

TELEVISION

Simplified

By **MILTON S. KIVER**

The most up-to-date, complete, and practical television handbook now available for repair men, radio workers and all who are interested in the practical opportunities a working knowledge of television affords. Basic principles are explained in language free from involved theory or mathematics—from the analysis of circuits and operating fundamentals of V.H.F.—U.H.F., TV and F-M, to repairing, troubleshooting and such important material as TV tuners, keyed A.G.C. systems, Cascode amplifiers, 45-mc video I.F. systems, a completely new chapter on COLOR TELEVISION.

NEW 5th Edition—Revised and enlarged



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Little is assumed of the reader beyond an elementary knowledge of home sound receiver operation. Upon this is built an understanding of the modern television set with its highly integrated synchronizing circuits. This up-to-date revision contains many new illustrations and schematic diagrams and a complete new chapter on U.H.F. Added material includes an expanded discussion of TV tuners, an explanation of keyed A.G.C. systems and how they are utilized, D.C. video amplifiers, and Cascode amplifiers and how they function. In addition to an enlarged Intercarrier receiver section, two television receivers are completely analyzed and the new 45-mc video I.F. systems as well as the older 25-mc I.F. circuits are discussed.

No other manual in the field offers so complete and basic a treatment of the subject, in as compact and easy-to-follow a presentation. As an additional aid for those who use this manual without formal instruction, a set of self-check questions for each chapter has been added.

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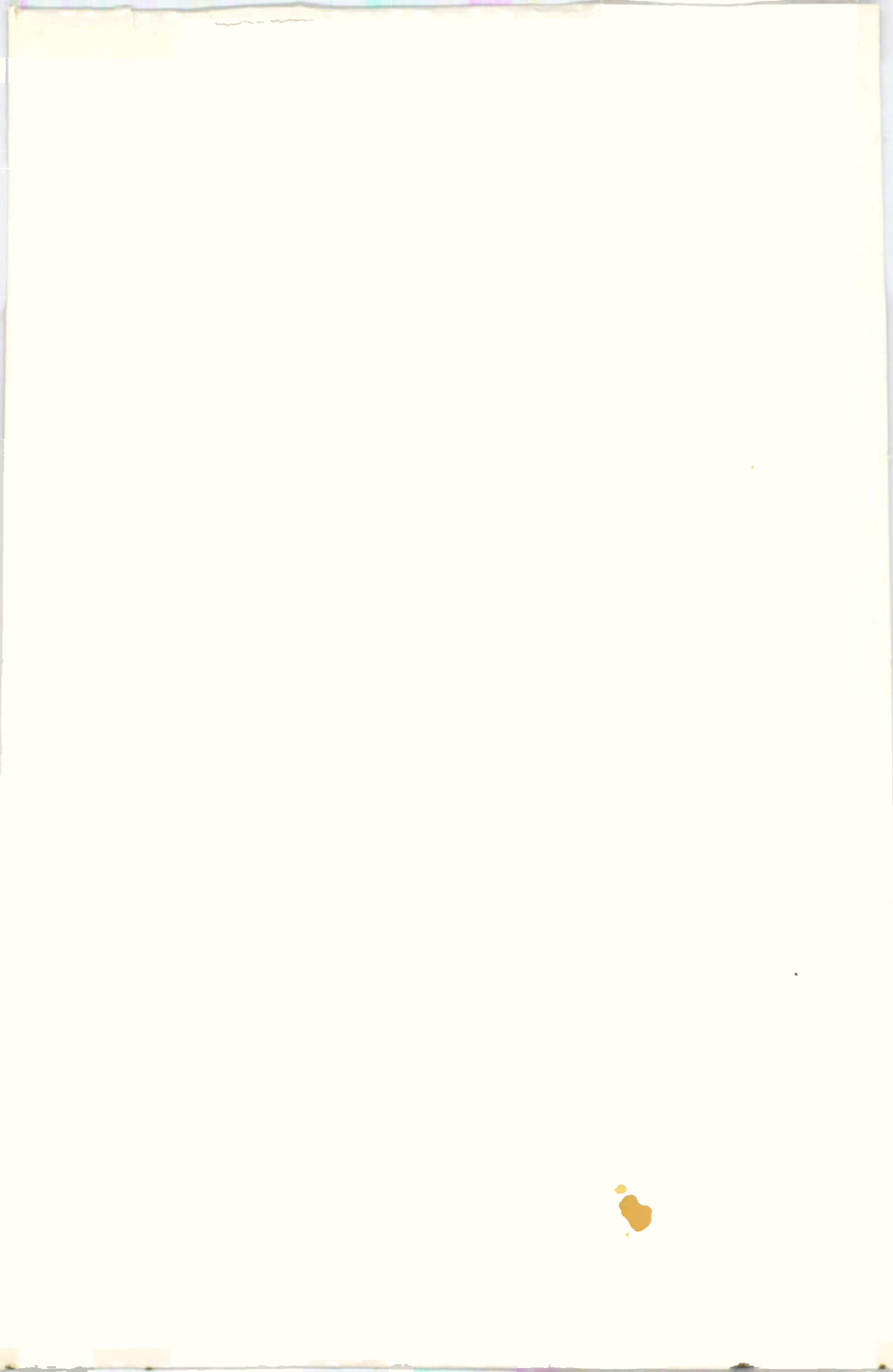
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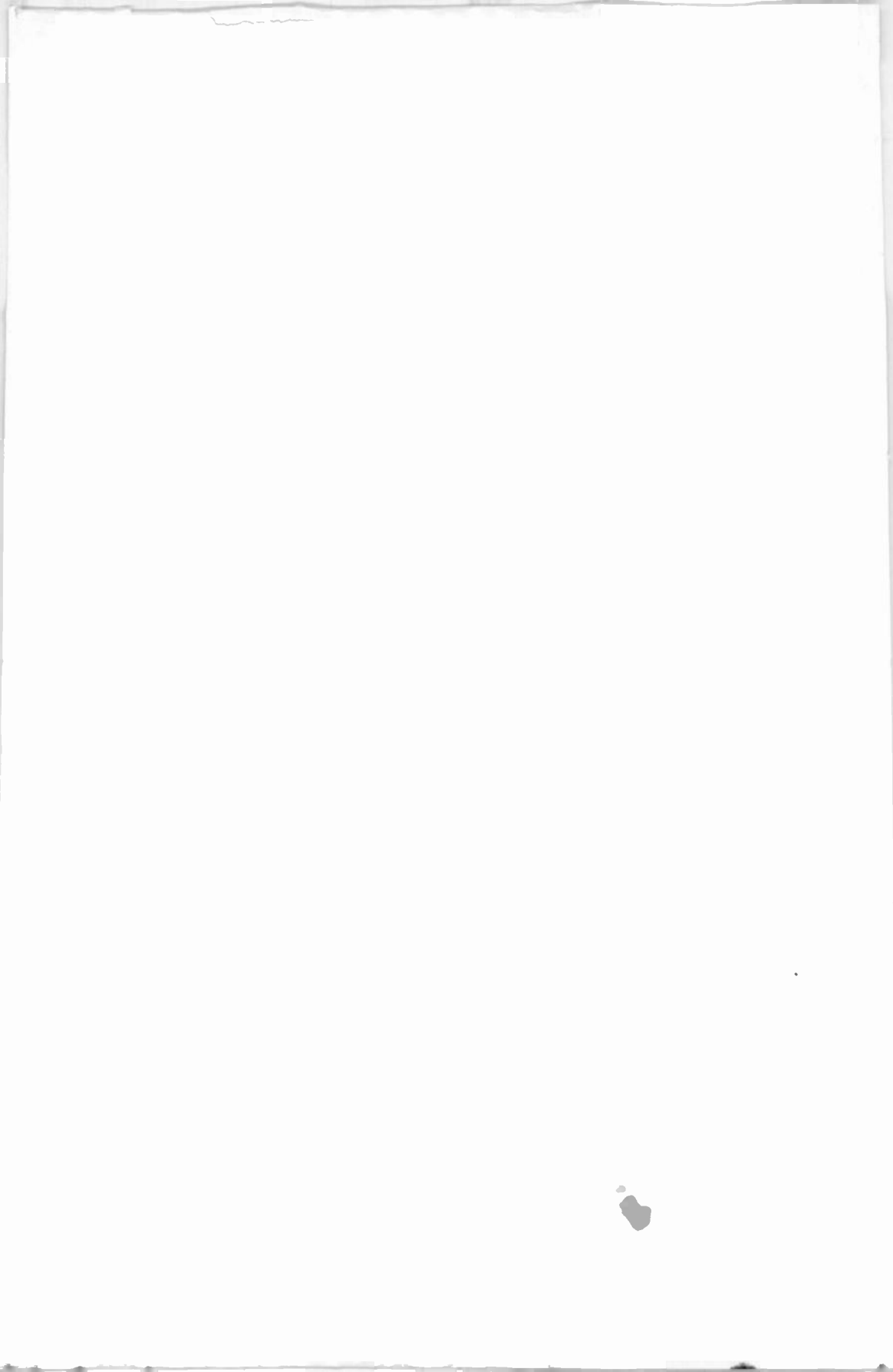
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BY THE SAME AUTHOR

F—M Simplified, 2nd Ed.

Television and F—M Receiver Servicing, 3rd Ed.

Introduction to UHF Circuits and Components

PREFACE TO THE 5TH EDITION

Changes are occurring so rapidly in color television that the material in Chapter 18 which was up-to-date at the end of 1953 is now sufficiently far behind current practice to warrant a complete rewriting. This portion of the 4th edition has been thoroughly revised and it now covers in detail the operation of the various circuits of a large screen color television receiver. With the information given, the reader will be in a position to service color receivers as effectively as black-and-white sets.

PREFACE TO THE 4TH EDITION

In the short space of eight years, the original volume of *Television Simplified* has undergone three complete revisions. Furthermore, each new edition contained at least 30 per cent more material than its predecessor. These simple figures reveal as well as any set of formal statistics the phenomenal growth of the television industry. This book has been geared to keep pace with that growth.

The presentation in this fourth edition follows the pattern established in previous editions. Little is assumed beyond an elementary knowledge of the operation of home sound receivers and upon this is built an understanding of the modern television receiver with its highly integrated synchronizing circuits. Chapter 1 presents an outline of the various units that combine to form a television system. It attempts to answer those pertinent

questions which always arise when any subject is first investigated and which, if left unanswered, soon begin to interfere with the smooth accumulation of subsequent information. With each succeeding chapter, a different section of the television receiver is discussed, starting at the input end of the set and traveling along the same path as the incoming signal. The function of every part, both within its stage and within the receiver as a whole, is carefully noted.

Because of the wealth of new material in this edition, several of the chapters in the previous (i.e., third) edition have been divided into two chapters. Thus, we have now separate chapters for video I.F. amplifiers and for power supplies, where formerly these were combined with other material. A new chapter (Chap. 17) is devoted to UHF television and the methods which are currently being employed to receive signals in this section of the frequency spectrum.

The genuine test of how well knowledge has been acquired is its application to everyday problems. Chapter 16 on servicing has, as one of its aims, the co-ordination and application of all the facts contained in the preceding chapters. The television receiver is divided into several major sections and, from the facts previously presented, the majority of troubles is systematically analyzed.

A set of questions is included at the end of each chapter for those who wish to gauge their progress through the book. The questions are straightforward and are drawn wholly from the text material.

No book represents the sole efforts of one person and this volume is no exception. Grateful acknowledgment is due to the Radio Corporation of America, the Allen B. DuMont Laboratories, Inc., the General Electric Company, Sylvania Electric Products, the Walter L. Schott Co., and Howard W. Sams & Co., Inc., for their generous aid in furnishing illustrations and data that were essential in the preparation of the book. The author also wishes to thank George Kearse for his permission to use the antenna chart appearing in Chapter 2.

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Chapter 1

THE TELEVISION FIELD

Introduction. Television, the science of transmitting rapidly changing images from one place to another by means of electromagnetic waves, is today as vital a factor in the home as the amplitude-modulated (A-M) sound receiver.

Accompanying this change in media has come a corresponding change in the operating technique of the radio serviceman. Not only was he impelled to extend his knowledge to include television circuits, but, just as important, he was brought face to face with an array of new test instruments and new servicing methods. Television has also forced other technicians, men and women, to become familiar with their basic operation. The change has not been a simple one, for present-day television receivers are intricate, critical mechanisms, and the person with an insufficient technical background will rapidly find the situation hopeless. The rise of television is indeed the final nail in the coffin of the semi-skilled screw-driver mechanic.

Television receivers are housed in either large console cabinets or in the smaller table enclosures. A television receiver is shown in Fig. 1-1 which is representative of the outward appearance of commercially manufactured sets. The number of tubes vary with the elaborateness of the layout, with as few as 15 tubes or as many as 30 or more. However, although this increase in tubes may be appreciable compared to A-M sets, the number of *additional* tuning controls at the front of the receiver is kept as low as possible. One control regulates the background brightness, another permits the proper station to be obtained, while the third control adjusts the intensity of the various objects in the image itself. The latter is known as the contrast control.

In addition to the three television controls just mentioned, others will be found. One is labeled "Fine Tuning." This is actually a correction control. The main tuning circuits used in television receivers are relatively



FIG. 1.1. A combination A-M, F-M, and television table model receiver.

fixed and the desired station is obtained either by push buttons or a selector switch. In addition to these main resonant circuits, the proper oscillator coil and condenser are likewise selected at the same time. If any change should occur in the resonant properties of these circuits and no adjustment were provided, it may readily be seen that distorted sound and optical

images would be obtained at the output of the receiver. To prevent such a situation, a fine tuning control is placed on the front panel. Within limits, this permits the observer to center the entering signal so that the proper frequencies are obtained at the video and audio I.F. amplifiers. Thus, we may consider this knob as a vernier adjustment.

Among the other controls on the front panel, we find volume and sometimes tone controls. These are associated with the sound portion of the television receiver and their function is the same as in the more common A-M receiver. Thus, while the mechanism of the television radio set may be quite complex, the necessary controls are few and readily learned, even by those entirely unfamiliar with the technical aspects of radio. It will be readily appreciated that this must be so, if television is to become popular.

In the following paragraphs of this chapter, the overall operation of the present-day television system is explained, with particular emphasis on the methods used to transform light rays into equivalent electrical impulses. After that, chapter by chapter, and section by section, the receiver operating principles are presented, assuming only a rudimentary knowledge of the operation of present superheterodyne A-M receivers.

Desirable Image Characteristics. Since the image is the final product of the television system, and because everything centers about the production of this image, here is the most logical place to begin. In order for the picture to be satisfactory from the observer's point of view, the following minimum requirements should be obtainable.

1. The composition of the image should be such that none of the elements that go into its make-up is visible from ordinary viewing distances. This requires that the image have the same fine, smooth appearance that we obtain with good photographs.

2. Flicker must be totally absent. To accomplish this, it is necessary for the cathode-ray beam to sweep across the fluorescent screen in time to cause light to be emitted before the previous image has lost its effect in the observer's mind. Then the scenes follow each other in rapid succession and the action appears continuous.

3. The size of the picture should be large enough to permit comfortable viewing by several people at distances of 10 feet or more from the screen.

4. To meet the changing requirements for viewing the screen either by day or by night, an adequate amount of light must be available from the cathode-ray screen. Naturally, less would be necessary when the room illumination is low than when it is high.

5. The final consideration, contrast, is less important than any of the previous conditions, but effective range is still desirable. Contrast refers to the ratio between points of maximum to minimum brightness on the same

screen. In broad daylight, for example, the contrast ratio between places in bright sunlight to shaded areas may run as high as 10,000 to 1. On fluorescent screens, however, the amount of light that can be emitted is definitely limited, and only contrast ratios of from 50 to 100 to 1 are obtainable ordinarily. These, however, prove quite satisfactory.

The foregoing requirements have been listed with only a slight explanation advanced for each. There are limitations which affect these conditions, but before any extensive discussion is undertaken, it is necessary to gain a more detailed knowledge of the overall operation of present-day television systems.

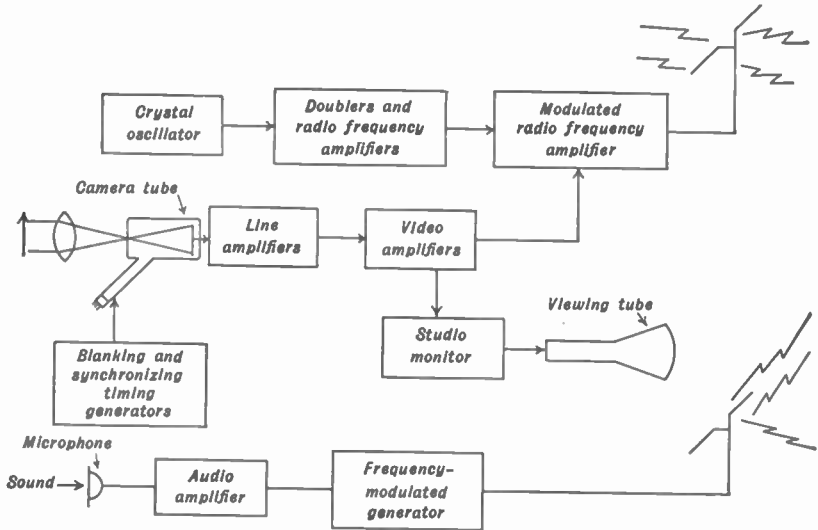


FIG. 1.2. An outline of a modern television transmitter.

Outline of Stages of Television Transmitters and Receivers. An outline of the various stages of a television transmitter is shown in Fig. 1-2. The scene to be televised is focused onto the photosensitive plate of the camera tube by means of a lens. At the tube, the light rays are transformed into equivalent electrical impulses. Thereafter, amplifiers and the regular amplitude-modulating sequences take place, the final television signal is formed and transmitted into space. To synchronize the position of the electron beam at the receiver-viewing tube with the beam in the camera tube, synchronizing pulses are inserted into the television signal as well.

We may pause for a moment and observe that, aside from the synchronizing pulses, the action in a television transmitter is entirely analogous to the corresponding action in a sound transmitter. In one, the object is to transform audio vibrations in the surrounding air to equivalent electrical varia-

tions. A microphone accomplishes this simply. In the other, light rays are changed into equivalent electrical variations and a camera tube is employed. In either case, once the currents have been formed, essentially the same procedure is followed to form the final amplitude-modulated R.F. signal. It is well to keep the correspondence between the purpose of the microphone and the camera tube in mind, for this will aid in visualizing the overall operation of television transmitters.

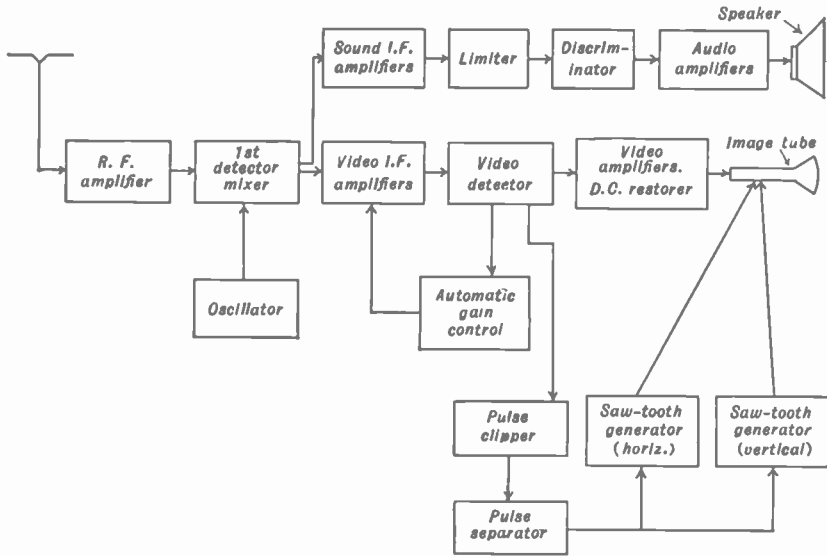


FIG. 1.3A. A block diagram of a television receiver operating under the conventional system. (The word "conventional" is used purposely because there are many receivers which operate slightly differently. These are known as Intercarrier sets and are discussed in conjunction with Fig. 1.3B.)

The sound that is spoken by the actors in the scene being televised is kept separate from the video electrical currents. The sound is frequency modulated and sent out by another transmitter at a frequency that lies close to the edge of the band of frequencies utilized by the image signals. So far as the transmitters are concerned, two separate units are necessary: one for the sound, the other for the image signals.

At the receiver, shown in block form in Fig. 1-3A, the video and audio carriers are received simultaneously by wide-band amplifiers. After amplification by an R.F. stage (if used), the composite signal is applied to the mixer tube where it is acted on by the high-frequency oscillator voltage. The desired I.F. values are produced by this action and, at the output of the mixer stage, the video and sound signals are separated and fed to their respective I.F. amplifiers.

The audio signal is frequency modulated and, although the I.F. amplifier stages found in F-M receivers do not differ radically in construction from the corresponding amplifiers in A-M sound superheterodyne receivers (except for frequency), the detector is entirely new. In the F-M set, a discriminator is necessary in order to convert the F-M signal into the equivalent audio variations. A brief description of the operation of F-M receivers will be given in Chapter 14. Once past the discriminator, the ordinary audio stages amplify the signal until it is suitable for application to a loud-speaker.

Returning to the video signal, we find that, after separation from the audio voltage, it passes through several I.F. amplifiers (the number ranging from three to five), before the diode detector is reached. Either half-wave or full-wave rectification is employed at the detector. At some point beyond the detector, a portion of the signal is applied to the synchronizing section of the receiver. Here, the synchronizing pulses are separated from the picture detail and used to actuate oscillators that directly control the position of the electron beam in the cathode-ray tube. In this manner, the exact point where the electron beam impinges on the fluorescent screen is kept related to the electron beam in the studio camera tube. Only vertical and horizontal synchronizing pulses are required for black and white images.

The remainder of the video signal, where the detail information is contained, is amplified by the video amplifiers and then applied to the control grid of the viewing tube. The amplitude of the input voltage varies the intensity of the electron beam while the deflecting plates (or coils) are swiftly moving the beam from one side of the screen to the other. The result is an image on the screen, produced by approximately 500 distinct lines. The eye of the observer integrates these lines so that they blend into each other, and the image assumes the appearance of a photographic picture.

After the scanning beam forms an image in this manner, a second picture, a third picture and so on are formed in such rapid succession that the blending of each into the next becomes even and continuous, as in the movies. When the system is operating properly the viewer is not aware of each individual picture.

Intercarrier TV Receivers. The foregoing discussion was centered about the block diagram in Fig. 1-3A. This is a conventional-type of television receiver and was the form employed by the first sets; it is still utilized by a number of manufacturers. However, in 1948, a television circuit operating on a somewhat different principle was introduced and now is extensively employed. The new system is known as the Intercarrier system. It is illustrated, in block form, in Fig. 1-3B.

The R.F. stages of the receiver are identical with the R.F. section of the conventional-type set. The incoming signal is received, amplified, and then

converted to the lower intermediate frequencies. The entire signal, sound and video, now enters the I.F. system where both are amplified. However, there is this important difference. The video signal receives its normal amount of amplification, but the sound signal is permitted to receive only 5 per cent of the total available amplification. This relationship is purposely maintained to prevent the sound signal from producing visible interference on the picture screen.

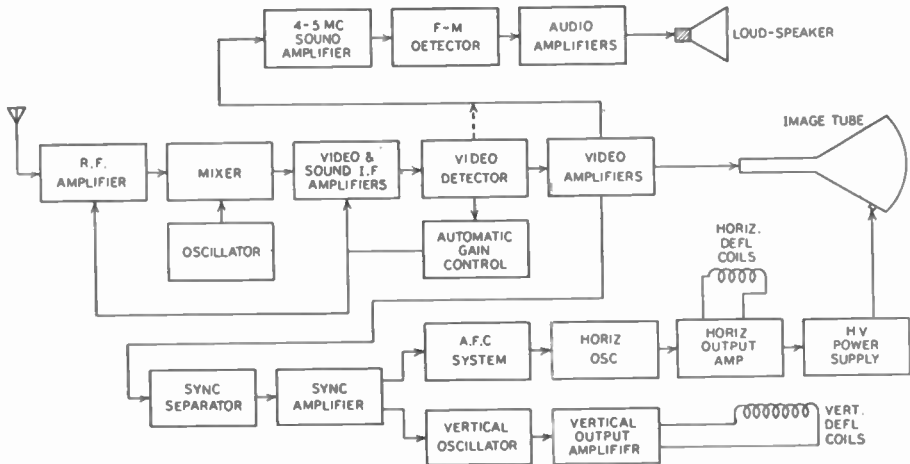


FIG. 1.3B. A block diagram of an Intercarrier set.

In the Intercarrier system, there is no sound take-off point in the mixer or the video I.F. system. Hence, both signals remain together until the video second detector. In this stage, a beating or mixing takes place between the video and sound I.F. carriers with the result that a difference frequency of 4.5 mc is produced.* If we wish, we can consider the video carrier as being equivalent to the local oscillator and the audio carrier acting as the incoming signal. The result of the mixing is a 4.5-mc beat note. In addition, we also obtain all of the 0-4 mc video frequencies from the amplitude-modulated video carrier.

The 4.5-mc beat note contains all of the sound information. This portion of the signal may now be transferred to its own sound system or the separation may be further delayed until both sound and video signals have passed through one or more video amplifiers. Once the separation is effected, the sound signal goes through one or two 4.5-mc sound I.F. amplifiers, through an F-M detector, and on to several audio amplifiers and the speaker.

* In every television signal, the frequency difference between the sound and video carriers is 4.5 mc.

The video signal, for its part, travels through the video amplifiers to the picture tube. Here, its information is displayed visually.

The vertical and horizontal deflection systems are synchronized by a portion of the signal obtained from some point in the video amplifiers. In this respect, both conventional and Intercarrier receivers are alike; hence, whatever was stated previously concerning the operation of these systems applies here, too.

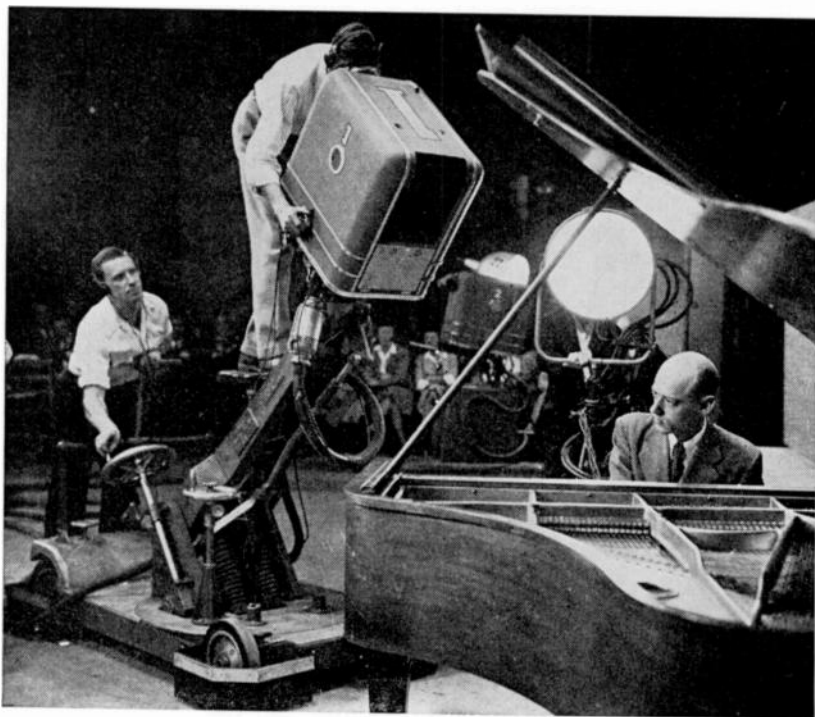


FIG. 1.4. A television studio scene developed from an ordinary negative.

The principal difference, then, between conventional-type and Intercarrier receivers lies in the point where the video and sound signals separate. In the conventional set, this must occur *before* the video detector; in the Intercarrier set, the separation takes place *after* the video detector.*

Television Camera Tubes. The preceding explanation is an outline of present-day television systems. With this in mind, let us investigate the important operation of the studio camera tube in greater detail, for it is what this tube "sees" and converts into equivalent electrical impulses that will determine the form of the final reproduced image at the receiver. Faithful

* A more detailed discussion of Intercarrier receivers is given in Chapter 11.

reproduction of the scene being televised is essential for high quality images at the receiver.

Consider an ordinary photograph, such as shown in Fig. 1-4. This was obtained from a negative that contained a large number of grains originally sensitive to light. So long as the picture or positive obtainable from the negative is not greatly enlarged or examined too closely, the granular structure of the photograph is not evident and the photograph appears smooth and continuous. However, if the picture is further and further enlarged,

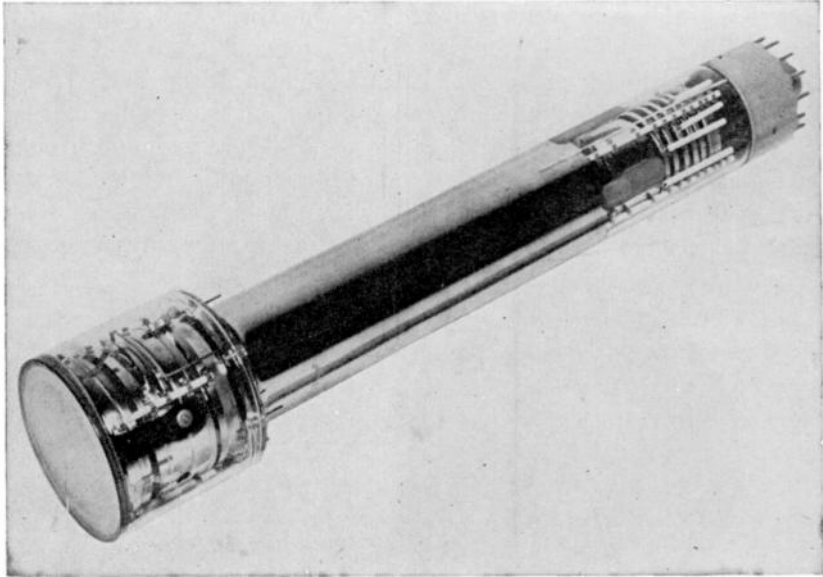


FIG. 1.5. An Iconoscope camera tube.

Courtesy Farnsworth

a point is soon reached where the granular structure of the picture does become visible. These grains, then, are the elements that combine to form the picture.

A fine grain photograph, in which there are many grains per unit area, is capable of greater enlargement than a coarse grain picture, before these elements become discernible. With television images, much the same sort of situation prevails. In the receiver, each picture element is just as large as the area of the circular beam impinging on the fluorescent screen of the cathode-ray tube. The light that is seen when observing a cathode-ray tube screen is derived from the energy given off by the impinging beam to the particles of the fluorescent coating on the inner face of the tube. If the points of light are closely spaced, the observer will integrate them and their separate character will disappear. Hence, one of the first considerations for a television picture that is to reproduce any amount of fine detail is an electron beam of small diameter. This requirement is just as important at the receiver screen as it is at the camera tube.



Courtesy RCA

FIG. 1.6. An Image-Orthicon Camera tube.

At present there are three types of camera tubes that are widely used in this country. They are known by the patented names of Iconoscope, Image Orthicon and Image Dissector tubes. A photograph of each is given in Figs. 1-5, 1-6, and 1-7. These tubes are commercially employed now; but, like the microphone, they are continuously being modified and revised as

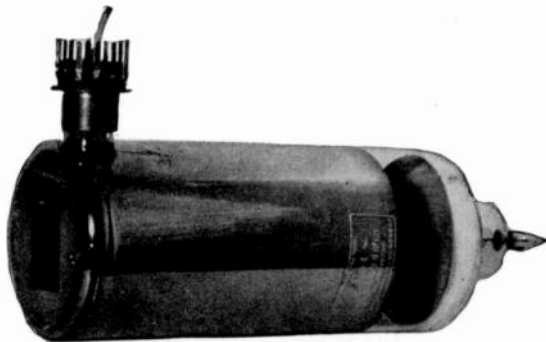


FIG. 1.7. Farnsworth Image Dissector tube.

better and more efficient methods are evolved that permit the same scenes to be televised under poorer conditions with better results.

The Iconoscope has the internal construction shown in Fig. 1-8. Within the tube is a relatively large rectangular plate upon which all the light from

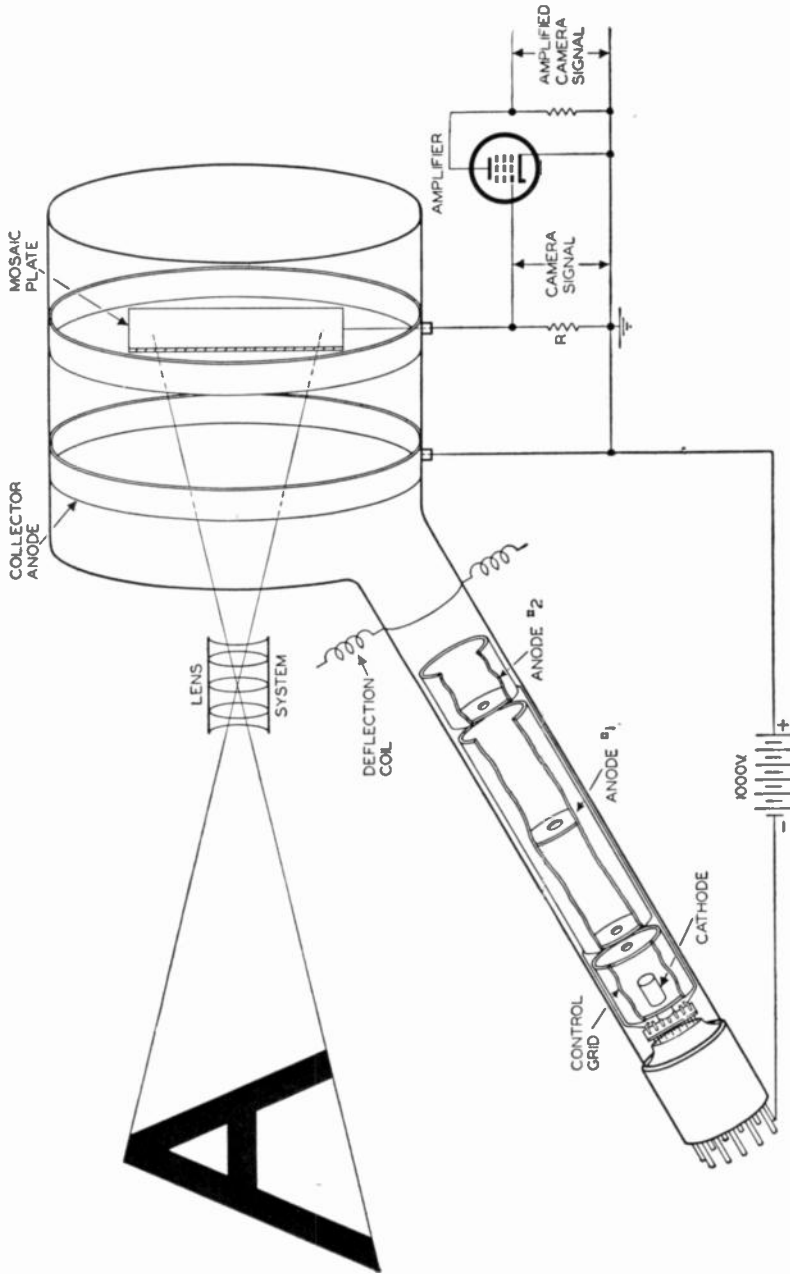


Fig. 1.8. The internal construction of an Iconoscope. The neck of the tube has been purposely enlarged out of proportion to the rest of the drawing to illustrate the construction of the elements of the electron gun.

the scene is focused. The plate consists of a thin sheet of mica (an insulator) upon the front of which has been deposited many microscopic globules of a sensitized caesium-silver compound. Due to the manner in which the globules have been placed on the mica plate, they do not come in actual contact with each other, each tending to form its own little island. Between these separate globules, of course, is the surface of the mica. On the reverse side of the plate, a continuous layer of some conducting substance is deposited and an electrical connection is brought from here to the external circuit.

It will be recognized that actually a condenser combination is formed by the foregoing method of construction. Each globule forms one separate plate, with the back side of the mica acting as the common second plate for all the globules. The dielectric is the mica.

The object is focused on the front face of the plate (commonly called the mosaic). Due to their silver-caesium composition, the globules emit electrons in proportion to the light intensity reaching that particular point. Thus each globule assumes a different positive charge due to this loss of negative electrons, with each element retaining this charge since it is insulated from all the other elements. The mica likewise prevents the charge from leaking off to the conducting layer on its other side. Essentially, we now have a charged condenser, but the charge varies from globule to globule because of the difference in light intensity that fell on these various points.

By having the amount of charge on each globule vary in proportion to the light at that globule, we have succeeded in accomplishing the first step of our process, namely, conversion of light rays into equivalent electrical charges. It remains for us to convert these charges into electrical currents.

The similarity between the above action and the corresponding photographic process of taking a picture is noteworthy. With more globules deposited on the mosaic, it should be possible to obtain a finer structure for the final reproduced image. This possibility will depend on the size of the electron beam and the scanning process used. The latter is associated with the method employed to convert the various differences in globule charge into corresponding electrical impulses. For the reproduction of fine detail, a fine grain structure is necessary, in photographic films. For the iconoscope mosaic, caesium-sensitized globules correspond to the grains on a film negative. The more globules that are deposited on the mosaic, the smaller the detail that may be distinguished. However, the number of globules, in itself, is not the only deciding factor. Important, too, is the diameter of the scanning electron beam. A large round beam covers many globules at one time, and an average current, determined by the average of the charges on all these globules, results. Any detail that is too fine will blend with the surrounding objects and become obscured. On the other hand, with a small

beam it is possible to contact smaller groups of globules and cause separate electrical currents to flow for each. The finer detail will be more evident now.

Electron Beam Scanning. In order to transmit a picture, it is possible to send all the elements that compose this picture at one time, or to send each element separately in some orderly sequence. Due to the complexity of

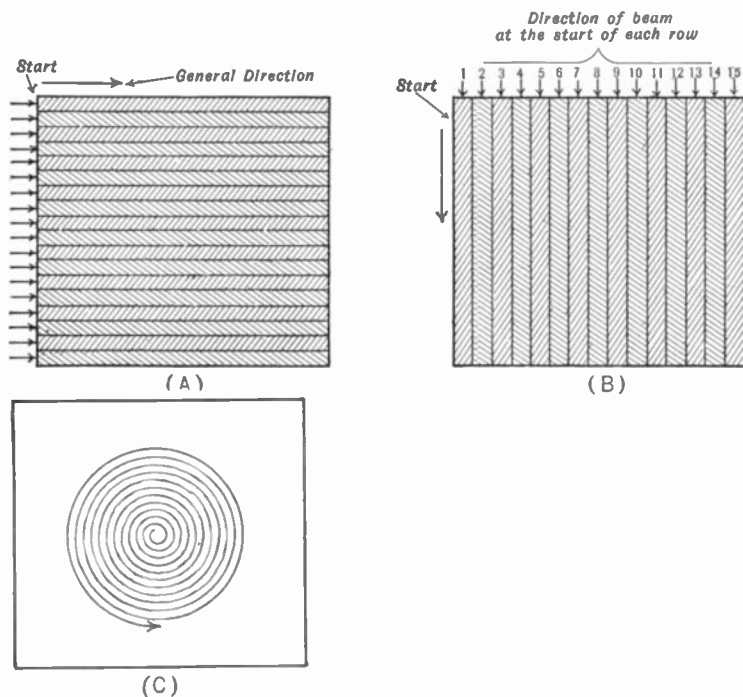


FIG. 1.9. Three possible methods of scanning an image: Horizontal scanning at (A), vertical scanning at (B) and spiral scanning at (C).

the system that would be required if an attempt is made to transmit all the elements simultaneously, the second method (sending each element separately) has been universally adopted. Even with these alternatives, there is still a choice of scanning sequence. For example, it is possible to divide the image into a series of narrow horizontal strips and transmit each after the other, starting at the left-hand side of the uppermost strip. Another method might dissect the image into vertical strips and transmit these in order, while a third means could employ spiral scanning. Each is illustrated in Fig. 1-9. Of practical interest, however, is horizontal scanning, since this is closest to the process currently employed in all American receivers.

At the start of the horizontal scanning process in the Iconoscope, an electron beam is formed, focused in the neck of the tube, and accelerated

toward the upper left-hand corner of the mosaic plate, point *A* in Fig. 1-10. There, under the influence of varying voltages applied to the Iconoscope's deflecting coils (contained on the neck of the tube), the electron beam moves to the right, passing over the charged globules that are located across the top of the image and which have been exposed to the focused rays of light from the televised scene. As each globule, or group of globules, is reached, enough electrons are supplied by the electron beam to restore the globule to its previously neutral potential.

This action automatically releases any charge on the opposite conducting surface of the mosaic that was held there by the positive globules.

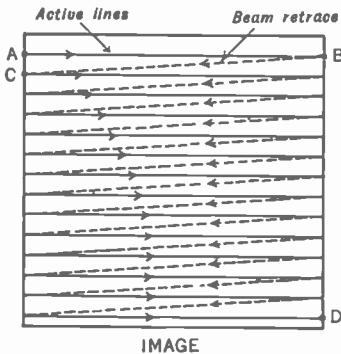


FIG. 1.10. The motion of the electron beam in one form of horizontal scanning.

With the release of this charge, a small pulse of current passes through resistor *R* of Fig. 1-8. The strength of this current is proportional to the amount of positive globule charge neutralized, which in turn was proportional to the intensity of the light striking this point of the mosaic plate. Thus the second phase of our task has been accomplished and we have transformed light rays into equivalent electric currents. The voltage developed across *R* will be proportional to the varying pulses of current passing through it. Tube *T* will then amplify the fluctuating voltage and forward it to the stages that follow.

Returning to the scanning process, the beam will continue along the first line until the end, point *B*, is reached. Here a generator connected to the camera tube will cut off or blank out the beam while the deflecting coils bring it rapidly back to point *C* at the left-hand side of the mosaic again. This point is slightly below the first line. The blanking voltage is removed now, and again the cathode-ray beam moves toward the right, neutralizing the positively charged globules along this horizontal line and causing electrical impulses to pass through *R*.

The sequence recurs until the end of the lowermost line is reached, at point *D*. The beam is blanked out and returned to the starting point *A*. The entire process is now ready for repetition. It should be noticed that each globule has been storing up a charge (or giving off electrons) during the time the electron beam is busy passing over other globules. Thus, if it takes the beam one minute to scan the entire image, during all of this time the globules are exposed to the focused light rays. The resultant emission of electrons causes the positive charge to increase. With the arrival of the beam, a neutralization takes place; but, at the next second, with the passage

of the beam, the storage process begins anew. While one minute is mentioned as an arbitrary period, in practice the beam passes over each globule every $\frac{1}{30}$ of a second. Hence 30 complete pictures are sent every second.

In actual equipment, the motion of the scanning electron beam, as described above, must be modified somewhat for two reasons. First, it is extremely difficult to generate a voltage that will cause the beam to drop suddenly from the end of one line to the level of the next one directly beneath it. It is simpler to have the beam move down to the level of the second line gradually, as illustrated in Fig. 1-11.

To obtain this type of motion for the electron beam, both horizontal and vertical deflection coils in the Iconoscope tube are utilized. Without going into any extensive discussion at this time of the operation of the electron gun located in the neck of the Iconoscope, let us state simply that the horizontal deflection coils can move the electron beams horizontally across the screen from left to right and back again. The vertical deflection coils can cause the beam to move vertically. Between them, and with different amounts of currents through each set of coils, it is possible to move the electron beam across the screen to reach any desired point.

In the foregoing type of motion, with the beam moving across the screen slantwise, we have the equivalent of a fast-acting voltage on the horizontal plates quickly forcing the beam straight across, while a slow-acting voltage at the vertical plates is forcing the beam down. The result is pictured in Fig. 1-11. When the beam reaches the end of a line, it is quickly brought almost straight across (with the blanking signals on) and thus finds itself in correct position to start scanning line 2 when the blanking voltage is removed. The remainder of the lines follow in similar fashion. At the bottom of the picture, after the last line has been scanned, a longer blanking signal is applied while the beam is returned to the top of the picture. The purpose of the blanking voltages is simply to prevent the beam from impinging on the screen when there is nothing to impart, but is merely moving into position for the next scanning run.

Many readers will probably note at this point that possible currents that could be used for the horizontal and vertical deflecting coils are of the familiar saw-tooth form illustrated in Fig. 1-12. These rise gradually to a

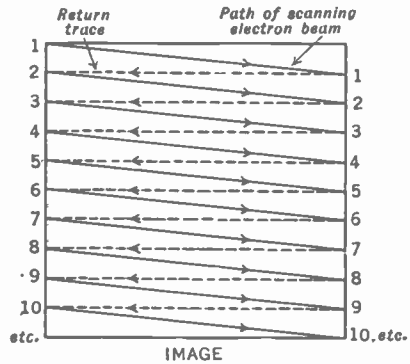


FIG. 1.11. In actual equipment, it is easier to have the electron beam travel in the manner indicated above, rather than as shown in Fig. 1.10.

fixed level and then suddenly drop (almost vertically) to zero to begin the process all over again. More will be mentioned about saw-tooth wave generators when the television receiver is discussed.

It was stated that there are two reasons why the horizontal scanning process as first explained had to be modified. The first reason has already been given. For the second reason, we must examine more closely the human eye and its action when observing motion on a screen.

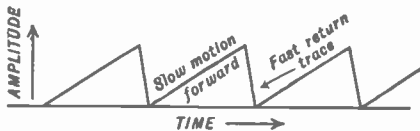


FIG. 1.12. A saw-tooth current, as illustrated here, when sent through a set of deflecting coils, will cause the electron beam to move slowly from left to right and then retrace rapidly from right to left.

Flicker. If a set of related still films follow each other fairly rapidly on a screen, the human eye is able to correlate them, and the motion appears continuous. The eye can do this because of the well-known phenomenon called persistence of vision. Due to this property of the eye, visual images do not disappear as soon as their stimulus is removed. Rather, the light appears to diminish gradually,

taking, on the average, about $\frac{1}{50}$ of a second before it disappears entirely. In motion pictures, this is very fortunate, for otherwise this method of entertainment would be impossible.

It has been found that when the theater films are presented at a rate of 15 stills per second, the action appears continuous. However, at this speed, a flicker is still detectable and detracts from the complete enjoyment of the film. The flicker is due to the sensation in the observer's mind reaching too low a value before the next film is presented on the screen. Increasing the rate at which the stills are presented will gradually cause the flicker to disappear. At 50 frames per second there is no trace of flicker, even under adverse conditions. The rate is not absolute, however, but depends greatly upon the brightness of the picture. With average illumination, lower frame rates prove satisfactory.

In the motion picture theater, 24 individual still films (or frames) are flashed onto the screen each second. Since at this rate, flicker is somewhat noticeable, a shutter in the projection camera breaks up the presentation of each frame into two equal periods. This is accomplished by having the shutter move across the film while it is being projected onto the screen. Thus we are actually seeing each picture twice; the fundamental rate has now been increased to an effective rate of 48 frames per second. By this ingenious method, all traces of flicker are eliminated.

In television, a fundamental rate of 30 images (or frames) per second was chosen because this frequency and the effective rate are related to the frequency of the alternating current power lines. Practically, this choice

of frame sequence rate results in less filtering in order to eliminate a-c ripple, which is called hum in audio systems. With 24 frames per second, for example, any ripple that was not eliminated by filtering would produce a weaving motion in the reproduced image. Less difficulty is encountered from a-c ripple when 30 frames per second are employed.

To eliminate all traces of flicker, an *effective* rate of 60 frames per second is employed. This is accomplished by increasing the downward travel rate of the scanning electron beam so that not every successive line, but

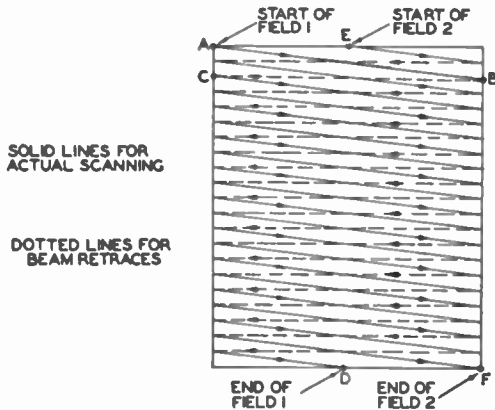


FIG. 1.13. The path of the electron beam in interlaced scanning.

every *other* line is sent. Then, when the bottom of the image is reached, the beam is brought back to the top of the image, and those lines that were missed in the previous scanning are now sent. Both of these operations, the odd and even line scanning, take $\frac{1}{30}$ of a second and so 30 frames is still the fundamental rate. However, all the even lines are transmitted in $\frac{1}{60}$ of a second and the same is true of the odd lines. Both add up, of course, to $\frac{1}{30}$ of a second. Hence, to the eye, which cannot separate the two, the *effective* rate is now 60 frames per second and no flicker is noticeable.

To differentiate between the actual fundamental rate and the effective rate, we say that the frame frequency is 30 cycles per second, whereas the effective rate (called the field frequency) is 60. This method of sending television images (see Fig. 1-13) is known as interlaced scanning.

Thus, as the standards for television images now stand, each complete scene is sent at a rate of 30 frames per second. In order to obtain the desired amount of detail in a scene, the picture is divided into a total of 525 horizontal lines. The technical reasons behind the choice of 525 lines are related to:

1. The frequency band width available for the transmission of the television signals. As will be shown later, the required band width increases with the number of lines.

2. The amount of detail required for a well-reproduced image.
3. The ease with which the synchronizing (and blanking) signals can be generated for the horizontal and vertical deflection plates.

With each frame divided into two parts (because of interlaced scanning), each field will have one-half of 525 lines or $262\frac{1}{2}$ lines from its beginning to the start of the next field. (As a matter of definition, a complete picture is called a frame.) With interlaced scanning, each frame is broken up into an even-line field and an odd-line field. Each field contains $262\frac{1}{2}$ lines whereas a frame has the full amount, or 525.

The Complete Scanning Process. From the foregoing discussion it becomes possible to reconstruct the entire scanning process. Although only the movement of the electron beam at the Iconoscope will be considered, an identical motion exists at the receiver screen.

At the start of the scanning motion at the camera tube mosaic, the electron beam is at the upper left-hand corner, point *A* of Fig. 1-13. Then, under the combined influence of the two sets of deflecting coils, the beam moves at some small angle downward to the right. When point *B* is reached, the blanking signal acts while the beam is rapidly being brought back to point *C*, the third line as required for interlaced scanning. The blanking signal then relinquishes control and the electron beam once again begins its left-to-right motion. In this manner every odd line is scanned.

When the end of the bottom odd line has been reached (point *D*), the blanking signals are applied while the beam is brought up to point *E*. Point *E* is above the first odd line of field 1 by a distance equal to the thickness of one line. The beam is brought here as a result of the odd number of total lines used, namely, 525. Each field has $262\frac{1}{2}$ lines from its beginning to the start of the next field and, when the beam reaches point *E*, it has moved through the necessary $262\frac{1}{2}$ lines from its starting point *A*. From here the beam again starts its left-to-right motion, moving in between the previously scanned lines, as shown in Fig. 1-13. The beam continues until it reaches point *F* and from here is brought to point *A*. From point *A* the entire sequence repeats itself.

Thus, as matters stand, the electron beam moves back and forth across the width of the mosaic $262\frac{1}{2}$ times in going from point *A* to point *D* to point *E*. The remaining $262\frac{1}{2}$ lines needed to form the total of 525 is obtained when the beam moves from point *E* to point *F* back to point *A*. The process may seem complicated but actually it is carried out quite readily and accurately at the transmitter (and receiver). A more detailed analysis, including the number of horizontal lines which are lost when the vertical synchronizing pulse is active, will be given in Chapter 12.

The Image Orthicon. Of the three camera tubes mentioned previously, the Iconoscope has already been described in some detail. It is not without

defects, the two most serious being its poor efficiency and its tendency to produce background shading that is not found in the original scene. The latter defect arises from the fact that the impinging scanning beam has sufficient force to dislodge secondary electrons from the surface of the globules over which it may be passing. Some of these electrons, once freed from their globules, may be attracted either to the positive collector ring or fall back on the mosaic plate. In either case, the charge distribution of the mosaic plate has been altered from its true, original form. This distortion, for that is what it is, appears on the screen, generally, as a darkened background. Correction voltages from a so-called shading generator are inserted into the signal in an effort to eliminate the distortion. In essence, the voltages from the shading generator are 180° out of phase with the distortion voltages, and their elimination is thus effected. It may appear to the reader that almost an infinite number of shading signals would be required. Fortunately this is not so. Experience with Iconoscopes reveals that relatively simple correcting voltages are required and these are readily generated and injected into the voltage wave.

A camera tube which is claimed to be one hundred times more sensitive than either the Iconoscope or the Image Dissector was developed by RCA engineers. The tube, shown in Fig. 1-6, is known as the Image Orthicon. The greater sensitivity of this tube gives it the following advantages:

1. The ability to televise scenes too dark to establish an acceptable image with other camera tubes.
2. A greater depth of field, permitting the inclusion of background that otherwise appears blurred or obscured on the receiver screen.

Physically, the tube looks like an elongated image projection tube, being approximately 15 inches long and 3 inches in diameter at the head. Electrically, the tube is divided into three parts: the image section, where the equivalent distribution of charge over a photosensitive surface is formed; a scanning section, consisting of the electron gun, the scanning beam and deflecting coils; and, finally, a multiplier section where, through a process of secondary emission, more current is generated than is contained in the returning beam. This action is closely akin to the electron multiplier contained in the Farnsworth Image Dissector. Fig. 1-14 illustrates these three sections of the Image Orthicon.

In operation, light rays from the scene to be televised are focused by an optical lens system onto a transparent photosensitive plate. At the inner surface of this plate, electrons are emitted from each point in proportion to the incident light intensity. Note that the light rays must penetrate the transparent plate to reach the photosensitive inner surface.

The emitted electron image (in which, at each point, the density of the electrons corresponds to the light at that point) is drawn to the target by a

positive wall coating. At the target, the arriving electrons produce secondary emission and thus develop a pattern of positive charges directly proportional to the distribution of energy in the arriving electron image. The target is not photosensitive, but is capable of emitting secondary electrons.

Note that by this method of forming a charge distribution on the target plate, we obtain a more intense degree of positive charge distribution than if the light rays themselves had been the activating agent, as in the Iconoscope.

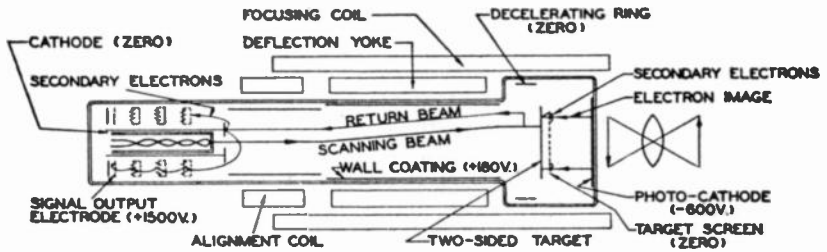


FIG. 1.14. The internal construction of the Image Orthicon.

The back of the target plate is scanned by a low-velocity electron beam. This beam is slowed down just short of the plate, and at each point gives up sufficient electrons to neutralize the positive charge at that point. The remainder of the electrons in the beam then return to an electron multiplier arrangement where several electrons are produced for each impinging electron. The result — at the output — is a current amplified many times over the current in the return beam.

It is evident that the most positive points on the plate return the least number of electrons from the original scanning beam. Hence, the voltage developed across the output load resistor is inversely proportional to the positive charge intensity on the target. As we shall see presently, this corresponds to negative phase polarity in the signal.

In order to function effectively, the two-sided target must be able to conduct electrons between its two surfaces but not along either surface. The logic of this is evident. Whatever charge appears on the one side of the target due to the focused image must likewise appear on the other side. It is this second side which is scanned and it is from here that the video signal is obtained. Hence, a conducting path must exist between the front and back sides. On the other hand, nothing must disturb the relative potential that exists throughout the charge pattern, as deposited on the front side of the target. Hence, no conduction is permissible between the various elements of any one side of the target plate. Where this occurs, the charge differences between the various points on the image disappear.

The two-sided target used in the Image Orthicon consists of a thin sheet of low-resisting glass. The resistivity between the front and back sides is sufficiently low so that if we were to place opposite charges on the sides, complete neutralization (by conduction) would occur in less than $\frac{1}{30}$ of a second. In this way, we prevent one frame from affecting the next frame, an effect which is known as "hangover."

The thin sheet of glass is about $1\frac{1}{2}$ inches in diameter. It is placed about two thousandths of an inch from a flat fine-mesh screen. The purpose of this fine-mesh screen is to collect secondary electrons that are knocked off the target when the photoelectrons impinge upon it. In order not to interfere with the oncoming photoelectrons, the mesh contains 500 to 1,000 meshes per linear inch, an open area of 50 to 75 per cent and a considerable accuracy of spacing.

The Image Dissector Camera Tube. The Iconoscope and Image Orthicon come under the heading of storage type camera tubes because the globules of the mosaic have the property of increasing their charge during the time that the electron scanning beam is acting elsewhere on the plate. In this way, each picture element, by the time it is scanned, possesses a greater charge than if it did not have the property of storage.

P. T. Farnsworth developed a camera tube that may be considered as an instantaneous scanner because storage of energy is not involved. The tube, shown in Fig. 1-7, consists of a flat photosensitive cathode located at one end of the tube. The scene to be televised is focused on this cathode by means of a lens system located outside the opposite end of the tube. When the light reaches the cathode, electrons are emitted in proportion to the amount of light striking any one point. Thus the electron distribution leaving the cathode is equivalent to the distribution of light focused by the lens.

The electron cloud is forced to move down the tube by high positive voltages applied to attracting electrodes situated at the other end of the tube. A scanning aperture (see Fig. 1-15) is also located at the other end of the tube, but its position is fixed. Here, deflecting coils move the electron cloud past the aperture and each electron, as it is swung in front of the opening, enters because of its forward motion. Notice that in this tube, the electron cloud or electron image is moved while the scanning device is stationary. In the Iconoscope, the opposite was true. As in the Iconoscope, however, the image is scanned by horizontal lines, 525 in all.

After the electrons enter the scanning aperture in turn, they hit specially constructed plates that have the property of emitting a number of electrons for each one that impinges on them from the electron cloud. These extra, or secondary, electrons thus emitted are accelerated to another plate which also gives off more electrons than strike it. The process is repeated approximately eleven times and a fairly large current is obtained at the output

A structure of this type is known as an electron multiplier; some idea of its internal structure may be gained from Fig. 1-15.

If the reader is puzzled as to what happens to the electrons in the electron cloud or image that are not, at any particular moment, being scanned, it may be stated that they are attracted to the anode coating located on the side walls. The image is not destroyed because electrons in a steady stream are continually being given off by the photocathode. Thus there is always an electron image present and the scanning aperture always has electrons

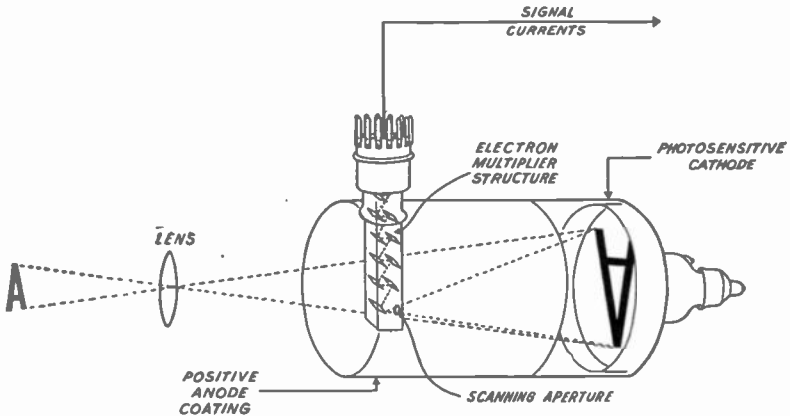


FIG. 1.15. An internal view of an Image Dissector tube.

surging in. The electrons all eventually get back to the photocathode by either of two routes. One is from the anode coating on the side walls; the other is through the scanning aperture, down the electron multiplier structure, through the load resistor and back to the cathode.

To move the electron cloud past the scanning aperture, horizontal and vertical deflecting coils are employed. Currents from timing generators control the fields set up by these deflecting coils and the entire electron image is moved back and forth and scanned. With sufficient light from the scene being televised, pictures of good quality are obtainable. The Image Dissector is employed in equipment manufactured by Capenhart-Farnsworth (a subsidiary of I.T. & T.).

All camera tubes are housed in large rectangular cases, then placed on dollies to allow the entire assembly to be moved from one position to another quickly and quietly. The necessity for employing the relatively large cases is due to the extremely small video currents generated in the camera tubes, even under the most favorable conditions. If these tiny currents were sent into the long connecting coaxial cables, they would be too small to override the inherent noise in the system by the time the transmitter was reached. To overcome this, several amplifiers are built into the camera assembly along

with the camera tube. Consequently, the small video currents are amplified immediately and then sent into the connecting transmission line.

Figs. 1-16 and 1-17 illustrate two examples of present-day television cameras. The first camera is for regular studio use; the second is designed for outdoor, on-the-spot programs.



Courtesy NBC

FIG. 1.16. A television studio camera in action.

Blanking and Synchronizing Signals. The cathode-ray beam at the receiver must follow the transmitter action at every point. For example, each time the camera-tube beam is blanked out, the same process must occur at the receiver and in the proper place. It is for this purpose that blanking signals are sent along with the video signals, those that contain the image details. These blanking pulses, when applied to the control grid of a cathode-ray tube, bias it to a large negative value, sufficient to prevent any electrons from passing through the grid and on to the fluorescent screen.

Blanking voltages, while preventing the electron beam from impinging on the fluorescent screen during retrace periods, do not cause the movement of the beam from the right- to the left-hand side of the screen, or from bottom to top. For this, another set of pulses, superimposed over the blanking signals, control oscillators at the receiver and these, in turn, control the

rupted, even while the vertical deflecting coils are bringing the electron beam to the top of the field, the long vertical pulse is broken into appropriate intervals. In this manner it is possible to send both horizontal and vertical pulses at the same time, each being accurately separated at the receiver and transferred to the proper deflection system. Greater detail is given on this point in Chapter 12. The term used for the series of synchronizing pulses that combine to make up the total vertical signal is "serrated vertical impulses." This type of wave form has been established as standard in the United States.

Under the action of the vertical deflecting coils the beam is brought to either point *A* or *E* (refer to Fig. 1-13) and then the usual camera action starts anew.

Negative and Positive Video Polarity. A closer inspection of a video signal, Fig. 1-20, reveals that of the total (100%) amplitude available, 75

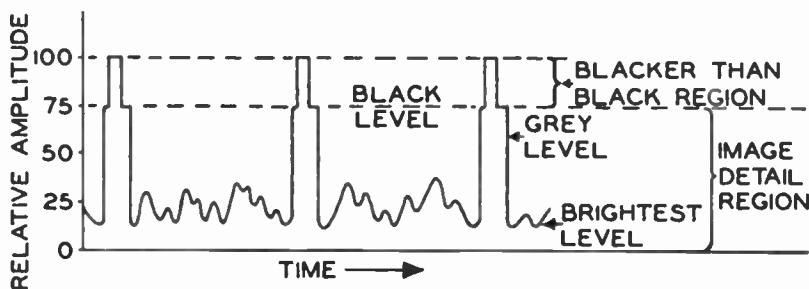


Fig. 1.20. The various proportions of a video signal.

to 80 per cent is set aside for the camera signal variations. At the level where the camera signal ceases, the blanking voltage is inserted. The remaining 20 to 25 per cent of the amplitude is reserved for the horizontal or vertical synchronizing pulses. It will be noticed that, no matter where the camera signal happens to end, the blanking level (and the synchronizing pulses) always reaches the same amplitude. This is done purposely at the transmitter, and several operations in the television receiver depend upon this fact. It must be remembered, however, that this does not necessarily have to be, but is specifically employed because of resulting simplicity at the receiver.

Fig. 1-20 illustrates the form of the video signal as it is used in the United States. From the relative polarity marked on the side (or vertical) scale, it is seen that the brightest portions of the camera signal cause the least amount of current to flow, or the voltage has the least amplitude. This is exactly opposite to the action at the Iconoscope as explained previously in the chapter. The signal voltage (or current) values have been completely reversed. The blanking voltage, which should be more negative than any

portion of the camera signal, is actually more positive. And the synchronizing pulses give the largest voltage and current of all.

Transmitting the signal in this form is known as negative picture transmission, and the picture is said to be in the negative picture phase. If the video signal were reversed so that it assumed the form of Fig. 1-21, it would be called the positive picture phase and, if transmitted, would be known as positive transmission. In the United States, negative R.F. transmission is employed, although in England the other form is preferred. It is claimed

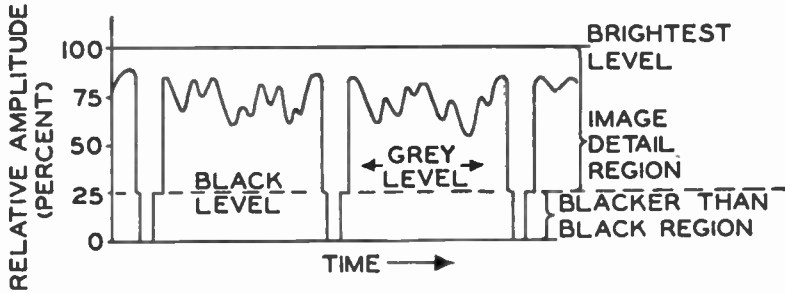


FIG. 1.21. The form of the video signal in positive transmission.

here that less interference is visible on the viewing screen with negative transmission, and better all-around reception is obtained under adverse conditions. Be that as it may, one standard has been decided upon, and all receivers must be constructed to receive this signal. If a receiver designed for negative picture phase signals receives a positive picture phase signal, all the image light values are reversed on the viewing screen. The bright portions would appear dark and vice versa. The result would be similar to a photographic negative, in which the values are likewise reversed.

In the receiver, before the video signal is applied to the control grid of the cathode-ray tube, the signal must be transferred into the proper or positive picture phase. The grid of the cathode-ray tube is then biased by enough negative voltage so that, whenever the blanking voltage section of the signal does act at the grid, the electron beam is automatically prevented from reaching the fluorescent screen. With the positive picture phase, the camera signal voltages are all more positive than the blanking pulse and, on these portions of the video signal, the electron beam is permitted to impinge on the screen with varying amounts of electrons. A bright spot in the received image causes the grid to become more positive than when the voltage of a darker spot is applied. More electrons in the beam mean more light emitted at the screen, and the various shades and light gradations of the image are formed by different voltages.

The purpose of the blanking voltage in the video signal is to prevent the electron beam from reaching the fluorescent screen. This is well known

by now. The point in the video wave where the blanking signal is located occurs in the region where the currents corresponding to the very dark portions of the image are found. By the time the blanking voltage acts at the control grid of the viewing tube, the beam is entirely cut off and nothing appears on the screen. The blanking level could then properly be called the black region, because nothing darker appears on the fluorescent screen. By nothing darker, we mean no light at all.

Now, consider the video signal of Fig. 1-21. With the blanking level we find the synchronizing pulses. When applied to the viewing tube control grid along with the rest of the wave, the pulses drive the grid to a negative voltage even greater than cut-off. The pulse region, for this reason, is labeled as blacker than black, because the position of the blanking signal has been called black. The unwanted synchronizing pulses that ride through the video amplifiers with the necessary video signal need not be removed because they do not interfere, in any way, with the action of the control grid at the cathode-ray tube. As will be shown presently the complete video wave is applied, after the detector, to the synchronizing and video amplifier circuits simultaneously. The synchronizing clipper tube permits merely the pulses to pass through, whereas the video amplifiers allow the entire signal to pass.

Why Television Requires Wide Frequency Bands. In dealing with television receivers, it will be found that extensive use is made of wide-band amplifiers designed to receive signals extending over a band width of 6 megacycles (mc). The different forms these amplifiers may assume and their characteristics are discussed in later chapters; however, the reason for the extremely wide band width may be appreciated now.

In the foregoing paragraphs on television images, it was brought out that the more elements in a picture, the finer the detail that could be portrayed. The picture could also stand closer inspection before it lost its smooth, continuous appearance. Each thirtieth of a second, 525 lines are scanned, or a total of 15,750 lines in one complete second. If each horizontal line contains 700 separate elements, then $15,750 \times 700$ or 11,025,000 elements or electrical impulses are transmitted each second. In order to attain full advantage of the use of this number of elements, it is first necessary to determine what relationship exists between the two quantities, number of elements and band width.

Consider, for example, that the mosaic plate in the Iconoscope is broken up into a series of black and white dots, each dot representing one element. The resulting pattern is shown in Fig. 1-22A. As the scanning beam passes over each element in turn, a pulse of current flows every time a white dot is reached, for this element has a large deficiency of electrons. At the next element, the current drops to zero, for theoretically a black dot

represents an element that has received no light at all and hence requires no additional replacement of electrons. In one complete horizontal line, the electric pulses of current would have the shape shown in Fig. 1-22B.

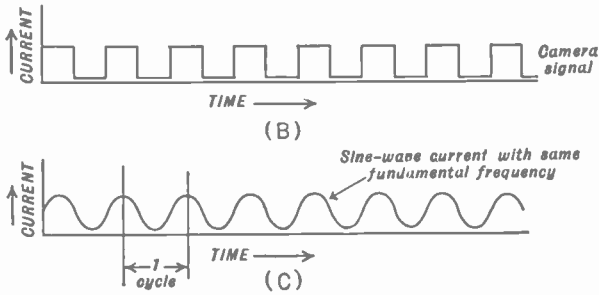
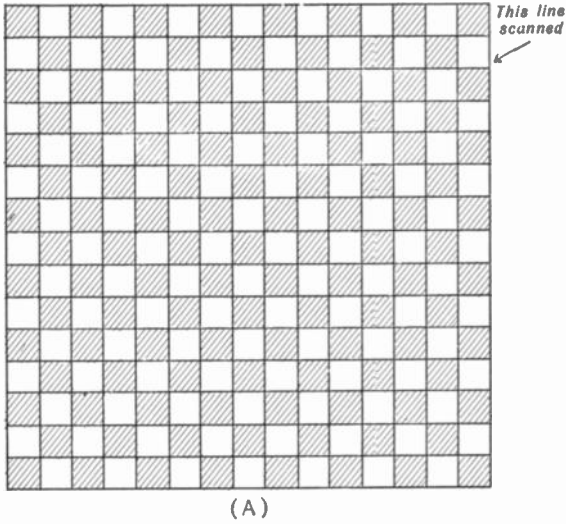


FIG. 1.22. The basic relationship between the number of elements in an image and the width of the frequency band required.

If we combine one maximum point in the wave with its succeeding minimum point, we obtain one complete cycle. The same situation prevails in any sine wave. (See Fig. 1-22C.) Since each white dot represents a maximum point and each black dot a minimum point, then by taking the total number of white and black dots on a line, and dividing their sum by 2, we obtain the number of cycles the current goes through when one horizontal line is scanned. With 700 elements (dots, in this case) on a line, a fundamental frequency of 350 cycles is generated.

Under present standards, 525 lines are scanned in $\frac{1}{30}$ of a second, or a

total of 15,750 in one second. Employing 700 elements per line, 11,025,000 picture elements are sent each second. This, from our analysis, results in a frequency of $\frac{11,025,000}{2}$ cycles per second or 5.51 mc. In actual practice, a band width of 4 mc is allowed. Thus, for the video section alone, this extremely large band width must be passed by all the tuned circuits of the television receiver.

The above situation would seldom, if ever, be found in practice. However, the figures obtained by this reasoning yield results that have been found satisfactory and so the method, from this viewpoint, is justified.

While 4 mc are required to accommodate the video information alone, the band width set in practice is 6 mc. Of the extra 2 mc, the F-M audio carrier uses 50 kilocycles (kc). Apparently considerable band width is not utilized. The reason for the extra space is found in the process whereby the television video carrier is generated.

On ordinary broadcast frequencies, it is common knowledge that most stations occupy a 10-kc band width, or ± 5 kc about the carrier position. Thus, if a station is assigned to the frequency of 700 kc, it transmits a signal that occupies just as much frequency space on one side of 700 kc as on the other. Under existing F.C.C. regulations, the maximum deviation is 5 kc (or 5,000 cycles) on either side of the carrier position of 700 kc. In radio language, we say that these side frequencies are side bands and, for the present illustration, each side band may have a maximum deviation of 5 kc about the mean or carrier position. The information of the signal is contained in the side bands, since they are not generated until speech or music (or other sounds) are projected into the microphone. At the receiver, the variations in the side bands are transformed into audible sounds and heard by the radio listener.

It can be shown that those side bands that are generated having frequencies higher than the carrier frequency contain the same information as the side bands with frequencies lower than the carrier. In other words, if we eliminated one set of side bands (either above or below the carrier), we could still obtain all the necessary information at our receiver. The only reason one side band is not eliminated is due to purely economical reasons. A transmitter naturally generates both side bands and it is cheaper to transmit both rather than try to eliminate one by expensive and complicated filters. However, single side-band transmission does exist for certain communication facilities.

Now, let us turn our attention to the video signal. It is generated by fundamentally the same type of apparatus that is employed at the sound broadcast frequencies. Since 4 mc are needed for the picture detail, a signal would be generated that extended for 8 mc, or ± 4 mc about the carrier.

And this does not include the sound. An 8-mc band is undesirable because of the ether space occupied and the difficulties inherent in transmitting a signal of this band width. Hence, the necessity arises for removing one side band, since only one is required.

The undesired side band is removed by filters that follow the last amplifier of the television transmitter. But filters are not easily constructed that will sharply cut off one side band completely and leave only the desired one. Furthermore, in the process of elimination, nothing must occur that would change the amplitude or phase of any of the components in the desired side band. As a compromise arrangement, most, but not all, of one side band is removed and in this way the remaining side band is least affected by the filtering. Thus part of the 2 mc (of the total 6-mc band width) is occupied by what may be called the remnants of the undesired side band. This method is known as "quasi-single-side band" or "vestigial-side band" operation.

In Fig. 1-23A is the television video signal as it appears with both side bands present, and Fig. 1-23B shows it as it appears after passage through filters that partially remove one side band. The carrier's frequency is found 1.25 mc above the low-frequency edge of the television signal. Then for 4 mc above this, we have the television video signal with the desired picture information. This is all indicated in Fig. 1-23. A 0.5-mc. band width separates the high-frequency edge of the video signal and the F-M carrier. The space is left for the purpose of preventing undesirable interaction between the two, such as cross-modulation, which would lead to distortion of the video signal. In this manner the allotted 6 mc are distributed.

Effect of Loss of Low and High Video Frequencies. While uniform response over a 4-mc band may be required in the picture I.F. and video amplifiers, this is not easily attained in practice. Special circuit designs must be resorted to which are more fully explained at their appropriate places in later chapters. For the moment, it is only necessary to point out the effects of poor response at either the high- or low-frequency ends of the band.

In the preceding analysis we have seen that a greater number of elements required a greater band width if advantage was to be taken of the increase. Since detail is determined mainly by the number of very small elements, any decrease in the response at the higher frequencies will result in less fine detail available at the receiving cathode-ray screen. The picture will lose some of its crispness and may even appear somewhat blurred if the high-frequency response is degraded enough. In commercial television receivers, a video passband of from 3.3 to 4.0 mc is generally considered good, while anything below 3.0 mc is not too desirable.

At the low-frequency end of the band, poor frequency response results in

obliterating the slow changes that occur in background shading. However, it is possible partially to counteract bad effects caused by the poor response with the manually adjusted brightness control to be described later.

Frequency Allocations. With a maximum band width set at 6 mc, it becomes obvious that in order to operate even as few stations as five in any

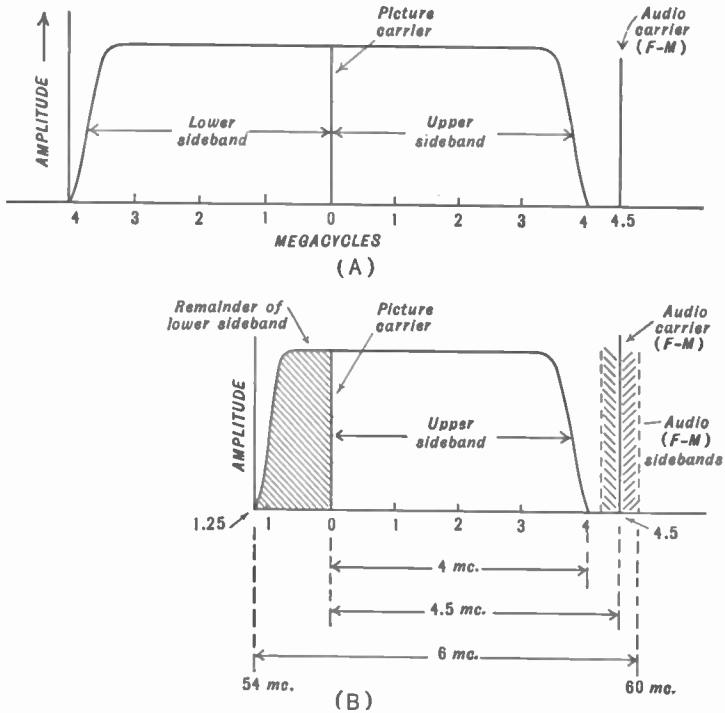
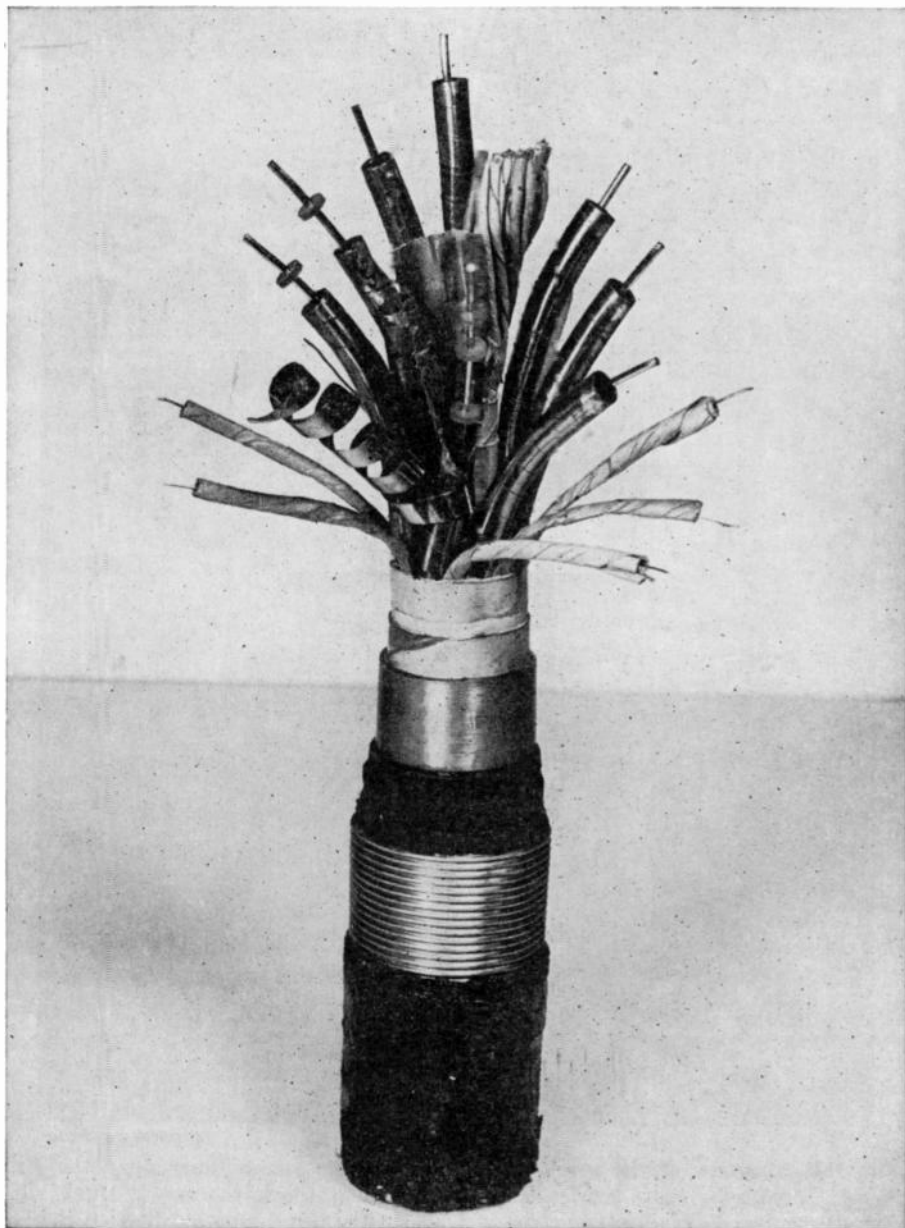


FIG. 1.23. Double-side band (A) and vestigial-side band (B) transmission of television signals. Method B is standard in the United States and represents the signal as sent.

one area, a band 30 mc wide must be provided. With most of the lower frequencies already occupied by existing services, television was allotted space at the high frequencies, from 54 mc up. The advantage of using the higher frequencies lies in the vast amount of free ether space that is available. This permits extensive expansion, which is certain to occur when television sets become as numerous as the present A-M receivers. The disadvantages of using the higher frequencies result from the semi-optical behavior of radio waves in this range and the engineering difficulties in building stable high-powered equipment at short wavelengths.

The latter problem is fast dissolving in the many research laboratories in this country and abroad. New tubes, more easily built components, bet-



Courtesy A.T.T.

FIG. 1.24. A fanned-out section of an 8-tube transmission cable.



Courtesy Bell System

FIG. 1.25. Typical of the seven intermediate stations on hilltops along the Bell System radio relay route between New York and Boston is the one at Birch Hill, near Patterson, N.Y. Electromagnetic lenses in the horns seen on the roof receive and beam long distance communication between the two cities.

ter design and many other factors have combined to give good results with receivers and transmitters at the high and ultra-high frequencies.

The other problem involving the short range in which high-frequency radio waves can be received means that eventually numerous relay stations will have to be installed throughout the country for wide population coverage. Such work is now in full swing and consists of a combination of coaxial cable and microwave relays. A section of the transmission cable (Fig. 1-24) actually being used contains eight independent concentric cables — together with 49 paper insulated wires for testing and maintenance purposes. A pair of the concentric cables, when fully terminated by the proper repeater stations, is capable of handling about 600 simultaneous telephone conversations, or one television program in each direction. Each of these copper tubes is about the diameter of a man's small finger. Running down the center of each tube is a copper wire about the size of a pencil lead, held in place by round insulating discs. Both the copper tube and the wire inside have the same axis, hence the name of concentric or coaxial.

Microwave systems, which may be used for long-distance telephone calls as well as television, make use of very short radio waves (in the neighborhood of 4,000 megacycles). These microwaves are free from static and most man-made interference and shoot off into space instead of following the earth's curvature. Therefore, the waves are gathered into a beam and aimed at towers, usually located atop high buildings or hilltops. Microwaves are focused in narrow beams from one point to another by huge horn-shaped antennas. Microwaves can be beamed over distances of 30 to 40 miles, just as long as there is no interference in its line-of-sight path. Typical of the microwave relay stations on hilltops is the one shown in Fig. 1-25. This is located along the Bell System radio relay route between New York and Boston and is the one at Birch Hill, near Patterson, N. Y.

Fig. 1-26 shows the Bell System's coaxial cable and radio relay system as it exists now. While not shown separately, some portions of the routes use microwave relay and other sections use coaxial cable. Thus, for example, the link between Boston and New York consists entirely of microwave relay stations, while from New York, southward and westward, there is coaxial cable. In addition, there is a microwave relay system between New York and Chicago to supplement the coaxial cable between these two cities. Other systems besides the Bell System are in operation, but the latter is the most extensive.

