in the picture to that of the deepest shadow. Because of scattering of light at the fluorescent surface, the black of the shadow can never be an absolute black, or absence of light. Since this ratio of maximum illumination to illumination at the cutoff point is directly proportional to the voltage swing of the control grid of the picture tube, the contrast can be controlled by varying the video-amplifier output. The output can also be controlled in the RF amplifier, the first detector, or the IF amplifier; and this was done in early TV receivers.

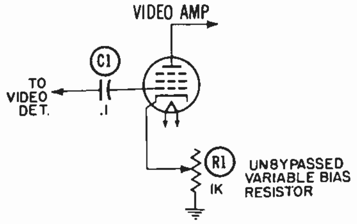
Contrast Control by Change of Video-Amplifier Gain

Figs. 15-14A and 15-14B show two methods of controlling the video-amplifier output. In Fig. 15-14A, the contrast control is identical to the familiar radio volume control, which is a potentiometer functioning as the second-detector load. circuit elements C1 and L1 are employed for video high-frequency compensation.

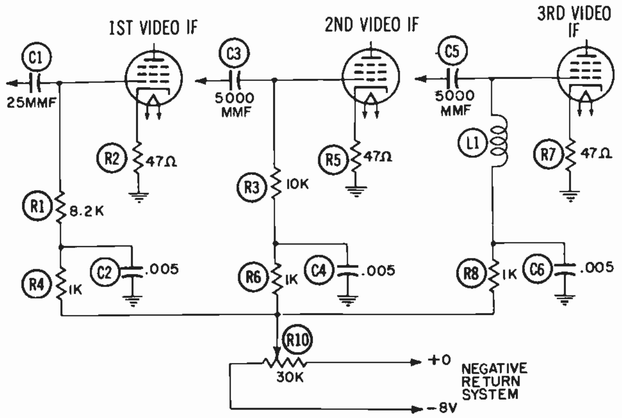
In Fig. 15-14B, a variable cathode-bias resistor in the video amplifier is employed to change the operating point of the tube and, hence, its transconductance or gain. The control is left unbypassed, so that the degeneration introduced by the resistor is the same at all frequencies and does not affect the amplifier bandwidth.



(A) Contrast control by potentiometer.



(B) Contrast control by video-amplifier bias.



(C) Contrast control by change of video-IF bias.

Fig. 15-14. Contrast-control circuits.

Contrast Control by Change of Video-IF Amplifier Bias

Fig. 15-14C shows an early method of contrast control in which the gains of three video-IF amplifiers were adjusted by changing the grid bias. This grid bias is developed below ground potential. Several methods of deriving a negative voltage for this control purpose are employed:

1. The ground point is connected to an intermediate point in a voltage divider, rather than to the center tap of the power transformer.
2. A selenium rectifier is used to derive approximately 8 volts DC from the 6.3-volt AC filament supply. (The peak of the 6.3-volt AC wave is 8.9 volts.)
3. The negative bias developed in the grid of the horizontal-scanning oscillator is filtered and used as a source of contrast-control voltage.

Locations of the Contrast Control

Two different locations of the contrast, or picture, control are illustrated in the circuits in Figs. 15-15 and 15-16. Both types of connections are widely used today. In Fig. 15-15, control R1 is connected between the cathode of the video amplifier and ground. In this particular circuit, the grid resistor is returned to the cathode end of the control. The bias on the tube, therefore, is not varied when the control is reset; instead, the effective B+ voltage on the tube is varied. The grid resistor is grounded in some circuits of this type, and a change in contrast setting does change the bias. In either case, the gain of the tube is altered according to the contrast setting.

Contrast control R1 in the circuit in Fig. 15-16 is connected in parallel with the plate load resistors. The gain of the video amplifier is constant. A variable proportion of the output signal is tapped off at the arm of the contrast control and is applied to the picture tube. The plate load resistors in this circuit are composed of R2 and R3 in series; L2 and L3 are the peaking coils, and C2 is for frequency compensation.

VIDEO COUPLING TO CRT

The output signal of the video amplifier in Fig. 15-15 is AC-coupled to the cathode of the picture tube through capacitor C6. The level of the DC voltage on the cathode is determined by the setting of the brightness control. When the video signal is applied to the cathode, the average value of the signal always coincides with this preset DC level. Theoretically, this is not a workable arrangement because the composite video signal is supposed to play some part

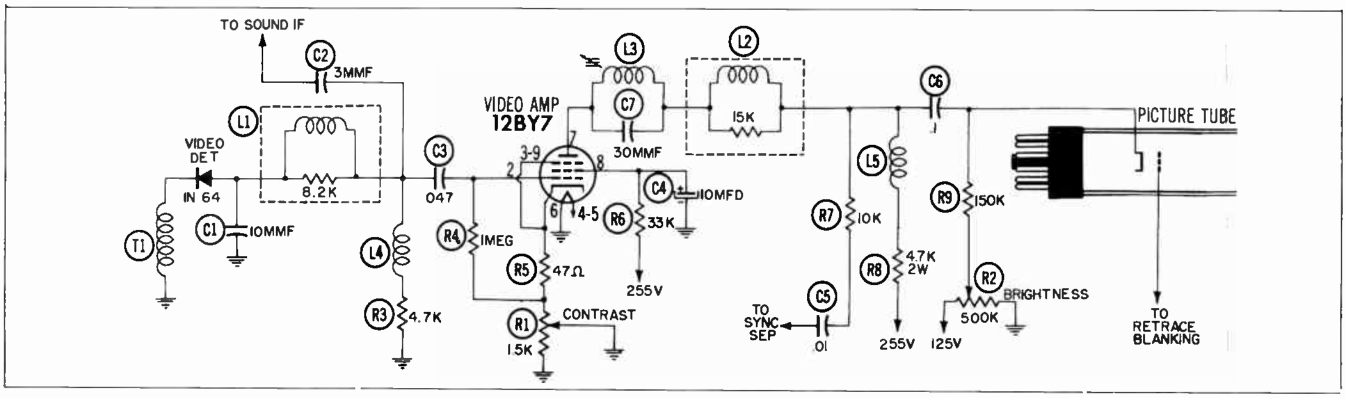


Fig. 15-15. Schematic of a typical single-stage video amplifier employing AC coupling.

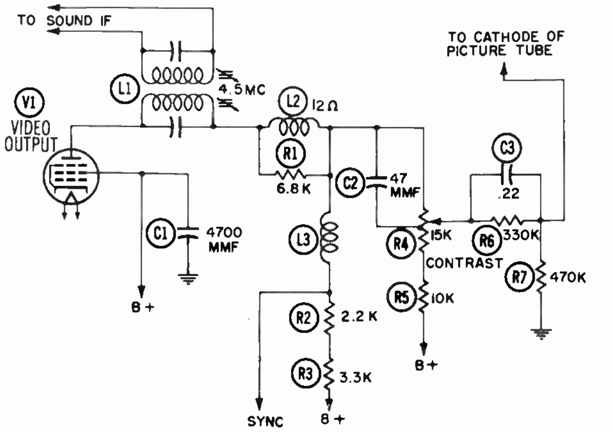


Fig. 15-16. Contrast-control circuit in coupling to picture tube.

in the determination of the DC cathode voltage. Regardless of shifts which may occur in the average value of the video signal, the pedestals of the sync pulses in that signal should remain at a constant DC level equivalent to the cutoff voltage of the picture tube. If they do, they establish a fixed reference point for all black portions of the picture.

When the video signal is allowed to arrange itself around a fixed average value, as is done in an AC-coupled amplifier, the level of the sync pedestals shifts somewhat whenever the average level of brightness in the picture changes. A given shade of gray, therefore, is not reproduced exactly the same in a predominantly dark scene as it is in a light scene.

In actual practice, a satisfactory picture can be obtained with an AC-coupled video amplifier. Most modern TV receivers contain this type of circuit. The appearance of retrace lines in the picture (potentially a serious drawback of AC coupling) is prevented by the use of an efficient retrace-blanking circuit. Sharp negative pulses with an amplitude of more than 50 volts are obtained from the vertical-deflection circuits and are applied to the grid of the picture tube. These pulses cut off the tube during the retrace period, and the retrace lines are not seen.

Many older models and a few of the new ones contain a diode which functions as a DC restorer. This tube re-establishes the fixed black level that is lost in capacitive coupling. The diode is generally used in two-stage amplifiers where the video signal is coupled to the grid of the picture tube. Part of the resistive path between the grid and ground is shunted by the diode. How much the diode conducts depends upon the peak value of the tips of the sync pulses in the signal. A capacitor is charged in proportion to the amount of conduction. The charge on this capacitor regulates the DC grid voltage of the picture tube, and the pedestals of the sync pulses are always held at a constant level. The DC restorer insures that all signals representing black objects will drive the picture tube completely to cutoff if the brightness control is set correctly.

The DC level of the video signal can be preserved without a DC restorer if direct coupling is used throughout the entire path of the signal from the video-detector output to the driven element of the picture tube. A direct-coupled, single-stage video amplifier is shown in Fig. 15-16. Note the absence of any series capacitors in the signal path.

PUSH-PULL VIDEO AMPLIFIER

Fig. 15-17 shows an unusual, but effective, video-amplifier circuit that utilizes both sections of a 6BH8 triode-pentode. The pentode section of the tube operates as a conventional video-output stage and delivers a signal with a peak-to-peak amplitude of 50 volts to the cathode of the picture tube. A portion of the pentode output is applied to the triode grid through a voltage-divider network comprised of R7, R9, and R10. The triode takes this sample of the video signal amplifies and inverts it, and delivers the resulting signal to the picture-tube grid at an amplitude of up to 55 volts peak-to-peak.

The 50 volts on the cathode of the picture tube and the 55 volts on the grid provide about the same signal drive as one signal of 105 volts peak-to-peak applied to a single element. This design, therefore, gives the same results as a push-pull circuit, even though it does not provide true balanced push-pull

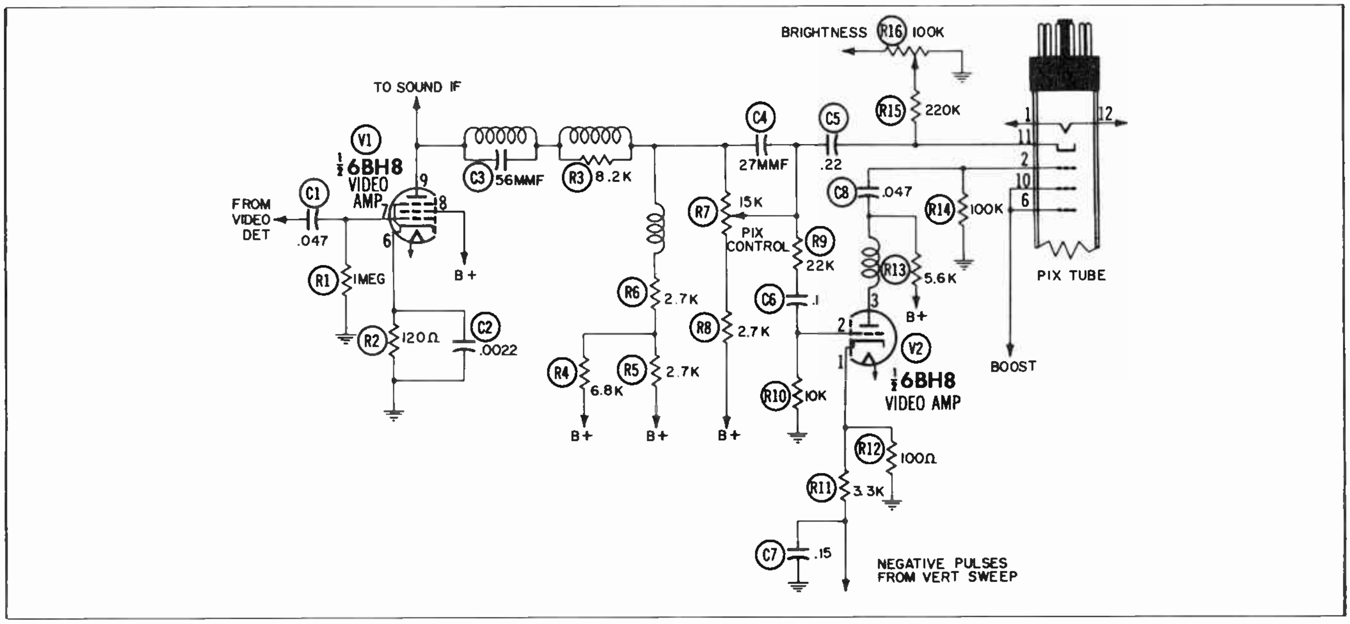


Fig. 15-17. A push-pull video amplifier.

operation. This video helper is added to the circuit to increase the amount of picture contrast obtainable in a receiver with a low-voltage B+ supply.

A retrace-blanking circuit is associated with the picture-tube control grid. Negative pulses from the vertical-sweep system are applied to the cathode of the triode section of the 6BH8. Amplified by the triode, these pulses appear at the picture-tube grid and cut off the electron beam during vertical retrace.

QUESTIONS

1. What happens to the picture when the video amplifier has poor low-frequency response?
2. How is the effect of the screen circuit on low-frequency response minimized?
3. What happens to the picture when the video amplifier has poor high-frequency response?
4. What causes the loss of high-frequency response in the video amplifier? How is this loss minimized?
5. What is the function, circuitwise, of the brightness control?
6. Name three ways the contrast control can be used in the video-amplifier circuit to control the amplitude of the video signal applied to the picture tube.
7. When the video signal is AC-coupled to the cathode of the picture tube and the grid is effectively grounded, what determines the level of the DC voltage on the cathode?

EXERCISES

1. Draw the basic circuit of a video amplifier with shunt peaking.
2. Show the equivalent circuit of Exercise 1.

Chapter 16

Automatic Gain Control

Automatic gain control (AGC) minimizes the effect of changes in signal strength at the receiver antenna. The gains of the RF and IF stages are so regulated that a strong signal is amplified less than a weak signal. As a result, the quality of the TV picture tends to be relatively constant.

Variations in signal strength are of two types—(1) variations between signals received on different channels, and (2) variations occurring from time to time on the same channel.

Both strong and weak channels are available in many locations. If AGC is provided in the receiver, the contrast control does not need to be reset each time a new channel is tuned in. AGC also compensates for extremely strong signals received in powerful station areas.

The AGC system levels out most of the periodic amplitude variations which would cause fading on a particular channel; therefore, a steady picture is obtained, even in moderate fringe areas. The rapid flutter caused by airplanes flying near the path of the transmitted signal is also corrected as much as possible through AGC action.