The allocation of channels for television broadcasting, as it now stands, is as follows:

TELEVISION SIMPLIFIED

<i>Channel</i>	<i>Megacycles</i>	
2	54-60	} lower VHF band
3	60-66	
4	66-72	
5	76-82	
6	82-88	
7	174-180	} upper VHF band
8	180-186	
9	186-192	
10	192-198	
11	198-204	
12	204-210	
13	210-216	

UHF Band

Channel No.	Channel Freq. (mc)	Channel No.	Channel Freq. (mc)	Channel No.	Channel Freq. (mc)
14	470-476	38	614-620	62	758-764
15	476-482	39	620-626	63	764-770
16	482-488	40	626-632	64	770-776
17	488-494	41	632-638	65	776-782
18	494-500	42	638-644	66	782-788
19	500-506	43	644-650	67	788-794
20	506-512	44	650-656	68	794-800
21	512-518	45	656-662	69	800-806
22	518-524	46	662-668	70	806-812
23	524-530	47	668-674	71	812-818
24	530-536	48	674-680	72	818-824
25	536-542	49	680-686	73	824-830
26	542-548	50	686-692	74	830-836
27	548-554	51	692-698	75	836-842
28	554-560	52	698-704	76	842-848
29	560-566	53	704-710	77	848-854
30	566-572	54	710-716	78	854-860
31	572-578	55	716-722	79	860-866
32	578-584	56	722-728	80	866-872
33	584-590	57	728-734	81	872-878
34	590-596	58	734-740	82	878-884
35	596-602	59	740-746	83	884-890
36	602-608	60	746-752		
37	608-614	61	752-758		

When television allocations were first made following the war, thirteen channels were assigned to immediate television broadcasting. These included the 12 VHF channels listed above, plus channel 1, at 44-50 mc. At the same time, non-government fixed and mobile radio services were permitted to share television channels 1 through 5 and 9 through 13. However, it didn't take long to demonstrate that this sharing was impractical

because of serious interference to television reception. It was then decided to allot channel 1 entirely to non-government fixed and mobile services and to eliminate sharing of television channels 1 through 5 and 9 through 13. Channels 8 and 9 are shared with government fixed and mobile services, but since these latter services function infrequently, no serious interference is obtained.

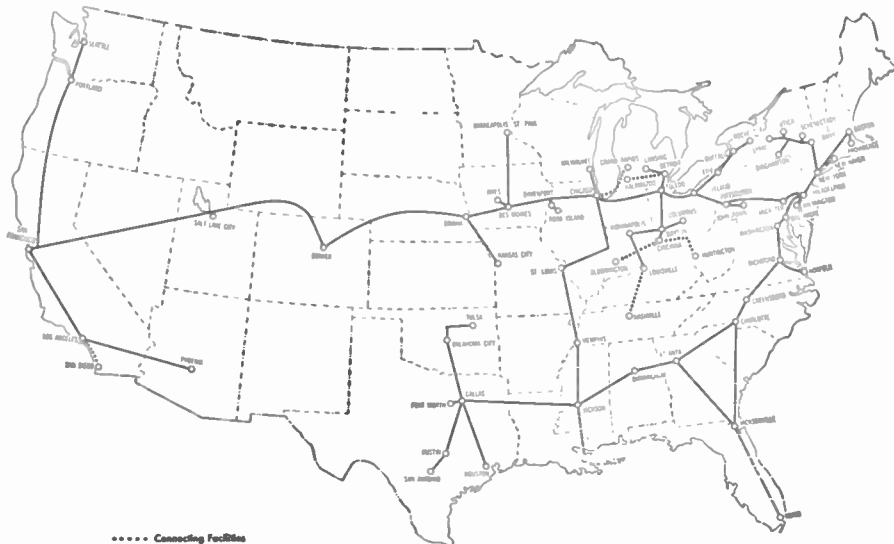


FIG. 1.26. The Bell System coaxial cable and microwave relay routes.

The UHF band, containing channels 14 to 83, was opened to commercial broadcasting on July 1, 1952. The allocation of stations within this region is now proceeding at a rapid rate, and there should be several thousand stations in operation within the next few years.

F-M for Audio Transmission. When standards were established for commercial television broadcasting, it was directed that F-M was to be utilized for the audio portions of the television signal, and amplitude modulation for the video portion of the signal. F-M for the audio offers noise-free reception and higher fidelity due to the possible use of audio frequencies up to 15,000 cycles. This type of modulation, however, has been found to give poorer results for the video signal, and amplitude modulation has been retained for two reasons.

Let us consider, for example, an antenna receiving two waves from the same transmitter. One ray traveled directly from transmitter to receiver, while the other ray (which we will call the reflected ray) arrived at the receiver by a longer, more indirect path. This could have occurred if the second ray was moving in some other direction, hit an obstacle in its path, and

was reflected toward the receiving antenna. Because the reflected ray traveled a longer path to reach the receiving antenna, it arrived some small fraction of a second after the direct ray. During the interval between received rays, the electron beam has traveled a small distance across the fluorescent screen. The end result is two similar images, slightly displaced from each other. This condition is known as "ghosts" and occurs when A-M is used. The same situation with F-M also produces a ghost image, but the two contrasts are more prominent and prove more distracting than the ghost images of A-M.

Another advantage of A-M over F-M for the video signal is the better synchronizing action observed. When there are several paths that a signal may follow in reaching the receiving antenna, or when there are other types of interference, there is less tendency for the synchronizing pulses to become obliterated in the A-M signal. Loss of the synchronizing pulses means no control of the motion of the electron beam as it moves across the screen. The image, under this condition, would appear with streaks in it at points where the synchronizing action was lost.

The distance over which the signal can be transmitted directly is the same, whether F-M or A-M is employed. Frequency, and only frequency, is the determining factor. As explained in Chapter 2, the usable range at these high frequencies is governed by the height of the receiving and transmitting antennas above the ground.

QUESTIONS

1. Why do television receivers have more panel controls than sound receivers?
2. What is a "Fine Tuning" control? Why is it necessary in most television receivers?
3. What are some of the desirable characteristics that an image should possess? Explain each briefly.
4. *a.* Draw the block diagram of a conventional television receiver. *b.* Where do Intercarrier and conventional-type television receivers differ?
5. Of what does a complete television signal consist?
6. State briefly the function of each stage in a television receiver.
7. What is the purpose of synchronizing pulses?
8. Name three types of television cameras. Explain briefly the operation of one of these tubes.
9. Define scanning. What type of scanning is standard today? Why?
10. Define frame frequency, field frequency, and line frequency. Give values for each.
11. What would happen to the receiver image if blanking voltages were not employed in the video signal?
12. Discuss negative and positive picture polarity.
13. How do English and American television signals differ? What effect does this have on the reproduced image? What would happen if an American-made receiver were to receive English television signals?

14. Explain why video signals require wide frequency bands.
15. What effect does the bandwidth of the television signal have on the form of the final transmitted signal? Explain.
16. List all the VHF channels currently assigned to television broadcasts, giving frequencies in each instance. How many UHF channels are there?
17. What is the visual effect of a loss of high video frequencies? Low video frequencies?
18. Why is F-M employed for sound transmission and A-M for video signal transmission?
19. What relationship exists between flicker and the method employed for scanning?
20. Discuss in detail the motion of the electron beam in interlaced scanning.
21. How is horizontal synchronization maintained while the vertical pulses are active? What is this called?
22. What is the horizontal scanning frequency? The vertical scanning frequency? Explain how each figure is arrived at.
23. What is meant by vestigial side-band transmission? How does this differ from the type of transmission employed in standard broadcast practice?
24. Are relay stations necessary for television transmission? Where are relay stations useful?
25. List the video front-panel controls that are generally used in commercial television receivers. State their function briefly.
26. Why must amplifiers be included with a camera unit?
27. Draw two lines of a complete video signal including two blanking and sync pulses. Draw the video in positive picture phase.
28. What is an electron multiplier? Where is it used?
29. What is the time, in microseconds, for one complete horizontal line?
30. Explain the difference between camera tubes employing the storage and non-storage principles.

Chapter 2

HIGH-FREQUENCY WAVES AND THE TELEVISION ANTENNA

The antenna for a television receiver requires much more attention and care, especially with regard to placement, than that used with the ordinary sound receiver. In order to obtain a clear, well-formed image on the cathode-ray tube screen, it is absolutely necessary that:

1. The maximum signal strength be developed at the antenna.
2. The signal be received from one source, not several.
3. The antenna be placed well away from man-made sources of interference.

In ordinary sound receivers, a certain amount of interference and distortion is permissible. If not excessive, reception of the broadcast is satisfactory. For television, however, the standards are more severe, and added precautions must be taken to guard against almost every type of interference and distortion. Hence, the need for more elaborate antenna receiving systems.

The position of the antenna must be chosen carefully, not only for additional signal strength, but also because of the appearance of so-called "ghosts" on the image screen which are due to the simultaneous reception of the same signal from two or more directions. For an explanation of this form of interference, refer to Fig. 2-1, in which a television dipole antenna is receiving one signal directly from the transmitting tower, while another ray strikes the same antenna after following a longer, indirect path. Reflection from a building or other large object could cause the indirect ray to reach the antenna.

Because of the longer distance the reflected ray travels, it will arrive at some small fraction of a second later than the direct ray. In sound receivers, the ear does not detect the difference. On a television screen, the scanning beam has traveled a small distance by the time the reflected ray arrives at the receiver. Hence, the image contained in the reflected ray

appears on the screen displaced some small distance from similar detail contained in the direct ray. The result is shown in Fig. 2-2. When the effect is pronounced, a complete double image is obtained and the picture appears

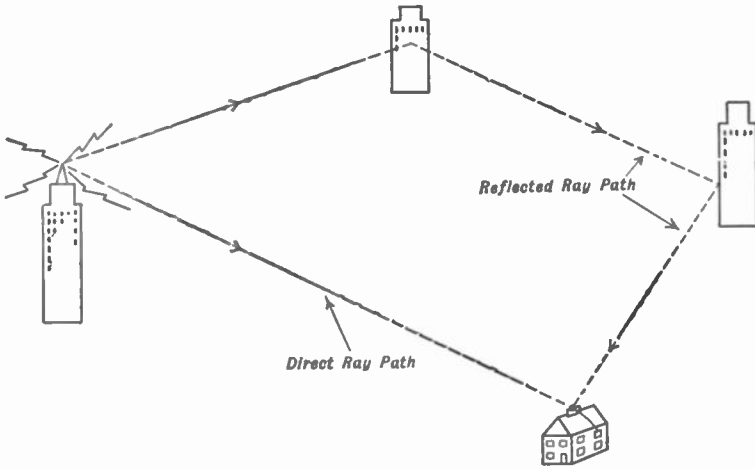


FIG. 2.1. The reflected ray, along with the direct ray, arrive at the receiving antenna to form double images, called "ghosts."

blurred. To correct this condition, it is necessary to change the position of the antenna until only one ray is received. The antenna should not be turned to favor the reflected signal unless it is impossible to obtain a clear image with the direct ray. The properties of reflecting surfaces change daily, and there is no certainty that a good signal will always be received.

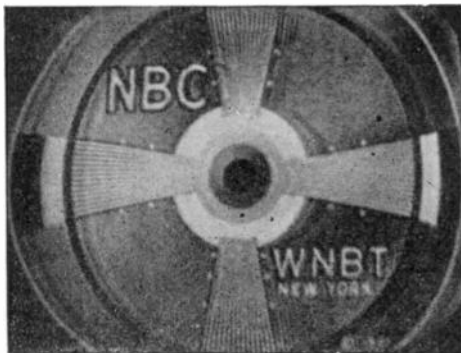


FIG. 2.2. A "ghost" image on a television viewing screen.

The placement of the antenna is generally the most difficult operation of a television installation. To obtain maximum results, it is necessary for the radio serviceman or other person erecting the receiving antenna to have a good knowledge of the behavior of radio waves at the high frequencies.

Radio Wave Propagation. Transmitted radio waves at all frequencies may travel in either of two general directions. One wave closely follows the surface of the earth, whereas the other travels upward at an angle which is dependent on the position of the transmitting radiator. The former is known as the ground wave, the latter as the sky wave. At the low frequencies, up to approximately 1,500 kc, the ground wave attenuation is low, and signals travel for long distances before they disappear. Above the broadcast band, the ground wave attenuation increases rapidly, and all extensive communication is carried on solely by means of the sky wave.

The sky wave leaves the earth at an angle that may have any value between 3° and 90° and travels in almost a straight line until the ionosphere is reached. This region begins at a distance of about 70 miles above the surface of the earth, and within this area are found large concentrations of charged gaseous ions, free electrons, and uncharged or neutral molecules. The ions and free electrons act on all passing electromagnetic waves and tend to bend these waves back to earth. Whether the bending is complete (and the wave does return to the earth) or only partial depends on several factors:

1. The frequency of the radio wave.
2. The angle at which it enters the ionosphere.
3. The density of the charged particles (ions and electrons) in the ionosphere at that particular moment.
4. The thickness of the ionosphere at the time.

Extensive experiments indicate that, as the frequency of a wave increases, a smaller entering angle is necessary in order for complete bending to occur.

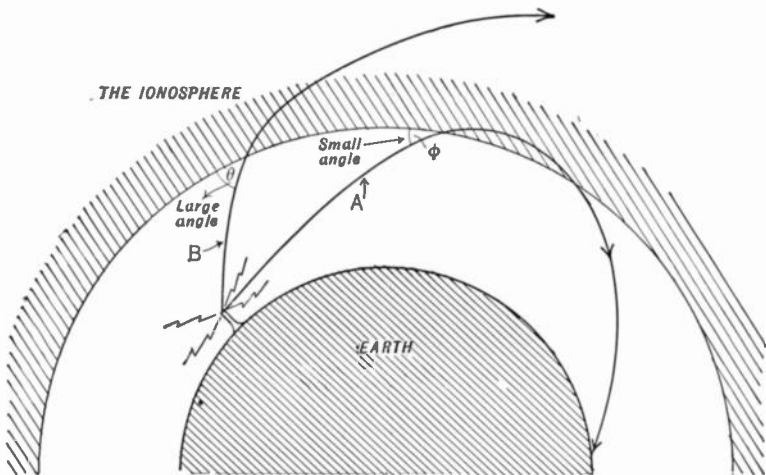


FIG. 2.3. At the higher frequencies, a radio wave must enter the ionosphere at small angles if it is to be returned to earth.

As an illustration of this, consider the two high-frequency waves, *A* and *B*, shown in Fig. 2-3. Wave *A* enters the ionosphere at a small angle and, hence, little bending is required to return it to earth. Wave *B*, subject to the same amount of bending, is headed outward, however, because its initial entering angle was too great. Naturally, this latter wave would not be useful for any communication purposes.

By raising the frequency still higher, the maximum incident angle at the ionosphere becomes smaller, until finally a frequency is reached where it becomes impossible to bend the wave back to earth, no matter what angle is used. For ordinary ionospheric conditions, this occurs at about 35 to 40 mc. Above these frequencies, the sky wave is useless so far as radio communication is concerned. Only the direct ray is of any use. Television bands, starting above 40 mc, would fall into this category. By direct ray (or rays), we mean the radio waves that travel in a straight line from transmitter to receiver. Ordinarily, at lower frequencies, the radio waves are sent to the ionosphere and, from there, to the receiver at some distant point. With high frequencies, the ionosphere is no longer useful, so the former sky waves must be concentrated into a path leading direct to the receiver. If not intercepted by the receiver, they finally hit the ionosphere and are lost. It is this restriction to the use of the direct ray that limits the distance in which high-frequency communication may take place.

There are, at times, unusual conditions present when the concentrations of charged particles in the ionosphere increase sharply. At these times, it is possible to bend radio waves of frequencies up to 60 mc. The exact time or place of these phenomena cannot be predicted and hence are of little value for commercial operation. They do explain to some extent the distant reception of high-frequency signals that may occur.

Line-of-Sight Distance. At the frequencies employed for television, reception is possible only when the receiver antenna directly intercepts the signals as they travel away from the transmitter. These electromagnetic waves travel in essentially straight lines, and the problem resolves itself into

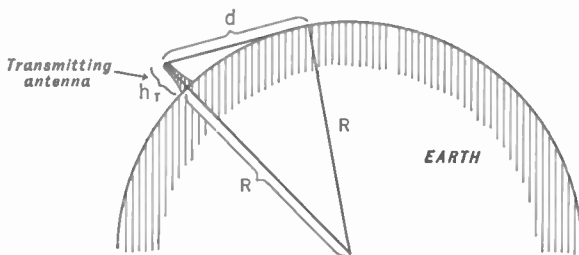


FIG. 2.4. Illustrating how to compute the line-of-sight distance for high-frequency radio waves.

finding the maximum distance at which the receiver can be placed from the transmitter and still have its antenna intercept the rays. This distance may be computed as follows.

In Fig. 2-4, let the height of the transmitting antenna be called h_t , the radius of the earth R , and the distance from the top of the antenna to the horizon d . This gives us a right triangle. From elementary geometry it is possible to write down the following equation

$$(R + h_t)^2 = R^2 + d^2 = R^2 + 2Rh_t + h_t^2$$

h_t is very small compared with the radius of the earth and the h_t^2 term may be neglected. This leaves

$$d^2 = 2Rh_t$$

The value of R is approximately 4,000 miles. Substituting this value in the above equation, and changing h_t from units of miles to feet, we obtain

$$d = 1.23\sqrt{h_t}$$

where d is in miles, h_t in feet. The relationship between d and h_t for various values of h_t has been put into graph form in Fig. 2-5.

The coverage for any transmitting antenna will increase with its height. The number of receivers capable of receiving the signals would likewise increase. This accounts for the placement of television antennas atop tall buildings (for example, the Empire State Building, New York City) and on high plateaus.

The signal range thus computed is from the top of the transmitting antenna to the horizon at ground level. By placing the receiving antenna at some distance in the air, it should be possible to cover a greater distance before the curvature of the earth again interferes with the direct ray. Such a situation is depicted in Fig. 2-6. By means of simple geometrical reasoning, the maximum distance between the two antennas now becomes

$$d = 1.23(\sqrt{h_t} + \sqrt{h_r})$$

where h_r is the receiving antenna height in feet.

Unwanted Signal Paths. While the foregoing computed distances apply to the direct ray, there are other paths that waves may follow from the transmitting to the receiving antennas. Each of these other rays is undesirable as they tend to distort and interfere with the direct-ray image on the screen. One method, by reflection from surrounding objects, has already been discussed. Another ray may arrive at the receiver by reflection from the surface of the earth. This path is shown in Fig. 2-7. At the point where the reflected ray impinges on the earth, phase reversals up to 180° have been found to occur. This phase shift thus places a wave at

the receiving antenna which generally acts against the direct ray. The overall effect is a general lowering of the resultant signal level and the appearance of annoying ghost images.

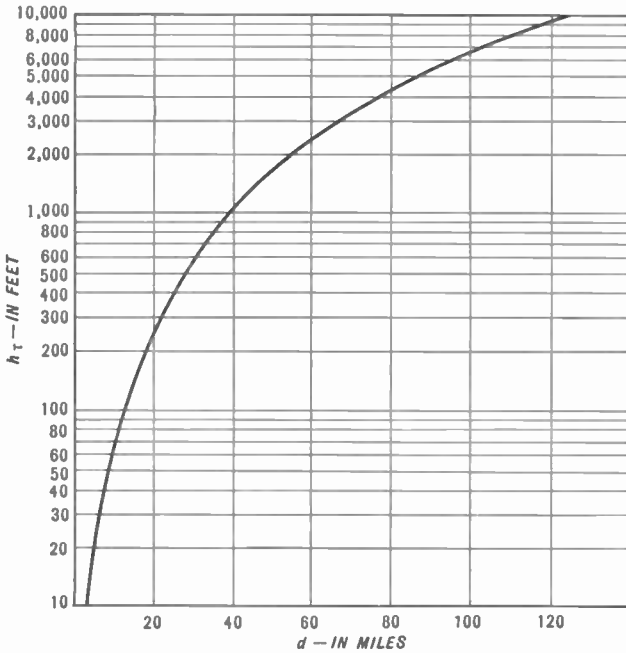


FIG. 2.5. The relationship between the height of the transmitting antenna (in feet) and the distance in miles from the antenna that the ray may be received.



FIG. 2.6. Increasing the line-of-sight distance from the receiving antenna to the transmitter by raising both structures as high as possible.

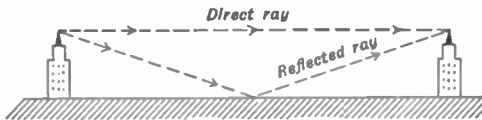


FIG. 2.7. The reflected radio wave, arriving at the receiving antenna after reflection from the earth, may lower the strength of the direct ray considerably.

However, there are compensating conditions acting against the decrease due to the ground reflected ray. One is the weakening of the wave strength by the absorption at the point where it grazed the earth. The other results

from the added phase change (not that just mentioned) arising from the fact that the length of the path of the reflected ray is longer than the direct ray path. Thus there is a ground phase shift plus whatever else may have been added because of the longer distance. All combine to lower the direct ray strength less than we would at first expect.

It has further been observed that the received signal strength increases with the height of either or both antennas. At the same time, a decrease in

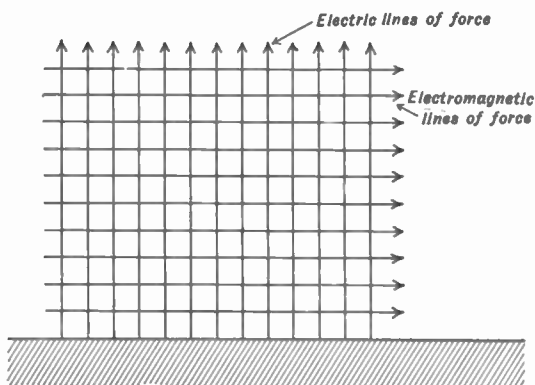


FIG. 2.8. The components of an electromagnetic wave. The wave travels at right angles to these lines of force. In this illustration, direction of travel is forward, out of the paper.

noise pick-up occurs. For television signals, this is most important. Placement of the antenna and utilization of its directive properties will help in decreasing (and many times eliminating) all but the desired direct wave.

Wave Polarization. The height of the antenna is important, but the manner in which it is held, either vertically or horizontally, must also be considered. The position of the antenna is affected by the nature of the electromagnetic wave itself.

All electromagnetic waves have their energy divided equally between an electric field and a magnetic field. In free space these fields are at right angles to each other. Thus, if we were to visualize these fields and represent them by their lines of force, the wave front would appear as in Fig. 2-8. The fields represent the wave, the arrows the direction in which the forces are acting. The mode of travel of these waves in free space is always at right angles to both fields. As an illustration, if the electric field lines are vertical and those of the magnetic field are horizontal, the wave travels forward.

In radio, the sense of a radio wave has been taken to be the same as the direction of the electric lines of force. Hence a vertical antenna radiates a vertical electric field (the lines of force are perpendicular to the ground),

and the wave is said to be vertically polarized. A horizontal antenna radiates a horizontally polarized wave. Experience has revealed that the greatest signal is induced in the receiving antenna if it has the same polarization (is held in the same manner) as the transmitting antenna.

Concerning the relative merits of horizontal versus vertical polarization, Dr. George H. Brown has found that, for antennas located close to the earth, vertically polarized rays yield a better signal. On raising the receiving antenna about one wavelength above ground, this difference generally disappears and either type may be employed. Further increase in height, up to several wavelengths, has shown that the horizontally polarized waves give a more favorable signal-to-noise ratio and are to be desired. In television, the wavelengths are short and the antennas are placed several wavelengths in the air. Hence, horizontally polarized waves have been taken as standard. All television receiving antennas are mounted in the horizontal position.

Tuned Antennas.* The need for good signal strength at the antenna has led to the general use of tuned antenna systems. A tuned antenna, which is a wire cut to the necessary length, is equivalent in its properties to any resonant circuit. The radio waves passing by the antenna will induce voltages along the wire. For equally powered radio waves, the maximum voltage is developed in the antenna when its resonant frequency is equal to that of the passing wave. A large signal at the antenna means a greater input to the receiver.

Half-Wave Antennas. An ungrounded wire, cut to one-half the wavelength of the signal to be received, represents the smallest length of wire that can be made to resonate at that frequency. The half-wavelength antenna is the most widely used since it represents the smallest antenna for its frequency and consequently requires the least amount of space. In troublesome areas it may be necessary to erect more elaborate arrays possessing greater gain and directivity than the simple half-wave antenna. They are, however, more costly.

A simple half-wave antenna is erected and supported as indicated in Fig. 2-9. Metallic rods are used for the antenna itself, mounted on the supporting structure and placed in a horizontal position (parallel to the ground). Each of the rods is one-quarter of a wavelength long, the total equal to the necessary half wavelength. In this arrangement, which is also known as a dipole antenna, the transmission lead-in wire is connected to the rods, one wire of the line to each rod. The line then extends to the receiver. Care must be taken to tape the line at several points to the supporting mast so that it does not interfere with the operation of the antenna. Taping also prevents the line from flapping back and forth in the wind. Any such motion could weaken the connections made at the rods.

* UHF antennas will be considered separately in Chapter 17.

When the properties of a dipole antenna are investigated, it is found that signals are received with greatest intensity when the rods are at right angles to the direction of the signal. This is illustrated in Fig. 2-10A. On the other hand, signals approaching the antenna from either end are very poorly received. To show how waves at any angle are received, the graph of Fig. 2-10B is commonly drawn. It is an overall response curve for a dipole antenna.

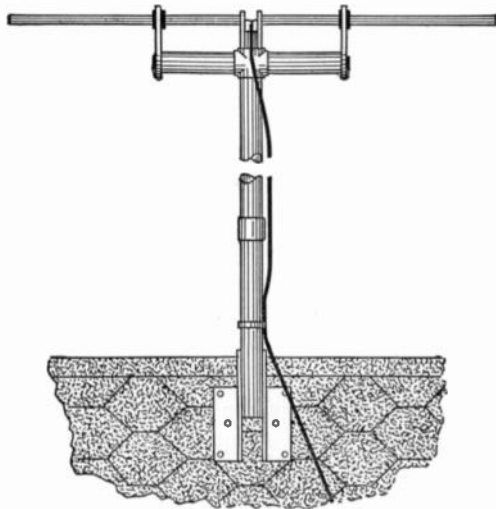


FIG. 2.9. Dipole antenna assembly used extensively for television receivers.

From the diagram, with the placement of the antenna as shown, the strongest signal would be received from direction *A*. As the angle made with this point is increased, the strength of the received signal decreases, until at point *B* (90°) the received signal voltage is at a minimum (or zero). The reader can determine the reception for waves coming in at other angles by inspection of the graph. Notice that good signal strength is obtained from two directions and, because of this, the dipole may be called bi-directional. Other systems can be devised that are uni-directional, non-directional or that have almost any desired properties. For each system, a response curve would quickly indicate its properties in any direction.

As stated, an antenna must be tuned in order to have the strongest signal develop along its length. Hence it becomes necessary to cut the wires (or rods) to a specific length. The length will vary with each different frequency, longer at the lower frequencies and shorter at the higher frequencies. It might be supposed, then, that a television set capable of receiving signals with frequencies ranging from 54 to 88 mc would need several antennas, one

for every channel. It is not necessary, however, to go to such extremes; in practice, one antenna is sufficient, if tuned to a middle frequency.

Antenna Length Computations. With the foregoing range of frequencies, a middle value of 65 mc might be chosen. While an antenna cut

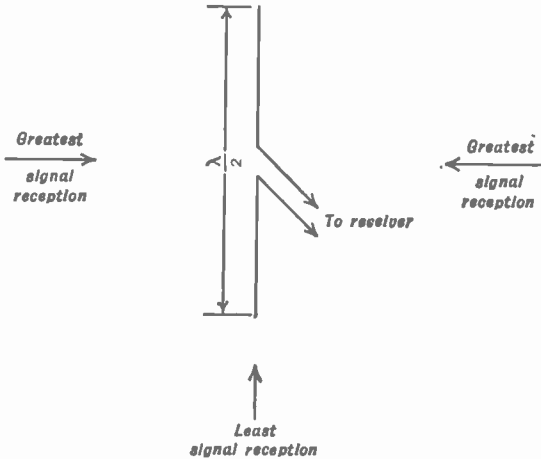


FIG. 2.10A. Dipole antennas, of the type shown, received signals best from the directions indicated.

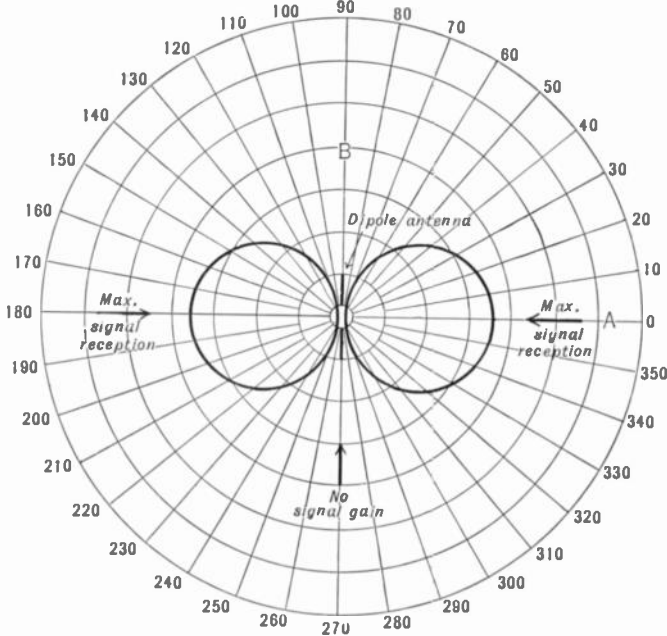


FIG. 2.10B. The directional response curve of a dipole antenna.

to this frequency would not give optimum results at the other bands, the reception would still be quite satisfactory.

To compute the length needed for the 65-mc frequency half-wave antenna, the following formula is used:

$$L \text{ in feet} = \frac{468}{f_{\text{mc}}}$$

With f set equal to 65 mc, the length would be equal to $\frac{468}{65}$ or 7.2 feet. Practically, 7 feet might be cut, with each half of the half-wave antenna 3.5 feet long. For a full-wavelength antenna, approximately 14 feet is needed. In congested areas, antenna length must be as short as possible, and only half-wave antenna systems are generally found. At the present time, most television stations are located in urban areas in order to reach the greatest number of sets. Emphasis, then, is on short antennas, such as the half-wave type. If longer lengths are desired, the equation should be modified by the proper factor. A full wavelength antenna requires a factor of 2; a wavelength and a half requires a factor of 3, etc.

Half-Wave Dipole with Reflector. The simple half-wave system provides satisfactory reception in most locations within reasonable distances of the transmitter. However, the signals reaching receivers situated in outlying areas are correspondingly weaker, and noise and interference have a greater distorting effect on the image. For these locations more elaborate arrays must be constructed — systems that have greater gain and directivity and provide better discrimination against interference.

A simple yet effective system is shown in Fig. 2-11. The two rods are mounted parallel to each other and spaced about $.15\lambda$ – $.25\lambda$ apart. The ac-

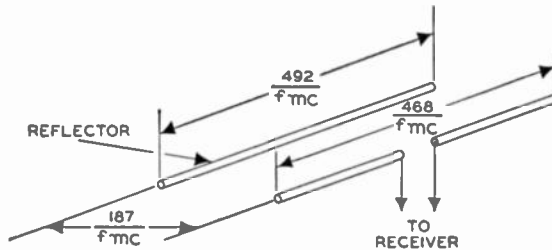


FIG. 2.11. Dipole receiving antenna and reflector.

tion of the second wire, which is not connected, is twofold. First, because of its position, it tends to concentrate signals reaching the front wire. Second, it shields the front antenna from waves coming from the rear. The gain of the array is generally 5 db greater than that obtainable from a single half-wave antenna.

Besides the additional gain that is observed with this two-wire system, the graph of Fig. 2-12 shows that the angle at which a strong signal may be received now is narrower. This is also advantageous in reducing the number of reflected rays that can affect the antenna. Finally, partial or complete discrimination is possible against interference, man-made or otherwise.

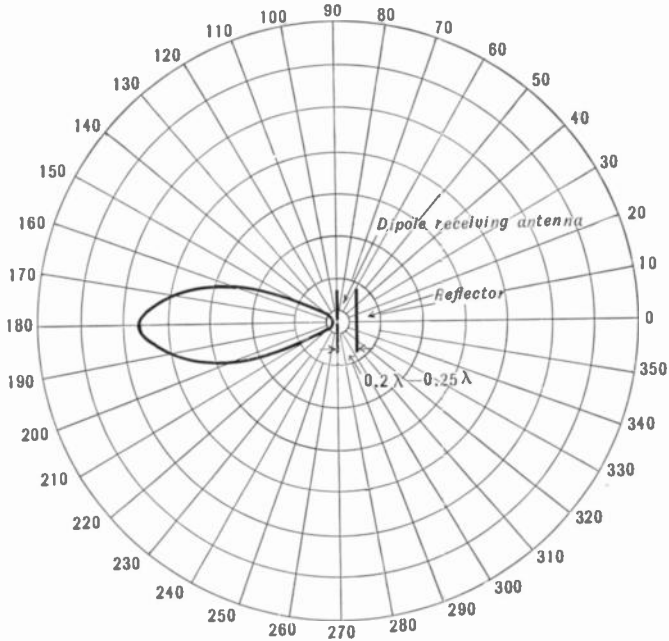


FIG. 2.12. The directional response curve for a half-wave antenna with a reflector.

The method of erecting the antenna is similar to that of the half-wave dipole, although the adjustment of the position of the wires is more critical. A small displacement, one way or another, will alter the strength of the received signal. Many commercial antenna kits do not provide adjustment of the spacing distance between the two wires. However, if an adjustment is possible, the spacing may be altered if experimentation indicates that it would result in better reception.

Another antenna widely used by servicemen and recommended by television receiver manufacturers is the folded dipole shown in Fig. 2-13A. This antenna consists essentially of two dipole antennas connected in parallel with each other. The separation between the two sections is approximately 3 to 5 inches. The folded dipole has the same bi-directional pattern as the simple dipole and approximately the same gain. The input impedance is now 300 ohms (at the frequency for which it was cut) against 72

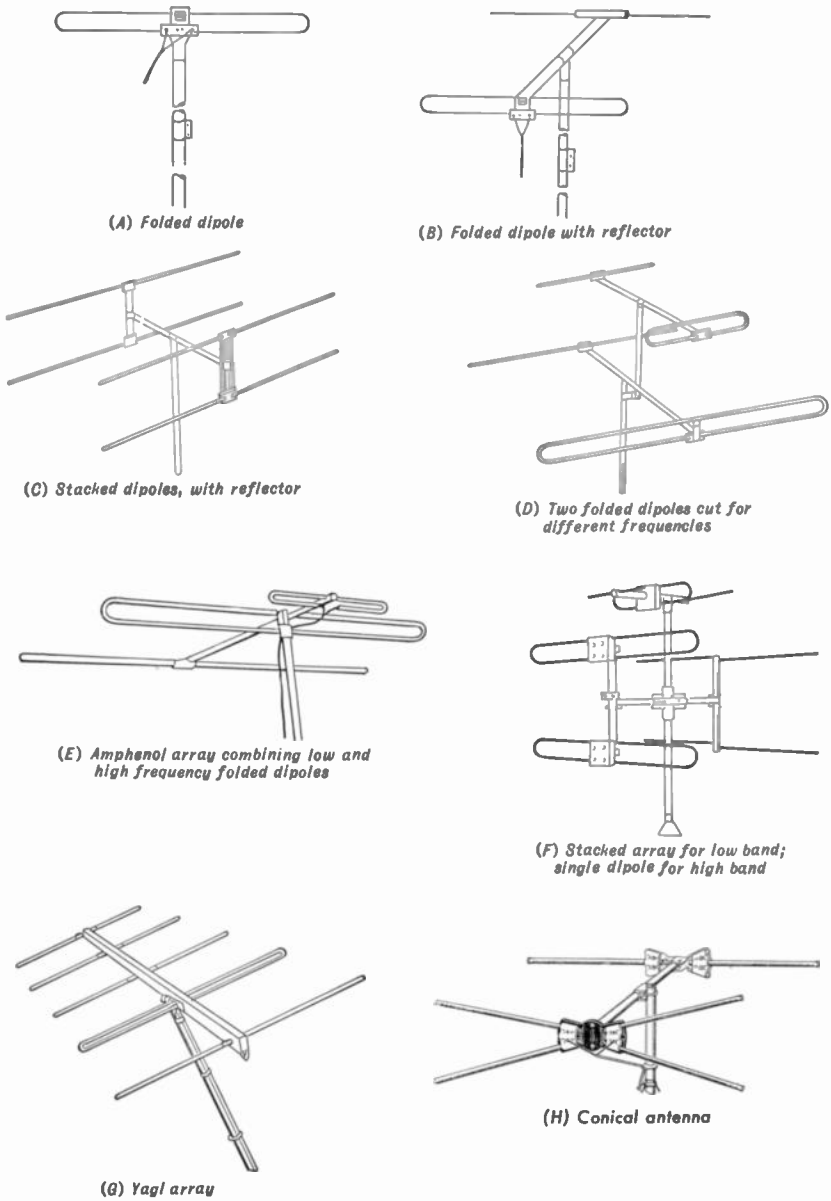


FIG. 2.13. Various types of popular television receiving antennas.

ohms for the simple dipole. However, the response of a folded dipole is more uniform over a band of frequencies than the simple dipole.

The directivity of the folded dipole can be increased by the addition of a reflector. (See Fig. 2-13B.) The unit now possesses the same directional pattern as the simple dipole with reflector. The length and spacing of the reflector can be determined by the same formulas previously given for the simple dipole and reflector.

In the remaining illustrations of Fig. 2-13, other popular types of television antennas are shown. Nearly all of these are elaborations of the basic dipole or folded dipole. In Fig. 2-13C, we have a stacked dipole array, with reflectors. (This antenna is sometimes referred to as a "Lazy H" because of its similarity to the letter H lying on its side.) Two half-wave dipoles are placed at the front of the assembly, one mounted above the other. The center terminals of each dipole are connected together by means of a parallel-wire transmission line. Each conductor of the lead-in line to the television receiver attaches to a conductor of this connecting transmission line at a point midway between the dipoles. A reflector is mounted behind each dipole.

In Fig. 2-13D, there are two folded dipoles, with reflectors, mounted one above the other. The upper dipole is cut for a resonant frequency approximately in the center of the upper VHF-TV band (174-216 mc), and the longer folded dipole is resonated at the center frequency of the lower television band. A short length of 300-ohm transmission line connects the upper dipole to the lower. From the lower antenna, a 300-ohm line feeds the signals to the receiver. With this assembly, each antenna can be oriented independently for best reception from stations within its band, providing the receiver with good coverage on both bands. Fig. 2-13E is essentially the same arrangement as Fig. 2-13D, except that the longer folded dipole acts as the reflector for the shorter folded dipole. The two folded dipoles are connected in the same manner as the array in Fig. 2-13D. Independent orientation of each folded dipole is not possible in the array of Fig. 2-13E. The unit in Fig. 2-13F is an elaboration of that shown in Fig. 2-13C.

The antenna of Fig. 2-13G contains a folded dipole with one reflector and three directors. It is known as a "Yagi" array and possesses sharp unidirectional directivity. Signals are received only from the direction broadside to the directors.

The final array (Fig. 2-13H) is one which has been extensively used because of its ability to receive low and high VHF band signals. The front elements are bent or veered forward while the rear elements (the reflectors) generally extend straight out. Due to the shape of the front elements, the response pattern contains only one major lobe on all channels (as in Fig. 2-12). This is an improvement over the conventional dipole where an ele-

ment cut for the low frequencies will have a multi-lobed pattern on the high channels and an element cut for the high frequencies will have a poor response on the low channels. With the conical antenna, one array suffices for all VHF channels.

Conicals may come either singly or stacked two or four high. The same is true of most other antennas. In installations within 10 miles of the transmitter, the simple dipole and the folded dipole, each aided perhaps with a reflector, will frequently prove completely satisfactory. Beyond 10 miles, stacked arrays are generally required. Conditions, of course, will vary in different localities and no set rule will always work. But, as a start, the technician can follow the foregoing suggestions.

Table 2-1 reviews the properties of the basic television antennas, together with their dimensions.

After the particular antenna has been chosen, the following points should be kept in mind before installing the antenna.

1. The higher the antenna, the stronger the signal received.
2. The antenna should be set-tested with an actual connection to its receiver before the supports are fixed in place permanently.
3. When more than one station is to be received, the final placement of the antenna must, of necessity, be a compromise. In extreme cases, it may be desirable or even necessary to erect several antennas.





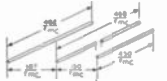


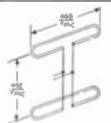
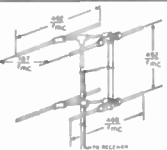
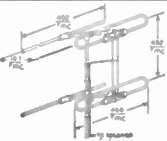
Transmission Lines. With the antenna system in position, the next problem is the transmission line that conducts the signal from the antenna to the receiver. Although many differently constructed transmission lines have been designed, only two general types find any extensive use in television installations. These are the parallel-wire types and the concentric or coaxial cable.

From the standpoint of convenience and economy, one antenna should be capable of receiving all the VHF television stations. It should have, therefore, a fairly uniform response from 54 to 216 mc. A resonant dipole presents an impedance, at its center, of 72 ohms. To obtain maximum transfer of power, the connecting transmission line should match this value. However, when we attempt to use the same dipole for a band of frequencies, we find that the 72-ohm value is no longer valid. A dipole cut for 50 mc presents a 72-ohm impedance. At 100 mc, the impedance has risen to 2,000 ohms. It is obvious that the best transmission line impedance is no longer 72 ohms, but a higher value which will serve as a compromise. It is desirable to use as high an impedance value as possible, because line loss is inversely proportional to characteristic impedance. On the other hand, such factors as the size of the line and the wire gage must also be considered, and it is current practice to design the input circuit of the television receiver for a 300-ohm transmission line. It has been found that a 300-ohm line used

with a half-wave dipole produces a broad frequency response without too great a loss due to mismatching. A folded dipole has an impedance close to 300 ohms at its resonant frequency and a much more uniform response is obtained with this antenna.

The flat, parallel-wire transmission line (Fig. 2-14A) is probably the most

TABLE 2.1. Characteristics of common television antennas.

ANTENNA TYPE	APPROXIMATE CHARACTERISTIC IMPEDANCE [ohms]	DIRECTIVITY	DIMENSIONS [feet]	GAIN [COMPARED TO DIPOLE]	
				db	voltage
Plain dipole	72	Bi-directional; broadside to elements		-	Unity
Folded dipole	300	Bi-directional; broadside to elements		-	Unity
Dipole and reflector	50	Uni-directional; broadside to radiator		5	1.78
Folded dipole and reflector	250	Uni-directional; broadside to radiator		5	1.78
Dipole, reflector, and director	25	Uni-directional; broadside to radiator		7	2.24
Folded dipole, reflector and director	100	Uni-directional; broadside to radiator		7	2.24
Stacked dipoles	40	Bi-directional; broadside to elements		5	1.78
Folded stacked dipoles	150	Bi-directional; broadside to elements		5	1.78
Stacked dipoles and reflectors	25	Uni-directional; broadside to elements		10	3.16
Folded stacked dipoles and reflectors	100	Uni-directional; broadside to elements		10	3.16

popular lead-in line in use today. The wires are encased in a plastic ribbon of polyethylene which is strong, flexible, and is not affected by sunlight, water, cold, acids, or alkalis. At 100 mc, the line loss is on the order of 1.2 db per hundred feet of line. Its characteristic impedance ranges from 75 ohms to 300 ohms. The line is balanced, which means that both wires possess the same average potential with respect to ground. It is, however, ungrounded and therefore not recommended for use in extremely noisy locations.

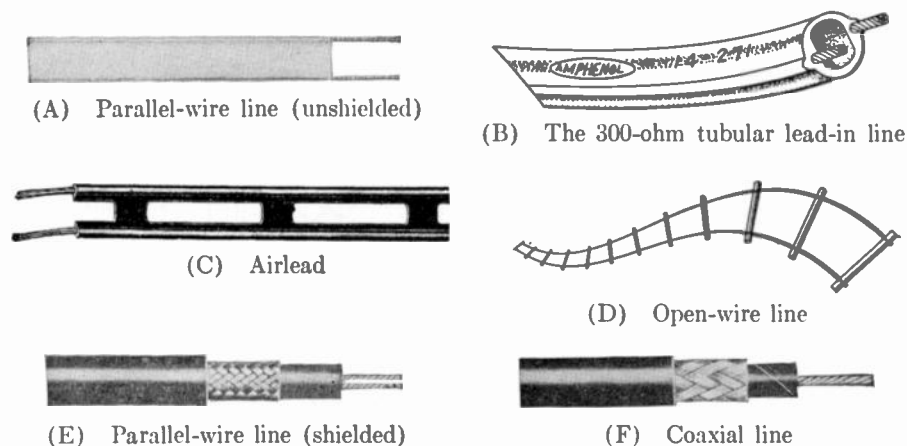


FIG. 2.14. Various types of popular transmission lines used for F-M and television installations.

A companion tubular twin-lead line is also available (Fig. 2-14B) which, while somewhat more expensive than the flat twin-lead, does possess the advantage of being less affected by adverse weather conditions than the flat line. Rain, sleet, snow, etc., may not physically affect the flat line, but electrically these serve to increase its attenuation. Thus, at 100 mc, flat and tubular lines under dry conditions possess equal attenuations of 1.2 db. When wet, however, the loss on a flat line rises to 7.3 db, whereas that on a tubular line is only 2.5 db. In strong signal areas this may not be important, but imagine what it would do to a picture in a weak signal area.

A third type of parallel-wire line is shown in Fig. 2-14C. Known commercially under the name of "Airlead," it has 80 per cent of the polyethylene webbing removed. This is said to reduce the loss or db attenuation by at least 50 per cent. Its performance in wet weather is also superior to the other lines. Line impedance is still 300 ohms.

The fourth parallel-wire line is one which is completely open (Fig. 2-14D), being held together by small polystyrene spacers placed approximately 6 inches apart. The attenuation of this line is only of the order of

0.35 db per hundred feet at 100 mc and it is relatively unaffected by changes in weather. Impedance of this line is 450 ohms.

A parallel-wire transmission line that is completely shielded is shown in Fig. 2-14E. The two wires are enclosed in a dielectric, possibly polyethylene, and then the entire unit is shielded by a copper-braid covering. As a protection against the elements, an outer rubber covering is used. Grounding the copper braid converts it into a shield which prevents any stray interference from reaching either conductor. Furthermore, the line is balanced against ground. It can be built with a variety of impedance values ranging from 50 ohms up, but a 225-ohm line has found greatest use in television installations. Attenuation of this line is 3.4 db at 100 mc and it is seen that this is considerably higher than the attenuation of any of the unshielded lines. Because of this, and because of its greater cost, the shielded line is used only where the surrounding noise is particularly severe.

The final transmission line is the coaxial or concentric cable, shown in Fig. 2-14F. It contains an insulated center wire enclosed by a concentric metallic covering which is generally flexible copper braid. The inner wire is kept in position by a solid dielectric which is chosen for its low-loss properties. The signal carried by the line is confined to the inner conductor, with the outer copper-braid conductor grounded so as to serve as a shield against stray magnetic fields. Due to this arrangement, the line is unbalanced and the input coil of the receiver must be connected accordingly. Coaxial cables are available in a range of impedances from 10 to 150 ohms.

At the receiver, the connections for balanced and unbalanced line differ, as shown in Fig. 2-15. For a balanced line, the input coil is center-tapped and grounded at this tap. Stray fields, cutting across both wires of a balanced line, induce equal voltages in each line. The similar currents that flow because of the induced voltages are in the same direction on the two conductors of the line and they neutralize each other.

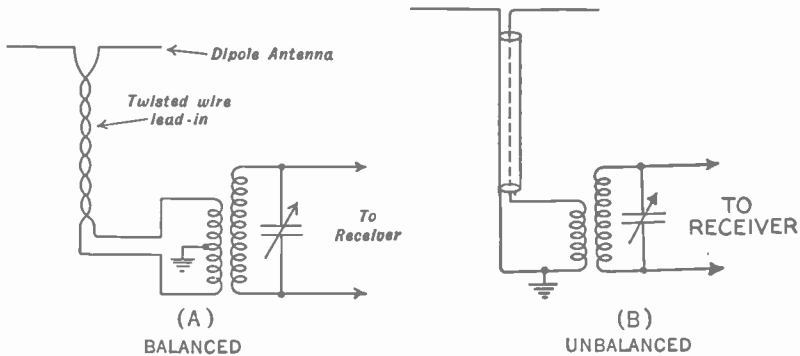


FIG. 2.15. Methods of connecting lead-in wires to the input coil of a receiver.

Antenna Installation. A complete installation is illustrated in Fig. 2-16A. Stand-off insulators should be mounted on the side of the building to prevent the transmission line from rubbing against the wall. It is good practice, also, to install a lightning arrester at the point where the line enters the building (see Fig. 2-16B). The other ends of the arrester should be connected securely to an iron pipe sunk into the earth. From the lightning arresters, the transmission line is led into the building to the receiver.

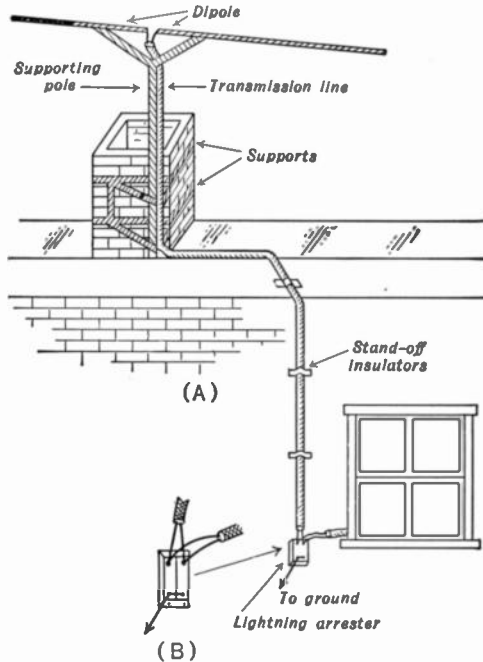


FIG. 2.16. An antenna installation for a television receiver. The ground from the lightning arrester should be as short and direct as possible.

QUESTIONS

1. Why are antennas more important to television receivers than to standard A-M broadcast receivers?
2. What happens if the same signal is received from several directions? What is this called?
3. What is the importance of the ionosphere?
4. What factors determine whether or not radio waves are returned to earth from the ionosphere?
5. How are television signals sent? Explain.
6. What is the horizon distance for a television antenna mounted atop a tower 450 feet high?
7. How much is the above distance increased if the antenna is raised an additional 100 feet?

8. A receiver is located 30 miles *beyond* the horizon distance computed in Question 6. How high should the receiver antenna be raised to receive signals from this transmitter?

9. What is meant by wave polarization? How does it affect the installation of a television receiving antenna?

10. What are the disadvantages of using any length of wire for the reception of television signal?

11. Indicate the materials required to construct and erect a half-wave dipole antenna.

12. A half-wave antenna is to resonate at 70 mc. What should its overall length be?

13. How are the directional characteristics of antennas obtained?

14. Name and sketch five different types of antennas that could be employed to receive television signals.

15. What precautions must be observed in choosing and installing a transmission line?

16. Name and describe four types of transmission lines.

17. Where could each type of line be used? Give reasons for each choice.

18. Define antenna gain and antenna directivity.

19. Must a signal always be received directly from the transmitter to be useful? Explain.

20. A half-wave dipole antenna designed for 80 mc is to be used on 192 mc. By how much should it be altered?

21. Illustrate a balanced and an unbalanced input system.

Chapter 3

WIDE-BAND TUNING CIRCUITS—R.F. AMPLIFIERS

The Bandwidth Problem. The television signal occupies a 6-mc bandwidth in the radio spectrum, a range far greater than anything we have had to receive with the ordinary sound set. The problem must be met at the television receiver in the R.F. and mixer stages, if both are used; otherwise, only at the mixer. The response of the tuned receiving circuit should be uniform throughout the 6-mc band and yet be selective enough to discriminate against unwanted image frequencies or stations on adjacent bands. Before the circuits of the R.F. and mixer stages are considered, it will be helpful to discuss wide-band tuning circuits.

Ordinary Tuning Circuits. A single coil and capacitor, connected as shown in Fig. 3-1A, form a parallel tuning circuit. At or near the resonant frequency, the variation of impedance which this combination presents is given by the graph of Fig. 3-1B. At frequencies below the resonant frequency, the parallel combination acts as an inductance with a lagging current; above resonance, the effect is capacitive with a leading current. At the resonant point, both capacitive and inductance reactances cancel each other, the impedance becoming high and wholly resistive.

While Fig. 3-1B shows the general shape of the resonant curve, more specific information is necessary. Hence, in Fig. 3-2, several resonant curves have been drawn, each for a circuit having a different value of Q . Q , which is the ratio of inductive reactance to coil resistance, may be taken to indicate two things:

1. The sharpness of the resonant curve in the region about the resonant frequency. This, of course, is the selectivity of the tuning circuit.
2. The amount of voltage that will be developed by the incoming signal across the resonant circuit at resonance.

For any given circuit, the greater its Q value, the more selective will be the response of the circuit and the greater the voltage developed. While these factors may be highly desirable, they are only useful if they do not

interfere with reception of radio signals. At the broadcast frequencies, each station occupies a bandwidth of 10 kc. Within this region, uniform response is desirable. However, the sharply peaked curve of Fig. 3-1B does not produce equal response at all points within this region. The portion of the signal exactly at the resonant frequency, for example, would

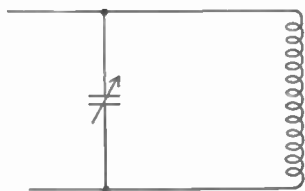


FIG. 3.1A. A parallel tuning circuit. The response curve for this circuit is shown in Fig. 3.1B.

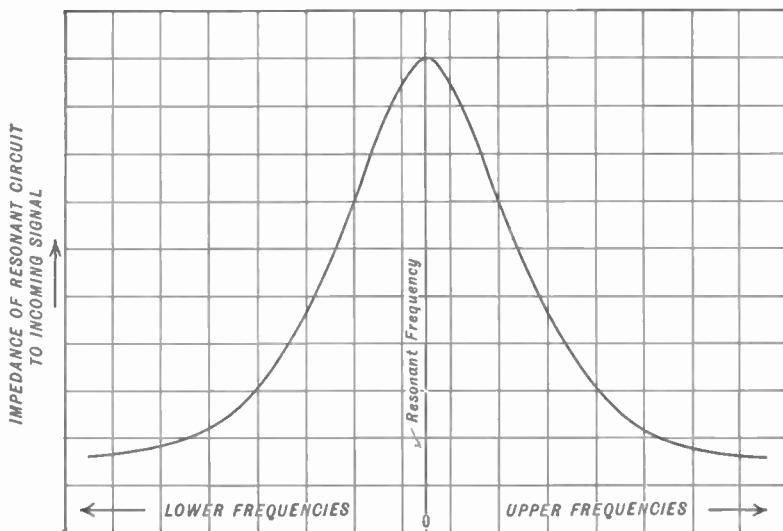


FIG. 3.1B. The response curve for the tuning circuit shown in Fig. 3.1A.

develop a greater voltage across the resonant circuit than those frequencies at the outer fringe, plus and minus 5 kc away. A coil and capacitor combination having a lower Q would give a more uniform response and might be chosen over one with a higher value of Q . Less voltage results from this change but, with the advent of high gain tubes, amplification is not too serious a problem. The emphasis now can be shifted to fidelity, which is especially necessary for the reproduction of images in television receivers.

Transformer Coupling. Whereas the simple circuit already described is sometimes used by itself for tuning, a more common combination is

shown in Fig. 3-3. Here we have an untuned primary coil inductively coupled to a tuned secondary. With this form of coupling, additional gain may result by having more turns in the secondary than in the primary coil. The stepped-up voltage applied to the grid of the next stage is larger than that obtained with only the single coil and capacitor by a figure dependent upon the design of the coils.

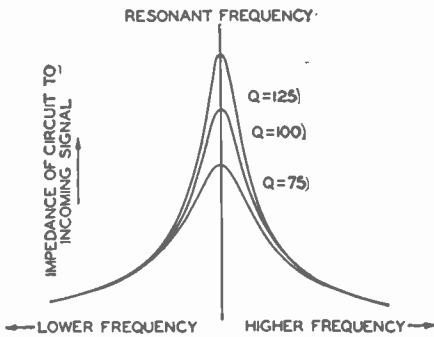


FIG. 3.2. The variation in the response curve with different values of Q .

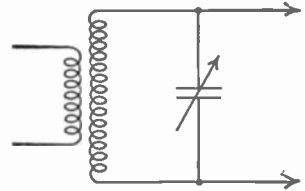


FIG. 3.3. A common form of transformer-coupled tuning circuit used in radio receivers.

The shape of the response curve of the primary circuit depends to a great extent upon the degree of coupling between the coils. When the coefficient of coupling k is low (i.e., when the coils are relatively far apart), the interaction between coils is small. The secondary response curve will retain the shape shown in Fig. 3-1B.

As the coefficient k is increased, the secondary circuit reflects a larger impedance into the primary. The primary current is affected more by variations in the tuning of the secondary capacitor. This, in turn, changes the manner in which the flux lines cut across the secondary coil and the end result is a gradual broadening of both primary and secondary response curves. With very close coupling, the secondary response curve may continue to broaden and even develop a slight dip at the center. The dip, however, will never become too pronounced. It must be remembered that the discussion, so far, has dealt with coupled circuits where the primary is untuned. Hence, no matter how close a coupling is effected, the secondary will retain essentially the same curve shape given in Fig. 3-1.

On the other hand, with two *tuned* circuits coupled together, such as I.F. transformers, the effect of each circuit on the other becomes more pronounced. With close coupling, the familiar double-humped curve of Fig. 3-4 is obtained. The closer the coupling, the broader the curve and the greater the dip at the center.

For television reception, none of these preceding combinations provide

the necessary uniform bandwidth. Loose coupling produces a curve that is too sharp and which lacks uniformity over its range. Tight coupling tends to decrease the voltage of the frequencies near resonance because of the dip. Between these two extremes we may obtain some semblance of uniform response about the center point of the curve, but never for a 6-mc spread. However, if a low-valued resistor is shunted across the coil and capacitor,

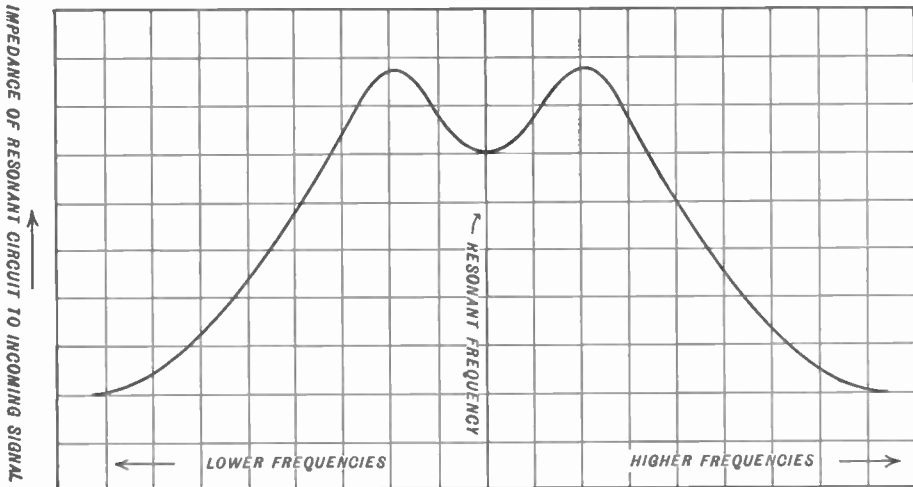


FIG. 3.4. Close coupling between two tuning circuits produces this type of response curve.

we can artificially flatten the curve to receive the necessary 6 mc. The extent of the flat portion of the response curve will depend inversely on the value of the shunting resistor. The higher the resistor, the smaller the width of the uniform section of the curve. Hence, what we could not accomplish with a coil and capacitor alone, we can do with a combination of these two with resistance.

One of the undesirable results of increasing the width of a response curve by the resistor method is the lowered Q that is obtained. As the value of Q decreases, the voltage developed across the tuned circuit becomes smaller for the same input. An inevitable reduction in output results. There are many ways of combining the tuned circuits and loading resistors to achieve the optimum gain and selectivity. Several of the more widely used circuits will be discussed in the section on R.F. amplifiers.

Special Tubes for Television. In commercial television circuits, resistors having values between 1,500 and 10,000 ohms are shunted across the tuning circuit to provide the necessary bandwidth. To compensate for the signal reduction due to the shunting resistors, pentode tubes having

large values of mutual conductance were especially designed for television. At present, the best gain obtainable is about 20 per stage. Compared to the amplification available with similar tubes at the broadcast frequencies (several hundred or more), the need for a greater number of stages becomes evident.

The reason for the low gain at television frequencies is directly related to the low-valued shunting resistor that is placed across the tuning circuit.* A tuning circuit, when connected in the output of a tube, is essentially in series with the plate resistance of the tube. This is illustrated in Fig. 3-5,

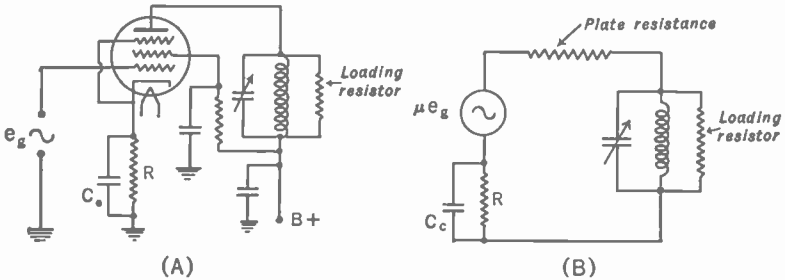


FIG. 3.5. An amplifier stage (A) and its equivalent circuit (B).

in which the actual schematic and its electrical equivalent are shown. At resonance, the resistance of the tuning circuit itself may be high, but due to the low shunting resistor, the total value of the combination becomes low. The plate resistance, on the other hand, is very high (in pentodes), and most of the output voltage is lost in the tube. Only a small portion of the total voltage appears across the tuning circuit to be transferred to the next stage.

Mathematically, the gain of the pentode stage can be expressed closely by the relation:

$$\text{Gain} = g_m \times Z_L$$

where g_m = mutual conductance of the tube (in mhos),

Z_L = load in output circuit (in ohms).

For a 6AG5 tube, g_m is 5,000 micromhos. With a plate load of 2,000 ohms, we obtain

$$\begin{aligned} \text{Gain} &= \frac{5,000}{1,000,000} \times 2,000 \\ &= 10. \end{aligned}$$

The 5,000 is divided by 1,000,000 to convert it from micromhos to mhos.

* With frequency increase, the input impedance (between grid and cathode) of a tube decreases. This, too, will load down the attached tuning circuit as surely as a resistor. More on this later.

To obtain more amplification per stage, the mutual conductance of the tube must be increased. g_m , it will be recalled, represents the change in plate current caused by a change in grid voltage. To effect an increase in this ratio, radio engineers designed tubes in which the grid is given greater control over the space charge near the cathode. This was done by moving the grid closer to the cathode. Although this caused an increase in grid-to-cathode capacitance, it increased the mutual conductance even more. This design is exemplified in tubes like the 6AG5, 6J6, 6AK5, 6CB6, and 6BC5.

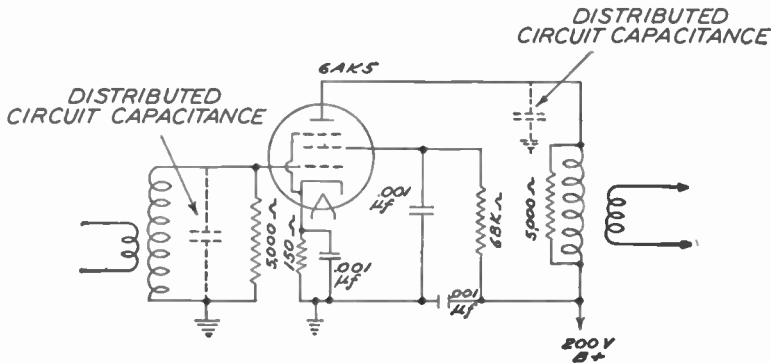


FIG. 3.6. A typical television R.F. amplifier.

As an example, the 6AG5 has a mutual conductance of 5,000; the 6AK5 likewise has a g_m of 5,000, and the 6CB6 has a g_m of 6,200. Compare these values with older R.F. and I.F. pentode voltage amplifiers, like the 6SK7, 6D6, 6S7, and the 6SJ7, which have mutual conductances of 2,000, 1,200, 1,750, and 1,600 micromhos, respectively. If the gain of the television stage is computed using these values of mutual conductance, a voltage amplification much less than 10 is obtained.

R.F. Amplifiers. The typical television R.F. stage, shown in Fig. 3-6, is very similar to the same stage in amplitude-modulated broadcast receivers. Its functions are threefold. First, it provides signal amplification in a portion of the set where the signal is at its lowest value. In outlying regions or noisy locations, this extra amplification may be the deciding factor in whether or not satisfactory reception is obtained. Second, it provides greater discrimination against signals lying in adjacent bands. This is especially applicable for image frequencies. A properly designed R.F. stage will help the signal override any small interferences that are produced in the tubes themselves. The latter boost applies only to the first tube or two where the signal may be comparable to the internal disturbance voltage. In audio systems, the internal tube disturbance is known as

noise. In television receivers, these disturbances are amplified along with the video signal and, if stronger than the received signal, will appear as small white spots on the image screen. (These spots are often referred to as "snow," or masking voltages.) Finally, the R.F. amplifier also reduces local oscillator radiation, which can be quite offensive to neighboring receivers.

The tube employed in the R.F. stage, besides having a low noise content and a high mutual conductance value, should also possess a remote cut-off characteristic. With remote cut-off properties, the stage does not distort

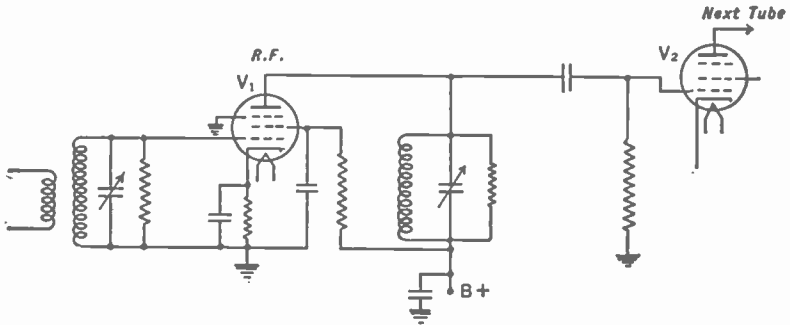


FIG. 3.7. Another R.F. amplifier stage. One tuned circuit is common to the plate of V_1 and the grid of V_2 .

as readily when large input signals are received. Furthermore, automatic gain control * voltage may be applied to the tube, materially aiding the amplifier stability and tending to maintain a steady signal output.

Some of the forms that the R.F. stage may assume are shown in the accompanying diagrams. In Fig. 3-6, transformer coupling is used in the input and the output circuits of the R.F. amplifier. Each transformer is loaded down by a shunting resistor so that its response will be fairly uniform over a 6-mc bandwidth. The resistor value is chosen with an eye toward keeping the stage gain as high as possible. On the upper VHF channels (7-13) and throughout the UHF band sufficient loading is usually provided by the tube itself so that external resistors are not needed.

In Fig. 3-7 we have an arrangement in which a single tuned circuit, instead of a transformer, is used between the plate of the R.F. amplifier and the mixer tube. The tuning capacitance shown in each of these diagrams might be either a small variable trimmer capacitor or the stray circuit wiring and tube capacitance always present in the circuit. In the latter instance, adjustment of the tuned circuit would not be accomplished by varying the capacitance (since the wiring and tube capacitances are not

* A.G.C. in a television receiver is similar to A.V.C. in a conventional radio receiver.

adjustable) but by using movable cores within the coil. Thereafter a selector switch or some other tuning arrangement is used. Although only one set of coils is shown in some of these diagrams, there would be similar arrangements for each of the channels.

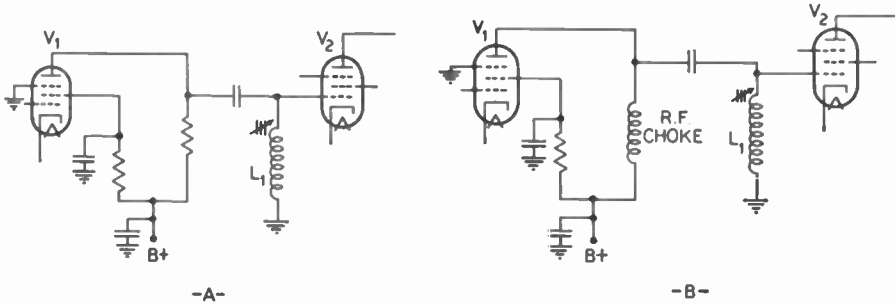


FIG. 3.8. Two additional interstage coupling networks between the R.F. amplifier (V_1) and the mixer (V_2). Either triodes or pentodes may be employed.

There are a number of variations of the coupling network between V_1 and V_2 of Fig. 3-7, two of which are shown in Fig. 3-8. In the first illustration, Fig. 3-8A, the plate load for V_1 is a resistor. L_1 is the resonant circuit between the two tubes and it is placed in the grid circuit of V_2 . In Fig. 3-8B, the plate load for V_1 is an R.F. choke. It might also be another resonant circuit. There are as many different approaches as there are designers, and only those that are more popular can be shown here.

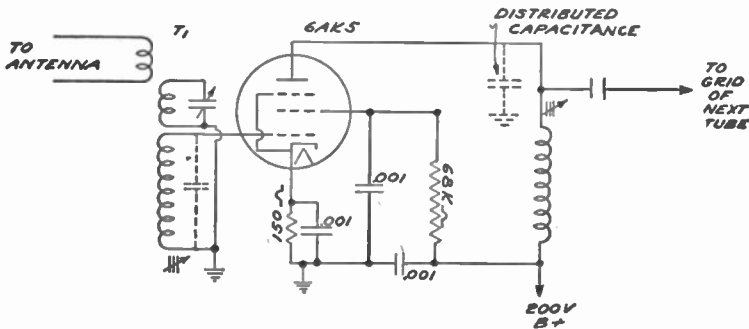


FIG. 3.9. An R.F. amplifier which combines the response characteristics of grid and plate tuned circuits to obtain a 6-mc overall spread.

An approach sometimes practiced is to insert an over-coupled transformer in the input circuit and a single-peaked circuit in the plate circuit of the stage. One such circuit is shown in Fig. 3-9. The primary winding of T_1 is untuned and matches the transmission line impedance. The grid winding is tuned by the grid input capacitance of the tube, plus whatever

stray capacitance is inevitably present in the circuit. The third winding contains a small trimmer to permit adjustment, although in some instances it is nothing more than a one- or two-turn winding which functions as a link coupling between the input and grid coils. The combination of these three coils results in a double-peaked response curve (see Fig. 3-10A). In the plate circuit of the stage, and serving as impedance coupling between circuits, is a single tuned coil. Its response is single-peaked, as illustrated

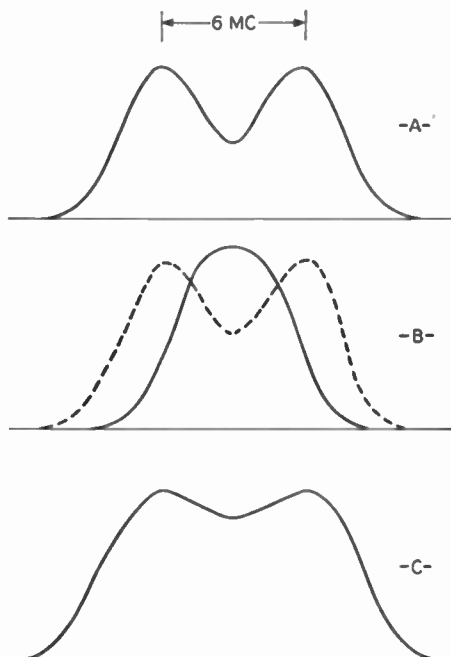


FIG. 3.10. The combination of two tuning circuits to produce a flat-topped overall response. (A) grid-circuit response; (B) plate-circuit response; (C) overall response.

in Fig. 3-10B. By properly adjusting the peaks of these circuits, we can achieve an essentially overall flat response of 6 mc for the stage (see Fig. 3-10C). (The word “essentially” is used because it seldom occurs that the R.F. response curve has an absolutely flat top. In practice, up to a 30% dip in the center of the curve is permissible.)

Another method of coupling between stages in order to achieve a broad bandpass is shown in Fig. 3-11. Here, a small capacitor connects the primary and secondary windings. The value of this capacitance is low (10 to 20 $\mu\mu\text{f}$) and governs the extent of the bandwidth; increasing the capacitance increases the bandwidth.

In Fig. 3-12, mutual capacitive coupling between R.F. amplifier and the mixer is achieved in still another way. L_A and L_B are two coils which

are coupled to each other *only* through the common capacitance C_4 . In other words, their magnetic fields do not interact. Each coil is pre-tuned to the same frequency by means of a brass slug. C_A represents the output capacitance of the 6AG5 R.F. amplifier plus other circuit capacitances; C_B is the input capacitance of the following 6J6 plus the distributed wiring capacitance.

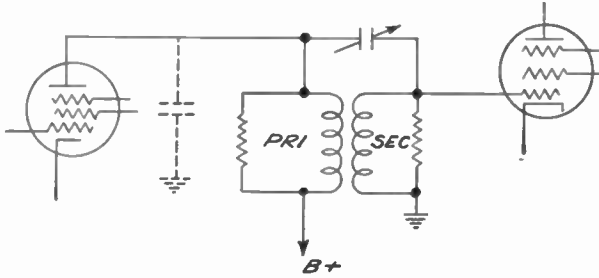


FIG. 3.11. One method of increasing the coupling between two tuned circuits to achieve broad bandpass.

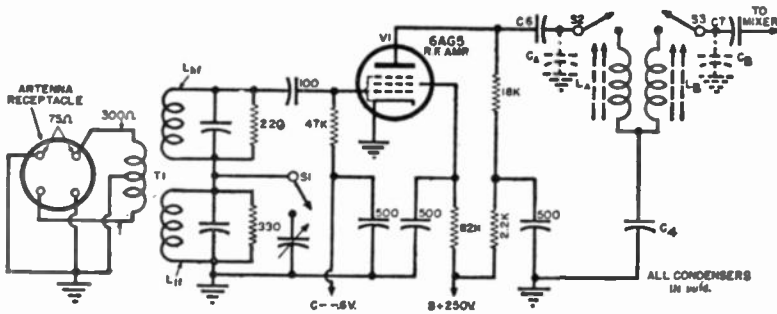


FIG. 3.12. Mutual capacitive coupling, C_4 , is used between L_A and L_B . (See text.)

In this type of tuned circuit, the bandwidth is determined by the degree of coupling and the Q 's of L_A and L_B . The degree of coupling is controlled by the value of C_4 . The smaller this capacitance, the greater the mutual impedance and the greater the bandwidth. The value of C_4 is chosen to provide a bandpass of approximately 6.0 mc. To maintain a constant bandwidth, C_4 has a value of 250 μf on the lower channels and a value of 140 μf on the higher channels. This compensates for the change in coil Q 's with frequency. For each channel a new pair of coils is switched into the circuit.

Another feature of the circuit in Fig. 3-12 is the provision for either 75- or 300-ohm input transmission lines. This is accomplished simply by using the full primary winding of the input transformer for the 300-ohm

line and half of the winding for the 75-ohm coaxial line. Inductance of a coil is proportional to the square of the number of turns. Doubling the number of turns produces four times the inductance and, at the same frequency, four times the impedance; 300 ohms is four times 75 ohms.

Push-pull rather than single amplifiers have also been used in the R.F. section of television receivers. The circuit schematic is shown in Fig. 3-13. The transmission line from the antenna is fed directly into the grids of a 6J6 push-pull triode amplifier. To match the impedance of the line, two 150-ohm resistors are connected in series to provide the total of 300 ohms. T_1 is a center-tapped coil used to prevent low-frequency signals from reaching the grids of the R.F. amplifier. C_1 and C_2 are antenna-isolating capacitors.

In the plate circuit of the R.F. amplifier, starting with L_{26} and progressing down to L_1 , we have a series of inductances that may be considered as sections of a quarter-wave transmission line. The switch, as it moves progressively to the left, brings in more inductances, thus decreasing the channel frequency. In position 13, only L_{25} and L_{26} are in the circuit and the receiver is set for the highest VHF-TV channel. At position 2, the set will receive the lowest channel. (Position 1 was for the now obsolete Channel 1.) At various points along the line, adjustments may be made by changing the position of the tuning slugs. The physical construction of each of the small inductances, L_{13} to L_{26} , is a small, fixed silver strap between the switch contacts. Each strap is cut long enough to introduce a 6-mc change in frequency. In order to make the transition from the lowest high-frequency channel, 174-180 mc, to the highest low-frequency channel, 82-88 mc, adjustable coils L_{11} and L_{12} are used. Coils L_1 to L_{10} are more substantial in appearance than coils L_{13} to L_{26} , being wound in figure-8 fashion on fingers protruding from the switch assembly.

Since each section of the 6J6 is a triode, neutralizing capacitors are necessary to counteract the grid-to-plate capacitance. This is the function of C_3 and C_4 .

Coupling between the quarter-wave line of the R.F. amplifier and a similar section in the grid circuit of the mixer tube is twofold: by direct capacitance connection and by link coupling. The response characteristic of these R.F. circuits extends the full 6 mc. In addition, a 10,000-ohm loading resistor is placed across a portion of the mixer tuning circuits to provide the necessary bandwidth. (In Fig. 3-13 the resistor is effective only for channels below 9.)

It is interesting to note that triodes are employed in television R.F. amplifiers as well as pentodes. To the radio serviceman who is accustomed to pentodes only in these stages, the change will come as a distinct surprise. The reasons for this switch are important and will be covered in Chapter 4.

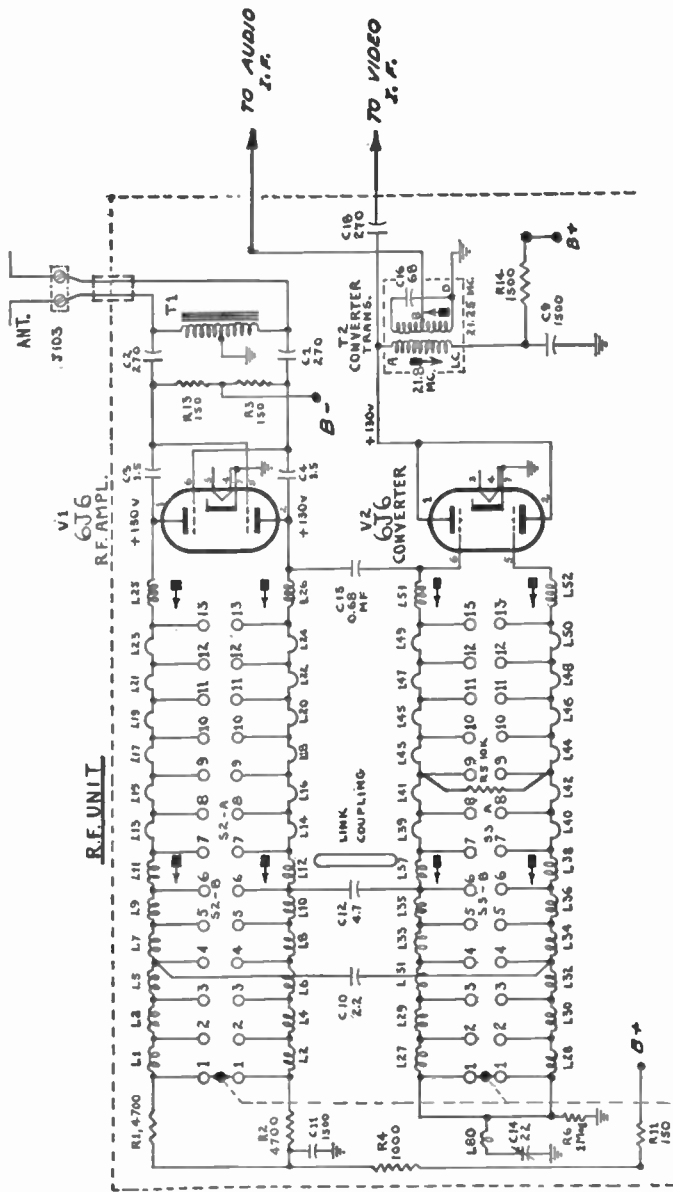


Fig. 3.13. A quarter-wave transmission line used for tuning the R.F. stages of a television receiver.

Grounded-Grid Amplifiers. Triode R.F. amplifiers are often employed in an arrangement known as the grounded-grid amplifier. This type of amplifier is contrasted with the conventional amplifier in Fig. 3-14. Note that the grid of the tube is at R.F. ground potential and that the signal is fed to the cathode. The tube still functions as an amplifier because the flow of the plate current is controlled by the grid-to-cathode potential. Instead of varying the grid potential and maintaining the cathode fixed, the grid is fixed and the cathode potential is varied. The net result is still the

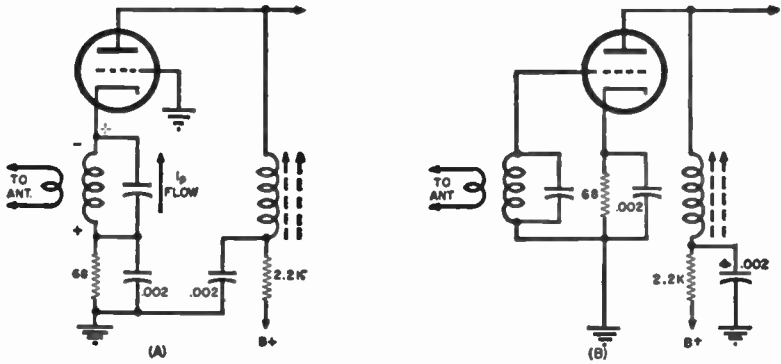


FIG. 3.14. A comparison between the grounded-grid (A) and conventional R.F. amplifier (B).

same. In addition, the grid, being grounded, acts as a shield between the input and output circuits, thereby preventing the feedback of energy which is so essential to the development of oscillations.

The grounded-grid amplifier also offers low input impedance, enabling the amplifier to match the antenna transmission line impedance. The low impedance provides a broader bandpass characteristic which is particularly desirable for 6-mc television signals.

In General Electric television receivers, the R.F. amplifier employs a 6AU6 high-frequency pentode connected as a grounded-grid triode amplifier. (See Fig. 3-15.) The antenna is connected into the cathode circuit of the R.F. amplifier. L_1 is a simple high-pass filter designed to reject all low-frequency signals, especially those at the intermediate frequency. The cathode chokes, L_2 to L_6 , are placed in series with the cathode resistor to prevent the input impedance from being lowered by the shunting effect of any stray capacitance to ground due to the cathode of the tube. The choke value is changed with frequency. R_1 and C_1 provide cathode bias.

The R.F. amplifier is coupled to the mixer tube through a wide-band transformer. One such unit is provided for each channel. The windings are self-tuned by the distributed and tube capacitances to provide maxi-

imum gain through a high L/C ratio. The R.F. coils for each channel are placed physically near the oscillator coils of the same channel (not shown) in order that both voltages will combine at the mixer grid.

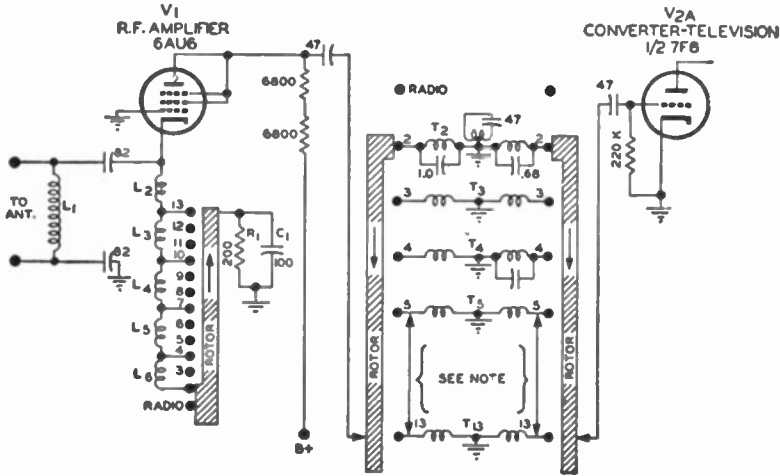


FIG. 3.15. The grounded-grid R.F. amplifier stage used in G.E. television receivers. (NOTE: R.F. coils and switch points for channels 6 through 12 not shown.) Coils T_6 through T_{12} correspond to channels 6 through 12 and are connected the same as T_5 .

In the conventional amplifier, the incoming signal is applied to the control grid of the tube. In the grounded-grid amplifier, it is the cathode that receives the signal. A combination circuit, in which both grid and cathode receive portions of the signal, is shown in Fig. 3-16. Half of the signal is developed between point A and ground across L_1 and half between point B and ground. Current flowing in the cathode circuit flows from the center-tap of L_1 to point B and thence through R_1 to the cathode. The d-c potential developed across R_1 represents the bias for the tube. In addition, the bottom end of R_2 can connect to an A.G.C. voltage for a regulatory bias, if desired. Otherwise this end of R_2 can be grounded.

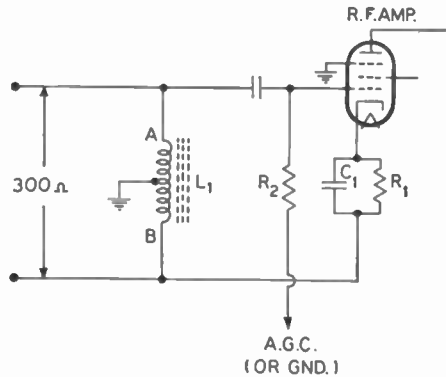


FIG. 3.16. A balanced input circuit in which grid and cathode receive portions of the incoming signal.

L_1 serves the same purpose here as T_1 did in Fig. 3-13. The input is balanced and will terminate a 300-ohm twin-lead line.

Internal Tube Capacitances. As equally important as the mutual conductance of a tube are its interelectrode capacitances. It has already been noted that the gain of a stage is equal to the product of the mutual conductance of the tube and the load impedance. The load impedance, in turn, is essentially equal to the value of the resistor shunting the tuning coil and capacitor. And, as we shall see in a moment, it is the value of the L to C ratio of the tuning circuit which determines how high a resistor we can use.

For greatest gain over any band, a high L to C ratio should be maintained in each resonant circuit. The capacitance which shunts the coil includes the interelectrode capacitance of the tube. As we make this capacitance smaller, the gain increases correspondingly. In addition, the value of the resistance R needed to load a tuned circuit is proportional to the reactance of the capacitance across the coil. Thus, with a smaller capacitance, we obtain a higher capacitive reactance and the loading resistor is higher in value. The end result is more gain.

For the R.F. input stage, the minimum capacitance is determined by:

1. The grid-to-cathode capacitance, C_{gk} .
2. The grid-to-plate capacitance, C_{gp} .
3. The stray capacitance, C_s .

The total capacitance is equal to

$$C_{\text{total}} = C_s + C_{gk} + C_{gp}(1 + G)$$

where G is the gain of the stage, usually about 10-20 in these amplifiers.

For the 6AK5, C_{gk} is equal to $4 \mu\mu\text{f}$, C_{gp} amounts to $0.015 \mu\mu\text{f}$, and the gain of the stage may be taken as 10. The stray capacitance will depend upon the manner in which the stage is wired and may amount to an additional $10 \mu\mu\text{f}$. The total, or $14.2 \mu\mu\text{f}$, would then represent the minimum capacitance of the stage and would have to be considered as an addition to any tuning capacitor inserted across the coil (see Fig. 3-17).

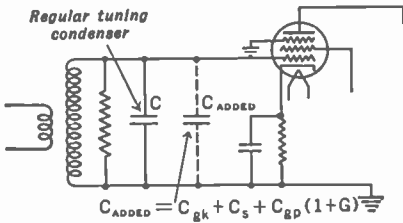


FIG. 3.17. At the higher frequencies, the stray wiring and internal tube capacitances represent an appreciable part of the total circuit capacitance and hence must be included in all computations.

At the broadcast frequencies (500 to 1,500 kc) in the ordinary home receiver, these tube and wiring capacitances are never serious when compared to the size of the tuning gang employed.

However, when frequencies as high as 90 to 100 mc are to be received, the tuning capacitor may be even smaller than these additional capacitances and they can no longer be disregarded.

Whereas the wiring and tuning capacitances remain fixed once the set has been completed, no such happy state of affairs exists for C_{pk} or C_{gp} ($1 + G$). The latter values will vary as the gain of the stage varies. This occurs every time the input voltage changes. C_{pk} will change its value as

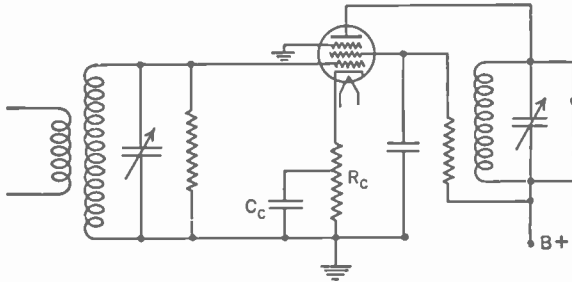


Fig. 3.18. By leaving R_c , either partially or totally unby-passed, it is possible to minimize the effect of tube capacitance changes with receiver operation.

the electron current is altered. The effect of the variation, if great enough, is sufficient to detune the stage. Again, these small items, insignificant in themselves, may become very influential as the frequency increases and the size of the coil and capacitor decreases.

It has been discovered that a small amount of negative feedback will minimize these variations. For this purpose, a portion of the grid-bias resistor is unby-passed. In doing this, however, the cathode is no longer directly connected to the capacitor C_c and is not at a-c ground potential. Under these circumstances, the screen capacitor and the suppressor grid should be tied directly to ground instead of to the cathode itself. A suitable circuit is given in Fig. 3-18.

We have considered only the minimum capacitance in the input circuit. A similar line of reasoning may be applied to the plate tuning circuit, where the total minimum capacitance is composed of the following:

1. The output capacitance, C_o , as obtained in any tube manual.
2. The wiring capacitance.

The list is short because it has been assumed that the output circuit is inductively coupled to the next grid. This coupling tends to separate the input capacitance of the next tube from the plate circuit of the preceding tube. However, if a direct connection is made to the next tube, the additional input capacitances must be taken into account.

From the foregoing brief discussion, it is quite evident that in designing R.F. television amplifiers of all types, tubes should be selected that have:

1. High mutual conductance values.
2. Low input and output capacitances.

It has been suggested that the usefulness of a tube may be determined by the ratio of (1) to (2), or

$$\frac{g_m}{C_{in} + C_o}$$

This ratio is called the "Figure of Merit" of a tube, and large values are desirable. It should be noted that both numerator and denominator of the ratio are important at the high frequencies. At the low frequencies, the tube capacitances have less importance and only g_m need be considered.

Tubes with Two Cathode Terminals. One final word about recent tubes which have been built with two cathode terminals. It has been found that the input impedance of vacuum tubes, which is ordinarily so high as to be considered infinite, begins to decrease as we raise the frequency of the signal. In the television channels above 50 mc, this tube loading on the attached tuned circuits causes a reduction in the gain and Q of the circuit. One of the causes for this reduction in tube input impedance is due to the inductance of the cathode leads within the tube itself. Why this is so can be seen from the following explanation.

The current of a tube must flow through the cathode lead wires and in so doing develops a voltage across the inductance of these wires. Note that this inductance is of importance only when the signal frequency is high. The average or d-c component of the current does not enter into this consideration. The voltage produced across the lead inductance, although due to the plate current, is impressed between the grid and the cathode. As a result, the effective signal voltage acting at the grid of the tube is lowered because of the opposition of the cathode lead voltage. The situation is analogous to inverse feedback, except that the lead-inductance voltage is present even though the cathode of the tube is grounded directly to the tube socket. The lead inductance occurs within the tube itself.

Note that the voltage which is developed across the cathode-lead inductance is due to the plate current. So far as the plate circuit is concerned, this voltage is of little significance. It is at the grid, where the signal is applied, that the voltage is important.

To eliminate the effect of the lead inductance voltage on the input grid circuit, tube manufacturers have designed tubes with two wires leading directly from the cathode structure inside the tube to the tube base. In this manner, one terminal is available for the grid circuit return and one

for the plate circuit and its current, and the two circuits are divorced from each other. In the circuit of Fig. 3-19 the 6AG5 R.F. amplifier tube possesses two cathode terminals. Even though both cathode terminals are grounded, pin 2 would be connected to the grid coil and capacitor. Pin 7 is the cathode connection for the plate circuit. To it would be connected the screen-grid and plate by-pass capacitors. The d-c plate current divides between both cathode terminals, but this is of no consequence since it does not contribute to the degenerative effect.

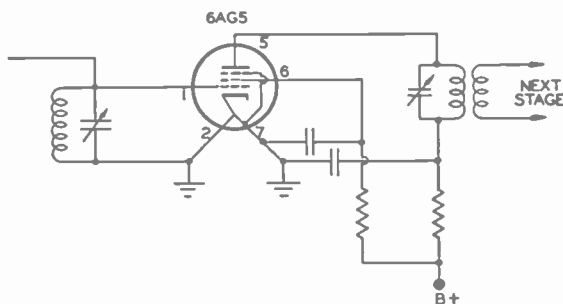


FIG. 3.19. The use of two-cathode wires to eliminate the adverse effect of cathode inductance.

Servicemen should be cautious, in this respect, in accepting the connections of the two cathode terminals as shown on the manufacturer's schematic diagram. The diagram is not always an exact representation of the circuit, as laid out in the chassis, especially with regard to the separate connections of the same cathode. In many schematics, the cathode is grounded, but the diagram does not indicate that a separation exists as explained for Fig. 3-19. If anyone who is not familiar with the reason for the separate cathode terminals connected them together, receiver sensitivity would decrease due to a lower input resistance.

In this chapter we have been concerned solely with the basic R.F. amplifier circuits found in television receivers. Nothing has been said about the tuner mechanisms themselves. This discussion will be deferred until after high-frequency oscillators and mixers have been covered, because all three stages are contained within the tuner housing.

QUESTIONS

1. How are wide-band tuning circuits achieved using conventional tuning circuits?
2. What is the difference in behavior between transformer-coupled tuning circuits using tuned and untuned primaries?
3. Why is the gain low in television circuits?

4. What is the usual purpose for including R.F. amplifiers in receivers?
5. Why are R.F. amplifiers especially useful in television receivers?
6. What desirable characteristics should R.F. amplifier tubes possess? Explain your answers.
7. Illustrate several types of coupling networks used in television receivers.
8. What is the advantage of using an overcoupled transformer in the input circuit and a single-peaked tuner in the plate circuit?
9. Explain the operation of the R.F. circuit shown in Fig. 3-13.
10. Draw the circuit of a grounded-grid amplifier. Explain briefly how it operates.
11. Explain the origin of all the capacitances associated with an R.F. amplifier.
12. Explain the term "Figure of Merit." Why is it useful?
13. Why is the cathode-lead inductance important in high-frequency tubes?
14. Draw the circuit of an amplifier in which the effect of cathode-lead inductance is minimized.
15. Explain what effect a narrow frequency response in the R.F. tuned circuits would have on the reproduced image.
16. What precautions should be noted when constructing an R.F. amplifier for use in a television receiver?
17. What effect would an inoperative R.F. amplifier tube have on the image? Explain your answer.

practice, although there exists a definite technical difference. A tube is a mixer only when a separate oscillator is used. Its action then merely mixes or combines the R.F. input signal and the oscillator signal to obtain the difference frequency, or I.F. A converter combines the action of mixing and generating the oscillator voltage within one envelope.*

With increase in frequency, the stability and output of the oscillator section of a converter decrease. At the relatively high frequencies required

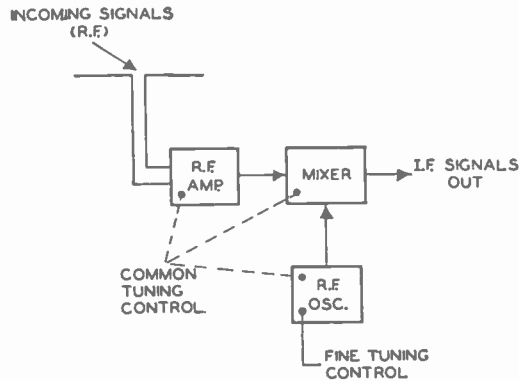


FIG. 4.2. A block diagram of the front-end section of television receivers.

for television, the *conventional* converter becomes unsatisfactory. The oscillator has a tendency to drift, and its output voltage is not constant. The only suitable method of obtaining sufficient oscillator voltage, without appreciable frequency drift, is to separate the oscillator from the mixer. (See Fig. 4-2.)

Miniature high-frequency triodes and pentodes appear to find equal application as mixers in current television receivers. Incoming signals, at this point in the receiver, are extremely weak, and all extraneous noise must be prevented from approaching signal strength. Triodes are superior to pentodes and the latter are superior to pentagrid converters because each element within a tube introduces a certain amount of noise into the circuit. In this respect, diodes would be the best tube of all to use, except that it is felt the additional amplification provided by triodes and pentodes outweigh whatever reduction in noise would result from the use of a diode. For the oscillator, triodes are almost universally used.

Energy from the oscillator may be capacitively or inductively coupled to the mixer. Two frequently used methods are shown in Fig. 4-3. Interaction between the input signal and the oscillator outside the mixer tube is kept as low as possible, to prevent any changes occurring in the oscillator

* Since the interchangeable use of these two words in industry is extensive, we will follow the same practice. No confusion should result.

frequency and to minimize oscillator radiations from appearing at the antenna. The latter tendency must be especially guarded against in sets that do not have an R.F. stage ahead of the mixer.

It has been observed that any considerable amount of radiated signal can produce a complete loss of contrast or even a negative picture in near-by television receivers. When the interfering frequency is close to the picture

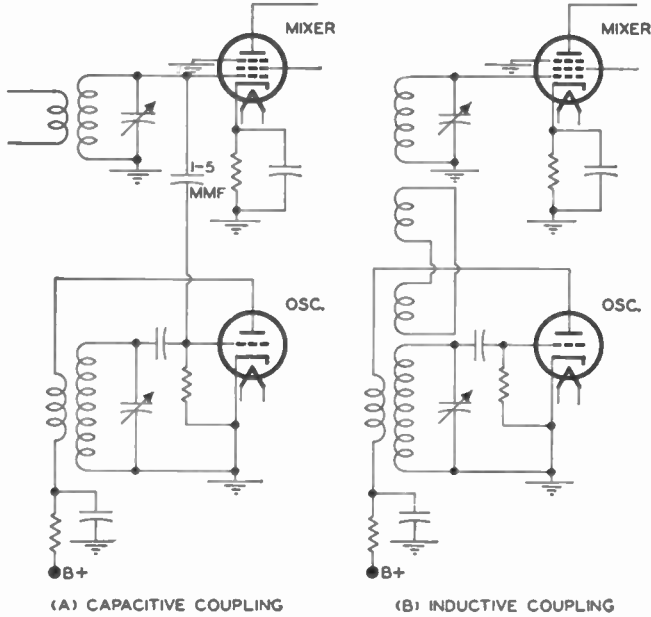


FIG. 4.3. Coupling energy from the oscillator to the mixer by inductive or capacitive means.

carrier of the station being received by the other sets, the “beat” interference produces vertical, horizontal, or slanted stripes across the screen.

Within the mixer of the television receiver, the received signal and the oscillator voltage both modulate the electron stream to form the desired audio and video I.F. voltages. In nearly all sets, the signal from the R.F. amplifier is transformer- or impedance-coupled to the mixer. The oscillator voltage, as indicated, is transferred to the mixer tube either capacitively or inductively.

Oscillators. Perhaps the most frequently employed oscillator circuit in current television receivers is the circuit shown in Fig. 4-4B. It is known as the ultraudion and is equivalent in its action to the well-known Colpitts circuit. (See Fig. 4-4A.) In the ultraudion, the voltage division across the tank circuit is accomplished through the grid-to-cathode (C_{gk}) and the plate-to-cathode (C_{pk}) capacitances within the tube. The feedback voltage which

sustains oscillations is developed across C_{gk} . In Motorola television receivers, the oscillator stage has the actual form shown in Fig. 4-5A and the equivalent network indicated in Fig. 4-5B. The voltage-dividing capacitance network consists of the effective capacitance of C_{gk} in series with the parallel combination of C_{pk} and C_1 , while C_t represents the combination of

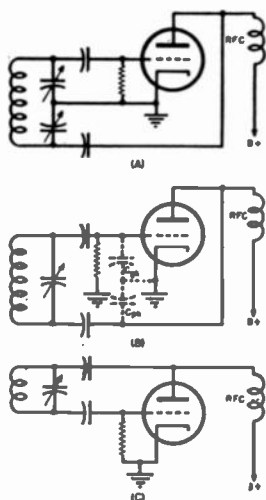


FIG. 4.4. Circuit diagrams show the similarities between the Colpitts (A) and the ultraudion (B) oscillators. The circuit diagram (C) illustrates how the ultraudion is usually drawn.

the grid-plate capacitance, the distributed capacitance, and C_2 . C_2 is a temperature compensating capacitor and helps reduce oscillator drift. In spite of this, some drift does occur, and C_1 is provided to permit the set user to adjust the oscillator frequency for the best sound output. Because C_1 is actually a vernier adjustment, it is labeled "fine-tuning control" and placed on the front panel. Any shift in oscillator frequency immediately alters the I.F. produced as a result of the mixing action. The effect is the same as detuning the receiver. By means of the "fine-tuning" control, the oscillator frequency can be readjusted to its proper value. Capacitors C_3 and C_4 keep the d-c plate voltage off the exposed coils. C_4 also makes it possible for the oscillator to develop grid-leak bias across R_1 . A separate coil is brought in for each channel, and each coil can be adjusted individually as to frequency. This is necessary because the oscillator frequency must be accurately set for each channel.

Modified Ultraudion. In the Colpitts oscillator of Fig. 4-4A, it is entirely possible to shift the placement of the ground connection from the cathode to the plate end of the tank coil without affect-

ing the operation of the oscillator. Now, however, we must provide a different d-c path from the cathode to ground, otherwise the tube electron flow will be interrupted. For this, we can use an R.F. choke which provides the necessary d-c path but which still maintains the cathode at the required R.F. potential. The corresponding shift in ground point can be applied to the ultraudion circuit, and an oscillator of this type is employed in G.E. television receivers. The actual circuit employed is shown in Fig. 4-6.

In sets using continuous tuning, no fine-tuning control is required. Since the station is tuned in manually, oscillator frequency drift can be compensated for directly by retuning.

Still another variation of the ultraudion circuit is shown in Fig. 4-7. This differs from the previous circuit in possessing the tuned circuit in its plate rather than in its grid circuit. Grid-leak bias is developed by the

30- $\mu\mu\text{f}$ capacitor across the 6,800-ohm resistor in the grid circuit. Grid current flows whenever the cathode assumes a potential which is more negative with respect to ground than the grid voltage present across the 30- $\mu\mu\text{f}$ capacitor. Actually, of course, this is the equivalent of driving the grid positive with respect to the cathode, whereupon the grid draws current. This action is similar to that in any conventionally connected tube.

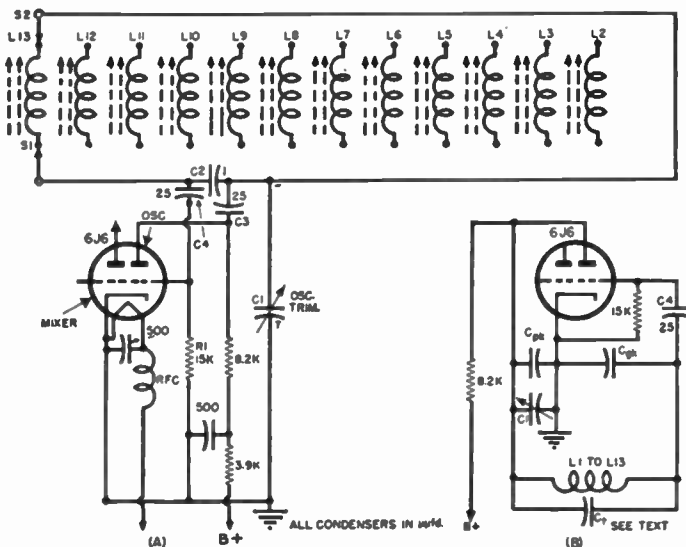


FIG. 4.5. The ultraudion oscillator arrangement (A) and its equivalent circuit (B) as employed in Motorola receivers.

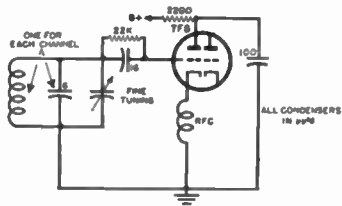


FIG. 4.6. The modified ultraudion circuit used by G.E. in their receivers.

In the circuit of Fig. 4-7A, the tuning coil L_1 remains in the circuit at all times. To change the resonant frequency of the oscillator, additional capacitors or coils are placed in parallel with L_1 . The added components used for different television channels are shown in Fig. 4-7B. A testing point is provided to permit servicement to measure the oscillator grid-leak bias voltage without actually going into the tuner.

Push-Pull Oscillator. In conjunction with the push-pull R.F. amplifier of Fig. 3-13, there is used a push-pull oscillator and a push-pull mixer. See Fig. 4-8A. The channel switch, as it progressively moves down the line, adds more and more inductance to the line, effectively lowering the tank frequency. For tuning, brass screws are placed near the high-frequency tuning straps L_{14} , L_{16} , L_{18} , L_{20} , L_{22} , L_{24} , and

L_{26} , and brass cores are adjusted through coils $L_2, L_4, L_6, L_8, L_{10}$, and L_{12} . For keeping the two sections of the line as closely balanced as possible, L_{11} and L_{25} are provided.

C_1 is a fine-tuning control which is connected across the entire tuning circuit. It provided a frequency variation of approximately plus or minus 300 kc on old channel 1, and this increases to plus or minus 750 kc on

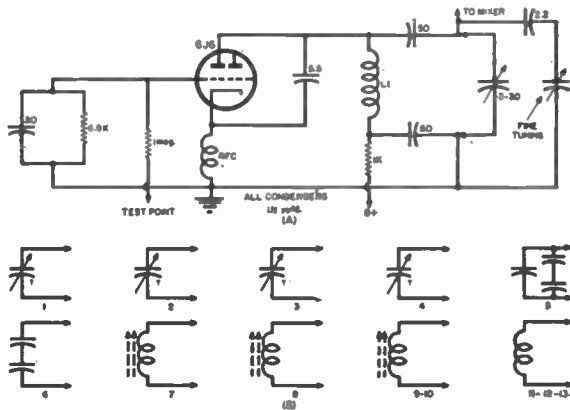


FIG. 4.7. (A) Another variation of the ultraudion oscillator. (B) The various tuning coils and capacitors that are switched into the oscillator circuit for the different television channels.

channel 13. Coupling between the oscillator and mixer is accomplished by a single turn of link coupling and by physically placing the oscillator tuning circuit close to the converter grid tuning circuit.

Oscillator Frequencies. In design, the oscillator frequency is generally placed above the incoming sound and video frequencies.* By being located above both signals, the highest I.F. produced will be that of the video signal. To illustrate, refer to Chapter 1, where it was shown that, in a 6-mc television channel, the audio carrier was 4.5 mc higher than the video carrier. For channel 2, 54–60 mc, the video carrier would be at 55.25 mc. (The remnants of the other sideband are from 54.00 to 55.25 mc, but these are useless and rapidly attenuated in the circuits.) From the video carrier, the picture sidebands extend for 4 mc up to 59.25 mc. The audio carrier would then be located at 59.75 mc.

* This is true in all receivers employing the conventional television system. In receivers using the Intercarrier system, the oscillator frequency is sometimes placed above the sound and video frequencies and sometimes below them. Where the latter is true, the sound and video signals retain the same relative position they had when broadcast. This does not affect operation of the receiver as long as the video I.F. system is designed for this condition.

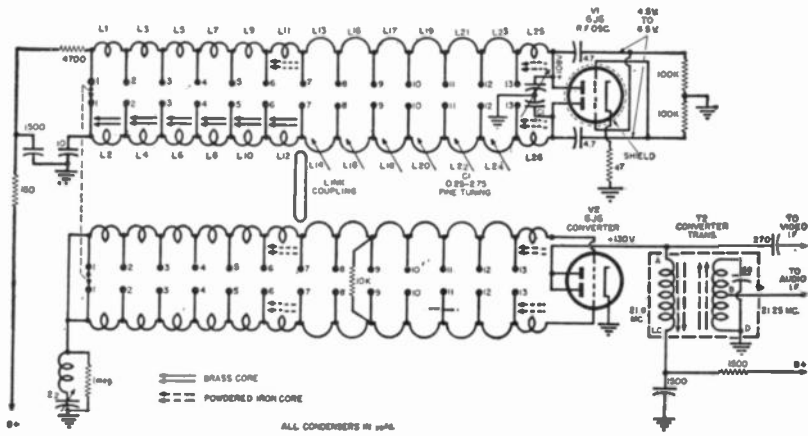


FIG. 4.8A. Push-pull oscillator and mixer circuits. These are used in conjunction with the circuit shown in Fig. 3.13.

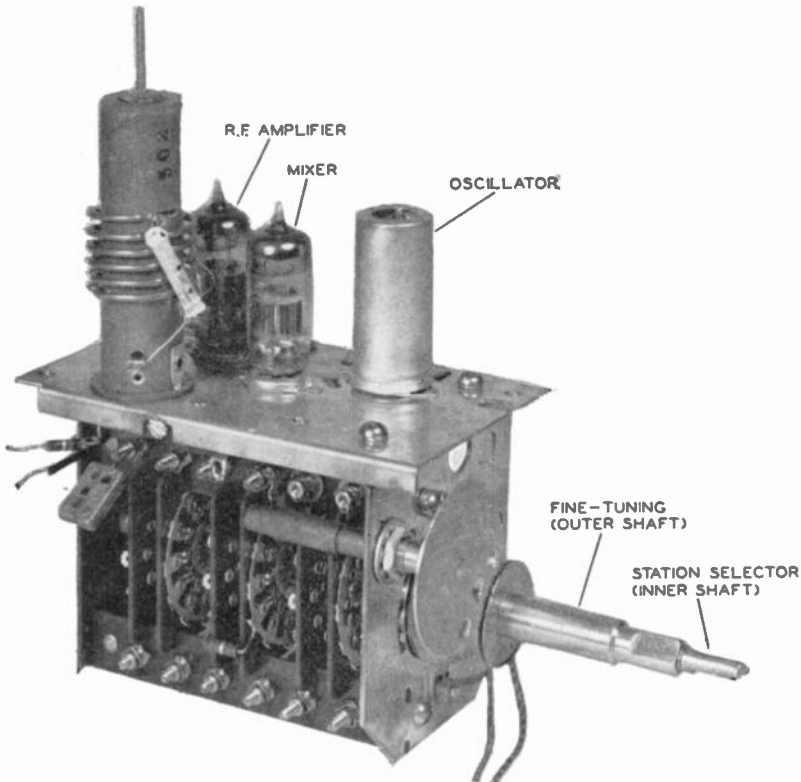


FIG. 4.8B. A rotary wafer switch tuner. The circuit of this unit is shown in Figs. 3.13 and 4.8A.

Now suppose that the oscillator frequency is 81.00 mc. In the mixer tube, the 54- to 60-mc signals would combine with the oscillator frequency to form the following I.F. signals:

1. For the video, the I.F. will range from 21.75 mc to 25.75 mc. This is the difference between 81.00 mc and 55.25 to 59.25 mc. Actually, the I.F. generated will extend to 27 mc. However, the vestigial sideband remnants are from 25.75 mc to 27 mc and are not desired. The I.F. bandpass tuning circuit eliminates them.

2. For the sound, the I.F. will be centered at 21.25 mc. F-M is employed for the audio transmission, resulting in a frequency variation of plus and minus 25 kc about this center (21.25 mc) position.

In Chapter 1, Fig. 1-23B, the standard video transmission characteristic curve is shown. The remnants of the lower sideband are permitted to remain because of the difficulties encountered in attempting to separate the lower sideband entirely from the upper sideband without affecting the phase or amplitude characteristics of the desired upper sideband. At the receiver, the remaining or vestigial lower sideband must be attenuated; otherwise it is found to produce unequal response at the video detector output. This latter form of distortion arises because the lower video frequencies are contained in both the upper sideband and the remnants of the lower sideband. The higher video frequencies are present only in the upper sideband, having been eliminated from the lower sideband. If the transmitted signal waveform is permitted to remain intact, there would be proportionally more low video frequency voltage produced at the second detector output than high video frequency voltage. To prevent this, a receiver response characteristic such as shown in Fig. 4-9 is employed. At the carrier frequency, the response is 50 per cent down, increasing linearly toward the higher frequencies and decreasing for the lower frequencies. Roughly speaking, the lower video frequencies, for which there are two sidebands, receive half the amplification accorded those higher frequencies for which there is only one sideband. The overall result is an equal response for both the low and the high video frequencies.

The characteristic shown in Fig. 4-9 is the response curve of the I.F. system of the receiver. In any superheterodyne, it is the I.F. stages which mostly determine the selectivity and sensitivity of the receiver.

Indicated, too, in Fig. 4-9 are other frequencies which are attenuated by means of trap circuits inserted in the various I.F. amplifiers. The reason for these circuits will be given presently.

R.F. Tuners. Now that R.F. amplifiers, oscillators, and mixers have been discussed we can turn to modern television tuners and see how these circuits are combined electrically and mechanically to receive one incoming signal to the exclusion of all others.

There are in use today a number of methods by which television stations are tuned in. All systems initially can be broken down roughly into continuous tuning or selector switch tuning. In the first method, which is somewhat similar to that used in radio receivers, the dial is rotated until the station is tuned in. Dial rotation is continuous and you move from station to station by turning the dial knob clockwise or counterclockwise. With selector switch tuning, each channel is assigned a fixed and definite set of

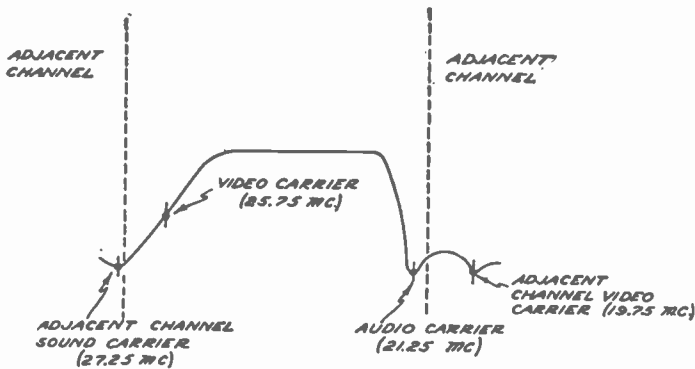


FIG. 4.9. An overall video I.F. response curve.

contacts on the switch. Each position of the switch is numbered; to receive a desired channel, the switch is rotated until the channel position is reached. Contact is here made with the tuning circuits of this channel, and any signal present is received.

Within each of these two categories there are a number of different types of tuners that have been developed and the more important of these will be discussed.*

SELECTOR SWITCH TUNING

Turret Tuner. The turret type of tuner, illustrated by the popular Standard Coil unit, is shown in Fig. 4-10. The coils for each of the 12 VHF channels are snapped into position on the rotatable circular drum. Separate coils are used for each channel. The antenna coil and the R.F. amplifier input coil are mounted on one form; the R.F. amplifier output coil, the oscillator coil, and the mixer input coil are all grouped together on another form. These two units then constitute the complete set of coils for one channel. The forms are of the snap-in type and can be removed readily if necessary. The popularity of this tuner is due to its rugged construction, freedom from trouble, and good sensitivity.

* UHF tuning methods will be described in Chapter 17.

A schematic diagram of this tuner is shown in Fig. 4-11. The input circuit is balanced with an impedance of 300 ohms. This is purposely designed to match the 300-ohm twin-lead line. However, by using one end terminal and ground, the input impedance becomes 75 ohms, and a shielded coaxial cable may be connected to the receiver without mismatch. The

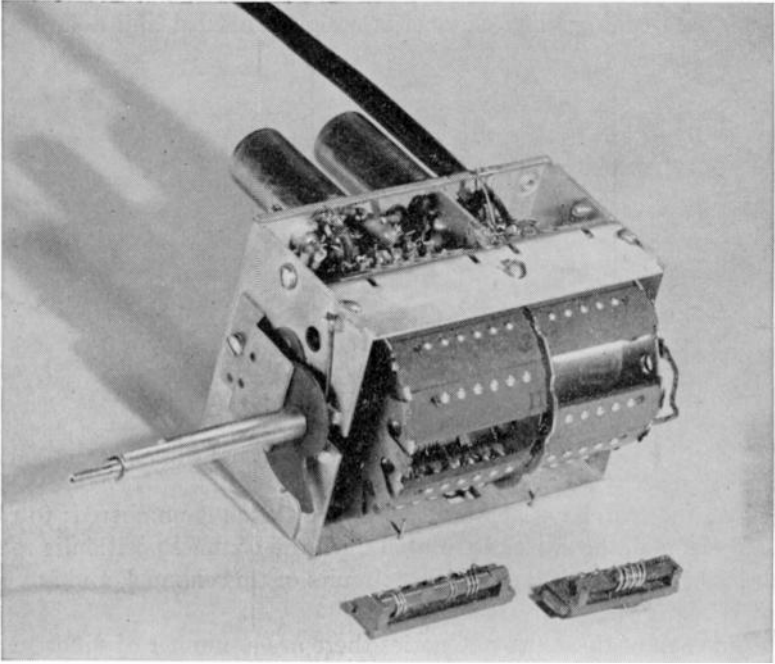


FIG. 4.10. The Standard Coil rotary turret tuner. The two sets of coils for one channel are shown being inserted.

R.F. amplifier tube may be a 6AG5, a 6CB6, or a 6BC5. These tubes have similar characteristics and are generally interchangeable, except for the slightly different pin connections of the 6CB6.

The secondary winding of the input circuit, L_2 , is tuned by the input capacitance of the R.F. amplifier tube in series with the parallel combination of C_1 and C_2 . Trimmer C_2 is used for alignment. The 3900-ohm resistor (R_8) across L_2 is inserted for the purpose of broadening the response of the input tuned circuit to the necessary bandwidth.

The plate load of the R.F. amplifier is L_3 in-conjunction with C_3 plus whatever tube and stray capacitance is present here. R_6 is again a loading resistor to widen the bandpass.

The gain of this stage is controlled by an A.G.C. voltage fed to the control grid of the R.F. amplifier through a 47,000-ohm resistor.

Signal transfer from the plate of the amplifier to the mixer grid is accomplished by inductive coupling between L_3 and L_4 . The combination of R_3 and R_4 across L_4 is not for loading, but rather to provide a testing point where an oscilloscope can be attached (at test point 9) to observe the response pattern of the R.F. amplifier tuned circuits. Also, test point 9 can be used as an injection point for video I.F. test signals.

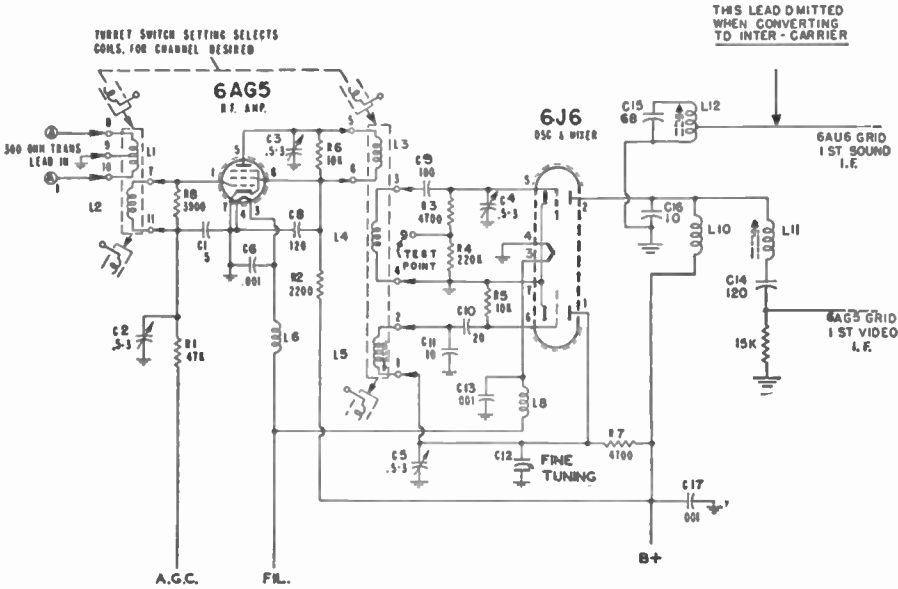


FIG. 4.11. Schematic diagram of the Standard Coil tuner.

R_3 and R_4 , together with C_9 , develop grid-leak bias for the mixer stage. Trimmer C_4 is used for R.F. alignment. The output of the mixer stage is coupled to the first video I.F. amplifier by means of the low-pass network composed for C_{16} , L_{10} , L_{11} , and C_{14} . Capacitor C_{14} is a d-c blocking capacitor.

Oscillator coil L_5 is inductively coupled to mixer grid coil L_4 to inject the oscillator signal into the mixer circuit. Capacitor C_{11} is in series with the parallel combination of C_5 and C_{12} to form the split-capacitor of a Colpitts oscillator. Trimmer C_5 is an R.F. oscillator adjustment. Variable-dielectric-type capacitor C_{12} is the fine-tuning control. Grid-leak bias for the oscillator is developed by R_5 and C_{10} . The oscillator plate is shunt-fed by means of R_7 .

L_{12} and C_{15} form a parallel resonant circuit tuned to the sound I.F. frequency (usually 21.25 mc). L_{12} is inductively coupled to L_{10} and absorbs a considerable amount of the sound I.F. signal present in L_{10} . A tap on L_{12} then provides direct coupling to the first audio I.F. amplifier grid.

In Intercarrier receivers, L_{12} and C_{15} are not present because in these sets the sound and video I.F. voltages do not separate until some point beyond the video second detector.

Zenith also has a turret tuner which is similar to that of the Standard Coil unit. See Fig. 4-12. Electrically the circuits differ, but the overall operation is the same.

Printed Circuit Tuner. It was inevitable that some day printed circuits would be adapted to television tuners just as they have been applied to other components in the television receiver. At the present time there is only

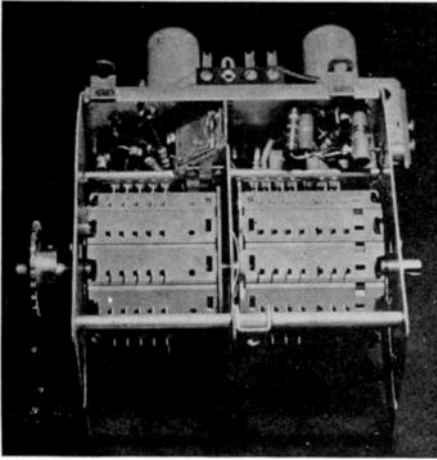
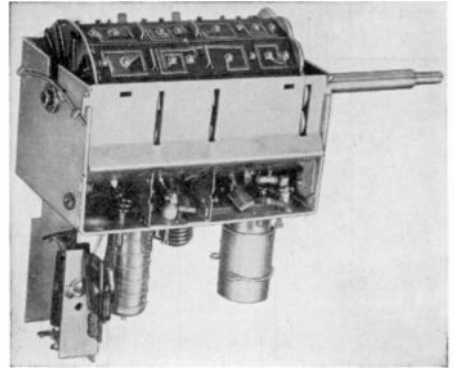


FIG. 4.12. The Zenith turret tuner.



Courtesy RCA

FIG. 4.13. The printed circuit TV tuner.

one printed circuit tuner in use. This is the turret-type unit which closely resembles the Standard Coil unit. See Fig. 4-13. However, in place of wound coils, this tuner possesses circuits obtained by a photoetching process on copper sheets.

Briefly, the photoetch process begins with the photographing of a circuit drawing. A contact print is then made from the negative in a copper-covered sheet of phenolic plastic which has been sensitized. The print of the plastic sheet is developed and placed in an etching solution. The solution dissolves away that part of the copper not covered by the circuit pattern. What is left is the copper circuit on the plastic sheet. This sheet is then placed on a die and cut into separate sections and pierced.

The thin copper strips are formed into individual coil strips for the various channels. All tuned circuits are printed in this manner except the oscillator coils for channels 2 to 6. These are wound in the normal manner.

A simplified circuit diagram of the tuner is shown in Fig. 4-14. The

input circuit contains a pair of elevator transformers (L_1-L_2 and L_3-L_4) which match the tuner to a balanced 300-ohm transmission line. They also tend to attenuate noise which may be introduced in the line. Following this is a high-pass filter with cut-off at approximately 47 mc and with maximum attenuation at approximately 23.5 mc. This is for rejection of I.F. frequencies.

A tuned input circuit at the grid of the R.F. amplifier provides gain and selectivity and reduces oscillator voltages reaching the antenna terminals.

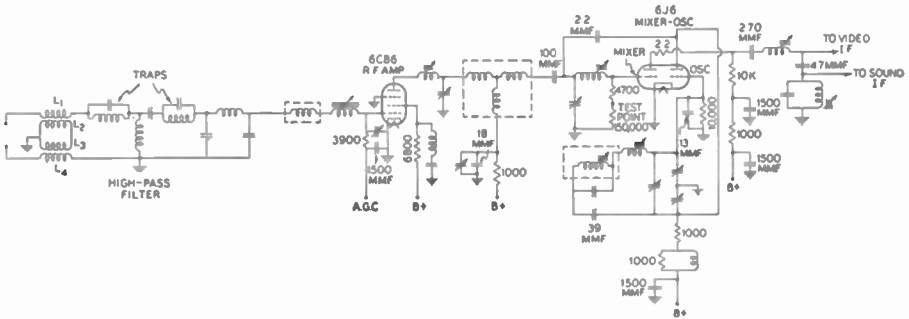


FIG. 4.14. Simplified schematic diagram of printed-circuit tuner. Components in dotted boxes are changed with each channel.

The R.F. amplifier is a type 6CB6 tube; its characteristics include high gain, low noise, and low grid-plate capacitance. The output of the R.F. amplifier contains a double-tuned bandpass filter with maximum attenuation at approximately the image frequency of each channel. The filter also cuts down voltages of oscillator frequencies at the plate of the R.F. amplifier. This is additional protection against any oscillator signal reaching the antenna and interfering with neighboring sets.

The oscillator is a temperature-compensated Colpitts. A fine-tuning control, adjusted by a shaft extending over the channel selector shaft, enables the viewer to make fine changes in the oscillator frequency. For each channel there is a separate oscillator screw adjustment which can be reached from the front of the cabinet.

The mixer plate circuit contains a tuned low-pass filter section for the video I.F. output and a high-Q trap for the sound I.F. output. This trap is tuned to 21.25 mc and attenuates sound I.F. frequencies at the video I.F. output.

Wafer-Switch Tuner. Another type of selector-switch tuning device is the wafer-switch tuner.* These units employ rotary switches with the

* Also known as wafer-tier tuners. The word "wafer" refers to the switching arrangement, whereas the word "tier" indicates that the switch contains several sections or tiers.

tuning coils mounted between each set of switch contacts. To accommodate the different R.F., oscillator, and mixer tuning circuits, multisection switches are employed. Each section would then contain the various coils for one tuning circuit over the 12 VHF television channels.

The RCA tuner, shown in Fig. 4-8B, was the first type of wafer-switch tuner to appear on the market after World War II. From the circuit diagrams of this unit, Figs. 3-13 and 4-8A, it can be seen that 6J6 double

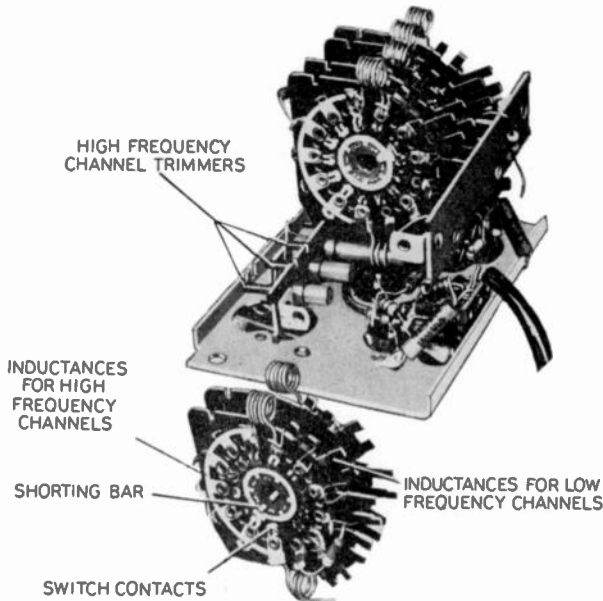


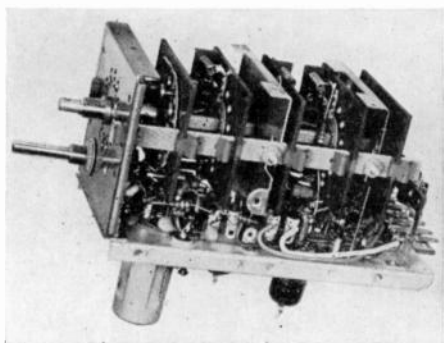
FIG. 4.15. An exploded view of the wafer assembly of the tuner shown in Fig. 4.8B. (Courtesy RCA.)

triodes serve as the R.F. amplifier, local oscillator, and mixer. Push-pull circuits are used throughout with appropriate tuning circuits.

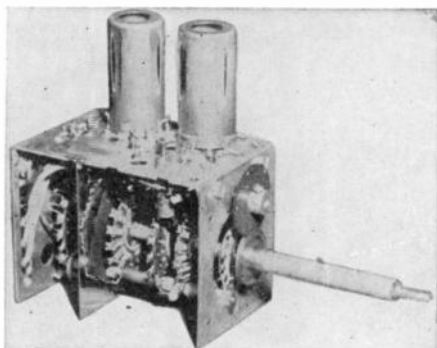
The exploded view of the wafer assembly of this tuner in Fig. 4-15 shows how the various low- and high-frequency inductances are mounted between the switch contacts. The low-frequency coils have several turns; the high-frequency inductances are nothing more than small silver straps between each set of contacts.

In later models of this tuner, the R.F. amplifier was first changed to a pentode for greater gain and then more recently to the cascode circuit (to be described presently). The mixer was made into a pentode and the oscillator remained a triode. The wafer switching arrangement together with the series connection of the inductances has been essentially retained.

Photographs of other wafer-switch tuners are shown in Figs. 4-16 and 4-17. In the G.E. unit, there are two R.F. amplifiers followed by an oscillator and a mixer. The first R.F. amplifier is a grounded-grid triode amplifier (a 6AB4) followed by a 6AK5 pentode R.F. amplifier. In the Sarks-Tarzian unit, a 6AK5 R.F. amplifier is followed by a 6J6 mixer. The other half of the 6J6 serves as the oscillator.



Courtesy G.E.



Courtesy Sarks-Tarzian

FIG. 4.16. The wafer-switch tuner of G.E.

FIG. 4.17. The Sarks-Tarzian tuner.

CONTINUOUS TUNING

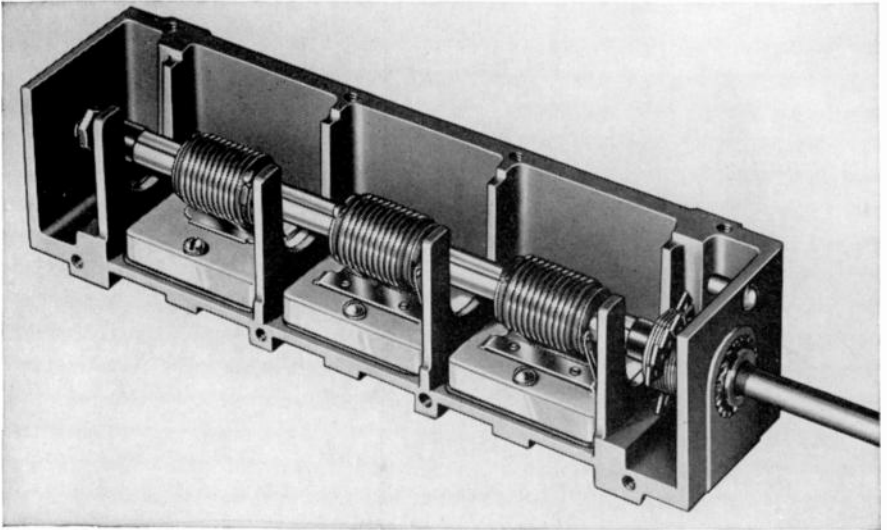
Continuous tuning over the twelve television channels has a certain amount of attraction because not only does it do away with the fine-tuning control but it also enables the receiver to pick up the FM signals which are broadcast in the 88-to-108 mc band.

A number of continuous tuners have been used. Of these, the Inductuner shown in Figs. 4-18 and 4-19 has enjoyed the greatest popularity. This unit consists of three separate variable inductances mounted on a common shaft. The coils, each containing ten turns, are wound on ceramic forms, and movable trolley sliders make contact at each point on the coil.

Each trolley is grounded, and, as it moves along, more and more turns of each coil section are shorted out, progressively raising the frequency of the circuit. The inductance can thus vary from 0.02 to 1.0 microhenry, permitting a tuning range from 54 to 216 mc.

The use of the "Inductuner" in television receivers is illustrated in Fig. 4-19. The input circuit leading to the antenna transmission line is an unbalanced arrangement having an impedance of 72 ohms. The signal is received in the cathode leg of the R.F. amplifier because the tube is being operated as a grounded-grid amplifier with both triode sections connected in parallel. The signal output of the 6J6 R.F. amplifier is transferred to the mixer via the complex coupling network of L_1 , L_{2A} , C_1 , C_2 , C_3 , L_3 and

L_{2B} . L_{2A} and L_{2B} form two sections of the three-section Inductuner. L_1 and L_3 are end coils, inserted in series with the Inductuner windings to obtain a better response at the high end of the band. Each of these coils has but a few turns, and, when the trolleys of the Inductuner are run out to the end of their windings, only the end coils (L_1 and L_3) represent the inductance



Courtesy P. R. Mallory Co.

FIG. 4.18. The physical appearance of the Inductuner.

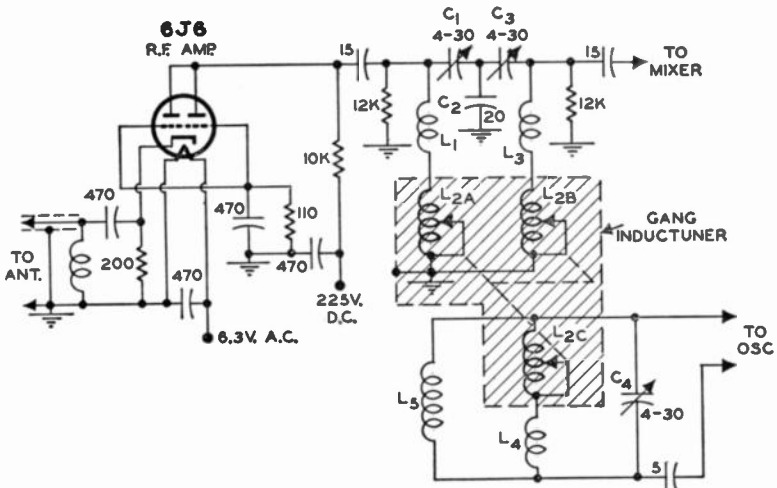


FIG. 4.19. Circuit diagram showing the use of the Inductuner in the R.F. section of a television receiver.

still remaining in the circuit. In this way the high-frequency response can be kept up.

The third winding of the Inductuner is employed in the oscillator stage together with an end coil, L_4 , a shunt coil, L_5 , and a shunt capacitor, C_4 . The shunt coil is required because the oscillator frequency is higher than the incoming signal frequency by an amount equal to the I.F.

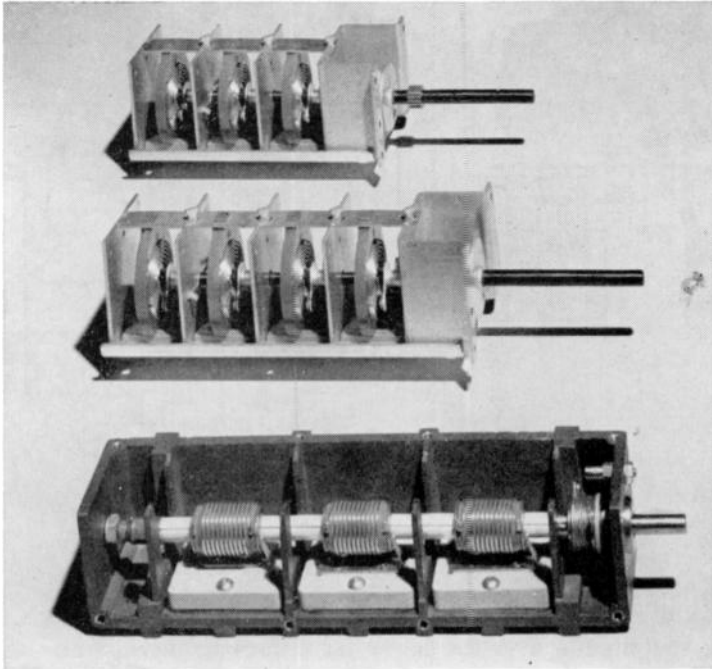


Fig. 4.20. A comparison of the new spiral Inductuners (top two) with the older unit (bottom).

Recently a modified Inductuner has appeared using a spiral type of winding (see Fig. 4-20). Operation of this unit is identical to that of the older Inductuner, except that now the contact arm moves around a spiral instead of along a solenoid. The advantages gained are greater compactness, lower cost, and increased mechanical stability. The smaller size permits a fourth winding that can be used at the input to the R.F. amplifier.

Fig. 4-21 contains the schematic diagram for an R.F. section using the new 4-section Inductuner. The input terminals of transformer T_1 allow the use of either 75-ohm or 300-ohm antenna systems. Connection is made to the center (grounded terminal) and one of the outside terminals for 75-ohm input, whereas the two outside terminals are used for 300-ohm input.

The secondary of input transformer T_1 is loosely coupled to the tuned circuit through capacitor C_1 .

The tuned input circuit comprises the inductors L_1 plus L_{2A} resonated by capacitors C_1, C_2 (with C_3 and L_3 in parallel), C_4 and the input capacitance of tube V_1 . In this circuit C_4 tracks channel 6, C_2 tracks channel 2, and inductor L_1 is employed to track channels 7 to 13. Damping of the circuit for proper bandwidth is accomplished by resistors R_1 and R_2 .

A double-tuned bandpass circuit couples the plate of the R.F. tube V_1 to the grid of the mixer tube V_2 . A compensated system of bottom side coupling comprising circuit elements L_5, L_7, C_{11}, L_8 and C_{13} provides cor-

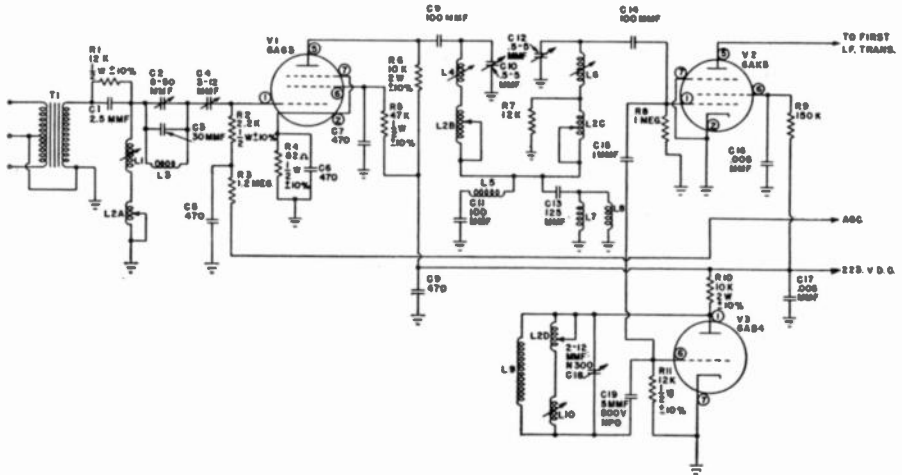


FIG. 4.21. Circuit employed with spiral Inductuner.

rect bandwidth and signal transfer over the entire tuning range. C_{10} and C_{12} track the circuits over the lower television channels, while L_4 and L_6 track the tuner circuits over the high television channels.

The oscillator circuit is of conventional type in which the feedback voltage from plate to grid is accomplished by means of the interelectrode capacitance of tube V_3 . The oscillator tuned circuit comprises the variable inductor L_{21} , with series inductor L_{10} and shunt inductor L_9 , this combination being resonated by adjustable capacitor C_{18} and the tube capacitance. The low television channels are tracked by C_{18} , while inductor L_{10} serves to track the high television channels.

The oscillator is coupled to the grid of the mixer through capacitor C_{15} .

Other types of continuous tuners include a permeability tuner and a capacitance tuner. In the permeability tuner, the movable slugs of each coil are mechanically ganged together and moved in or out of the coils when the front panel knob is rotated. There is one set of coils for the low band and one set for the high band. A special switch on the front panel affects the change-over from one band to the other. The capacitance tuner

is similar to radio receiver tuners, employing a number of variable capacitors ganged together by a common shaft.

Cascade Amplifiers. The ability of a receiver to amplify a signal is not limited by the amplification which can be obtained from vacuum tubes but by the noise which arises from the tubes and the associated receiver networks. Furthermore, the noise that is developed by the first stage (the R.F. amplifier) is actually the most important because at this point in the

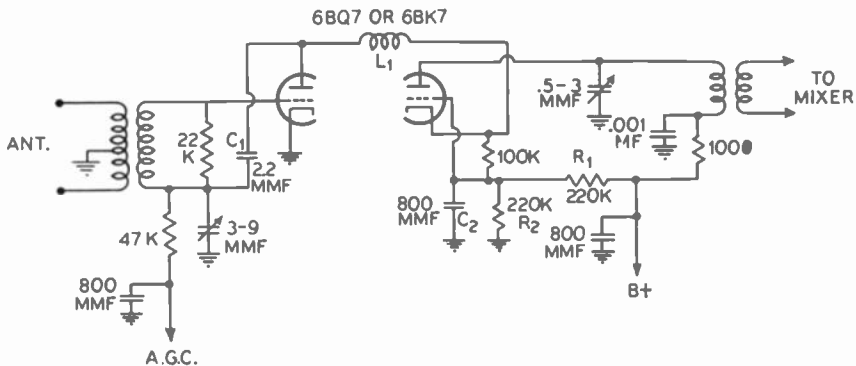


FIG. 4.22. A cascode R.F. amplifier.

system the level of the incoming signal is more nearly on a par with the noise level than it is at any other point in the receiver. Whatever noise voltage appears at the grid of the R.F. amplifier is amplified along with the signal and, to obtain the best noise-free picture, we want to have as much signal and as little noise as possible at the front end of the set.

The best choice for low noise is a triode R.F. amplifier. This is because noise originating in a tube varies directly with the number of positive elements within that tube and a triode has fewer elements than a pentode. Unfortunately, however, the gain of a triode is less than that obtainable from a well-constructed high-frequency pentode.

A recent circuit development makes it possible to achieve low noise with good gain. The name of this circuit is "cascode" and its schematic is shown in Fig. 4-22. Basically it consists of two triodes connected in series, that is, the plate of the first section goes directly to the cathode of the second section. The same current flows through both tubes, and the amount of current is controlled by the bias on the first triode.

The input tuned circuit of this series amplifier connects to the control grid of the first triode; the output circuit is in the plate lead of the second triode. The first stage is operated as a conventional amplifier; the second stage is employed as a grounded-grid amplifier. The coil L_1 between both stages serves to neutralize the grid-to-plate capacitance of the first triode

(with help from capacitor C_1) and it is designed to resonate with the grid-cathode capacity of the second section on the high VHF channels. While L_1 thus aids the stability of this combination, it is also largely responsible for the low noise qualities of the cascode circuit.

Direct coupling is used between the first triode plate and the second triode cathode. With cathode feed to the second triode, C_2 is used to place the grid at R.F. ground potential. Since the two triode sections are in series across a common plate supply, the cathode of the second triode is 125 volts positive with respect to chassis ground. A divider across the plate supply, consisting of R_1 and R_2 , places the grid of the second triode at a sufficiently positive potential (with respect to its cathode) for proper operating bias.

The cascode circuit is being widely adopted by the various receiver manufacturers, both in their VHF tuners (such as those described) and in their UHF circuits, too.

Television Boosters. Television receiver owners located in areas where the signal strength is weak often attempt to improve the quality of their pictures by adding an external booster to their sets. Boosters are basically nothing more than R.F. amplifiers, and when one of these is attached to your set, you are, in effect, adding one or more R.F. amplifiers to that already existing in the receiver.

Your purpose in doing this is to strengthen the incoming signal to such an extent that it will produce a picture possessing the full contrast range and, at the same time, improve the signal-to-noise ratio so that the picture will be clear and free of annoying noise spots. Of these two objectives, the improvement of the signal-to-noise ratio is the more difficult to attain — and it is the more important. So let us pause and examine the relationship between noise and weak signals.

As stated previously, the ability of a receiver to amplify a signal is not limited by the amplification which can be obtained from vacuum tubes but by the noise which arises from the tubes and the associated receiver networks. This noise is known as random noise because it possesses no fixed frequency, but extends from zero to frequencies far above any being used today.

The noise that is developed in a receiver is due to two sources, thermal agitation in conductors and electron flow through tubes. Thermal agitation arises from the random motion of electrons within a conductor. There is no external voltage applied, but the electrons, using their own energy, move to and fro along a conductor. This movement of electrons constitutes a current flow. Since, at any given instant, a few more electrons are moving in one direction than in the other, a voltage is set up in the conductor which is proportional to the net current flow and the value of the conductor re-

sistance. The polarity of the voltage due to thermal agitation changes constantly, electrons moving first in one direction and then in another. Because of this, there is no definite pattern to the random voltage, or, for that matter, any one frequency at which the energy changes. It has been found that the energy of this disturbance is distributed uniformly throughout the entire frequency spectrum used for communications.

The second source of receiver noise is developed in the tubes. While there are several components to this noise, the most important component is due to the shot effect. The current that flows in a tube is not a continuous fluid but a moving aggregation of separate particles, the electrons. Noise voltages are produced, even when so-called steady currents are flowing, because at any single instant, the number of electrons impinging on the plate differs from the number reaching this electrode at any other instant. Over a measurable period, the current is steady, but instantaneously it fluctuates rapidly due to the non-fluid nature of electrons. It is these instantaneous fluctuations that represent the noise.

In a receiver, the noise that is developed by the first stage (the R.F. amplifier) is actually the most important because at this point in the system the level of the incoming signal is more nearly on a par with the noise level than it is at any other point in the receiver. Whatever noise voltage appears at the grid of the R.F. amplifier is amplified along with the signal; hence, to obtain the best noise-free picture, we want to have as much signal and as little noise as possible at the front end of the set.

From these facts concerning noise, we see immediately that in choosing a booster we want one which has a low noise figure. This is certainly as important as gain. You can have all the gain in the world, yet if a large noise voltage exists at the grid of the first amplifier stage of the booster, you will get a high noise voltage output. And nothing you can do thereafter will reduce the noise.

Thus, boosters are designed with two aims in mind: To improve the signal-to-noise ratio and to amplify the weak incoming signal. Both are important and both are needed. A booster capable of high gain but incapable of providing a good signal-to-noise ratio will give a picture filled with disturbing noise spots. A booster possessing a minimum of internal noise but capable of little gain will not amplify the signal sufficiently to permit it to override the set noise. So again the picture will be covered with noise spots. Thus your booster must have both attributes or it might as well have none.

Right here we have the reason why a booster may not always help the set. Assume that in the regular TV receiver (without a booster) the noise voltage existing in the R.F. amplifier is 10 microvolts and the received signal is 30 microvolts. This is a signal-to-noise ratio of 3 to 1. We figure

that, if we place a booster ahead of our set, we should get a clearer picture. This may not be so.

Suppose our booster, in its input circuit, develops a noise voltage of 20 microvolts. The incoming signal is still 30 microvolts. If the booster gain is 10, then what the R.F. amplifier will receive is 200 microvolts of noise and 300 microvolts of signal. What is the signal-to-noise ratio now? 3 to 2, which is not as good as 3 to 1.



Courtesy Regency

FIG. 4.23. A television booster.

Thus, in this instance, this booster will not help you improve the quality of your picture.

On the other hand, if the noise existing in the R.F. amplifier of the booster is less than 10 microvolts and the same signal of 30 microvolts is received, then the signal-to-noise ratio will improve and, with it, so will the quality of your picture.

Before we leave this subject of noise, it should be pointed out that we have said nothing about noise generated outside the set or the booster. This noise, if present, comes down the transmission line with the signal and it is indistinguishable from the signal as far as the booster is concerned. To overcome this noise we must attack it at its source, or, if this is not feasible,

to try to keep as little of it as possible from reaching the signal via the antenna or the lead-in line. Standard methods of attack include increasing antenna height, antenna replacement, and the use of shielded lead-in line. It has also been found helpful to position the booster at the antenna (or at least as close to the antenna as possible). This serves to strengthen the signal before it has been subjected to the noise and thus enables it, with its amplified strength, to better overcome the adverse effects of the noise. The reader will recognize that here again we are acting to improve the signal-to-noise ratio.

A commercial booster is shown in Fig. 4-23. It is connected between the receiver and the antenna. All boosters are self-powered, developing their own d-c voltage from the a-c power lines. The R.F. booster circuits are similar to the R.F. amplifiers already discussed. The trend, at present, is toward the use of cascode circuits, but single or push-pull amplifiers are also found. Designers seem to be evenly divided between triodes (for low noise) and pentodes (for gain), although, where the cascode circuit is used, the tube is invariably a duo-triode.

QUESTIONS

1. Why are pentagrid converters seldom, if ever, found in television receivers?
2. What is the difference, technically, between a mixer and a converter?
3. Draw the circuit of an ultradion oscillator.
4. Indicate the differences in circuit between an ultradion oscillator and a Colpitts oscillator.
5. Explain the purpose of a "Fine Tuning" control.
6. Where is the "Fine Tuning" control placed, electrically, in the circuit? Draw a circuit using this control.
7. Illustrate two methods of coupling the oscillator signal to the mixer.
8. How is the oscillator frequency for each channel obtained, using a single tube? Illustrate your answer.
9. What is the mathematical relationship between the oscillator frequency and the video and audio frequencies?
10. Explain why the video carrier is placed where it is on the video I.F. response characteristic.
11. What is the difference between a turret type of tuner and a wafer-tier type of tuner?
12. Name several types of continuous TV tuners.
13. Explain briefly how a printed circuit tuner is formed.
14. Draw the circuit of a cascode R.F. amplifier.
15. What is a booster and where is it used?
16. Explain why a booster may not always be beneficial to a television receiver.
17. Identify two sources of noise within a television receiver which determine the quality of the reproduced picture.

Chapter 5

VIDEO I.F. AMPLIFIERS

Choice of Intermediate Frequencies. The television signal at the output of the mixer has been reduced to its I.F. values. The change occurred in the mixer where the incoming signals were beat against the local oscillator signal. The I.F. amplifiers will now amplify these voltages and then transfer them to a detector where the intelligence originally imparted to the signal will be removed.

In order to understand why the particular I.F. values were chosen, let us pause and determine what factors are most important in the design of an I.F. system. Basically there are three factors to consider. These are:

1. Frequency of the I.F. stages.
2. Gain.
3. Selectivity.

The procedure is to choose the operating frequency first and then to consider the problems of gain and selectivity together.

The choice of an intermediate frequency may appear, at first, to be quite simple since we know that at the lower frequencies it is easier to construct amplifiers which have high gain. However, there is a limit to how low a frequency can be used because of the stability of the circuits ahead of the I.F. system and because of the bandwidth required by the television signal itself. When a set is first turned on, it may require as much as an hour before the oscillator frequency stops drifting. When ordinary parts are used in the construction of a receiver, the oscillator may drift as much as 0.2 per cent in frequency. At 60 mc, this means a drift of 120 kc. Although a shift of this magnitude may not noticeably affect the reproduced image, it can certainly affect the television sound. This is especially true of conventional-type television receivers where the sound and video signals separate before the video second detector. Hence, the stability of the oscillator and R.F. circuits will be governed by the F-M audio signal considerations.

In the interests of stability, a low I.F. value is indicated. However, there are the video I.F. amplifier requirements to consider.

The I.F. tuned circuits must pass a band of frequencies 4 mc wide. Suppose we use the I.F. values which were used in the television receivers of 1939—8.75 mc to 12.75 mc. At the second detector, the demodulated video frequency voltages extending from 0 to 4 mc would have to be separated from the I.F. values, 8.75 to 12.75 mc. To effect a clear-cut separation between the video frequencies and the I.F., it is desirable to have their ratio as high as possible. At the low broadcast frequencies, the sound "spread" is only 5 kc, which is a small fraction of the 465 I.F. Thus, no difficult problem exists here. But in a television receiver, as noted above, the separation between the desired and undesired frequencies is considerably less and the problem becomes more difficult. A high I.F. is desirable as this would simplify the problem of separation. As the I.F. value rises, the gain and stability decrease.

In addition to the above, there are various types of spurious responses capable of affecting a receiver and they, too, influence the choice of an intermediate frequency. The more important of these spurious responses are: *

1. Image response.
2. Response of two stations separated in frequency by the I.F. value.
3. Direct I.F. response.

1. *Image Response.* Image response is due to the mixing of an undesired signal with the local oscillator signal in the converter stage to produce a voltage at the intermediate frequency. Since a frequency equal to the intermediate frequency is produced, this signal will be accepted and passed by the I.F. amplifiers. As an illustration, suppose a television receiver had an I.F. carrier value of 12.75 mc. This means that its bandpass extends from 12.75 mc to 8.75 mc. Further, suppose the set is tuned to the 54–60 mc television channel, No. 2. With the I.F. value specified, the local oscillator would be operating at 55.25 mc plus 12.75 mc or 68.00 mc. If a powerful station is, at the same time, operating in the 76–82 mc channel, its signal will, in some measure, appear at the mixer stage input. Mixing of this signal with the oscillator voltage within the mixer tube will produce signal voltages which will be at and sufficiently close to 12.75 mc to be accepted by the I.F. amplifiers. The result, at the cathode-ray tube, is distortion.

By choosing an I.F. value which is greater than half the entire band to be covered, it is possible to eliminate image response from that band entirely. For the widest VHF television band, 174–216 mc, this requires an I.F. value in excess of 21 mc. (The UHF band, of course, is even wider, being 420 mc wide. Half of this is 210 mc. However, the I.F. chosen for

* One more spurious response which stems from neighboring television receivers is discussed on p. 105.

a receiver should not be greater than the lowest frequency to be received, in this case, Channel 2, 54 mc.)

2. *Stations Separated by the Intermediate Frequency.* The second listed source of interference is due to stations separated by the intermediate frequency value. In this situation, one incoming signal acts as the mixing oscillator for the other signal, their difference frequency appearing at the output of the mixer or converter stage at the intermediate frequency.

There are two solutions to this problem. One is to provide sufficient discrimination in the circuits preceding the mixer so that they will reject two signals so widely separated in frequency. The other is to provide a high I.F., one which is slightly greater than the entire band. In the upper-frequency (VHF) television band, this would be 42 mc.

3. *Direct I.F. Response.* The third form of spurious response is due to the direct reception of a signal equal in frequency to the I.F. itself. To avoid the need of incorporating special filters, wave traps, and shielding to prevent interference from this source, an I.F. is chosen whose frequency is not used to any appreciable extent for commercial or amateur transmissions. This accounts for such seemingly odd values as 10.7, 9.1, etc.

The foregoing discussion has, by no means, exhausted the subject of spurious responses. However, the important contributing factors have been covered and it is possible to see how they affect receiver design and operation. For the video I.F., a high value is indicated, and to a certain point, the higher the better. This will not only reduce interference from spurious responses but also simplify the problem of filtering the video signal in the detector output from the I.F. Opposing the use of a high I.F. are the disadvantages of reduced gain, necessity for greater care in selecting components to prevent excessive losses, additional shielding, and greater tendency of feedback through the tubes and adjacent circuits and the relatively narrow bandwidths of the audio F-M circuits. For the latter, stability in the local oscillator is highly important, and since stability is more readily achieved at the low frequencies, a low I.F. is desirable. With all these considerations and in view of the fact that, at the moment, it is more important to reduce spurious responses and other outside interferences from reaching the screen, a relatively high I.F. is used.

When television was introduced commercially in 1946, the Radio and Television Manufacturers Association (RTMA) recommended a video carrier I.F. frequency between 25.75 and 26.4 mc. If 25.75 mc were chosen (and it was extensively used), then the complete video signal would extend from 25.75 mc to 21.75 mc. This is a 4-mc range. The corresponding sound I.F. value would be 4.5 mc lower than the video carrier I.F. value. In this instance 25.75 mc — 4.5 mc gives a value of 21.25 mc. At the time these recommendations were made it was felt that these frequencies were high

enough to minimize the amount of interference to which a TV set would be subjected and yet not so high that manufacturing obstacles would be difficult to overcome.

On the whole, the choice of these particular frequencies has been found to be a good one. There has been a certain amount of interference from medical and industrial electronic equipment operating in the 21 to 27 mc range and from harmonics of near-by amateur stations. FM stations (88–108 mc) have also caused image interference. However, what was somewhat unexpected, and what has proved to be very annoying, has been the interference that comes from neighboring television receivers.

Suppose your set is tuned to channel 5 (76–82 mc) and your neighbor's set is tuned to channel 2 (54–60 mc). If his video I.F. value is 25.75 mc, then his local oscillator is operating at 81.0 mc. This is so because 55.25 mc (video R.F. carrier) beating against 81.0 mc produces the difference frequency of 25.75 mc. Now, if some of this 81.0-mc voltage is radiated from your neighbor's antenna to your antenna, a series of fine stripes will appear across your screen.

By a similar type of reasoning it can be shown that interference can be produced also on channels 6, 11, 12, and 13. In each case, of course, the interfering receiver would have to be tuned to a different channel. With the high concentration of television receivers in large cities and suburban areas, this problem can be a very serious one to the TV set owner. There are ways by which a certain amount of relief can be achieved by altering the I.F. values of the interfering receiver. However, this represents an added expense to the set owner and from a commercial standpoint is not very desirable.

In view of this situation, studies and research were carried on to determine I.F. values which would eliminate some of these undesirable interferences. The results revealed that a sound I.F. of 41.25 mc and a video I.F. of 45.75 mc would serve much better, and these are the frequencies now recommended by the RTMA. The circuits for either set of frequencies are similar and will be used interchangeably in the discussion to follow.

Separation of Video and Sound Signals. The television signal in the plate circuit of the mixer has been reduced to its I.F. values. What happens to the video and sound signals now depends upon the type of system being employed in the receiver. In the Intercarrier type of set, both sound and video signals remain together, at least until the video second detector. In the conventional type of receiver, separation of the video and sound signals takes place *before* the video second detector. The separation may occur at the plate of the mixer, or it may occur in any one of the video I.F. amplifiers. But wherever the separation, only the video signal remains by the time the video second detector is reached.

Since signal separation in conventional-type receivers must take place before the video second detector, it may be instructive to pause here briefly and examine typical separation methods. In one popular method, the converter transformer (T_1 in Fig. 5-1) is a combination video I.F. transformer, sound trap, and sound I.F. transformer. The primary winding of T_1 is peaked at 21.8 mc, but since its frequency response is fairly wide, all the video and sound frequencies appear across it. The secondary of T_1 is a

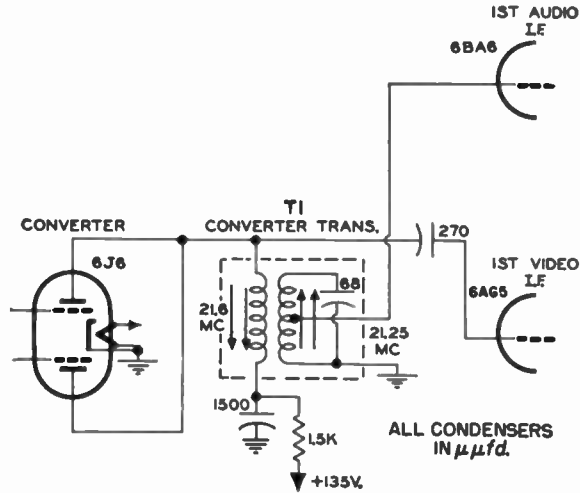


Fig. 5.1. A signal separation network used by RCA and many other television manufacturers.

high- Q resonant circuit, sharply peaked to 21.25 mc, the sound carrier I.F. value. The resonant circuit absorbs the sound I.F. component from the primary, transferring it to the grid of the first I.F. amplifier. The rest of the signal on the primary of T_1 is unaffected by the trap and continues on into the video I.F. system.

It should be noted that the secondary of T_1 , although it absorbs a considerable portion of the sound I.F. voltage appearing across the primary, does not absorb all of it. Hence, a second, and sometimes a third, sound trap will be found at a subsequent point in the video system.

Another means of signal separation is shown in Fig. 5-2. The tuned circuit in the plate of the mixer tube is designed to pass signals having frequencies between 21.5 and 26.4 mc. The sound I.F. frequency, together with a portion of the video signal, is picked off at the tuned circuit and applied to the grid of the first sound I.F. amplifier. Although the full signal (sound and video) is applied to the sound system, only the F-M sound signal passes through because of the limiter stage which precedes the F-M detector

and the limited response of the tuned circuits in this section. The limiter removes all amplitude modulation, permitting only the frequency variations to pass.

Signal separation by means of a series resonant circuit is shown in Fig. 5-3. C_1 , in conjunction with L_1 , forms a series resonant path from the plate of the mixer to ground at the sound intermediate frequency. (C_2 , shunted across L_1 , is used to achieve another effect. This will be discussed in a subsequent section dealing with trap circuits.) In any series resonant circuit,

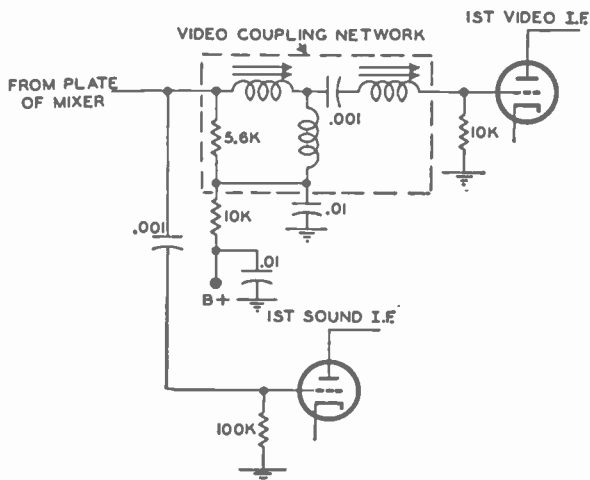


FIG. 5.2. In the circuit above, the entire signal is applied to the sound I.F.; however, because of the design of this section, only the F-M sound portion of the signal reaches the sound detector.

the opposition offered to an applied voltage is low when its frequency is equal to the resonant frequency of the circuit. Maximum current then flows through each component in this series branch. Moreover, and this is important, the voltage across each component at resonance is Q times the applied voltage, where Q is the Q of the circuit (generally this means of the coil). Thus, if the applied voltage is 5 and the Q is 5, the potential across the series capacitor and inductance, each, will be 25 volts. It is true, of course, that the voltage across the coil is 180 degrees out of phase with the voltage across the capacitor, and these, taken together, nullify each other. However, if we consider each component by itself, the rise in voltage is considerable. In the circuit of Fig. 5-3, the voltage fed to the grid of the first sound I.F. amplifier is developed across the coil of the series resonant circuit. At the sound intermediate frequency, a resonant rise in voltage occurs, providing a sizeable voltage for the grid of the first sound I.F. amplifier. At

the same time, C_1 , C_2 , and L_1 form a trap, reducing the amount of sound I.F. signal penetrating the video I.F. system.

Still another method of signal separation is exemplified by the circuit shown in Fig. 5-4. Two tuned circuits are connected in series in the plate circuit of the mixer. The top resonant circuit receives the video I.F. signal while the lower resonant circuit is sharply tuned to the sound I.F. Currents of both signals pass through both coils, but the greatest voltage drop in the

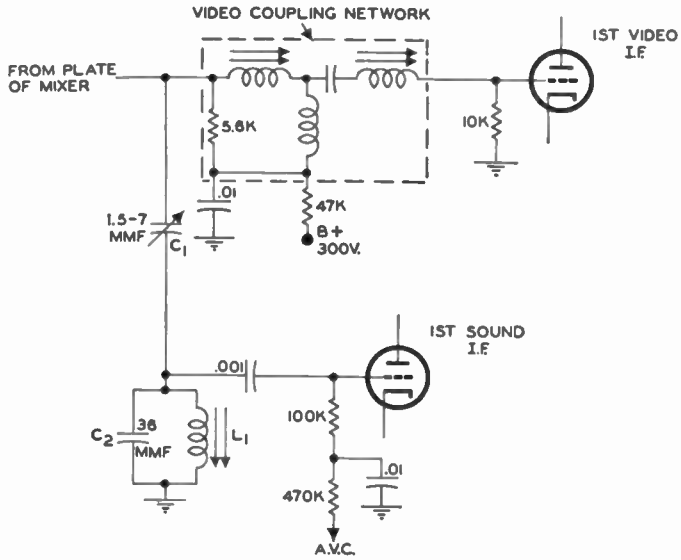


FIG. 5.3. A series resonant circuit, C_1 and L_1 (the latter in conjunction with C_2), separates the sound and video I.F. voltages.

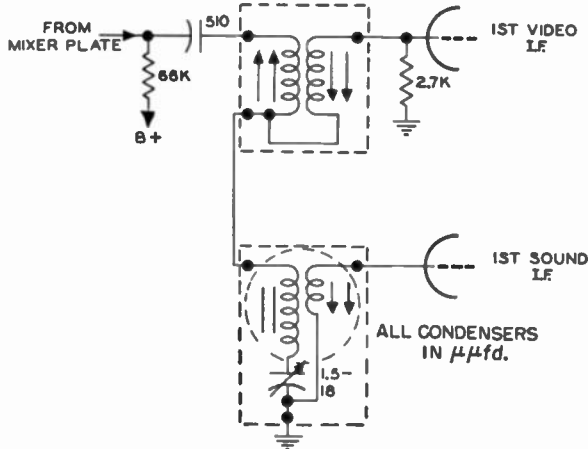


FIG. 5.4. Sound and video signal separation using two tuned circuits in series.

top coil will be at the video I.F. and the greatest voltage drop in the lower coil will be at the sound I.F. The energy in each resonant circuit will then transfer to its respective system.

In some receivers, separation of the signals does not occur until the output of the first I.F. amplifier or even later. Thus, in Fig. 5-5, the signal is first passed through the first I.F. amplifier, whereupon a third winding is coupled to the interstage transformer and the energy it absorbs is fed to the control grid of the first sound I.F. Sometimes the manufacturer labels this first stage as the first video I.F. Actually, a better designation would be, simply, first I.F., since both video and sound I.F. signals are amplified.