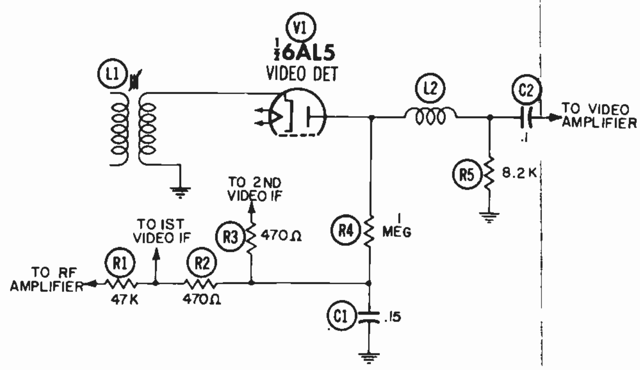
RECTIFIED AGC

A typical AGC circuit composed of a simple filter network can be seen in the schematic in Fig. 16-1A. The filter is composed of R4 and C1. The time constant of this filter circuit is approximately 0.15 second.

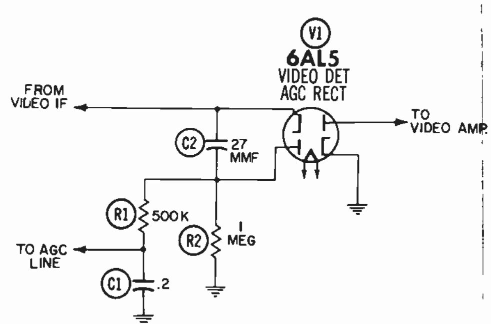
The performance of this system can be refined by the use of a special AGC diode and a shorter charging time constant than the discharging time constant. A circuit with these modificatioins will not be affected by variations in scene brightness. Such variations cause shifts in the amplitude of the carrier in the television signal.

The average value of voltage of a video signal is obviously not an absolutely true indication of signal strength. If the AGC voltage were developed from this average voltage, AGC filter output would tend to increase during the transmission of scenes containing many large, dark objects or a dim background.

The only portion of the composite video signal which has a constant amplitude regardless of picture content is the sync-pulse signal. If the strength of the received signal does not change, a consistent peak value of voltage is reached by the tips of these pulses.



(A) AGC circuit composed of a simple filter network with a time constant of approximately 0.15 second.



(B) AGC circuit with improved action obtained by a lengthened time constant in the discharge path.

Fig. 16-1. Schematics of simple AGC systems.

This peak is comparable to the maximum voltage attained during 100 per cent modulation of a carrier by an audio signal. Improved AGC action will be obtained if the AGC filter capacitor can be charged to this peak voltage and if most of this charge can be maintained between pulses.

If the charging and discharging time constants of the AGC filter are of nearly equal lengths, the system can never build up a charge approaching the peak amplitude of the signal voltage. The discharging time constant can be lengthened in order that a greater charge can be retained on the filter capacitor. This feature has been included in the circuit in Fig. 16-1B.

Resistor R2 in Fig. 16-1B corresponds to R5 in Fig. 16-1A, but the value of the resistor in Fig. 16-1B has been increased to one megohm. As a result, the charging time constant is 0.1 second, but the discharging time constant is increased to 0.3 second.

A separate diode must be used to rectify the AGC voltage if a resistor of high value is used in the discharge circuit of the AGC capacitor. The reason for this requirement will be clear if it is noted in Fig. 16-1A that the voltage applied to the video amplifier is developed across R5. Most of the high-frequency portions of the video signal would be lost if that resistor were large in value because the shunting effect of stray capacitance in the video-amplifier input would be exaggerated. R2 in Fig. 16-1B can be as large as necessary because the video detector is separate from the AGC rectifier.

It should be repeated that AGC action can be obtained without special concern for changes of brightness in the picture. However, correcting this condition is important enough that many of the relatively simple AGC systems and all of the more complex ones develop the AGC voltage from the peak voltage of the sync pulses.

AMPLIFIED AGC

An improvement of the basic AGC system was the addition of an amplifier tube to boost the amplitude of the rectified AGC voltage. A two-stage, amplified-AGC circuit can develop an adequate control voltage when the changes in amplitude of the video signal are so slight that a simple AGC system would not respond to them. Amplified AGC is, therefore, more efficient than ordinary simple, or filtered, AGC.

An amplified-AGC circuit is shown in simplified form in Fig. 16-2. An almost pure DC voltage that is

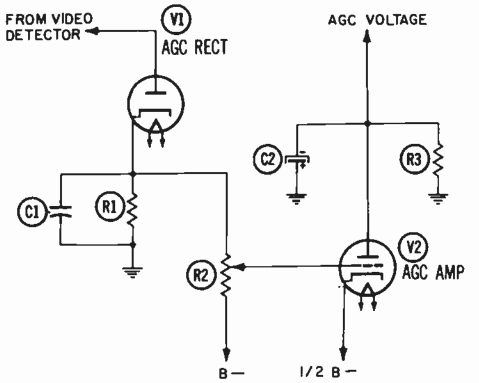


Fig. 16-2. Schematic of amplified-AGC circuit.

positive in polarity is produced across C1 and R1 when a video signal is applied to the plate of V1. This signal is directly coupled to the grid of AGC amplifier V2.

Since the AGC voltage is taken directly from the plate of the AGC amplifier, a negative voltage of low amplitude must be present in the plate circuit of V2.

It is assumed that the circuit in Fig. 16-2 can be supplied with a voltage of approximately −100 volts. A voltage equal to one-half the B− potential is applied to the cathode of the amplifier. The average grid voltage is determined by the setting of potentiometer R2. This control is a voltage divider between the full B− voltage and the slightly positive voltage at the cathode of V1.

The bias on V2 automatically varies in step with the voltage developed across R1 and C1 by the video signal. An increase in amplitude of the incoming sync pulses causes the cathode voltage of V1 to go in the positive direction; therefore, the voltage at the arm of R2 is less negative than before, and the bias on V2 is reduced. Conduction increases in the amplifier. The plate voltage of V2 swings in a negative direction, and C2 in the plate circuit is heavily charged in the polarity shown in Fig. 16-2. The time constant of R3 and C2 is so long that C2 discharges very slowly. The AGC control voltage is, therefore, nearly equal to the voltage which charges C2.

KEYED AGC

Keyed AGC is the most complex form of automatic gain control used in TV receivers: it is also the most efficient. The circuit in Fig. 16-3 is typical of most circuits of this type.

The AGC keying tube used in these circuits is

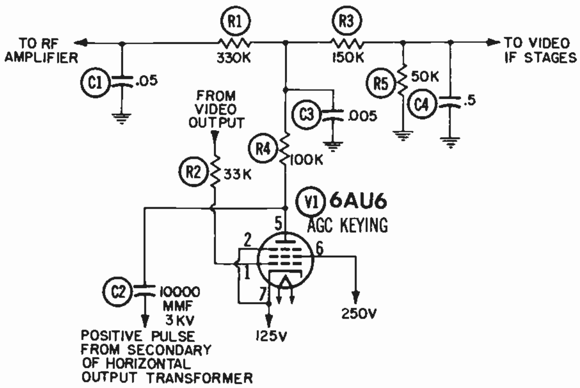


Fig. 16-3. Schematic of keyed-AGC circuit.

typically a pentode, but it could be a triode. Two input signals are required. A composite video signal which contains positive-going horizontal-sync pulses is applied to the control grid, and positive pulses of high amplitude are coupled to the plate from a winding on the horizontal-output transformer. The conduction of the tube depends upon the arrival of a pulse at the plate at the same time that a horizontal-sync pulse arrives at the grid. Since the plate pulses are timed by the horizontal oscillator, they have the same frequency and phase at the sync pulses if the receiver is correctly synchronized.

A short burst of conduction occurs 15,750 times each second in response to the arrival of each pulse. Since the level of the sync tips determines the bias of the keying tube during the burst of conduction, the amount of conduction that occurs during the pulses depends upon the DC level reached by the tips of the horizontal-sync pulses at the grid. The level of the sync tips, in turn, depends upon the strength of the incoming signal. The keying tube is cut off between pulses, and the plate voltage assumes a negative value proportional to the amount of conduction that takes place during the pulses. The greater the conduction, the more negative the average plate voltage. The plate voltage is passed through an RC filter, and the resultant DC voltage is applied to the RF amplifier and to one or more IF amplifiers as grid bias.

Since the keying tube is cut off in the interval between horizontal-sync pulses, noise and video in the input signal cannot affect the production of AGC voltage. In addition, the keyed AGC circuit does not respond to the vertical-sync pulses. The keying tube conducts only during alternate equalizing and serration pulses. The amount of conduction is not sufficient to cause a periodic rise in AGC voltage at the vertical rate of 60 cps. Such a rise is characteristic of other AGC systems. The AGC filter in the keyed circuit, unlike the filters in other AGC circuits, does not have to be designed to remove vertical pulses from the output voltage. The time constant of the filter is, therefore, appreciably shorter in a keyed circuit than in other circuits. This is a desirable feature because it tends to make the keyed circuit largely immune to airplane flutter.

The DC cathode voltage is made several volts more positive than the average DC grid voltage in order that the keying tube will be correctly biased during conduction. The difference between the grid and cathode voltages is more important than their absolute values.

These two voltages have high positive values in most actual circuits. Regardless of this fact, the tube conducts normally because the keying pulses on the plate have a much higher positive value. The reason behind the use of a positive grid voltage is that the grid of the keying tube must be directly coupled to the stage supplying an input video signal to the AGC system. Frequently, the signal source is the plate circuit of a video amplifier, and the high voltage at the amplifier plate also appears at the grid of the keying tube. Direct coupling is used because the variations in grid voltage which occur in response to changes in the DC level of the sync pulses are fundamental to the operation of the AGC system. This reference level would be lost if a capacitor were inserted in the input lead to the grid of the keying tube.

As shown in Fig. 16-3, an isolating resistor R2 is included in the grid circuit. This resistor limits the amount of current in the circuit should the grid draw current. The impedance of the resistor isolates the input capacity of the keying tube from the video circuit; therefore, the loading effect of the AGC circuit on the video circuit is greatly reduced.

Recall that the amount of conduction through the tube is determined by the peak amplitude of the sync pulses. If the video signal is weak, the positive voltage on the grid of the keying tube will be relatively low. The difference between the grid voltage and the cathode voltage is great, and the heavily biased tube passes less than average current. Little AGC voltage is produced under these conditions because the charge on the capacitors in the plate circuit is comparatively small. On the other hand, a strong video signal develops a high grid voltage which is closer to the value of the cathode voltage. The bias on the tube is low, conduction through the tube is heavy, and considerable AGC voltage is thus produced in the plate circuit.

Fig. 16-4 shows an AGC circuit and a noise inverter combined in one tube with a sync-separator circuit. The sync circuit and noise inverter were discussed in Chapter 10. The input signal for the AGC section of the 6BU8 tube is a composite video signal with positive-going sync pulses. The composite video signal is taken from the plate of the video amplifier and applied to one of the second control grids (pin 9) of the 6BU8. This signal is direct-coupled because the DC level of the sync tips is important to the operation of the AGC tube. If the composite video signal is strong, the sync tip level at pin 9 of the 6BU8 will be less negative than if the video signal were weak. In other words, a strong input signal places a relatively low bias upon the left-hand sec-

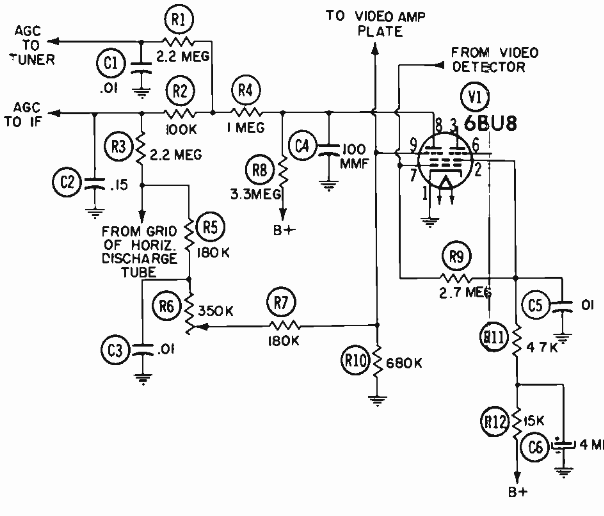


Fig. 16-4. Schematic of the 6BU8 AGC circuit.

tion of the 6BU8. This section of the tube then conducts heavily, and the voltage on the left-hand plate drops. The lower the plate voltage becomes, the more AGC voltage is produced.

Since the plate is maintained at a positive potential of approximately 35 volts in order that the tube will conduct, the AGC voltage cannot be obtained directly from the plate. Instead, it is derived in the following manner. The plate of the 6BU8 is connected to the grid of the horizontal-discharge tube through a voltage divider made up of resistors R3, R2, and R4. The average DC potential at the grid is −75 volts. Variations in this grid voltage are leveled off by the AGC filters. The voltages at the intermediate points on the divider can be applied directly to the AGC line. Note that the AGC bias voltages for the tuner and for the IF strip are taken from separate points. When no input signal is being applied to the AGC section of the 6BU8, there is a slight positive voltage at the junction of R2 and R4 and a slight negative voltage at the junction of R2 and R3. These voltages are fed to the tuner and IF stages, respectively. Both AGC voltages are driven in a negative direction when an input signal is applied to the 6BU8, but the IF voltage always remains more negative than the tuner voltage. This arrangement amounts to a simple delay circuit for the AGC line to the tuner.

The negative potential at the grid of the discharge tube is also utilized by another voltage divider. This voltage divider is in the grid circuit of the AGC section of the 6BU8 and is composed of R10, R7, AGC control R6, and R5. A negative DC voltage is fed from the junction of R10 and R7 to the grid of the 6BU8, and the input signal from the video amplifier is superimposed upon this DC voltage. The AGC control permits the range of bias on the tube to be adjusted.

If noise pulses in the AGC input signal were uncontrolled, they would cause the AGC section of the 6BU8 to conduct excessively. The voltages in the AGC system would then become too negative. The noise inverter (see discussion on page 95) keeps these unwanted pulses from increasing the conduction of the 6BU8 through the action of the inverted signal on pin 7. The AGC voltages are maintained at normal values even when considerable noise is present in the video signal.

The values of many of the components in the 6BU8 circuit are critical. This fact applies especially to the noise inverter. The bias on the inverter grid (pin 7) must be maintained at such a value that the sync tips will almost reach the cutoff voltage of the grid. If the bias is too low, the noise inverter will not accomplish its purpose. If the bias is too high, the 6BU8 will be driven into cutoff by each sync pulse. The latter condition would cause low AGC voltage and unstable synchronization.

No provision is made for adjusting the inverter circuit. This is actually an advantage because then the operation of the inverter cannot be upset by misadjustment of some control.