Most conventional-type receiver designers effect the video and sound I.F. separation at the output of the first I.F. stage. However, there are some receivers which delay this operation until the second I.F. or even third I.F. amplifier. By delaying the separation, the advantages of additional amplification are gained, which can be employed to reduce the number of sound I.F. amplifier stages. On the other hand, there is danger that enough sound F-M voltage will reach the video second detector to produce interference across the video screen.

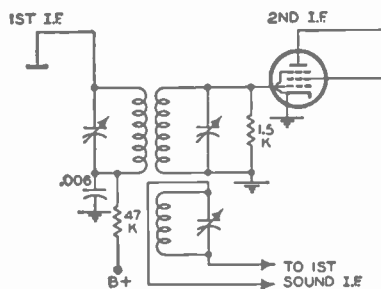


FIG. 5.5. Sound and video signal separation by means of a third winding on the interstage transformers.

Video I.F. Amplifiers.* In any superheterodyne circuit, the major portion of the overall gain and selectivity is contributed by the I.F. amplifiers. (See Fig. 5-6.) Consequently, it is most important for all those working with television receivers to be familiar with the shape of the I.F. response characteristic and to understand why this particular form was chosen. The reason, as we will see, lies with the signal transmitted by the broadcast station.

A television signal, when broadcast, possesses the form shown in Fig. 5-7. It is amplitude-modulated, but differs from conventional A-M signals in having essentially only one sideband. The other sideband, of which some remnants are still present, has been effectively suppressed. This is known as vestigial sideband transmission and is the standard in modern television. When any carrier is amplitude-modulated, an upper and lower sideband form automatically. However, because identical information is contained in each sideband, only one sideband is required at the receiver.

* The video I.F. amplifier section remains essentially the same for conventional and Intercarrier operation. Such differences that do exist will be brought out as we proceed.

Complete suppression of the lower sideband is the goal, but it is not economically achievable. It is impossible to eliminate completely one sideband using simple filters without, at the same time, distorting near-by portions of the remaining sideband. Hence, as a compromise between economy and easily adjustable circuits on the one hand, and minimum distortion

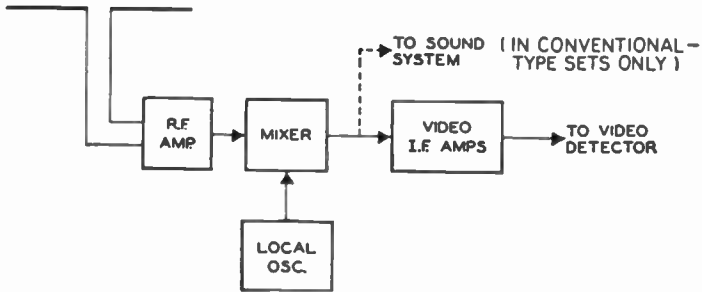


FIG. 5.6. The position of the video I.F. amplifiers in the television receiver.

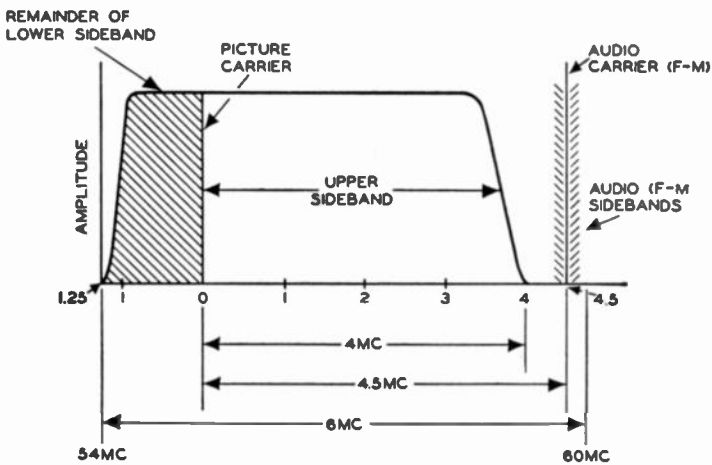


FIG. 5.7. The form of the television signal as broadcast by the television station.

and bandpass on the other, it was decided to remove all but 1.25 mc of the 4.0-mc lower sideband of the video signal. The transmitted signal, then, consists of this 1.25 mc plus the carrier plus 4.0 mc of the upper sideband. With the addition of the near-by audio carrier and its sidebands, the full 6.0 mc allotted to each television station are obtained.

Within the receiver we must take the upper sideband, together with the remnants of the lower sideband, and provide a response characteristic in which all sideband frequencies will have available an equal amount of amplification. In sound A-M sets this presents no great problem because both

sidebands are alike. But things are different in the television signal. Here, there is 1.25 mc of the lower sideband and 4.0 mc of the upper sideband. The lower video frequencies (those having frequencies close to the carrier) are contained in both the upper sideband and the remnants of the lower sideband. On the other hand, all video frequencies above 1.25 mc are present only in the upper sideband, having been suppressed in the lower sideband. If both the low and the high video frequencies are accorded equal amplification in the receiver, proportionately more low video frequency voltage will be developed at the second detector output than high video frequency voltage.

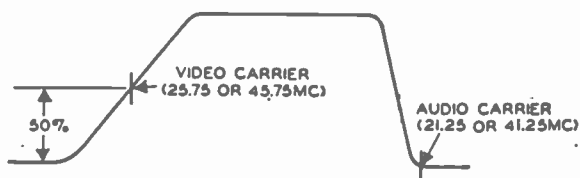


FIG. 5.8. The desired response curve of the video I.F. system.

It is to prevent this that the receiver response characteristic shown in Fig. 5-8 is employed. At the carrier frequency the response is 50 per cent down, increasing linearly toward a maximum for the higher frequencies and decreasing for the lower frequencies. Roughly speaking, the lower video frequencies, for which there are two sidebands, receive half the amplification accorded all video frequencies above 1.25 mc. In this way, we equalize the response for the low and the high video frequencies. To the serviceman, the shape of this curve is important in his work on television receivers. When aligning the I.F. stages, he must be careful to place the video carrier close to the 50 per cent point. At the same time, the circuits should be adjusted to provide the maximum bandpass. Detail in a television image is dependent upon the strength of the high video frequencies present. When the response curve drops at the upper end of the curve, fine detail becomes fuzzy and indistinct. Poor low-frequency response gives rise to poor synchronizing action, smearing, and a generally darker image.

Video I.F. systems can be conveniently classified according to the type of interstage coupling employed. All systems can be placed into at least one of the following categories; some make use of two.

1. Transformer coupling.
2. Stagger-tuned coupling.
3. Complex coupling.

Transformer Coupling. Typical of the transformer-coupled video I.F. systems is the circuit shown in Fig. 5-9. Transformers T_1 and T_2 are over-coupled and loaded with resistance to provide approximately a 4-mc band-

pass frequency characteristic. A third winding on T_2 functions as a trap to attenuate certain undesirable frequencies. (A complete discussion of trap circuits will be given presently.) The overall response of the video I.F.

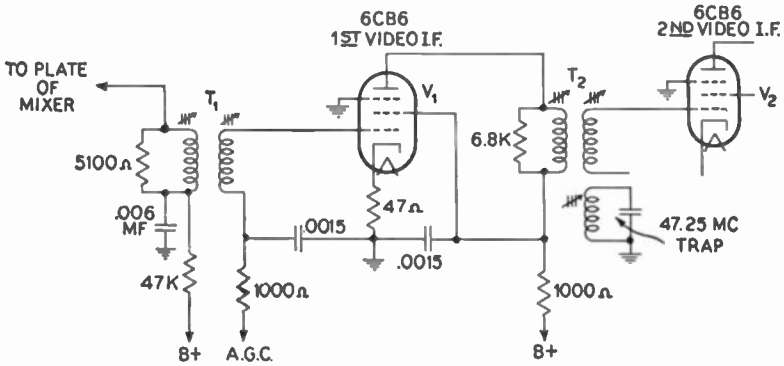


Fig. 5.9. A typical transformer-coupled video I.F. amplifier. The addition of the trap circuit to T_2 does not alter the operation of this stage.

system (of which Fig. 5-9 is one stage) is shown in Fig. 5-10.* The curve is approximately 4.0 mc wide, extending from 41.75 mc to 45.75 mc. The I.F. value of the video carrier is 45.75 mc and as such receives only 50 per cent of the total amplification. The video frequencies containing the image information then extend from this point to 41.75 mc.

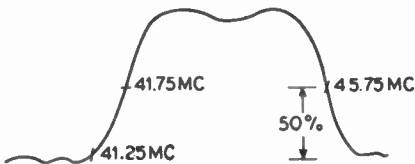


Fig. 5.10. The overall response of a complete transformer-coupled I.F. system.

While a 4.0-mc bandpass is to be preferred, little loss in picture quality will be noted for pass bands of 3.5 mc. In many sets, the response begins to fall at 3.0 mc. A narrower bandpass permits the manufacturer to effect economies in design and, in a field as competitive as television, cost is always an important consideration.

Of course, this practice cannot be carried too far, or picture quality will be materially degraded.

In Fig. 5-9, only one complete I.F. stage is shown. Actually, there would be three or four, this number being required to provide sufficient amplification for the signal. The first and second stages would generally receive A.G.C. bias to help maintain a constant output signal (at the detector)

* Video I.F. response curves may be shown either with the frequency increasing from left to right (as in Fig. 5-10) or with the frequency rising from right to left (as in Fig. 5-8). Both types of presentations are used extensively and interchangeably throughout the industry. Also, while the ideal curve is shown in Fig. 5-8, slight variations (as in Fig. 5-10) are common.

for fluctuating input signals. The use of automatic gain control is now widespread, although in early receivers it was not always found.

Since trap circuits are an important adjunct to video I.F. systems, it might be worth while to consider their form and mode of operation at this point.

TRAP CIRCUITS FOR TELEVISION RECEIVERS

Television receivers are continually beset by a host of interfering signals, any one of which is easily capable of distorting or completely destroying the desired image. Fortunately, many of these interfering signals never get beyond the input tuner and consequently are suppressed before any harm is done. Some, however, are so close to the channel frequency to which the set is tuned that they are able to penetrate the R.F. defense and reach the video I.F. system. This system actually constitutes the main bulwark against all interference and once a signal passes through these stages successfully, there exist few stop-gaps between the video detector and the cathode-ray tube. Every effort must be made to suppress any signal which is capable of distorting the reproduced image.

In order to receive a 6-mc band of frequencies, the R.F. and converter tuning circuits are designed with a low Q . This means that the sides of the input curve are not very steep, like the sides of a rectangle, but tend to taper off gradually. Fig. 5-11A illustrates a typical input characteristic of a modern television receiver. With a response of this type, voltages at the frequency of the sound carrier of the next lower channel or the picture carrier of the next higher channel could penetrate through the R.F. stages and reach the video I.F. system. Unless trap circuits are inserted in the video I.F. system (or even beyond), the effect of these interfering signals will be detrimental to the image.

To ascertain the frequency of the most important interfering signals, suppose the receiver is tuned to channel 3 (60-66 mc) and the video carrier I.F. value is 26.4 mc. The sound carrier of the next lower channel (54-60 mc) is at 59.75 mc. A signal at this frequency when mixed with the local oscillator (87.65 for channel 3) will develop a difference frequency of 27.9 mc. This is one interference signal frequency.

The second interference signal is the picture carrier of the next higher channel (66-72 mc). When this beats with the same local oscillator, a difference frequency of $87.65 - 67.25$ mc, or 20.4 mc, is produced.

The two frequencies — (1) sound from the next lower channel (27.9 mc) and (2) picture carrier from the next higher channel (20.4 mc) — must be eliminated from the video signal before it reaches the cathode-ray tube. Channel 3 is used in this illustration, but any other channel could have been

chosen. In all cases where closely adjacent channels exist, the two interfering frequencies will be 27.9 and 20.4 mc for this receiver. Note, however, that there are channels which are not subject to such interference. Thus, channel 2 (54-60 mc) does not possess an adjacent lower channel; on the

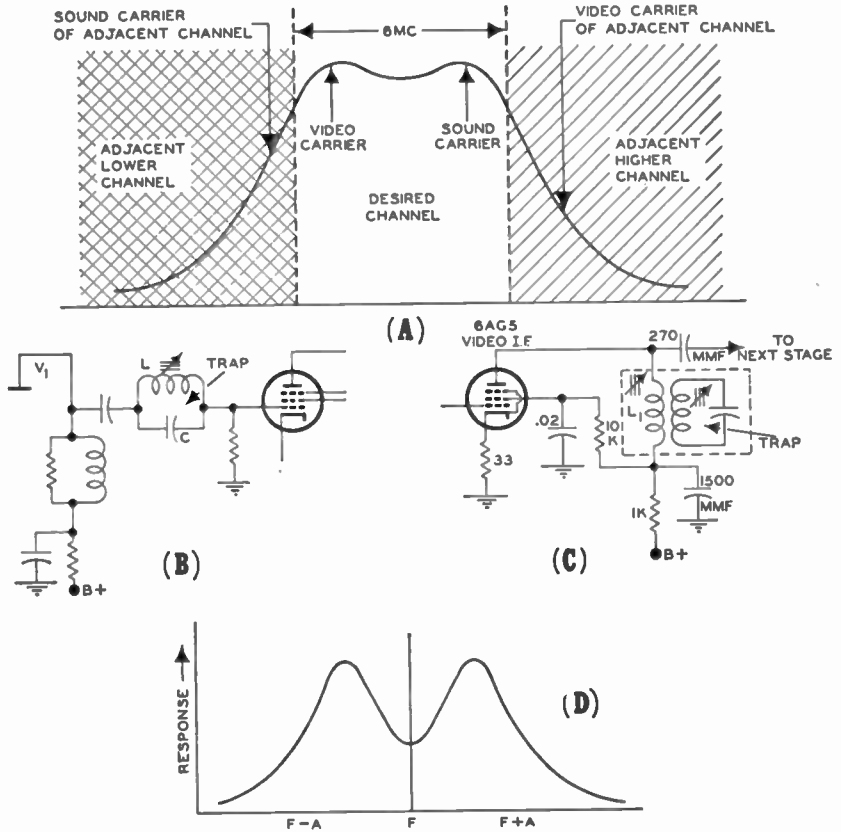


FIG. 5.11. (A) The R.F. response curve of most television receivers. Note that signals from adjacent channels can be received. (B) A series trap circuit. (C) An absorption trap circuit. (D) The double-humped response curve characteristic of closely coupled circuits.

other hand, channel 4 does not possess a closely adjacent higher channel. By "closely adjacent" we mean channels which follow each other without any frequency separation. Channel 4 is followed by channel 5, but the end of channel 4 and the beginning of channel 5 are separated by 4 mc. Channel 4 is 66-72 mc; channel 5 is 76-82 mc. This 4-mc separation is sufficient to prevent any of the frequencies in channel 5 from adversely affecting channel 4. However, channel 2 (54-60 mc) is closely followed by channel 3 (60-66

me), and interference is possible. The same is true of many of the other channels — hence the importance of using these traps. The trap frequencies will vary with the video and audio I.F. values employed in the circuit. The purpose of the traps, however, remains unchanged.

The objection may be raised that traps are not really necessary because under existing regulations adjacent channels are not assigned to *any one community*. This is true. However, communities which are located comparatively near to one another are assigned to adjacent channels. Thus, consider New York and Philadelphia, only 90 miles apart. New York is assigned VHF channels 2, 4, 5, 7, 9, 11, and 13; Philadelphia is assigned VHF channels 3, 6, and 10. Any set situated between these two cities would certainly be subject to considerable interference and would definitely require trap circuits. This same situation is true in many other parts of the country.

Adjacent channel traps could be used in conventional and Intercarrier type receivers.* On the other hand, there is one additional trap that is always used in conventional type receivers and occasionally used in Intercarrier sets. This is a trap designed to attenuate the sound I.F. voltage and prevent it from reaching the picture tube. It will be remembered that sound and video signal separation in conventional receivers occurs before the video second detector. However, while a substantial portion of the sound signal is removed at the separation point, enough energy does remain to cause interference to the picture. Indications of sound interference is the appearance of either a wormy, fine-line pattern (Fig. 7-23) or a series of dark horizontal bars (Fig. 7-24), or both. (See p. 178 for details on these patterns.)

In conventional-type receivers this trap is more important than any of the other traps because adjacent channel stations are not always present, but the sound of the same channel is.

In Intercarrier receivers a sound trap is sometimes used. Here its purpose is not to reject the sound I.F. signal completely, but rather to make certain that it is kept at its proper low level.

Type of Traps. (a) *Series Traps.* A series type of trap circuit is shown in Fig. 5-11B. It is placed between two I.F. stages and tuned to the frequency to be rejected. Trap circuits are sharply tuned networks, designed to reject one frequency or, at most, a narrow band of frequencies. When a signal at the trap frequency appears at the plate of V_1 (Fig. 5-11B), the impedance offered by LC is high, absorbing most of this interference voltage. A negligible amount appears across the input circuit of the following I.F. amplifier. At all other frequencies, the tank offers negligible impedance, and the desired signal passes easily.

* Whether or not a receiver contains these traps is largely a matter of design and economics. Some sets have one or the other trap; some have both and some have none.

(b) *Absorption Traps*. The absorption trap, shown in Fig. 5-11C, is the most widely used type of rejection circuit. It consists of a coil (and parallel fixed capacitor) inductively coupled to the plate load of an I.F. amplifier. When the I.F. amplifier receives a signal at the resonant frequency of the trap circuit, a high, circulating current develops in the trap (as a result of the coupling between the trap and the plate coil), and the voltage in the video coil, at the trap frequency, becomes quite low. Consequently, very little of this interference voltage is permitted to reach the next amplifier.

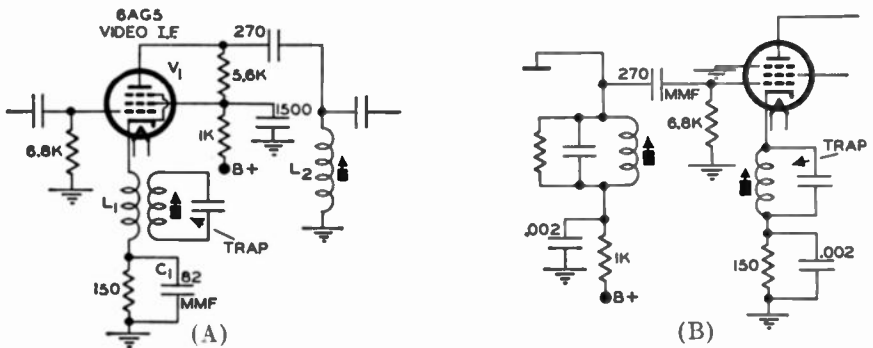


FIG. 5.12. Two types of cathode or degenerative trap circuits.

To explain this action more fully, two resonant circuits closely coupled will give a double-humped curve, such as shown in Fig. 5-11D. Note the sharp decrease in primary current at the center frequency. In the case of the two tuned circuits of Fig. 5-11C, L_1 is tuned to a band of frequencies while the secondary is sharply set at one frequency. Since the primary band coverage includes this one frequency, there is a sharp drop in primary voltage at this frequency, due to the presence of the trap. It is this interaction between coils which produces the marked decrease in voltage at the trap frequency. The other frequencies in this signal are unaffected by the trap.

(c) *Cathode or Degenerative Traps*. A cathode trap, shown in Fig. 5-12A, is a sharply tuned parallel resonant circuit coupled to a coil in the cathode leg of a tube. The coil L_1 in series with C_1 forms a broadly tuned series resonant circuit to the same range of frequencies as L_2 . This permits V_1 to function as a normal amplifier for all signals within this frequency range. However, at the resonant frequency of the trap, a high impedance is reflected into the cathode circuit (by the trap), and the gain of this stage is reduced by degeneration.

Another type of cathode trap, illustrated in Fig. 5-12B, places the parallel circuit directly into the cathode leg. At the resonant frequency of the trap the impedance in the cathode leg will rise, producing a large degenerative

voltage and reducing the gain of the amplifier. At all other frequencies, the impedance of this parallel network is low, and only a small degenerative voltage appears. There is thus only a slight loss in gain.

(d) *Parallel Traps.* Parallel traps are series resonant circuits which are placed across (or in shunt with) the circuit. (See Fig. 5-13.) At the frequency for which the trap is set, it acts as a short circuit, by-passing all

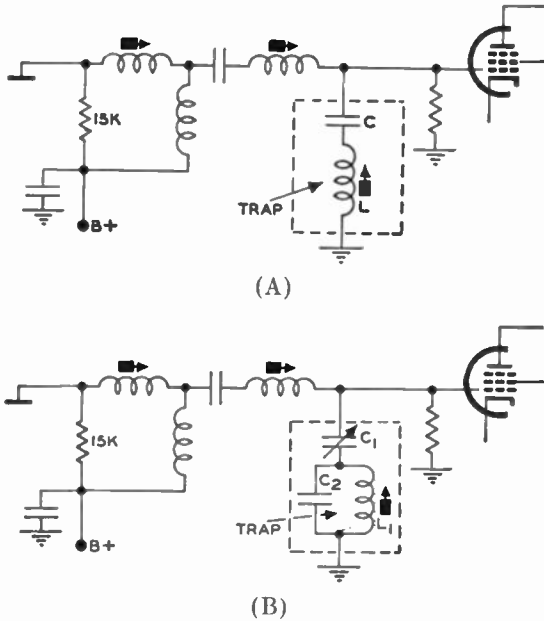


FIG. 5.13. Two types of parallel trap circuits.

signals of the same frequency to ground and preventing their further penetration into the circuit. At other frequencies the trap circuit presents a relatively high impedance, permitting these signals to proceed farther into the circuit.

The simplest type of parallel trap is illustrated in Fig. 5-13A. A capacitor and a variable coil, in series with each other, are connected across the signal path. Any signal possessing the same frequency as the trap will "see" a virtual short circuit upon reaching this portion of the circuit. The simple capacitor and coil trap is effective if the Q of this network (principally the coil) is quite high. However, when the Q is low, the tuning of the network is broad and more than one frequency is attenuated. A more efficient parallel trap is shown in Fig. 5-13B.

At first glance, the trap network appears as a parallel resonant circuit rather than a series resonant circuit. Let us, however, analyze its operation.

The illustrative example, Fig. 5-13B, is taken from a DuMont television receiver. It is designed to by-pass or trap the audio carrier voltage of the signal being received. In this instance the I.F. frequency is 21.9 mc.

If we consider first L_1 and C_2 , ignoring C_1 for the moment, then we have a parallel resonant circuit, tuned to 22.9 mc. The coil Q is 200, and a fairly large voltage is developed across the combination at this frequency. Now, it is easily demonstrated that, for all frequencies lower than its resonant frequency, a parallel circuit appears inductive. (At the resonant frequency, of course, it presents a purely resistive impedance.) For frequencies above resonance, the impedance presented is capacitive. This latter fact can be understood by noting that for higher frequencies, the parallel capacitor offers less impedance than the coil. Consequently most of the current flows through the capacitor, and the circuit current possesses a leading phase. Since the audio I.F. of 21.9 mc is below the 22.9-mc resonant frequency of L_1 and C_2 , the parallel combination appears inductive to the audio I.F. signal. By resonating this inductance with C_1 , we obtain a series resonant path for the audio I.F. signal, and the audio I.F. voltage is by-passed.

By providing a parallel resonant circuit (L_1 and C_2) for 22.9 mc, we obtain a sharp rise in voltage just beyond 21.9 mc. Since the 22.9-mc value is included in the range of the desired video frequencies (they extend from 26.4 mc down to 22.4 mc), we insure that the video frequencies are passed by the trap with negligible attenuation while, at the same time, forcibly suppressing the undesired audio carrier I.F.

(c) *Bridged-T Traps*. A fifth trap, which is more complex than any of the foregoing circuits but also more effective, is the Bridged-T trap shown in Fig. 5-14A. L_1 , C_1 , and C_2 are resonated to the frequency of the signal to be rejected. Now, if the value of the resistor R is properly chosen, the attenuation imposed upon a signal to which L_1 , C_1 , and C_2 are resonated will be very great. Ratios of 50 and 60 to 1 are easily attainable using components which possess manufacturing tolerances.

Some understanding of the operation of this trap can be obtained if we take the Bridged-T network shown in Fig. 5-14B and transform it into the equivalent network of Fig. 5-14C. This is called a delta-wye transformation and can be readily accomplished with well-known electrical theorems. If the various components of the Bridged-T network C_1 , L_1 , and C_2 are properly chosen, Z_1 will have a negative value. If we make R equal to Z_1 , then the total impedance between points 1 and 2 will become zero, effectively short-circuiting signals of the frequency to which the Bridged-T network is tuned. For all other frequencies the trap offers little shunting effect and the signals pass through with negligible attenuation.

Tuning the Traps. To adjust the traps to their proper frequencies, the following procedure is employed.

1. Connect an A-M signal generator to the grid of the mixer or converter tube.
2. Connect a vacuum-tube voltmeter across the load resistor of the video second detector.
3. Set the signal generator to the trap frequency.

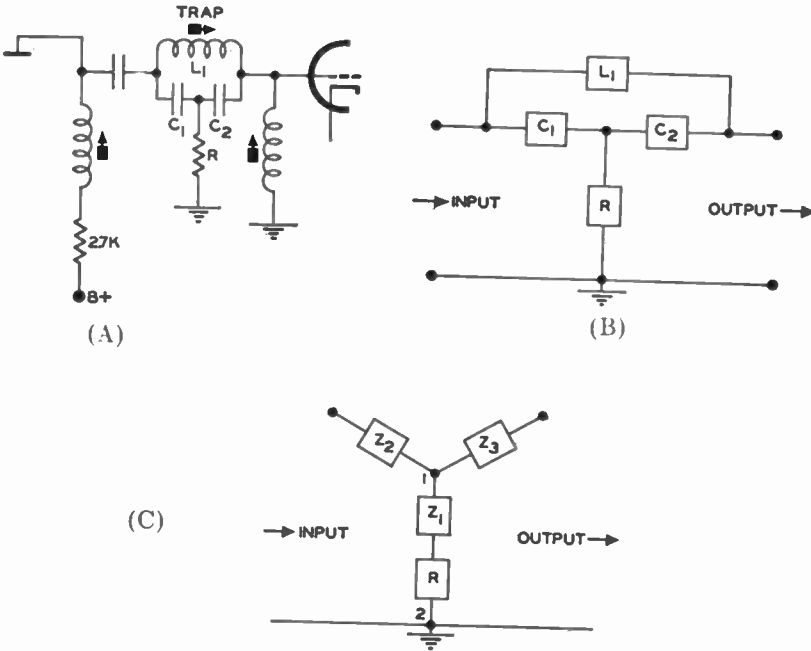


FIG. 5.14. (A) A Bridged-T trap. (B) The same circuit shown in block form. (C) An equivalent electrical network.

4. Using an alignment screwdriver or wrench, change the capacitance or inductance of the proper trap until a minimum voltage is registered at the detector. Consult the manufacturer's instructions for the resonant frequency of each trap circuit.

5. Follow the same procedure for each trap, each time altering the signal generator frequency to correspond.

(Always check the manufacturer's instructions to determine whether a certain bias should be applied to the controlled video I.F. stages during trap alignment. This is frequently done.)

For most television receivers, the traps are adjusted before any video I.F. alignment is begun. During the video alignment process, the traps' adjustment should be rechecked because they are affected to some extent by changes in the associated video I.F. tuning coils and transformers.

The order of importance of the various trap circuits are:

1. Sound signal of the same channel.
2. Sound signal of the next lower adjacent channel.
3. Picture carrier of the next higher adjacent channel.

If, for the sake of economy, the number of trap circuits are limited, No. 3 could be dispensed with first, and then No. 2. Trap No. 1 is never omitted from receivers employing the conventional television system.

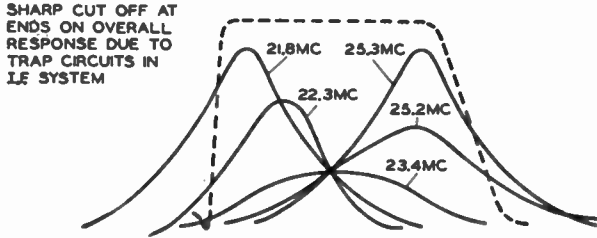


FIG. 5.15. In the circuit of Fig. 5.17, five stagger-tuned coils combine to produce an overall response of 4.0 mc.

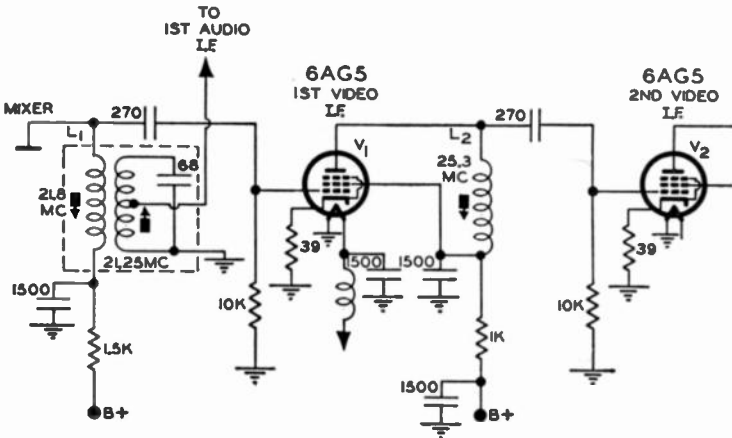


FIG. 5.16. A stagger-tuned video I.F. amplifier.

Stagger-tuned I.F. Systems. The majority of sets today employ single tuned coils between each pair of video I.F. amplifiers. The coils themselves are stagger-tuned in frequency; that is, each one is peaked to a different frequency within the bandpass of the system. Fig. 5-15 illustrates how five separate coils, each tuned to a different frequency, combine to give a 4.0-mc bandpass.*

*It might be mentioned that it is not necessary for each coil to be peaked to a different frequency. Two of the coils may be peaked to the same frequency, but these coils must not be placed in adjacent stages, otherwise oscillations will occur.

A simple stagger-tuned I.F. system is shown in Fig. 5-16. The first tuned coil, located between the mixer and first video I.F. amplifier is peaked (in this circuit) to 21.8 mc. Inductively coupled to this coil is an absorption trap resonated at 21.25 mc. This is the I.F. value of the sound signal and the trap will absorb this frequency voltage from the incoming signal and transfer it to the sound I.F. amplifiers. The remainder of the video signal will be unaffected by the sound trap and continue on through the system. The second coil, between the first and second video I.F. amplifiers, is peaked to another frequency (here 25.3 mc) within the video bandpass. In a similar manner, by adding two or three more I.F. stages, each possessing a coil tuned to a different frequency, we can achieve the desired 4.0-mc video bandpass. Fig. 5-17 illustrates, in semi-block form, a stagger-tuned video I.F. system used in many television receivers. The various peaking frequencies are:

Converter coil	21.8 mc (L_1)
1st I.F. coil	25.3 mc (L_2)
2nd I.F. coil	22.3 mc (L_3)
3rd I.F. coil	25.2 mc (L_4)
4th I.F. coil	23.4 mc (L_5)

Fig. 5-15 indicates the relative gains and selectivities of each coil and the overall shape of the combined response (with the effect of the trap circuits included). Note that each coil has a fairly wide response and this is due chiefly to the low plate and grid resistors which are used in the circuit. In Fig. 5-16, for example, the grid resistor of V_1 is only 10,000 ohms. Since this is effectively shunted across L_1 , it acts to reduce the Q of this coil and broaden its response. Just how broad this response is, is indicated in Fig. 5-15 by the curve marked 21.8 mc.

Four trap circuits are found in this system, of which the first three are absorption type traps. The first trap, coupled to L_1 , is tuned to 21.25 mc, the audio I.F. of this system. The voltage developed across this trap is fed directly into the sound I.F. system. The second trap is coupled to L_2 and is set for 27.25 mc, the frequency of the lower adjacent channel sound carrier. The third trap, coupled to L_3 , is peaked to 19.75 mc, the higher adjacent channel picture carrier frequency. The final trap is a degenerative one, located in the cathode leg of the fourth I.F. amplifier and tuned to 21.25 mc. A contrast control network (not shown) regulates the amount of negative voltage applied to the first three I.F. amplifiers and the R.F. stage.

The number of video I.F. amplifiers employed in any set is determined chiefly by the price of the set. In quality receivers, possessing the full 4.0-mc bandpass, there might be five stagger-tuned coils, using four video I.F. stages. In other sets, where a 3.0-mc spread is common, three video I.F. amplifiers might be employed. Regardless of the number of stages used, the basic operation remains the same.

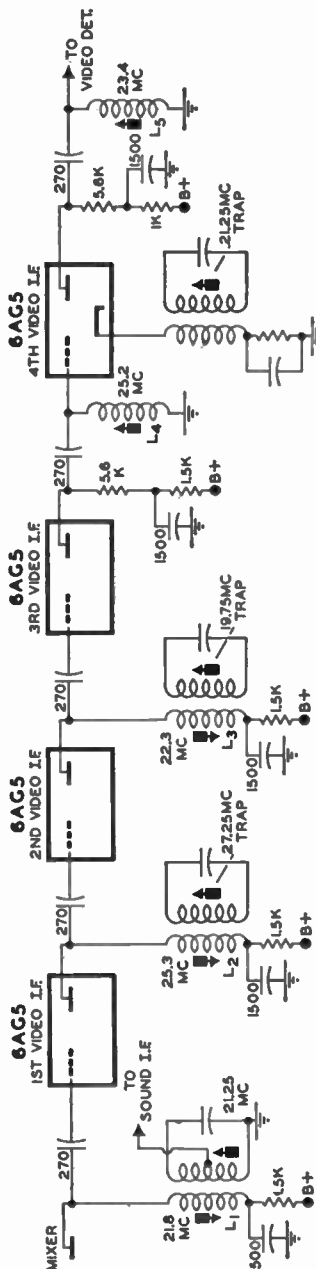


Fig. 5.17. A semi-block diagram of a complete stagger-tuned video I.F. system.

Sometimes a combination of single-coil stagger-tuned and transformer video I.F. amplifiers is employed. Thus, in Fig. 5-18, the signal from the mixer is transformer-coupled to the first video I.F. amplifier while, beyond this stage, single stagger-tuned coils are used throughout.

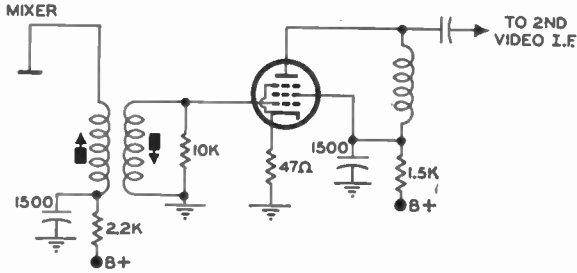


FIG. 5.18. A combination of transformer-coupled and stagger-tuned video I.F. amplifier.

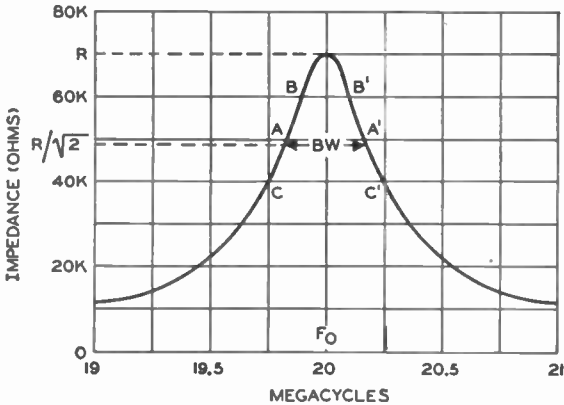


FIG. 5.19. The accepted definition for bandwidth of a tuned system.

Stagger Tuning. Stagger tuning is employed so extensively in video I.F. systems that it might be instructive to analyze these circuits in greater detail. As a first step in this analysis, let us define bandwidth.

A typical resonance curve for a parallel tuning circuit is shown in Fig. 5-19. The response is not uniform but varies from point to point. At the resonant frequency (labeled F_0 in the diagram) the response of the circuit is at its peak — or maximum. From this point, in either direction, the response tapers off until it soon becomes negligible. With a characteristic of this type, what would you say was its bandwidth? Obviously the answer to this question is arbitrary. We could say, for example, that all frequencies between the points B-B' on the curve should be considered as part of the bandwidth. Or, we could choose points C-C' and say that all fre-

quencies which receive an amplification equal to that of C-C' or greater should be considered as within the bandpass of the circuit. Note that this does not prevent other frequencies — those that receive less amplification — from passing through the circuit.

The arbitrary definition generally accepted for bandwidth is illustrated in Fig. 5-19. The bandwidth of a circuit is equal to the numerical difference in cycles between the two frequencies at which the impedance presented by the tuning circuit is equal to 0.707 of the impedance presented at F_0 (i.e., the maximum impedance). Thus, in the response curve shown in Fig. 5-19, the impedance at points A-A' is 0.707 (or $\frac{1}{\sqrt{2}}$) of the impedance offered by the circuit at F_0 . In this particular illustration, Fig. 5-19, the bandwidth is 0.4 mc.

A further note of importance is the fact that, if the gain of the circuit is considered as equal to 1 at F_0 , it is down 3 db at points A-A'. That this is so can be seen from the following.

The definition of decibel is given by: $\text{db} = 20 \log (E_1/E_2)$ where E_1 here would be the voltage at F_0 , and E_2 would be the voltage at points A-A'. For the sake of simplicity, let us assign a value of 1 volt to E_1 . At either point A or point A' the impedance offered to the same signal is $\frac{1}{\sqrt{2}}$ or 0.707 times as great. Hence, the voltage developed at either of these two points will be 0.707 volts. Substituting these values in the formula we have: $\text{db} = 20 \log (1/0.707)$ or $\text{db} = 20 \log 1.414$, or about 3 db.

Points A-A' are also known as the "half power" points because $P = E_1^2/R$, and since $E_2 = \frac{E_1}{\sqrt{2}}$, then $P_{A-A'} = E_1^2/2R$ which is one-half the power developed across R at F_0 , the peak of the curve.

With this concept of bandwidth in mind, let us consider two single-tuned amplifiers, both tuned to the same frequency. If these two amplifiers are in cascade (i.e., follow each other), then the overall bandwidth is not equal to the bandwidth of either circuit, as one might expect, but to 64 per cent of this value. The reason for the shrinkage in bandwidth will be apparent from the following:

The response curve of the first amplifier, shown in Fig. 5-20A, has a maximum value of amplification of 1 at F_0 , its peak, and 0.707 at the ends of the bandpass. Let us say that the mid-frequency is 10 mc, while the end frequencies of the bandpass are 9 and 11 mc respectively. If each of these three frequencies has an amplitude of 1 volt at the input to this tuned stage, then at the output they would possess the following values: at 9 mc — $1 \times 0.707 = 0.707$ volts; at 10 mc — $1 \times 1 = 1$ volt; and at 11 mc — $1 \times 0.707 = 0.707$ volts.

These same three frequencies are now passed through the second tuned circuit. Since this second circuit possesses the same characteristics as its predecessor, here is the result at its output: at 9 mc — $0.707 \times 0.707 = 0.49$ volts; at 10 mc — $1 \times 1 = 1$ volt; and at 11 mc — $0.707 \times 0.707 = 0.49$ volts. After passage through the two amplifiers, 9 and 11 mc are no longer within

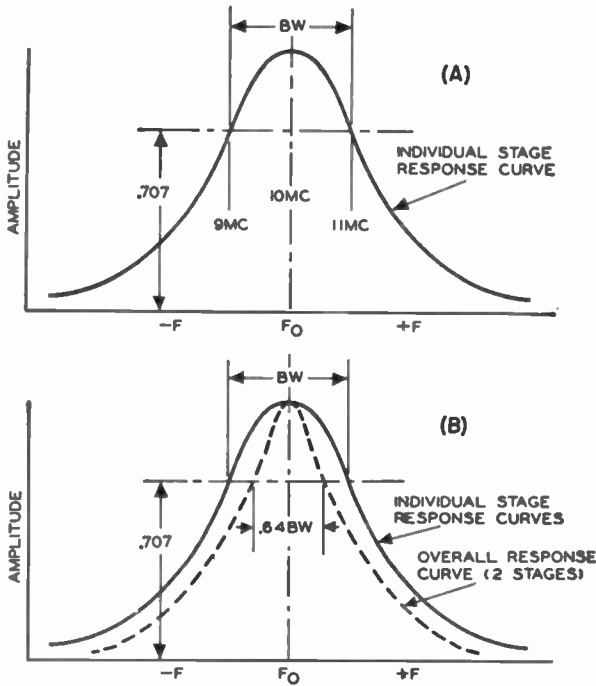


FIG. 5.20. Two tuned circuits, each peaked to the same frequency, produce an overall response in which the bandwidth is less than that of either curve taken separately.

the 0.707 region about the resonant frequency of 10 mc. To find frequencies with voltages equal to at least 0.707 of the 10-mc voltage we must move *closer* to 10 mc. The result, of course, is a narrower bandpass; more accurately, 36 per cent narrower. (See Fig. 5-20B.)

Now let us consider two single-tuned amplifiers, each with the same bandwidth, but with their peaks separated (or staggered) by an amount equal to their bandwidth. (See Fig. 5-21.) The result is a response in which the overall bandwidth (to the 0.707 points) is 1.4 times the bandwidth of a single stage. The overall gain, however, is now only one-half that of the two stages tuned to the same frequency. This is so because at the center frequency of the overall response curve, the individual stage responses

are only 0.707 of their peak response. The product of the stage gains is approximately one-half ($0.707 \times 0.707 \cong 0.5$). This would be stagger tuning.

Now, to progress one step further. We have seen that by stagger-tuning two tuned circuits, we achieve 1.4 times the bandwidth of a single stage but with only one-half of the gain. Suppose, however, we retain stagger tuning, but we decrease the bandwidth of each individual tuned circuit. The overall bandwidth of the stagger-tuned system will still be 1.4 times the bandwidth of the individual stages. However, because we have decreased the individual coil's bandwidth, 1.4 times this new figure will be less than 1.4 times

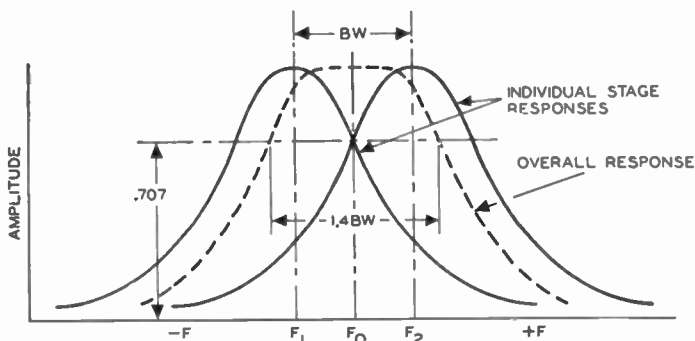


FIG. 5.21. By stagger-tuning two tuned circuits we obtain a wider bandpass. (See text.)

the previous figure when each individual bandwidth was greater. The advantage of this is that we still get a greater bandwidth than if we hadn't stagger-tuned the circuits, and the overall gain remains high.

A simple illustration will make this clearer. Suppose that the bandwidth of each individual stage is decreased to 0.707 of its original value. To do this, we raise the individual circuit Q 's to 1.4 times their previous value which will provide an increase in gain by 1.4 times. Now, when the stages are staggered by an amount equal to this reduced bandwidth, the overall gain is one-half the product of 1.4 times 1.4. The answer is 1. Thus the overall gain is now the same as with the previous amplifier with both circuits tuned to the same frequency.

There are other systems which can give high gain and bandwidth, but the prime advantage of the stagger-tuned amplifier is its ease of alignment.

An important relationship and one which should be remembered by all radio men is the fact that bandwidth of any parallel resonant circuit (or an ordinary resistance-coupled amplifier) is inversely proportional to the amplification of that system. Expressed a little differently, we can say that *bandwidth \times gain = constant*. Thus if we increase the bandwidth of a system by 1.5 times, we decrease its gain by the same amount. For any

A second complex-coupled network is shown in Fig. 5-23. A common inductance, L_2 , couples the signal energy from the mixer plate to the grid of V_1 . The mixer load circuit is L_1 plus L_2 . The input circuit for V_1 is L_2 plus L_3 . L_2 is thus the common element in both arms of this network and functions as the transferring agent. L_1 and L_3 possess variable cores in order

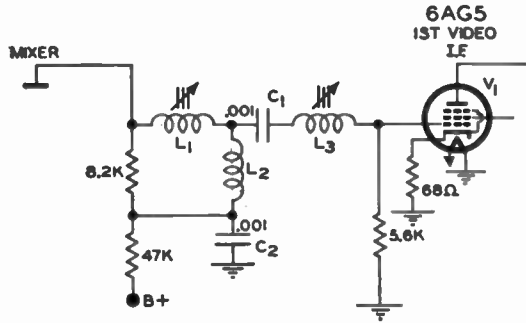


FIG. 5.23. A complex-coupled stage using a common inductance, L_2 , to transfer the signal from the plate of the mixer to the grid of V_1 .

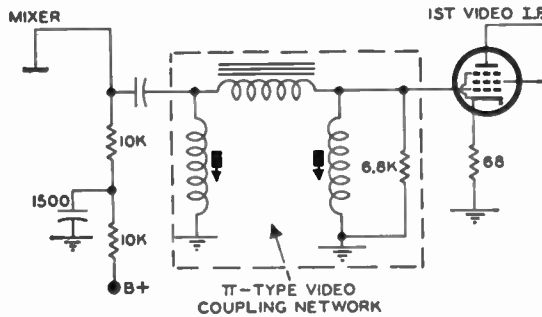


FIG. 5.24. A pi-type video coupling network.

that the bandpass of this network may have the proper width (4.0 mc). C_1 is a d-c blocking capacitor, preventing the mixer B+ plate voltage from reaching the grid of V_1 . C_2 is an ordinary by-pass capacitor. An 8,200-ohm plate resistor broadens the response of the network to insure a 4.0-mc spread.

Some readers will recognize the foregoing circuit as the familiar "T" network employed in some filters. Another complex arrangement would be the "pi" circuit shown in Fig. 5-24. Although an infinite number of such circuits are possible, few, besides the ones shown, are ever employed. Circuits more complex than these are usually too costly and too difficult to adjust to permit their adaptation to mass-produced television receivers.

Bifilar I.F. Coils. A new type of interstage coupling has recently appeared that resembles transformer coupling, yet differs from it in having only one tuning adjustment instead of the usual two. An illustration of a bifilar coil is shown in Fig. 5-25. The two coils comprising the unit are very closely wound, so much so, in fact, that the degree of coupling is close to unity. This means that practically all of the voltage developed across the primary is transferred to the secondary winding.

Tuning of the bifilar coil is achieved with a single iron core inside the coil form. In most instances, the tuning rod extends above the top side of the chassis.

It can be shown by the proper circuit analysis that a 1 to 1 bifilar coil will produce the same selectivity curve and act in all other respects as a single-tuned coil having the same capacitances shunted across it and loaded down by the same resistors. Why, then, may you ask, should we desire to use bifilar coils? The reason, as we shall see in a moment, arises from the economic and operating advantages of these bifilar coils.

Before we consider these advantages, let us compare a video system using single-wound coils and a video system using bifilar coils. In Fig. 5-26 we

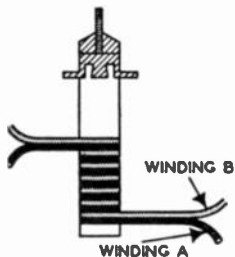


FIG. 5.25. An illustration of a bifilar coil.

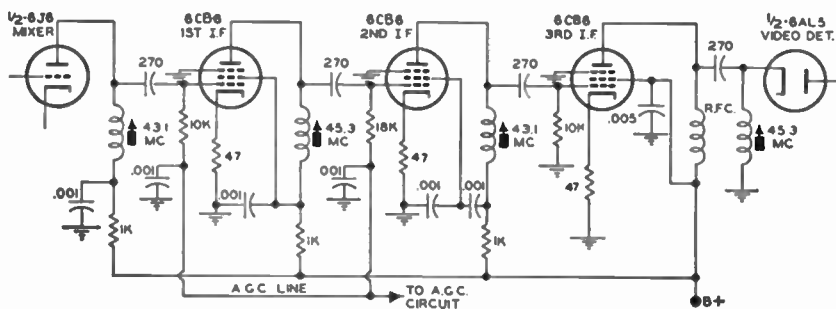


FIG. 5.26. A video I.F. system using single-tuned coils.

have an example of the former; in Fig. 5-27 we have an example of the latter. Both systems were purposely chosen with the same number of stages so that this factor would not enter into the discussion. A comparison of Figs. 5-26 and 5-27 reveals that, by using bifilar coils in place of single-tuned coils, it is possible to eliminate each of the interstage coupling capacitors. In addition, the plate R.F. choke in the last I.F. amplifier may also be dispensed with. Thus a system employing bifilar coils is more economical than a system utilizing single-tuned coils. The operating advantage achieved through

the use of bifilar coils is improved ability of the television receiver to overcome the effect of noise. Here is how this comes about.

When a sufficiently strong noise pulse comes along, it drives the I.F. amplifier grid positive and grid current flows through the grid circuit. Electrons flowing in any of the grid circuits of Fig. 5-26 will charge up the

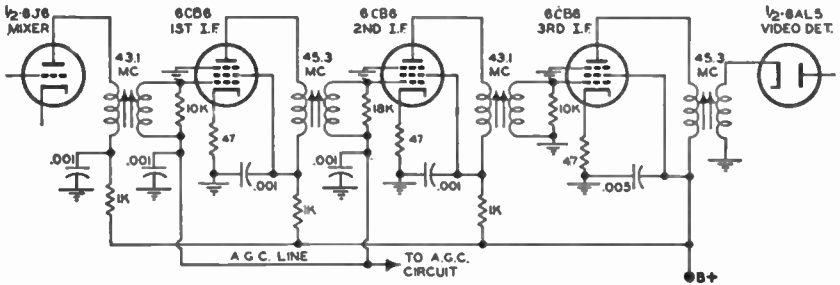


FIG. 5.27. A video I.F. system with bifilar coils.

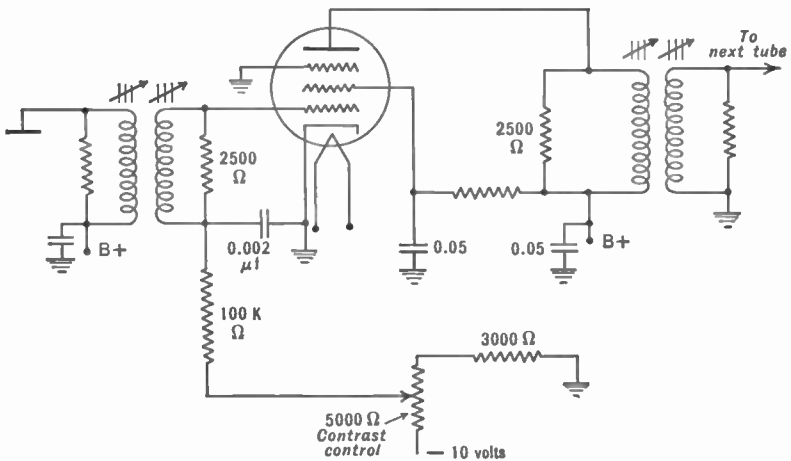


FIG. 5.28. A method for varying manually the gain of a television receiver.

coupling capacitor. When the noise pulse passes, the grid capacitor does not lose its charge immediately, but instead discharges gradually, all the while maintaining a fairly large bias on the tube. Until all this additional negative bias is removed, the gain of the tube remains lower than normal. It may even happen that a severe noise pulse will drive the I.F. tubes to cutoff. When the noise pulse is such as to produce a black dot on the screen under the conditions indicated above, it will be found to be followed by a white tail.

Now let us consider the same circuit using bifilar coils. When a large noise pulse occurs and drives the grid of an I.F. amplifier positive, grid current flows. Now, however, there is no grid coupling capacitor to charge up and so the grid returns to its normal bias just as soon as the noise pulse passes. The noise pulse may still produce a black or white dot on the screen (depending on its polarity at the cathode-ray tube), but black dots are not followed by white tails. Hence, the picture is not disturbed to the same extent.

Alignment of video I.F. systems using bifilar coils is achieved in the same manner as alignment of I.F. stages possessing single-tuned coils. Stagger tuning of the coils plus loading resistors provides the necessary bandwidth. Although trap circuits are not shown in either Fig. 5-27 or 5-28, they are used.

Usually, a system either employs all single-tuned coils or all bifilar coils. Occasionally, however, combinations of the two may be encountered. Thus, one manufacturer employed single-tuned coils in the first, second, and third I.F. stages and a bifilar coil in the grid circuit of the fourth I.F. stage. This last I.F. stage is the most susceptible to noise pulses since the signal has been considerably amplified by this time and by placing a bifilar coil here the grid is prevented from developing a large negative bias on noise pulses.

It may be mentioned again that all of the video I.F. systems discussed may be used for the older 21-26 mc I.F. band or for the newer 41-46 mc range. The form remains essentially the same; only the inductance and capacitance values of the tuning circuits change. Also Intercarrier and conventional type receivers use basically the same video I.F. systems with only minor differences. These consist chiefly in whether the sound signal is to be completely suppressed before it reaches the video second detector (in conventional-type sets) or whether a small amount of sound signal will be permitted to reach the video detector (in Intercarrier sets). Chapter 11 will indicate fully why this difference exists between the two systems and their relative advantages and disadvantages.

Video I.F. Gain Control. In most present-day television receivers it is customary to control the gain of the first two video I.F. amplifiers by means of A.G.C. Figs. 5-26 and 5-27 are typical of this arrangement. In addition, there is also a potentiometer which is located beyond the video second detector where it controls the amount of signal reaching the picture tube. This control is called the contrast control and it is to a television set what a volume control is to a sound receiver.

When a television receiver does not contain A.G.C., then it is customary to have the contrast control regulate the bias on the video I.F. stages directly. One such arrangement is shown in Fig. 5-28. A d-c biasing voltage is obtained from the negative side of the power supply, and the position of the

center arm of the potentiometer determines how much of this negative voltage reaches the grid of the controlled tube.

QUESTIONS

1. Draw the standard response curve for the I.F. system of a television receiver. Indicate the position of the video carrier.
2. What basic factors govern the design of an I.F. system?
3. How is the I.F. chosen for a system?
4. What is a spurious response?
5. List and explain two types of spurious responses.
6. Illustrate two methods for separating audio and video I.F. voltages from each other in conventional type sets.
7. When are video and audio I.F. signals separated from each other? Explain for both conventional and Intercarrier systems.
8. Why are trap circuits used in video I.F. amplifiers?
9. In a conventional receiver employing 25.75 mc for the video carrier I.F. and 21.25 mc for the audio carrier I.F., what trap frequencies should be employed? Explain how your answers were obtained.
10. When is the audio signal of the adjacent television channel not important? Why?
11. Draw three types of trap circuits commonly found in television circuits.
12. Given that the video carrier I.F. is 26.4 mc for a certain receiver. List the oscillator frequencies for each of the thirteen channels.
13. Draw the circuits of two different types of I.F. amplifiers.
14. Explain why condenser tuning is seldom found in television I.F. tuned circuits.
15. A video I.F. signal extends from 28.3 mc to 24.3 mc. What effect would attenuation of the frequencies around 28.0 mc have on the image? Around 24.3 mc?
16. Explain the reason for the recent RTMA recommendation that the video I.F. carrier be raised to 45.75 mc and the sound I.F. carrier value be raised to 41.25 mc.
17. To which type of receiving system are traps in the video I.F. stages most important? Explain.
18. What is the most widely used method of controlling the gain of the I.F. system? Illustrate how the controlling voltage is applied to the tubes.
19. How would you connect a contrast control into the video I.F. system for manual gain control?
20. Wherein does a bifilar coil differ from conventional transformer coupling?

Chapter 6

DIODE DETECTORS AND A.G.C. CIRCUITS

Detection of the Video Signal. In accordance with the general line-up that is peculiar to superheterodynes, the second detector follows the last I.F. amplifier. Detection in television receivers is carried out in much the same manner as in any ordinary sound broadcast receiver. The single diode connected as shown in Fig. 6-1 is typical. The demodulated video signal with its blanking and synchronizing peaks is developed across R_L . The form of the signal, when it enters the second detector, is shown to the left of the figure. The rectified resultant is illustrated at the right of the figure.

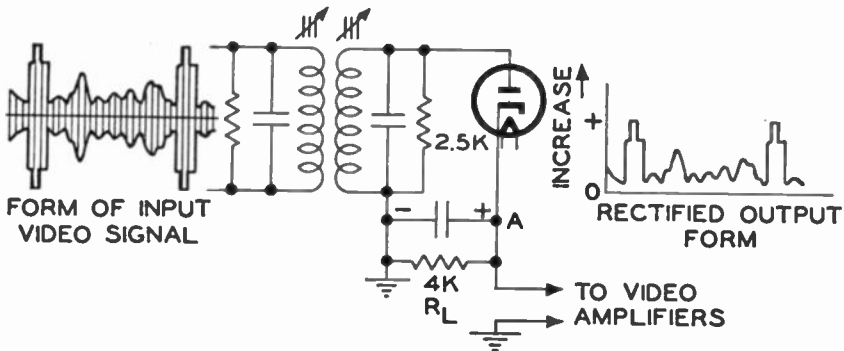


Fig. 6.1. A diode detector for a television receiver.

As is true of diode operation, plate current flows only when the plate is positive with respect to the cathode. The effect of this action is to eliminate the negative portion of the incoming signal. This is of little interest here, since the positive and negative sections of the modulated video signal are exact duplicates of each other and either one may be used. Instead of a half-wave rectifier it is also possible to use the full-wave arrangement shown in Fig. 6-2 with a 6AL5 double diode.

Positive and Negative Picture Phases. At this point it is necessary to pause and consider the effect of the relative polarity of the voltage drop across the load resistor, R_L . It will be remembered from Chapter 1 that, for American television systems, negative picture transmission is standard. This means that the brightest elements cause the least amount of current to flow while maximum current is obtained when the blacker than black re-

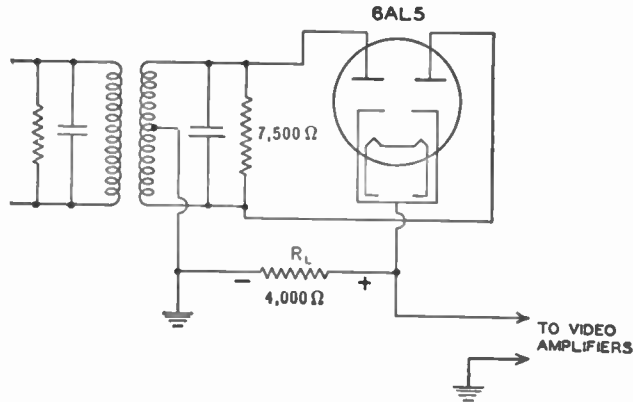


FIG. 6.2. A full-wave detector.

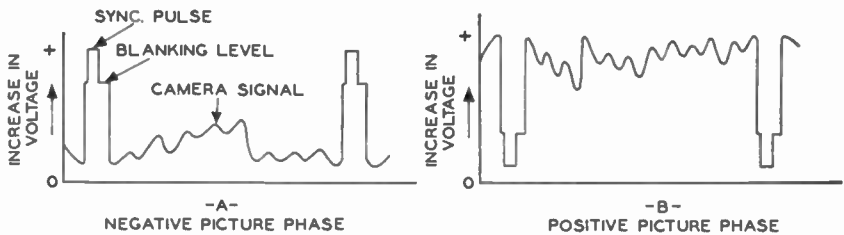


FIG. 6.3. Rectified video signals may be obtained from the output of the detector in either one of the two forms shown, depending upon how the detector is connected.

gion of the synchronizing signal is reached. This method of transmission was adopted because it was felt that better overall reception would be obtained.

The signal in the negative picture phase form, as shown in Fig. 6-3A, could not be applied directly to the grid of the picture tube, however, but first has to be reversed to the form of Fig. 6-3B. That this is essential is easily seen, for the blanking and synchronizing signals, when applied to the control grid of a picture tube, must bias it to cut-off. The objective can be attained only if the signal has the form given in Fig. 6-3B. The radio engineer calls this latter form of the television signal the positive picture phase. It is interesting to note that, if the negative phase of the signal

were applied to the control grid of the picture tube, all the picture values would be reversed and the observed scene would be similar to a photographic negative.*

In sound receivers, no attention is given to the relative phase of the audio signal because our ears are insensitive to all but gross phase differences. Television, on the other hand, deals with visual images, and reversal of phase produces noticeable effects. Possible ways of altering the phase of the video signal are discussed in the following paragraphs.

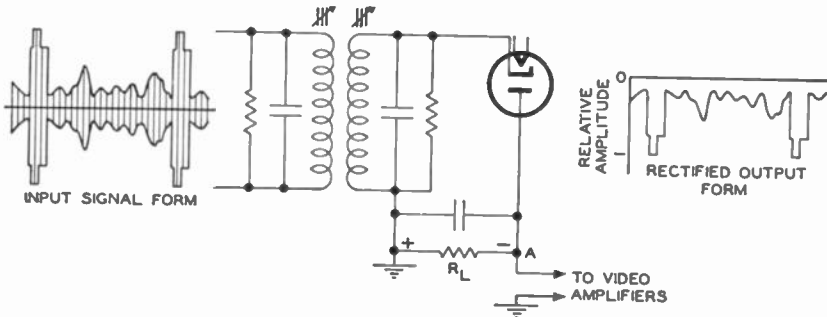


FIG. 6.4. A diode detector connected to give a positive picture phase output signal. Note that, with an inverted diode detector, only the negative half of the input signal is rectified. In Fig. 5.1, the opposite is true.

Turning to the half-wave detector circuit of Fig. 6-1, let us investigate the voltage developed across R_L . The incoming signal has the same form as at the antenna, with the synchronizing pulses giving rise to the greatest voltages. At the diode rectifier, these synchronizing signals cause the plate to become the most positive, resulting in a greater voltage drop across R_L and having the polarity as shown. On the other hand, those portions of the video signal representing the bright segments of the image will have the least positive voltage at the diode plate, with a smaller resultant voltage drop at R_L . Thus, with this circuit hook-up, point A of resistor R_L will still give rise to a large positive voltage for the synchronizing signals, which means that the signal is still in the negative picture phase. The signal is unsuitable for direct application to the grid of the viewing tube.

The direction of the current flow through R_L may be altered to give the opposite polarity quite easily. Merely reverse the connections between the diode tube and the input transformer, as in Fig. 6-4. Rectification now eliminates the positive half of the modulated carrier and leaves only the negative half. Since both contain the same information, nothing is lost. Point A

* Another widely used method of distinguishing between signal polarities is by employing the designation of sync pulse positive (Fig. 6-3A) or sync pulse negative (Fig. 6-3B).

becomes more strongly negative for the blanking and synchronizing portion of the video signal while the bright elements cause A to become less negative. When the signal is applied in this form between the grid and cathode of the image tube, the largest current will flow for the bright sections of the image and a bright spot will appear on the fluorescent screen. For the blanking and synchronizing parts of the signal, the voltage at the grid (from point A) will be highly negative and the electron beam will be cut off, as it should be.

The strength of the signal that is developed at the diode load resistor is not strong enough to use directly at the picture tube. Hence, further am-

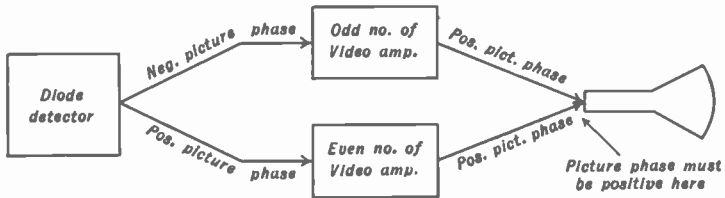


FIG. 6.5. An illustration of why the number of video amplifiers after the detector is dependent upon the polarity of the signal obtained from the detector.

plification is necessary. The following video amplifiers, which are generally of the resistance-coupled type, are capable of reversing by 180 degrees the polarity of any signal sent through them. (This will be proved in Chapter 7.) Thus, if the video signal had a positive picture phase at the diode load resistor, it would have a negative picture phase at the output of the first video amplifier. With another stage of amplification, the picture would be brought back to the positive phase again. As a general rule, then, an even number of video amplifiers is required if the picture phase across R_L in the detector is positive. For a negative picture phase at R_L , an odd number of video amplifiers is needed, this time for a positive picture to appear at the grid of the image tube. These conditions are given in block form in Fig. 6-5.

The circuit of the video detector and the video amplifier used in a General Electric television receiver (Fig. 6-6) appears, at first glance, to violate the foregoing rules. Examination of the video signal developed across R_1 reveals it to be positively phased. This would require 0, 2, 4, or some other even number of video amplifiers. Actually, only one is present. The mystery is resolved when we note that the output of the video amplifier is fed not to the control grid of the cathode-ray tube but to its cathode. The foregoing rules were drawn up with the tacit understanding that all incoming signals are applied to the control grid of the picture tube. This is the normal point of signal application. Signals applied to the cathode of a

tube should differ by 180 degrees from the same signals applied to the control grid in order to produce similar results.

Detector Filtering and Peaking. The frequencies present in the detector circuit include the intermediate-frequency values and the actual video signals themselves, 0-4 mc. The latter voltages are to be passed on to the video amplifiers and be strengthened to the point where they are

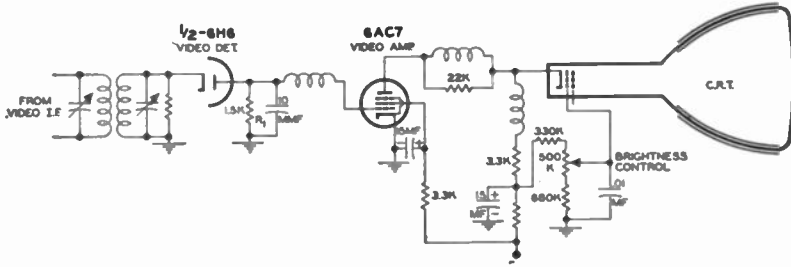


FIG. 6.6. A video detector and video amplifier feeding the signal to the cathode of the image tube.

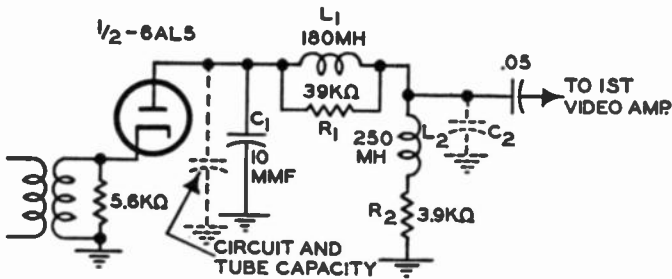


FIG. 6.7. A video detector circuit with a low-pass filter and load resistor.

able to modulate the electron current in the cathode-ray tube to produce an image on the screen. At the detector output, the intermediate frequencies must be properly shunted around the load resistor to prevent their reaching the following video amplifiers. In the receivers currently being produced, the problem of filtering the I.F. voltages has been made comparatively simple through the use of fairly high I.F. values. The rectified video signal has a maximum frequency of 4 mc. In early television receivers, the I.F. values ranged from 8.75-12.75 mc, and considerable filtering was required because of the low order of separation between the desired frequencies (0-4 mc) and those which were to be by-passed (8.75-12.75 mc). However, by increasing the separation between the two, we have simplified the problem considerably. Current recommended values for the video I.F. are between 40 and 46 mc. Adequate filtering can be obtained through the arrangement shown in Fig. 6-7. The rectified current passes through the

low-pass filter composed of C_1 , L_1 , R_1 , L_2 , R_2 , and C_2 . C_1 is a small fixed capacitor of 10 $\mu\mu\text{f}$, but actually there exists additional capacitance across this point produced by the tube and the wiring. At the other end of the filter, C_2 is shown in dotted form because no such component is inserted. However, the sum of the stray wiring capacitance plus the input capacitance of the following video amplifier produces the equivalent of an actual capacitor of 10–15 $\mu\mu\text{f}$. The two coils, L_1 and L_2 , while forming part of the low-pass filter, at the same time maintain a good frequency response to 4 mc, thereby counteracting any tendency of the circuit to attenuate these higher video frequencies. More will be noted on this point in the succeeding chapter on video amplifiers. The 39,000-ohm resistor shunted across L_1 is to prevent the response of the coil from rising abruptly at the higher video frequencies because of a natural resonant circuit formed by the coil and its inherent capacitance. The detector load resistor is R_2 , 3,900 ohms.

In place of a diode tube, many television receiver manufacturers have substituted one of the new germanium crystal detectors, 1N34, 1N54, 1N55, 1N56, 1N64, and 1N65, among others. These new crystals, which were developed for radar, are well suited for operation up to 100 mc. Germanium crystals, which are rectifiers, should not be confused with quartz crystals. These are known as piezoelectric crystals and have an entirely different function. Piezoelectric crystals are used for oscillators and are not interchangeable with rectifier crystals. The old cat's-whisker crystal detectors which many radiomen still recall were much too critical in adjustment to be used for any such purpose. Furthermore, their large size produced a high internal capacitance which was detrimental to high-frequency operation. In the new germanium crystals the total capacitance from cathode to anode is only 3 $\mu\mu\text{f}$, which compares favorably with any of the miniature diodes.

Briefly, the theory of operation of a rectifier crystal is as follows: When two metals are brought into contact, electrons can flow as readily in one direction as in another. However, when a metal is brought into contact with a certain class of materials known as semi-conductors, electrons flow more easily in one direction than in the other. When we apply an alternating wave to a circuit containing a crystal, more current will flow in one direction, with the result that an average current will be produced. Hence, rectification has occurred. In this respect the crystal is less efficient than a diode, since current is permitted to flow in only one direction through a tube. In the crystal, current flows in both directions, although considerably stronger in one direction than the other. The smaller current, which is not desired, offsets some of the rectified voltage developed across the load resistor, thereby lowering the ratio of voltage output to voltage input. With this ratio decrease, the detector efficiency drops.

A typical circuit using a germanium crystal is shown in Fig. 6-8.

A.V.C. and A.G.C. Automatic volume control (actually this should be called automatic gain control) in a broadcast receiver serves to keep the output constant while wide variations occur in the input signal. Once the manual volume control has selected the output level that is desired, the A.V.C. system tends to keep it there. In addition, when tuning to other

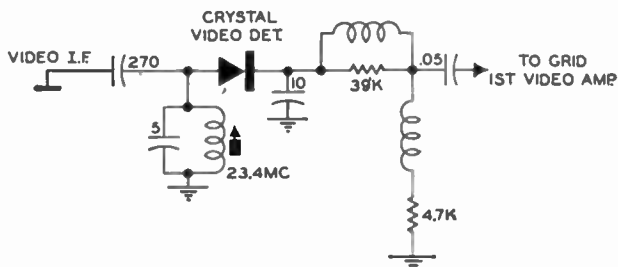


Fig. 6.8. A germanium crystal as a video detector.

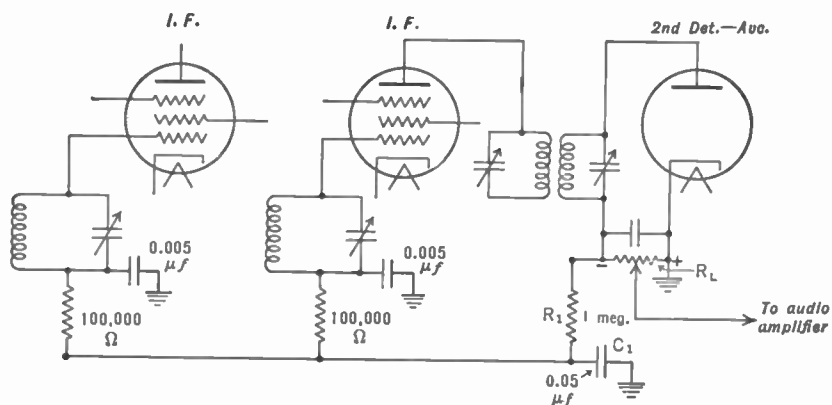


Fig. 6.9. A conventional A.V.C. circuit found in present sound receivers.

stations, no adjustments are necessary to prevent blasting. For television receivers, automatic gain control is advantageous in keeping the picture intensity fixed at one level while the actual video signal at the input of the set may be varying. The eye is far more critical of changes than the ear, and anything that would minimize unwanted variations in image intensity is very desirable. A.G.C. would be advantageous when switching from one station to another, for again input signal strengths may differ. Finally, more stable synchronizing operation is obtained if the signal fed to the synchronizing circuits is constant in amplitude.

In a broadcast receiver, the A.V.C. voltage is obtained at the second detector. The necessary audio signal is developed across the load resistor. A circuit frequently used is illustrated in Fig. 6-9, with the polarity of the

A.V.C. voltage and the filter circuit included. It will be remembered that the object of the A.V.C. is not to feed the instantaneous audio variations but rather an average voltage that depends upon the carrier level back to the R.F. and I.F. stages. If the audio variations were sent back to the preceding tubes, all musical passages, for example, would arrive at the speaker with the same intensity, in itself a form of distortion.

The desired A.V.C. voltage is obtained from the average voltage developed across the load resistor by the rectified carrier signal. The incoming

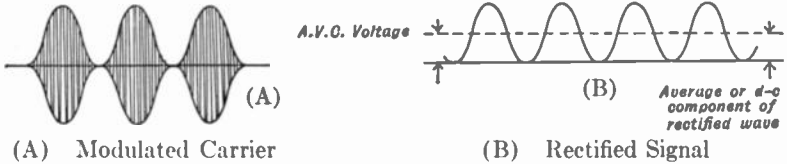


FIG. 6.10. The A.V.C. voltage developed across R_L represents the average value of the incoming wave.

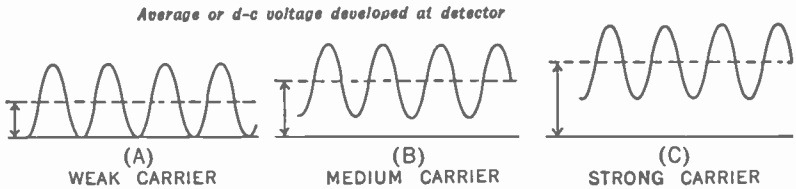


FIG. 6.11. These diagrams illustrate how the A.V.C. voltage is dependent upon the strength of the incoming signal, and not its modulation.

signal both before and after rectification by the diode detector is shown in Fig. 6-10. The detector output consists of a pulsating d-c voltage that contains a varying audio voltage, and an average d-c voltage dependent upon the carrier strength. The audio signal, since it is varying as much above the average value as below, contributes nothing to the d-c voltage. The only way that the average d-c voltage can be changed is to alter the strength of the carrier. This is demonstrated in Fig. 6-11 where different levels of carrier signals are shown, all having the same audio component. The d-c voltage is then fed to the various controlled I.F. and R.F. stages. The purpose of the filter capacitor C_1 and resistor R_1 of Fig. 6-9 is to prevent (or at least minimize) any audio voltage from reaching the controlled tubes.

The objectives in television receivers, through the use of automatic gain control, are similar to those of broadcast sound sets. The means of obtaining the necessary voltage, though, is slightly different, due to the difference in the make-up of the video signals. A study of the television-modulated signal in Fig. 6-12 reveals that, so far as A.G.C. is concerned, the rapidly varying camera signal is of no use to us. We desire some point

which will be indicative of the strength of the carrier and which does not change with anything but the carrier.

With the present system of transmission, the carrier is always brought to the same level when the synchronizing pulses are inserted. Thus, as long as the signal being received is constant in strength, the level of the synchronizing pulses will always reach the same value. If something should affect the carrier level, these pulses would likewise change. With the change, the gain of the set would require adjustment to maintain the previous level at the detector. Hence, the strength of the synchronizing pulses will serve nicely as a reference level for the A.G.C. system. It should be noted that the

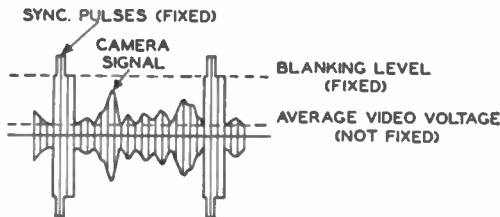


FIG. 6.12. An amplitude-modulated television signal. Only the fixed voltage levels are suitable for A.G.C. control since only these voltages vary directly with signal strength.

level of the blanking pulses (immediately below the top of the synchronizing pulse) is likewise fixed and may also be used.

Analysis of some of the circuits found in commercial receivers readily indicates how automatic gain control is obtained.

A simple yet effective method of developing a suitable A.G.C. voltage is employed in some television receivers. (See Fig. 6-13A.) The A.G.C. diode is one section of a 6AL5 and receives the incoming signal from the video I.F. system through a $0.5\text{-}\mu\text{f}$ capacitor, C_1 . The load for the A.G.C. tube is R_1 , a high-valued resistor of 1 megohm. To understand better the operation of this circuit, consider the equivalent diagram shown in Fig. 6-13B. (The cathode resistor and capacitor are omitted from the equivalent diagram because they do not affect the A.G.C. voltage development.) The A.G.C. tube will not conduct until its plate is driven positive with respect to its cathode. When this occurs, electrons flow from the cathode to the plate of the diode and into C_1 where the negative charge is stored. Very few electrons attempt to go through R_1 because of its high value. On account of the low impedance offered by the tube when it is conducting, C_1 charges up to the peak of the applied voltage, which is the value of the synchronizing pulses.

During the negative excursion of the incoming signal, the plate of the diode is driven negative with respect to its cathode and no conduction

through the tube occurs. However, if we examine Fig. 6-13B, we see that a complete circuit exists with C_1 , R_1 , and the input coil all in series. Since a voltage exists across C_1 and a complete path is available, current will flow, with the upper end of R_1 becoming negative with respect to ground. Because of the long time constant of R_1 and C_1 , the charge accumulated across

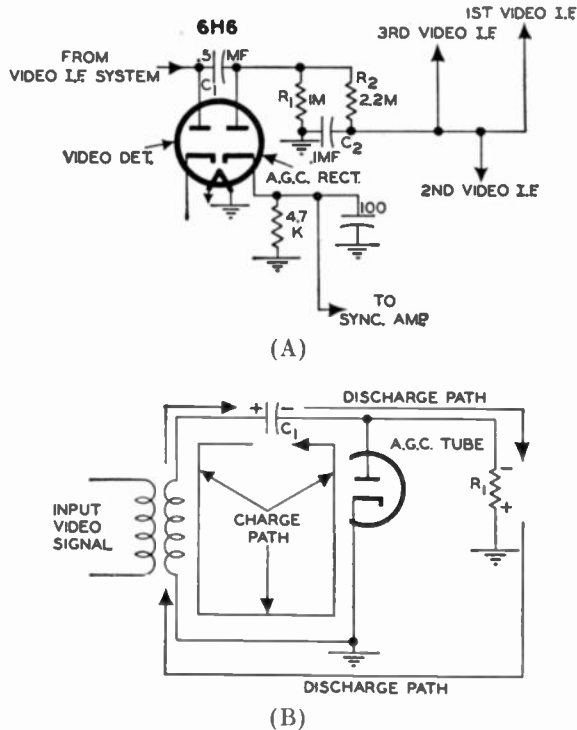


FIG. 6.13. (A) A simple A.G.C. network. (B) The equivalent circuit, showing the charge and discharge paths of C_1 .

C_1 will discharge slowly through R_1 , so slowly, in fact, that only a small percentage of the voltage across C_1 will be lost during the interval when the tube is not conducting.

When the incoming signal becomes positive again, the tube does not immediately conduct because the applied signal voltage must first overcome the negative voltage existing across C_1 . Since C_1 has lost but little of its voltage, tube conduction will occur only at the very peaks of the positive cycle. These peaks, of course, are the synchronizing pulses. Thus, the voltage across C_1 is governed entirely by the sync pulses, which is what we desire. The negative voltage across R_1 is filtered by R_2 and C_2 to remove the 15,750-cycle ripple of the horizontal sync pulse and then fed to the video

I.F. amplifiers as the control voltage. It is interesting to note that the same diode also supplies the sync pulses to the horizontal and vertical sync systems. Since current flows through this diode only at the sync pulses, voltage pips will appear across the 4,700-ohm resistor in the cathode leg of the tube at these times. These pips are tapped off and applied to the sync-separating circuits.

A novel yet simple arrangement for the development of A.G.C. voltage is shown in Fig. 6-14. The entire video signal (containing signal voltages and sync pulses) is tapped off from the output circuit of the first video-

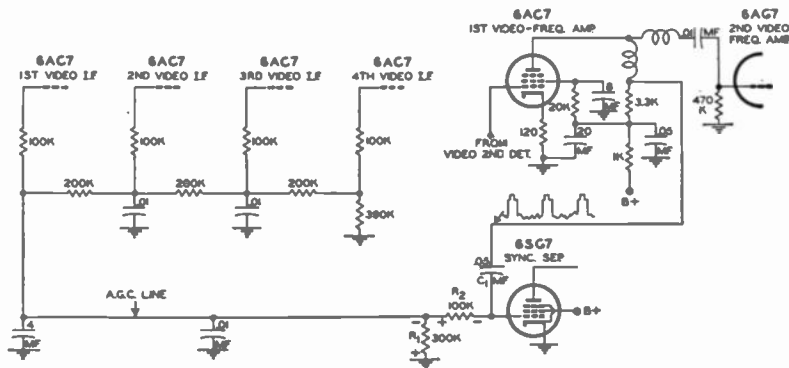


Fig. 6.14. An A.G.C. network employed in some television receivers.

frequency amplifier (6AC7) and applied to a 6SG7 sync-separator tube. The A.G.C. voltage is developed at the grid of the 6SG7 tube as a result of the grid-leak biasing arrangement employed here. The 6SG7 tube has no fixed bias of its own. Hence, when a signal is applied through C_1 , grid current flows for the positive portions of the signal, the amount dependent upon the strength of the signal voltage. Because of the high value of R_1 and R_2 , the grid electrons flow into C_1 , charging this to the peak value of the applied signal. This peak value, of course, is set by the sync pulses. The voltage developed across C_1 then discharges slowly through R_1 and R_2 , developing a bias across these two resistors which is negative with respect to ground. Because of the long time constant of the grid network, the charge on C_1 leaks off slowly, and only the sync pulses of the signal are able to overcome its negative charge and cause current to flow in the 6SG7. Thus, the voltage developed across R_1 and R_2 is governed by the sync pulses. The voltage across R_1 is applied, through a suitable filter network, to the last four video I.F. amplifiers as an automatic bias. The 6SG7 tube is known as a sync separator because only the sync pulses appear in its output. All the rest of the video signal is without effect because of the high negative grid-leak bias.

voltage which is finally employed as the A.G.C. voltage to control the gain of the R.F., input I.F., and first video I.F. amplifiers.

When the signal level increases, the A.G.C. rectifier develops more voltage across C_1 and R_1 . This means more positive voltage for the A.G.C. amplifier (7B6) and a greater output from this tube because of the increased gain. The horizontal deflection voltage fed into the tube remains constant at all times and a more positive bias for the tube will produce a greater output. This means more rectified A.G.C. voltage fed to the controlled stages. However, since the final A.G.C. voltage is negative, the gain of the controlled stages decreases, counteracting the increased video signal and reducing the output to the normal level.

When the signal level decreases, less positive voltage appears across C_1 and R_1 and the A.G.C. amplifier bias becomes correspondingly more negative. Less gain in this tube provides less voltage at its output, and the negative A.G.C. voltage decreases. As a result, the gain of the controlled stages rises, and the signal level is again brought back to the normal level.

Whereas in previous A.G.C. systems, the output of C_1 and R_1 was used as the controlling A.G.C. voltage, here it is used to control the bias of the A.G.C. amplifier. The final product of both systems is the same, but this arrangement is more sensitive to carrier changes because of the addition of the amplifier. A divider network at the output of the diode section of the 7B6 feeds less A.G.C. voltage to the R.F. amplifier than to the I.F. stages. The potentiometer in the A.G.C. system permits the system to be adjusted for desired sensitivity.