DELAYED AGC FOR RF STAGE

Even the weakest usable input signal develops some AGC voltage, which reduces the gain of all stages controlled by the AGC system. Unfortunately, the amplitude of a weak signal at the grid of the RF amplifier is barely more than the amplitude of the noise in that stage. When reception is poor, full gain in the RF stage is essential, so that the signal-to-noise ratio can be kept high. Once the signal has been amplified above the level of the noise, it can be acted upon by AGC in the IF amplifier without bad effects.

Therefore, provisions are made for delaying the action of the AGC bias on the RF amplifier in most receivers employing keyed AGC systems. The delay circuit includes a connection through a high resistance to B+. When a weak signal is being received, the tuner bias is sharply reduced because of this B+

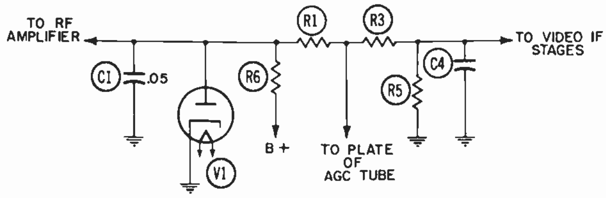


Fig. 16-5. AGC delay circuit.

connection. Included in most delay circuits is a clamper diode which conducts and shorts the AGC line to ground whenever the bias voltage tends to become positive. The circuit in Fig. 16-5 has this delay feature in the RF branch of the AGC line. The components included in the delay circuit are resistors R1 and R6, and clamper diode V1.

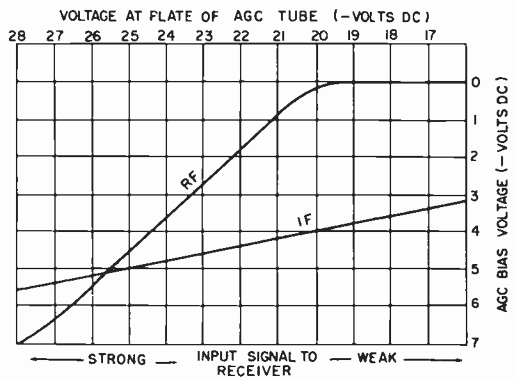


Fig. 16-6. Graph of the RF and IF bias voltages when an AGC delay circuit is employed.

As shown in the graph of Fig. 16-6, the delay circuit holds the AGC bias applied to the RF amplifier to about zero until the incoming signal is strong enough to develop −4 volts of bias in the IF section of the AGC line. The RF bias appears at this signal level, increases more rapidly, and eventually becomes greater, than the IF bias. When the incoming signal is strongest, the RF amplifier is biased most heavily, and the signal is promptly reduced before it has a chance to overload any of the IF amplifiers.

QUESTIONS

1. What is the purpose of an automatic gain control?
2. What is the difference between rectified AGC and amplified AGC?
3. How many signals are required in keyed AGC? What are these signals?
4. What determines the grid voltage on the AGC keying tube during the burst of conduction?
5. What is the purpose of the resistor in the grid circuit of the keying tube?
6. In the AGC circuit that employs a 6BU8, why isn't the AGC voltage taken directly from the plate? How is the AGC voltage obtained in this case?
7. What is the purpose of the clamper diode in the AGC delay circuit?

EXERCISES

1. Draw the basic circuit of a keyed AGC system. Show the input and output signals.
2. Explain the operation of the circuit in Exercise 1.

Chapter 17

Receiver Controls – Application and Adjustment

Preceding each day's programing, most stations transmit a test pattern which usually carries the station's call letters or distinguishing insignia. A test pattern is frequently transmitted when a regular broadcast is interrupted. By using the test pattern, the technician can properly adjust the controls in the receiver; the owner also can adjust the front-panel controls for the best picture. The horizontal- and vertical-hold, centering, linearity, and focus controls can all be precisely adjusted while the test pattern is being viewed. We will now review these controls and show their effect upon the received test pattern.

Fig. 17-1 shows a typical transmitted test pattern on the screen of a correctly adjusted television receiver. The pattern is similar to a more complex test chart (Fig. 17-2) developed by the television transmitter committee of the Electronic Industries Association. This standard resolution chart is used to test the performance of television transmitters as well as receivers. A study of its features will explain the use of the less complicated test pattern in Fig. 17-1.

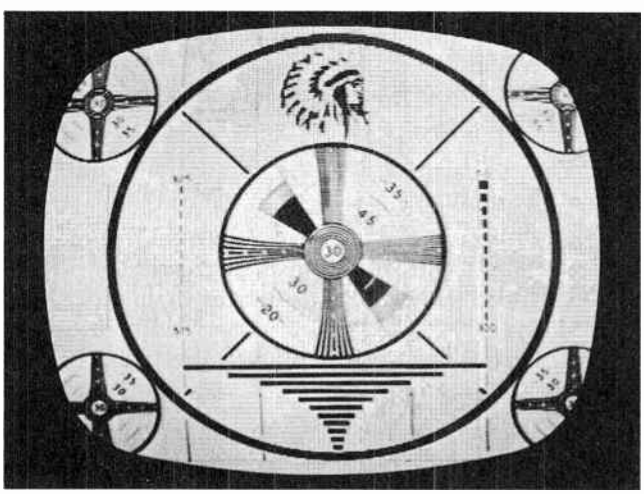


Fig. 17-1. Normal test pattern as it appears on a receiver that is operating normally.

Fig. 17-2 shows the television resolution chart, with added explanatory letter symbols. The chart consists of a series of geometric forms and a number of tones ranging from black through gray to white. The gray scales determine whether all elements of the television system are preserving the correct ratios of light intensities (as video modulation) to accurately reproduce the televised scene at the picture tube. The circles are used in checking horizontal and vertical size and linearity in the picture.

The horizontal and vertical fan-shaped wedges are composed of lines which gradually narrow as they approach the center. An estimate of the "resolving power" of the television system, including the receiver under test, can be determined by observing the point at which the lines can no longer be distinguished from one another. Numbers beside the horizontal and vertical wedges show the corresponding number of lines being reproduced when the individual lines of the fan can just be distinguished from each other. The vertical fans are used to determine the horizontal resolution of the system; conversely, the horizontal fans are used to determine the vertical performance.

Tests for vertical and horizontal linearity and for such requirements as interlace are described in Fig. 17-2. We will discuss their application to the simplified pattern in Fig. 17-1 as we consider the effects on the reproduced pattern caused by maladjusted controls or malfunctioning circuits.

In our study of the circuits employed to produce the raster, the control of scanning by signal pulses, and the modulation and focusing of the cathode-ray beam, we covered the action of a number of variable controls. These controls can be classified into two distinct groups according to whether or not they are readily accessible by the owner.

FRONT-PANEL OR OWNER-OPERATED CONTROLS

The front-panel controls are on the outside (front, top, or side) of the cabinet and can be operated by

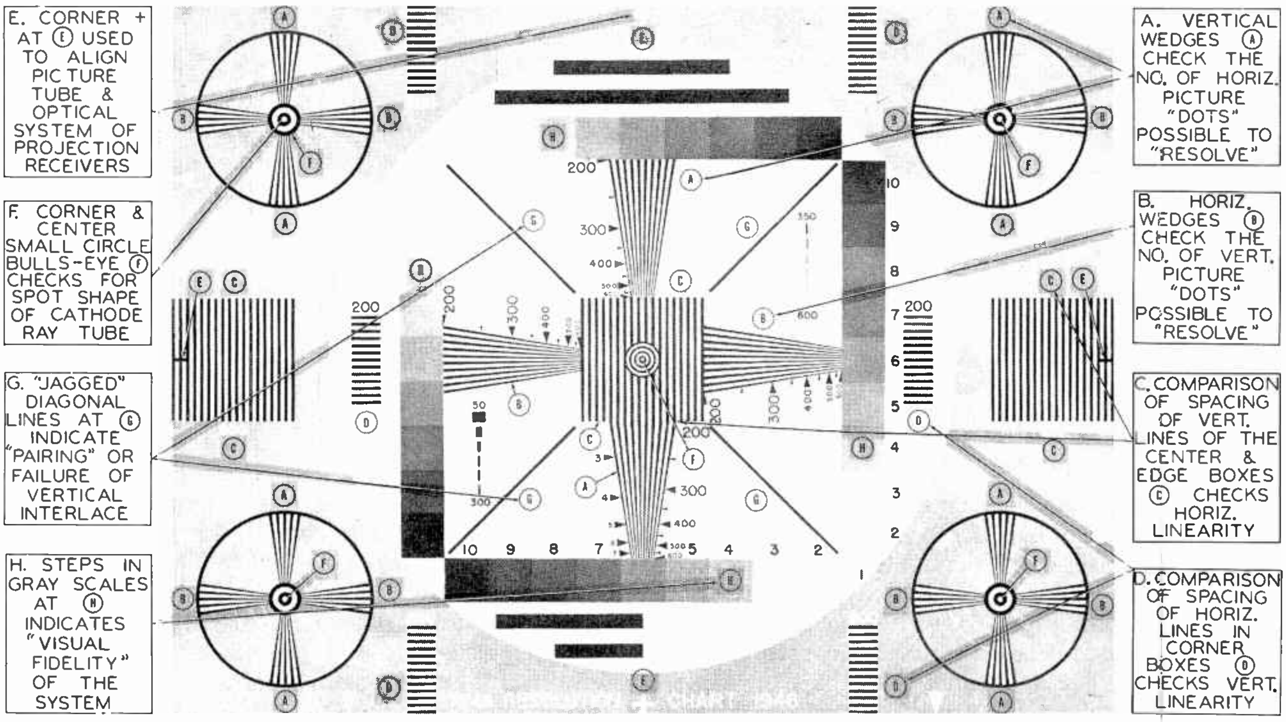


Fig. 17-2. A transmitter test chart.

the owner. They normally include the volume control which also turns the receiver on and off, a channel selector for tuning to the desired channel, and a group of controls for adjusting the appearance of the picture. The picture-control group lets the owner adjust brightness, contrast, and stability of the picture. Fig. 17-3 shows the front-panel controls on various receivers. Notice that concentric knobs are often used.

PRESET CONTROLS

The preset controls require adjustment during original installation or at infrequent intervals only and are not readily accessible to the owner. Their number and complexity differ greatly among the manufacturers. The preset controls most likely to require readjustment can be reached without the chassis being removed. They are adjusted by a knob or screwdriver. Fig. 17-4 shows the location of many of these controls.

CLASSIFICATION ACCORDING TO FUNCTION

The controls of a television receiver can also be classified into four groups according to their functions:

1. Those which adjust the operating conditions of the cathode-ray picture tube.

   a. Adjustment of ion-trap position—directs the beam through the gun to the screen.

   b. Adjustment of the deflection yoke—positions the raster correctly.

   c. Focus—for sharp definition.

   d. Adjustment of picture-tube operating voltages—for proper contrast between black level and highlight brightness.

   e. Scene brightness.

2. Those which establish the correct lock-in, or hold, of the horizontal- and vertical-scanning oscillators:

   a. Horizontal hold — sets the free-running frequency of the horizontal-scanning oscillator.

   b. Vertical hold — sets the free-running frequency of the vertical-scanning oscillator.

   c. Horizontal-oscillator frequency adjustment in AFC systems.

   d. Horizontal-discriminator phase control in AFC systems.

   e. Horizontal-waveform adjustment in pulse-width systems.

3. Those which adjust the dimensions and position of the picture:

   a. Width control—adjusts horizontal size.

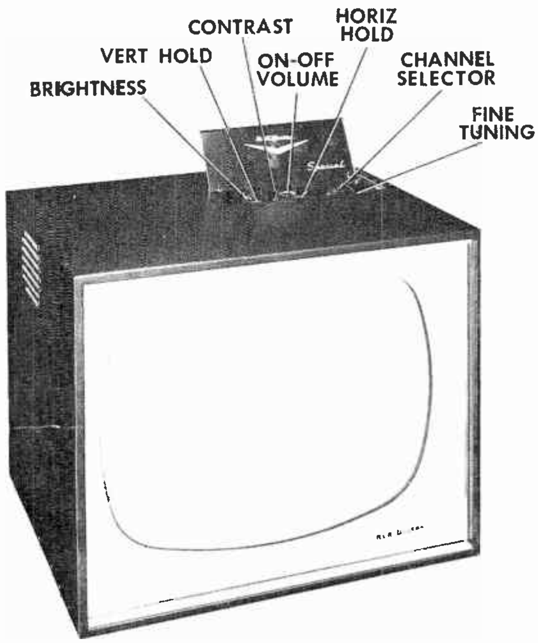
   b. Height control—adjusts vertical size.

   c. Horizontal centering—moves the picture horizontally.

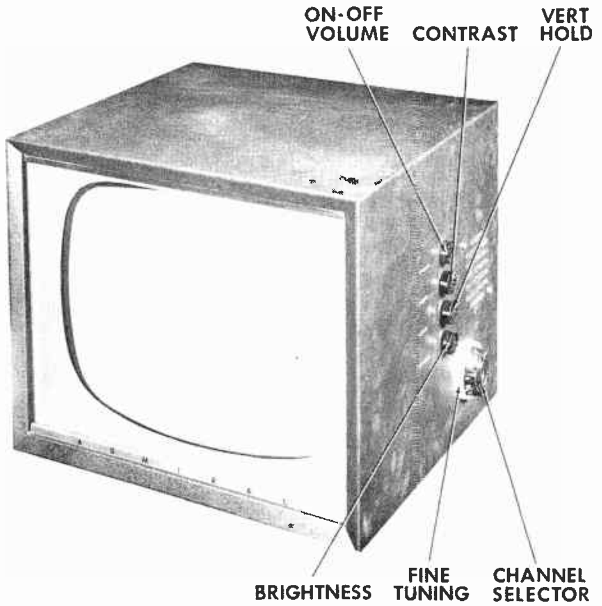
   d. Vertical centering—moves the picture vertically.

4. Those which determine the shape of the scanning-voltage waveforms to produce an undistorted picture:

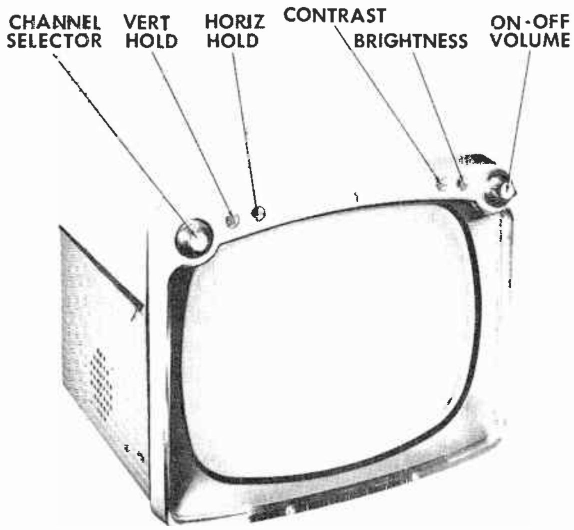
   a. Horizontal linearity — controls the shape of the horizontal-scanning wave.



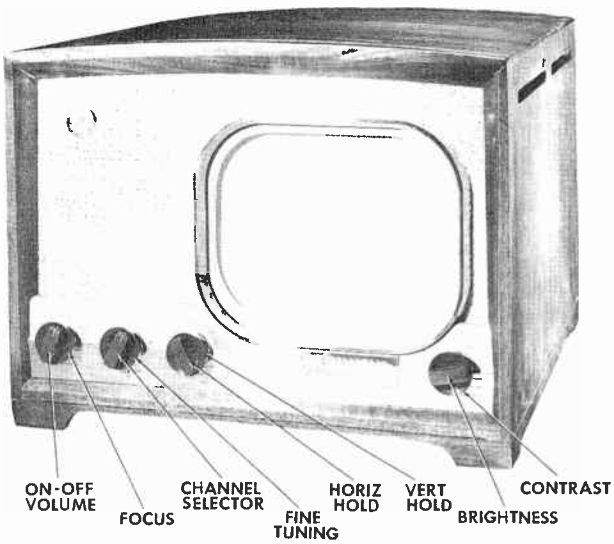
(A) Receiver employing seven panel controls on top of cabinet.



(B) Receiver employing six panel controls on side of cabinet.



(C) Receiver employing six panel controls on front of cabinet.



(D) Receiver employing eight panel controls on front of cabinet.

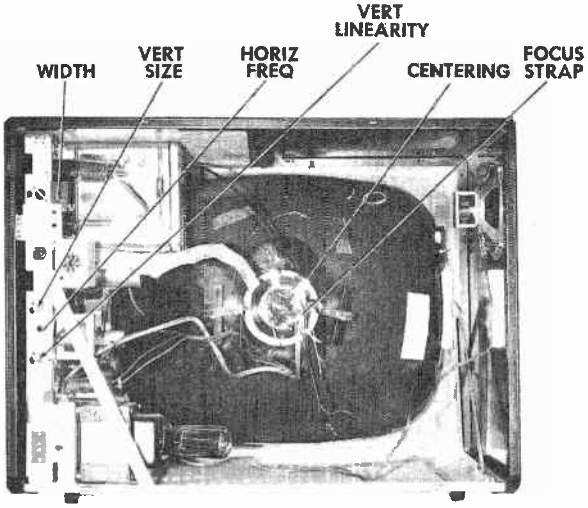
Fig. 17-3. Typical examples of panel controls.

b. Horizontal drive — determines the ratio of pulse to linear sawtooth for the voltage wave in magnetic deflection.

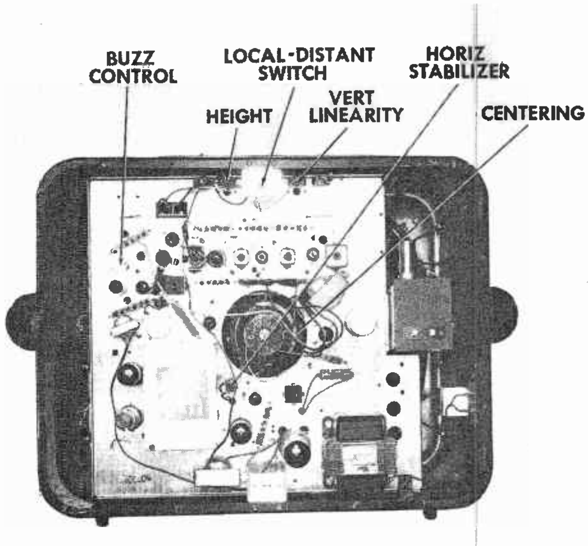
c. Vertical linearity—controls the shape of the vertical-scanning wave.

Focus-Device and Ion-Trap Adjustment

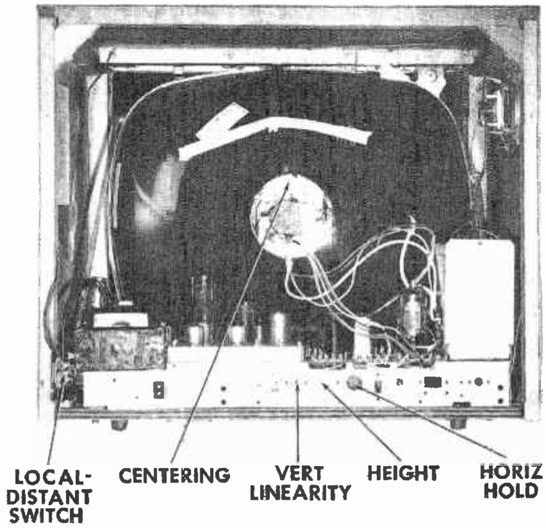
Fig. 17-5 shows the test pattern when the focus device and the ion trap are incorrectly positioned on the neck of the picture tube. The focus device and the ion trap operate interdependently, and an adjustment of one may require readjustment of the other. The shadowed corner and poor vertical positioning in Fig. 17-5 indicate that the electron beam is striking the neck of the tube. The focus device should be adjusted in its mounting until the picture is properly centered. No raster on the picture-tube screen may indicate an improperly positioned ion trap.



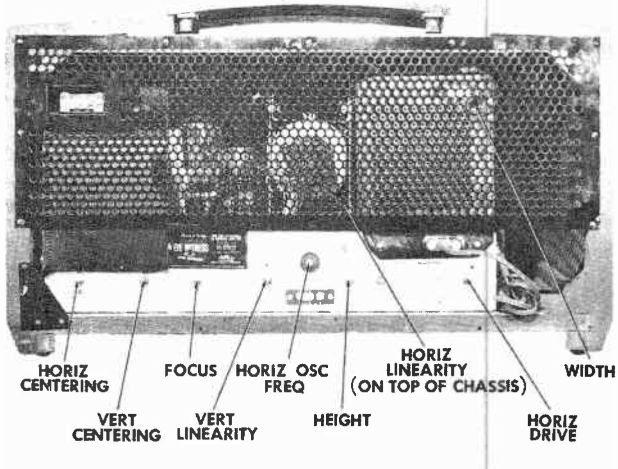
(A) Service controls located at back of chassis. Back cover is removable for servicing picture-tube controls.



(B) Receiver employing vertical chassis. Back cover must be removed to adjust controls, except for two controls accessible from top of cabinet.



(C) Service controls accessible from rear of chassis. Back cover is removable for servicing picture-tube controls.



(D) Service controls at back of the chassis. Controls on top chassis and picture tube are accessible by removing protective screen.

Fig. 17-4. Typical examples of preset controls.

Deflection-Yoke Adjustment

Fig. 17-6 shows the test pattern when the deflection yoke is improperly positioned. The control of the electron beam by the deflection coils was explained in Chapter 2. If the lines of the raster are not horizontal and squared with the edge of the picture mask, the deflection yoke is incorrectly positioned. The yoke adjustment screw is loosened and the yoke is rotated until the raster lines are properly aligned with the edges of the picture mask. The yoke-adjustment screw is then tightened. The position of the deflection yoke along the picture-tube neck will affect the deflection sensitivity (amount of scanning voltage for a given deflection).

Focus-Control and Focusing Adjustments

Fig. 17-7 illustrates the test pattern when the electron beam is out of focus. The image is not sharply defined, as it is in the normal picture in Fig. 17-1.

Electrostatically-focused picture tubes require several types of focusing methods. Some require a variable source of focus voltage, and a variable-

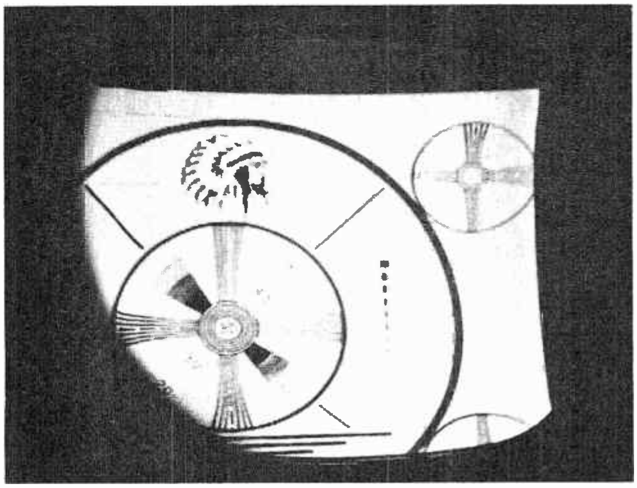


Fig. 17-5. Focus coil and ion trap misaligned.

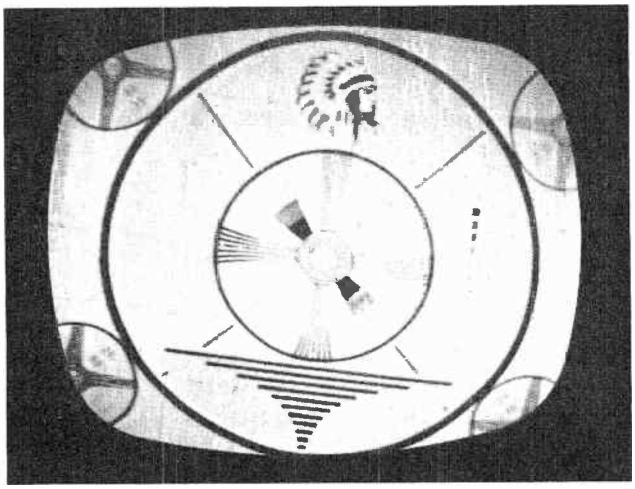


Fig. 17-6. Deflection yoke not properly adjusted.

resistance control is usually provided. The focus electrode is sometimes connected to a fixed voltage source. The voltage is varied by the lead or connection being moved to a different voltage point.

Some early electrostatically-focused tubes required focusing voltages of 3,000 to 4,000 volts. Because these tubes required a separate high-voltage supply, they were soon replaced by low-voltage tubes.

Magnetically-focused tubes can be focused by either a permanent magnet or an electromagnet. Either magnet requires two adjustments—position and fine focus. The fine focus and the position of the magnet are adjusted while the technician watches the test pattern for the sharpest picture.

Brightness Control

The brightness control normally is located on the front, side, or top of modern receivers and can be adjusted by the viewer. The beam can be cut off with this control, and the tube will remain dark.

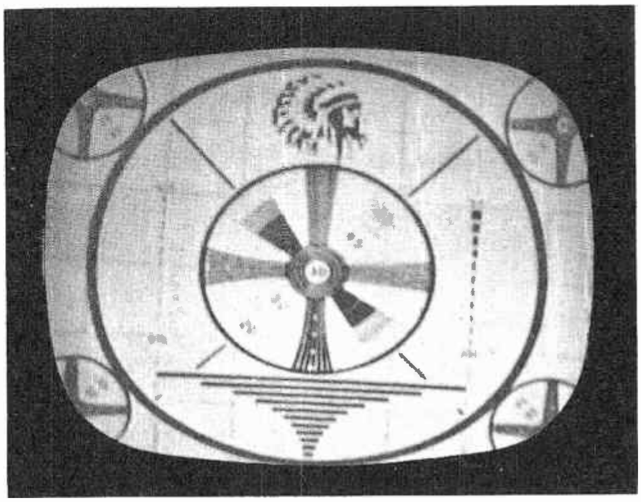


Fig. 17-7. Focus misadjusted.

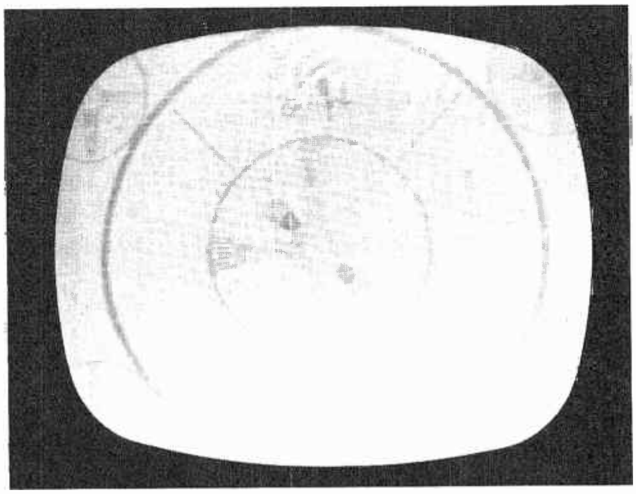


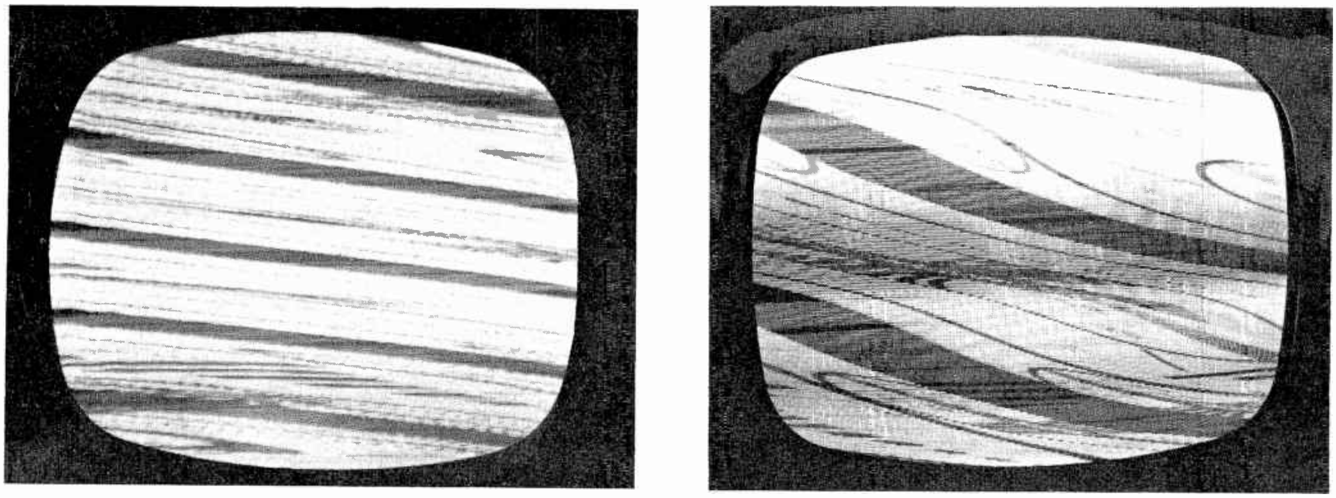
Fig. 17-8. Brightness control misadjusted. (Brightness too high.)

Setting the brightness control too high in older models causes a light, washed-out picture, as shown in Fig. 17-8. The shadows and lower-key halftones have disappeared, and the vertical-retrace lines have become visible. In modern receivers, the vertical-retrace lines are blanked out and more brightness can be attained before any loss in picture quality becomes noticeable. In the circuit, the brightness control establishes the control-grid bias of the picture tube.