Keyed A.G.C. Systems. Keyed A.G.C. systems represent an entirely new approach to the problem of adequately controlling the gain of a receiver. Although of fairly recent origin, they are already widely used, and therefore it behooves the serviceman to be familiar with their operation and servicing.

If we examine the A.G.C. filter network in any of the circuits discussed thus far, it will be seen that fairly high-valued resistors and capacitors are used. This means that the capacitors charge up and discharge slowly. High-valued resistors and capacitors are required because not only must they smooth or filter out a 15,750-cycle ripple arising from the horizontal pulses which actuates the circuit, but they must also smooth out a 60-cycle ripple produced by the vertical pulses. If the 60-cycle fluctuation is not removed, the bias on all controlled tubes will rise during the vertical sync interval and act to depress these pulses.

Thus, for slow changes in signal strength, the conventional A.G.C. bias arrangement is satisfactory. But what happens when a very fast change in signal level occurs? If such changes do not occur frequently, they will have very little noticeable effect on the A.G.C. bias. But if the changes

occur rapidly and continuously (or frequently), they will affect the bias and, through this, the picture. Thus, for example, when an airplane passes overhead, the picture intensity will rise and fall (flutter). Again, if the noise level is very high, the conventional A.G.C. system will react to large noise pulses by developing more bias than it would if it responded solely to the signal. The result will be less amplification for the signal; if the signal itself is quite weak, this decreased amplification may cause it to be lost altogether.

Both of these disturbances will adversely affect the television image. In fringe areas, the weak signal and the surrounding noise are the most important considerations; in areas near airports, airplane flutter is important. Before we describe how a keyed A.G.C. system overcomes both these annoyances, let us briefly determine the reason for airplane flutter.

Picture flutter occurs whenever an airplane passes overhead (or nearby). The picture intensity rises and falls, becoming light and dark in turn at a fairly rapid rate. This effect might last for 15 to 30 seconds, depending upon how long it takes the airplane to pass.

The intensity pulsation is caused by the airplane acting as a reflector. Some of the television signals striking the metallic surface of the airplane bounce off and reach the television antenna. If these reflected television signals arrive in phase with the normal signal that the set antenna receives, they will add to the desired signal and strengthen it. If the reflected signal arrives out-of-phase with the normal signal, the strength of the normal signal will be reduced. This rapid increase and decrease in strength of the received signal cannot be counteracted by the usual slow-acting A.G.C. filter. As a result, the picture on the screen will vary in intensity, producing the aforementioned flutter.

A keyed A.G.C. system is able to overcome this flutter because of two facts: First, the A.G.C. system is receptive to incoming signals only at certain specific times. Second, the resistance and capacitances comprising the A.G.C. filter are lower in value than the corresponding components in conventional A.G.C. systems. This enables a keyed A.G.C. network to react instantly to a fairly rapid signal fluctuation, such as that produced by airplanes, and to change the A.G.C. bias fast enough to counteract this signal change.

A simplified illustration of a keyed A.G.C. system is shown in Fig. 6-16. A 6AU6 pentode is so connected that a portion of the detected video signal is applied to its control grid. The signal is in the negative picture phase, which means that the sync pulses are most positive. The plate of the 6AU6 is connected to a winding on the horizontal output transformer and receives from this transformer a positive pulse of voltage at the end of each horizontal line.

When the 6AU6 conducts, current flows from the tube cathode to plate, then through R_1 and R_2 to ground, and finally back to the cathode of the 6AU6. C_1 is charged to the voltage developed across R_2 . The values chosen for the R_2, C_1 combination are designed to remove the 15,750-cycle ripple of the plate current pulses. No provision need be made to filter out a 60-cycle ripple since none exists in this A.G.C. system. It is due to the use of small $R-C$ values that largely accounts for the ability of a keyed A.G.C. system to act fast in overcoming airplane flutter.

Before we progress further, there is probably one question which will occur to many readers. That is: Why do the other A.G.C. networks develop a 60-cycle ripple, yet the keyed A.G.C. networks do not? To understand this, refer to Fig. 1-19 (Chapter 1) where horizontal sync pulses and vertical sync pulses are shown. Note that the duration of a horizontal sync pulse is much shorter than the duration of the vertical sync pulse. In the conventional A.G.C. system, this inequality in pulse duration results in a larger charge being developed across the A.G.C. capacitors by the vertical sync pulses. Hence, every 60th of a second, more current will flow through the A.G.C. tube due to the greater width of the vertical sync pulses. As a result, the A.G.C. biasing voltage will have a 60-cycle ripple in addition to the 15,750-cycle variation caused by the horizontal pulses themselves.

In the keyed A.G.C. system, the 6AU6 is "fired" by a combination of a positive plate pulse and a positive grid pulse. The plate pulse, however, is constant in duration since it is obtained from the horizontal output transformer and the pulses in this circuit do not change. Hence, whether the grid pulse is a horizontal pulse or a vertical pulse, the 6AU6 conducts for the same length of time. It is because of this behavior that only a 15,750-cycle ripple is present and need be filtered in keyed A.G.C. systems.

In applying the sync pulses to the grid of the 6AU6 tube, care must be taken to see that they are all aligned to the same level. This is necessary because the amount of current flowing through the A.G.C. tube is determined in large measure by the amplitude of the sync pulses. As these vary, so will the current and, with this, the A.G.C. bias developed across R_2, C_1 and fed to the controlled tubes. Hence, it is important that the video signal contain its d-c component, for when the d-c component is removed, the sync pulses of the same signal are no longer aligned. To feed the latter type of signal to the A.G.C. tube would result in a variation in bias with picture background and would not provide true reproduction of the original signal.

Commercial application of the keyed A.G.C. system is shown by the circuit of Fig. 6-17. The video signal developed at the second detector has a complete d-c path to the grid of the video amplifier tube and therefore the d-c component is present in the video signal. Hence this requirement is satisfied.

ohm resistors and the 100,000-ohm resistor in its plate circuit. The polarity of the voltage developed at points A and B is thus negative with respect to ground and this is the A.G.C. regulating bias sent to the grids of the controlled tubes. Note that the R.F. amplifier receives a slightly larger negative bias than the I.F. stages.

Under normal conditions, the A.G.C. voltage measured at the control grid of any of the video I.F. amplifiers in this circuit will be approximately -4.5 volts. This voltage will vary slightly with signal input and contrast control setting. The latter control adjusts the screen grid voltage of the 6AC7 between the limits of 65 to 140 volts. At 65 volts the video signal is cut off completely; at 140 volts it receives its maximum amplification.

The foregoing has indicated the advantage of a keyed A.G.C. system and its manner of operation. There are some disadvantages to this system, too. First, when the horizontal sweep is not in step with the incoming signal, the pulses at the A.G.C. tube will be also out of step and the A.G.C. bias will vary rapidly. Hence it is most important that the horizontal sync system be stable.

Another factor which will influence the operation of this A.G.C. network is the timing of the horizontal sweep pulse with respect to the incoming sync pulses of the signal. The interval during which both must be active is very limited, being at most 10 microseconds, and this tends to make the operation of a keyed A.G.C. system more critical than an ordinary A.G.C. system. However, this criticalness cannot be reduced since it is inherent in the operation of this circuit. Actually to make the system less critical would mean that the 6AU6 would have to conduct for a longer period of time. This, in turn, would make the set more susceptible to noise pulses. Thus, a choice must be made and the operation of the circuit, as outlined above, provides very good results when it is functioning properly.

Zenith television sets employ a keyed A.G.C. system which differs slightly from the foregoing circuit. The Zenith circuit is shown in Fig. 6-18. The video second detector is a germanium crystal, 1N64. The detected video signal is developed across R_{a1} (4,700 ohms) and applied via L_{55} to the grid of V8A. Although the manufacturer labels this tube as a noise limiter and inverter, it is essentially an amplifier, with the signal which is developed at the plate being fed to a second video amplifier.

The cathode resistor for V8A ($R_{33,820}$ ohms) is also the cathode resistor for V12A, which is the A.G.C. amplifier. Hence, the full video signal is applied to the cathode of the 6SN7 A.G.C. amplifier tube. The polarity of the video signal here is such that the sync pulses are negative.

(Note: To tube V12A, V8A functions as a cathode follower because V12A receives the video signal from the cathode of V8A. However, to the following video amplifier, V8A is a conventional amplifier because it receives the plate signal of V8A.)

The plate of V12A is connected to the cathode of the horizontal sweep oscillator. The voltage waveform developed here is a sine wave. When this oscillator is properly synchronized to the incoming signal, the positive peak of the sine wave occurs at the same time that the sync pulses appear at the cathode of V12A. Since negative pulses at the cathode are equivalent to positive pulses of voltage at the grid of V12A, the tube is thus triggered into conduction and plate current flows. Capacitor C_{29} charges up to a

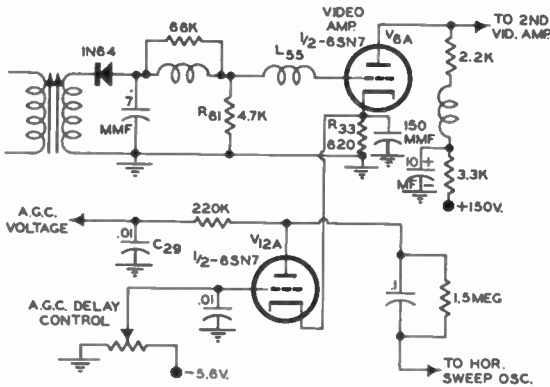


FIG. 6.18. A modified keyed A.G.C. system employed by Zenith.

negative value determined by the strength of the current flowing. When the signal is strong, the negative voltage developed here is high; when the signal is weak, less negative voltage appears. This A.G.C. voltage is then fed to the 1st, 2nd, and 3rd I.F. amplifiers and to the R.F. amplifier and converter as a control voltage.

An A.G.C. delay voltage is applied to the grid of V12A. Its purpose is to permit adjustment of the amount of negative A.G.C. voltage which V12A develops in order that this voltage may be set for the particular signal conditions prevailing at the receiver location.

The procedure for correctly setting this control is as follows. Connect a calibrated oscilloscope through a 10,000-ohm isolation resistor to the grid of V8A. This will show you the video signal at this point. Select the strongest TV signal available and observe it on the screen. Adjust the A.G.C. delay control for 2.5-volt peak output.

Satisfactory adjustment can also be made by observing the picture and backing off the A.G.C. control from its maximum clockwise position to a point comfortably below picture distortion, improper sync, or excessive contrast when the contrast control is at maximum setting.

A.G.C. Networks Containing Diode Clampers. Since it is frequently desirable for the reception of weak signals that the negative bias on the

of the contrast control. The gain of the R.F. stage is reduced when it becomes necessary to prevent distortion in the first I.F. amplifier.

When the contrast control is in the maximum gain position, or when the movable arm B is at point C, the I.F. bias is approximately -1 volt. The R.F. bias is taken from the plates of the diodes. Since the diode plates are positive with respect to the grounded cathodes (due to the presence of the $+270$ volts), the tube conducts heavily. The plate resistance of a diode is inversely proportional to the amount of voltage between the plate and

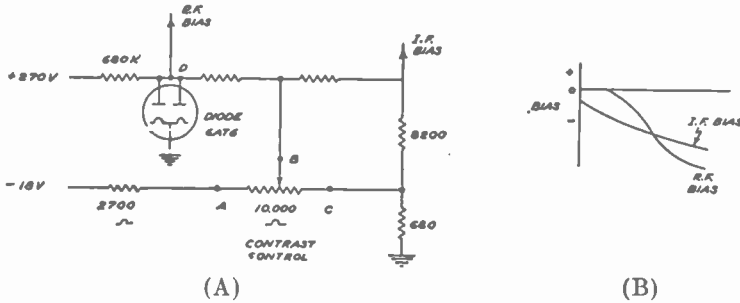


FIG. 6.20. (A) The contrast control circuit of Fig. 6.19. (B) The variation in R.F. and I.F. grid voltage due to this control.

cathode. When the tube conducts heavily, its plate resistance is quite low, and, for all practical purposes, point D is at ground potential and the full 270 volts is dropped across the 680,000-ohm resistor. Since the R.F. amplifier grid connects to this point, the R.F. grid bias is also zero. This is shown in Fig. 6-20B.

As we move the control arm B of the contrast potentiometer away from point C, the negative bias on the grids of the I.F. amplifiers increases. The contrast potentiometer is part of a series network (of the 2,700-ohm and 680-ohm resistors) which connects from the negative 18-volt power supply terminal to ground. The closer point B moves to the -18 volt terminal, the more negative it becomes. While this is occurring, the voltage at point D, where the R.F. grid bias is established, has changed very little because (and this is important) the diode is still conducting. In fact, it might appear from the circuit diagram, Fig. 6-20B, that the diode would always conduct. Such, however, is not the case. When the contrast control voltage is reduced still further, point D actually becomes slightly negative and the diode stops conducting. Thereafter, the R.F. bias voltage changes rapidly and becomes even more negative than the I.F. grids.

The circuit is unusual and merits further analysis. To see how point D can become negative even though it is attached (through a 680-000-ohm resistor) to $+270$ volts, let us simplify the diagram somewhat by removing

the diode and the connection to the I.F. grids. The result is shown in Fig. 6-21. The points A, B, C, and D are still marked as before. We now see that the leads from the plus 270-volt and the minus 18-volt power supplies have the contrast control and the 680-ohm resistor in common. For the -18 volts, there is the series path consisting of the 2,700-ohm, the 10,000-ohm (contrast control), and the 680-ohm resistors to ground. Point A is then approximately -14 volts, and point C is -0.9 volt. The current from the

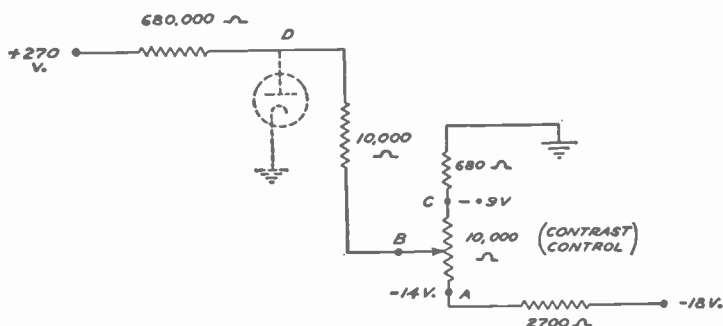


FIG. 6.21. A simplified diagram of the contrast control circuit.

positive 270-volt supply flows through the 680,000-ohm resistor, the 10,000-ohm resistor, a portion or all of the contrast control, and the 680-ohm resistor. Most of the 270 volts is dropped across the 680,000-ohm resistor, leaving so little for the remaining resistors that, when arm B is moved toward point A, the negative voltage overcomes the positive voltage and point D actually becomes negative.

Connecting a diode from point D to ground reduces the potential of this point to approximately zero when the diode conducts. However, when, as just noted, the center arm of the contrast potentiometer is moved toward the negative supply, point D becomes negative and the diode ceases to conduct.

If the manner in which the clamping tube functioned in the foregoing circuit is understood, then an understanding of its use in an A.G.C. network will come readily. In Fig. 6-22 is an A.G.C. system which is used by Stromberg-Carlson. A 6AU6 operates as a keyed A.G.C. tube, developing a negative voltage in its plate circuit which is filtered by R_1 and C_1 and then directly applied to the first and second video I.F. amplifiers.

At point A the A.G.C. line leaves the video I.F. amplifier circuit and heads toward the R.F. amplifier. It is at this point that the clamping diode is inserted. Also connecting to point A is a 3.3-megohm resistor and a 22,000-ohm resistor. 310 volts is applied to this combination with the result that most of the B+ voltage is dropped across the 3.3-megohm resistor.

Just enough voltage appears at point A to cause the clamping diode to conduct and thereby maintain point A at essentially zero (or, here, actually -1 volt) potential. R_2 (39,000 ohms) isolates point A from the rest of the A.G.C. line. Consequently the controlled video I.F. amplifiers are not affected by the clamping diode. They operate directly from the A.G.C. tube.

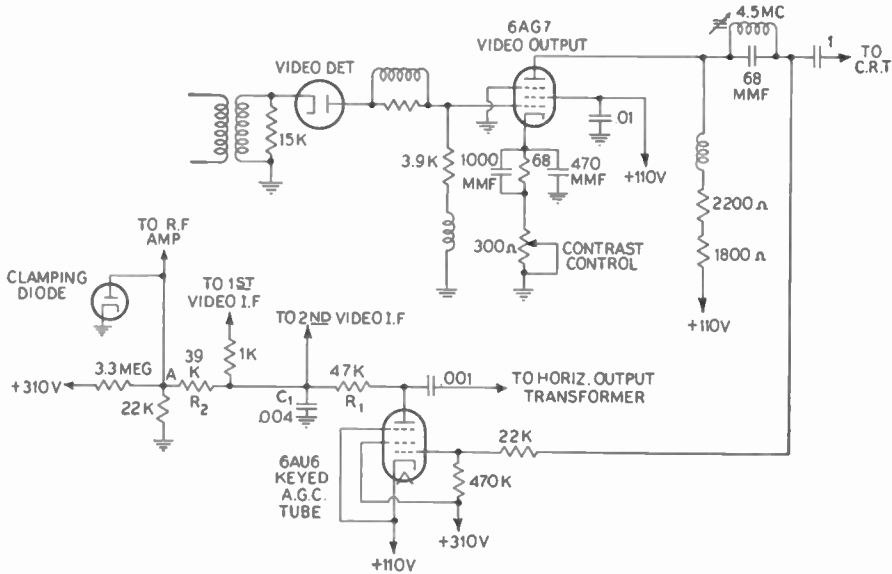


FIG. 6.22. An A.G.C. system containing a clamping diode.

Now, when the incoming signal is weak, the A.G.C. negative bias is small and the potential at point A, due to the presence of the clamping diode, is zero. The controlled video I.F. amplifiers, under the same signal condition, have a bias close to -1 volt. But as the signal strength increases, so does the negative A.G.C. bias, and part of it overcomes the slight positive potential at point A, drives the clamping diode into non-conduction, and raises the negative grid bias of the R.F. amplifier.

Clamping diodes have been used only to regulate the bias of R.F. amplifiers. They have not been used, to date, on the video I.F. system.

QUESTIONS

1. Draw the circuit of a television diode detector which will produce a negative picture phase signal.
2. Show how the above circuit can be modified to produce a positive picture phase signal.

3. Explain the difference between positive and negative video signals. What phase must the video signal possess when applied to the grid of the cathode-ray tube? Why?
4. How can a video signal, which is phased positively, be converted to the negative phase? Explain your answer.
5. Besides picture phase, what other precautions must be observed in video detector circuits?
6. What advantages would be gained by the use of A.G.C. in a television receiver?
7. What portion of the incoming signal is useful in regulating the A.G.C. voltage? Why?
8. Contrast the methods used to obtain A.V.C. in a sound receiver and A.G.C. in a television receiver.
9. Draw the schematic diagram of an A.G.C. circuit.
10. What relationship exists between the polarity of the signal at the video detector output and the number of permissible video amplifiers?
11. Would your answer remain unchanged if the signal is applied to the cathode of the image tube rather than its control grid? Explain.
12. Why is it important to maintain a good frequency response in the coupling network between the video second detector and the video amplifiers?
13. Describe briefly how a keyed A.G.C. system operates.
14. What advantages does a keyed A.G.C. system possess over the system of Fig. 6.13?
15. What is the purpose of adding a clamping diode to an A.G.C. system?

Chapter 7

VIDEO AMPLIFIERS

Some Considerations for Video Amplifiers. Up to this point, the television signal has been received and amplified by an R.F. stage, converted to another frequency by means of a mixer, further amplified by the I.F. stages, and rectified by the diode detector. As has been pointed out previously, the amplitude of the video signal at the output of the detector is not capable of providing the necessary contrast variation at the cathode-ray viewing screen. Hence, further amplification is necessary. In this chapter the requirements, operation, and structure of video amplifiers are described.

The rectified video signal, which contains the blanking, synchronizing and picture information, has a bandwidth that extends from 30 cycles to 4 mc. Every video amplifier must therefore be able to pass these frequencies without attenuation or phase distortion. These requirements are unusually severe, as may be appreciated by comparison with any high fidelity audio amplifier where the response is uniform for merely 15,000 cycles, and very little attention is given to phase distortion except in high fidelity systems. The ordinary amplifier, therefore, in its present form, is not suitable for the amplification of video signals. If it is to be suitable, revision is necessary.

It was indicated in Chapter 1 that, when the frequency response of a video amplifier fell off at the higher frequencies, the picture detail became impaired. The finer the detail to be reproduced, the higher the associated frequencies. The care that is taken in the design and construction of an amplifier will determine the type of image received on the screen of the picture tube. However, this is not the only governing factor. Another consideration must also be kept in mind, one that relates to the size of the viewing screen. While it is true that 525 lines can be placed on a 5-inch screen as well as on a 12-inch screen, the amount of detail necessary for the smaller tube is not as great as for the larger screen. The reason for this may be quite simply found in the resolving power of the human eye.

Relation between the Eye and the Detail on a Viewing Screen.

The resolving power of the eye is the ability of the eye to distinguish between objects that are placed close together. As an example, consider the card shown in Fig. 7-1, having two narrow lines located side by side. As long as the card is held fairly close to the eye, it is possible for an observer to see each line separately. As the card is slowly moved farther and farther away, it becomes increasingly difficult to see each line distinctly. Eventually a point is reached where the eye is just capable of distinguishing between them. This is the limit of the resolving power of the eye for these two lines.

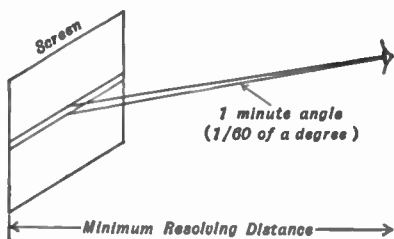


FIG. 7.1. The power of the human eye to resolve or separate two objects that are closely spaced depends upon the distance from the eye to the objects. If the objects subtend a 1 minute angle at the eye, they may be seen separately.

Quite obviously, the farther apart the lines are, or the wider they are, the more easily they may be separated from any given viewing distance. For the average person, it is claimed that as long as the two objects subtend an angle of 1 minute or more at the observer's eye, they may be seen as distinct units. This is known as the minimum resolving angle of the eye and is illustrated in Fig. 7-1. The reader can determine how wide a 1-minute angle is by divid-

ing any circle into 21,600 equal wedge-shaped parts. The angle of any small section at the wedge end would then equal 1 minute or $\frac{1}{60}$ of a degree. The distance that the observer must be from the objects in order to have the 1-minute angle subtended at his eye is known as the critical resolving distance. If the observer is farther away than this distance, the two objects merge into one. With television, it is necessary for the observer to remain outside the critical resolving distance. Coming closer only reveals the separate scanning lines and this hampers the illusion of continuity.

From the foregoing line of reasoning, it would seem possible to calculate the exact viewing distance for any size object; actually, with television images, an observer may approach closer to the screen than the calculated figure and still be unable to distinguish one line from another. This becomes possible because the resolution of two lines, for example, depends not only on their separation, but also on the amount of illumination of the lines and their relative motion. The stronger the light, the more clearly they stand out. Under these conditions, the critical resolving distance increases.

On the other hand, the introduction of motion tends to make the line of demarcation less clear-cut and the objects blend into each other at much smaller distances than if they were stationary. The latter condition pre-

vails for television images and hence the observer may view the screen from closer distances than if the motion were absent. In addition, due to the impossibility of obtaining perfect synchronizing action, the positions of the lines of the picture tend to change slightly during each scanning run, and this further obscures any clear division between lines.

Placing the same 525 lines on a 20-inch screen as on a 7-inch screen means that the proper viewing distance for the larger screen is greater than for the smaller screen. With the smaller screen, the ideal viewing distance is generally so small that ordinarily the observer never comes this close to the screen. Therefore many of the finer details of the picture are not seen, even though they are present on the screen. Manufacturers took advantage of this fact to design small screen receivers with bandwidths less than 4 mc. By the same token, as the picture tube screens became larger, it becomes more important to have the bandwidth wider.

The Low-Frequency Response of Video Amplifiers. Although it is possible to sacrifice some response at the high-frequency end of the 4-mc signal, the amplifier should possess a flat characteristic at the low end. This means uniform response to 30 cycles. Since amplifiers do not cut off sharply at any one frequency, but rather tend to decrease gradually, it is necessary for flat response at 30 cycles to have the curve extend downwards to 10 cycles, or even less.

To ascertain what loss of response at the low frequencies would mean, let us examine in some detail the camera signal, for it contains the information of the picture. In Fig. 7-2 there is a section of the signal which might

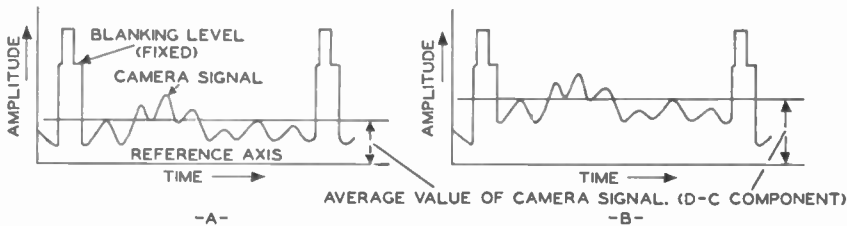


FIG. 7.2. The height of the camera signal variations above the reference axis represents the amount of background illumination that the line (or scene) will possess.

This average value is known as the d-c component of the video signal.

be obtained from the scanning of one line. On either end of the line we find the blanking and synchronizing pulses. These have a fixed level, always reaching the same voltage (or current) value whenever they are inserted into the signal. The elements of the image itself are represented by the varying voltages between the pulses and naturally differ from one line to the next. The engineer refers to these changing voltages as the a-c variations of the television signal.

In addition to the a-c variations, the synchronizing and blanking pulses of the video signal, there is another component, referred to as the d-c component. Examine the two video signals placed side by side in Fig. 7-2. The blanking levels of both are the same height and the a-c variations of each signal are identical. The only difference is in the average level of the a-c variations of Fig. 7-2A as compared with the average of the a-c portion of the signal of Fig. 7-2B. That of B is the greater of the two. This average value represents the background illumination of the scene at that line and is the d-c component of the video signal. The background illumination may vary from line to line, but this situation is unusual. Generally it changes slowly over the entire scene, and adjacent lines will have almost equal d-c components.

When the value of the d-c component is high, as in Fig. 7-2B, the people and objects of the scene being televised will appear against a dark background. This is true because with negative transmission every value is reversed. The darker the scene (or element), the greater the current. As the scene becomes brighter, there is correspondingly less current, and the a-c variations of the video signal move closer to the zero axis. Hence, as the d-c value is less in Fig. 7-2A than in Fig. 7-2B, the background illumination of A will be brighter. Neither the people nor the objects, however, have changed. A lighted background will convey to the viewer of a television scene the impression of daylight, sunshine, and clear weather. A darker background, on the other hand, will give the viewer the impression of night, fog, etc.

At the transmitter, the d-c component may be inserted manually by an operator viewing the scene from a monitor, or automatically by using the average current derived from the viewing tube, where this is possible. If the latter cannot be accomplished, the light from the scene is allowed to fall onto a photoelectric tube and the d-c component is derived in this manner. Once obtained, it is inserted into the video signal, raising the a-c component to the desired level.

From the discussion of the d-c component, which for the present is sufficient, we can see that the average illumination of a scene may change with each frame, or 30 times a second. Of course, if the exact scene is televised without any variations, the average illumination remains constant. Actually, however, this condition occurs very infrequently and each frame scanned at the camera has a different average value. In order to obtain the correct shading of the image background at the receiver, it is necessary that all transmitting and receiving circuits be capable of passing 30 cycles per second without too great attenuation. Any poor response would result in incorrect values for the background illumination and, as shown later, left-to-right stretching or smearing of large objects.

Phase Distortion. Frequency response is an important consideration in video amplifiers, but not the only one. Phase distortion, which can be tolerated in an audio amplifier, is capable of destroying the image on the cathode-ray tube screen and must also be given careful attention when an amplifier is designed. Since phase distortion is very seldom referred to in audio amplifiers, a brief discussion at this point may be helpful.

Phase distortion is produced when the time or angle relationship of electric waves to each other changes as they pass through any electrical system. For a simple example, let us consider the two sine waves shown in Fig. 7-3. Curve *A* reaches its maximum value some small time before curve *B* and curve *A* is said to lead curve *B* by a certain number of degrees. The actual number depends on the manner in which these curves were generated. Suppose that, in this case, curve *A* leads curve *B* by 45° . If, after the output of the electrical system is reached, there is still this relationship between the two waves, then no phase distortion has been introduced. On the other hand, if the value of the angle is altered, the network has introduced some phase distortion.

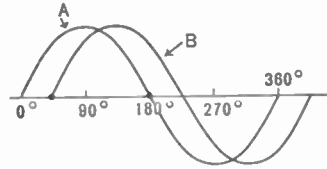


FIG. 7.3. Two waves, 45° out of phase with each other.

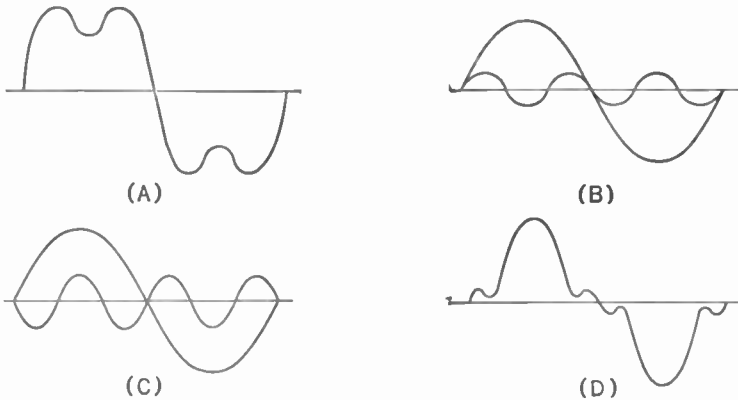


FIG. 7.4. The effect of phase distortion in changing the shape of a wave. The composite wave at (A) can be broken down into the two waves at (B). If these two component waves change their relative phases, then (D) shows one result. There may be other combinations than the ones shown.

As another example, consider the wave shown in Fig. 7-4A. This wave is actually composed of a fundamental wave in combination with its third harmonic. See Fig. 7-4B. If the effect of the network on each of these waves is different, the two waves may appear as in Fig. 7-4C, where the third harmonic wave has changed its position with respect to the fundamental,

i.e., its phase has changed. The resultant of the latter waves now assumes the shape given in Fig. 7-4D, which is certainly different from the original form of Fig. 7-4A.

How Phase Distortion Is Introduced. Now that the effect of phase distortion in changing the shape of a wave has been illustrated, let us see how the circuits found in video amplifiers may bring about such distortion. A resistance-coupled amplifier is used, as it is the only type that can be easily and economically adapted to satisfy the stringent requirements of wide-band amplifiers. A typical circuit is shown in Fig. 7-5A. For the first part of the discussion only the low-frequency response of the amplifier will

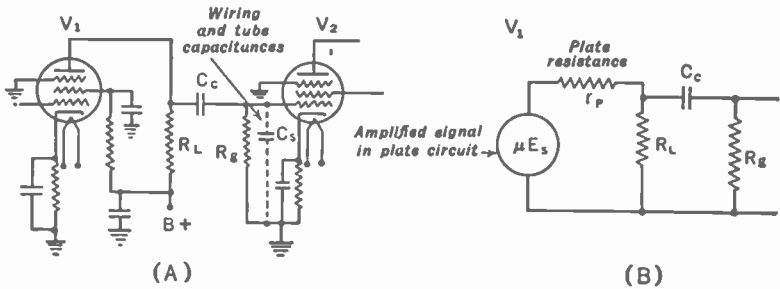


FIG. 7.5. A resistance-coupled amplifier (A) and its low-frequency equivalent circuit (B).

be considered. With this assumption, a simplification in the number of components of the amplifier may be made, as indicated in Fig. 7-5B. This is the equivalent low-frequency circuit.

When an alternating voltage is applied to the input of V_1 , an amplified version of this voltage will appear across R_L , due, of course, to the usual amplifier action of a tube. It is desired now to transfer this a-c voltage to the grid of V_2 , and this is accomplished through the series combination of C_c and R_p . How much of the total voltage of R_L will appear across R_g is dependent on how great an opposition (or impedance) C_c presents to the a-c current flowing in this circuit. At low frequencies, the opposition of the capacitor is high and a large part of the a-c voltage is lost. Less is available for R_g . This condition, as every radioman knows, is responsible for the poor low-frequency response of resistance-coupled amplifiers. Increasing the frequency will result in less voltage being lost across C_c and more will be available for R_g .

The phase of the voltage at R_g is governed by the amount of opposition C_c offers to the a-c wave passing through the circuit. Consider, for example, what the phase of the a-c current would be if only C_c were present in the circuit. The current flowing would be 90° ahead of the voltage. Now, add a resistor in series with the capacitor. The current flowing in the circuit

becomes less than 90° out of phase with the applied voltage. The voltage drop across the resistor is in phase with the current flowing through it and hence would also be less than 90° out of phase with the applied voltage. The situation is shown in Fig. 7-6, where E_g is a little less than 90° out of phase with E_L .

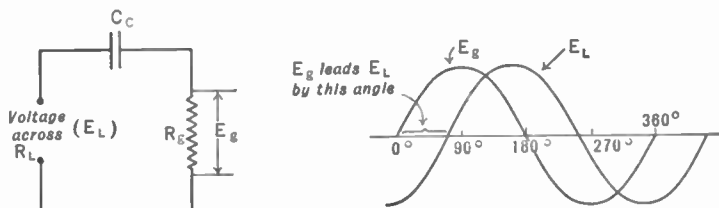


FIG. 7.6. At the low frequencies, E_g is out of phase with E_L because of the coupling capacitor C_o .

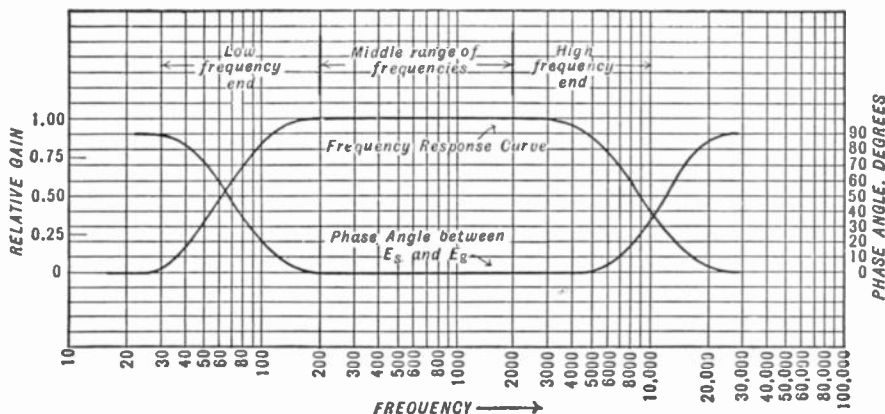


FIG. 7.7. The frequency and phase response of a resistance-coupled amplifier.

As the opposition that C_c offers to the current in the circuit becomes less and less (say, with increasing frequency), R_g becomes more important and the current approaches closer and closer in phase with E_L . At the middle range of frequencies, the opposition of C_c may be neglected entirely and E_g is in phase with E_L , similar to any other completely resistive circuit.

We see, then, that when voltages of many frequencies, and with no phase difference, are applied at E_L , the voltages appearing across E_g have different phase relationships, the degree dependent upon each frequency.

The lowest frequency will have the greatest phase angle introduced while it is traveling from the output of the tube to the input of the next stage. As the frequency rises, the phase difference becomes less, gradually reaching zero. This characteristic of a resistance-coupled amplifier may be seen from the curve in Fig. 7-7.

A complex wave, which contains many frequencies, would have its shape altered when it passed through the resistance-coupled amplifier network. With a change in shape, the effect of the wave at the grid of the picture tube must certainly be different and the resulting image is distorted to some extent. By distortion, we mean that the image is not an exact duplicate of the original. The amount of change introduced into the picture detail depends on the degree of phase distortion.

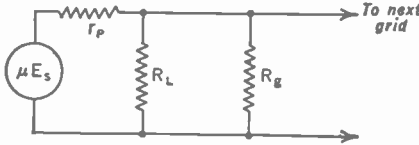


FIG. 7.8. The equivalent circuit, at the middle frequencies, of the resistance-coupled amplifier of Fig. 7.5.

In the middle range or frequencies, from 200 to 2,000 cycles, C_c has no effect on the passing waves and can be disregarded. The equivalent circuit for the middle range now assumes the form shown in Fig. 7-8. Since only resistances are involved, there is no phase shift introduced between the voltages at R_L and R_g .

At the high frequency end of the band, the input capacity, C_T , between the grid and cathode of T_2 , becomes important and must be considered. C_T has the effect of offering an easy shunting path for the a-c voltages around R_g , and the voltage appearing across this resistor decreases with increasing frequency. Note, in Fig. 7-9, that C_T forms a parallel combination with R_g . As the frequency increases, more and more current flows through C_T rather than R_g and soon the current becomes wholly a capaci-

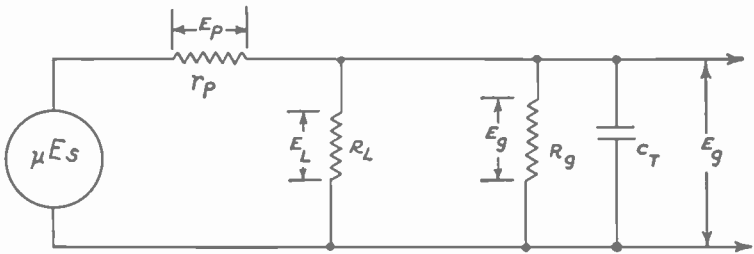


FIG. 7.9. The equivalent circuit, at the high frequencies, of the resistance-capacitance-coupled amplifier of Fig. 7.5.

tive one. Again we see E_g and E_s differing by 90° , but in this case the phase angle increases with frequency, whereas for the low frequency action the opposite was true. However, for both cases, the result is phase distortion. The graph of Fig. 7-7 shows how the phase angle between input and output voltages of the resistance-coupled network changes with frequency.

Results of Phase Distortion. To correlate the subject of phase distortion and its effect on the television picture, let us study the dependence of phase distortion and time delay. It has been noted that at the low

frequencies the phase angle between input and output voltages increased to a maximum of 90° as the frequency decreased. Suppose that a video signal is sent through this $R-C$ network containing (among others) two frequencies, say of 40 cycles and 90 cycles. From the preceding discussion, we know that the 40-cycle wave will receive a greater phase delay than the 90-cycle wave. Assume that the 40-cycle wave is shifted 45° and the 90-cycle wave, 10° . Obviously the two waves will no longer have the same relationship at the output that they had at the input, and by simple mathematics it is possible to compute their difference.

A 40-cycle wave takes $\frac{1}{40}$ of a second to complete one full cycle, or 360° . With $\frac{1}{40}$ of a second for 360° , it will take $\frac{1}{3 \cdot 20}$ of a second for the wave to change 45° ; $\frac{1}{3 \cdot 20}$ of a second is approximately 0.003 sec. Thus there will be this time difference between a maximum occurring at the input to the next tube and that occurring at the output of the preceding tube. The appearance of one will lag behind the other by 0.003 sec.

The 90-cycle wave, we know, has a 10° phase angle introduced into it. One cycle, or 360° , of a 90-cycle wave occurs in $\frac{1}{90}$ of a second. Ten degrees would require only $\frac{1}{3 \cdot 2 \cdot 40}$ of a second, or approximately 0.0003 sec. Thus the input and output variations will differ by this time interval for the 90-cycle wave.

At the cathode-ray screen, the electron beam moves across a 12-inch screen a distance of one inch from left to right in about 0.000,007 sec. The time interval is extremely short and, if waves containing the 40 and 90 cycles receive the time displacements computed above, the end result is a displacement of the picture elements that they represent. In actual television practice, the background illumination is determined by the low frequencies, and phase distortion in the video amplifiers causes a change in this shading. If, for example, the background transmitted from the studio were perfectly white, by the time it appeared at the receiver screen phase distortion would have altered it. It would now vary from white to gray, or even be black in some portions.

Any large objects or letters in the picture are distorted, too, by poor low frequency response. They appear to smear across the image like the smearing of fresh paint. The smearing effect is derived from the action described in the last paragraph, where we learned that phase distortion and time delay are directly related. A slight time delay causes certain parts of the object to be displaced from the correct position. The visible consequence of this displacement is smearing. Since the beam moves from left to right, the extended stretching of large objects will always be toward the right, or in the direction that the beam is moving. Only large objects are affected, because they are the only ones represented by the lower frequencies.

At the high-frequency end of the video signal, phase distortion results

in the blurring of the fine detail of the picture. The larger the size of the cathode-ray tube screen, the more evident this defect. Although it may be tolerated on a small screen, any enlargement immediately causes it to become apparent. Here is another reason why the larger sets require more careful design and construction. Phase distortion may be eliminated if the phase difference between the input and output voltages is zero, or if a proportional amount of delay is introduced for each frequency. Thus, a phase delay of 45° at 60 cycles is equivalent to a 90° delay at 120 cycles, etc. The first introduces a delay of approximately 0.002 sec, similar to 90° at 120 cycles. The net result is that all the picture elements are shifted the same amount, and correction is attained by positioning the picture. Phase shifts introduced by the electrical constants of one stage are additive to those of any other stage. The total phase delay of a system is equal to the sum of all the individual phase delays.

Video Amplifiers and Their Design. The preceding paragraphs have indicated what requirements are necessary for high fidelity transmission and reception of television images. The methods whereby these requirements are met in practice represent an important consideration in modern television.

The type of amplifier that can be used to give the necessary 4-mc bandwidth is restricted, almost without exception, to resistance-capacitance coupled networks. Transformers and inductances, even when they are built to possess a 4-mc width, involve a disproportionate expense. On the other hand, *R-C* amplifiers have the advantage of small space and economy and are universally employed.

From knowledge of conventional resistance-coupled amplifiers that the radioman possesses he knows that a flat response is obtained in the middle range of frequencies, say from 200 cycles up to approximately 2,000 cycles with ordinary circuits. A frequency response curve is illustrated in Fig. 7-7 and applies to any ordinary *R-C* amplifier. As we are also interested in phase response, this, too, is indicated in Fig. 7-7. The frequency and phase characteristics of the amplifier, throughout the middle range, are suitable for use in video amplifiers, and this section of the curve requires no further improvement. However, the responses at either end of the curve are far from satisfactory and corrective measures must be taken. Fortunately, any changes made in the circuit to improve the high- or low-frequency responses of the curve will not react on each other (with one limitation noted later), and each end may be analyzed separately and independently. Let us begin first with the high-frequency compensation.

When determining the high-frequency operation of a resistance-coupled amplifier, we include the internal and external plate resistances, the grid input resistor of the next stage, and any shunting capacitances that are

present in the circuit. The coupling capacitor, C_c , offers negligible opposition to high-frequency alternating currents and can be disregarded. Fig. 7-9 shows the equivalent circuit of the $R-C$ amplifier applicable under these conditions.

Since R_L and R_g are both resistances and are both constant in value, any change in high-frequency response must be due to the shunting capacitance, C_T . The reactance of a capacitor decreases with frequency and, in effect, the total impedance of the parallel combination of R_L , R_g and C_T becomes less as the frequency increases. The alternating voltage that is developed at the tube will divide between r_p , which is the internal tube plate resistance, and the parallel combination of R_g , R_L , and C_T . Since the value of the impedance of the parallel combination decreases, it means that more and more of the output voltage will be lost across the tube's plate resistance, r_p . If less voltage reaches the grid of the next tube, less is available for amplification.

To increase the gain at the high-frequency end of the response curve, it is obvious that the value of the shunting capacitances should be decreased. The shunting capacitance is composed of three components: (1) the output capacitance of the preceding tube, (2) the input capacitance of the following tube, and (3) the wiring capacitance to ground. For a typical video amplifier tube the output capacitance is approximately $5 \mu\mu f$. The wiring capacitance may run from about $5 \mu\mu f$ to 15 or $20 \mu\mu f$, whereas the input capacitance of the next tube might well be about $10 \mu\mu f$ and, unless tube construction changes radically, will remain close to this figure for most tubes. The wiring or stray capacitance can be reduced if care is taken when constructing the amplifier. The wiring capacitance will be kept at a minimum (about $5 \mu\mu f$) if all leads are kept short, low-loss sockets are used, and the parts are intelligently placed.

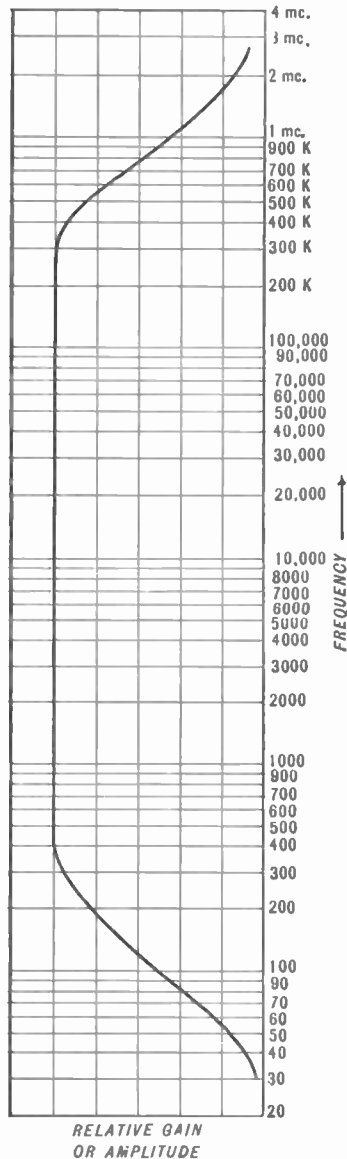


FIG. 7.10. Frequency response of $R-C$ coupled amplifier with decrease in shunting capacitance.

With the foregoing reduction in the value of the capacitance, due perhaps in part to the use of tubes of small internal capacitance and in part to careful wiring, it is possible to increase the frequency response of an amplifier to 1 or 2 mc. The gain, however, especially near the end of the curve (shown in Fig. 7-10), is not uniform. To improve the response uniformity, the value of the load resistor can be lowered, probably close to the impedance presented by the shunting capacitance. In this manner, the effect of capacitor C_T is less and it does not begin to destroy the linearity of the frequency response curve until some higher frequency. The results for several resistor values are shown in Fig. 7-11. As the resistance becomes

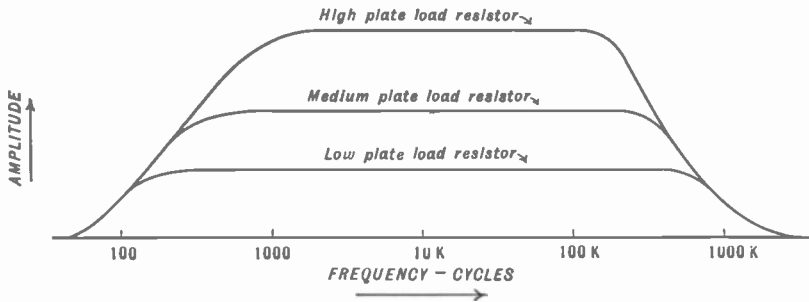


FIG. 7.11. By lowering the plate load resistor, it is possible to increase the extent of the flat portion of the response curve.

less, the flat portion of the curve increases, but the stage gain decreases. For many of the amplifiers found in commercial receivers, values of R_L as low as 1,500 ohms are used. This results in gains in the neighborhood of 20 or so per stage, which is not very high. Obviously, further extension of the frequency range of the amplifier by lowering the load resistance value is not very feasible.

A more satisfactory method of increasing the flat response of the amplifier can be obtained through the insertion of a small inductance in series with the load resistor. The inductance is so chosen that it will neutralize the effect of the shunting capacitances at least to the extent that we may improve the amplifier response at the upper frequencies. This method is known as shunt peaking.

A circuit diagram using this compensating inductance is given in Fig. 7-12. L is chosen to resonate with C_T at or above the highest frequency at which flat response is desired. In this manner, the peaking inductance tends to compensate for the loss caused by the capacitance C_T and the curve remains flat. If too much peaking is resorted to, the curve will rise sharply near the resonant point of L and C_T and result in a hump which is undesirable. The appearance of the response curve for several values of L is shown in Fig. 7-13. In practice the value of L is given by

$$L = (\frac{1}{2})C_T R_L^2 \text{ henrys}$$

where C_T = the total shunting capacitance in farads,
 R_L = the load resistance in ohms.

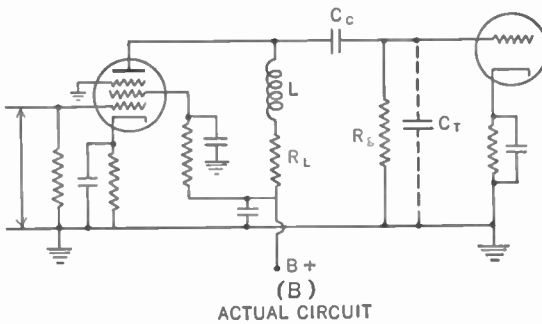
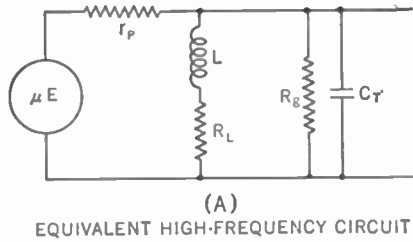


FIG. 7.12. High-frequency compensation for an $R-C$ amplifier.

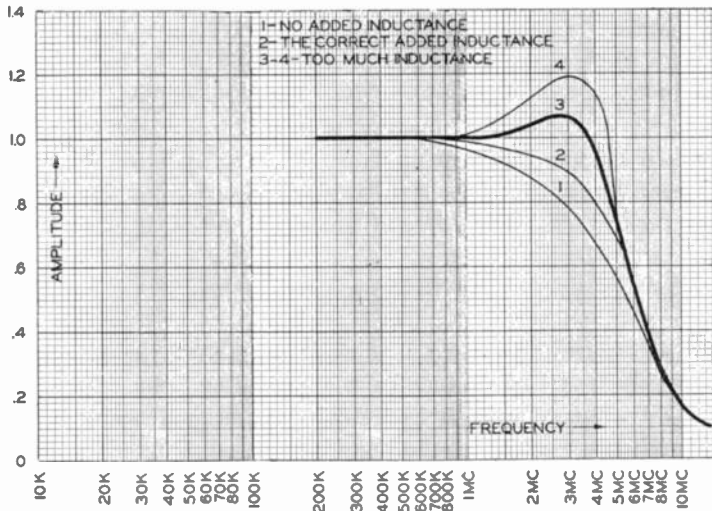


FIG. 7.13. The effect on amplifier response of the insertion of various amounts of peaking inductance in the plate load.

With this value for L , curve 2 is obtained. It has also been found that the introduction of L improves the phase angle response of the network.

A second method whereby the high-frequency response can be improved is to insert a small coil in series with the coupling capacitor, as illustrated in Fig. 7-14. This method gives higher gain and better phase response than shunt peaking. The added advantage of this type of coupling is due to the fact that the components of C_T are no longer lumped together in one unit, but have been separated. On the left-hand side of the series inductance we have the output capacitance of the preceding tube, while

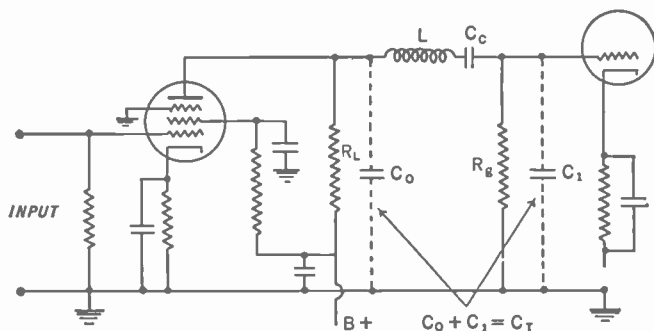


Fig. 7.14. High-frequency compensation by means of a series peaking coil.

on the other side we find the input capacitance of the next tube. With this separation, the load resistor R_L may be chosen higher in value because only C_0 is directly across it and not the larger C_T . As C_0 is smaller than C_T , its capacitive reactance is greater and it will have less of a shunting effect on R_L . Hence, a larger value of R_L is possible. The series combination of the inductance and the total capacitance is designed, by proper choice of L , to have a resonant frequency above the highest video frequency desired, which is generally 4 mc.

It has been found that best results are obtained when the ratio of C_1 to C_0 is approximately 2. The value of the series coil, L , is given by

$$L = \frac{1}{8\pi^2 f^2 C_0}$$

where f = the highest frequency it is desired to have the amplifier pass.

C_0 = the output capacitance of the preceding tube. To this we also add whatever wiring and stray capacitance would be associated with this portion of the coupling circuit.

It is further possible to combine shunt and series peaking and obtain the advantages of both. The shunt coil is designed to neutralize the output capacitance of the preceding tube while the series coil combines with the

input capacitance (and stray wiring capacitance) of the next tube. With this double combination, it is possible to achieve 1.8 times more gain than can be derived through the use of shunt peaking alone. Furthermore, the phase distortion of the coupling network is lower than either of the previous two types. An amplifier using combined shunt and series peaking is shown in Fig. 7-15. A resistor is shunted across the series coil to minimize any sharp increase in circuit response due to the combination of the series coil inductance and its natural or inherent capacitance. The coil is designed

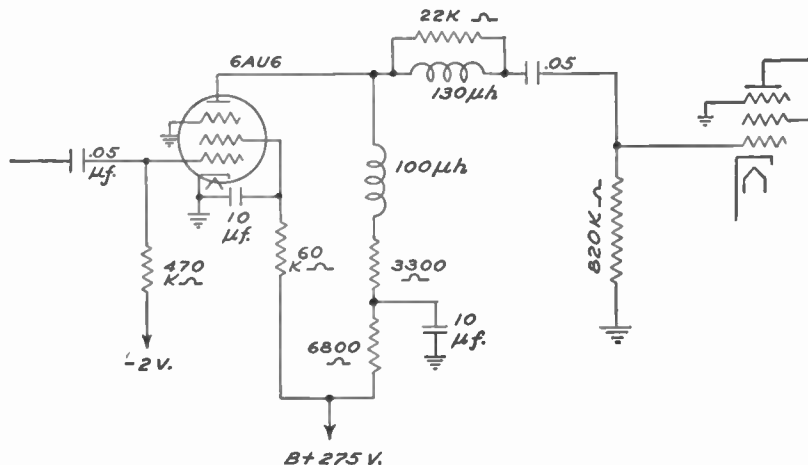


Fig. 7.15. A video amplifier employing series and shunt peaking.

to have a natural frequency considerably above the highest video frequency. In production, however, a certain number of coils will be produced with natural resonant frequencies within the range covered by the amplifier. The effect is a sharp rise in response, similar to curve 4, Fig. 7-13. It is to prevent this peak, if it occurs, that the shunting resistor is used. Its value is generally four to five times the impedance of the series coil at the highest video frequency.

For the combination circuit, the values of L_s , L_c , and R_L are obtained from the following relationships:

$$R_L = \frac{1.8}{2\pi f C_t} \text{ where } C_t = C_i + C_o$$

$$L_s = .12 C_t R_L^2 \text{ (shunt coil)}$$

$$L_c = .52 C_t R_L^2 \text{ (series coil)}$$

f is the highest frequency it is desired to have the amplifier pass.

Low-Frequency Compensation. With the high-frequency end of the response curve taken care of, let us determine what changes can be made

to improve the low-frequency response. At this end of the band, it is possible to disregard the shunting capacitances since their reactance, given by

$$X_c = \frac{1}{2\pi fC}$$

is very high, and they do not affect the low-frequency signal voltages in any way. Now, however, it becomes necessary to include the coupling capacitor. The equivalent low-frequency circuit was previously given in Fig. 7-5. The operation of the circuit, as explained, shows that the lower the frequency,

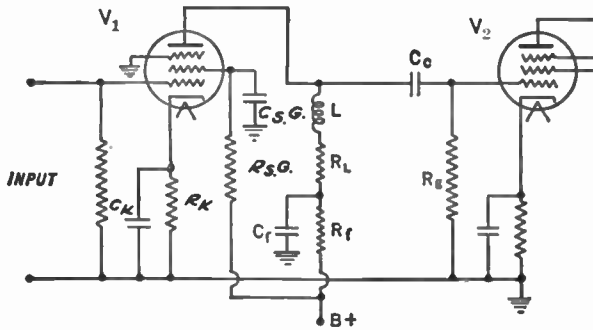


Fig. 7.16. Low-frequency compensation (R_f , C_f) of an R - C amplifier.

the greater the effect of the coupling capacitor. The response gradually falls off because the reactance of C_c soon becomes dominant and a large portion of the output voltage of V_1 is lost here. The phase delay of the signal begins to change, eventually approaching 90° . As a result, the back-ground illumination of the reproduced image is affected.

To increase the linear response at the low frequencies, either C_c should be made larger so that it will have less reactance, or R_g should be made larger. The limit of the size of either C_c or R_g is governed by several factors:

1. Too large a value of C_c increases the stray capacitance to ground and is certain to interfere with the high-frequency response.
2. A large coupling capacitor generally has an appreciable leakage current. This would permit the positive power supply voltage on the preceding plate to affect the grid of the following tube and bias it positively.
3. A large value of R_g could prove detrimental if the tube to which it is attached has even a slight amount of gas.
4. Finally, high values of R_g and C_c result in motorboating (or oscillations) due to the slow building up and leaking off of charge across the combination.

It is possible to improve the low-frequency response without making either R_g or C_c too large by inserting a resistor and capacitor in the plate circuit of tube V_1 , as indicated in Fig. 7-16. R_f and C_f are the two added components and they form the low-frequency compensation circuit. Through the addition of this resistor and capacitor, the impedance in the plate circuit is increased for the lower frequencies and greater gain results. At the high frequencies, C_f by-passes R_f and effectively nullifies it. Furthermore, C_f and R_f serve as a decoupling filter which aids in stabilizing the stage by preventing any low-frequency oscillations or motorboating from feedback between stages by way of the power supply.

The value of C_f in Fig. 7-16 is obtained from the expression:

$$R_L C_f = C_c R_g$$

where R_L , C_c , and R_g have previously been assigned values. R_L will be determined by the highest frequency to be passed by the amplifier and C_c and R_g will be as large as possible but within the limitations noted above. Finally, R_f should have a resistance which is at least twenty times greater than the impedance of C_f at the lowest frequency to be passed.

C_f and R_f provide the greatest amount of compensation, but there are additional factors which influence the extent of the low-frequency response. One of these is the screen-grid dropping resistor and by-pass capacitor. For best results, R_{sg} and C_{sg} should have a time constant which is at least three times as long as the period ($1/f$) of the lowest video frequency to be passed by the amplifier. A second governing factor is the cathode resistor, R_k , and the cathode by-pass capacitor, C_k . These should be chosen so that they satisfy the following expression:

$$R_k C_k = R_f C_f$$

Admittedly, the latter two circuits are not quite as important as the decoupling resistor and capacitor, C_f and R_f , but they should be considered in the amplifier design.

In the design procedure of video amplifiers, the values of the high-frequency compensating components are chosen first. These include R_L , L_s , and L_c . Next, the low-frequency compensating components, C_f and R_f , are computed, then R_{sg} and C_{sg} , and finally R_k and C_k . The values of each of the latter three resistors must fall within the operating characteristics of the tube as recommended by the manufacturer. This imposes a limitation. However, since we are concerned with a time constant in each instance (as $C_f \times R_f$, $R_{sg} \times C_{sg}$, and $R_k \times C_k$) rather than the individual value of each part, we can usually satisfy all the required conditions.

When the high- and low-frequency compensating circuits are applied to a video amplifier, the result appears as shown in Fig. 7-15. The frequency

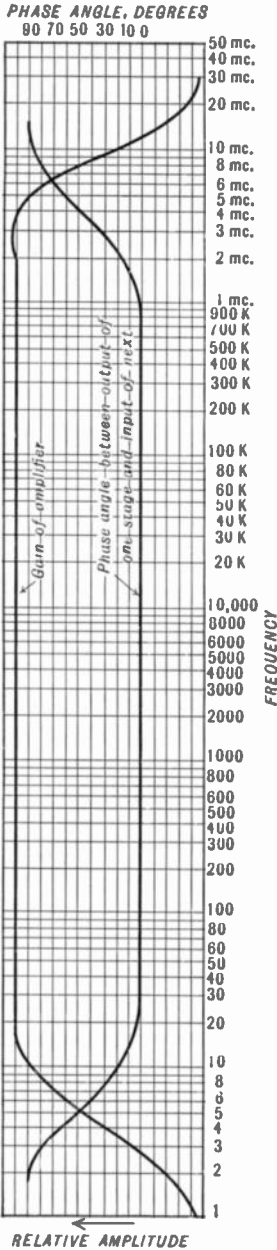


FIG. 7.17. The frequency response of a fully compensated resistance-coupled amplifier.

and phase response of this amplifier is plotted in Fig. 7-17. The number of such stages required between the video detector and the cathode-ray tube will depend upon the polarity of the signal at the output of the detector. If the picture phase is negative, its current or voltage increases as the picture elements become darker; an odd number of video amplifiers must be used because each amplifier changes the signal by 180°. Hence, if the picture phase is negative, one stage will convert it to a positive phase and the signal in this form can now be applied to the grid of the viewing tube. Of course, three, five, or any odd number of stages will also answer the purpose. For a positive picture polarity at the output of the detector, an even number of stages is necessary.

If there is any doubt as to how a signal is changed 180° when it passes through a tube, consider the following action. An a-c signal is applied to the grid of a tube, as indicated in Fig. 7-18. As the grid becomes increasingly positive, more plate current flows, resulting in a greater voltage drop across R_L , with the plate end of the resistor becoming more negative. Thus an increasingly positive grid gives rise to a decreasing plate voltage. These two voltages are 180° out of phase. For the opposite case, with the grid voltage going negative, the plate current decreases with a resultant decrease in the voltage drop across R_L . The plate voltage will increase.

It is well to remember that this 180° phase reversal in a tube has nothing at all to do with any phase distortion caused by the coupling capacitor or shunting tube capacitances. The tube reversal merely has the effect of changing a positive picture phase into a negative picture phase or vice versa. The voltage output is still in step with the wave at the input and there is no time delay introduced at all.

Contrast Controls in Video Amplifiers. The purpose of a contrast control is to regulate the amount of video signal reaching the picture

tube. It is manually operated and is adjusted by the viewer until the picture intensity suits his tastes. If the room is light, he may want to turn the contrast control up. If the room is darkened, he may want to decrease the picture intensity.

Regulation of picture intensity may be accomplished in several ways. Variation of the bias of one or more video I.F. amplifiers will vary the gain

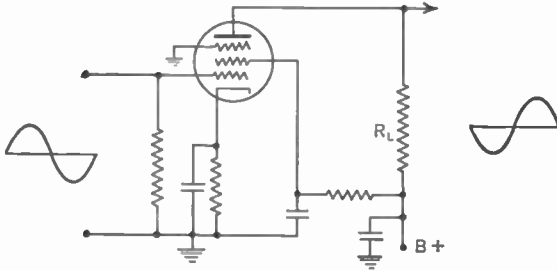
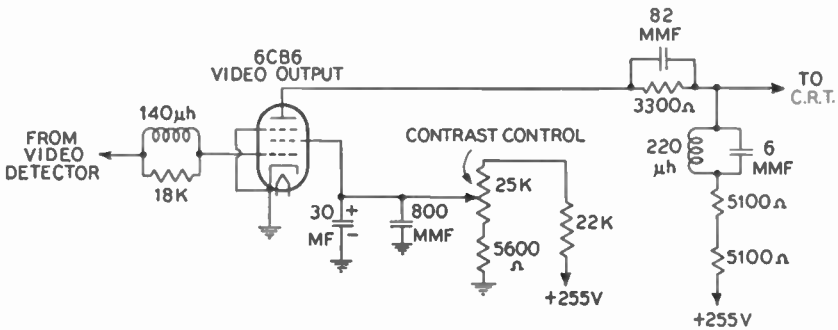


FIG. 7.18 The input and output voltages of a tube showing the 180° phase reversal.



Courtesy G.E.

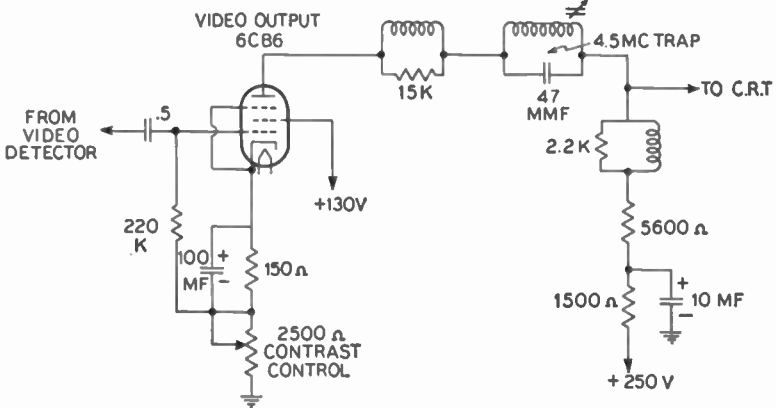
FIG. 7.19. A video amplifier stage containing a contrast control in the screen grid circuit.

of these stages and this, in turn, will control the signal amplitude at the picture tube. A suitable method for doing this was shown in Chapter 5.

Signal gain can also be varied in the video amplifier stages following the video second detector. In the circuit of Fig. 7-19, a contrast control is placed in the screen grid lead of the 6CB6 video amplifier. By regulating the screen-grid voltage, we can vary the gain of the amplifier and of the signal that passes through.

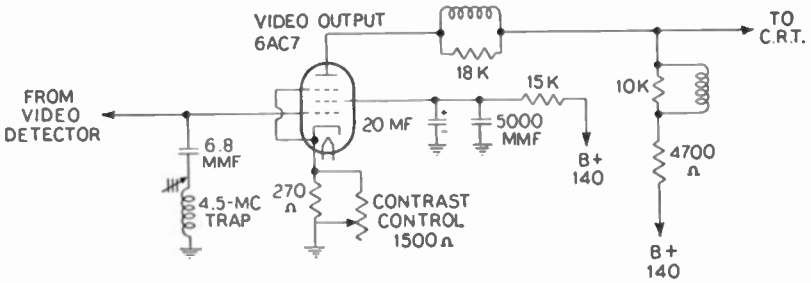
Another approach to video signal amplitude control is shown in Fig. 7-20. Here the contrast potentiometer is placed in the cathode leg of the stage. At first thought it might appear that the control varies the bias on this stage and thereby varies signal gain. Actually, because the 220,000-

ohm grid resistor ties in to the bottom of the 150-ohm cathode resistor, the bias of the stage is only slightly affected by the contrast control. What the control really does is to determine how much of the video signal appears across the 5600-ohm resistor in the plate circuit and how much of it appears across the 2500-ohm contrast control. Naturally, the more the con-



Courtesy Hallcrafters

FIG. 7.20. A video amplifier with the contrast control in the cathode leg of the 6CB6.



Courtesy Admiral

FIG. 7.21. Bias variations to achieve control of signal amplitude.

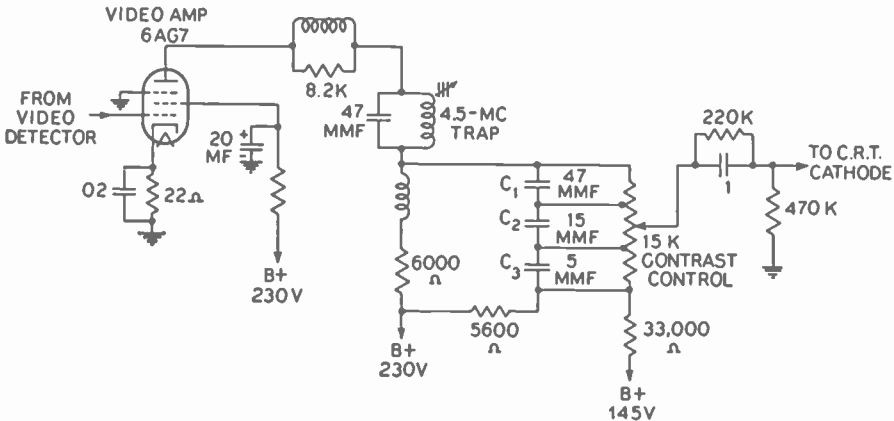
trol receives, the less there is available for the 5600-ohm resistor and the cathode-ray tube to follow.

A variation of negative feedback voltage to achieve gain control is obtained in the circuit shown in Fig. 7-21. The 270-ohm resistor is shunted across the 1500-ohm contrast control to prevent the total cathode resistance from exceeding 225 ohms (approx.). To do so would result in picture distortion. Maximum gain is accorded the signal when the cathode resistance is zero.

A number of sets have employed the contrast control arrangement shown in Fig. 7-22. The video signal which is amplified by the 6AG7 video am-

plifier appears across the contrast control (and the 33,000-ohm resistor below it). The amount of the signal which then actually reaches the picture tube depends upon the position of the center arm of the control. This action is quite similar to that of the volume control in a radio receiver.

Inspection of the diagram will also reveal that the control has branched across it three small, fixed capacitors. These serve to equalize the response of this unit so that all a-c and d-c voltages receive similar treatment. A potentiometer (or any other resistor, for that matter) acts as a pure re-



Courtesy DuMont

Fig. 7.22. A compensated contrast control arrangement.

sistance only at very low frequencies. At high frequencies, stray capacitance across sections of the control offer alternate paths for high-frequency signals with the result that the proportion of this voltage reaching the picture tube is not the same as it is for d-c and low-frequency voltages. For example, when the control arm is two-thirds of the way up the potentiometer, the picture tube receives two-thirds of the d-c and low-frequency voltage present across the control. However, it receives less than half of the high-frequency voltage because of the shunting capacities.

To equalize the response of the control, additional small capacitances are branched across the potentiometer in the manner shown in Fig. 7-22. These units are so chosen that the picture tube receives the same proportion of low- and high-frequency voltages when the movable arm is at the top of the potentiometer as when it is at the two tapped positions.* At intermediate positions, the equalization is not complete, but the fall-off is not sufficient to produce a noticeable effect on the picture.

The addition of the three capacitors across the video circuit tends to de-

* Note that this does not mean the tube receives the same amount of voltage in all three positions. Obviously, it does not.

grade its high-frequency response. However, this is overcome by increased peaking in the video frequency amplifier.

4.5-mc Traps in Video Amplifiers. In Intercarrier receivers, both the video and sound I.F. signals reach the video second detector. Within this tube a beating or mixing takes place between the video and sound carriers, and a 4.5-mc beat note is produced. The sound intelligence is contained within this signal and is subsequently removed from the video system and



Courtesy G.E.

Fig. 7.23. An enlargement of a section of a screen showing the wormy, fine-line pattern produced by a 4.5-mc beat signal.

transferred to the sound system. As a precautionary measure, it is customary to insert a 4.5-mc trap in the video amplifiers beyond the separation point to insure that none of this 4.5-mc voltage reaches the picture tube. (Should this occur, it would produce a fine-line pattern on the screen similar to that shown in Fig. 7-23.)

Trap circuits used are either parallel resonant as shown in Fig. 7-20, or series resonant as indicated in Fig. 7-21. Note again that they are placed in the video system *beyond* the sound separation point.

A number of conventional type receivers also utilize a 4.5-mc trap because it is usually impossible (or impractical) to prevent completely some sound I.F. carrier voltage from reaching the video detector. This in spite of the sound traps which these sets have in their video I.F. system.

It is interesting to note that when the sound I.F. signal reaches the video detector, two actions may occur. First, there is the formation of the 4.5-mc

beat note (provided the video carrier is also present). Any amount of sound I.F. voltage will produce this effect. Second, the F-M sound signal may, by itself, be detected to release the audio voltages it contains. This occurs by a process known as slope detection.* These audio voltages are comparatively low in frequency, and so the visual effect they produce on the screen will differ from that of the 4.5-mc signal. Fig. 7-24 shows the face of a



FIG. 7.24. The visual result when sound voltages reach the cathode-ray tube.

picture tube afflicted with low-frequency sound voltages. It is seen that the voltages appear as a number of horizontal black bars. The intensity of the bars will vary with the sound amplitude, and the thickness of the bars will depend upon the sound frequency. No bars will be present when no sound is being heard.

The appearance of the sound bars will occur only when the amplitude of the sound signal reaching the video detector is quite large, certainly much larger than required for the operation of an Intercarrier system. Hence, when these bars are observed on the screen, it is an indication that either the video I.F. system (and its traps) is misaligned or that the Fine-Tuning control is grossly mis-set.

* See the author's, *F-M Simplified*, D. Van Nostrand Co., Inc., 1951.

Direct-coupled Video Amplifiers. The video amplifiers which have been discussed thus far have employed resistance-capacitive coupling between stages. The capacitor served to prevent the d-c voltage at the plate of the preceding stage from reaching the grid of the following stage. At the same time, it did permit the a-c portion of the signal to pass.

In television, the signal not only possesses an a-c portion, but what can be called a d-c component as well. The a-c component represents the detail of the picture, while the d-c component provides information concerning the background brightness. Now, when the complete video signal is passed

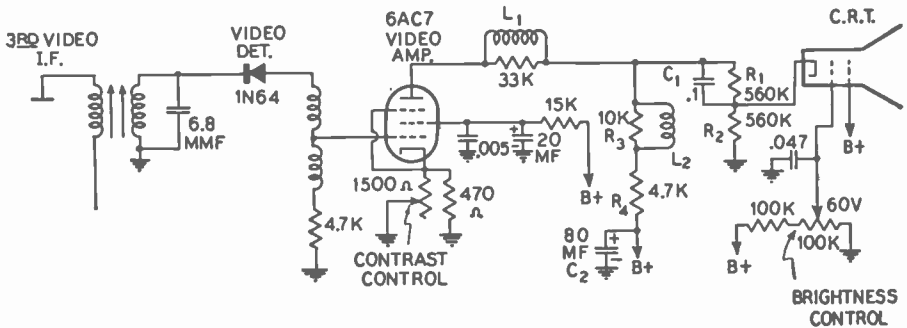


Fig. 7.25. A direct-coupled video amplifier system.

through an $R-C$ network, the capacitor removes the d-c component. This can be reinserted subsequently by a special d-c restorer. (Chapter 8 is devoted exclusively to a description of this process.) However, if we remove the coupling capacitor, then the d-c component will not be affected, and a special d-c restorer circuit becomes unnecessary.

Direct-coupled amplifiers have been known and used for many years even before the advent of television. Recently they have been employed in television receivers, too, and it might be instructive to see what form they take and how they differ from $R-C$ coupled amplifiers.

A video system in which there exists a d-c path from the video second detector to the picture tube is shown in Fig. 7-25. The 6AC7 amplifier receives the video signal direct from the video detector, amplifies it, and then feeds it to the cathode of the picture tube. Because of the direct coupling, the cathode of the picture tube has a fairly high positive potential, in this case, 82 volts. Partially to offset this, the control grid of the picture tube is also given a positive voltage. (The exact value will be determined by the setting of the brightness control. The 60 volts indicated in the diagram is an average value.) In this way we still maintain the control grid negative with respect to the cathode, as required for proper operation of the tube.

Elimination of the coupling capacitor extends the amplifier response down to d-c or zero frequency. It was the presence of this capacitor, it will be remembered, which caused the response of an amplifier to drop off at the low frequencies. High-frequency compensation is still necessary, and L_1 and L_2 of Fig. 7-25 serve this purpose.

The presence of C_1 , R_1 , and R_2 requires some explanation because they are tied in with the fact that direct coupling is used between the video amplifier and the picture tube.

If we tied the plate of the 6AC7 directly to the cathode of the picture tube, the positive voltage at this cathode would be close to 195 volts (in this circuit). To offset this, the control grid of the picture tube would also have to have a high positive voltage. And finally, since the second grid (or first anode) of the picture tube should be 300 to 350 volts more positive than the control grid, its voltage in this case would have to exceed 500 volts. In order to avoid using voltages this high (which are obtained from the low-voltage power supply), the following system is employed.

In the plate circuit of the 6AC7, the load presented to the a-c component of the video signal is a combination of L_2 , R_3 , and R_4 . The load for the d-c portion of the signal, however, includes not only these three components but also the resistance formed by C_2 and the power supply. C_2 , an electrolytic capacitor, may represent a short circuit to a-c, but to d-c its impedance is very high. Therefore this impedance in parallel with the power supply impedance must also be considered in computing the load seen by the d-c component of the video signal.

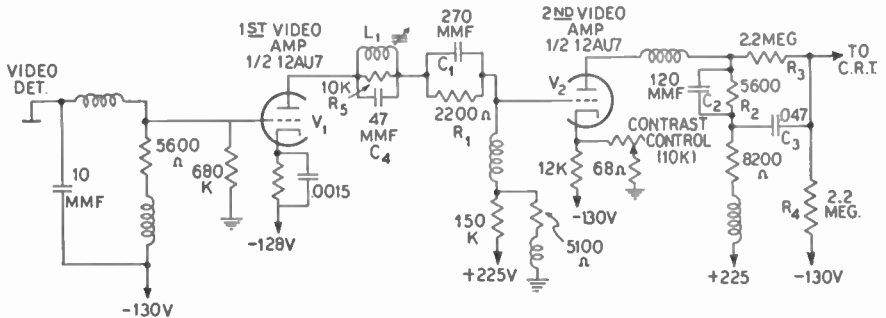
Since the d-c load of the amplifier is greater than its a-c load, the d-c component will receive more amplification than the a-c component. To equalize this response, a d-c voltage divider consisting of R_1 and R_2 is formed. From the value of these resistances it is seen that slightly less than half of the d-c component reaches the picture tube. The $.1\text{-}\mu\text{f}$ capacitor prevents the a-c signals from being attenuated by shunting them around R_1 .

The voltage-dividing action of R_1 and R_2 also reduces the d-c plate voltage reaching the cathode of the picture tube and thereby reduces the amount of B+ required by the control grid and second grid of the C.R. tube.

A two-stage, direct-coupled video amplifier system is shown in Fig. 7-26. Note how the voltages start at -130 volts at the grid of V_1 and work up to $+225$ volts at the plate of V_2 . This increase in operating voltages from stage to stage is characteristic of direct-coupled amplifiers. If we added another stage after V_2 , the grid of this third stage would be positive because of its connection to the plate of V_2 . This, in turn, would mean that the cathode and plate of this third tube would have to be even more positive. By adding enough stages, the B+ voltage required would soon rise to extremely high values. Fortunately, television receivers seldom

require more than two stages of video amplification, and hence direct coupling may be successfully employed.

In the circuit of Fig. 7-26, the balance between the amplification accorded the a-c and d-c components is carefully maintained. C_1 and R_1 form one equalizing network wherein the a-c component is left alone while the d-c component is reduced. $C_2, C_3, R_2, R_3,$ and R_4 form another such network. Finally, $L_1, C_4,$ and R_5 constitute a 4.5-mc trap to remove any 4.5-mc signal that may be present in the circuit.



Courtesy RCA

FIG. 7.26. A two-stage direct-coupled video amplifier system.

QUESTIONS

1. Why must we modify a high fidelity audio amplifier before it can be used as a video amplifier?
2. Must the full 4.0-mc video signal be used in all television receivers? Explain.
3. What governs the minimum viewing distance of a television screen?
4. What purpose does the d-c component of a television signal serve?
5. What occurs to the image when the d-c component is removed?
6. Why is phase distortion important in television? Why is it unimportant in sound receivers?
7. Explain how phase distortion occurs in a television circuit.
8. What particular components are responsible for low-frequency phase distortion? Why?
9. What is the effect of high-frequency phase distortion?
10. What is the visual effect of low-frequency phase distortion?
11. What factors tend to reduce the high-frequency amplification of an audio amplifier?
12. Draw the equivalent high- and low-frequency circuits of an audio amplifier.
13. What is a peaking coil? Why is it useful in video amplifiers?
14. Draw the circuit of a video amplifier containing high-frequency compensation.
15. Explain and illustrate the differences between series peaking, shunt peaking, and a combination of the two.
16. Why can we disregard all shunting capacitances when designing the low-frequency compensation network?

17. Without adding any additional components to an audio amplifier, how can we partially improve its low-frequency response? What limitations exist to this method?

18. Draw the circuit of a video amplifier containing low-frequency compensation.

19. Specify the various points in an amplifier where low-frequency compensation can be applied. Indicate the compensation suggested in each instance.

20. What would be the visual effect of over-peaking? Under-peaking?

21. A video amplifier is to use a single shunt peaking coil. The response is to extend to 4.0 mc. If the load resistor is 2,000 ohms and the total shunting capacitance is $20 \mu\mu\text{f}$, what value should the peaking coil have?

22. In the same video amplifier, the coupling capacitor (C_c) has a value of $0.1 \mu\text{f}$ and the grid resistor of the following stage a value of 250,000 ohms. What value should C_f have in the low-frequency compensation network?

23. Explain how the low-frequency compensation networks accomplish their purpose.

Chapter 8

D-C REINSERTION

The Need for D-C Reinsertion into Video Signals. Although the signal at the output of the final video amplifier may have sufficient amplitude to be applied directly to the cathode-ray tube, its form may not be entirely suitable. The television signal, we know, contains an a-c component, a d-c component, and blanking and synchronizing impulses. Of importance, at the moment, is the d-c component, its function within the video signal, and what happens when it is removed from the signal.

Several lines of a typical video signal are shown in Fig. 8-1. Between every two successive synchronizing and blanking pulses, we have the camera signal variations, ranging from white (at the most positive value) to black at the level of the blanking pulse. The signals are shown in the posi-

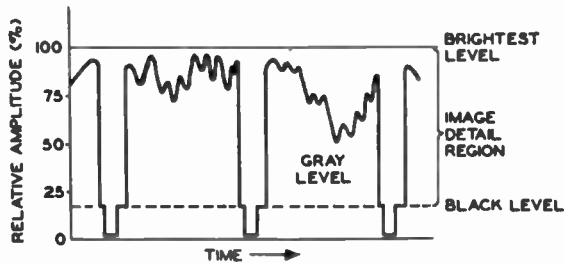


Fig. 8.1. Several lines of a typical video signal.

tive picture phase form. When applied to a cathode-ray tube, each different value of video voltage produces a different spot intensity on the cathode-ray tube screen and from all these light gradations we obtain the image.

Suppose, now, we take a video signal and, while maintaining the same camera signal variations, we first move these variations closer to the blanking pulse level (Fig. 8-2A) and then shift the same variations as far away as possible from the blanking pulses (Fig. 8-2B). What would be the visual

result in each instance? Since the blanking level represents the point at which the cathode-ray tube beam is supposed to cut off, moving the video signal closer to this level means that the overall background of the image will become darker. On the other hand, when the video signal variations are farther away from the blanking level, the background of the image becomes brighter. Note, however, that because the video signal variations are identical in each instance, the same scene is obtained. The only thing we have altered by shifting the relative position of the video signal is the background brightness. In the first instance it is dark; in the second illustration it becomes bright. We can simulate the same conditions in a room by in-

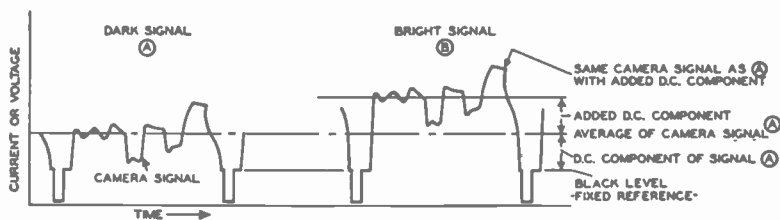


Fig. 8.2. Two video signals containing the same detail (a-c component) but different background brightness (d-c component).

creasing or decreasing the intensity of the electric lights. This change does not affect the objects themselves in the room but merely the overall brightness of the scene.