Hold Adjustments

The horizontal- and vertical-hold adjustments enable the free-running frequencies of the two receiver scanning systems to be adjusted for synchronism, or lock-in, with the transmitted sync pulse.

In triggered sync systems, these controls are variable resistors in the scanning-oscillator circuit. In flywheel or AFC sync systems, the hold control is the same type, but is placed in the grid circuit of a



(A) Picture out of sync. (B) Closer to sync.

Fig. 17-9. Horizontal-hold control adjusted to bring picture into sync.

sine-wave oscillator. Once the picture is in sync, the hold adjustment controls the phase of the oscillator with respect to the sync pulses.

Horizontal Hold—Fig. 17-9 illustrates the effect on the test pattern when the picture is out of sync and the horizontal-hold control is adjusted to bring it back into sync. The diagonal bars become fewer and fewer until the picture snaps into synchronization. Proper hold-control setting is achieved when the picture does not go out of sync when the channel selector is switched off channel and then back on channel again.

Vertical Hold—Fig. 17-10 shows the test pattern when the vertical-hold control is misadjusted. The effects on the picture are similar to those discussed for the horizontal hold, except that the image moves vertically instead of horizontally before lock-in occurs.

The vertical hold must be carefully adjusted. Otherwise, "pairing" of horizontal lines of alternate fields will occur. Pairing reduces the vertical definition of the picture.

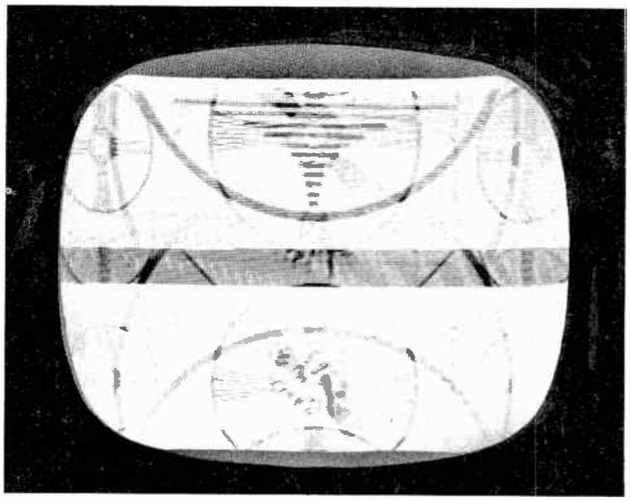


Fig. 17-10. Vertical-hold control misadjusted.

Horizontal-Oscillator Frequency Adjustment (AFC Systems)

Fig. 17-11 shows the test pattern when the horizontal-oscillator frequency is misadjusted. In the flywheel or AFC system of horizontal-sync control, the free-running frequency of the oscillator is controlled by an iron core in an inductor. The horizontal-hold

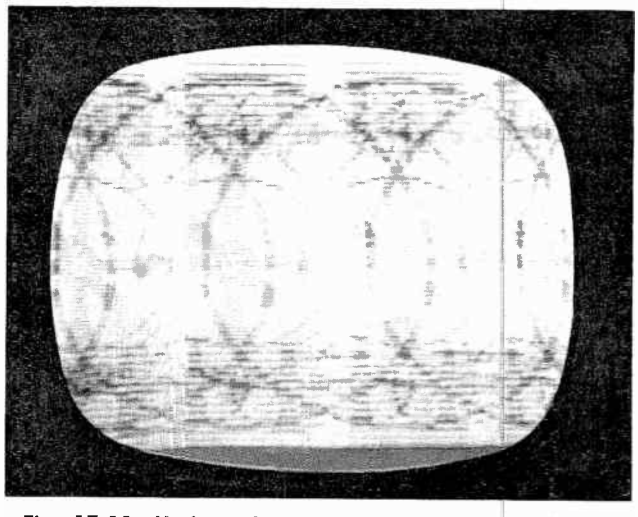


Fig. 17-11. Horizontal-oscillator frequency misadjusted. (AFC system.)

and the discriminator-phase controls also affect the oscillator frequency. For this reason, the adjustment of the horizontal-oscillator frequency must be rechecked if the discriminator-phase control is changed. Service literature contains explicit instructions concerning the order in which these adjustments are to be made in the particular AFC circuit design.

The picture should remain horizontally synchronized when the hold control is turned to either extreme. To test for this condition, tune to a signal while the control is in the middle. Then turn the hold control to its extreme position in either direction. Next, remove the signal by detuning the receiver.

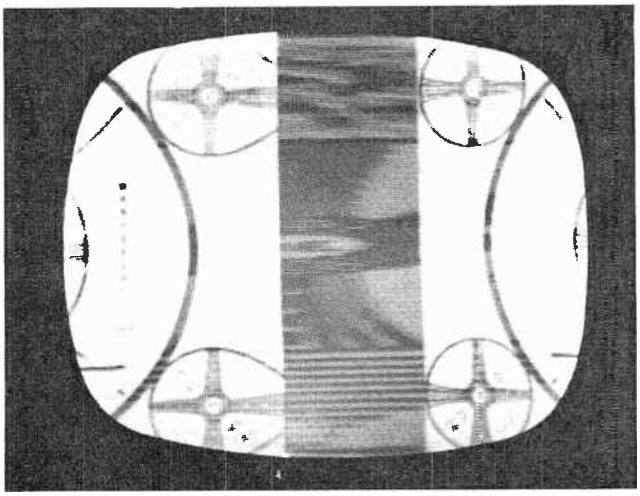


Fig. 17-12. Horizontal-discriminator phase misadjusted. (AFC system.)

When the receiver is retuned, the system should pull into sync. Make the same check at the other end of the control range. If the receiver does not pull into sync at both ends of the hold range, readjust the horizontal-oscillator frequency.

Horizontal-Discriminator Phase Adjustment

The discriminator stage compares the sync-pulse rate with the horizontal-oscillator frequency and delivers a DC voltage to control the oscillator. If the voltages impressed by the oscillator at the plates of the two diodes are not equal, the DC output will not be zero at the correct time for retrace. This off-balance condition will result in the test pattern

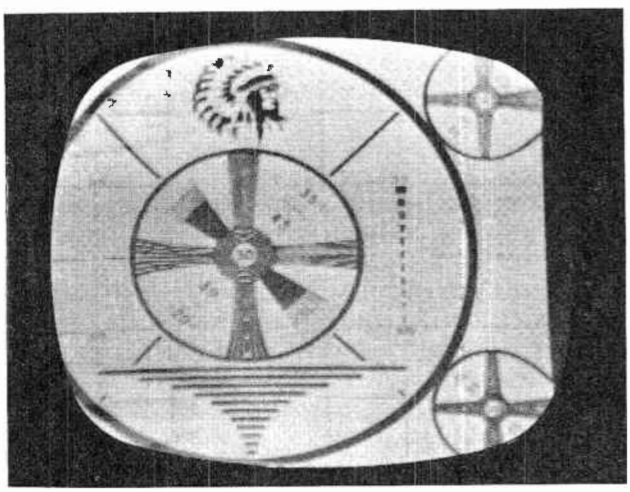


Fig. 17-13. Horizontal centering incorrect.

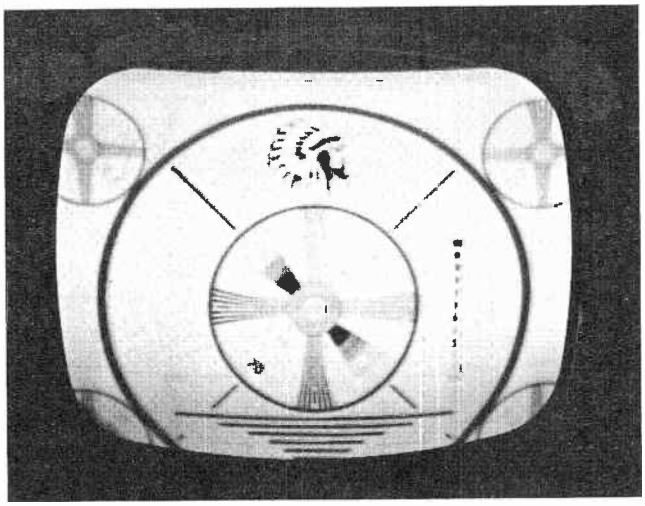


Fig. 17-14. Vertical centering incorrect.

shown in Fig. 17-12. The picture is locked in and steady, but retrace has occurred at the wrong time. The black band in the center of the picture is the blanking period during which horizontal retrace should have occurred.

The adjustment for retrace at the proper instant also affects the setting of the horizontal-oscillator frequency. The service technician should follow the instructions in the service literature for the proper sequence of these adjustments.

Centering Controls

The adjustments which center the picture are usually at the rear of the set. They are called the horizontal- and vertical-centering controls. In the early receivers, varying the DC voltage to the deflection plates of the electrostatic tube or the DC current through the deflection coils of the electromagnetic-deflection circuit was one method of centering the picture. Another method was by moving the focus coil.

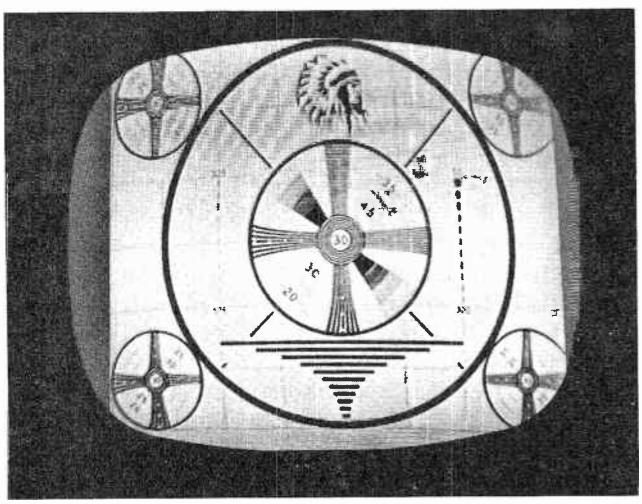


Fig. 17-15. Width control misadjusted.

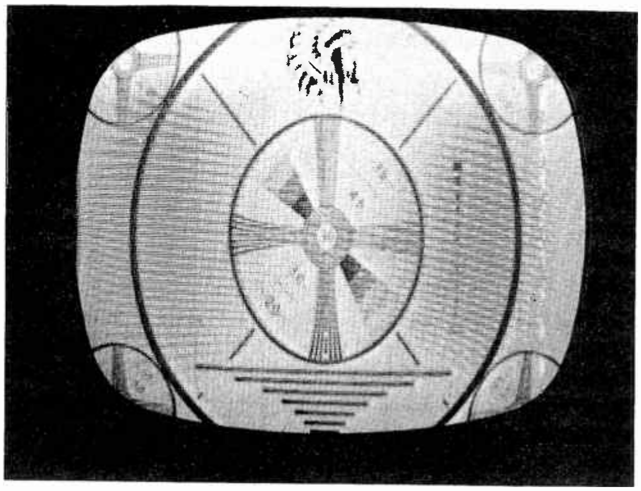


Fig. 17-16. Height control misadjusted.

Almost all modern receivers use two permanently-magnetized rings on the neck of the picture tube to accomplish centering. Fig. 17-13 shows the effect on a test pattern when the horizontal centering is misadjusted. An example of a picture which is not centered vertically is shown in Fig. 17-14.

Width Control

The width adjustment varies the energy applied to the horizontal-deflection plates or to the deflection coils. The effect on the test pattern when this control is not properly adjusted is shown in Fig. 17-15. Regulating the output of the horizontal oscillator, the discharge tube, or the horizontal-output amplifier controls the width.

Height Control

The height control functions similarly to the width control, but in the vertical direction. The vertical-oscillator or vertical-amplifier output is controlled. Fig. 17-16 shows the effect on the test pattern when the height control is incorrectly adjusted. The picture is too tall, but is symmetrical. Sometimes the picture is incorrect in size and is off center. The size controls (width and height) must be adjusted simultaneously with the horizontal- and vertical-centering controls.

Controls Which Affect Scanning Waveshape

In the discussion of scanning, we pointed out that the electron beam must trace at a linear rate both horizontally and vertically if picture distortion is to be avoided. The shape of the voltage wave (electrostatic) or current wave (electromagnetic) required to produce the linear beam motion was explained. In the discussion of beam-deflection systems, the various circuits and controls for achieving scanning linearity were discussed in detail.

The output voltage wave of the cathode-coupled multivibrator or blocking oscillator can be made linear enough for electrostatic deflection. For this reason, linearity controls are not required. For electromagnetic deflection, the scanning current consists of a combined linear sawtooth and pulse. The following variable controls are needed for proper time and amplitude relationships of the various sections of the waveshape: linearity controls (in early sets, as many as three for horizontal scanning) and drive controls (which adjust the amplitude of the pulse portion of the wave).

Horizontal Linearity—Fig. 17-17 shows the effect of misadjusted horizontal linearity on the test pattern. Note that the circle of the test pattern is now egg-shaped; the picture has been extended on the left side.

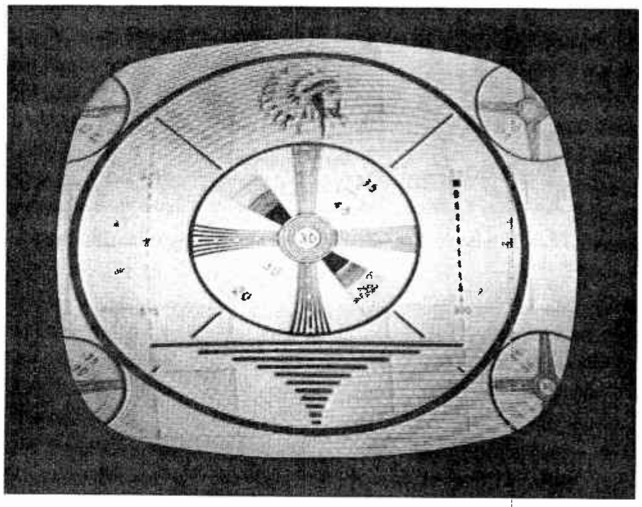


Fig. 17-17. Horizontal-linearity control misadjusted.

Since the number and circuit function of linearity controls differ among receiver designs, the technician is advised to study service literature for the particular linearity adjustment and for the interaction of the control adjustment with other linearity and width controls.

Horizontal Drive—The horizontal drive adjusts the ratio of the pulse to the linear portion of the horizontal-sawtooth, scanning-current wave. This action controls the point on the scanning trace at which the horizontal-output tube conducts. The effect of drive misadjustment is shown in Fig. 17-18. Feedback is sometimes used in the output stage to provide a negative pulse from the horizontal-output transformer to the grid of the output tube. In this case, the drive control is a voltage divider across a winding on the output transformer. The voltage pulse is fed back in series with the output-tube grid return.

Vertical-Linearity Control—Fig. 17-19 shows the effect of a misadjusted vertical-linearity control on the test pattern. The vertical-linearity control is usually an adjustable cathode-bias resistor, which adjusts the operating point of the vertical amplifier.

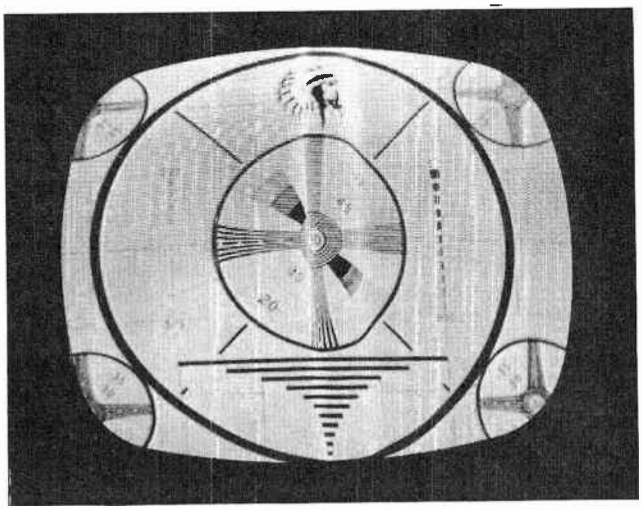


Fig. 17-18. Horizontal drive misadjusted.

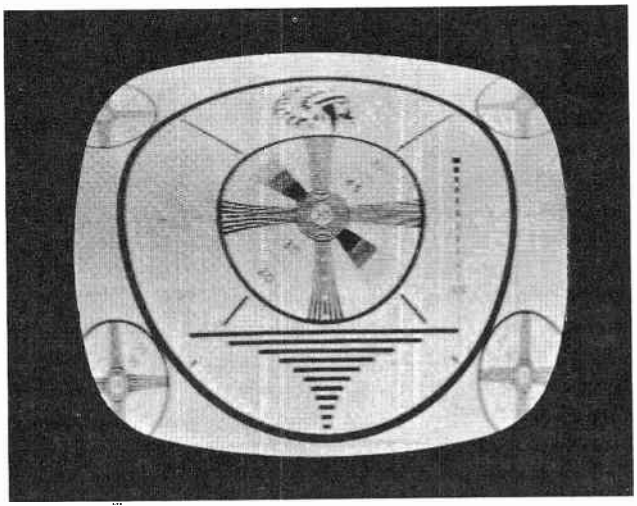


Fig. 17-19. Vertical-linearity control misadjusted.

In early receivers with more than one linearity control, the additional control was a variable resistor in the peaking circuit, which constituted the plate load of the vertical-discharge tube. This control functioned much like a horizontal-drive control.

The linearity adjustment in either the horizontal or the vertical circuit is interdependent with the width or height controls. Because of this interlocking action, one control may have to be readjusted if the setting of the other is changed.

QUESTIONS

1. What can be determined from the horizontal wedges of a test pattern? From the vertical wedges?
2. What are the purposes of the circles in the test pattern?
3. What adjustments are necessary when the raster is tilted, the lines are blurred, and a black shadow is at a corner of the raster?
4. When the circle in the test pattern is oblong (egg-shaped) in the horizontal direction, what controls need adjustment?
5. When there is a narrow white vertical line (foldover) near the center of the picture, what control needs adjustment?
6. Name the controls that are adjusted for defects in the picture in the vertical direction.

Appendix

Glossary of Terms

A

Absorption Trap. A parallel tuned circuit coupled either magnetically or capacitively to absorb and attenuate interfering signals.

Accompanying Audio (Sound) Channel. The RF carrier frequency which supplies the sound that accompanies the picture. Also known as co-channel sound frequency.

Active Lines. The lines which produce the picture, as distinguished from the lines occurring during blanking (horizontal- and vertical-retrace lines).

Adjacent Audio (Sound) Channel. The RF carrier frequency which carries the sound modulation associated with the next lower-frequency television channel.

Aluminized-Screen Picture Tube. A picture tube in which a thin layer of aluminum has been deposited on the back of the fluorescent surface. This layer improves the brilliance of the image and prevents ion-spot formation.

Amplitude Modulation (AM). A method by which an RF carrier is modulated. The instantaneous amplitude of the carrier is proportional to the instantaneous value of the modulating voltage.

Amplitude Separation. The method by which part of a voltage wave (particularly the sync pulse) is sorted from the rest of the wave by differences in amplitude. A bias clipper is usually employed.

Antenna Array. A system of antennas which are coupled together to produce a desired directional pattern or to increase signal pickup.

Aquadag Coating. A conductive coating on the surface of the glass envelope of picture tubes. The coating is formed by a colloidal solution of carbon particles. It is placed internally to collect secondary electrons emitted by the fluorescent screen. In some tubes it is also placed externally so that the tube can be used as a final capacitor of the high-voltage filter circuit.

Aspect Ratio. The ratio of picture width to picture height. The television standard is 4/3.

Asymetrical-Sideband Transmission. (See Vestigial-Sideband Transmission.)

Audio Channel. (See Accompanying Audio Channel.)

Automatic Gain Control (AGC). A method by which the over-all amplification (gain) of a television receiver is automatically regulated, so that relatively constant output (contrast of the picture) is produced for varying input signals.

Automatic Contrast Control. (See Automatic Gain Control.)

Automatic Phase Control. A method by which the frequency and phase of the horizontal-scanning oscillator are automatically held in step with the repetition rate and phase of the horizontal-sync pulses.

B

Background. In television, background is the average illumination of the scene and is represented by the DC component of the video signal. (Also see DC Video Component.)

Back Porch. The portion of the synchronizing signal (at blanking or black level) which follows the horizontal-sync pulse and precedes the start of the next horizontal active line. (Standard duration for this porch is 3.81 microseconds.)

Bandwidth. The difference in frequency between the highest and lowest frequencies involved. A television channel has a bandwidth of 6 megacycles.

Beam. In a cathode-ray tube, the stream of electrons moving toward the screen.

Bidirectional. The shape of the reception pattern of an antenna that responds equally well to stations located 180° apart with respect to the receiver.

Blacker-Than-Black (Infrablack). That portion of the composite signal that is higher in amplitude than the black level (from 75% to 100% of maximum signal). This region is occupied by the synchronizing pulses.

Black Level. The amplitude of the composite signal at which the beam of the picture tube is extinguished (becomes black) to blank retrace of the beam. This level is established at 75% of the signal amplitude.

Blanking Pulses. The portion of the composite signal during which the beam of the picture tube is extinguished. (See Black Level.)

Blocking Oscillator. A type of relaxation oscillator for the generation of sawtooth waves. It employs an iron-core feedback transformer and a long time-constant capacitor-resistor network in the grid circuit.

Blooming. The phenomenon of the picture being defocused because of excessive brilliance. The spot size is enlarged to halation (spreading of light at the fluorescent screen).

Booster. A separate RF amplifier connected between the antenna and the television receiver to amplify weak signals.

Brightness Control. This receiver control sets the operating point of the picture tube, so that beam retrace is not visible.

C

Camera Tube. The device at the television transmitter that changes the light variations of the scene into electrical variations.

Cascode Circuit. A circuit which consists of a grounded-cathode triode driving a grounded-grid triode. Both triodes are connected in series, as far as plate current is concerned.

Cathode Follower. A circuit in which the input signal is applied between control grid and ground and the output is taken between cathode and ground. The plate is bypassed to ground. A cathode follower exhibits high input impedance and low output impedance.

Cathode Input. (See Grounded-Grid Amplifier.)

Centering Control. Controls the position of the picture-tube screen. (See Horizontal-Centering Control and Vertical-Centering Control.)

Channel. The band of frequencies assigned for the transmission of a television signal.

Characteristic Impedance. The ratio of the voltage to the current at every point on an RF-transmission line. This ratio does not apply unless the line is properly terminated and there are no standing waves.

Clamping Circuit. A circuit which maintains the amplitude of a voltage wave at a predetermined DC level. (Also see DC Restoring.)

Clipping Circuit. A circuit that removes one or both extremities of an impressed voltage wave. This produces a flat-topped output.

Coaxial Cable (Coax). A high-frequency transmission cable consisting of a central inner conductor and a cylindrical outer conductor, the two being separated by an insulator. Coax is often employed as a television antenna lead-in.

Composite Signal. The transmitted television signal, which is composed of video modulation, blanking pedestals, and synchronizing pulses. Blanking pedestals occur at 75% of maximum signal, and the sync pulses occupy the remaining 25%.

Contrast. The range of light-to-dark values of the image, proportional to the voltage swing at the picture-tube input.

Contrast Control. Controls the voltage swing to the picture-tube input, so that the most acceptable image contrast is produced.

D

DC Restoring. The combining of the DC component of the video signal (lost in capacitance-coupled amplifiers) with the AC component, so that the average light value of the reproduced picture is re-established.

DC Video Component. The part of the video signal caused by the average steady background illumination of the scene being televised.

Damping Tube. A vacuum tube (diode or triode) that momentarily short-circuits the stored energy in a deflecting coil system and thus prevents transient oscillations.

Definition. The sharpness of the fine detail in the reproduced picture.

Deflection. The process whereby the electron beam of a picture tube is deviated from its axial path (either electrostatically or electromagnetically) to produce the raster.

Deflection Yoke. An assembly consisting of the horizontal- and vertical-deflection coils of a magnetically-deflected receiver.

Detail. The least perceptible elements or areas of an image which can be recognized as being different from one another.

Diathermy. An interfering signal caused by certain types of medical appartus. The interference usually produce a "herringbone" pattern on the picture.

Differentiating Circuit. A combination of circuit elements which produces an output pulse proportional to the rate of change of the input signal.

Dipole. An antenna whose length is approximately one-half the electromagnetic wavelength to which it is resonant. The antenna is usually divided in the middle for connection to the transmission line.

Director. An antenna element placed in front of a dipole element (toward the transmitter) to increase the pickup and obtain a directional pattern.

Discharge Tube. A tube which discharges a wave-shaping capacitor. The tube is biased to cut off until triggered by a positive pulse.

Discriminator. An FM-detector circuit using a pair of diodes. The audio output is proportional to the frequency deviation.

Dissector. A camera tube in which an electron image is moved across the aperture of an electron multiplier.

Drive Control. (See Horizontal-Drive Control.)

Driven Element. The antenna element that is connected to the transmission line.

Dynode. In a photomultiplier tube, an additional electrode which produces secondary emission when bombarded by electrons.

E

Echo. A delayed signal. This may refer either to a reflected television carrier (ghost signal) or to effects in the video amplifier.

Electron Focus. The reduction in size of the electron beam in a picture tube by variation of an electrostatic or electromagnetic field.

Electromagnetic-Deflection Coil. A circular coil placed near the neck of a picture tube, producing a magnetic field which deflects the electron beam.

Electron. The elementary electrical charge of negative polarity (1.6×10^{-19} coulombs).

Electron Gun. An arrangement of electrodes which produces and controls a small beam of electrons in a cathode-ray tube.

Electron Multiplier. An electron tube in which a number of electrodes are arranged in cascade. Each electrode delivers more electrons to the next electrode than it receives. The increase is due to secondary emission.

Electron Scanning. The deflection of an electron beam to form a regular pattern or raster.

Electrostatic Field. A strain in space which exerts a force on an electrical charge (electron) within its region of influence.

Electrostatic Scanning. The deflection of an electron beam by means of an electrostatic field.

Equalizing Pulses. A series of pulses (usually six) which occur at twice line frequency and which precede and follow the serrated-vertical synchronizing pulse. These pulses cause vertical retrace at the correct instant for proper interlace.

F

Field. One complete scanning operation of the picture from top to bottom, including vertical retrace. This scanning takes 1/60th of a second and occurs twice per frame.

Field Frequency. The number of fields transmitted per second. Sixty fields per second is standard.

Field-Frequency Synchronizing Pulse. (See Vertical Blanking.)

Field Period. The time required for one field to be transmitted. This period is the reciprocal of the field frequency, or 1/60th of a second.

Fluorescent Screen. The phosphor coating on the inside of the picture-tube faceplate, which emits light under the impact of the electron beam.

Flyback. The return of the electron beam after tracing a horizontal line. (See Retrace.)

Flywheel Sync. A method of horizontal synchronization in which scanning is controlled by the average timing of the sync pulses rather than by each individual pulse, as in triggered sync.

Focusing Control. Brings the electron beam to the smallest possible spot on the fluorescent screen.

Folded Dipole. An antenna consisting of two half-wave dipoles parallel to each other and having their extremities connected. One of the dipoles is open at its center point for connection to the transmission line.

Frame. The complete single picture contained in an image. A frame consists of two fields and has a repetition rate of 30 frames per second.

Frame Frequency. The number of times per second the picture is completely scanned—the standard is 30 frames per second.

Front Porch. The portion of the synchronizing signal (at blanking or black level) which precedes the horizontal-sync pulse and which occurs at the end of each active horizontal line (1.27 microseconds is standard duration for the front porch).

Front-to-Back Ratio. The antenna sensitivity ratio of signals arriving from the front (desired direction) to the signals arriving from the back (180° from the front).

G

Ghost Image. A second, or echo, image superimposed on the picture because of a reflected carrier wave.

Grid Limiting. Grid-current bias derived from the signal, through a large series grid resistor, which cuts off the plate current and thus levels the output wave for all input signals above a critical value.

Grounded-Grid Amplifier. A circuit in which the input signal is applied between a grounded grid and the cathode. The grid thus acts as a shield between the input circuit (cathode-to-ground) and the output circuit (plate-to-ground).

H

Halation. A halo surrounding a point of high brilliance on the fluorescent screen. The halo may be due to light scattering in the phosphor or to multiple reflections between the front and back surfaces of the glass.

Half-Wave Dipole or Doublet. (See Dipole.)

Height Control. Controls the amplitude of the vertical output and hence the height of the picture.

Hold Control. Controls the scanning-oscillator phase and frequency (horizontal or vertical), so that it is synchronized with the transmitted sync signal.

Horizontal Blanking. The process of cutting off the electron beam during retrace between successive active horizontal lines.

Horizontal-Blanking Pulse. The rectangular pedestal of the composite television signal, which occurs between active horizontal lines. This cuts off the beam current of the picture tube during horizontal retrace.

Horizontal-Centering Control. Moves the image horizontally on the picture tube screen.

Horizontal Directivity. In a receiving antenna, the reception characteristic of the antenna in the horizontal plane.

Horizontal-Discharge Tube. In the horizontal-deflection circuit, a vacuum tube which discharges the sawtooth-forming capacitor. (See Discharge Tube.)