To distinguish between the camera signal variations and the average level of these variations (or the average distance of these variations from the blanking level), it has become standard to call the latter the d-c component and the former the a-c component of the video signal. The average level of the signal can be altered by the insertion of a d-c voltage, thereby raising or lowering the average level of the video signal and changing the background brightness of the image.

At the transmitter, the level of the blanking pulses is established as the dark level, at which point the electron beam in the receiver cathode-ray tube is to cut off and the screen, for that point, to become dark. When the a-c video signal variations obtained from the camera tube are combined with this blanking voltage and the sync pulses, we have a complete video signal. At any point along the program line, the distance between the average level of the a-c video signal and the blanking level may be varied (through insertion of a d-c voltage) to produce the desired shading or background brightness as dictated by the program director. Note that, since the d-c voltage moves the video signal variations closer to or farther away from the blanking level, we are using this level as a reference. Therefore, the level of the blanking pulses

must remain fixed, and the signal is transmitted with this relationship maintained.

The second detector output in the receiver contains the full video signal, as shown in Fig. 8-1. The blanking pulse of each line is aligned to the same

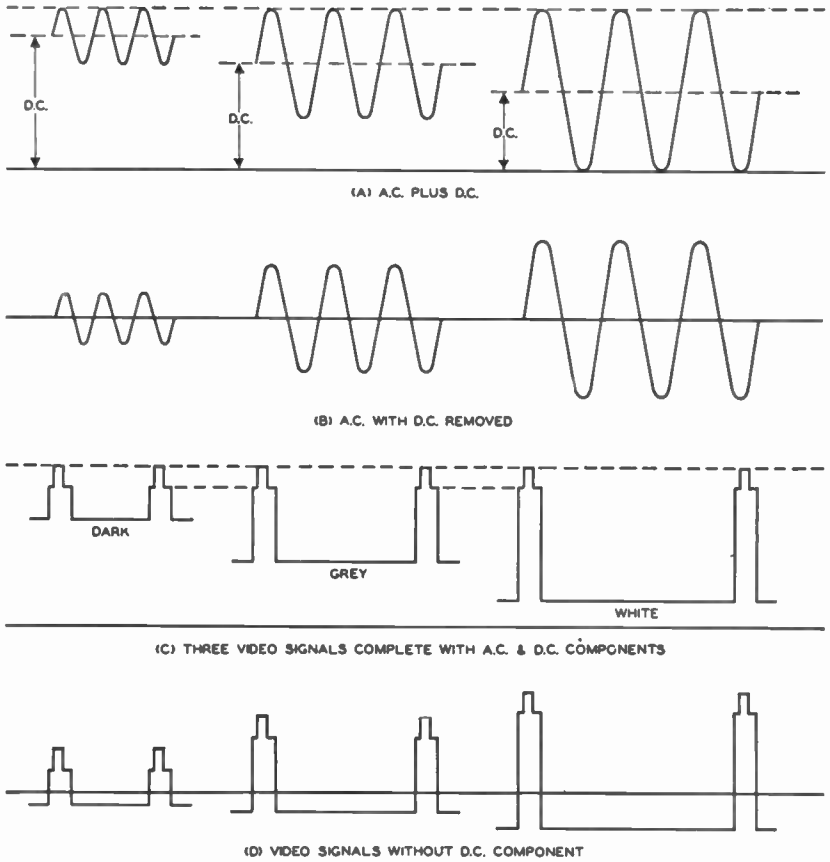


Fig. 8.3. An illustration of the effect of removing the d-c component from the video signal.

level. However, when the signal is passed through $R-C$ coupled video-frequency amplifiers, the blanking pulses of the various lines are no longer lined up because the coupling capacitors cause the video signal to possess equal positive and negative areas about the zero axis.

This situation has been encountered by the serviceman, although in slightly different form. Suppose we take three 60-cycle a-c voltages and three d-c voltages and combine them to form the signals indicated in Fig. 8-3A. (Voltages of this type are frequently found in power supplies

where the a-c wave represents the ripple.) For the sake of this discussion, we have provided enough d-c voltage so that the positive peaks of all three waves reach the same level. Now, let us pass these voltages through a capacitor. The result is shown in Fig. 8-3B. By removing the d-c voltages, each wave has as much area above the axis as below and, because of this, the positive peaks of the waves are no longer at the same level.

Let us look at the equivalent situation in a television system. In Fig. 8-3C there is shown three video signals taken at different moments from a television broadcast and representing three lines. One line is essentially white, one is grey, and one is dark or black. As they come out of the video second detector, all the blanking voltages are aligned to the same level. After passing these three signals through a coupling capacitor, the signals possess the form indicated in Fig. 8-3D. For each signal, the area above the axis is equal to the area below the axis. But because of this, the blanking voltages of the signals are no longer at the same level. In this condition, we say that the d-c component of the video signal is missing. The question now is: What effect will this variation in blanking level have on the image produced on the screen?

Each blanking pulse represents the dark level of each line. Since all lines in an image should have the same reference (or black) level, all blanking pulses should have the same voltage value. This was true of the video signal when it left the transmitter and it was true in the receiver just before we passed the detected video signal through a coupling capacitor in the video-frequency amplifier system. After passage through this capacitor, the blanking pulse levels were no longer aligned to the same level. If now we apply the three signals to an image tube (reverse polarity of Fig. 8-3D), here is what happens.

When the signal corresponding to a white line reaches the cathode-ray tube grid, we manually adjust the brightness control (which controls the bias for the image tube) to the point where the blanking pulse level just drives the tube into cut-off. Thus, as long as this signal remains, the negative voltage of the blanking pulse, added to the negative bias set by the brightness control, will just darken the screen at the blanking pulse level.

If now the grey video signal comes to the cathode-ray tube, we see that its blanking pulse level is less negative than the blanking pulse level of the previous video signal. Hence, here, the beam will not cut-off at the blanking pulse, and the beam retrace will be visible. We could produce the proper cut-off conditions by increasing the negative bias on the image tube, but this is impractical for several reasons. First, in any changing scene, the background shading changes too rapidly to be manually adjusted by the viewer. As a result, if the brightness control is set for a very bright picture, we will see the retrace lines when a darker picture arrives. Conversely, if the bright-

ness control is set for a darker image, then, when a lighter image is viewed, part of the detail will be lost because of the greater cathode-ray tube grid bias.

Finally, the situation is even more aggravated when a dark video signal arrives. Now, we require an even greater negative bias and, when the brightness control is set correctly for this signal, it is much too negative for any of the two previous signals. If either of these two other signals is viewed with the bias set for this last signal, the image will appear too dark. On the other hand, when it is correctly set for a white picture, a black picture will appear too light, with even the retrace lines visible. The only solution to this state of affairs is to return all blanking voltages to the same level again, just as we found it in the incoming signal. *This, then, is the function of the d-c restorer in the receiver.*

Every cathode-ray tube has a definite characteristic curve. For a certain input voltage, a definite amount of light appears on the screen. All blanking pulses are purposely placed on the same level in order that the cathode-ray tube will react to them in the same manner throughout the entire reception of the signal. The same is true of white, grey, black, or any other shade that is transmitted for the scene. Any one shade must produce the same illumination on the cathode-ray screen each time its corresponding voltage is present on the tube's control grid. However, this cannot occur unless all video signals have the same reference level. It is here that the usefulness of the d-c component becomes apparent. Through the use of this inserted voltage, all blanking and synchronizing pulses are leveled off and the image detail attached to these pulses is likewise correctly oriented.

To operate the television receiver properly, then, some method must be devised whereby the a-c video signals which appear at the cathode-ray tube are again brought to the same relative level that they had before the removal of the d-c component in the intervening video amplifiers. The problem resolves itself into one of reinserting a d-c voltage that will take the place of the one removed. Note that in direct-coupled video amplifiers there are no coupling capacitors and hence the d-c is not removed. Consequently there is no need here for special d-c reinsertion networks.

Reinserting the D-c Components. To understand why d-c restoration is possible, it is necessary to know that removing the d-c component from a video signal does not change its shape but merely its reference level. This is evident when Figs. 8-2A and B are compared. The same variations in the a-c components still occur and the relationship of the a-c signal to the blanking and synchronizing pulses remains the same, with or without the d-c component. It is also seen that the brighter the line, the greater the separation between the picture information variations and the pulses. As the scene becomes darker, these two components move closer together.

It is from these relationships that we are able to reinsert the d-c component; for, if we could develop a variable bias that would effect each change in blanking and synchronizing pulse voltage and act in such a manner that all pulses would be brought to one common level, our purpose would have been achieved. It would mean, for example, that if a video signal in its a-c form were applied to the input of a tube where the process of d-c restoration was to occur, a variable grid bias, developed here, would return them to the same level again in the tube's plate circuit. The bias would automatically adjust

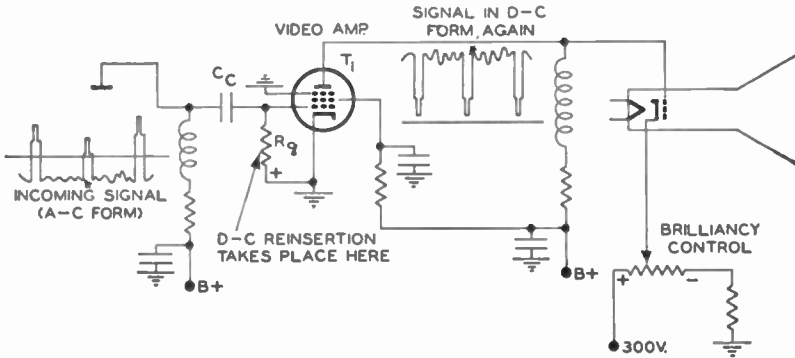


FIG. 8-4. A simple d-c reinsertion circuit. The bias developed across R_g varies with the incoming signal.

itself to suit each individual case. Then, with all the signals lined up again, they could be applied to the cathode-ray tube.

D-c Reinsertion Circuits.* There are several methods whereby the d-c component may be reinserted into the video signal. Perhaps the simplest is the circuit diagram of Fig. 8-4. Here the final video amplifier is operating at zero fixed bias, with no signal applied to the grid. As soon as a signal does arrive, grid current flows, the amount dependent upon the strength of the signal voltage. Thus, one of the conditions specified above, namely the signal determining its own bias, is obtained.

The form of the a-c signal applied to the grid of this last video amplifier tube must be of a negative phase, as shown in Fig. 8-4. It must be negative at the input in order that the proper positive phase will be obtained at the output where it is applied to the grid of the picture tube. Because there is no fixed bias on T_1 , the grid will swing positive whenever the a-c signal is positive (above the zero line). Making the grid positive causes electrons to flow in this circuit, charging the capacitor C_c . C_c , in turn, discharges through R_g .

* The name "d-c reinsertion" circuit is common throughout the television field. However, sometimes the name of "clamping" circuit is also heard. Both refer to the same thing and may be used interchangeably.

The electrons, in passing through R_g , will develop a voltage, the amount dependent on how positive the grid is driven by the signal. This voltage across R_g is the operating grid bias and, in effect, acts in series with a-c signal applied to the tube. Since the current flowing in the grid resistor will depend on the extent the applied a-c signal goes positive, it is evident that the grid current will vary from one pulse to another. A large positive voltage (corresponding to a bright line) will cause a large current to flow through R_g and hence a large biasing voltage will develop here and will be applied in series with the signal. For a small positive pulse, such as is obtained for a dark line, only a small biasing voltage will appear across R_g .

Now let us see how this variable bias brings each synchronizing pulse to the same level. Consider first a pulse of small amplitude. The pulse extends a small distance above the zero or reference line and, under the influence of this signal, the grid will go slightly positive. Consequently, only a small negative biasing voltage will develop across R_g . Let us assume that for the blanking level of the signal 5 ma of plate current will flow.

Now a large pulse, due perhaps to a bright line, arrives at the grid. Since its level is much farther above the zero line, the grid will be driven more positive and a greater grid current will flow. The result is a larger negative bias across R_g to counteract the increased positive value of the signal. At the blanking level of this signal, 5 ma of plate current should also flow. With the same amount of plate current flow for each blanking pulse, the output signals are all lined up again as indicated in Fig. 8-4. One further point should be kept in mind throughout this entire process. The bias developed across R_g does not vary each instant, but remains relatively constant from one pulse to another. Since the horizontal pulses are separated from one another by the camera signal, the bias will be constant for this detail. Thus, the detail of the picture is not smoothed out, as it would be if the bias on R_g changed with every single current variation. It merely changes at each pulse.

The time constant of the grid resistor R_g and the grid capacitor C_c must be long enough so that the bias developed will last for at least one complete horizontal line, or from pulse to pulse. In practice, however, it may last longer, perhaps for several lines, since the average brightness of the background illumination seldom changes that rapidly.

Values of R_g range from about 400,000 ohms up to one megohm. C_c would be chosen so that the time constant ($T = R \times C$) is equal to the duration of one or more lines. Each line lasts for approximately 1/16,000 of a second.

The output of T_1 , shown in Fig. 8-4, is applied directly to the control grid of the cathode-ray tube. Direct coupling is necessary since a capacitor would remove the d-c component just inserted. At the grid of the cathode-

ray tube, a fixed bias between the grid and cathode is obtained from the power supply. This bias sets the operating point for the tube and, in conjunction with the video blanking and synchronizing pulses, cuts off the electron beam at the proper moments. The setting of this bias will depend

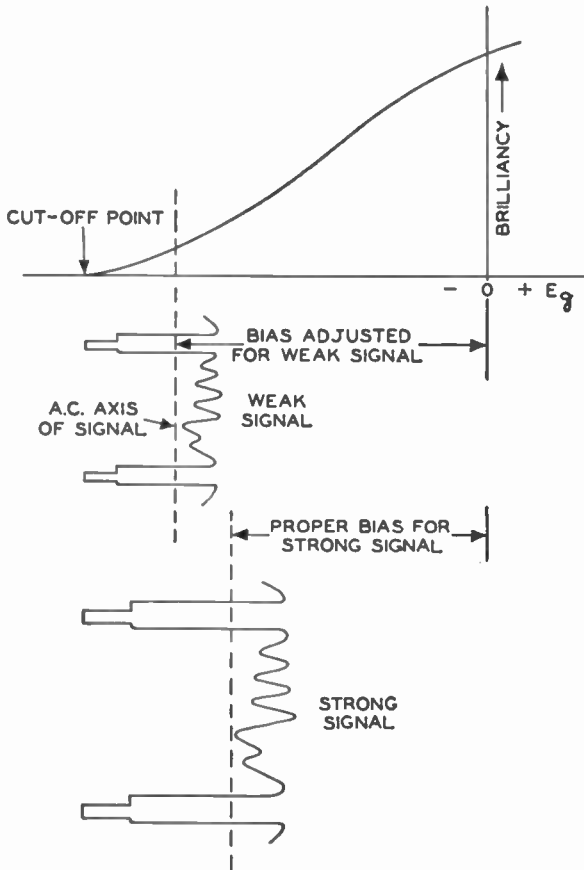


Fig. 8.5. Illustrating why the brilliance control must be adjusted to suit the incoming signal.

upon the strength of the signal reaching the grid. A signal of small amplitude, say from some distant station, requires more fixed negative bias on the grid than a stronger signal.

The dependency of the cathode-ray tube grid bias on the strength of the arriving signal is illustrated in Fig. 8-5. For a weak signal, the bias must be advanced to the point where the combination of the relatively negative blanking voltage plus the tube bias drives the tube into cut-off. However, with a

strong signal, the negative grid bias must be reduced; otherwise some of the picture detail is lost.

The Brilliancy Control. Since the bias of the cathode-ray tube may require adjustment for different stations, or even for various conditions on the same station, a potentiometer is connected into the bias circuit, brought out to the front panel, and called the brilliancy or brightness control. By its use, the observer is able to adjust the bias on the grid of the picture tube in order that blanking pulses just drive the grid to cut-off and so that no retrace is visible on the screen.

The effects of the brightness control and the contrast control previously described overlap to some extent. If the setting of the contrast control is increased so that the video signal becomes stronger, the brightness control must be adjusted to meet the new condition, which means, of course, that no retraces are visible. Too small a value of negative grid bias allows the average illumination of the scene to increase and permits some of the return traces to become visible. In addition, the image assumes a thin, watery, washed-out appearance. Too low a setting of the brightness control, which will result in a high negative bias on the picture tube grid, will cause some of the darker portions of the image to be eliminated, and the average illumination of the scene will decrease. To correct this latter condition, either the brilliancy control may be adjusted or the contrast control setting can be advanced until the correct position is obtained. Finally, the focusing action of the tube may also be affected by either the brightness control or the contrast control and require a slight adjustment, too, although normally this will not occur.

D-c Reinsertion with a Diode. The foregoing method of d-c reinsertion is attained simply and produces good results. One limitation of this method, however, is the fact that the B+ potentials on the screen grid and plate must be reduced in order that excessive current does not flow when no signal is being received. This reduces the overall gain available from this stage. Second, it is quite important that the screen-grid voltage be well regulated, otherwise the d-c restoring action is impaired.

Another method exists which requires the addition of a diode tube, but which removes the highly positive voltage from the control grid of the cathode-ray tube (see Fig. 8-6). The signal here is in its a-c form until it reaches the input to the d-c restorer, composed of capacitor C_1 , resistor R , and the diode tube. The form of the signal, at this point, is the positive phase since no further reversals take place before the grid of the cathode-ray tube is reached.

In the signal applied to the restorer, the blanking and synchronizing pulses are below the zero line. When applied to points 1 and 2, the signal will cause point 1 to become negative with respect to 2. This follows from the action of an a-c wave. The other portion of the signal, which contains the image

information, is above the line and, when it is applied across points 1 and 2, will make 1 positive with respect to 2. The diode in the circuit conducts only when its plate is positive with respect to its cathode, or when point 2 is positive with respect to point 1.

The action of the d-c restorer is simple. When the polarity of the video signal at point 1 is negative, point 2 and the plate of the diode are positive. A flow of current will occur through the tube, and capacitor C_1 will charge to a value dependent upon the strength of the signal acting at points 1 and 2. The polarity of the charge is indicated in Fig. 8-6. During the positive por-

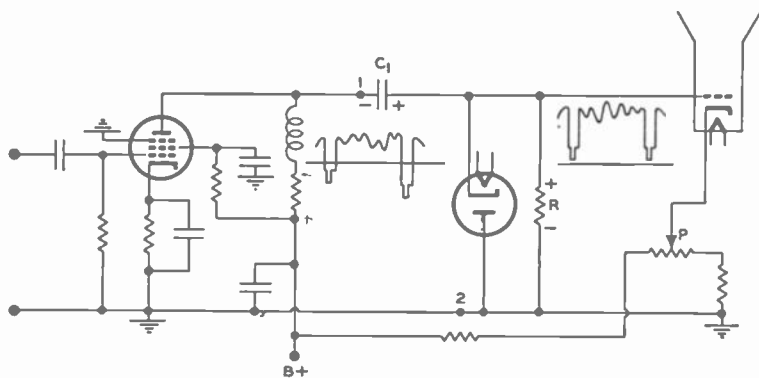


FIG. 8.6. D-c reinsertion with a diode.

tions of the video signal at the input of the circuit, capacitor C_1 will discharge through R , since the diode plate is now negative, and the tube is non-conducting. The value of R is high, about 1 megohm, and C_1 will discharge slowly.

The values of C_1 and R are so designed that the voltage on the capacitor remains fairly constant throughout an entire horizontal line, or during the time that the positive a-c signal is acting on the picture tube grid. Note that this charge is between the grid and ground, or cathode, and hence acts as a variable bias in series with the a-c signal. When the negative portion of the signal (which is due mostly to the blanking and synchronizing pulses) acts at the input, the plate of the diode again becomes conductive. The charge on C_1 will now be automatically adjusted to the amplitude of the negative pulse. A bright line will place a larger positive voltage on the capacitor C_1 than a darker line (positive picture phase here). The positive voltage will cause the grid to become more positive and the line will receive its correct value. The bias will raise each line until the blanking pulses are lined up again. Thus, in this instance, we have a bias developed which is proportional to the impulse amplitudes, which are in turn governed by the average brightness of the line, as previously explained. Potentiometer P is avail-

able, and its adjustment will cause the grid to cut off on the application of all blanking pulse.

A slightly modified version of the foregoing circuit is used in many television sets. The circuit, given in Fig. 8-7, reveals that the diode tube is not placed across the entire plate output of T_1 , but merely across a portion of it, obtained from resistor R_3 . The action of the a-c video signal across C_5 , R_5 and the diode results in C_5 charging to the peak value of the pulses. It then discharges partially through the 1-megohm resistor, and the effect of the capacitor charge is to place its stored voltage in series with the a-c video signal so that the necessary d-c component is reinserted into the signal.

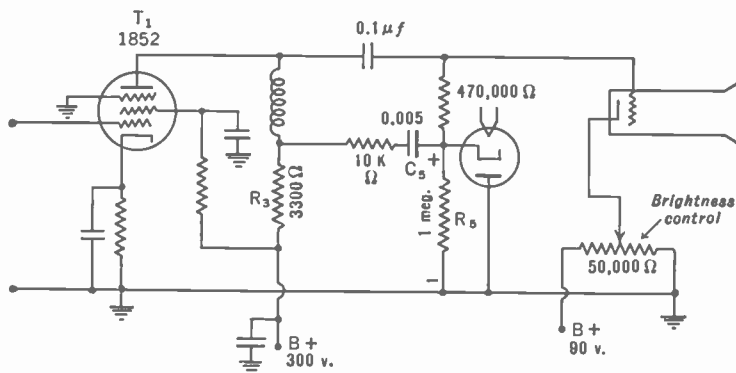


FIG. 8.7. A d-c reinsertion circuit that has been employed in commercial receivers.

The amount of charge on the capacitor will naturally vary from line to line.

In some sets, tubes other than diodes are used, but in most cases, the tube elements are tied together in such a manner as to form the equivalent of a diode. Instead of using a diode tube for the d-c reinsertion, a 1N34 crystal may be substituted. Since crystals and diodes function in the same manner, the entire previous discussion can be applied here.

Diode restorers also can be utilized to function as clippers to separate the sync pulses from the rest of the video signal and then to transfer these pulses to the horizontal and vertical sweep systems. A typical circuit is shown in Fig. 8-8. Between every two horizontal sync pulses, C_1 discharges slightly through R_1 , setting up the d-c restoration voltage here. Upon the arrival of a sync pulse, a short flow of current takes place through the tube and R_2 since this latter resistor is in series with the tube. These pips of current represent the sync pulses since they appear only when the peak pulses are present across the circuit.

There are variations of the foregoing diode restorer and clipper circuit which accomplish the same job but function slightly differently. Consider, for example, the circuit shown in Fig. 8-9.

V_2 , the diode d-c restorer, receives the video signal from the plate V_1 through R_1 and C_1 . Because of the way in which it is connected, V_2 will conduct only for the negative sync pulses of the video signal, charging C_1 in the manner already described. The d-c restorer biasing voltage is then de-

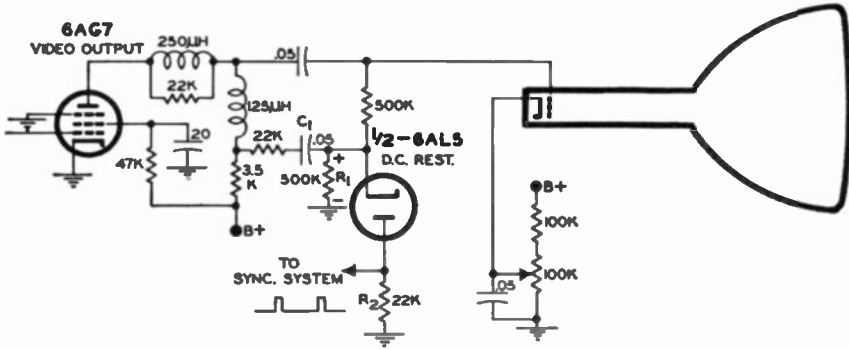


FIG. 8.8. A diode d-c restorer serving also as a sync clipper.

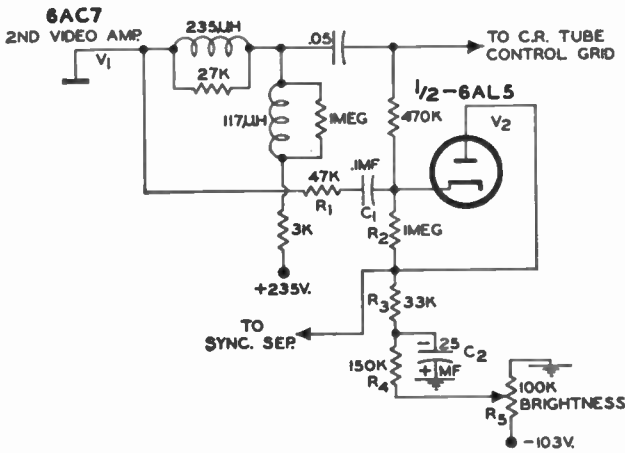


FIG. 8.9. Another combined d-c restorer and sync clipper.

veloped across R_2 and, by varying the bias of the cathode-ray tube, reinserts the absent d-c component. Thus far the circuit is concerned with the d-c restoration.

The sync voltage is the voltage developed across R_3 since R_4 and R_5 are by-passed by C_2 . When the positive half of the video signal is applied to V_2 , this tube does not conduct since its plate is negative with respect to its cathode. This positive voltage, which represents only the picture information and not the sync pulses, is divided between R_1 (47,000 ohms), R_2 (1 meg-

ohm), and R_3 (33,000 ohms). The ratio of the voltage across R_3 to the total applied video signal is:

$$\frac{33,000}{1,080,000} \quad \text{or} \quad \frac{33}{1,080}$$

This ratio is very small; hence not much voltage is fed to the sweep system.

On the negative half of the video signal, when the sync pulses are present, V_2 conducts, shunting out R_2 . Now, the total applied video voltage is divided between R_1 and R_3 , and the sweep system receives $3\frac{3}{8}\%$ of the applied voltage. This means that approximately 14 times more sync voltage is transferred to the sweep system than picture voltage. Because of the presence of some picture voltage, further separation is usually found in these systems.

Triodes and even pentodes may be connected to serve the dual function of d-c restorers and partial or total clippers. More will be given on this subject when the section concerned with clippers is reached.

Additional Notes on D-c Restoration. Economics is a strong determining factor in the design of commercial television receivers and if it is possible to reduce the cost of a set without compromising picture quality too much, this sacrifice is frequently made. There are a number of receivers on the market today that do not employ d-c restoration, nor do they possess a d-c path between the video second detector and the picture tube. In other words, the d-c component is removed from the signal and never reinserted.

As the preceding discussion has indicated, loss of the d-c component will tend to make the overall picture darker. To counteract this, the viewer generally turns up his brightness control. This, in turn, frequently causes the vertical retrace lines to become visible. The continued presence of these lines during normal broadcasts will prove to be annoying. To rid the screen of these retrace lines, it has become standard practice to apply a negative pulse to the grid of the picture during the vertical retrace interval. (A positive pulse fed to the picture tube cathode will achieve the same results. Generally the pulse is applied to the element not receiving the video signal.) The pulse biases the tube to cut-off, prevents electrons from passing through the tube, and effectively removes the vertical retrace lines for any normal position of the brightness and contrast controls.*

It is true that removal of the d-c component will reduce the contrast range of the image. However, this has been partially offset by the development of screen phosphors possessing wider contrast ranges, and it is doubtful whether any viewer can tell the difference when the d-c component is missing.

* More information concerning the origin of these pulses will be given at a subsequent point.

QUESTIONS

1. What is meant by d-c reinsertion? Why is it necessary?
2. Explain the difference between the a-c and the d-c components of a video signal.
3. How is the d-c component removed? Why is it possible to reinsert this voltage?
4. Must a television receiver contain a d-c restorer? Explain.
5. Explain the operation of the grid-leak bias method of d-c reinsertion.
6. What is the brilliancy or brightness control? Where is it situated in the circuit?
7. Explain the need for the brightness control.
8. What is the difference between the brightness and contrast controls?
9. What would happen to the image if the grid-leak d-c reinsertion network has a time constant of 1 microsecond?
10. Draw the schematic circuit of a diode d-c reinsertion network.
11. Explain how the circuit drawn in the previous question operates.
12. Explain what happens when the d-c component of a video signal is removed.

Chapter 9

CATHODE-RAY TUBES

Introduction. The cathode-ray tube, which is the very heart of the television receiver, is in many respects quite similar to the ordinary receiving tube. Like the receiving tube, it, too, has a cathode that emits electrons because of heat received from the heater wires. The flow of electrons, and hence the number, is regulated by the voltage on the electrodes in the same manner as any other tube. Once past the control grid, however, the electrons are narrowed down to a fine beam and subjected to focusing and deflecting forces until the beam strikes a fluorescent screen located at the far end of the tube. At each point where the electron beam impinges on the screen, a spot of light appears. If the points follow in rapid succession, the motion can be made to appear continuous due to the persistence of vision phenomenon of the human eye.

For television receivers, the video signal containing the image detail is applied to the control grid or cathode, while the synchronizing impulses control saw-tooth oscillators that connect to deflecting plates or coils. Under the influence of these changing voltages, the beam is swept across the screen in step with the scanning beam in the camera tube at the studio. In the smaller cathode-ray tubes, focusing and deflection of the electron beam are accomplished by electrostatic means. For the larger tubes, say those that are 10 and 12 inches or more in diameter, electromagnetic deflection coils are found. Focusing may be electromagnetic or electrostatic.

The formation of the electron beam starts naturally at the cathode. The emitting surface, composed of thoriated tungsten or barium and strontium oxides, is restricted to a small area in order that the emitted electrons progress only toward the fluorescent screen. They would serve no useful purpose in any other direction. The emitting material is thus deposited on the end of the nickel cathode cap that encloses the heater in the manner shown in Fig. 9-1 for a typical construction. The electrons, after emission, are drawn by the positive anode voltages into electric or magnetic lens systems. These

form and focus the electrons into a sharp, narrow beam that finally impinges on the fluorescent screen in a small round point.

The use of the word "lens" may puzzle the reader who thinks of this term only in connection with light rays, not electron beams. The purpose of a glass lens is to cause light rays either to diverge or to converge to a point. Electronically, the same results can be achieved and hence the reason for the carry-over of the name.

The First Lens System. In the first lens we find the cathode, the control grid, and the first anode arranged in the manner shown in Fig. 9-2. The grid, it is noticed, is not the familiar mesh wire arrangement found in ordinary tubes. For the present purpose it is a small hollow cylinder. The end nearest the cathode is partially closed by a round baffle, with only a small pinhole through which the electrons may pass. This restricts the area of the cathode that is effective in adding electrons to the beam and aids in giving the beam sharpness. The other end of the grid cylinder is open and leads to the first anode. Here again a baffle permits only those electrons near the axis of the tube to pass through.

Because of the energy imparted to them by the cathode heating, the electrons leave the cathode surface with some small velocity. With no positive electric force (or field) to urge them forward, the electrons would tend to congregate in the vacuum space just beyond the cathode and form a space

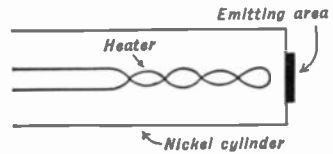


FIG. 9.1. Cathode and heater construction for a cathode-ray television tube.

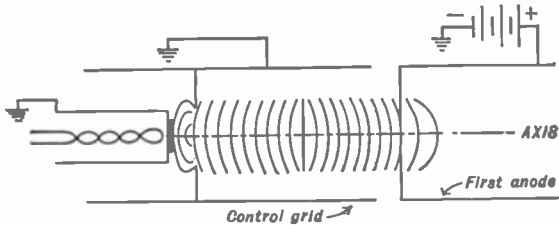


FIG. 9.2. The first lens system of a television viewing tube.

charge. Eventually, just as many electrons leave the heated cathode surface as are repelled by the negative space charge, and a state of equilibrium exists. This condition can be overcome and a flow of electrons allowed to take place down the tube if a high positive voltage is placed on the first anode.

The first anode, which is a hollow cylinder, does not have its electric field contained merely within itself; it also reaches into the surrounding regions. To be sure, the farther away we get from the anode, the weaker the strength of the field. With zero and low negative potentials on the control grid, the

influence of the positive anode field extends through the baffle of the control grid right to the cathode surface. Electrons leaving this surface are urged on by the positive electric field and accelerated down the tube, with the baffle restricting the direction of the electrons to very small angles with the axis of the tube.

The distribution of the electric equipotential lines outside the first anode is shown in Fig. 9-2. In the diagram, the lines are drawn through points that have the same electric potential, as determined by actual measurements with a probe throughout the region. It is interesting to note that these lines are not straight, but tend to curve, the amount of curvature being influenced by

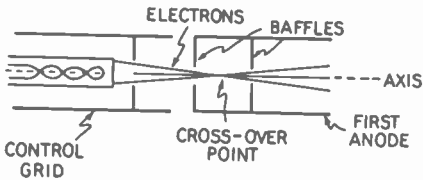


FIG. 9.3. A simplified diagram of the cross-over point in the first lens system. The subsequent electronic lens system is designed with this cross-over area serving as the starting point.

the distance from the first anode and the control grid. Cathode-ray tube design engineers use such field distribution diagrams to determine the effect of each electrode on the electrons at the cathode and in the beam.

As a result of the bending of the electric field at the cathode, it can be proved by means of vectors that all electrons passing through the small hole in the control grid baffle will come to a focus or converge to a small area located just inside the first anode. This region is on the axis of the tube and is known as the cross-over point. The effect of the electric field is such that electrons near the outer edges of the control grid opening travel at an angle in order to get to the cross-over point, whereas electrons on the axis of the lens move straight forward to this point. The direction of some of the electrons is shown in Fig. 9-3.

It is well to keep in mind that the shape of the electric field is determined by the placement of the electrodes and the voltages applied to them. The electrons are forced to converge to the cross-over point because this point can more readily serve as the supply source of the beam electrons than the cathode from which they initially come. The area of the cross-over point is more clearly defined than the relatively larger cathode surface, and it has been found that the electron beam is easier to focus if the cross-over area is considered as the starting point, rather than the cathode itself. The electrons that compose the final beam are then drawn from the cross-over point while other electrons come from the cathode to take their place. The greater the number of electrons drawn from this point, the brighter the final image on the fluorescent screen.

The control grid, in Fig. 9-3, is at zero potential. For ordinary purposes, this value of grid bias would permit too many electrons to pass into the beam.

Hence a negative bias is placed on the grid. In the larger cathode-ray tubes, the bias may rise as high as -60 volts. With a negative voltage on the control grid, the extent of the positive electric field is modified and it no longer affects as large an area at the cathode surface as it did previously with zero grid volts. Now, only electrons located near the very center of the cathode are subject to the positive urging force, and the number of electrons arriving at the cross-over point is correspondingly less. The intensity of the final electron beam likewise decreases. In the television receiver, the video signal is applied to the control grid and the resulting variations in potential cause similar changes in electron beam intensity.

For the beam arriving at the screen to remain in focus once the controls have been set, the position of the cross-over point must remain fixed. With normal variations of control grid voltage, this condition is obtained. With large variations, however, the position of the cross-over point tends to change, moving closer to the cathode as the grid becomes more negative. Thus a certain amount of defocusing will take place. Proper design generally keeps this at a minimum and, for most of the voltage variations encountered in television work, defocusing does not become too noticeable.

To summarize the purpose of the first lens system, we see that electrons leaving the cathode surface are forced to converge to a small area near the anode. This offers a better point for the formation of the beam and its subsequent focusing.

The Second Lens System. The second lens system draws electrons from the cross-over point and brings them to a focus at the viewing screen. The system consists of the first and second anodes, as shown in Fig. 9-4. The second anode is operated at a higher potential than the

first anode, is larger in diameter, and frequently overlaps the first anode to some extent. It is at the point of overlap of the two anodes that the second lens is effective, and it is here that the focusing action of the electron beam takes place. Electrons, when drawn from the cross-over point established by the first lens system, are not all parallel to the axis of the tube. Some leave at various small angles, as shown in Fig. 9-3. The beam thus tends to diverge and it is due to the second lens that these diverging electrons alter their path and meet at another point on the axis. This second point is at the screen. Those electrons moving straight along the axis of the tube are not affected, in direction, by the focusing action of the second lens.

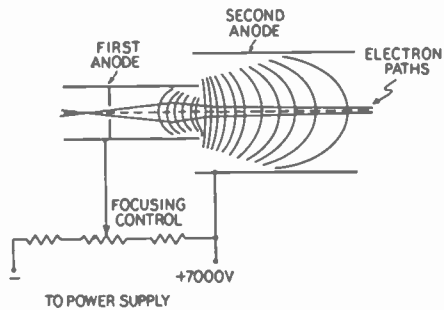


FIG. 9.4. The second lens system. Focusing of the electron beam at the viewing screen is accomplished by varying the voltage at the first anode.

The operation of the second lens depends upon the different potentials that are applied to the first and second anodes and the distribution of the resulting electric field. The equipotential lines for this lens are drawn in Fig. 9-4, and it is to be noted that the curvature of these lines changes at the intersection of the two anodes. On the left-hand side, the electric field lines are convex to the approaching electron beam, while to the right of the intersection the lines are concave. Without resorting to mathematical reasoning, it can be stated that the effect of these oppositely shaped electric lines on the beam is likewise opposite. Since we have seen that some of the electrons tend to diverge after they leave the cross-over point, the field distribution

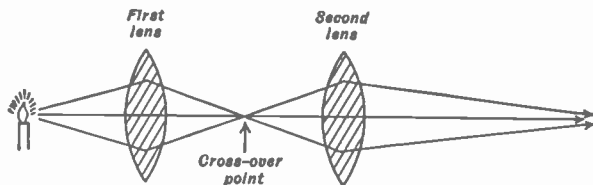


FIG. 9.5. The glass lenses used in focusing light rays illustrate the similarity between light wave and electron beam focusing.

must be designed to overcome such a tendency. In action, the convex equipotential lines force the electrons to converge to a greater extent than the concave lines cause the electrons to diverge. Inasmuch as the convergence exceeds the divergence, the net result is a focusing of the electrons on the screen.

The ratio of the voltages, the size of the anode cylinders, and their relation to each other will determine the distribution and curvature of the electric lines of force; the latter, in turn, will determine the amount and the point at which the focusing takes place. In current cathode-ray tubes, the ratio of the first to the second anode voltages ranges from 3 to 1 to 6 to 1 or more.

In order that the electron beam leaving the cross-over point shall not diverge too greatly, a baffle is placed just beyond this point, similar in construction to the baffle previously described for the control grid. The baffle again limits the width of the electron beam to the desired size. Practically, focusing control is accomplished by varying the voltage on the first anode by an arrangement shown in Fig. 9-4. This is the simplest way of altering the voltage ratio between the first and second anodes and, with it, the distribution of the electric lines of force of the lens system. An approximate optical analogy of the lens system is shown in Fig. 9-5 and may prove helpful in indicating the operation of the electric system.

Electrostatic Deflection. Once past the second anode, the electron beam speeds toward the fluorescent screen. However, the beam must first pass through two sets of deflecting plates mounted at right angles to each

other. One set of plates is known as the horizontal deflecting plates, the other as the vertical deflecting plates. In television receivers, saw-tooth oscillators are electrically connected to these plates, and the electron beam is subjected to changing voltages that force it to move across and up (or down) on the screen. The 525 lines of each frame are swept out in this manner. The synchronizing pulses control the action of the saw-tooth oscillators in order to keep the original and reproduced images in step with each other. The same action of deflection can also be accomplished with coils and will be presently described.

In commercial television receivers, two methods of applying the deflecting voltages to the deflecting plates may be employed. In one method, we find that the one deflecting plate of each set is connected directly to the second anode while the other plate receives the varying deflecting voltages. When this voltage is zero, both plates (of each set) are at the same potential, which in this case is equal to the second anode voltage. The electron beam is therefore not subjected to any deflecting force and passes unmolested through the center of the system. The reason that the deflecting plates are at second anode potential is due to their position in the tube. The beam, in leaving the second anode, must not be slowed down. Any voltage difference between the second anode and the deflecting plates would not only change the velocity of the beam but would also produce defocusing at the screen. To avoid all these difficulties, the deflecting plates and the second anode are electrically connected.

With one plate permanently attached to the second anode and the other to the deflecting voltage, we have an unbalanced arrangement as in Fig. 9-6A. The plate with the varying potential will deflect the beam, an amount proportional to the acting voltage. For small voltage variations, the operation is satisfactory. At higher deflecting voltages (in larger tubes), a defocusing of the beam occurs. This effect is sometimes called astigmatism.

The reason for the defocusing action may be explained as follows: The electron beam, once it leaves the second anode, is travelling toward the screen where it should come to a focus. Any electrical disturbances not part of the focusing action will tend to destroy the sequence. When varying voltages are applied to the deflecting plates, their average potential varies above and below the second anode voltage. The result, when the difference between the two becomes sufficiently great, is to throw the beam out of focus. In small tubes the voltage difference is not large enough to cause defocusing, but in larger tubes definite defocusing is observed. This is especially noticeable when the beam is at the ends of the screen, for it is at such points that the required deflection voltage is greatest.

The more desirable method, which is balanced, is shown in Fig. 9-6B. The voltages on both plates change and uniform force is exerted by the plates

on the beam. As both deflection plate voltages vary in opposite manner, the average *change* in voltage is zero. However, now two tubes are required (push-pull), which increases the cost of the set. Despite the latter fact, this method is the one used in receivers that employ electrostatic deflection.

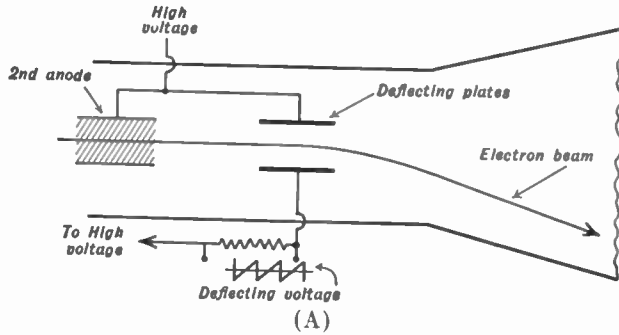


FIG. 9.6A. Deflection of electron beams. This unbalanced method, while simple in construction, does not always give a clearly defined beam on all parts of the fluorescent screen.

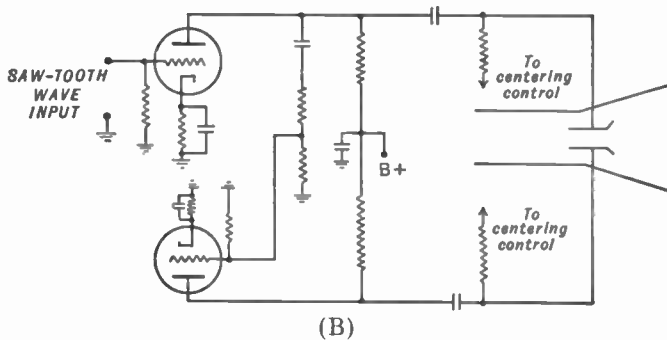
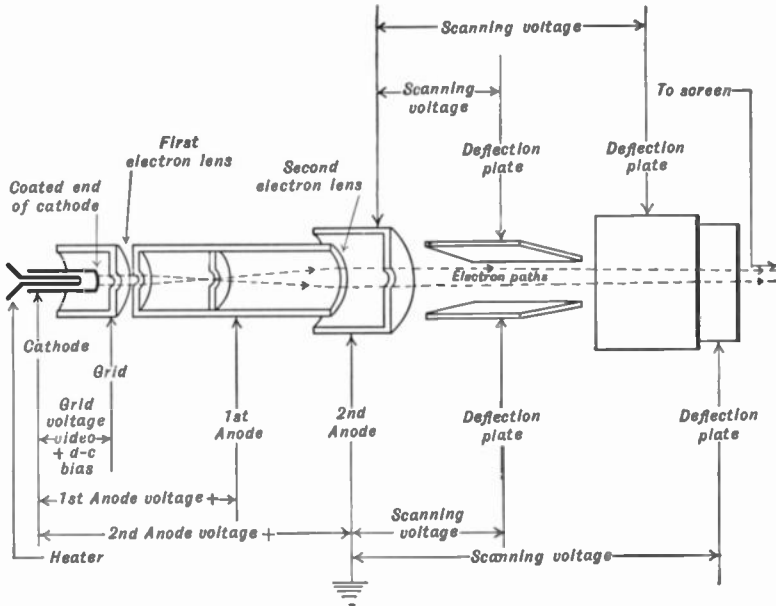


FIG. 9.6B. A balanced deflection system.

To obtain maximum deflection of the beam without having it strike the edges of the plates, cathode-ray tube designers have flared the plates slightly at the ends. The basic elements of a cathode-ray tube employing electrostatic deflection and focusing are shown in Fig. 9-7A. An example of this type of tube is the 5JP4. Seven- and 10-inch electrostatic deflection tubes generally incorporate one additional element not present in Fig. 9-7A. The new element is placed between the control grid and the focusing (first) anode. Internally, it connects to the accelerating (or second) anode. See Fig. 9-7C. The introduction of this new element serves to reduce further interaction between beam focus and voltage variations at the control grid.

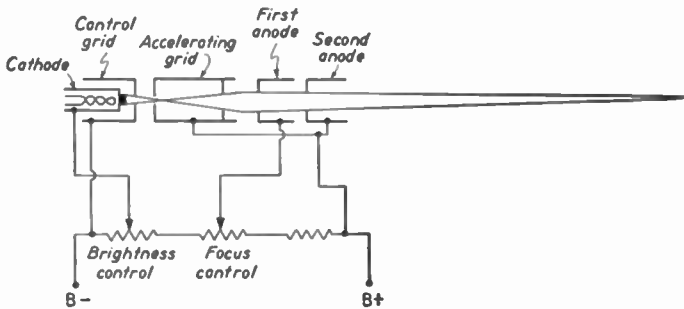


(A)



RMA BASING

(B)



(C)

Courtesy Sylvania Electric Products Corp.

FIG. 9.7. The basic elements of a cathode-ray tube using electrostatic focusing and deflection (A). The great number of elements found in these tubes necessitate a larger tube base, as shown in (B). Modified construction of electron gun in 7- and 10-inch electrostatic deflection tubes (C).

As the screen area becomes larger, beam defocusing defects become more apparent and greater precautions must be taken to prevent them.

Cathode-ray tubes also possess a conducting aquadag coating entirely around the inside of the glass, extending from the first or second anodes almost (but not quite) to the fluorescent screen. An important function of this aquadag coating is to prevent the collection of free electrons that would otherwise accumulate on the glass walls of the tube. In addition, it serves as a collecting anode for the secondary electrons which are emitted from the fluorescent screen when the electron beam impinges on the screen. If we removed the aquadag coating, these secondary electrons would land on the glass walls of the tube and remain there. Eventually, sufficient charge would develop to prevent proper functioning of the beam. By coating the inside of the tube with aquadag and giving it a positive potential, we remove all this undesirable charge. Internally, the coating is connected to the second anode. Furthermore, through the use of this dark coating, less light is reflected from the screen into the tube and then back to the screen again. This reduction in reflection aids the contrast range available from a fluorescent screen.

Amount of Deflecting Voltages Necessary. The amount of voltage that must be applied to the deflection plates in order to force the beam to travel from one side of the screen to the other will depend to a great extent upon the accelerating voltage of the second anode and the width of the screen. A larger voltage will accelerate the electrons more strongly and hence necessitate a large voltage on the deflection plates. The faster an electron travels, the less time it spends between the plates, and the greater the deflection voltage required to bend it. The deflection factor, which is given for each electrostatic tube in its characteristic data, then increases. On the other hand, lowering the second anode voltage will permit the deflecting plates to exert greater control over the electron beam for the same deflecting voltage.

In characteristic charts of cathode-ray tubes, the deflection factor may be stated directly or another unit, known as the deflection sensitivity, may be given. Many times both can be found, although it is possible to compute one if the other is given. To illustrate, consider the 7JP4 tube. The deflection factor is given as 31 to 41 d-c volts/inch/kv. This means that, with 1,000 volts on the second anode (1 kilovolt or 1 kv), 31 to 41 d-c volts' difference between a set of deflection plates will move the beam 1 inch on the screen. The range of 31 to 41 volts is given because of manufacturing tolerances; 36 volts could be used as an average value. If 2,000 volts are placed on the second anode, the deflection force becomes less effective, and now we require 62 to 82 d-c volts to produce the same 1 inch of deflection. In general, increasing the second anode voltage by a certain amount de-

creases the distance the beam is deflected by a proportionate amount (assuming no voltage change on the deflection plates). To overcome this decrease, we must increase the deflection voltage accordingly.

The deflection factors of the vertical and horizontal plates need not necessarily be the same. In the 7JP4, for example, the foregoing figures (31 to 41 volts) represent the deflection factor of the vertical plates. The horizontal plates have a deflection factor of 25 to 34 d-c volts/inch/kv.

Deflection sensitivity is given as 0.71 mm/volt/kv for the vertical plates of a 7JP4. This means that with 1,000 volts on the second anode, 1 volt of difference between the vertical deflection plates will move the beam 0.71 millimeter on the screen. The figure of 0.71 is an average value equivalent to a deflection factor of 36 volts.

To change from one set of units to the other, take the deflection factor in d-c volts/inch/kv and divide it into 25.4 (the number of millimeters in one inch). The result is the deflection sensitivity in mm/volt/kv. To convert from the deflection sensitivity (in mm/volt/kv) to the deflection factor, we divide 25.4 by the deflection sensitivity. In other words,

$$\frac{25.4}{\text{Deflection Factor}} = \text{Deflection Sensitivity}$$

D-c volts are specified to avoid the confusion which might exist if a-c voltages were stated without specifying whether the values are peak, average or rms. As an indication of the total amount of deflecting voltage necessary, the 7JP4 tube requires 186 to 246 d-c volts to move the electron beam 1 inch horizontally, with a second anode voltage of 6,000 volts. To cover the entire 7 inches, the large value of 1,302 to 1,722 volts would be required. With alternating voltages, this is the peak-to-peak value, because the positive portion of the signal swings the beam across one half the screen and the negative portion of the signal swings it through the other half.

Centering Controls for Electrostatic Deflection Tubes. The electron beam, when not subject to any deflecting voltages, should hit the fluorescent screen at its center. In this position, a symmetrically placed image will result when the deflecting voltages are applied to the plates. In practice, stray electric and magnetic fields, or distortions of fields within the tube itself, may interfere and cause the beam to be displaced from the center position. In order to correct this condition, positioning controls for the vertical and the horizontal plates are generally available at the back of the receiver. A slotted drive screw permits adjustments to be made when necessary. Ordinarily, this should not be very often.

A popular method for applying correcting voltages to electrostatic deflecting plates to center the beam is shown in Fig. 9-8. A high fixed voltage is placed on one vertical and one horizontal deflecting plate from a tap be-

tween two 50,000-ohm resistors. In parallel with these two resistors are two 500,000-ohm potentiometers, the center arm of each going to the other vertical and horizontal deflection plates. When the arms of the potentiometers are in the center position, there is no d-c potential difference *between* the plates of the horizontal and vertical sets. A balance exists. Any change in the position of these potentiometer arms, however, will make one plate more positive than the other of either set and bend the electron beam in the

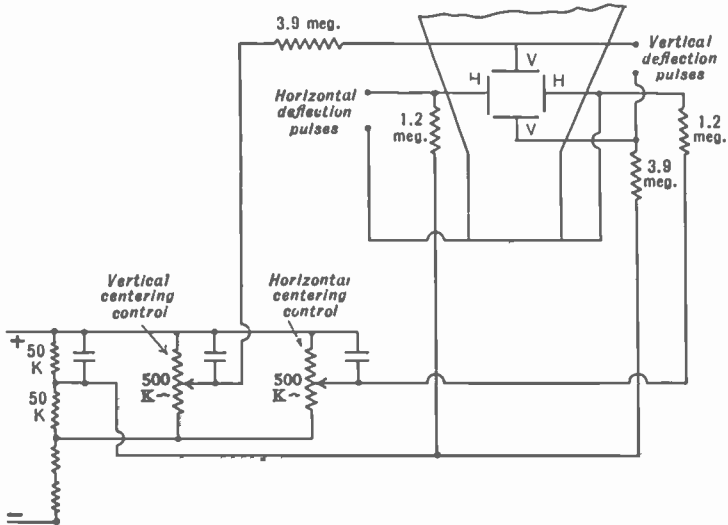


FIG. 9.8. Vertical and horizontal centering controls.

desired direction. The deflecting voltages for the image are applied separately as shown, with large resistors placed in the centering leads to act as connecting resistors to couple the deflecting voltages to the plates themselves.

Magnetic Focusing. While the preceding discussion has been concerned with electrostatic methods of focusing and deflecting the electron beam, the same operations may be performed as well magnetically. However, before any circuits are discussed, it would perhaps be advisable to review the action of magnetic fields on moving electrons.

From elementary electricity, it is well known that a wire carrying a current has a circular magnetic field set up around it, as shown in Fig. 9-9A. Suppose that the wire is placed in a magnetic field parallel to the magnetic lines of force. See Fig. 9-9B. There will be no interaction between the magnetic lines of the field and those set up by the wire. Why? Because the two *fields* are at right angles to each other.

For the opposite case illustrated in Fig. 9-9C, the current carrying wire is placed at right angles to the field lines of magnetic force. Above the wire the lines of both fields add, whereas underneath the wire they oppose

and tend to cancel. Experiment indicates that a resulting force will act on the wire in such a way that it moves from the stronger part of the magnetic field to the weaker portion. This is indicated in the figure. The illustration represents the two extreme angles that the wire and the field can make with each other. Intermediate positions (those between zero and 90 degrees) will cause intermediate values of force to act on the wire.

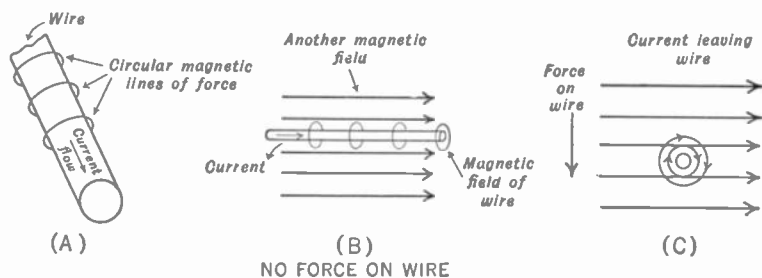


FIG. 9.9. The action of a wire carrying current when placed in a magnetic field.

The transition from a wire carrying electrons to the electrons themselves, without the wire, is quite simply made. With only electrons moving through space, the same circular magnetic field is set up about their path. From the preceding discussion, we know that electrons travelling parallel to the lines of force of an additional magnetic field experience no reaction from this field. On the other hand, if they enter the magnetic field at an angle to the flux lines, a force will be brought to bear on them and their path will be altered.

It is well to reiterate that for an electron to react with a magnetic field: (1) the electron must be moving, otherwise it does not generate a magnetic field; and (2) the moving electron must make an angle with the magnetic field in which it is travelling.

Now let us apply the foregoing facts to magnetic focusing. The focusing coil is slipped over the neck of the cathode-ray tube and placed just beyond the first anode. The first lens system remains essentially as in the previously described electrostatically controlled tubes; it still converges electrons to the cross-over point. From this point, the electrons spread out and the focusing action of the coil begins to function. The focus coil, then, represents the second lens in electromagnetic tubes. An accelerating anode (anode No. 2) is positioned close to anode No. 1 to accelerate the beam down the tube and also to provide a means of removing ions from the beam. (More on this presently.) The accelerating anode is connected internally to the aquadag coating (see Fig. 9-10) and receives its voltage from the coating. The latter, in turn, obtains its potential via a metal cavity or ball insert terminal located on the side of the glass bulb.

The field of the focusing coil is parallel to the axis of the tube and is generated by direct current flowing through the coil. As long as the electrons leave the cross-over area and travel down the tube along the axis, the magnetic lines do not interfere with their motion. However, many electrons tend to spread out beyond the cross-over region, and it is on these electrons that the magnetic force reacts because they are moving at some small angle to the magnetic flux lines.

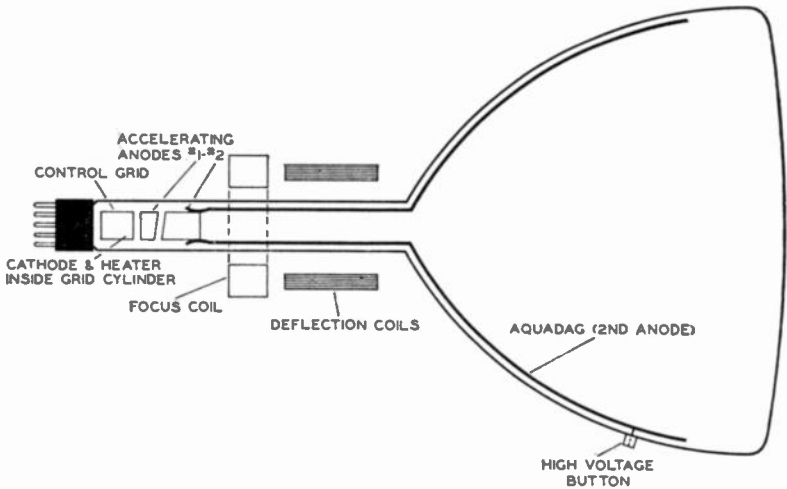


FIG. 9.10. The internal construction of an electromagnetic deflection and focus tube.

The path taken by electrons that are acted on by a magnetic field can be more easily understood if it is recalled that the resulting force on the electrons is at right angles to both its motion and the magnetic field. The result of this force, as shown in Fig. 9-11A, is to cause the electrons to move in a circular path. In this way the force on the electrons, the electronic motion, and the magnetic force are always at right angles to each other.

Apply these ideas to the action inside the cathode-ray tube. As the electrons leave the cross-over point at small angles to the magnetic field, they are subjected to a force that tends to make them turn in a circle. But at the same time that they are being forced to travel this circular path, they are also speeding forward. The resulting motion of the electrons is known as helical and is similar to the action of a screw being turned into a piece of wood. It rotates while also moving forward. Fig. 9-11B may aid the reader to visualize the motion.

The electrons that are acted on by the magnetic field all come from the cross-over point that is situated on the axis of the tube. The minute they

leave this point at some angle, the magnetic force starts to act, forcing them to move in a circular path back to the axis again. In the cathode-ray tube they are, at the same time, also moving forward and hence when the circular path is completed the electrons will again be on the axis of the tube some distance away. The exact position down the tube where the electrons return to the axis is dependent upon the strength of the magnetic force and the forward velocity.

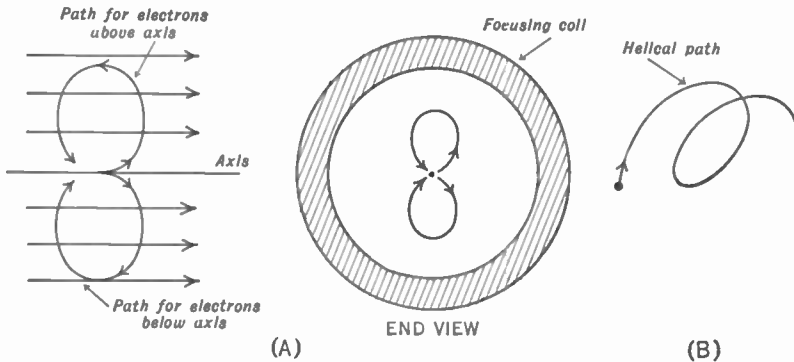


FIG. 9.11. Electrons, when cutting across magnetic lines of force, are made to move in a circular path (A). If, however, they are also subjected to an electrical force urging them forward, then their resultant path will be helical (B).

By suitable variation of the intensity of the magnetic field, it is possible to have the electrons return to the axis of the tube exactly at the screen. The beam is now focused. The greater the speed of the electrons, the stronger the magnetic field required. Thus, any changes that affect the velocity of the electrons, such as varying the first anode voltage, will also require readjustment of the current through the focusing coil.

To review the process, we find that the magnetic field causes the outgoing electrons from the cross-over region to travel in helical paths that will force them back to the axis again. With proper adjustment of the magnetic coil current, the electrons complete their circular path at the screen. Here they meet the other electrons that travelled straight along the axis (and not affected by the magnetic field) and a well-defined spot will result.

At other values of the magnetic field, defocusing occurs. As an exception to the last statement, it should be mentioned that by continually increasing the strength of the magnetic field the electrons can be made to do two (or more) complete revolutions before striking the screen. As each complete revolution brings the beam to the screen, a focused spot will appear. This process may be continued for as long as the magnetic coil will carry current.

It would appear from the preceding discussion that the magnetic field must extend all along the tube in order that the electrons are always under

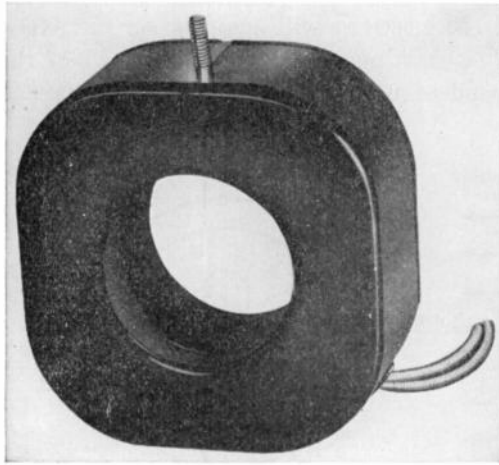


FIG. 9.12. A focusing coil.

its influence. Their path would then be helical as described. However, for practical applications, only a small iron-core coil is slipped over the neck of the tube. This is shown in Fig. 9-12. The electron beam is thus subjected to the magnetic force for only a short time. During this period it is given enough of a twist so that it will move toward the axis; the forward motion then keeps it traveling along this path. The motion now is not truly helical, but the end result is satisfactory.

It will occur to the reader that magnetic fields need not necessarily be obtained from coils only. Permanent magnets (PM) are also suitable and these have been extensively used. One type of PM focus magnet is shown in Fig. 9-13. The unit consists of three (sometimes four) small bar magnets

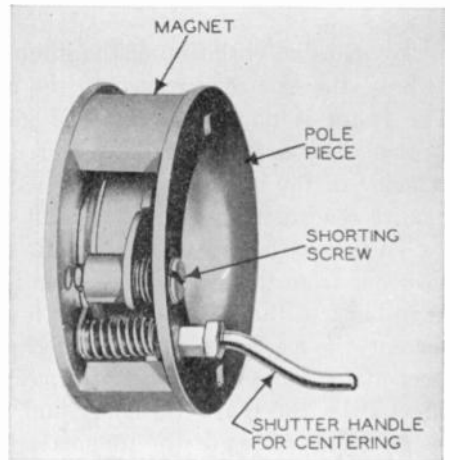
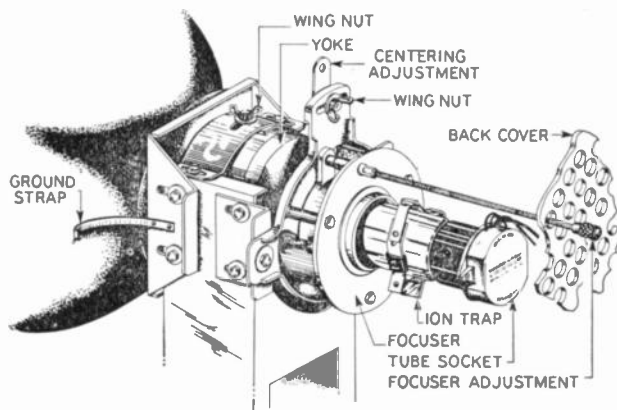


FIG. 9.13. A permanent magnet focus unit. The additional shutter handle permits centering of the picture.

which are placed along and around the axis of the tube neck. The magnets are equally spaced and held in position by two disc-like pole pieces

made of a low carbon steel. The magnets are mounted with similar poles at the same end. Thus, the flux lines joining the ends of the magnets pass through the tube along its axis. Focusing is accomplished by the interaction of these flux lines with the electron beam passing down through the tube.

To permit variation of the magnetic focusing field, one or two shorting screws are threaded into one of the end discs. The screw provides a path of lower magnetic resistance (i.e., less reluctance) for the lines of force, thereby tending to shunt them away from the tube. The magnetic field through the center of the device decreases as the screw is turned in. Turning the screw out has the opposite effect.



Courtesy Fada Radio & Electric Co.

FIG. 9.14. A P-M focusing unit using a movable steel collar for focus adjustment.

Another type of permanent focus magnet is shown in Fig. 9-14. In essence it is similar to the previous unit except that flux variation through the tube is achieved by a sliding steel collar controlled by a lead screw at the end of a flexible shaft. By rotating the knurled screw at the end of the shaft, either clockwise or counterclockwise, the steel collar can be moved back and forth. The end of the screw drive extends beyond the back cover of the receiver, permitting focus adjustments to be made without the necessity of removing the back cover.

Electromagnetic Deflection. It is possible to deflect electrons by either magnetic or electrostatic fields. Electrostatic deflection has already been described and magnetic deflection will now be considered. Actually, little new need be added to understand the action of deflecting coils on the electron beam. Two sets of coils are placed at right angles to each other and mounted on the section of the tube neck where the electron beam leaves the focusing electrode and travels toward the screen. There are four coils in all (two in each set), with opposite ones comprising one set. These are connected in series in order to obtain the proper polarity (see Fig. 9-15A).

A soft iron shell is placed around both sets of coils to act as a shield. The entire assembly is known as a yoke (see Fig. 9-15B). For horizontal deflection, the coils are vertically placed whereas, for vertical deflection, the coils are horizontally mounted. This reverse placement of the coils is due to the fact, stated above, that the force on travelling electrons in a magnetic field is at right angles to both the direction of motion and the lines of the field.

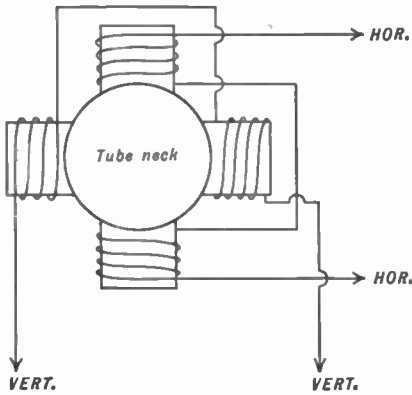
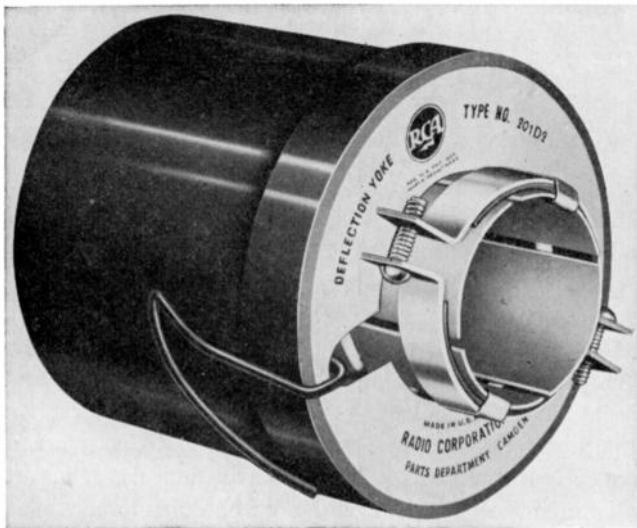


FIG. 9.15A. Electromagnetic deflection coil. This represents the physical placement of the windings.

After the coils have been oriented, saw-tooth shaped current variations are sent through them. The magnetic field flux follows these current changes and causes the electron beam to move back and forth (or up and down) across the screen, sweeping out the desired pattern.

The reader should not become con-



Courtesy RCA

FIG. 9.15B. A deflection yoke containing both horizontal and vertical deflection windings.

fused by the seemingly different actions of the focusing and the deflection coils. At first glance it may appear that one coil (the focusing coil) twists

the electron beam around so that it ends up at the screen in focus, while the other coils (the deflecting coils) only cause the beam to move either to the right or left or up and down. Actually the action of all the coils is the same; the only difference lies in the manner in which they affect the beam. With the focusing coil the magnetic lines of flux are parallel to the axis of the tube and the electrons that are moving away from the axis of the tube are subjected to a strong twisting force that turns them back to the axis. Their forward motion, given to them by the positive first anode, and sometimes by an intensifier ring, keeps them moving toward the screen.

At the deflecting coils, the magnetic fields are at *right angles* to the path of the beam. The beam, in moving through these fields, has a force applied which is at right angles to the forward motion of the electrons and the direction of the magnetic lines of force. Here the effect of the field is not as great as at the focusing coil, and the beam is merely deflected rather than bent all the way around into a circular path. The influence of the field ends when the electrons pass the yoke, but any sideward or up-and-down motion imparted to the electrons while in the field is retained. By varying the direction of the flow of current through the vertical and horizontal deflecting coils, it is possible to reach all points on the screen. This type of deflection is used with the larger cathode-ray tubes.

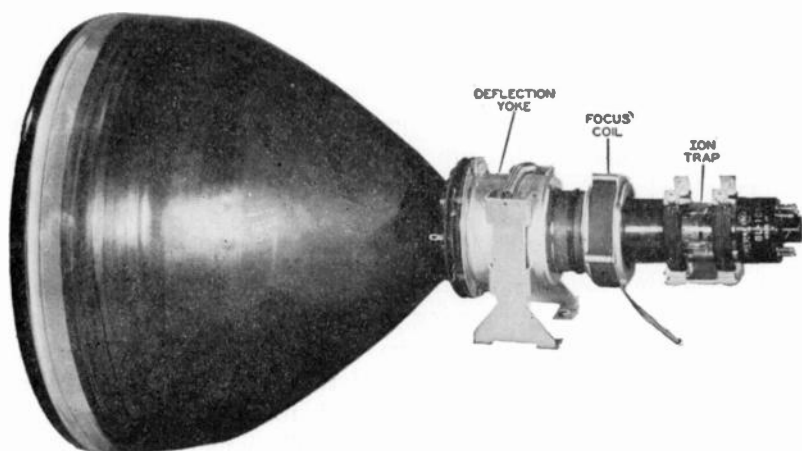
When a yoke is inserted over the neck of the picture tube, it is very easy to position it so that the image is not properly oriented. This is indicated in Fig. 9-16B. In this case, correction may be accomplished by rotating the yoke until the image is again properly positioned.

Fig. 9-16A illustrates the proper positions for focus, deflection, and ion-trap coils on the neck of a cathode-ray tube. The ion trap will be discussed presently.

Beam Centering in Electromagnetic Tubes. In electrostatic deflection and focus tubes, beam centering was accomplished as shown in Fig. 9-8. This is essentially the only method employed. In electromagnetic tubes, the deflection arrangement differs as do the beam centering methods.

In one system, a small d-c voltage is applied to the horizontal deflection windings of the yoke. The amount of voltage actually impressed across the coil windings is determined by a potentiometer. The current resulting from this voltage will produce a magnetic field, and the strength and direction (or polarity) of this field will cause the beam to move. If the current flows in one direction, the beam will be shifted one way; if the current flows in the other direction, the beam shift will likewise reverse. A similar arrangement in the vertical deflection winding will enable the set user to shift the beam up or down.

Note that this d-c centering current is in addition to the vertical and horizontal deflection currents that also flow through the yoke windings and



Courtesy G.E.

FIG. 9.16A. The positions for a deflection yoke, focus coil and ion trap on the neck of a cathode-ray tube.



Courtesy NBC

FIG. 9.16B. Incorrectly positioned image due to improper placement of deflecting coil yoke.

swing the beam across the screen. The d-c current is steady, and the beam shift it introduces is fixed. The reader may consider this as analogous to the d-c bias on a tube. Incoming signals then vary above and below this bias.

Picture centering can also be accomplished by mechanically positioning the focus coil. In the arrangement shown in Fig. 9-17, three centering screws enable the serviceman to tilt the axis of the focus coil (or magnet) in one of several different directions until the focus magnet field directs the beam to the center of the screen.

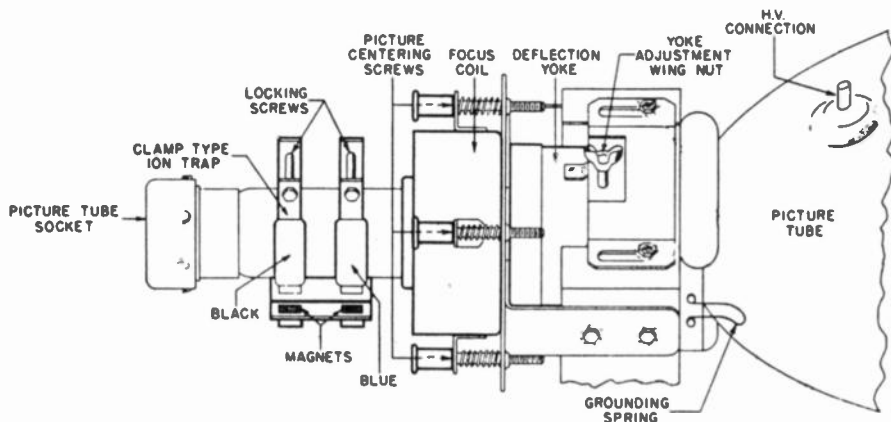


FIG. 9.17. Three centering screws permit the axis of the focus coil to be tilted in one of several directions until the beam is centered.

A system of centering widely used with permanent magnets is shown in Fig. 9-14. A metal plate is fastened onto the focus magnet. Some centering plates contain a locking screw which must be loosened before centering, and others are held in position by friction. Up-and-down adjustment of the plate moves the picture from side to side; sidewise adjustment shifts the picture up and down. Electrically what the plate does is to modify (i.e., distort) the distribution of the PM field until the desired beam shifting is achieved.

Magnetic Deflection with Electrostatic Focus. When commercial television was just getting started in the United States, in 1946, electrostatic focus and deflection tubes were manufactured in great quantity. The most popular of these tubes was the 7JP4 and, to a lesser extent, the 10HP4.

The chief drawback of electrostatic deflection (as compared to electromagnetic deflection) is that, for tubes with screens larger than 7 inches, the amount of deflecting voltage required is considerable. Since it is cheaper to develop high currents rather than high voltages, the trend turned toward electromagnetic deflection.

However, in late 1950, the outbreak of hostilities in Korea forced the

government to issue orders restricting the use of certain critical metals. Among these were aluminum, nickel, cobalt, and copper, all of which are required in the manufacture of focusing magnets. Faced with this limitation on the manufacture of receivers, the television industry set about finding ways of eliminating magnetic focusing. The result was a tube which retained electromagnetic deflection, but combined this with electrostatic focusing.

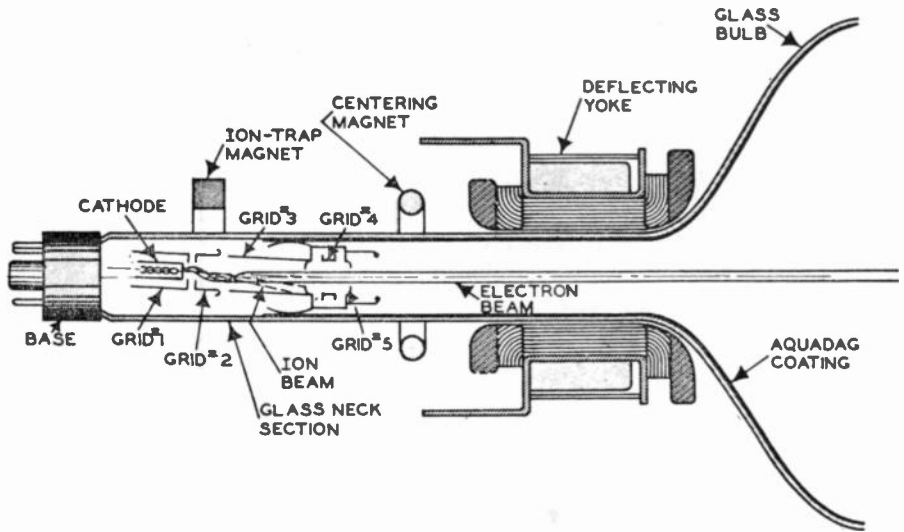


FIG. 9.18. Internal structure of new electrostatic focus tube. Deflection is accomplished magnetically.

A cross-sectional view of the new electrostatic focus gun is shown in Fig. 9-18. Grid No. 1 is the control grid. Grid No. 2 is the same as anode No. 1 previously shown.* Grid No. 3 serves the same purpose as the accelerating anode in Fig. 9-10. It contacts and operates at the high potential of the aquadag coating.

The new elements, grids Nos. 4 and 5, provide the focusing field which focuses the electron beam. The voltage applied to grid No. 4 is lower than that which grids Nos. 3 or 5 receive and it is frequently made variable to permit adjustment of the focus voltage to the proper value. Grid No. 5 (which structurally surrounds grid No. 4) is connected internally to grid No. 3 and operates at the same potential as No. 3.

*The labeling of the various elements in the gun assembly as consecutive grids is done frequently to simplify identification. Thus, the control grid is grid No. 1. The first anode following this is grid No. 2. The second anode is grid No. 3, etc. Note that this change in name does not in any way alter the function or construction of the electrodes.

The voltage which is applied to grid No. 4 depends upon the manner in which grids No. 3, 4, and 5 are constructed. The first electrostatic focus tubes manufactured required that the grid No. 4 potential be on the order of 20% of the accelerating (or second anode) voltage. This meant that voltages between 2,000 and 3,000 volts had to be made available. A special potentiometer, inserted in this circuit, permitted adjustment of this voltage for sharpest picture focus.

In a subsequent design it was found that, by constructing grids No. 3, 4, and 5 to closer tolerances, the necessary focusing action could be obtained using voltages on the order of 300 to 400 volts. These latter values can be obtained directly from the low-voltage power supply, avoiding the special circuit required when several thousand volts are needed. Because of this simplification and consequent economy, very few focus tubes of the first type are being manufactured today; the preponderance of electrostatic focus tubes falls into the second category.

It is also possible, by modifying the structure of grids No. 3, 4, and 5 to the form shown in Fig. 9-19, to obtain the proper focusing action with zero potential on grid No. 4. Now, no external voltage or focusing potentiometer need be used. DuMont engineers developed this tube and the firm is manufacturing it in quantity. As yet no other tube manufacturers have adopted their design or have built similar tubes, although this might happen. Manufacturing tolerances are somewhat more critical in this zero focus or self-focus tube, and this has tended to discourage their widespread use. It is entirely possible, however, that this difficulty will be overcome.

Since the focus magnet is dispensed with on electrostatic focus tubes, a new centering magnet is provided to center the picture on the screen. See Fig. 9-20. The magnet assembly is in the form of two wire rings mounted on a non-magnetic form which is placed around the neck of the picture tube and at a distance of about three-eighths of an inch in back of the deflection yoke.

Electrostatic tubes, although born of necessity, have given a good account of themselves and undoubtedly will take their place with all-electromagnetic tubes.

Cathode-Ray Tube Screens. Everything that has been done to the electron beam in the discussion thus far has been done with two ideas in mind, namely, to have it focus properly on the screen and to send it to different parts of the screen as well. Now let us consider the screen itself.

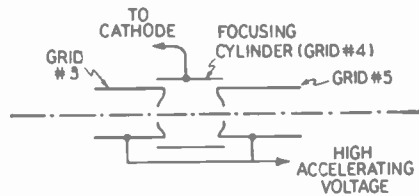


Fig. 9.19. The modified arrangements of grids nos. 3, 4, and 5 in the zero focus tube.

An electron gun, once constructed, can be subjected to considerable misuse without being permanently affected. On the other hand, failure to grasp the significance of certain precautions required to protect the fluores-

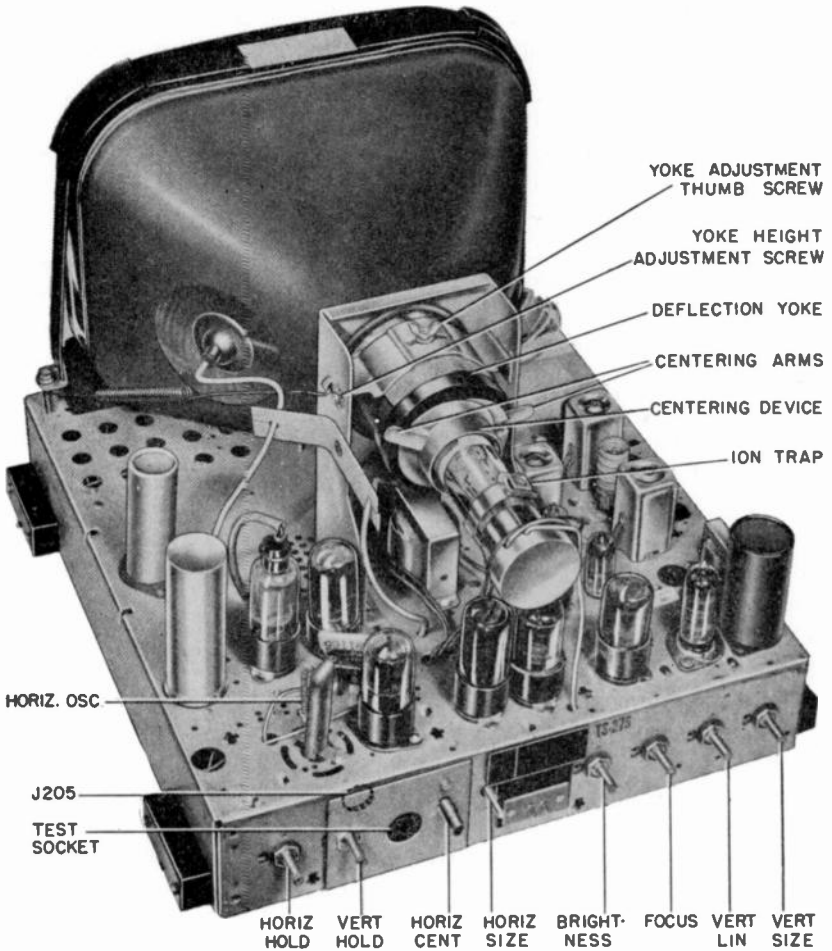


FIG. 9.20. A television receiver with one of the newer electrostatic focus tubes. *Courtesy Motorola*

cent screen can readily result in a shortened period of usefulness and/or unsatisfactory operation throughout the life of the tube.

The phenomenon by which certain substances convert the energy of an electron beam into visible light is known as luminescence. Luminescence is further divided into fluorescence and phosphorescence. Fluorescence is luminescence which ends when the exciting agent is removed. Phosphores-

cence is luminescence which exists after the exciting agent is removed. Technically, then, the screens used in television should be called phosphorescent screens, and indeed the crystalline substances used for these screens are known as phosphors. Unfortunately, however, the word "fluorescent" has become so widespread that one seldom hears the other, proper name. A tabulation of the most common phosphors in use in oscilloscopes, television receivers and radar equipment is shown in Table 9-1. For television, a

TABLE 9-1. The most common phosphors used for oscilloscopes, television receivers and radar equipment

RMA Designation — Substance	Activator	Fluorescent Color	Phosphorescence (seconds)
P1 — Zinc silicate	Manganese	Green	Med. — 0.03-0.05
P2 — Zinc sulphide	Copper	Blue-green	Long
P3 — Zinc beryllium silicate	Manganese	Yellow-Gr.	Med. — 0.05
P4 — P3 and zinc sulphide	Silver	White	Med. .02-.03
P5 — Calcium Tungstate		Blue	Very short 5 μ sec. med. 0.005
P6 — Zinc sulphide	Silver	White	
Zinc cadmium sulphide	Silver		
P7 — Zinc sulphide	Silver	Blue	Med. — 0.006
Zinc cadmium sulphide	Copper	Yellow	Long
P11 — Zinc sulphide	Silver with a nickel quencher	Blue	Very short 10 μ sec.

combination of zinc sulphide and zinc beryllium silicate is used, this combination giving higher conversion efficiency than most other known compounds. It is interesting to note that the electron beam remains at any one point for approximately 0.1 of a microsecond, yet the light emission continues for two to three hundredths of a second, indicating that the zinc sulphide and zinc beryllium silicate are responsible through their phosphorescence for essentially all of the emitted light seen by the observer. A typical persistence curve for zinc sulphide is shown in Fig. 9-21.

The principal objective in the design of a cathode-ray tube is the production of an image having good brightness and high contrast. When the electron beam strikes the backside of the fluorescent screen, the light which is emitted distributes itself in the following approximate manner:

50% of the light travels back into the tube.

20% of the light is lost in the glass of the tube by internal reflection.

30% reaches the observer.

Thus, of all the light that is produced by the electron beam (and this, itself, is a highly inefficient process), only 30 per cent reaches the observer.

Image contrast is impaired because of the interference caused by light which is returned to the screen after it has been reflected from some other points. Some of these sources of interference are:

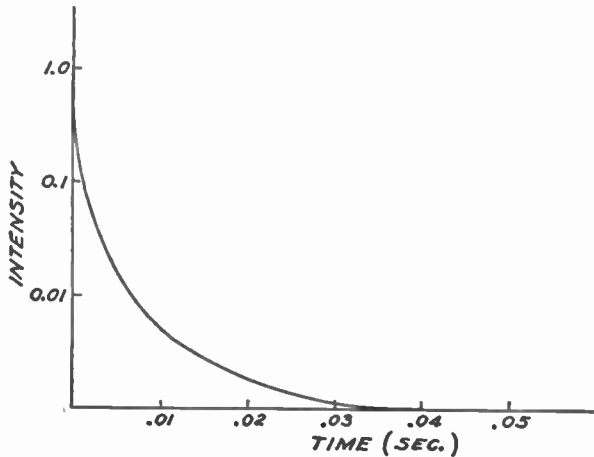


FIG. 9.21. A persistence curve for zinc sulphide, commonly used in P4 television screens.

1. Halation.
2. Reflections due to the curvature of the screen.
3. Reflections at the surface of the screen face.
4. Reflections from inside the tube.

These are listed in the order of their importance.

Halation. If we take a cathode-ray tube and minutely examine the light pattern produced by a stationary electron beam, we find that the visible spot is surrounded by rings of light. These rings of light are due to a phenomenon known as halation (see Fig. 9-22). The light rays which leave the

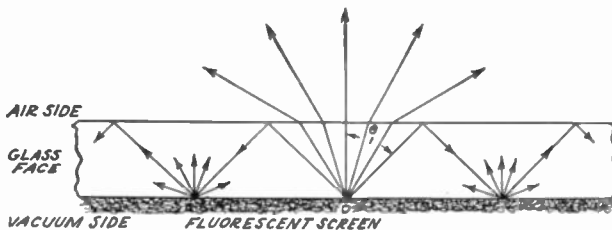


FIG. 9.22. Illustration of how reflections between the two sides of the glass can cause halation.

fluorescent crystals at the inner surface of the tube face travel into the glass and are refracted. Those rays which make an angle greater than θ do not leave the glass when they reach the outer surface, but instead are totally reflected back into the glass. At each point where these reflected rays strike the fluorescent crystals, they scatter and it is this scattering of the rays that produces visible rings on the screen. These rings cause a hazy glow in the region surrounding the beam spot and reduce the maximum possible detail contrast. Contrast, it will be remembered, is the ratio of the brightness of two points, one of which is being bombarded by the electron beam, the other, which is under cut-off conditions. It is desirable to have this ratio as high as possible in order to achieve "rich-looking" or high-quality images. Due to the scattering of the light, however, areas which should be in total darkness receive some light and the result is a reduction in the contrast ratio. A distinction is usually made between the detail contrast ratio, which is defined above, and the overall field contrast. The field contrast ratio compares two sections of the screen which are widely removed from each other. Halation affects only detail contrast.

Reflections Due to the Curvature of the Screen. Loss in contrast due to reflections arising from the curvature of the screen is shown in Fig. 9-23. The remedy for this is the use of a flat screen. Much progress has been made in this direction since the screen curvature greatly restricts the useful image area.

Reflections at the Surface of the Screen Face. Light rays, when they travel from one medium to another, always lose a certain amount of energy at the intersection of the two media. At the cathode-ray tube screen, some light is reflected when it reaches the dividing surface between the air and the glass of the tube. The reflected light travels back to the inner surface and then back to the outer surface again. At each dividing surface, some of the light continues onward and some is reflected back into the glass. Absorption and dispersion quickly reduce the strength of these rebounding rays.

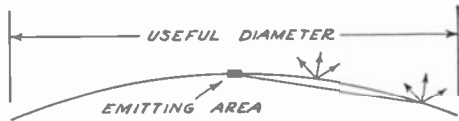


FIG. 9.23 Diffusion effects in nonflat screen.

Reflections from Inside the Tube. In Fig. 9-24 we see how reflections from the inside surfaces of the tube can act to decrease the field contrast of the image. The loss in contrast from this source of interference can be made quite low by special shaping of the bulb walls, as shown in Fig. 9-24, and the use of the black aquadag coating. The aquadag coating is also useful for electrical purposes, acting as a shield and a path for the return of the secondary electrons emitted from the fluorescent screen. Secondary electrons must be emitted by the screen, otherwise the negative charge accumu-

lation on the screen would soon become great enough to prevent the electron beam from reaching it.

One step taken toward improving screen brightness and contrast has been the addition of an extremely thin film of aluminum on the back of the fluorescent screen.

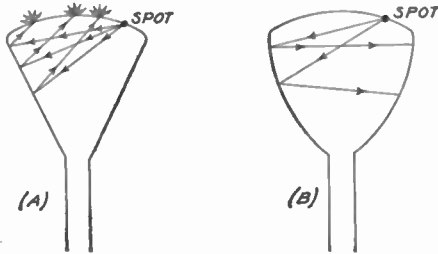


FIG. 9.24. The shape of the tube built can reduce internal reflections.

is one improvement. In addition, the overall field contrast is improved by as much as ten times. However, the detail contrast is not noticeably affected since it is governed primarily by halation, and the addition of the aluminum layer does not affect this.

The film is sufficiently thin to permit the electrons in the scanning beam to reach the fluorescent crystals. It will prevent, however, any of the light which is generated by the screen crystals from travelling back into the tube. This is shown in Fig. 9-25. The light which previously went back into the tube is now reflected toward the observer. This

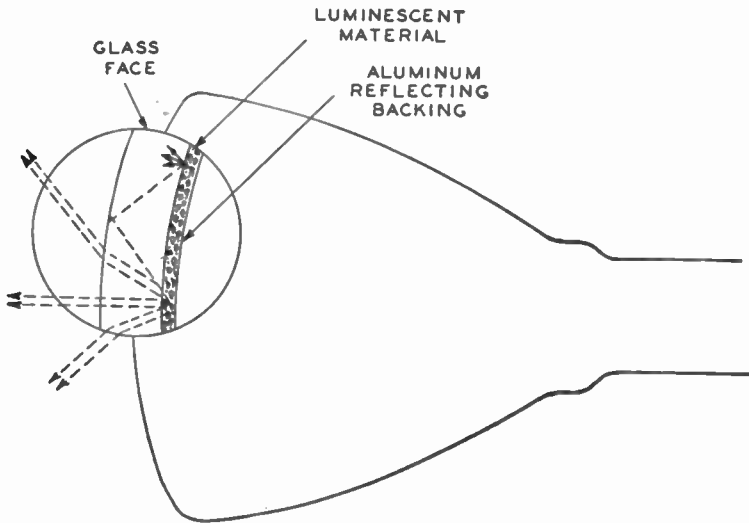


FIG. 9.25. An aluminum backing over the fluorescent screen prevents light from traveling back into the tube.

The metal film is extremely thin, being on the order of $3,500 \times 10^{-8}$ centimeters thick. Since even a layer this thin does interpose a barrier in the path of the electron, it is essential that sufficiently high accelerating voltages

be used. In Fig. 9-26, a comparison between the efficiency curves of screens having the metallic layer and those which do not is shown. At low accelerating voltages, the loss of energy by the electrons in penetrating the layer decreases their efficiency * below that of similar electrons in tubes not possessing this layer. The poorer efficiency continues until we reach the point where the curves intersect.

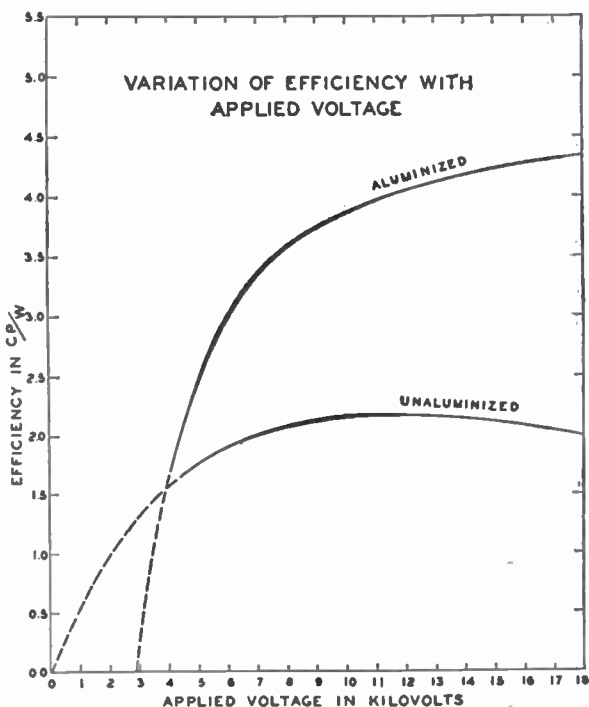


Fig. 9.26. Variation in efficiency of aluminized and unaluminized screens.

Beyond this region, the screen with the metallic layer proves to be quite superior to the ordinary tube. The rapid rise in efficiency is due to a decrease in energy lost at the metallic barrier plus an increase in the overall brightness due to the light-resisting characteristics of the layer itself.

Sticking Potentials. An additional purpose which the aluminum film serves is to avoid undesirable effects due to poor secondary emission from the screen. The electrons in the beam, where they strike the screen, must somehow be brought back to the cathode. The fluorescent crystals themselves are essentially non-conductors. If the electrons from the beam were

* Efficiency is expressed here in candle power of illumination attained for each watt of electrical power expended. CP/W is an abbreviated notation for candle power per watt.

allowed to accumulate on these crystals, a point would soon be reached where a negative charge would accumulate sufficiently to prevent any additional electrons from reaching the screen. Originally, when tubes were first built, a thin metal film was deposited on the glass face of the tube and the fluorescent screen coated on this. The metal film was connected to the second anode, thereby assuming the same potential. When the electrons hit the fluorescent coating, they continued through to the metal film.

Further investigation revealed that, without the metal film, the tube would still work because of the secondary emission from the screen. The beam electrons, when they hit the fluorescent screen, imparted sufficient energy to the screen electrons to cause them to leave the screen. These emitted electrons reached the aquadag coating on the walls of the tube and by this path were conducted back to the cathode. The removal of the metal film was a considerable step forward toward obtaining a brighter image.

The return of the electrons by secondary emission was not without limitations. First, the number of secondary electrons emitted depended upon the velocity of the arriving beam electrons. If their velocity was too low, there was no secondary emission and the tube would not function in any satisfactory manner. As we step up the beam velocity, secondary electrons are emitted, with good tube operation to the point where there are just as many electrons arriving at the screen as are leaving. Beyond this, additional increase in beam velocity will produce no corresponding increase in light output. If it is found, for example, that the number of arriving and emitted electrons at the screen is equal when the second anode potential is 8,000 volts, then raising this voltage to 12,000 volts will produce no greater light output, despite the higher voltage. This critical potential, at which the ratio of secondary electrons to beam electrons becomes equal to one, is known as the sticking potential. Any further increase in beam velocity will cause the ratio to decrease below one with the result that the screen accumulates sufficient negative charge to effectively reduce its potential to the critical point. No matter what the accelerating potential on the second anode may be, the effective screen potential cannot exceed its critical value. Thus, if the accelerating voltage is 12,000 volts, and the critical potential of the screen is 8,000 volts, then the light emitted from the screen will be on the basis of 8,000 volts, not 12,000 volts.

From the standpoint of screen manufacture it is desirable to have the critical potential as high as possible, certainly above the operating potentials of the tube. The method of providing an aluminum film corrects many of the defects due to secondary emission difficulties and greatly increases the range of substances which can be used for screen phosphors. Previously, each material had to be carefully examined to determine whether its sticking potential was of a suitable value.

Ion Spots. Another matter of considerable importance is the elimination of the ion spot in tubes using electromagnetic deflection. No matter how carefully a tube is degassed or how well cathode coating is applied, it will be found that ions are present in the electron beam. These ions are either gas molecules which have acquired an electron or else molecules of the outside coating material of the cathode. These ions possess the same charge as the electrons and are sensitive to the same accelerating voltages. In tubes employing electrostatic deflection, the ions and the electrons are similarly deflected and for all practical purposes may be considered as one. However, when electromagnetic deflection is employed, it will be found that these heavier ions are hardly deflected. As a result, they tend to impinge on the center of the screen in a steady stream and produce a thin film of deactivated material on the area exposed. When the electrons in the scanning beam pass over this area, no light is produced. To the observer this appears as a dark patch.

The reason for the difference in deflection characteristics of the ion can be obtained from an inspection of the equations governing electromagnetic and electrostatic deflection. For the tube which employs electromagnetic deflection, we have

$$d = \frac{DLeH}{mv}$$

where d = the distance the beam is deflected on the screen,
 D = distance from the deflection field to the screen,
 e = charge of particle deflected,
 H = strength of magnetic field,
 v = velocity of traveling particle,
 L = length of magnetic field,
 m = mass of particle.

For electrostatic deflection, the expression is

$$d = \frac{1}{2} \cdot \frac{V}{E} \cdot \frac{S}{h} \left(D + \frac{S}{2} \right)$$

where d = the distance the beam is deflected on the screen,
 V = potential difference between the deflecting plates,
 E = forward accelerating voltage of the tube (i.e., second anode),
 S = length of the deflection plate,
 h = the separation of the deflection plates from each other,
 D = distance from the end of the deflection plate to the screen.

With electromagnetic deflection the mass of the deflected particle appears in the equation; in electrostatic deflection it does not. Hence, the ions,

because of their greater mass, will receive less displacement than electrons in electromagnetic systems. However, when the mass of the particle does not enter into consideration, ions and electrons receive similar treatment.

a. **Bent Gun Ion Trap.** There are several methods available for preventing the ions from reaching the screen. First there is the bent electron gun, as shown in Fig. 9-27. The cathode, when heated, will emit ions and

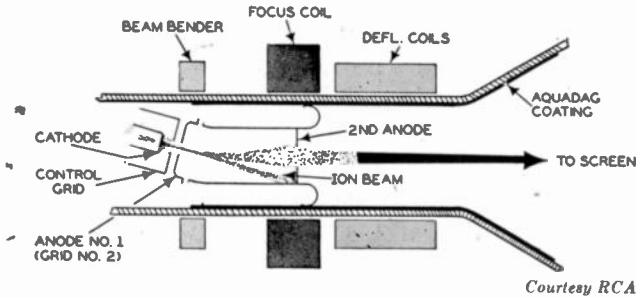


FIG. 9.27. A bent-gun ion trap.

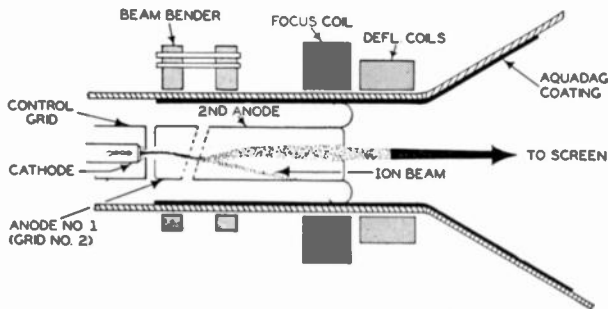


FIG. 9.28. A diagonal-cut (or slash-field) type ion trap.

electrons and these will be accelerated to the first and second anodes. However, the cathode is inclined at an angle to the rest of the gun structure and both particles would, if permitted to travel in a straight line, impinge on the side of the electron gun and never reach the screen. However, if a strong magnetic field is placed in the path of the particles, it is possible to alter the paths of the electrons sufficiently so that they travel toward the screen. The heavier ions, however, are not sufficiently deflected and as a result they hit the side of the electron gun. The magnetic field which causes this separation of ions and electrons is obtained from a small coil or permanent magnet placed on the outside of the neck of the tube, above the cathode. The ion trap magnet (or beam bender, as it is sometimes called) is clamped onto the neck of the tube in the position shown in Fig. 9-27.

b. **Diagonal-cut Ion Trap.** Another approach to the prevention of ion burns is the diagonal-cut ion trap. See Fig. 9-28. The electrons and ions

are emitted by the electron gun and are accelerated forward. The first and second anodes are so designed that the gap between them is oblique. The first anode has a low positive voltage; the second anode has a high positive voltage. The electrons, as they leave the cathode, are attracted forward by the first anode. However, the oblique gap between the first and second anode causes the electric field here to become warped with the result that the electrons and ions crossing the gap are bent in toward the second anode. With no other forces applied, the electrons and ions would strike the second anode and be prevented from reaching the screen.

However, if a magnetic field is introduced in the tube at right angles to the electrode, the electrons receive a counter force deflecting them upward and permitting them to continue through the gun. The ions, because of their greater mass and the fact that the magnetic field scarcely deflects them, strike the second anode and are removed from the beam path.

The magnets are clamped on the tube stem in the manner illustrated in Fig. 9-29.

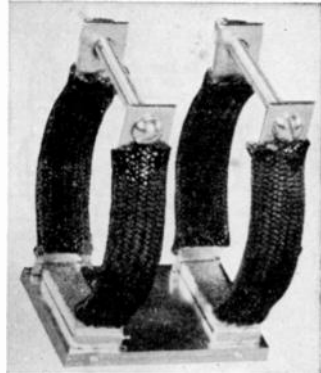
The ion trap may operate electromagnetically, in which case current flowing through the front and rear ion trap coils provides the necessary magnetic field, or the trap may contain fixed magnets of Alnico (Fig. 9-29).

It should be noted that the diagonal-cut (or slash-field) type of gun requires a double magnet whereas the bent gun utilizes a single magnet. In every complete picture tube listing, information concerning the type of magnet to use is generally included.

Four types of PM double magnet ion traps are in use. These are shown in Fig. 9-30.

1. One type is held in place with two clamps, colored black and blue, and tightened onto the neck of the tube with two screws.
2. A ring type containing a large and a small circular magnet which slip over the neck of the tube.
3. A bar type wherein two square bar magnets slip over the neck of the tube.
4. A friction-clamp type.

The ion trap unit should be placed on the neck of the tube with the stronger of the two magnets over the first anode. On some units, the stronger magnet will also be the larger one. On other units, the color code markings indicated in Fig. 9-30 will be used to guide the serviceman. Sometimes an arrow will be found stamped into the magnets' holding plate and,



Courtesy G.E.

FIG. 9.29. A fixed magnet type of ion trap.

in such instances, the unit should be positioned so that the arrow points toward the tube screen.

After the magnet is in position, the set is turned on and allowed to warm up.* Now turn up the brightness control and adjust the magnet by moving it forward or backward, at the same time rotating it around the tube neck until the brightest raster is observed on the screen. Keep reducing the brightness control as the system is brought into line. When ad-

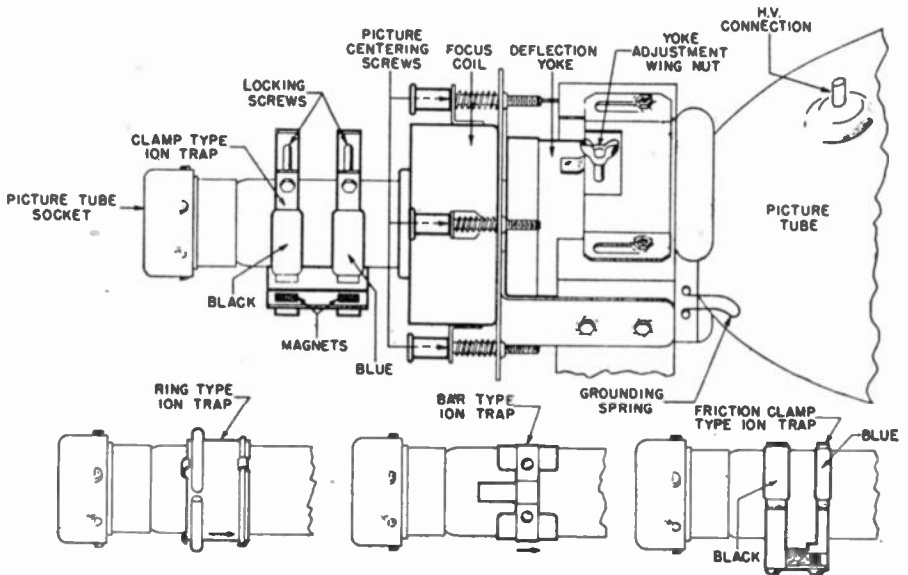


Fig. 9.30. Various types of double magnet ion traps.

justment is complete, make certain that the ion trap is held tightly in position.

It may happen that, when the raster is brightest, shadows will appear at the sides or corners of the raster. Never correct for a shadowed raster with the ion-trap magnet if such correction results in decreased brightness. The ion-trap magnet must always be adjusted for maximum brightness and, if shadows occur at this setting, they should be eliminated by adjusting the focus and deflection coils.

Sometimes it will be possible to obtain two brightness maxima in moving the ion trap back and forth along the neck. The correct position to use is the one which is closest to the base of the tube. The second maximum is usually obtained when the ion-trap magnet is close to the focus coil or magnet. The interaction of the two magnetic fields is such that a brightness

* The adjustment procedure given here is, the same for single or double magnet traps.

maximum is obtained in this incorrect location. Tubes should not be operated with the ion-trap magnet in this second position since possible damage to the tube may result.

c. **Metal-backed Screen Tubes.** The third method of preventing ions from reaching the screen is accomplished by the aluminum layer mentioned previously. The depth of penetration of any particle is governed by the relationship,

$$\text{depth of penetration} = \frac{K(Ve)}{m}$$

where K = constant,

Ve = energy of particle,

m = mass of particle.

Since the ion has considerably more mass than an electron, its depth of penetration is less. By properly proportioning the thickness of the metallic screen, the ions are excluded, but the electrons in the beam are able to pass through.

A metal-backed screen will be fully effective if the second anode or accelerating voltage is not too high. Thus, the 12KP4 with an accelerating voltage of 10,000 volts does not require an ion trap because of its metal-backed screen. However, the 19AP4C, with an accelerating potential of 16,000 volts, does.

Some of the more common cathode-ray tubes currently in use are listed in Table 9-2, together with their characteristics. The 3NP4 and the 5TP4 are both projection tubes producing an extremely intense image on their screen and requiring accelerating voltages of about 27,000 volts. All the other tubes listed are of the direct-viewing type, so-called because the images produced on their screen are viewed directly by the set user. Under these circumstances the image need not be as intense as that required by projection tubes, and the accelerating voltages used are correspondingly lower. Tubes with 7-inch screens require between 5,000-6,000 volts, while the 10-, 12-, 15-inch and larger tubes operate satisfactorily with 9,000-16,000 volts. It is to be noted that in all these tubes the accelerating voltages are far higher than normally encountered, and extreme caution should be exercised when working with them.

In electromagnetic deflection tubes used before World War II, the deflecting angle of the electron beam was 40 degrees. (See Fig. 9-31.) Following the war, the deflection angle was raised to 50°, then 60°, 70°, and now many of the 24- and all of the 27- and 30-inch tubes utilize 90° deflection angles. The increased deflecting angle results in a shorter overall length and therefore permits the use of relatively large screen tubes in rea-

TABLE 9.2. TV Picture Tube Characteristics

TUBE TYPE	FACE SHAPE	ENVE-LOPE	LENGTH (IN.)	TYPE DE-FLECTION	DEFLECT. ANGLE (°)	ION TRAP	ANODE CONNCG.	H.V. (KV)	FOCUS	MISC.
3KP4	Round	Glass	11 1/2	Elec	---	None	---	2	E 450 v.	C
3NP4	Round	Glass	10	Magnetic	---	None	Ball	24	M 120 ma.	Projection
5TP4	Round	Glass	12 1/2	Magnetic	---	None	Cavity	27	E 4.9 kv.	Projection
7AP4	Round	Glass	13 1/2	Magnetic	55	None	---	3.5	E 600 v.	C
7CP4	Round	Glass	13 1/2	Magnetic	57	None	---	6	E 1.1 kv.	C
7DP4	Round	Glass	14 1/2	Magnetic	50	Double	Cavity	6	E 1.4 kv.	C
7EP4	Round	Glass	15 1/2	Elec	---	None	---	2.5	E 650 v.	C
7GP4	Round	Glass	14 1/2	Elec	---	None	---	3	E 1 kv.	C
7HP4	Round	Glass	13 1/2	Magnetic	50	None	Ball	6	M ---	C
7JP4	Round	Glass	14 1/2	Elec	---	None	---	6	E 1 kv.	C
8AP4	Round	Metal	14 1/2	Magnetic	54	Single	Rim	9	M 112 ma.	Filterglass
8AP4A										
8BP4	Round	Glass	16 1/2	Elec	---	None	---	6	E 2 kv.	C
9AP4	Round	Glass	21 1/2	Magnetic	---	None	Cap	6	E 1.2 kv.	C
10BP4	Round	Glass	17 1/2	Magnetic	50	Double	Cavity	10	M 130 ma.	C
10BP4A										
10CP4	Round	Glass	17 1/2	Magnetic	50	None	Ball	10	M 150 ma.	
10DP4	Round	Glass	18	Magnetic	50	None	Cavity	10	E 2.9 kv.	C
10EP4	Round	Glass	18	Magnetic	50	Double	Cavity	9	M 132 ma.	C
10FP4	Round	Glass	18	Magnetic	50	None	Cavity	10	M 115 ma.	Filterglass
10FP4A										
10GP4	Round	Glass	18 1/2	Elec	---	None	---	5	E 1.5 kv.	Discontinued
10HP4	Round	Glass	19 1/2	Elec	---	None	---	5	E 1.5 kv.	Discontinued
10MP4	Round	Glass	17 1/2	Magnetic	52	Double	Cavity	10	M 150 ma.	No 2nd grid
10MP4A										
12AP4	Round	Glass	25 1/2	Magnetic	50	None	Cap	7	E 1.5 kv.	Discontinued
12CP4	Round	Glass	18 1/2	Magnetic	50	Double	Cap	7	M ---	No 1st anode.C
12JP4	Round	Glass	18	Magnetic	50	None	Ball	12	M 146 ma.	C
12KP4	Round	Glass	18	Magnetic	54	None	Cavity	12	M 135 ma.	Filterglass
12KP4A										
12LP4	Round	Glass	19 1/2	Magnetic	57	Double	Cavity	12	M 125 ma.	Filterglass
12LP4A										
12QP4	Round	Glass	17 1/2	Magnetic	55	Single	Ball	10	M 135 ma.	C
12QP4A										
B1034	Round	Glass	17 1/2	Magnetic	55	Single	Cavity	10	M 135 ma.	Repl for 12QP4.
12RP4	Round	Glass	18	Magnetic	56	Single	Ball	10	M 135 ma.	C
12TP4	Round	Glass	19 1/2	Magnetic	54	Double	Cavity	10	M 110 ma.	C
12UP4	Round	Metal	19	Magnetic	54	Double	Rim	10	M 120 ma.	Filterglass
12UP4A										
12UP4B						Single				Etched face
12VP4	Round	Glass	18 1/2	Magnetic	55	Double	Cavity	10	M 150 ma.	No 2nd grid.C
12VP4A										
12WP4	Round	Glass	17 1/2	Magnetic	55	Single	Special	12	M 110 ma.	Filterglass
12YP4	Round	Glass	18 1/2	Magnetic	54	Single	Cavity	12	Automatic	Cyl face
14AP4	Round	Glass	24 1/2	Elec	---	None	---	8	E 4 kv.	Discontinued
14BP4	Rect.	Glass	16 1/2	Magnetic	70 diag.	Double	Cavity	10	M 110 ma.	Filterglass
14CP4	Rect.	Glass	17 1/2	Magnetic	70 diag.	Single	Cavity	12	M 115 ma.	F
14DP4	Rect.	Glass	17 1/2	Magnetic	70 diag.	Double	Cavity	11	M 100 ma.	F
14EP4	Rect.	Glass	16 1/2	Magnetic	70 diag.	Single	Cavity	12	M 110 ma.	F
14FP4	Rect.	Glass	16 1/2	Magnetic	70 diag.	Single	Cavity	12	M 115 ma.	F
14GP4	Rect.	Glass	17 1/2	Magnetic	70 diag.	Single	Cavity	12	E 2.5 kv.	Filterglass
14GP4A										
15AP4	Round	Glass	20 1/2	Magnetic	57	None	Ball	12	M 160 ma.	C
15CP4	Round	Glass	21 1/2	Magnetic	50	Double	Cavity	12	M 110 ma.	C
15DP4	Round	Glass	20 1/2	Magnetic	57	Single	Cavity	13	M 145 ma.	C
81014P4	Round	Glass	20 1/2	Magnetic	57	Single	Ball	13	M 145 ma.	Repl. for 15DP4.C

M = Magnetic focus.

E = Electrostatic focus.

C = Tube requires use of 1000 micromicrofarad capacitor between 2nd anode and chassis ground.

F = Requires RTMA #109 focus coil (470 ohms) or equivalent.

All cylindrical-faced tubes have Filterglass.

TABLE 9.2. TV Picture Tube Characteristics (cont.)

TUBE TYPE	FACE SHAPE	ENVELOPE	LENGTH (IN.)	TYPE DE-FLECTION	DEFLECT. ANGLE (°)	ION TRAP	ANODE CONNEX.	H.V. (KV.)	FOCUS	MISC.
16ABP4	Rect	Glass	18 1/2	Magnetic	70 diag	Single	Cavity	14	Automatic	Filterglass
16ACP4	Round	Glass	20 1/2	Magnetic	60	Single	Cavity	14	Automatic	
16AEP4	Rect.	Glass	18 1/2	Magnetic	70 diag	Single	Cavity	13	E -60 to 350 v	Filterglass
16AP4										
16AP4A	Round	Metal	22 1/2	Magnetic	53	Double	Rim	12	M 90 ma F	Filterglass Etched face C
16AP4B										
16CP4	Round	Glass	21 1/2	Magnetic	52	Double	Cavity	12	M 110 ma F	
16CP4A	Round	Glass	22 1/2	Magnetic	53	Double	Cavity	14	M 115 ma F	C
16DP4										
16DP4A	Round	Glass	21	Magnetic	60	Single	Cavity	12	M 115 ma F	Filterglass, C
16EP4										
16EP4A	Round	Metal	20	Magnetic	60	Double	Rim	12	M 100 ma F	Filterglass Etched face C
16EP4B										
16FP4	Round	Glass	20 1/2	Magnetic	62	Single	Ball	13	M 145 ma F	C
16GP4										Filterglass
16GP4A	Round	Metal	17 1/2	Magnetic	70	Single	Rim	12	M 100 ma F	Etched face Etched face
16GP4B										
16GP4C										
16HP4										
16HP4A	Round	Glass	21 1/2	Magnetic	60	Double	Cavity	12	M 110 ma	Filterglass
16JP4										
16JP4A	Round	Glass	21 1/2	Magnetic	60	Double	Cavity	11	M 115 ma	Filterglass
16KP4	Rect	Glass	19 1/2	Magnetic	70 diag	Single	Cavity	14	M 105 ma F	Filterglass Filterglass
16KP4A										
16LP4	Round	Glass	22 1/2	Magnetic	52	Double	Cavity	12	M 110 ma F	Filterglass
16LP4A										
16MP4										
16MP4A	Round	Glass	22 1/2	Magnetic	60	Double	Cavity	12	M 110 ma F	Filterglass
16QP4	Rect	Glass	19 1/2	Magnetic	70 diag	Double	Cavity	11	M 130 ma	Filterglass, C
16RP4	Rect	Glass	19 1/2	Magnetic	70 diag	Double	Cavity	12	M 100 ma F	Filterglass
16SP4										
16SP4A	Round	Glass	17 1/2	Magnetic	70	Double	Cavity	12	M 110 ma F	Filterglass
16TP4	Rect	Glass	18 1/2	Magnetic	70 diag	Single	Cavity	12	M 110 ma F	Filterglass
16UP4	Rect	Glass	18 1/2	Magnetic	70 diag	Single	Cavity	12	M 100 ma F	Filterglass, C
16VP4	Round	Glass	17 1/2	Magnetic	70	Single	Cavity	12	M 110 ma F	Filterglass, C
16WP4										C
16WP4A	Round	Glass	18 1/2	Magnetic	70	Double	Cavity	12	M 110 ma F	Filterglass
16XP4	Rect	Glass	19 1/2	Magnetic	70 diag	Double	Cavity	12	M 100 ma F	Filterglass, C
16YP4	Round	Glass	17 1/2	Magnetic	70	Single	Cavity	12	M 100 ma F	Filterglass
16ZP4	Round	Glass	22 1/2	Magnetic	52	Double	Cavity	12	M 110 ma F	Filterglass
17AP4	Rect	Glass	19	Magnetic	70 diag	Single	Cavity	12	M 100 ma F	Filterglass, C
17BP4										Filterglass, C
17BP4A	Rect	Glass	19 1/2	Magnetic	70 diag	Single	Cavity	12	M 100 ma F	Filterglass Filterglass Etched face, C
17BP4B										
17BP4C										
17CP4	Rect	Metal	19	Magnetic	70 diag	Single	Rim	12	M 95 ma F	Etched face Filterglass
17CP4A										
17FP4										Filterglass
17FP4A	Rect	Glass	19 1/2	Magnetic	70 diag	Single	Cavity	12	E 2-3 kv	Filterglass
17GP4	Rect	Metal	18 1/2	Magnetic	70 diag	Single	Rim	12	E 2-3 kv	Etched face
17HP4	Rect	Glass	19 1/2	Magnetic	70 diag	Single	Cavity	14	E -50 to 270 v	Filterglass
17JP4	Rect	Glass	19 1/2	Magnetic	70 diag	Single	Cavity	14	M 95 ma F	Filterglass
17KP4	Rect	Glass	19 1/2	Magnetic	70 diag	Single	Cavity	14	Selfocus	Filterglass
17LP4										
17LP4A	Rect.	Glass	19 1/2	Magnetic	70 diag	Single	Cavity	14	E -50 to 270 v	Cyl. face
17QP4	Rect	Glass	19 1/2	Magnetic	70 diag	Single	Cavity	14	M 110 ma F	Cyl. face
17RP4										Filterglass
17RP4A	Rect	Glass	19 1/2	Magnetic	70 diag	Single	Cavity	14	E -100 to 100 v	Filterglass

Courtesy Radio and Television News Magazine.

TABLE 9.2. TV Picture Tube Characteristics (cont.)

TUBE TYPE	FACE SHAPE	ENVE-LOPE	LENGTH (IN.)	TYPE DE-FLECTION	DEFLECT. ANGLE (°)	ION TRAP	ANODE CONNEX.	H.V. (KV)	FOCUS	MISC.
175P4	Rect	Glass	19 1/2	Magnetic	70 diag	Single	Cavity	12	Automatic	Cyl face
177P4	Rect	Metal	19 1/2	Magnetic	70 diag	Single	Rim	14	E 0 to 500 v	Etched face
17UP4	Rect	Glass	19 1/2	Magnetic	70 diag	Single	Cavity	12	M 95 ma F	Cyl face
17VP4	Rect	Glass	19 1/2	Magnetic	70 diag	Single	Cavity	14	E 50 to 300 v	Cyl face
17YP4	Rect	Glass	19 1/2	Magnetic	70 diag	Single	Cavity	12	M 90 ma	Cyl face
19AP4	Round	Metal	21 1/2	Magnetic	66	Single	Rim	16	M 140 ma	Filterglass Etched face Filterglass Etched face
19AP4A										
19AP4B										
19AP4C										
19AP4D	Round	Glass	21 1/2	Magnetic	66	Double	Cavity	13	M 140 ma	Filterglass Filterglass, C Etched face, C Filterglass, C Filterglass, C Filterglass
19BP4A										
19DP4										
19DP4A										
19EP4	Rect	Glass	21 1/2	Magnetic	70 diag	Double	Cavity	14	M 100 ma F	Filterglass, C
19FP4	Round	Glass	22	Magnetic	66	Double	Cavity	14	M 100 ma F	Etched face, C
19GP4	Round	Glass	21 1/2	Magnetic	60	Single	Cavity	14	M 100 ma F	Filterglass, C
19JP4	Rect	Glass	21 1/2	Magnetic	70 diag	Single	Cavity	12	M 95 ma F	Filterglass, C
19QP4	Rect	Glass	21 1/2	Magnetic	70 diag	Single	Cavity	12	E 50 to 350 v	Filterglass
20AP4	Round	Glass	27 1/2	Elec	---	None		8	E 4 kv	Discontinued, C
20BP4	Round	Glass	28 1/2	Magnetic	54	None	Cap	15	M 135 ma	Discontinued, C
20CP4	Rect	Glass	21 1/2	Magnetic	70 diag	Single	Cavity	12	M 95 ma F	Filterglass Etched face, C C
20CP4A										
20CP4C										
20DP4										
20DP4A	Rect	Glass	21 1/2	Magnetic	70 diag	Single	Cavity	12	M 95 ma F	Filterglass
20FP4	Rect	Glass	21 1/2	Magnetic	70 diag	Single	Cavity	12	E 2-3 kv	Filterglass, C Filterglass
20FP4A										
20GP4	Rect	Glass	21 1/2	Magnetic	70 diag	Single	Cavity	12	E 2-3 kv	Filterglass
20MP4	Rect	Glass	21 1/2	Magnetic	70 diag	Single	Cavity	12	E 50 to 270 v	Filterglass C
20MP4A										
20MP4B										
20JP4	Rect	Glass	21 1/2	Magnetic	70 diag	Single	Cavity	13	Automatic	Filterglass Filterglass
20JP4A										
20MP4	Rect	Glass	22 1/2	Magnetic	70 diag	Single	Cavity	14	E 50 to 350 v	
21AP4	Rect	Metal	22 1/2	Magnetic	70 diag	Single	Rim	14	M 105 ma F	Etched face
21DP4	Rect	Metal	22 1/2	Magnetic	70 diag	Single	Rim	16	E 2-3 kv	Etched face C
21EP4	Rect	Glass	23	Magnetic	70 diag	Single	Cavity	16	M 95 ma F	Cyl face Cyl face C
21EP4A										
21EP4B										
21FP4										
21FP4A	Rect	Glass	23	Magnetic	70 diag	Single	Cavity	14	E 50 to 350 v	Cyl face Filterglass C
21GP4	Rect	Metal	22 1/2	Magnetic	70 diag	Single	Rim	18	Automatic	Filterglass C
21KP4	Rect	Glass	22 1/2	Magnetic	70 diag	Single	Cavity	14	Automatic	Cyl face Etched face
21KP4A										
21MP4	Rect	Metal	22 1/2	Magnetic	70 diag	Single	Rim	14	E 50 to 350 v	
22AP4	Round	Metal	23 1/2	Magnetic	70	Single	Rim	14	M 105 ma F	Filterglass
22AP4A										
24AP4	Round	Metal	24 1/2	Magnetic	70	Single	Rim	16	M 100 ma F	Filterglass Filterglass
24AP4A										
24BP4	Round	Metal	24 1/2	Magnetic	70	Single	Rim	14	E 50 to 270 v	
27AP4	Rect	Metal	21 1/2	Magnetic	90 diag	Single	Rim	15	E 40 to 300 v	Filterglass
30BP4	Round	Metal	23 1/2	Magnetic	90	Single	Rim	27	M 95 ma F	Filterglass

sonably sized table model receivers. True, the increase in deflecting angle means greater deflection currents, but this has been met in large measure by improved operating efficiency in the circuit components. Electrostatic deflection, under the same conditions, would entail a significant rise in cost. Further, the use of electromagnetic deflection and focusing simplifies the internal construction of the cathode-ray tube. This may not lower the overall cost of the set because deflection and focusing coils are needed, but it does lower the replacement cost of the cathode-ray tube.

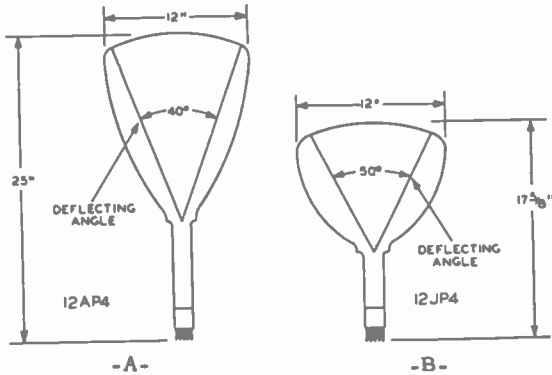


FIG. 9.31. (A) Pre-war television tubes. (B) Current television tubes. By increasing the deflecting angle, the overall tube length has decreased.

In the small 3NP4 and 5TP4 tubes, the deflection is electromagnetic instead of electrostatic for several reasons.

1. The second anode accelerating voltage is 27,000 volts while the first anode voltage is 6,000 volts. Since each tube is small, the separation between the deflection plates (with their average potential of 27,000 volts) and the first anode would be small, necessitating special precautions to prevent corona and arc discharges.

2. The system required to develop the extremely large deflection voltages needed with a 27,000-volt accelerating potential would involve a cost entirely out of proportion to the cost of the rest of the set. Special tubes, 27,000-volt isolating capacitors, and additional amplifying stages are only some of the major items necessary.

3. Electromagnetic deflection not only permits a simplification in tube design, but requires no more power than that needed by a conventional 10-inch tube.

Metal-Cone Tubes. A metal-cone tube, Fig. 9-32, contains a metal-cone envelope, producing a final product which weighs approximately 60 per cent less than an all-glass tube. When metal-cone tubes are used, it is important to keep in mind that the metal cone itself serves as the second anode,

from 9,000 volts upward, and therefore should not be touched while the receiver is in operation. After the power has been turned off, the metal cone should be grounded before human contact is made with it.

Rectangular Screens. It had been recognized for many years that a rectangular image on a circular screen is wasteful not only of screen area, but of cabinet space as well. The obvious solution was a rectangular screen



FIG. 9.32. A metal-cone picture tube.

dimensioned in the standard 4 to 3 ratio of the transmitted image. Until quite recently, manufacturing difficulties and obstacles presented by the rectangular tube shape prevented their mass production, but these have now been overcome and rectangular tubes are widely used. A rectangular picture tube, the 20CP4, is shown in Fig. 9-33.

To have the screen size numbering system of rectangular tubes conform to that employed for circular screens, it is standard procedure to use the diagonal length of the rectangular screen as the screen size. Thus, the 20CP4 shown in Fig. 9-33, has a diagonal length (from one corner to the opposite corner) of 20 inches. The reader can, from elementary geometry, then figure out that the screen sides will be 12 inches high and 16 inches wide.

Nomenclature of Cathode-Ray Tubes. In an effort to standardize the nomenclature used for cathode-ray tubes, the following system has been adopted:

1. The first number, whether of one or two digits, will represent the diameter of the screen.
2. The letter P, with the number following it, will indicate the type of fluorescent screen that the cathode-ray tube contains.

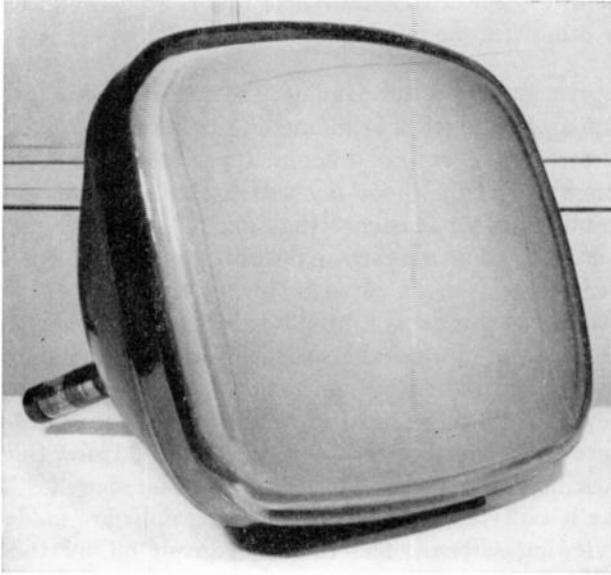


FIG. 9.33. A 20-inch picture tube.

3. Any other letters found between the first number and P will be used to distinguish between tubes that may be just as large as each other, with identical screens, but possessing other differences. These differences may consist of the addition of another element (for example, an accelerating grid), another shaped-bulb, another tube socket, etc.

4. The letters A, B, C, and D may frequently be found following the screen type (P4). These indicate that the screen glass of the tube is either of special texture or construction or that it has been specially treated to help improve picture contrast. When the letter A is used (as in 16HP4A), it frequently indicates that a special light-absorbing glass (known commercially as Grey Glass, Filterglas, or Teleglas) is being used for the screen plate. When the letter B is found, it indicates that the screen face has been frosted. This tends to diffuse the reflections of bright lights in the room and thereby lessen their annoyance to the viewer. This frosting, in combination with Filterglas, helps to improve picture contrast considerably.

Occasionally the letters C and D will be found. In the 19AP4C, for example, Grey Glass is used and, in addition, the fluorescent side of the screen is coated with a thin layer of aluminum. This tends to increase the light output considerably. In the 19AP4D, a clear frosted glass is used. This is in distinction to the 19AP4B where a frosted Grey Glass is employed.

Some variations of the above letter additions (A, B, C, D) will be found because it is becoming standard practice to incorporate some of these additional features in the newer tubes. Where this is occurring, the letters may be omitted or otherwise modified.

With the preceding rules as a guide, the 16DP4A is a tube having a screen approximately 16 inches in diameter, made of Filterglas, and with a P4 fluorescent screen. There are different types of screens, depending upon the purpose to which the cathode-ray tube is put. In television, a white trace is desirable with a persistence that might be termed medium. This would be labeled P4. For oscillographic work, either a P1 or a P2 screen, having a green retrace, might serve better. Green is used here, in preference to white, because it gives a brighter trace with the same accelerating potentials. This feature is desirable as it is often necessary to observe oscillogoscopes where the surrounding illumination is high. For long periods of viewing, however, white results in less strain on the eyes.

The P1 trace has a shorter persistence than the P2 trace, the latter being employed in oscillogoscopes where transients are to be viewed. The P5 fluorescent coating is especially suited where photographs are made of the pattern on the viewing screen. The radiation given off by this fluorescent screen is blue in color and of short persistence. Other types of screens have different properties that make them suitable for other particular applications (see Table 9-1).

Projection Television Systems. In the picture tubes discussed thus far the images swept out on their screens are designed for direct viewing by the set owner. Under these circumstances the size of the picture will be governed by the size of the screen. It is possible, however, to use this image to produce a larger picture by projecting it through a lens system onto a screen. Projection systems have been designed specifically for television and have been commercially used to a limited extent.

In the first system that appeared, the image was formed on the screen of a small cathode-ray tube and then enlarged, using several projection lenses similar to those employed in motion-picture equipment. (See Fig. 9-34.) The chief disadvantage of this method is its poor efficiency. Specially treated lenses, having a large aperture and good transmission of light, deliver to the viewing screen only about 6 per cent of the light present on the screen of the cathode-ray tube. Consequently, even with intense images

developed on the cathode-ray screen, the final image is bound to be dim and difficult to see.

As an illustration, consider the 5-inch projection tube, with its 3- by 4-inch image. The total area here is 12 square inches. If the image is enlarged to fill a screen 15 by 20 inches, the total area becomes 300 square inches, and the light which was originally concentrated in an area of 12 square inches is now spread over an area of 300 square inches. The brightness is thereby reduced by the ratio of 300/12 or 25 to 1. This, of course, assuming 100 per cent transmission. If now we take into account that the transmission efficiency is only 6 per cent, then the brightness is decreased by a factor of 400 to 1 rather than 25 to 1.

Increasing the brightness of the image developed on the tube screen would be one partial solution to this problem, but the cost involved can readily place the equipment in the same expensive class as large cathode-ray tubes. A second approach to the problem, and one which had greater possibilities, was to increase the efficiency of the lens system.

The first attempt in this direction was directed toward the lens system shown in Fig. 9-34 and several difficulties were immediately encountered. In order to capture as much light as possible being emitted from the screen, the lens should be as large as possible. However, as the lens becomes larger, the number of distortions or "aberrations" multiply. (There are principally six aberrations that must be corrected; namely, chromatic aberration, spherical aberration, coma, astigmatism, curvature of field, and distortion.) To correct for these aberrations in a fairly satisfactory manner, we require several lens elements, say possibly 3. However, in a wide-angle system, more than 3 lens elements should be used. As more elements are added, the cost of the system rises. The greater the number of lenses in a system, the greater the total loss of light at the intersecting surfaces and the lower the overall efficiency. Finally, with large lenses, the elimination of some aberrations is accompanied by a greater accentuation in others.

This was the impasse that faced television engineers until the adaptation of the Schmidt reflective optical system to television. Schmidt, an instrument maker at the Hamburg, Germany, Observatory, invented his optical

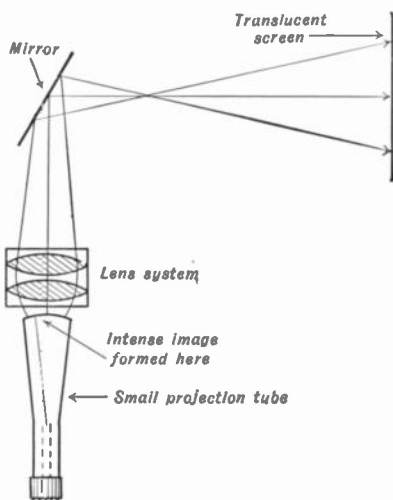


FIG. 9.34. Projection system to obtain large images.

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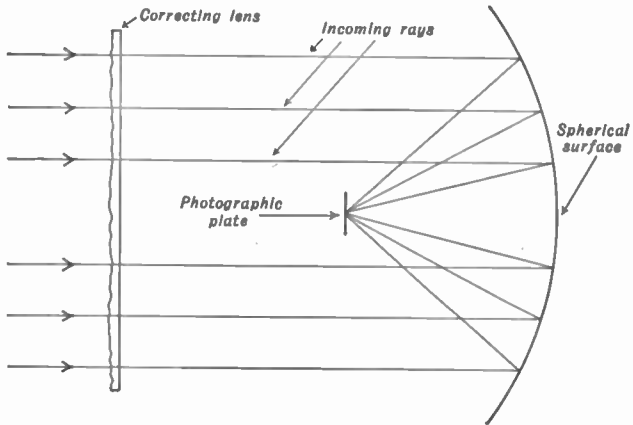


FIG. 9.35. Original arrangement of the Schmidt lens system.

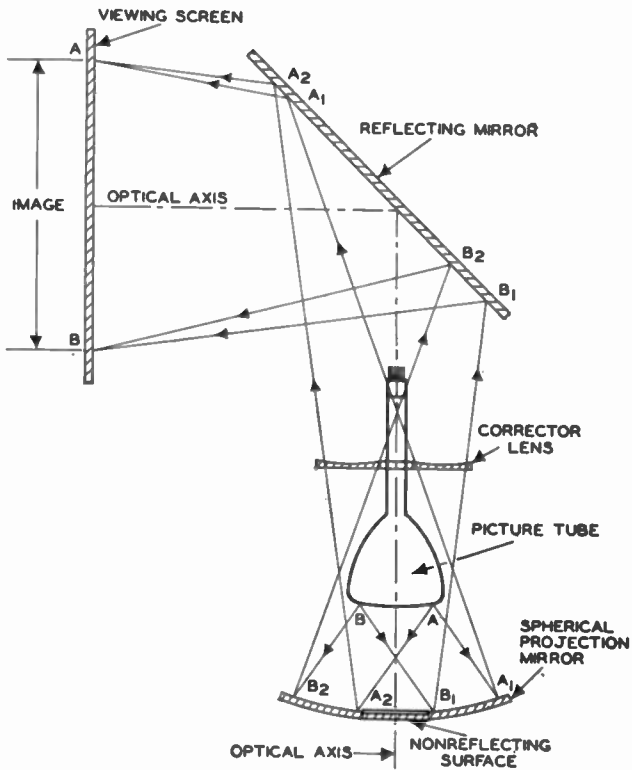


FIG. 9.36A. The adaptation of the Schmidt optical system to G.E. and RCA projection television receivers.

Courtesy RCA

system in 1931. The system, originally designed for astronomical telescopes, was built around a large spherical reflecting surface or mirror. (See Fig. 9-35.) Use of this type of reflecting surface offered several immediate advantages:

1. Mirrors are completely free of chromatic aberrations.
2. Under comparable design of focal distance and diameter, a spherical mirror has a spherical aberration one-eighth that of a single lens. The need



Courtesy RCA

FIG. 9.36B. A projection television receiver. The image from a small projection-type tube is enlarged to permit comfortable viewing by many people.

for correction is still present, but the problem is now considerably simplified.

3. By placing a small aperture or opening at the center of curvature of a spherical mirror, all the monochromatic aberrations (except spherical aberration) are eliminated.

We are thus left with essentially two problems to solve. One is spherical aberration and the other is curvature of the field. The latter is quite simply resolved by curving the screen of the projection tube to suit the curvature of the mirror. To eliminate spherical aberration, a special correcting lens is employed. The correcting lens introduces into the beam an amount of spherical aberration which is equal to that introduced by the mirror but is opposite in sign. As a result, the two neutralize each other, effectively removing the last great defect of the spherical mirror. With this lens in place, we have optical system possessing an efficiency of 25 per cent with magnifica-

tion of 5. Compare this to the meager 6 per cent obtainable using the refractive lens system of Fig. 9-34.

For use in television receivers, there are several modifications of the original Schmidt lens system as designed for astronomical use. In RCA and G.E. projection receivers, the optical mirror is mounted at the bottom of the cab-

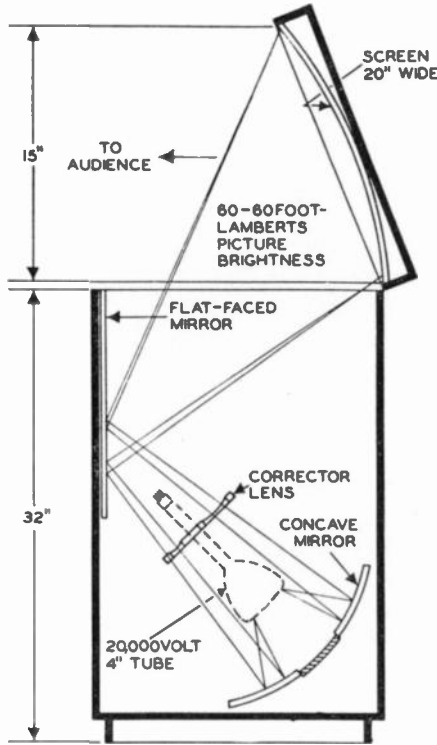


Fig. 9.37. The optical system in the Philco projection television receiver.

inet with its axis vertical, projecting the image straight up and onto a flat mirror inclined at 45 degrees to the beam of light and throwing the image onto a translucent screen. (See Fig. 9-36 A and B.) The throw or distance between the correcting lens and the viewing screen will depend upon the diameter of the correcting lens and the spherical mirror. To increase the size of the projected image, the distance or throw must increase, necessitating large cabinets, a larger mirror, and a larger correcting lens. Eventually the optical system becomes awkward and bulky. A compromise is thus necessary between the size of the final image and the cost and size of the set. For each different throw or magnification, we require a different correcting lens. This, it will be noted, differs from a refractive lens system where the size of the

projected image can be increased merely by changing the position of the lenses and the screen.

In Philco projection television receivers the arrangement of the optical system components is slightly different as shown in Fig. 9-37. The distance between the corrector lens and the screen is $33\frac{1}{2}$ inches, and variations of plus or minus 10 per cent are permissible without leading to any serious deterioration of image quality. The reflecting plane mirrors in all these optical systems are front-surfaced mirrors to prevent ghosts which would occur from reflections at the surface of the glass of a rear-surfaced mirror.

The translucent screens upon which the final image is projected has, in itself, directional properties which concentrate the light in certain desired directions. If the screen was a perfect diffuser of light, it would produce illumination which was equally visible all over the room. Graphically this could be shown as indicated in Fig. 9-38. At all points throughout a hemisphere whose center coincides with the screen equal illumination would be received from the screen. Since many of the extreme angles of this hemisphere are never (or very seldom) used for viewing due to foreshortening of the image, it is advantageous to concentrate the light that would normally go to these points toward those angles that are most used for viewing. To do this, the translucent screen is made directional in the vertical and horizontal directions. Not only does this cause the final image to be brighter than it would be using the perfect diffuser screen, but it also presents the added advantage of greatly reducing the susceptibility of the screen to any stray light from lamps located in the viewing room.

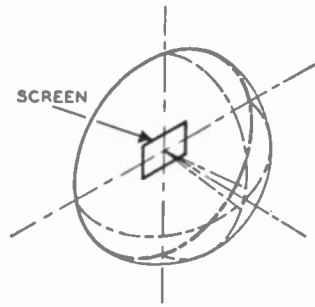


FIG. 9.38. The distribution of light from a non-directional screen is uniform throughout a hemisphere whose center coincides with the screen.

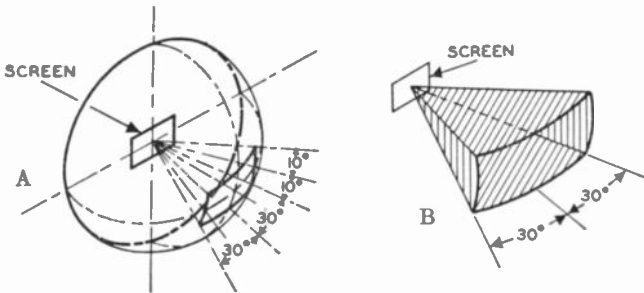


FIG. 9.39. (A) The vertical and horizontal directivity of the Philco viewing screen. (B) The 60° horizontal viewing angle.

In the Philco projection receiver, the screen is designed to have a viewing sector which extends 60 degrees horizontally and 20 degrees vertically. (See Fig. 9-39.) To achieve this directivity, the screen contains a large number of vertical grooves, random-shaped. These vertical grooves are responsible for the horizontal directivity of the screen. To achieve the 20-degree vertical directivity, the screen surface is made concave. (See Fig. 9-37.) The screen, in addition to its directional properties, also possesses a great many minute or lenticular elements, each of which redistributes or diffuses the light reaching it uniformly throughout the desired sector. The overall brightness of this screen is about 50 foot-lamberts.

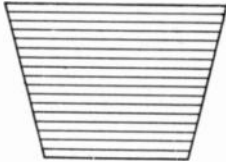


FIG. 9.40. A trapezoid.

Since the optical system is mounted at an angle and projects on the screen at an angle, a rectangular image projected from the face of the picture tube would appear on the screen as a trapezoid (the image would have sloping sides with the top larger than the bottom). (See Fig. 9-40.) On the other hand, by projecting a trapezoidal image from the tube, we obtain a rectangular image on the screen. This latter method is the one used in the Philco system.

Formation of the trapezoid pattern is achieved by applying a magnetic field at right angles to the electron beam. To produce this magnetic field, two oppositely polarized permanent magnets are mounted opposite each other on the end of the projection tube. (See Fig. 9-41.) An iron pole piece,

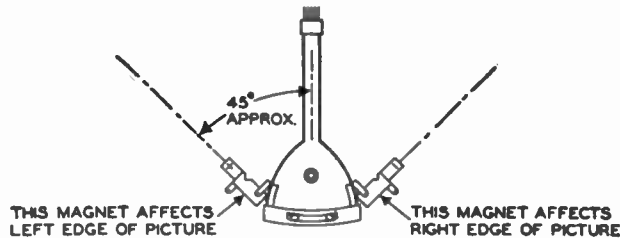


FIG. 9.41. The special magnets mounted on the cathode-ray tube in Philco projection receivers.

curved to fit the sides of the tube, is attached to each magnet and is used to produce a strong field for deflecting the electron beam upward near the tube face. The oppositely polarized ends of the magnets farthest from the tube face cause a lesser and downward deflection of the beam before it is deflected upward. The result is the same as that which would be produced if the face of the tube were tilted inward; the distance the beam travels to the bottom of the image is reduced and the distance to the top is increased. This creates the desired trapezoidal pattern. The magnets are adjusted for the proper

keystoning pattern by moving them toward (parallel to the tube neck) or away from the tube (at right angles to the tube neck).

In the RCA and G.E. projection receivers the additional bar magnets are

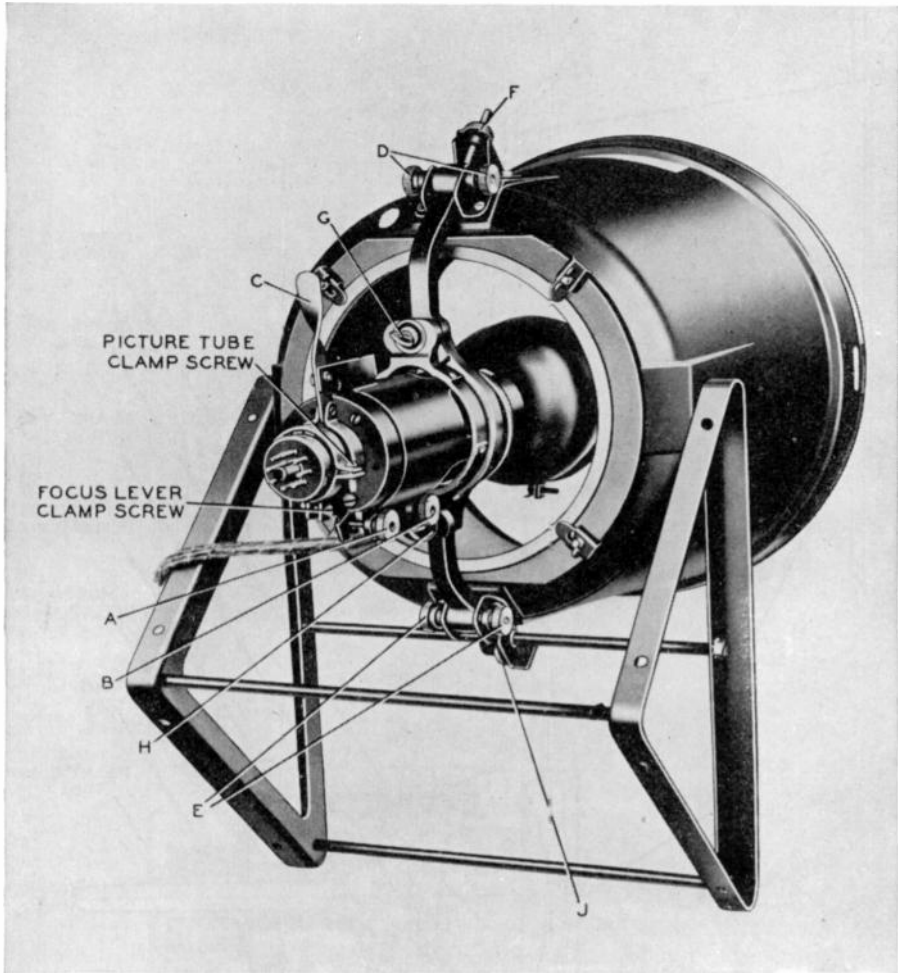


FIG. 9.42. The "optical barrel" which contains the projection tube, the spherical mirror, and the correcting lens. Points lettered A through J indicate physical positioning adjustments. *Courtesy Philco*

not required because the translucent screen is not slanted and because it is perpendicular to the axis of the optical system. Hence, a rectangular image on the projection tube face appears as a rectangle on the screen. The translucent screen used in these sets is composed of two lucite sheets with a partial

diffusing layer between them. The back sheet has a fresnel lens molded into its rear surface. The front sheet has vertical ribs molded into its outer surface. The fresnel lens functions to concentrate the light into a narrow view-

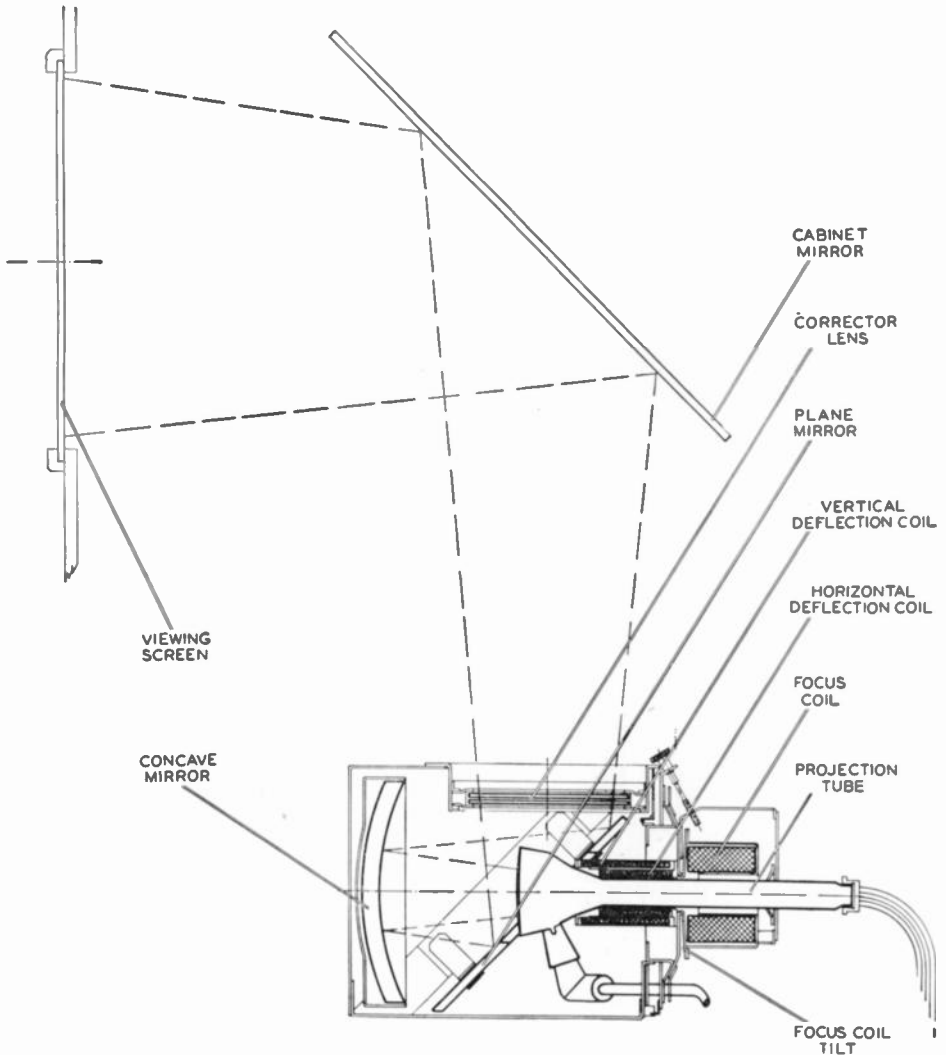


Fig. 9.43. The various components of the optical unit of the North American Philips projection system.

ing angle. The vertical ribs act to increase the horizontal directivity above that obtained with a flat surface. The diffusing layer eliminates interference patterns between the fresnel lens and the vertical ribs. The screen and

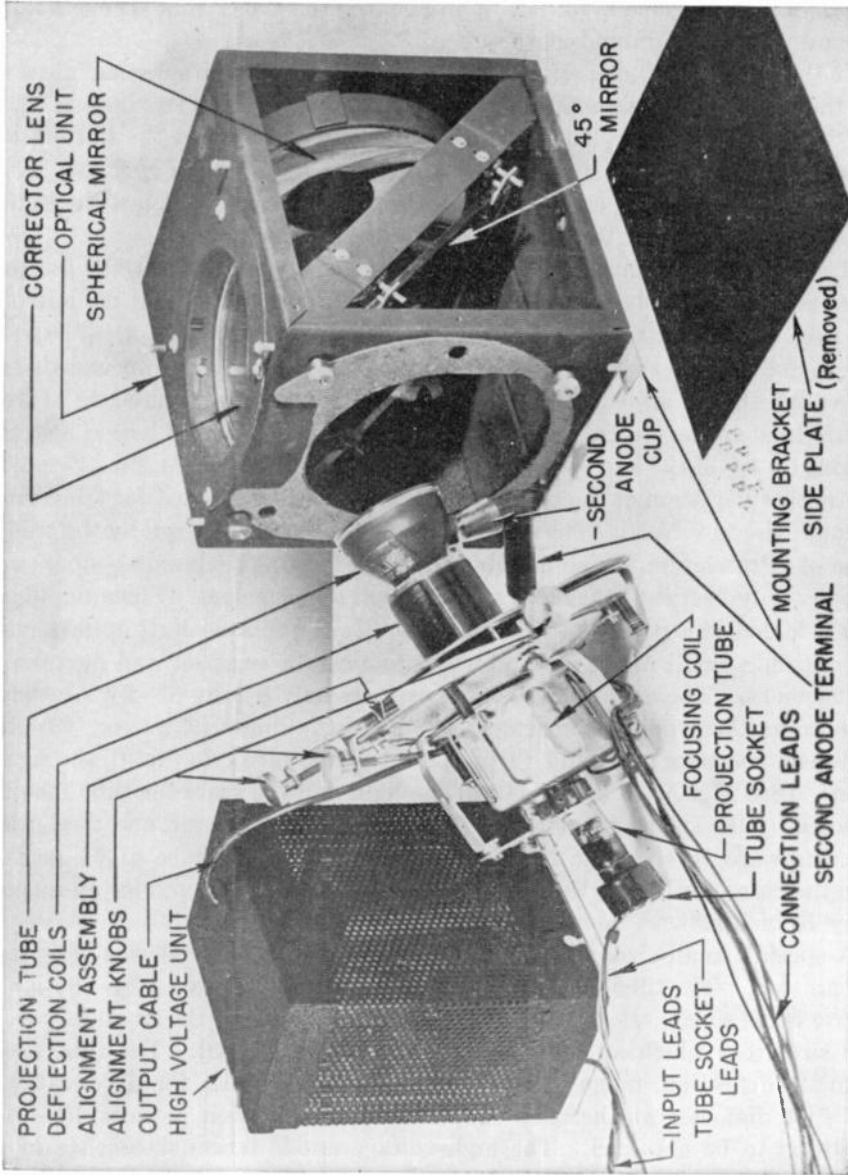


Fig. 9.44. The complete assembly, including power supply, of the North American Philips projection system.

lens combination gives a gain of approximately five over that which would be obtained from a ground-glass screen.

In the RCA and Philco receivers, the projection tube, the spherical mirror, and the correcting lens are all mounted in what is known as an "optical barrel." (See Fig. 9-42.) The spherical mirror is mounted in the bottom of the optical barrel and held against a knife edge by 3 sets of springs. The spring pressure is strong enough to hold the mirror securely against the knife edges, but not sufficient to distort the spherical mirror.

The 5TP4 is held in a retaining ring within the optical barrel, the face of which is approximately half way between the correction lens and the mirror. The tube is adjusted to position by controls on the side of the optical barrel. Optical focusing is accomplished by moving the picture tube up and down vertically. Other adjustments are needed to secure proper centering of the picture tube on the optical axis of the mirror. The correction lens is held in the ring on top of the barrel and secured by three spring fingers.

Another variation of the Schmidt optical system as adopted for television is that devised by North American Philips. The system, known by the trade name of "Protelgram," is an adaptation of the "folded" Schmidt system and occupies only half the space of a conventional arrangement. Since the light path is folded, it is possible to mount the projection tube with its optical system within a small metal box, thereby producing a compact and dustproof arrangement. The actual metal case measures only $8\frac{1}{2}$ by $8\frac{1}{2}$ by 9 inches. It contains three optical elements: (1) a 6-inch spherical mirror, (2) an aspherical corrector lens, and (3) a special plane mirror to "fold" the light beam. (See Figs. 9-43 and 9-44.) The light emitted from the tube face is gathered by the spherical mirror, reflected to the plane mirror, and then projected upward through the corrector lens. A throw distance of 31 inches from the corrector lens to the viewing screen is required to produce an image 12 by 16 inches.

A special, small-sized cathode-ray projection tube (3NP4) was designed for this unit. The tube screen diameter is 2.5 inches, and a 1.4 by 1.86-inch picture is obtained. It appears that 2.5 inches constitute the smallest practical size from which an enlarged image can be obtained. The tube uses magnetic deflection, magnetic focusing, and 25,000 volts for acceleration. The spot diameter at the tube face is 0.003 inch which permits 450-line resolution to be obtained. The high-voltage anode terminal consists of a button in a glass cup sealed to the cone of the tube. The glass cup lengthens the external leakage path from the high-voltage contact to the coils, thereby minimizing any tendency for arc-over to occur. The outside of the cone and part of the neck are covered with a conductive coating that can be grounded. This outer coating, together with the conductive coating inside the tube, forms a $300\text{-}\mu\text{f}$ capacitor which can be utilized for filtering of the high volt-

age. The neck of the tube is quite narrow and, in conjunction with a deflection angle of only 40 degrees, permits full deflection to be achieved using only as much deflection current as ordinarily supplied to a 10BP4 direct-viewing tube operating at 9 kv.

QUESTIONS

1. Indicate briefly how a cathode-ray tube differs from a conventional pentode.
2. Which elements of the cathode-ray tube are contained in the first lens system?
3. Explain what occurs to the electron beam in the first lens system.
4. Which elements are contained in the second lens system? What occurs to the electron beam in this section of the electron gun?
5. What is an electric equipotential line?
6. What type of electrodes is employed in the electron gun? Why?
7. Explain, with illustrations, how the deflection plates bend the electron beam.
8. Name all the basic elements of an electrostatic deflection cathode-ray tube.
9. What is the purpose of an aquadag coating inside the walls of the cathode-ray tube?
10. Why is balanced deflection preferred over unbalanced deflection? List the advantages of each method.
11. Draw the circuit of a balanced electrostatic deflection system.
12. Why are vertical and horizontal centering controls necessary?
13. Show how centering controls are placed in the circuit.
14. Define deflection sensitivity and deflection factor. How can one be converted to the other?
15. Explain what happens when an electron enters a magnetic field.
16. Explain how an electron beam is focused magnetically.
17. Explain how a beam is deflected magnetically. Indicate the placement of the deflection coils at the cathode-ray tube.
18. What is the purpose of the fluorescent screen? What is its composition? What is the difference between fluorescence and phosphorescence?
19. What happens to all the light that is generated at the fluorescent screen?
20. Define image contrast. Why does the scattering of light reduce image contrast?
21. Explain halation briefly.
22. How can reflections inside the cathode-ray tube interfere with the image on the screen?
23. What is the reason for coating the backside of the fluorescent screen with aluminum?
24. What do we mean by sticking potential? When does it occur?
25. What is an ion spot? Why does it occur only in tubes using electromagnetic deflection?
26. Illustrate and explain two methods frequently used to remove ion spots.
27. How are cathode-ray tubes identified? Explain in detail the system used.
28. Explain the difference between electromagnetic focus and deflection tubes and electrostatic deflection and focus tubes. Illustrate the difference in their internal construction.
29. Name the elements in an electromagnetic deflection, electrostatic focus tube. What advantage does this tube offer over an all-electromagnetic tube?

30. How are ion trap magnets adjusted? What precautions must be carefully observed?
31. Explain, with illustrations, how a simple lens type of projection system operates.
32. What is the Schmidt optical system? How has it been adapted for television?
33. Why is the Schmidt optical system superior to a simple lens type of projection system?

Chapter 10

TELEVISION RECEIVER POWER SUPPLIES

The power requirements of a television receiver differ from those of radio sets. This is due, in part, to the higher voltages required for the operation of the cathode-ray tube. In a television receiver we are confronted with the task of supplying 300 to 400 volts to the plates of the ordinary tubes, while the picture tube must have voltages that range up to 30,000 volts (for projection tubes).

It is possible to construct one supply for both or to employ two separate supplies. For the latter case, one would be used for the image tube and the other for the remainder of the set. If one supply is decided upon, it must be capable of an output of 400 volts with 200 to 300 ma for the ordinary tubes and up to 30,000 volts at 1 ma for the picture tube. In one case we have low voltage, high current, while in the other instance there is the opposite combination. For one composite unit, then, there would have to be 30,000 volts available with 1 ma and 400 volts at 300 ma. The unit would be bulky, expensive, and quite out of proportion with other sections of the set.

A more effective solution is two separate supplies. The low-voltage, high-current unit could then take its familiar form as in Fig. 10-1. A 5U4G operates as a full-wave rectifier, developing an output voltage of 335 volts. (This value will vary with the set.) In place of one of the filter chokes the field coil of an electrodynamic speaker may be substituted. Voltages of 230, 130, and 90 volts are also made available by using suitable dropping resistances. Note that there are five filter capacitors in this supply. The more filtering the voltage receives, the more nearly it will approach the desired pure d-c state. A poorly filtered d-c voltage in the audio stages will produce an audible hum. The same voltage in the video amplifiers will distort the picture.

Examining Fig. 10-1 further, we notice that the primary contains an on-off switch plus an interlock switch. The interlock is a safety feature de-

signed to prevent unauthorized tampering with the set and also to protect anyone who accidentally removes the back cover of the cabinet. The interlock is so devised that when the back cover to the set (or, in some instances, the back cover of the high-voltage section) is removed, the a-c plug pulls out along with it. For the serviceman who has work to do on the set, there are available extension cords with female plugs which will fit onto the male plug left in the set and serve to reintroduce power into the set.

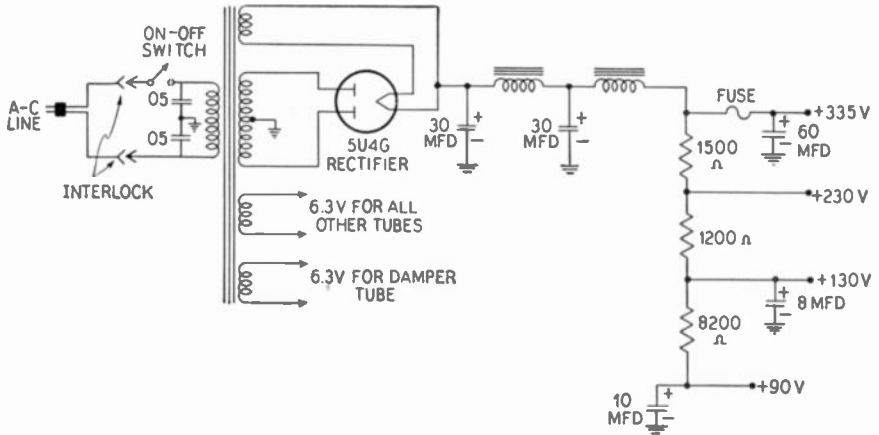


FIG. 10.1. The low-voltage power supply for a television receiver.

Filament power for the tubes are, in this instance, obtained from two 6.3-volt windings on the power transformer. This means that the tubes are connected in parallel. The tubes in the R.F. section of the receiver plus some or all of the tubes in the video I.F. stages generally require R.F. chokes and/or by-pass capacitors in their heater circuits to prevent any of the high-frequency currents present in these tubes from interacting with each other or with other tubes. By the same token, these filters prevent disturbances in other portions of the set from affecting the critical R.F. and video I.F. stages. This precaution is not required in the broadcast A-M radio because of the low-frequency signals dealt with in these circuits. Filament R.F. chokes and by-pass capacitors generally appear when the signal frequencies exceed 10 mc.

A separate 6.3-volt heater winding is required for the damper tube because of the high d-c voltage which is present on the cathode of this tube. To forestall possible breakdown between heater and cathode, the two are frequently connected together and then the heater must be provided with a separate winding in order that the heaters of other tubes are not subjected

to the same high d-c voltage. In most tubes the cathode is only slightly above ground potential and, were we to establish a high difference of poten-

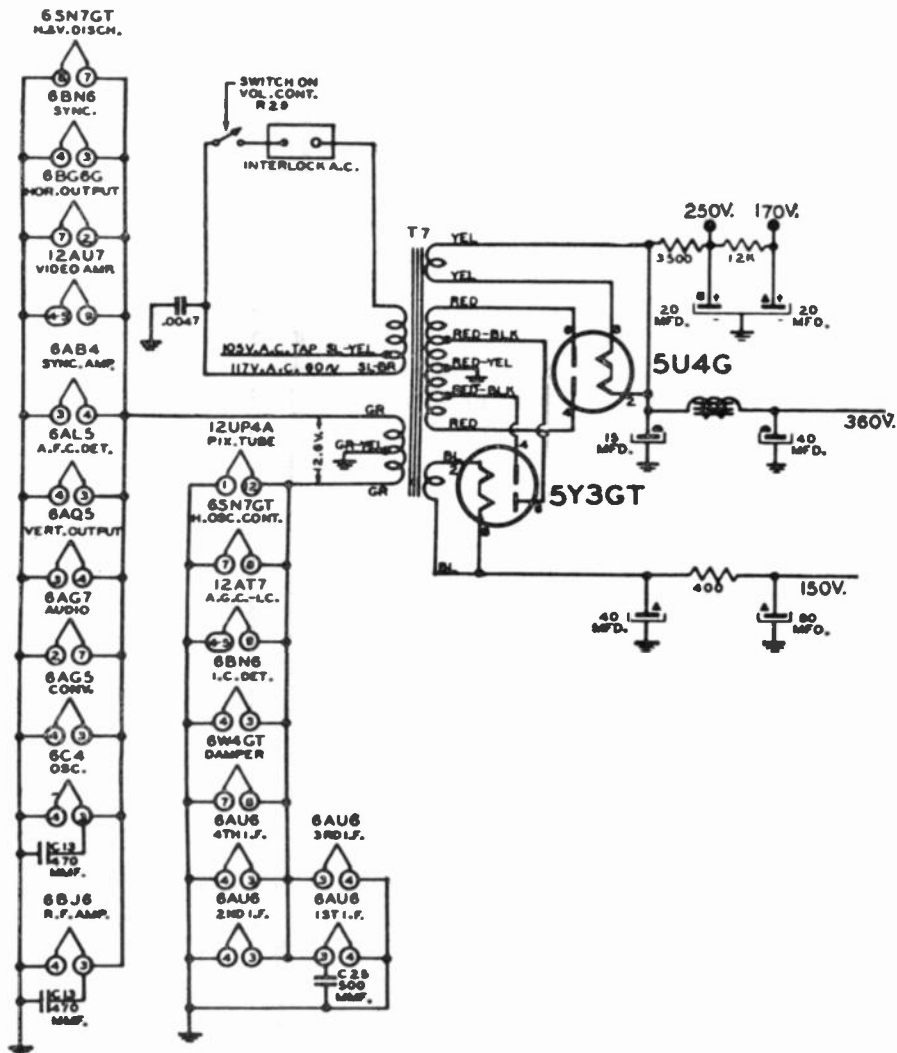


FIG. 10.2. Another low-voltage power supply.

tial between the heater and the cathode, breakdown of the insulation resistance between these elements would frequently occur.

A somewhat more extensive power supply is shown in Fig. 10-2. The supply employs 5U4G and 5Y3GT rectifier tubes, each acting independently of the other. The 5U4, being connected across the full secondary of the

power transformer, develops a d-c output voltage of 360 volts. The 5Y3GT, connected across a smaller portion of the secondary winding, develops only 150 volts. In this lower voltage section, a series resistor serves in place of a filter choke.

Also connected to the 5U4G is a resistance-capacitance filter network which is used exclusively to feed voltage to the audio power amplifier. Values of +250 volts and +170 volts are obtainable from this particular section. Separation of the two filter sections of the 5U4G helps to prevent interaction between those audio and video stages which receive their power from this source.

Filament power for the various tubes is obtained from a single 12.6-volt winding on the transformer. (Note: Although this winding is shown on the same side as the primary winding, it still develops only 12.6 volts.) The center of the winding is grounded, producing two equal halves containing 6.3 volts each. Then, one string of tubes is fed from one half and one string of tubes from the other half. In this way any one tube receives only its required 6.3 volts.