Horizontal-Drive Control. Controls the ratio of the negative pulse amplitude to the linear portion of the scanning-voltage wave.

Horizontal Flyback. (See Retrace.)

Horizontal-Hold Control. (See Hold Control.)

Horizontal Polarization. A transmitted signal is horizontally polarized when its electrostatic component is parallel to the earth's surface.

Horizontal Repetition Rate. The number of horizontal lines per second (15,750).

Horizontal Resolution. The number of horizontal picture elements which can be distinguished from each other in a single active line.

Horizontal Retrace. (See Retrace.)

Horizontal-Scanning Frequency. (See Horizontal Repetition Rate.)

Horizontal-Sync Discriminator. In the flywheel method of synchronization, a circuit which compares the phase of the horizontal-sync pulses with that of the horizontal-scanning oscillator. The output of the discriminator controls the frequency and phase of the oscillator by means of a reactance tube.

Horizontal-Sync Pulse. The rectangular pulse which occurs, above blanking level, between each active horizontal line. Because of these pulses, horizontal scanning of the receiver coincides with that of the transmitter.

Hum Bar. A dark band extending across the picture, caused by excessive 60-cycle hum (or harmonics thereof) in the input signal of the picture tube.

I

Iconoscope. A camera tube in which an optical image on a mosaic of photosensitive elements causes charges which are proportional to the image intensity at each point. A scanning beam releases these charges as a video signal.

Image Dissector. (See Dissector.)

Image Interference. A spurious response in a television receiver due to signals at a frequency twice the intermediate frequency. Usually due to an FM carrier or another television carrier.

Image Orthicon. A camera tube combining the orthicon principle with that of the electron multiplier. This tube is extremely sensitive to light.

Implode. The term applied to the bursting of a picture tube. Because of the high vacuum in the tube, the glass fragments move inward with terrific force.

Infrablack. (See Blacker-than-Black.)

Integrating Circuit. A combination of circuit elements which produces an output potential proportional to the stored-up value of a number of pulses of input signal.

Intercarrier Sound. A method by which the 4.5-mc beat between video and sound carriers is used as an intermediate frequency for the sound signal.

Interlace Scanning. A method by which an image is scanned in several successive fields. Each field contains part of the horizontal line structure only, the fields being arranged so that successive fields supply the lines missed by the preceding fields. The lines of the second field fall exactly between the lines of the first. This system overcomes the flicker which would occur if areas on the screen were illuminated at a rate of only 30 cps.

Ion. A charged atom. In the picture tube, ions produce a dark spot at the center of the fluorescent screen unless they are blocked out of the electron beam.

Ion Spot. A dark spot on the fluorescent screen of a picture tube, caused by ion bombardment.

Ion Trap. An arrangement of picture-tube electrodes whose external magnetic field passes the electrons but blocks the ions.

K

Keyed AGC. An AGC circuit in which a pulse from the high-voltage transformer is applied to the plate, so that the stage can conduct during horizontal sync time only.

Keystone. A trapezoid. A television picture with one side shorter than the other has keystone distortion.

Keystone Distortion. (See Keystone.)

L

Limiter. A stage preceding an FM detector. All input signals above a predetermined amplitude drive this stage to cutoff and thus limit the value of the output signal. The limiter removes amplitude modulation.

Line Frequency. (See Horizontal Repetition Rate.)

Line-Frequency Blanking Pulse. (See Horizontal-Blanking Pulse.)

Linearity. The distribution of picture elements over the image field, as determined by the shape of the horizontal- and vertical-scanning waves. In a linear picture, the elements are uniformly and correctly distributed. If the scanning motion is nonlinear, the picture will be distorted.

Linearity Control. Corrects distortion of the sawtooth voltage or current waves used for deflection. Linearity controls are employed for both horizontal and vertical correction.

M

Magnetic Deflection. The method by which a magnetic field, produced by a coil external to the picture tube, moves the electron beam. A linear sawtooth motion is produced when the current through the coil has a sawtooth form.

Marker Pip. A frequency index mark used with a sweep generator during alignment of TV receivers.

The marker pip is produced by a fixed-frequency oscillator which is coupled to the output of the sweep generator.

Magnetic Field. The space near a magnet or electromagnet in which a force is exerted upon an electron.

Monitor. A picture tube and its associated circuits which are usually connected directly to the video system of the transmitter and which permit the transmitted picture to be viewed.

Monoscope. A cathode-ray tube which produces a video signal for testing purposes. An internal test pattern is employed.

Mosaic. A photosensitive surface in some camera tubes and consisting of an insulating surface covered with numerous photosensitive "islands." A plate behind the insulator collects the charges when the mosaic is scanned by an electron beam.

Multipath Reception. The reception of a direct wave from the television transmitter, accompanied by one or more reflected and delayed waves. (See Ghost Image.)

Multiplier. (See Electron Multiplier.)

Multivibrator. A relaxation oscillator which produces sawtooth waves. Two tubes are used; the output of each tube is coupled to the input of the other through R-C networks, which determine the period of oscillation.

N

Negative Transmission. The polarity of modulation of the standard television signal. The sync pulses and signals corresponding to black drive the carrier toward maximum amplitude; signals of highest brilliance (white) drive the carrier toward zero amplitude.

Neutrode Circuit. A neutralized triode amplifier in which a capacitor feeds back voltage to the control grid from the low side of the plate coil.

Noise. Interference that causes a salt-and-pepper or "snowy" pattern on the picture. It is due to atmospherics, tube-fluctuation effects, or man-made interference.

Nonlinearity. The crowding of picture elements at the sides, top or bottom of the picture because scanning is no longer linear. (See Linearity.)

O

Odd-Line Interlace. A double interlace system in which there is an odd number of lines per frame. Therefore, each field contains a half line.

P

Pairing. Improper interlace, in which the lines of alternate fields do not fall exactly between those of the preceding field. When pairing is most pronounced, the lines of alternate fields fall on one another and result in separated lines with half the possible vertical resolution.

Parasitic Element. An antenna element not coupled to the transmission line.

Peaking Coil. In a video amplifier, an inductance which resonates with the circuit capacitance near the upper limit of the passband. Thus, high-frequency loss of gain is compensated for, and the amplifier phase shift is corrected.

Pedestal. The level of the video signal at which blanking of the picture-tube beam occurs. (See Blanking Pulses and Black Level.)

Persistence of Vision. A characteristic of the eye and brain whereby the sensation of an image remains after the light causing it has vanished. This effect lasts for approximately an eighth of a second and makes television as well as motion pictures possible.

Phase. A point in the cycle of an alternating current or voltage with respect to zero time.

Phase Distortion. A condition of different phase delays for different video frequencies. This action causes distortion of the peak values of the video signal and results in poor contrast and resolution.

Photoelectric. The phenomenon whereby electrons are emitted from certain substances by absorption of light.

Pickup Tube. Same as Camera Tube.

Picture Element. The smallest of picture areas that can be distinguished from each other.

Picture Tube. In television receivers, a cathode-ray tube which translates the video signal into a picture.

Polarization. The direction (either horizontal or vertical) of the electric field of a radiated wave. The magnetic field is perpendicular to the electric field.

Pre-emphasis. The practice in which the high frequencies of the audio spectrum are amplified more than the low frequencies. Employed in FM transmission of the television sound channel.

Q

Q. The figure of merit of a capacitor, an inductor, or a tuned circuit. Q is equal to the ratio of the reactance to the resistance.

Quasi-Optical. Having properties similar to light waves. Thus, the propagation of waves in the television spectrum is said to be "quasi-optical," i.e., cut off by the horizon.

Quasi-Single Sideband Transmission. Same as Vestigial-Sideband Transmission.

R

R-C Circuit. A time-determining network of resistors and capacitors in which the time constant is defined as the product of the resistance times the capacitance.

Ratio Detector. An FM detector which discriminates against amplitude modulation. A pair of diodes is so connected that the audio output is proportional to the ratio of the FM voltages applied to the two diodes.

Reactance-Tube Circuit. A circuit in which a high transconductance tube is so connected that it appears as a reactance (inductive or capacitive) to a circuit across which it is connected. The value of the reactance can be controlled by changes in the DC grid bias. Used for AFC.

Reflection. The reflected carrier wave and the ghost image on the picture, caused by the reflected carrier.

Reflector. An element placed behind the pickup element of a receiving antenna to intensify the received signal and improve the shape of the directional pattern.

Registry. The superposition of one image on another, as in the formation of an interlaced raster.

Reinserter. (See DC Restoring.)

Relaxation Oscillator. An oscillator which generates periodic waves in which a sudden excursion of plate current from cutoff to saturation and back is followed by a relatively long period of quiescence, or relaxation.

Resolution. In television, the ability of a receiver to reproduce picture detail. It is usually expressed as the number of lines which can be seen on a reproduced test chart.

Resolution Pattern. (See Television Test Chart.)

Retrace. The return of the electron beam after a horizontal or vertical scan.

Retrace Time. The time which elapses during retrace. The time is approximately 7 microseconds for the horizontal retrace and 500 to 750 microseconds for the vertical retrace.

Rhombic Antenna. A diamond-shaped arrangement of conductors, all of the same length (rhombus) and joined at three corners with a transmission line connected at the open corner. The length of the conductors must be more than a wavelength. Used for fringe-area or low-signal reception.

S

Sawtooth. The waveform employed in television scanning.

Scanning. The process by which the light values of the picture elements which constitute the entire scene being televised are successively analyzed according to a predetermined method.

Scanning Generator. A circuit which produces the sawtooth wave of voltage or current required for proper beam deflection in the camera tube or picture tube.

Scanning Line. A single active horizontal line of the picture.

Scanning Spot. The cross section of the electron beam at the fluorescent screen of the picture tube.

Screen Persistence. The property by which the fluorescent screen continues to radiate light for a short time after the electron beam has passed.

Secondary Electron. An electron which is knocked out of a metal surface under the bombardment of another electron called the primary electron.

Secondary Emission. The phenomenon by which secondary electrons are produced. (See Secondary Electron.)

Series Peaking. An inductance in series with the video-amplifier plate, which compensates for loss of high-frequency gain and corrects high-frequency phase shift.

Serrated Pulses. In the long vertical-sync pulse, "notches" which keep the horizontal oscillator in synchronism during vertical retrace.

Shunt Peaking. An inductance in parallel with a video-amplifier load. It compensates for the high-frequency loss due to the shunt capacitance and corrects the high-frequency phase shift.

Single-Ended. Input circuits of television receivers in which one side of the transmission line is connected to the chassis or ground.

Single Sideband. A method of RF transmission in which the carrier and only one sideband are radiated.

Smear Ghost. A spurious image caused by multiple reflections or by phase shift in the video amplifier.

Snow. Television slang for the effect of random noise in the reproduced picture. (See Noise.)

Spurious Signal. Refers either to the effect of reflections of the carrier wave or to undesirable shading signals in the camera tube.

Stacked Arrays. Antenna systems in which two or more antenna arrangements are positioned above each other at a critical spacing and connected by transmission line. Stacking increases the pickup and improves the directional pattern.

Staggered Circuits. Interstage coupling circuits of a video-IF amplifier are staggered when they are tuned to different frequencies. Staggered tuning permits broadband response.

Surge Impedance. Same as Characteristic Impedance.

Sweep. Refers to motion of an electron beam at right angles to its direction. (Also see Scanning.)

Sweep Voltage. The voltage applied to the deflection plates or coils of a cathode-ray tube.

Sync. Abbreviation for synchronizing.

Sync Clipper. A vacuum-tube circuit which is biased so that the sync signals are removed from the composite video signal.

Sync Inverter. A circuit which produces a 180° phase shift of the sync pulses. Thus, necessary polarity for control of the scanning oscillator is provided.

Sync Leveler. (See Sync Limiter.)

Sync Limiter. A vacuum-tube circuit which produces sync pulses of uniform height.

Sync Separator. (See Sync Clipper.)

Synchroguide. A type of control circuit for horizontal scanning in which the sync signal, oscillator voltage pulse, and scanning voltage are compared and kept in synchronism.

Synchronizing Pulses. The portions of the transmitted signal which control horizontal and vertical scanning of the receiver. (See Horizontal- and Vertical-Sync Pulses.)

T

Tearing. A fault of the synchronizing system in which groups of lines are displaced; this causes the appearance of a torn picture.

Televise. To convert a scene or image field into a television signal.

Television Test Chart. A televised chart with geometric patterns by which the service technician or set user can determine the operating condition of the receiver.

Test Pattern. (See Television Test Chart.)

Time Constant. The time required for the voltage or current of a circuit to rise to 63% of its final value or to fall to 37% of its initial value.

Time Delay. The elapsed time between the start of a modulation wave or pulse at the transmitter and the start of its reproduction on the picture-tube screen.

Transient Response. The manner in which a circuit responds to sudden changes in applied potential.

Transmission Line. A two-conductor circuit with uniformly distributed electrical constants, used for transmitting radio-frequency signals.

Triggering. Starting of an action in a circuit, which then continues to function for a predetermined time under its own control.

U

Ultra High Frequencies (UHF). The portion of the television spectrum from 300 to 3,000 megacycles.

V

Vertical Blanking. The blanking signals at the end of each vertical field. They blank out the picture tube during vertical retrace.

Vertical-Blanking Pulse. (See Vertical Blanking.)

Vertical-Centering Control. Moves the picture vertically on the picture-tube screen.

Vertical Directivity. In a receiving antenna, the reception characteristic of the antenna in the vertical plane.

Vertical-Hold Control. (See Hold Control.)

Vertical Oscillator. The sawtooth-scanning generator which furnishes the required voltage or current wave for vertical scanning.

Vertical Polarization. A transmitted signal is vertically polarized when its electrostatic component is at right angles to the earth's surface.

Vertical Resolution. The number of picture elements which can be resolved vertically.

Vertical Retrace. The return path of the electron beam (blanked out) from bottom to top at the end of each field.

Vertical Scanning. The vertical movement of the beam on the picture tube.

Vertical-Scanning Generator. (See Vertical Oscillator.)

Vertical-Sync Pulses. Between each field, a series of six pulses which synchronizes the vertical-scanning oscillator.

Very High Frequencies (VHF). The portion of the television spectrum from 30 to 300 megacycles.

Vestigial-Sideband Transmission. The standard system of video modulation. The carrier is modulated by a complete upper sideband and a vestige (1.25 mc) of the lower sideband.

Video. (Latin for "I See.") The frequencies employed for mdulation of a television carrier.

Vidicon. A camera tube which has a photoconductive mosaic and which operates similarly to the Image Orthicon.

W

Width. The horizontal dimension of the picture or the time duration of a pulse (pulse width).

Width Control. Controls the horizontal dimension of the picture so that it fills the picture-tube screen.

Y

Yagi Array. An arrangement of dipole antenna elements employed for television reception. One element acts as the antenna; and the others act as parasitic elements (directors and/or a reflector), so that gain and the directional reception pattern are improved.

Answers to Questions

CHAPTER 1
(Pages 5-11)

1. Cathode, control grid, focus anode (anode 1), and acceleration anode (anode 2).
2. An indirectly-heated cathode.
3. The quantity of electrons in the beam, which is is governed by the control grid.
4. Focusing anode.
5. Deflection sensitivity.
6. (1) The velocity of the beam. (2) The strength of the deflecting field.
7. They provide a means of moving the picture frame for proper horizontal and vertical positioning.
8. Raster.

CHAPTER 2
(Pages 12-17)

1. Right angle.
2. When the electron stream runs parallel to the lines of force of the magnetic field.
3. (3) Focusing anode.
4. By adjustment of the fine-focus screw which moves a soft-iron magnetic shunt back and forth.
5. Two sets of coils placed around the neck of the tube. One set is for horizontal deflection, and the other set is for vertical deflection.
6. A sawtooth current.
7. (2) The yoke.
8. The ions must be removed. If the ions are allowed to strike the phosphor screen, a brownish or burned area will result because they are much heavier than the electrons.

CHAPTER 3
(Pages 18-24)

1. It translates the scene to be transmitted into electrical impulses.
2. The amount of light on that particle in the image.
3. A semitransparent photocathode.
4. (3) Monoscope.
5. 30 frames per second and 525 horizontal lines per frame.
6. Vertical—60 cycles per second.
   Horizontal—15,750 cycles per second.

CHAPTER 4
(Pages 25-30)

1. Current is higher and good regulation is necessary.
2. The heavy and bulky power transformer is not necessary. Without this transformer, however, there is a shock hazzard because the chassis is connected to one side of the power line.
3. Voltage-divider networks are used, whereby voltage-dropping resistors reduce the voltage to the desired value. The plate resistance of the audio output tube also may serve as a voltage-dropping resistor.
4. It is first stepped up by a transformer and then rectified.
5. From a winding of only one or two turns on the high-voltage transformer.
6. A high-voltage pulse that is developed during retrace time.
7. The filtering requirement is *small* because of the high frequency and the low current drain.
8. 25,000 to 30,000 volts produced by a tripler circuit.

CHAPTER 5
(Pages 33-38)

1. In a charged capacitor, one plate contains more free electrons than the other plate. When a capacitor is discharged, both plates contain the same number of free electrons.
2. The voltage on the capacitor increases, and the voltage across the resistor decreases.
3. Nonlinear—exponential curves.
4. At the first instant, the voltage across the capacitor and resistor are equal, and then gradually decrease as time progresses.
5. The time required to charge a capacitor to 63.2% of the applied voltage.

   $$t \text{ (in seconds)} = R \text{ (in ohms)} \times C \text{ (in farads)}$$

6. Integrator voltage; differentiator voltage.
7. The voltage across the capacitor decreases, and the voltage across the resistor approaches that of the input.

CHAPTER 6
(Pages 39-49)

1. Neon and thyratron.
2. By use of a capacitor connected from the output of the second stage to the input of the first stage.
3. The discharge of the coupling capacitors.
4. The time constant of the R-C circuit of one tube is made greater than that of the other tube.
5. A common-cathode resistor is employed, and a coupling capacitor from the plate of the second stage to the grid of the first stage is not used.
6. The sawtooth-forming capacitor is allowed to discharge through the tube.
7. The current through the primary sets up a magnetic field, and a secondary voltage is induced across the grid winding, causing a positive potential to appear at the grid.
8. The magnetic field of the transformer collapses, and the voltage induced in the secondary is such that it causes the grid to go more negative.
9. It conducts plate current and discharges the sawtooth-forming capacitor.
10. (a) Symmetrical square waves.
    (b) A short rectangular pulse followed by a long gap.
    (c) A short sine-wave pulse followed by a long interval of relaxed oscillation.

CHAPTER 7
(Pages 50-57)

1. 1/60th of a second.
2. 15,750.
3. Triggers the vertical oscillator, blanks out the screen during retrace, keeps the horizontal oscillator in step during retrace, and provides for proper interlace.
4. (c) Equalizing pulses.
5. A positive pulse controls the blocking oscillator. A negative pulse controls the cathode-coupled multivibrator.
6. A sawtooth of current is passed through the coils. The voltage applied is a combination sawtooth and square wave.
7. By adding a resistor in series with the sawtooth-forming capacitor. This circuit is called a peaking circuit.

CHAPTER 8
(Pages 58-77)

1. To stop, or damp out, oscillations in the horizontal-sweep circuit and help produce the required linear current sawtooth through the deflection coils.
2. Inductive.
3. The first half.
4. Horizontal-linearity control, width coil, and horizontal-drive control.
5. It matches the plate impedance of the vertical-output tube to the impedance of the vertical-deflection coils.
6. An integrator network rejects the horizontal-sync pulses. A differentiating network accepts the horizontal-sync pulses.
7. The horizontal-sync pulse, a signal fed back from the output of the oscillator, and a pulse from the horizontal-deflection coils.
8. The stabilizing coil and a capacitor in parallel produce a sine wave of voltage which, when superimposed on the grid-voltage waveform, steepens the slope of the grid-voltage waveform immediately before conduction time.
9. It locks the phase and frequency of the horizontal oscillator.

CHAPTER 9
(Pages 81-89)

1. 30 cycles per second to over 4,000,000 cycles per second.
2. 4:3.
3. The brightness of the spot becomes darker.
4. 6 megacycles.
5. Vestigial-sideband modulation.
6. At 1.25 megacycles above the lower limit of the channel. The limit of the upper sideband is 4 megacycles above the picture carrier.
7. 75%. Not to exceed 15% and not to go below 10% of maximum carrier amplitude.
8. Maximum.

CHAPTER 10
(Pages 90-100)

1. (a) At the video-detector input.
   (b) At any of the video-amplifying stages.
   (c) At the point of DC restoration.
2. To remove the sync signals from the composite video signal.
3. Positive.
4. The gain is high enough that the output signal does not need amplification.
5. Across a resistor. Across a capacitor.
6. They differ in time duration.
7. Smaller. Total time constant calculation is the same as for resistors in parallel:

$$\frac{1}{T} = \frac{1}{T1} + \frac{1}{T2} + \frac{1}{T3}$$

CHAPTER 11
(Pages 101-113)

1. Horizontal; horizontal polarization.
2. The characteristic impedance of the transmission line must be matched to the impedances of the receiver input circuit and the antenna.

3. When the antenna is rotated, the ghost images due to line reflections will remain stationary; but the ones due to signal reflections will vary.
4. Uniform response, high gain, and sharp directional pattern.
5. When the dipole is horizontal and at right angles to the transmitting antenna. In other words, a line drawn from the receiving antenna to the transmitter should be at right angles to the length of the dipole.
6. The director is a parasitic element that is placed in front of the antenna and is used to direct the signal to the antenna. The reflector is a parasitic element that reflects to the antenna a portion of the signal that passes the antenna. The director element is the one that is located on the side toward the transmitter, and the reflector is the one on the side away from the transmitter. The reflector is longer than the director.
7. A stacked array consists of more than one antenna mounted one above the other and spaced the proper distance. The advantages are:
   (1) Additional gain is obtained.
   (2) Some vertical directivity is contributed by the mutual interaction of the antennas.

CHAPTER 12
(Pages 114-126)

1. RF amplifier; converter or first detector; local oscillator.
2. RF amplifier.
3. By a capacitor connected from the low side of the plate coil to the grid of the tube.
4. It uses a twin-triode tube in which the first section has a grounded cathode and the second section has an RF-grounded grid.
5. The input coils are mounted on one strip; and the RF plate, mixer grid, and oscillator coils, on the other strip.
6. Features of the turret and switch-type tuners are used.

CHAPTER 13
(Pages 127-136)

1. Fine detail of the picture is lost.
2. (a) Overcoupled transformer with shunt-resistance loading. (b) Staggered-tuned circuits.
3. The response curve flattens out and the Q and gain become lower.
4. Instead of all the stages being tuned to the same frequency, each stage is tuned to a different frequency about the center frequency.
5. The video response must be reduced at the position which would be occupied by the sound carrier of the channel being received, the video carrier of the next higher adjacent channel, and the sound carrier of the next lower adjacent channel.
6. To amplify two signals of widely different frequencies in the same tube.
7. Negative; plate.
8. The crystal diode.

CHAPTER 14
(Pages 137-143)

1. 4.5 mc.
2. 25 kc.
3. The video-detector output and the video-amplifier output.
4. Series.
5. Because the ratio detector is not sensitive to amplitude modulation.
6. Limiter, audio detector, and 1st audio amplifier.
7. Limiting in the 6DT6 depends on the damping of strong signals in the grid circuit. A strong input signal causes the control grid to draw considerable current, which loads down the tuned circuit connected to the grid. The oscillation is suppressed by the grid loading.

CHAPTER 15
(Pages 144-153)

1. Improper contrast of large areas in relation to the smaller objects in the picture.
2. By using a large electrolytic bypass capacitor.
3. Fine lines or small picture elements will be missing, or will be blurred.
4. The shunting effect of the stray capacitance of the various circuit elements to ground. The loss is minimized by using shunt- or series-peaking coils. A combination of the two may be used.
5. The brightness control is an adjustment of the bias between the control grid and cathode of the picture tube; this control establishes the correct blanking or black level.
6. (a) The contrast control can be connected between cathode and ground, and can be used to change the bias on the stage.
   (b) If the control-grid resistor is connected to the cathode end of the control, the effective B+ voltage on the tube is varied.
   (c) The contrast control can be connected in parallel with the plate load resistor to tap off a portion of the video signal.
7. The setting of the brightness control.

CHAPTER 16
(Pages 154-158)

1. To minimize the effect of changes in signal strength at the receiving antenna.
2. In amplified AGC, the rectified voltage is amplified before being applied to the controlled stages.

3. Two; a composite video signal which contains positive-going horizontal-sync pulses; and positive pulses from a winding on the horizontal-output transformer.
4. The level of the sync pulses of the composite signal applied to the grid.
5. It limits the amount of current in the circuit should the grid draw current. The impedance of the resistor also isolates the input capacitance of the keying tube from the video circuit, hence the loading effect on the video circuit is greatly reduced.
6. Because the plate has an approximate potential of 35 volts in order that the tube will conduct.

   The plate of the 6BU8 is connected to the grid of the horizontal discharge tube through a voltage divider. The voltages at the intermediate points of the voltage divider are applied to the AGC line.
7. When the bias voltage tends to become positive, the clamper diode conducts and effectively shorts the AGC line to ground.

CHAPTER 17
(Pages 159-167)

1. The horizontal wedges are used to determine the resolution of the vertical section of the receiver. The vertical wedges are used to determine the resolution of the horizontal system of the receiver.
2. The circles are viewed while adjustments are made for vertical height and linearity and for horizontal width and linearity.
3. The yoke, the focus device and/or control, and the ion trap must be adjusted.
4. Horizontal width and linearity controls.
5. Horizontal-drive control.
6. Controls for vertical hold, height, and linearity.

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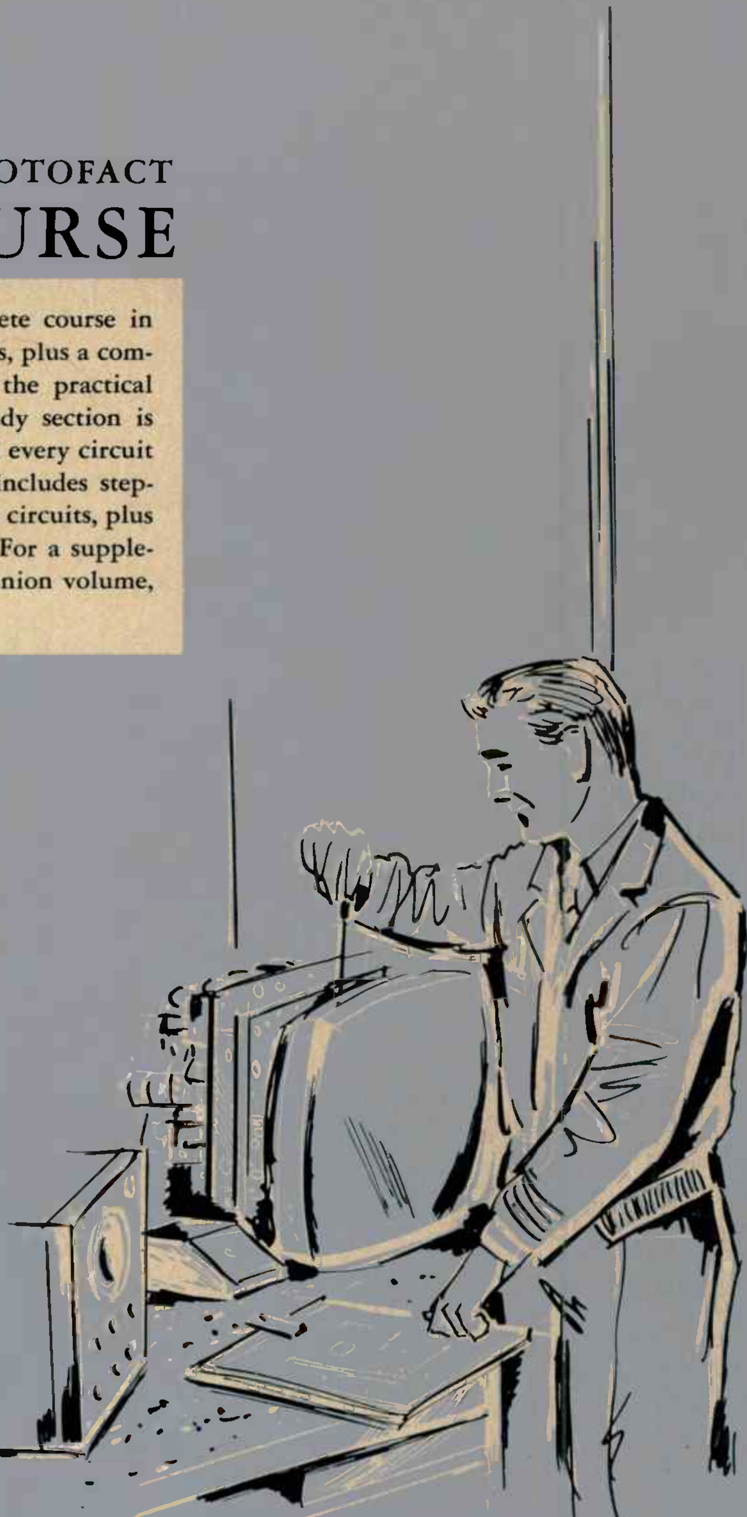

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