In many television receivers, both negative and positive voltages are required. Voltages of both polarities can be obtained from the same supply by the method shown in Fig. 10-3.

A bleeder is connected across the output terminals of the power supply, and then a ground is placed at some point along the bleeder. That ground point represents zero potential. From that point up, the voltages are positive; from the ground point down, the voltages are negative. By dividing the bleeder into a series of resistors, various positive and negative voltages can be obtained. The most positive voltage that can be obtained from the supply, as shown in Fig. 10-3, is +280 volts. The most negative voltage is -100 volts. This gives us a total of 380 volts over the entire bleeder.

Now, what would happen if we shifted the ground point up, say from point *A* to *B*? We would have less resistance between the ground point and the top of the bleeder string. Since the total amount of resistance in the bleeder is unchanged, the current flowing through it will also remain unaltered. This means that the amount of positive voltage now existing from the ground point to the top of the bleeder string will be less. By the same token, the negative voltage will have increased because of the greater amount of resistance between the ground point and the bottom end of the bleeder. Thus, all we have accomplished by moving the ground point up is to transform some of the positive potential into negative potential.

We could have increased the positive voltage and decreased the negative voltage by moving the ground point down.

To anyone who has been accustomed to the fairly simple A-M sets, especially the midget radios, the large number of filter capacitors frequently

found in television receiver power supplies may appear to be a waste of money. However, there is a good reason for these capacitors. The television picture seen on the screen is developed by the split-second timing of

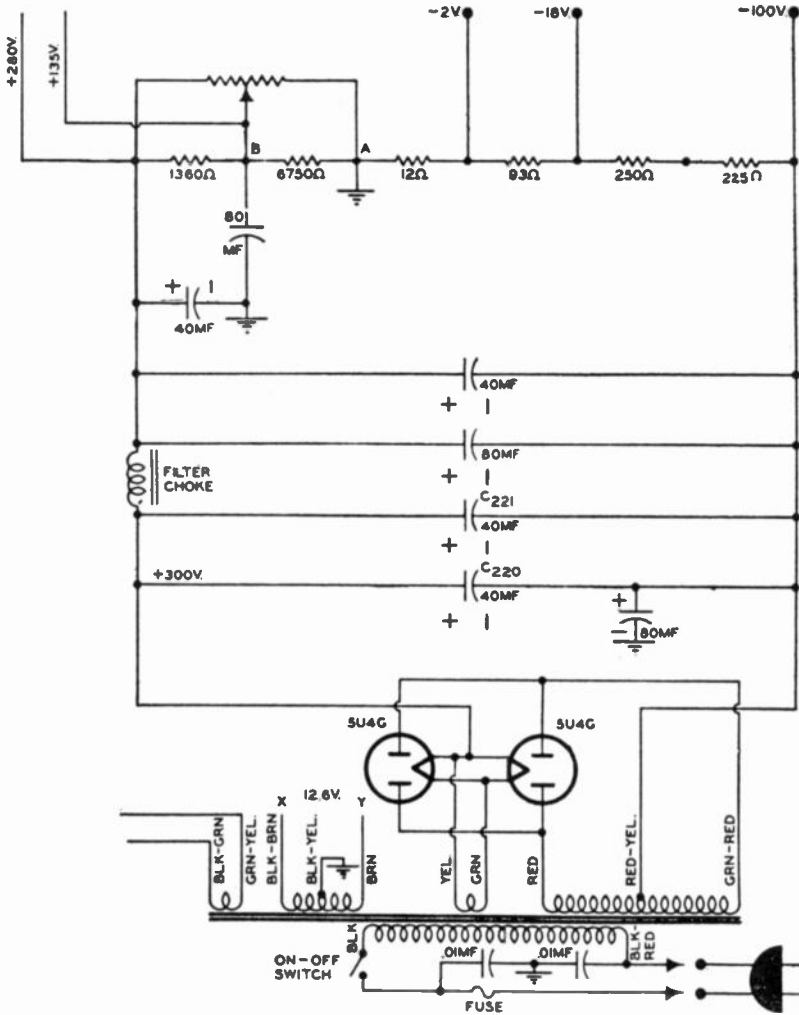


FIG. 10.3. A power supply capable of providing positive and negative voltages.

many circuits and this can be achieved only if each circuit is functioning properly. One way to achieve this goal is by powering the circuits with a well-filtered d-c voltage, filtered not only to remove the alternating component of the line voltage but also to prevent one circuit from affecting any of the other circuits through the common power supply. We want signals

from one circuit to travel to adjacent circuits through the proper channels, and coupling achieved through the power supply is decidedly not desired. It is the purpose of these filter capacitors to see that this undesirable coupling is kept at a minimum.

In nearly all of these power supplies, one or two small 0.01 to 0.05 μf capacitors are generally bridged across the a-c line leading into the power transformer. They serve to remove or filter out interference that may be present on the line due possibly to other electrical devices.

Selenium Rectifiers. Selenium rectifiers are being used in increasing numbers in the low-voltage power supply of a number of television receivers. A typical selenium rectifier is shown in Fig. 10-4. These units take the place of vacuum-tube rectifiers and offer the advantages of compactness, reduced power consumption due to absence of filament power for the rectifier, and generally better voltage regulation than comparable rectifier tubes.

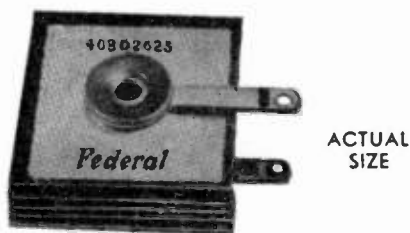


FIG. 10.4. A selenium rectifier.

The selenium rectifiers have two distinct poles, positive and negative. They correspond to the plate and cathode of the vacuum tube. The

positive side, indicated by a red dot of paint, or a plus sign, is equivalent to the cathode. The negative side, indicated by a minus sign or yellow dot, functions as the plate.

Selenium rectifiers are most frequently used as half-wave rectifiers (Fig. 10-5A) or as voltage doublers (Fig. 10-5B). The small protective series resistor in these circuits is used to prevent damage to the rectifiers and filter components from surge currents when the power is first turned on in a set.

A typical application of the selenium rectifier in a television receiver is shown in Fig. 10-6. The two rectifiers are connected to function as a voltage doubler. Briefly this action is as follows: Assume initially that S_2 is disconnected from the circuit. When point B of the power line is more positive than point A , current will flow around the circuit as indicated by the solid arrows. During this time, C_1 will charge up with the polarity as shown. When the a-c input voltage reverses in polarity so that point A is positive with respect to point B , S_1 does not conduct. However, now the a-c voltage adds to the voltage across C_1 , making point X positive with respect to ground by a value equal to the sum of both these voltages. If now S_2 is connected into the circuit, the voltage between point X and ground tends to charge C_2 to the peak voltage existing at X . This is generally in the neighborhood of twice the peak of the applied a-c voltage.

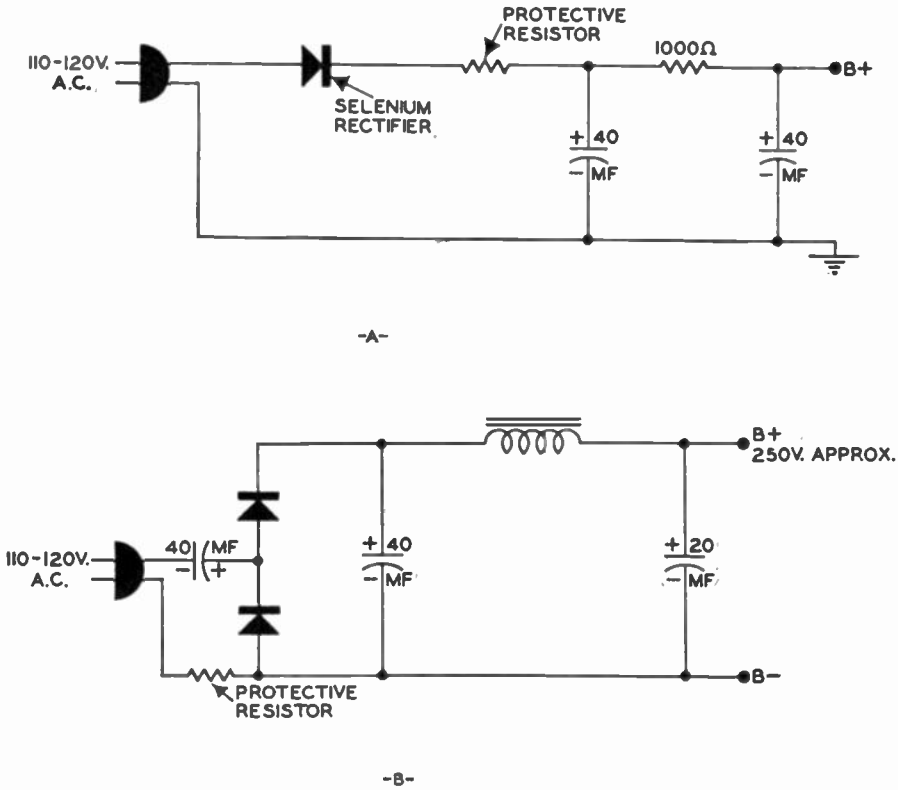


FIG. 10.5. Selenium rectifiers used in a (A) half-wave power supply and (B) in a voltage doubler circuit.

L_1 , C_3 , and C_4 provide filtering for the voltage across C_2 . The use of C_3 and C_4 in series not only permits capacitors with smaller voltage ratings to be used (for economy) but it also provides a 140-volt line for use in the set. Actually the voltage division provided by C_3 and C_4 in series is determined more by the resistance of the circuits shunted across these capacitors than by the relative microfarad values of the units themselves.

Selenium rectifiers are frequently used in transformerless receivers, although transformers can and are being used. When there is no transformer

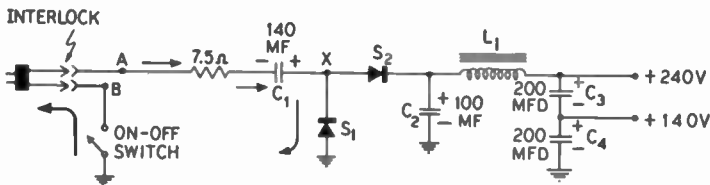


FIG. 10.6. A typical application of selenium rectifiers in a television receiver.

across the circuit. As before, this acts to stabilize the voltage output of the power supply and permit various taps to be made; for example, for the focusing anode and the cathode brightness control. The value of the bleeder resistor depends upon the current drain of the several electrodes of the tube. Since the current drain is small, and the voltage large, the resistance will be correspondingly large, generally several megohms in value. Because of the small current, a low wattage rating will suffice.

Whereas one tapped resistor might conceivably be used, this is never the case. It is more economical to insert small resistors in series, each with

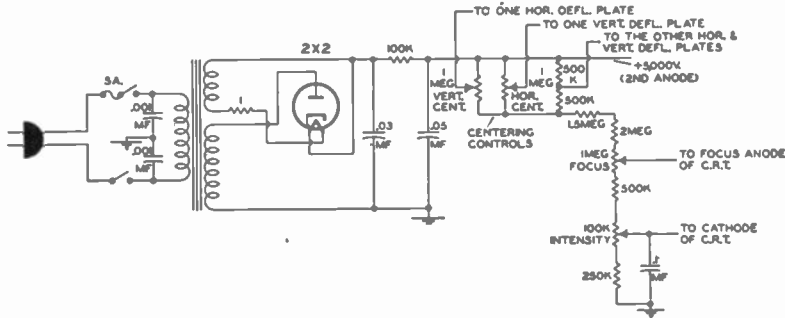


FIG. 10.8. A high-voltage power supply designed in the same manner as the conventional low-voltage power supply. This supply would be suitable for electrostatic picture tubes.

1-watt rating. A disadvantage of using one resistor results from the tendency of a carbon resistor to develop an internal arc if subjected to this high voltage. With several resistors, the voltage across each is proportionately less and arcing does not appear. Potentiometers of suitable value giving the desired voltage variation are used for the focusing control, the brightness control, and the horizontal and vertical positioning controls. A typical high-voltage supply evolved by the foregoing design is shown in Fig. 10-8.

R.F. Power Supplies. As we increase the value of the voltage desired from the high-voltage supply, we find that the cost and the bulkiness of the unit increase, too, but at a much greater rate. In an effort to evolve a more economical solution, research was directed toward other avenues of approach. From these investigations, two power supplies have been developed which offer relief from the cost and bulkiness of the conventional supply. One unit develops R.F. oscillations, sustained by a relatively small 60-cycle power supply. The output of the R.F. oscillator is then rectified, and the high voltage thus is obtained. The other unit obtains its voltage from the inductive kick-back of the horizontal deflection coils.

The first type or R.F. power supply is shown schematically in Fig. 10-9. A 5V4G full-wave rectifier operating from the 60-cycle line supplies the

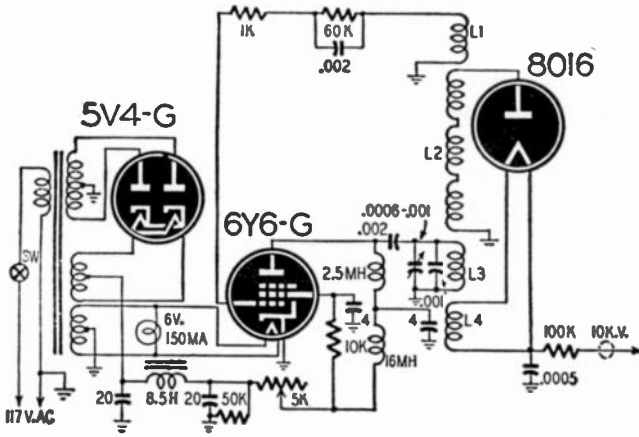


FIG. 10.9. An R.F. power supply. The filament of the 8016 rectifier also operates on R.F.

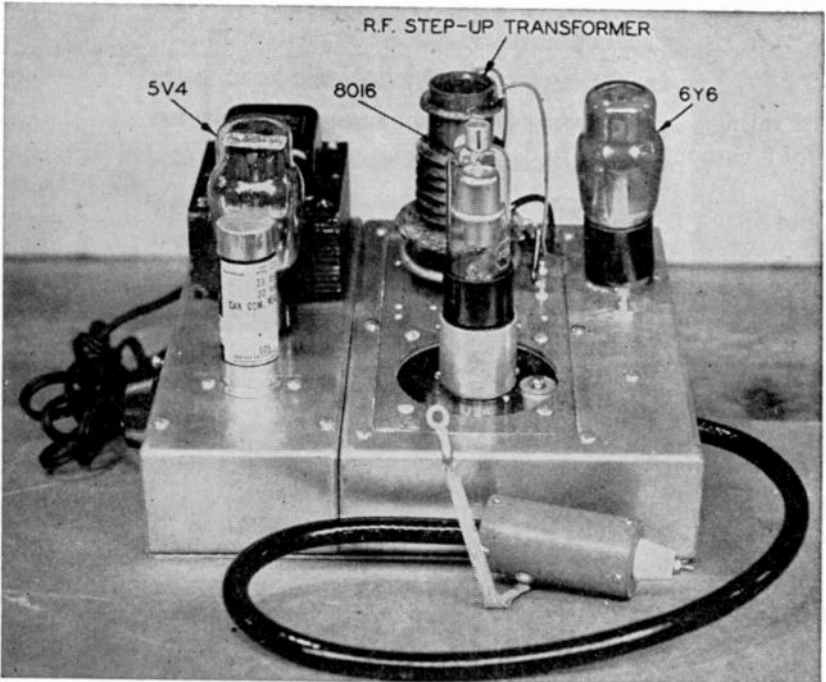


FIG. 10.10A. The entire R.F. power supply.

300–350 volts necessary to drive the 6Y6G oscillator tube. The oscillator itself is a conventional tuned plate, untuned grid tickler coil arrangement. Frequencies of oscillation vary anywhere from 85 kc to 300 kc. The secondary coil, L_2 , which contains more turns than the tuned primary, steps up the low oscillator voltage to approximately 10 kv. Voltage step-up is set at one-half maximum obtainable in order to provide high efficiency and good voltage regulation.

The feedback coil, L_1 , to sustain oscillations is coupled to L_2 , instead of L_3 directly, to obtain greater stability. The oscillator tube is biased for class C operation and hence has relatively low plate voltage loss. This is in the interests of efficiency. The 6Y6 (or 6L6) beam power tube is capable of developing 15 watts of power with 80 per cent efficiency at 350 volts. The screen-grid voltage is made self-regulating by a series resistor. Screen-grid voltage, under operation, varies from approximately 65 volts at no load to 120 volts at full load.

The high-voltage rectifier is a half-wave unit employing a specially designed 8016 or 1B3, tube.* At the high-frequency of the oscillator, a 500- μ mf capacitor and a 100,000-ohm resistor provide sufficient filtering. A photograph of the entire R.F. unit is shown in Fig. 10-10A; in Fig. 10-10B the high-voltage step-up transformer is shown separately.

The Inductive Fly-back Power Supply. The second high-voltage power supply is based on an idea conceived by P. T. Farnsworth about 1930. It was, however, with the advent of television that a good, practical model was evolved. The voltage induced in any inductance is governed by the relationship

$$e_L = -L \frac{di}{dt}$$

where e_L = the induced voltage

L = the inductance

$\frac{di}{dt}$ = the change in current in the time interval of dt .

* This tube is referred to interchangeably as the 8016 or the 1B3. The latter is more common and will be used in this text.

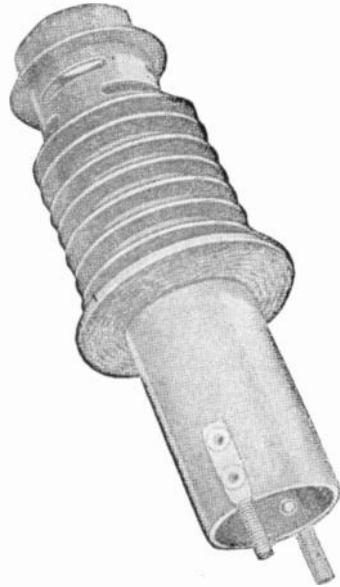


FIG. 10.10B. The high-voltage step-up transformer.

As the time interval, dt , is made smaller, e_L becomes greater. In the horizontal deflection coils, the retrace interval dt is on the order of 7 microseconds and a large voltage is produced. By rectifying the pulse, voltages to 30 kv can be obtained.

A circuit schematic of an "inductive kick" or flyback type of power supply is shown in Fig. 10-11. The horizontal discharge tube, V_{120B} , is

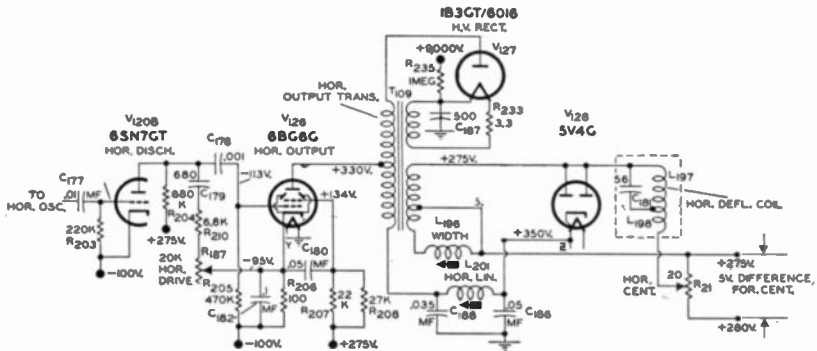


FIG. 10.11. A circuit schematic of an "inductive kick" or fly-back type of high-voltage power supply.

actuated by a positive pulse applied to its grid and discharges capacitor C_{179} , $680 \mu\text{f}$. At the end of the positive pulse, the discharge tube lapses back into cut-off and C_{179} starts to charge for the next cycle again. The deflection waveform produced by C_{179} , R_{210} , and R_{187} in series will, when applied to the deflection coils, produce a linear left-to-right motion of the electron beam across the face of the screen and a rapid retrace. The

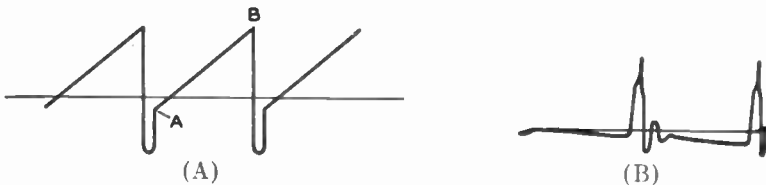


FIG. 10.12. Voltage waveforms in horizontal amplifier circuit of Fig. 10.11.

output tube, V_{126} , is the driving tube for the deflection transformer, the high-voltage rectifier, the deflection coils, and the damping tube. The output tube, a beam tetrode, is controlled by the deflection voltage which it receives from the discharge tube.

The shape of the deflection wave applied to the grid of V_{126} is shown in Fig. 10-12. During the interval labeled A-B, this voltage is rising, causing the current through the horizontal output transformer to rise

steadily, and forcing the beam to move across the screen of the image tube at an even rate of travel. At point B, the grid voltage drops sharply, cutting off the tube's plate current. The magnetic field, which has been steadily building up in the output transformer, collapses quickly, inducing a large positive pulse of voltage in the primary winding. (See Fig. 10-12B.) The plate of the 1B3 rectifier is connected to the top of the primary wind-

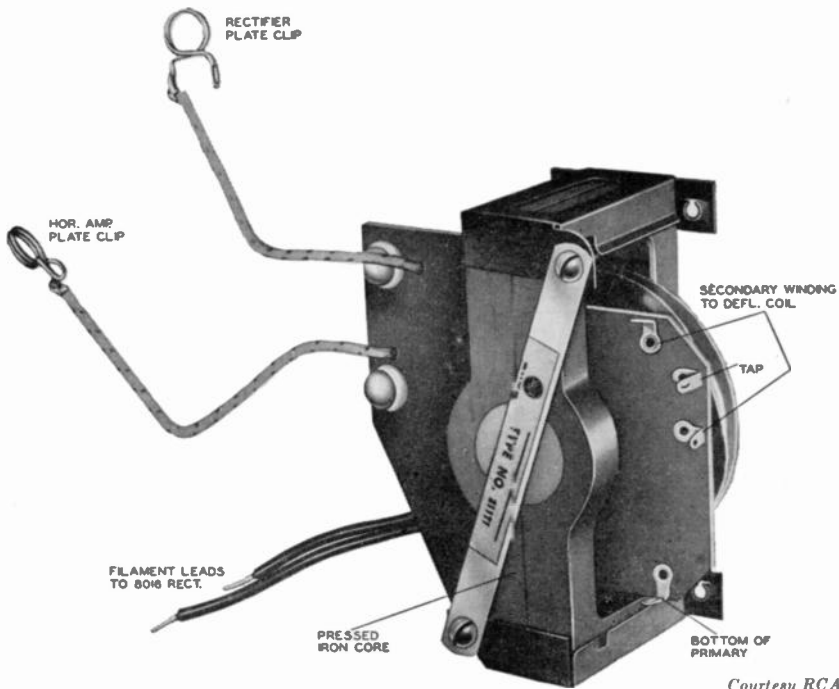


FIG. 10.13. The horizontal output transformer in the high-voltage fly-back system.

ing, and the positive pulse of voltage is rectified, filtered, and fed to the second anode of the cathode-ray tube. Since 15,750 such pulses appear each second, adequate filtering can be achieved using a single 500- $\mu\mu$ f capacitor. Filament power for the 1B3 is obtained from a 2-to-3-turn secondary winding on the same transformer. A photograph of the output transformer is shown in Fig. 10-13.

In both the R.F and flyback methods of developing high voltages, failure of the low-voltage power supply will automatically remove the high voltage from the circuit. As an added precaution in the flyback system, failure of the horizontal sweep system to develop the proper deflection wave will have the same effect. This eliminates the possibility of a

stationary spot appearing on the image screen, with subsequent burning and insensitizing of the screen.

High-voltage Doubling. It is possible to double the output high voltage by connecting two 8016 rectifier tubes as shown in Fig. 10-14.* During the retrace interval, the voltage developed across the full secondary of the output transformer rises sharply to, say, 9,000 volts. This causes V_2 to conduct, and C_2 charges to 9,000 volts (after the first few cycles) with the polarity as indicated. In the longer interval between retraces, C_2 and C_3

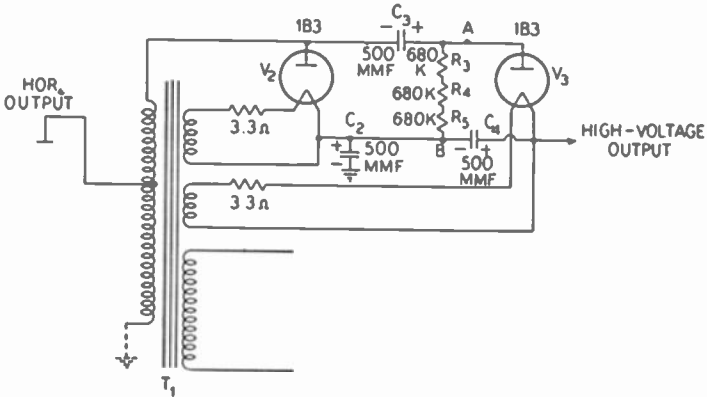


FIG. 10.14. A high-voltage doubler circuit. The bottom end of the primary of T_1 is not actually grounded but is assumed to be for purposes of circuit explanation.

are seen to be essentially in parallel with each other through the primary winding of T_1 and R_3 , R_4 , and R_5 . Hence, C_3 also charges up to the full 9,000 volts.

At the next retrace interval, 9,000 volts once again appear across the transformer. If we pause at this moment and add up the voltages existing between points A and B, we see that the transformer voltage and the voltage across C_3 are equal to 18,000 volts. Opposing this are the 9,000 volts across C_2 . Thus, between points A-B there are 9,000 volts with point A positive with respect to point B. This potential is applied to V_3 , causing this tube to conduct, and C_4 charges to 9,000 with the polarity indicated. If now we use the voltages across C_2 and C_4 , we can obtain an output voltage of 18,000 volts. Losses in the circuit plus the current drain on the power supply by the picture tube usually reduce the output voltage to some value less than twice the peak applied pulse.

* The circuit is shown here in simplified form. Aside from the additional filament winding for the second 1B3, the horizontal output transformer would be similar to the one shown in Fig. 10-11.

Pulse-type High-voltage Supply. North American Philips have developed a pulse-type high-voltage supply for use with their projection system described in the previous chapter. This supply differs from the fly-back system in that a separate pulse generator is utilized, operating at a frequency which is considerably lower than the horizontal sweep frequency.

The circuit of the power supply shown in Fig. 10-15 consists of a blocking oscillator, a driver amplifier, and a three-tube cascaded high-voltage rectifier. The blocking oscillator is conventional in form and operates at a frequency of about 1,200 cycles per second. It produces a saw-tooth

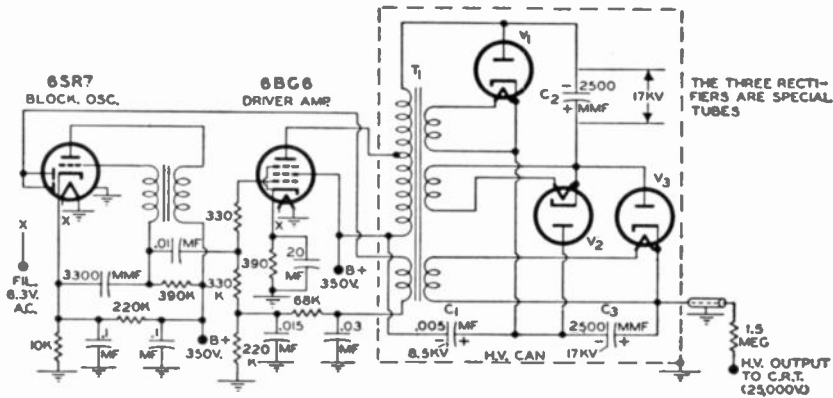


Fig. 10.15. Schematic diagram of the high-voltage unit of the North American Philips projection system.

voltage which is applied to the grid of the following 6BG6 driver amplifier. The grid of this tube is biased beyond cut-off so that its plate current flows only at the peaks of the saw-tooth. (See Fig. 10-16.) At the tip of the saw-tooth, the grid voltage of the 6BG6 drops sharply back into cut-off, stopping the flow of plate current and shock-exciting the transformer. Because of the inductance in the transformer windings and the stray capacitance across them, the system is shocked into oscillations. The values of these components were chosen to produce transient oscillations having a frequency of about 30 kc. These oscillations continue until the next flow of plate current from the 6BG6. This is indicated in Fig. 10-16. In the flyback system, a damping tube placed across the output transformer, damped out all but the first cycle of oscillations. In this circuit no such damping occurs, and the oscillations are present throughout the interval between plate current pulses of the 6BG6. At these moments the conducting 6BG6 loads down the circuit, stopping the oscillations. The sudden stoppage of the plate current at the end of each plate current pulse

shock-excites the transformer into oscillations at its natural frequency of 30 kc.

The voltage developed across the full primary winding of the output transformer is rectified by the three 1B3 tubes to provide an output voltage of 25,000 volts. Since the high voltage is developed here a little differently than in the previous high-voltage systems discussed, a brief explanation follows.

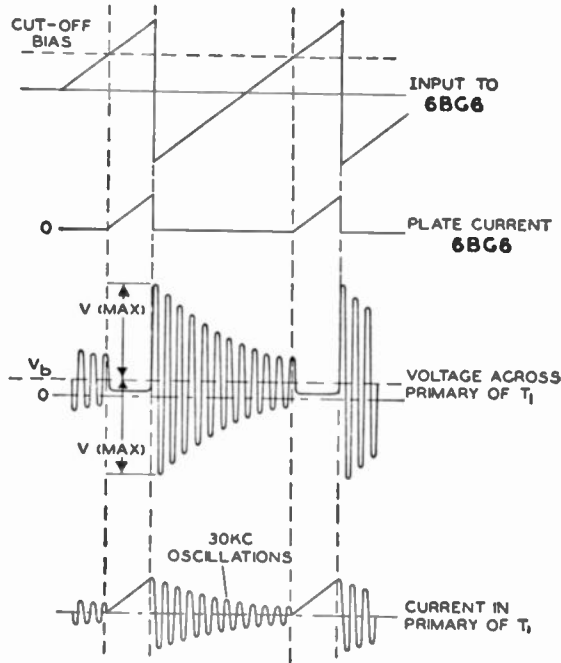


Fig. 10.16. Current and voltage waveforms in circuit of Fig. 10.15.

On the first positive oscillation peak (8,500 volts), current flows through V_1 and C_1 charging the latter capacitor to the full peak voltage with the polarity as shown. On the first negative peak, the voltage across the primary transformer winding adds to the voltage across C_1 to cause V_2 to conduct and charge C_2 to a peak value which is twice the transformer voltage, or 17,000 volts. Current, in this instance, flows from the cathode of V_2 to its plate, through C_1 and the transformer primary, to C_2 and thence back to the cathode of V_2 again. On the next positive peak, C_3 is charged by current flowing through V_3 and C_2 , the transformer primary, and C_1 to C_3 and then back to the cathode of V_3 again, completing the circuit. The voltage across C_2 adds to the voltage across the transformer primary to

feed a positive voltage to the plate of V_3 . Opposing this voltage is the potential across C_1 . The total positive voltage at this instant is $2V$ (from C_1) plus V (from the transformer) or $3V$. Opposing this is V from C_1 . Hence, C_3 receives $2V$ ($3V - V$), or 17,000 volts. By using the voltages across C_3 and C_1 we can obtain an output voltage of 25,500 volts. Actually, the output voltage is somewhat lower than this.

The output transformer contains another winding which is used for regulation purposes. The voltage developed in this winding is fed to the diode section of the 6SR7 tube where it is rectified and applied as a negative voltage to the grid of the 6BG6 by virtue of the voltage drop across the 220 K resistor in the grid circuit of this tube. If the output voltage tends to increase, more negative bias is applied to the 6BG6, reducing its interval of conduction and thereby reducing the amount of energy imparted to the transformer. This will tend to lower the amplitude of its oscillations and reduce the high voltage. On the other hand, a reduction in high voltage will cause less negative bias to appear at the grid of the 6BG6, increasing its interval of conduction and resulting in oscillations of greater amplitude in the transformer.

With this self-regulating arrangement, the voltage output is maintained constant within the limits set for this design. An external low-voltage power supply is needed to provide B+ at 350 volts for the 6SR7 and 6BG6 plus 6.3 volts a.c. for their filaments.

QUESTIONS

1. Draw the circuit for a low-voltage power supply (using vacuum-tube rectifiers) suitable for a television receiver.
2. Indicate how the same power supply can provide positive and negative voltages.
3. Explain the operation of a selenium rectifier voltage doubler.
4. What differences exist in the choice of components between 60-cycle low-voltage and high-voltage power supplies?
5. Explain briefly the operation of an R.F. type of high-voltage power supply.
6. Draw the schematic diagram for a suitable R.F. high-voltage power supply.
7. On what principle does the "flyback" type of power supply operate?
8. What is the purpose of placing a damping tube across the horizontal deflection coils?
9. How does a high-voltage doubler operate?
10. What differences exist between the operation of the circuit of Fig. 10-11 and the circuit of Fig. 10-15?
11. In the circuit of Fig. 10-7, explain how the value of the 60-ohm, 10-watt resistor was obtained. Do the same for the 43-ohm, 1-watt resistor.
12. Why is it possible to use a 500- $\mu\mu\text{f}$ filter capacitor in the flyback power supply when low-voltage power supplies require much larger filter capacitors?

Chapter 11

ADDITIONAL NOTES ON INTERCARRIER SYSTEMS

In general use today are two television receiving systems known as *conventional* and *Intercarrier* systems. The choice of the word "conventional" for the first system is perhaps unfortunate since it conveys the impression that it is the system most widely used, whereas Intercarrier sets would be only occasionally encountered. Such is not the case. The majority of receivers now being manufactured use the Intercarrier principle and the percentage is increasing steadily. Probably the origin of the word "conventional," as applied to television receivers, arose from the fact that the conventional system came first and was in operation for several years before the Intercarrier system was devised. Thus, when the Intercarrier system first appeared it was compared to the then existing receiver which was dubbed the conventional system. To avoid confusion the reader should regard each of the names as simply an identification tag for a particular system. In this way the word "conventional" will lose the significance that it ordinarily might have.

The operating differences that exist between these two systems have already been brought out in Chapter 1. At that time it was mentioned that in the conventional receiver the sound and video signals are separated *before* the video second detector. In the Intercarrier system, this separation takes place *after* the video second detector.

Now that we have followed the video signal from antenna to picture tube we can pause and examine in greater detail the differences between these two systems. It must be puzzling, for example, to note that in conventional receivers great precautions are taken to prevent the sound signal from reaching the video second detector, while in Intercarrier sets these precautions are seemingly ignored. And yet both systems operate. To appreciate the reasons for this apparent contradiction, let us analyze both systems.

Consider first the conventional system, Fig. 11-1. The incoming audio and video signals, after passage through a mixer, are reduced to their I.F.

values. At some point after the mixer and before the video second detector, the sound I.F. signal is removed from the video I.F. system and fed to a separate set of I.F. amplifiers which are peaked to the sound I.F. value. After two or three I.F. stages, the audio signal is fed to an F-M detector where it is converted to audio frequencies. This is then followed by one or more stages of audio amplification after which the signal is powerful enough to drive a loudspeaker. Once the two signals have been separated, trap circuits are placed in the video system to prevent any sound voltage from reaching the video detector and the subsequent cathode-ray tube.

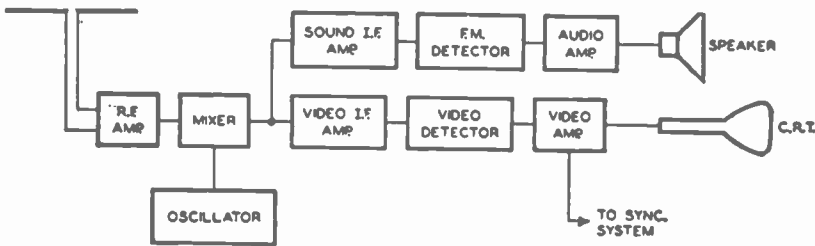


Fig. 11.1. A block diagram of a receiver employing the conventional television system.

While this system will operate successfully when properly designed, it does possess several disadvantages when cost becomes an important design factor. Thus, one of the most important stages in the television receiver, the local oscillator, is also one of its weakest points. We know that the sound I.F. value is the difference between the frequency of the local oscillator and the frequency-modulated sound carrier. If the frequency of the local oscillator should drift, either during the warm-up period when the set is first turned on or during the operation of the receiver, then the difference frequency produced by the mixer will change too. Thus, consider a television receiver tuned to channel 3 and designed for a sound I.F. of 21.25 mc. On channel 3, the audio carrier frequency is 65.75 mc and since the oscillator frequency is above this by an amount equal to the sound I.F., the oscillator frequency will be 65.75 mc plus 21.25 mc, or 87.00 mc. The sound carrier is frequency-modulated, and the modulation shifts the audio carrier frequency plus and minus 25 kc. In other words, the total audio signal for channel 3 can be expressed as 65.75 mc plus or minus 25 kc.

In the sound I.F. system (including the input discriminator coil), the bandpass is about 200–300 kc wide, although generally the discriminator response is linear only for about 150 kc. If the local oscillator should drift by as little as 75 kc, it is possible for the sound I.F. signal to shift to a non-linear portion of the discriminator characteristic, thereby causing audio distortion. Now, a drift in frequency of 75 kc, with the oscillator functioning at 87 mc,

represents a change of less than 0.09 per cent; at 200 mc, the same frequency drift represents a change of 0.03 per cent. Using ordinary components, combined with the station selector rotary switches, it is difficult to prevent the local oscillator from drifting even this amount during warm-up and subsequent operation. It is due to this very situation that oscillators are provided with "fine-tuning" controls which, extended to the front panel, permit the observer to correct for oscillator drift. While such controls remedy the drifting temporarily, it is annoying to have to use this control several times during each set operation. Furthermore, the oscillator frequency is extremely sensitive to tube capacitance, and it is not uncommon to find that the oscillator frequency is altered sufficiently by a tube change to require a complete realignment of the oscillator circuit. Finally, in time, the resistive and capacitive components themselves will change sufficiently in value, even without a tube change, to require a complete realignment also.

The local oscillator is the source of other annoyances. Any inadequate filtering of the B+ voltage used by the local oscillator will cause the oscillator frequency to vary at a 60-cycle or 120-cycle rate. This results in phase and frequency modulation of the local oscillator, which passes through the sound I.F. system and F-M detector and appears as an audible hum. The only remedy for this is complete filtering of the power supply. We also obtain phase and frequency modulation when acoustic feedback from the speaker causes oscillator coils, condensers, and tube elements to vibrate. Vibrations of the cores in permeability or slug-tuned coils or the plates in a condenser can be a difficult annoyance to eliminate.

Solutions to all of the foregoing problems are not unknown, but they usually involve sufficient additional expense to make them unattractive in today's competitive market. The Intercarrier system does offer some remedy to this situation and consequently it has gained wide favor.

Basic Principles of Intercarrier System. In sets employing the Intercarrier system, the various stages follow the sequence shown in Fig. 11-2. A comparison of this illustration with that of Fig. 11-1 reveals that separation of the audio and video signals does not occur until both signals have passed through the video second detector and possibly one or more video amplifiers. Thus, at first glance, every precaution which was previously taken to insure that the sound voltage did not reach the cathode-ray tube is now seemingly ignored. How, then, does the Intercarrier system prevent the sound signals from reaching the picture tube and interfering with the image presentation? The answer is given in the paragraphs that follow.

In Fig. 11-2 it is seen that the audio and video signals are received by the R.F. stages of the receiver, lowered in frequency at the mixer by beating with the local oscillator voltage, and then passed together through the I.F. system.

The carrier I.F. values after passage through the mixer are the same as

they would be using the conventional system of Fig. 11-1. The intermediate frequencies assigned to each carrier are a function only of the local oscillator setting and to this point in the receiver both systems are identical. The first departure appears in the I.F. system. Although the video I.F. response char-

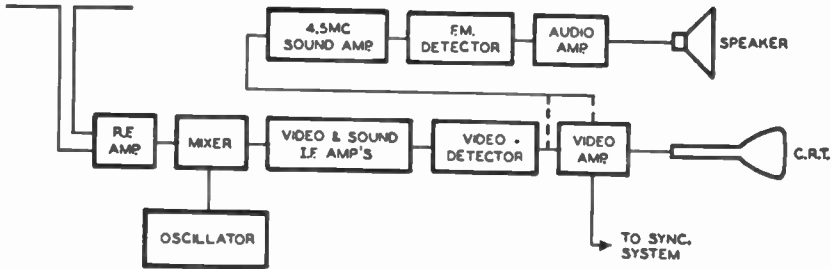


FIG. 11.2. A block diagram of a receiver using the Intercarrier system.

acteristic for a conventional receiver is as shown in Fig. 11-3, it must now be widened to include the sound I.F. signal as well. However, instead of providing the sound carrier with as much amplification as the video signal receives, we provide it with considerably less amplification, and it is specifically because the sound carrier receives this lesser amount of amplification that the system is able to function. This is important, for if the level of the sound carrier is not kept considerably below the level of the video signal, the latter signal becomes completely unusable.

To understand why the sound carrier level must be considerably below the video signal level, let us look ahead in Fig. 11-2 to the video detector. The

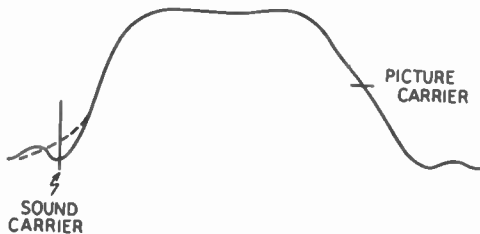


FIG. 11.3. The solid line curve represents the response for the I.F. system in a conventionally operated television system. The dotted curve represents the modification needed to adapt the response to a receiver operating on the Intercarrier system.

video and sound I.F. signals reach this point, in the Intercarrier system, after passage through the I.F. system. Now, a detector (such as used ordinarily for the video second detector) operates basically in the same manner as the previous mixer. In both, a mixing process occurs. Thus, in the mixer stage (which has long been known as the first detector), the local oscillator voltage

beats with the incoming signal or signals to produce the I.F. signal. At the video second detector, a beating or mixing takes place between the video and sound I.F. carriers with the result that a 4.5-mc beat note is produced. In this case we could consider the video carrier as being equivalent to the local oscillator and the audio carrier as the incoming signal. The result of the mixing is a 4.5-mc beat note. In addition, we also obtain all of the 0-4 mc video frequencies from the video signal, produced when the video carrier beats with each of the video sidebands.

The reason for keeping the level of the sound carrier low can now be given. Consider the two carriers shown in Fig. 11-4A and B. Assume the larger

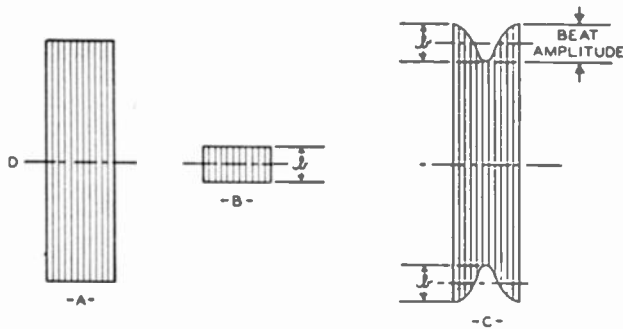


FIG. 11.4. (A) The video carrier. (B) The sound carrier. (C) The resultant wave.

one to be the video carrier and the smaller one to be the audio carrier. If these two signals are mixed together, a ripple will be produced in the resultant wave (Fig. 11-4C) which will have a frequency equal to the difference between the carrier frequencies — in this case, 4.5 mc. As long as the *amplitudes* of the two carriers differ considerably, the amplitude of the 4.5-mc beat note will remain small and unaffected by any amplitude modulation contained in the video carrier. Here is one of the pivotal points in the Inter-carrier television sound system. *By maintaining the sound carrier small in comparison to the video signal, we obtain a 4.5-mc beat note which contains only the frequency modulation of the original sound R.F. carrier and practically none of the video modulation.*

The sound carrier is continually shifting back and forth between the limits of plus and minus 25 kc. In the mixing process in the video second detector this frequency modulation is imparted by the sound I.F. carrier to the 4.5-mc beat note. The 4.5-mc signal represents the difference between the video and sound I.F. carriers. The frequency of the video carrier is fixed, but the frequency of the sound carrier continually shifts back and forth. This changes the value of the beat note. Thus, while we speak of 4.5 mc as the beat note, we actually mean a beat note with a center frequency of 4.5 mc,

but which shifts back and forth about this value to the limits of plus and minus 25 kc. This frequency shifting, remember, represents the sound intelligence of the television broadcast.

Now let us see how the foregoing is employed to make the Intercarrier television sound system workable. When the video carrier (with its amplitude modulation) and the audio carrier are received at the receiver, they have substantially the same amplitude. (The F.C.C. regulations state that the power of the sound carrier must be 50 to 150 per cent of the power of the video carrier when transmitting synchronizing pulses. However, most stations transmit equal sound and maximum amplitude video carriers.) After

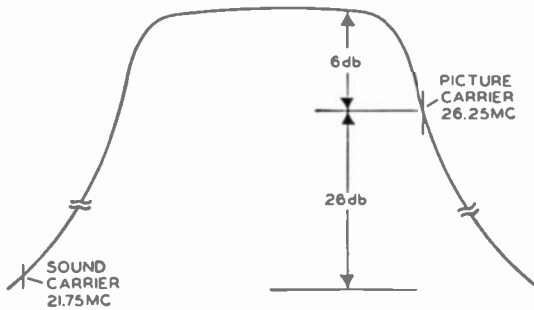


FIG. 11.5. The proper I.F. response curve for Intercarrier receivers.

passage through the mixer stage, the two signals are converted to their I.F. values and are ready to pass through the I.F. system. If we are to keep the degree of amplitude modulation of the beat note signal to a minimum, it is desirable to keep the level of the sound carrier at least 26 db below the peak video carrier level at the second detector. This means that the amplitude of the sound carrier should be reduced to about 5 per cent of the maximum picture carrier amplitude. In order to achieve this, the I.F. system through which both signals pass is designed to have the response characteristic shown in Fig. 11-5. At the frequency of the sound I.F. carrier, the response curve is 26 db down from the video carrier frequency point.

After passage through the I.F. system, the signals are fed to the video detector, where the I.F. frequencies are removed, leaving only the video frequencies, which extend from 0 to 4.0 mc, and the 4.5-mc beat note that is produced when the two carriers beat against each other.

At the output of the video second detector the complete signal, including the 4.5-mc voltage, is usually passed through the video-frequency amplifiers. The sound signal, it must be remembered, was reduced considerably in the I.F. system and thus requires a corresponding increase in amplification following the video second detector in order to bring it back to a usable value. Even after passage through the video-frequency amplifiers, its average level

seldom exceeds one or two volts. Thus, what is done is to transfer this voltage out of the video system into an audio I.F. amplifier, tuned to 4.5 mc, and after one stage of amplification, apply it to an F-M detector for conversion into audio frequencies. Thereafter, one or two audio amplifiers bring the signal to the desired strength for normal operation of a loudspeaker.

To remove the 4.5-mc voltages from the video system, a series resonant trap is connected from the plate of the final video-frequency amplifier to ground. (See Fig. 11-6.) The voltage developed across the coil is amplified by a 6BA6 4.5-mc amplifier and then made available to the F-M detector.

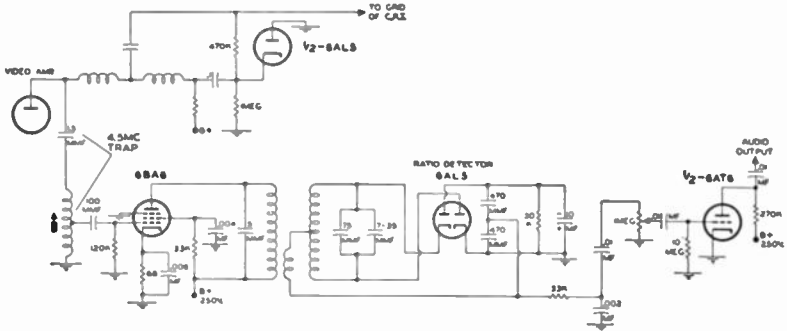


FIG. 11.6. The method of separating the sound signal (4.5 mc) from the video signal. Shown also is the sound system.

Through the use of a series resonant circuit, essentially all the 4.5-mc voltage is removed from the video path and thereby does not reach the cathode-ray tube. As a further precaution some manufacturers insert another 4.5-mc trap beyond the separation point. This was illustrated previously in Fig. 7-20, Chapter 7. The 6BA6 operates as a limiter and serves to present a fairly constant output when widely varying signals are applied to its input. This is desirable, especially if some form of tuning control is provided, otherwise the set user might attempt to tune for loudest sound, which would seriously distort the picture.

We are now in a position to appreciate fully the reasons for the reduction in sound carrier prior to the video second detector. First, if the video and audio carriers are comparable in strength when they reach the video second detector, the resultant 4.5-mc beat note will contain a considerable amount of amplitude modulation. Also, the amount of this amplitude modulation will vary as the modulation of the video carrier changes. Since no F-M detector now in practical use is able to remove completely or to ignore amplitude modulation in an F-M signal, the audio output will be distorted to an extent dependent upon the amount of amplitude modulation present in the signal. Even the use of a limiter prior to the detector does not altogether

alleviate this trouble. With the amplitude of the sound carrier reduced to about 5 per cent of the video carrier, the 4.5-mc beat-note amplitude remains substantially constant as the level of the video carrier varies between 15 and 100 per cent of its peak value.

Another consideration in conjunction with the use of comparable audio and video carriers is the manner in which the audio voltage can affect the picture. Since the sound carrier is frequency modulated, one way to convert an F-M signal into an audio voltage is by means of slope detection. Now, if the sound signal receives 26 db less amplification than the video signal in the I.F. system, it will be quite small and, even if 30 per cent of this signal is converted into audio voltage* at the video second detector, the interference from this source will be approximately 36 db below the video signal and generally unnoticeable. However, if the audio signal is large at the detector, then even 10 per cent slope detection will produce enough audio voltage to develop horizontal bars across the screen. It will probably be found that with a large audio signal, more than 10 per cent slope detection occurs, further aggravating the situation.

If the 4.5-mc signal itself reaches the control grid of the cathode-ray tube, a fine line pattern will appear on the screen. This will occur if the audio signal is strong at the video detector output and does not receive the proper amount of reduction. However, if the audio signal is reduced by the 26 db already specified, then its effect, after the shunting effect of the series resonant pick-off circuit, will again be so small as to be negligible.

Thus, we see from the foregoing why it is mandatory for the video carrier to be considerably stronger than the audio carrier, at the video detector, if the Intercarrier system is to function properly. It also explains why the presence of the sound signal, even as far as the last video-frequency amplifier, does not materially affect the image on the screen.

In receivers employing the conventional system, it has recently become common practice to insert a 4.5-mc trap in the grid or plate circuits of the first video-frequency amplifier following the video second detector. This is done to prevent any 4.5-mc voltage, which may be generated in the video detector, from reaching the cathode-ray tube. In the mass production of many television receivers, the sound carrier may not receive the full attenuation required to suppress it completely. Hence this additional precaution. In Intercarrier sets, the sound I.F. receives some amplification, as noted.

H.F. Oscillator Frequency Changeover. Since the sound and video signals both pass through the video I.F. system, special trap circuits are not

* F-M signals can be detected by detuning an A-M detector and using the slope or sides of the A-M response curve to obtain the audio modulation from the F-M signal. This is known as slope detection. It is not a very efficient method but it will work.

amplifiers is shown in Fig. 11-8. (This includes the mixer transformer.) The video carrier I.F. is 26.2 mc and the sound I.F. is at 21.7 mc.

In Fig. 11-9, we have the I.F. response curve for channels 7 to 13. The video and sound I.F. carriers receive the same amplification as in channels 2 to 6, but they are now located on different sides of the response curve. This can be explained if we examine the local oscillator frequency for each of the channels. (See Table 11-1.) For the lower channels, the oscillator

TABLE 11-1. A TABULATION OF THE VARIOUS CHANNELS, VIDEO CARRIER, SOUND CARRIER, AND LOCAL OSCILLATOR FREQUENCIES IN THE TELEVISION RECEIVER SHOWN IN FIG. 11-7

<i>Channel Number</i>	<i>Channel Frequency</i>	<i>Picture Carrier Frequency</i>	<i>Sound Carrier Frequency</i>	<i>Receiver R.F. Osc. Frequency</i>
2	54-60	55.25	59.75	81.45
3	60-66	61.25	65.75	87.45
4	66-72	67.25	71.75	93.45
5	76-82	77.25	81.75	103.45
6	82-88	83.25	87.75	109.45
7	174-180	175.25	179.75	152.35
8	180-186	181.25	185.75	158.35
9	186-192	187.25	191.75	164.35
10	192-198	193.25	197.75	170.35
11	198-204	199.25	203.75	176.35
12	204-210	205.25	209.75	182.35
13	210-216	211.25	215.75	188.35

frequency is above the carriers; for the upper channels, the oscillator frequency is below. Oscillator frequency switching is not possible in conventional television sets because of the sound take-off trap and other traps usually employed in the I.F. stages. It is also not possible in those Intercarrier sets which have traps in the video I.F. system. But where an Intercarrier set has no video I.F. traps, it can be done. All traps are fixed tuned, and their frequencies cannot be altered as they would have to be if the oscillator frequency position (with respect to the two carriers) were shifted. For the upper channels, 7-13, Fig. 11-9, the video carrier I.F. value is 22.9 mc, and the sound carrier I.F. is 27.4 mc. In each instance the sound carrier is 15.6 db below the level of the video carrier. Although this is not the 26 db suggested, it will still provide satisfactory operation.

Advantages of the Intercarrier System. Now that we have examined the underlying principles of the Intercarrier system and noted its application to commercial receivers, it might be well to evaluate its advantages and limitations. One big advantage offered through the use of Intercarrier system is its relative independence of many of the difficulties which beset the local oscillator and which affect the quality of the output in conventional television

receivers. Thus, consider what happens when the oscillator frequency drifts. In both systems, the resultant video and sound I.F. values will shift. Suppose that this shift is 200 kc. In a receiver employing the conventional television system, the video signal will be negligibly affected because a 200-kc shift is unimportant when compared to the total video spread of 4.0 mc. In the sound I.F. system, however, the discriminator linearity extends usually for ± 100 kc on either side of the I.F. carrier frequency, and a shift of 200 kc will remove the signal from the linear portion of the characteristic, distorting or causing the complete loss of audio output.

Compare this with the same effect in a receiver employing the Inter-carrier system. Since separation of the two signals does not occur in the I.F. system, the only effect this produces is to shift the position of the two signals along the I.F. response curve. This will alter somewhat the amplification accorded these signals, but not to an appreciable extent since 200 kc is negligible compared to the 4.5-mc bandpass. When the two signals reach the video second detector, the resultant 4.5-mc beat note will be obtained since the difference between the two carriers is fixed at the transmitter and nothing that occurs at the receiver can alter this relationship. The 4.5-mc beat note will then pass through the video-frequency amplifiers and into the sound system, unaltered by the local oscillator change in frequency.

Shifting of the local oscillator frequency due to hum or microphonics will likewise have negligible effect in the Intercarrier system because, no matter how the oscillator frequency changes, the difference between the two carrier frequencies remains 4.5 mc, and it is from this difference that the sound output is obtained.

Fine-tuning controls are employed in conventional television receiver sets to counteract any drift of the oscillator frequency. Since oscillator drift — within nominal limits — does not appreciably affect the operation of a set using the Intercarrier system, this control may be dispensed with. However, many of the Intercarrier receivers on the market still include a fine-tuning control. With the control available, less precautions need be observed in designing the front-end tuning system, with resultant economy.

Disadvantages of the Intercarrier System. The Intercarrier system is not without disadvantages, however. The appearance of the sound at the loudspeaker is dependent upon the mixing action that occurs in the video second detector. If something should happen to prevent the transmission of the video carrier — and this can and has happened — then no output at all is obtained at the receiver. Note that this is peculiar to Intercarrier systems and is not true in sets employing the conventional system. While it may be argued that the foregoing set of conditions seldom occurs, yet there is another manifestation of the same effect that appears more frequently. Thus, when a white line or series of white lines is being transmitted, the video signal level

drops down to a very low value. If care is not observed at the transmitter, the level may easily drop to zero (corresponding to 100 per cent modulation), resulting momentarily in no video signal. This causes the sound to disappear, also momentarily. This is usually repeated at line frequency (15,750 cycles) or field frequency (60 cycles) producing a 60- or 15,750-cycle buzz in the receiver. (The 60-cycle buzz is more evident because most people cannot detect a note as high as 15,750 cycles.) To prevent this, it is necessary to check transmitter modulation carefully. Current F.C.C. regulations merely specify that the amplitude of the video carrier, when transmitting maximum white, may not be more than 15 per cent of the maximum video carrier amplitude (which is attained when the sync pulses are sent). If this is amended to specify a definite minimum, say 10 per cent, then we would insure that the video carrier would never be lost.

Another factor that requires careful observance is the incidental phase or frequency modulation that sometimes develops in the video carrier when it is receiving the video intelligence. It has been found that under some circumstances, the video carrier will receive a small amount of phase or frequency modulation while it is being amplitude-modulated at the transmitter. In the subsequent mixing that occurs in the video second detector at the receiver, this phase or frequency modulation is passed on to the 4.5-mc sound signal and is not removed at the F-M detector. The result again is a 60-cycle buzz or a 15,750-cycle note.

Precautions are necessary in receivers, too. Here, the greatest single factor affecting proper operation is the relative magnitude of the video and sound carriers at the time they are mixed in the video detector. If the amplitude of the video carrier is very large compared to that of the sound carrier, then the amplitude of the beat note is very nearly independent of the amplitude of the video signal. This means that the F-M 4.5-mc beat note will not be affected by any modulation contained on the video carrier. Now, this stress on keeping the amplitude of the beat note as constant as possible arises chiefly because the F-M detector and the limiter which precedes it are not insensitive to amplitude modulation when such modulation appears in a sizable amount. Most servicemen believe that an F-M detector and limiter will not respond at all to amplitude modulation, which is true if the amplitude modulation is small. However, if it becomes large, then it will affect the output of the F-M detector and distort whatever audio signal is present.

The value of 26 db as the difference between the amplification accorded the video and sound I.F. carriers will insure that the 4.5-mc beat note receives little amplitude modulation.

Finally, there is the problem of attempting to maintain the sound limiter and detector tuned circuits at precisely 4.5 mc. Any deviation from this value, due to altered components, usage, humidity, etc., will cause the sound

signal to be displaced to a non-linear portion of the detector characteristic, thereby producing a distorted output. Nothing short of a complete realignment of these stages will correct this condition. In sets using the conventional system, this condition can be corrected by retuning the local oscillator, using the fine-tuning control.

QUESTIONS

1. Indicate, in block diagram form, the difference between conventional television receivers and those employing the Intercarrier system.
2. What has the phenomenon of signal mixing to do with the operation of Intercarrier television sets?
3. What are some of the disadvantages of the conventional system and how does the Intercarrier system avoid them?
4. Compare the I.F. response curves of conventional and Intercarrier sets and explain any differences.
5. In Intercarrier sets, what prevents the audio signal from reaching the cathode-ray tube?
6. Where does video and audio signal separation occur in an Intercarrier receiver? Draw a typical separation circuit.
7. Which would require the greatest number of additional parts: Transforming a conventional receiver to the Intercarrier system or converting an Intercarrier set to the conventional system? Explain your answer in detail.
8. Upon what major fact is the Intercarrier system based?
9. Detuning a receiver has the greatest effect on which of the two types of receivers? Why?
10. Under what conditions would a conventional receiver produce a sound output whereas an Intercarrier set would not? Explain.
11. What precautions must be observed at television broadcast stations in order that Intercarrier sets will operate properly?
12. List the advantages and disadvantages of Intercarrier sets.
13. Why can the designer of Intercarrier sets place the local oscillator frequency above the video carrier frequency on the low band and on the low side of the carrier on the high band and yet not do this in conventional receivers?
14. What happens in Intercarrier sets when the audio carrier becomes comparable in strength to the video carrier? What precautions are taken to avoid this condition?
15. List several ways in which the servicing procedure of an Intercarrier set might differ from that of a conventional set.
16. Why can the sync systems of both of these receivers be identical?
17. Describe the passage of a signal through an Intercarrier set, starting at the antenna and ending at the image tube and loudspeaker.
18. In what way does the mixing of a frequency-modulated signal with a single frequency voltage differ from the mixing of an amplitude-modulated signal and the same single frequency signal?
19. Why is the value of the audio I.F. frequency the same in all Intercarrier sets while this is not true in conventional television receivers?
20. What troubles might conceivably arise in an Intercarrier set that could not occur in a conventional receiver?

Chapter 12

SYNCHRONIZING CIRCUIT FUNDAMENTALS

Synchronizing Pulses. Up to this point we have studied the action of the various stages of the television receiver in amplifying and changing the form of the video signal so that it was finally suitable for application to the grid of the picture tube. Nothing, however, has been said so far about the method of supplying the proper voltages to the deflection plates (or coils) so that the image will be swept out properly on the cathode-ray screen. To accomplish this, we must obtain the synchronizing pulses from the video signal and apply them to other circuits that will eventually connect directly to the deflecting plates of the picture tube. Since each line has a separate synchronizing pulse, it becomes possible to lay them out on the screen in their proper position exactly as they were scanned on the camera tube mosaic. The synchronizing pulses that are responsible for the correct positioning of the various lines are referred to as the horizontal synchronizing pulses, or, perhaps more simply, as the horizontal pulses. These pulses are diverted to amplifiers that control the action of the horizontal deflecting plates and coils.

After the electron beam sweeps out the correct number of horizontal lines and arrives at the bottom of the picture, a vertical synchronizing pulse is applied to the vertical deflection plates, and the beam is rapidly brought back to the top of the screen again. This vertical pulse is transmitted along with the horizontal pulses in the video signal, separated by filters at the receiver, and applied to a set of amplifiers that end at the vertical deflection plates. The block diagram of Fig. 12-1 illustrates the general path of all the synchronizing pulses within a television receiver.

Pulse Separation from the Rest of the Signal. To use the pulses of a video wave, they must first be separated from the other portions of the signal. The separation may occur anywhere, from the video detector to the last video stage before the cathode-ray tube. In practice, commercial set designers have generally chosen to obtain the input for the synchronizing

stages from some point beyond the video detector, usually at the output of one of the video amplifiers. At these points, the signal has sufficient amplitude and is in proper form so that it can be made to control the horizontal and vertical deflecting oscillators with a minimum of additional stages. For example, set designers often do not apply the video signal to the separating tube until it has passed through the first video amplifier. In this way an extra pulse amplifier is eliminated.

Since it is necessary to obtain the synchronizing pulses from the incoming wave, it is first imperative that the signal be in its d-c form. This should be evident by reference to the figures of Chapter 8 where the a-c

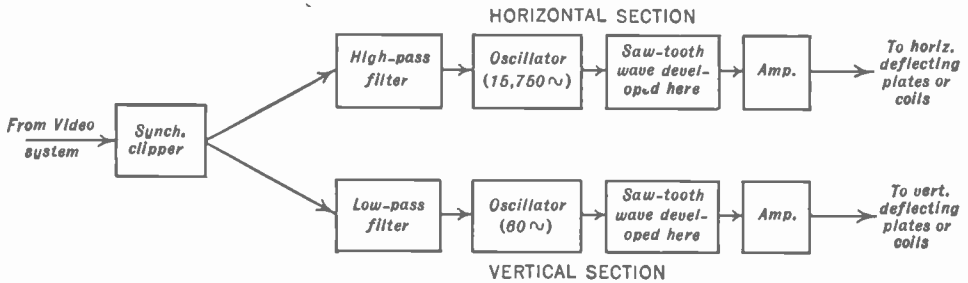


FIG. 12.1. A basic diagram of the synchronizing section of a television receiver.

and d-c forms of a video signal are illustrated. While the signal is always in its d-c form at the output of the detector, it may not be so if obtained from the plate circuit of some following amplifier. In this case, d-c restoration is necessary, and the method of achieving it is shown in the circuits that follow.

The tube that separates the sync pulses from the rest of the video signal is called the clipper or sync separator. Both horizontal and vertical synchronizing pulses are clipped by this tube, the further separation of these two pulses then occurring at another point beyond this stage. The type of tube that may be utilized for the synchronizing separation is not restricted. Practically every type is suitable since the action consists merely in biasing the tube so that only the top portions of the video wave (where the pulses are found) affect the tube and cause current to flow.

Diode Clippers. A possible diode clipper circuit is shown in Fig. 12-2. The video signal is applied between plate and ground while the output voltage is developed across the diode load resistor R_L . The small battery is inserted with its negative end toward the plate. This prevents current from flowing until the video signal acting on the tube becomes sufficiently positive to counteract the negative biasing voltage. Current then flows. With the circuit constants properly chosen, current should flow only at the

synchronizing pulses which are the most positive for a signal having negative phase, and the output will consist only of these short pulses of current. The picture phase at the input of this diode must be negative, as in Fig. 12-2.

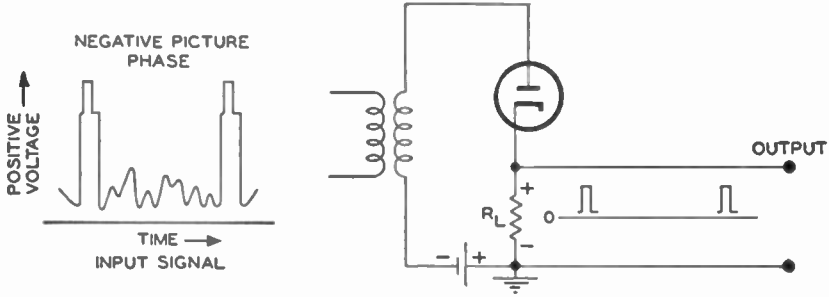


FIG. 12.2. A diode clipper. This operates with input signals having negative phases.

By inverting the diode, as in Fig. 12-3, it becomes possible to apply a positive picture phase to the tube and again obtain only the pulse tips across R_L . The d-c biasing voltages necessary for these diodes may be taken from the low-voltage power supply.

It is not very practical to use a biasing battery or power-supply d-c voltage for the diode clipper tube. We require some arrangement that is completely automatic in its operation, altering its operating point as the

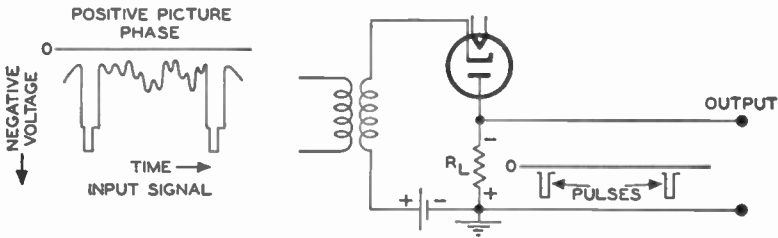


FIG. 12.3. An inverted diode clipper, suitable for input signals having a positive picture phase.

amplitude of the received carrier varies. A simple, yet effective, circuit is shown in Fig. 12-4A. The diode clipper uses the time constant of R and C to bias the tube so that all but the synchronizing pulses are eliminated. Capacitor C and resistor R form a low-pass filter with a comparatively long time constant, equal to approximately 10 horizontal lines. Therefore, the voltage developed across R (and C) will be determined by the highest voltage applied across the input terminals. This, of course, means the

synchronizing pulses. Throughout the remainder of the line, while the video voltage is active, the plate is never driven sufficiently positive to overcome the positive cathode bias.

A commercial application is shown in Fig. 12-4B. One half of a 6AL5 is used for picture signal detection (not shown) while the remaining half is devoted entirely to pulse rectification and clipping. R_1 and R_2 are the

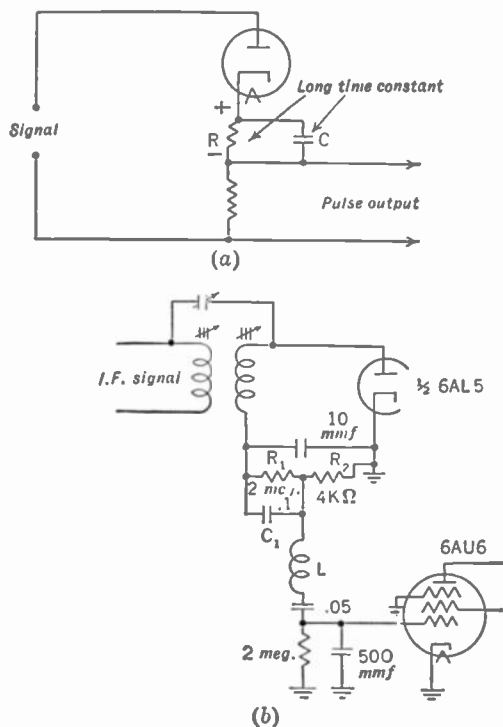


FIG. 12.4. (A) Fundamental diode clipper circuit. (B) The commercial application.

pulse detector load and here the rectified signal is developed. The time constant of the load is set by R_1 and C_1 . At the arrival of each pulse to the tube, a short flow of current occurs, recharging C_1 , and, at the same time, producing a pulse across R_2 . This voltage is passed on to a 6AU6 synchronizing pulse amplifier. The series inductance L maintains a good response in the connecting network to the higher frequency components of the square-shaped pulses. Any decrease in high-frequency response here would have the effect of rounding out the steep sides of the synchronizing pulses, thereby destroying the effectiveness of their triggering action. It can readily happen that the synchronizing oscillators will trigger at slightly differing intervals, causing sections of the image to "tear out." The appear-

ance of such an image is identical to that obtained when interference is active in the circuit.

A more frequent application of the diode sync separator uses the circuit shown in Fig. 12-5. The composite video signal is applied across terminals A-B. As a start, we will assume that C_1 is uncharged. Upon the application of a signal, electrons flow from cathode to the plate of the tube and thence to the nearest plate of C_1 . From the other plate of C_1 , an equal

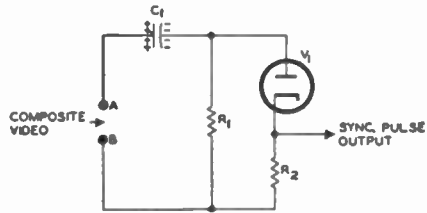


FIG. 12.5. A second and more frequent application of a diode as a sync separator.

number of electrons will flow to terminal A, through whatever circuit is connected between this point and terminal B, up through R_2 , and back to the tube again. C_1 will become charged, its potential equal to the peak potential of the applied video signal. During the second and all subsequent lines, V_1 will not conduct until the applied signal is able to overcome the negative charge existing across C_1 , and this will be true only when the sync pulses are active. At these moments, current flows through the circuit in the manner described above and a pulse of voltage appears across R_2 . At all other times V_1 will be kept at cut-off by the negative charge across C_1 . When V_1 is not conducting, C_1 discharges through R_1 , and it is actually this discharge through R_1 that makes the plate of the diode negative with respect to its cathode.

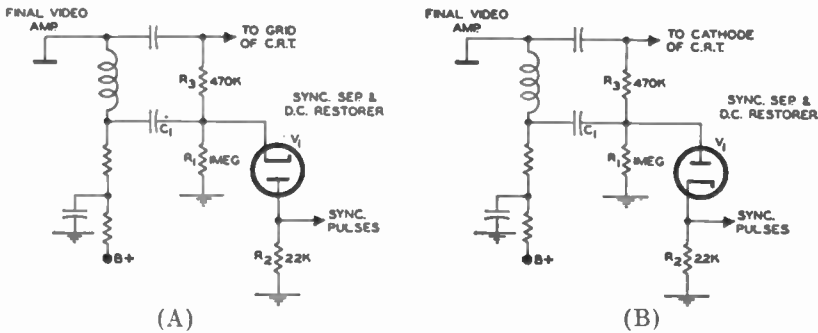


FIG. 12.6. Combining sync separation with d-c restoration.

As before, the time constant of C_1 and R_1 is made sufficiently long so that only the sync pulses cause current flow through the circuit.

If the reader will now refer to Chapter 8, he will see that the diode d-c restorer circuits which also functioned as sync clippers utilized a similar circuit. The d-c restorer action was obtained from C_1 , R_1 , and V_1 (Fig. 12-6). The negative voltage which developed across R_1 was fed through

a resistor, R_3 , to the cathode-ray tube, where it combined with the video signal, bringing all sync pulses to a common level. At the same time, R_2 , in series with V_1 , received a pulse of current each time the tube conducted. Since this occurred only when the sync pulses were active, only the sync pulses appeared across R_2 .

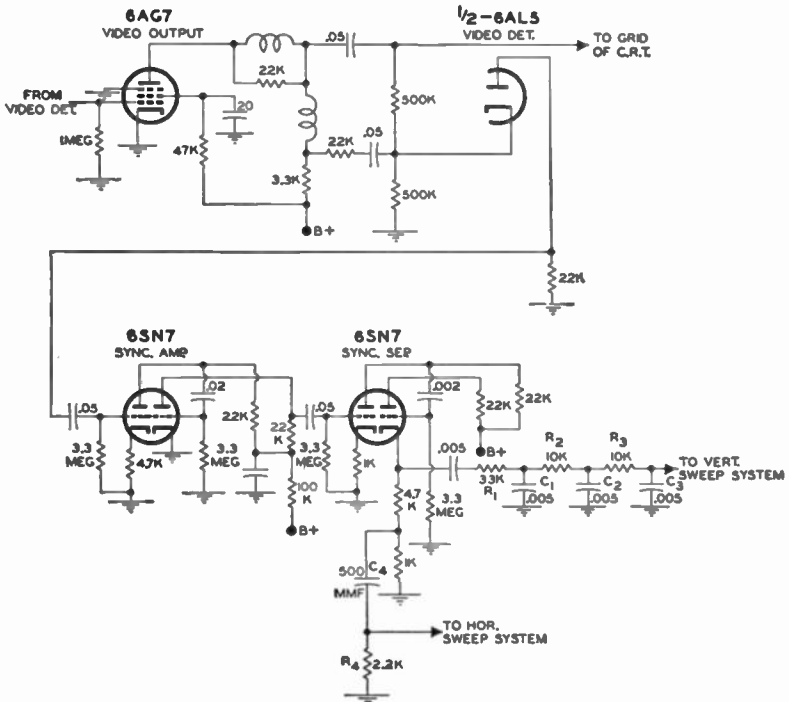


FIG. 12.7. A complete sync separation and amplifying network using the diode sync separation of Fig. 12.6(A).

When the video signal at the plate of the last video-frequency amplifier is fed to the grid of the image tube, signal phase is positive, which means that the sync pulses possess the most negative potential of the signal. Therefore the diode V_1 must be inverted. (See Fig. 12-6A.) On the other hand, when the video signal is fed to the cathode of the image tube, the sync pulses have the most positive potential of the video signal, and the diode is placed as shown in Fig. 12-6B. Circuit operation, in either instance, is the same.

In some television receivers, the sync separator shown in Fig. 12-6A is combined with two 6SN7 tubes, for the purpose of clipping and squaring off the sync pulses, before these are applied to their respective sweep oscillators. (See Fig. 12-7.) In spite of the different names applied to these

duo-triodes on the manufacturer's diagram (i.e., sync amplifier and sync separator), all perform essentially the same jobs of amplifying and clipping the sync pulses. The final tube is called a sync separator possibly because the pulses appearing in its cathode circuit are applied to the vertical and horizontal sweep oscillators. Note, however, that the separation of the vertical and horizontal sync pulses from each other is not performed by the tubes but by the high- and low-pass filters leading to each oscillator. All the sync pulses — and these include both vertical and horizontal — are applied to the low-pass filter composed of R_1 , R_2 , and R_3 plus C_1 , C_2 , and C_3 . The low-frequency (60-cycle) vertical pulses are permitted to pass while the effect of the 15,750-cycle horizontal pulses is made negligibly small.

By the same token, the pulses reaching the horizontal blocking oscillator must pass through the high-pass filter consisting of C_4 (500 $\mu\mu\text{f}$) and R_4 (2,200 ohms). This circuit is known as a differentiating network and produces a sharp pip at the leading and lagging edges of each rectangular pulse. The pip due to the leading edge is then used to sync the horizontal oscillators.

Maintenance of synchronization is an important consideration in the design of even the cheapest television receiver, and one way to achieve sync stability is by successfully amplifying and shaping the sync pulses obtained from the video signal. A point to remember is that *complete* separation of the sync pulses from the image detail voltages is seldom achievable using a single diode or triode. Hence, whenever these tubes are employed as sync separators, they are usually followed by one or more amplifiers and shaping circuits. A pentode with a sharp cut-off characteristic may be used by itself if economy is important.

Triode Sync Separators. Since a triode (or a pentode) can do anything a diode can, and provide amplification as well, it is natural to expect to find these tubes used as sync separators. An application involving a triode sync clipper is shown in Fig. 12-8. The triode is biased by grid-leak bias developed across C_1 and R_1 . The pulses in the video signal fed to the sync clipper triode V_1 possess the most positive polarity of the signal. Electrons will then flow in the grid circuit, charging the coupling capacitor. Because of the high value of R_1 , the charge on C_1 will leak off slowly, causing a fairly steady bias voltage to develop across the grid resistor. This biasing voltage prevents plate current from flowing except for the most positive values of the incoming signal, which are the synchronizing pulses. A low plate voltage (14 volts) causes the tube to saturate readily, thereby tending to square off the sync pulses and to limit any noise pulses that may appear in the signal.

The signal from V_1 is fed to two circuits: One portion is fed to the grid of V_3 through an integrating network (R_2 and C_2) which is a low-pass filter. The wide vertical pulses pass through the filter, are amplified by

V_3 , and then cathode-coupled to the following vertical multivibrator sweep oscillator. The horizontal sync pulses, on the other hand, do not develop sufficient voltage across C_2 appreciably to influence V_3 . Further integration of the sync signal is provided in the cathode circuit of V_3 .

The second circuit to which the output of V_1 is fed is the horizontal sync amplifier V_2 . This tube operates into an automatic frequency control

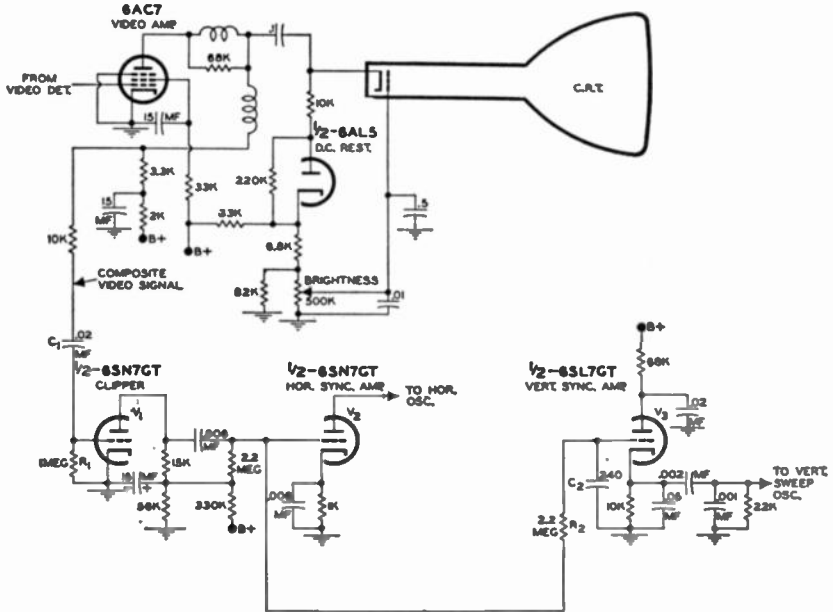


FIG. 12.8. A triode (V_1) sync clipper.

(A.F.C.) network which automatically regulates the frequency of the horizontal sweep oscillator. Not shown in Fig. 12-8, but indicated in Fig. 13-16 is coupling transformer T_1 which acts as a high-pass filter or differentiator by virtue of its low inductance; that is, in the secondary, the original sync pulses are transformed into positive and negative pips. Only the pip which is produced by the leading edge of each pulse is used.

The combination of a triode d-c restorer and sync separator is a common one. (See Fig. 12-9.) The grid and cathode of the triode V_1 function as a diode d-c restorer, providing a positive d-c voltage across R_1 which is added to the video signal applied to the control grid of the image tube through a 1-megohm resistor. Since the applied video signal has a positive picture phase (the sync pulses are most negative), the grid of the tube is grounded and the signal is received by the cathode. Current flows through the triode only when the negative applied voltage is sufficiently large to

overcome the positive potential developed across the cathode resistor R_1 . This occurs only during the sync pulse interval, producing amplified sync pulses across the plate load resistor. The pulses receive further amplification by triode V_2 and then are transferred to separate vertical and horizontal sync amplifiers. The vertical system contains two additional amplifiers while the horizontal system contains what is essentially a cathode-follower and phase inverter feeding into an A.F.C. circuit.

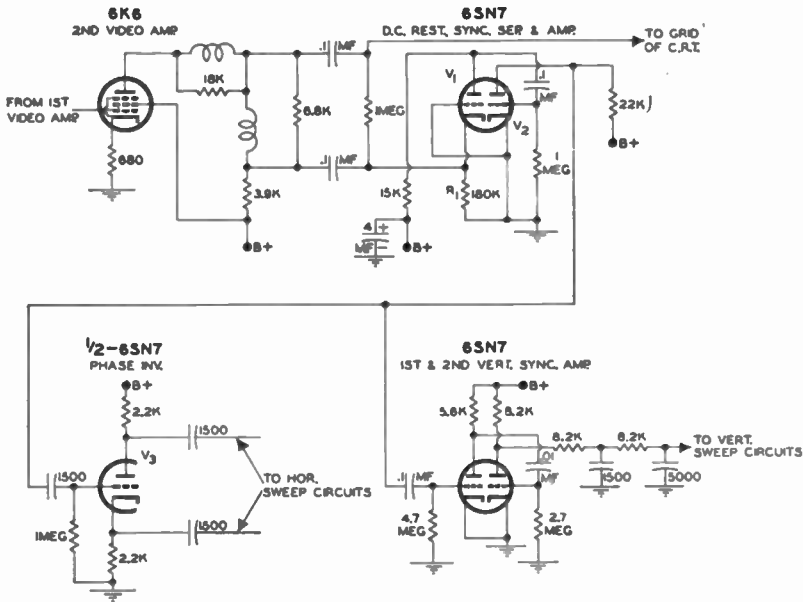


Fig. 12.9. A triode (V_1) employed as a sync separator and d-c restorer.

Another triode sync separation circuit is shown in Fig. 12-10. The first triode V_1 is a normal voltage amplifier. Its output is the complete video signal and is applied to the grid of V_3 with the sync peaks in the positive direction. In the input circuit of V_3 , a diode, V_2 , is connected from grid to ground. This diode has -14 volts applied to its plate terminal and will not conduct until the 14 volts are neutralized by a signal which is at least 14 volts above zero reference level (ground). When this occurs, the diode conducts, effectively preventing the signal at the grid from rising any further. The grid of V_3 is biased by the same -14 volts, preventing the tube from conducting until part of this voltage has been neutralized. The video detail voltage variations of the incoming signal are unable to overcome this negative bias and thus the video detail voltage up to approximately the blanking level is removed. Only sync pulses, which extend in

the positive direction, are able to force V_3 to conduct, producing positive pulses in the cathode circuit which are then fed through appropriate filters to the vertical and horizontal blocking oscillators.

V_2 serves as a sync leveler, tending to provide a constant pulse output. Because of the amplification accorded by V_1 , the sync pulses of all normal signals reaching V_3 extend more than 14 volts above the zero reference axis. Hence, all sync pulses drive the diode into conduction, preventing further

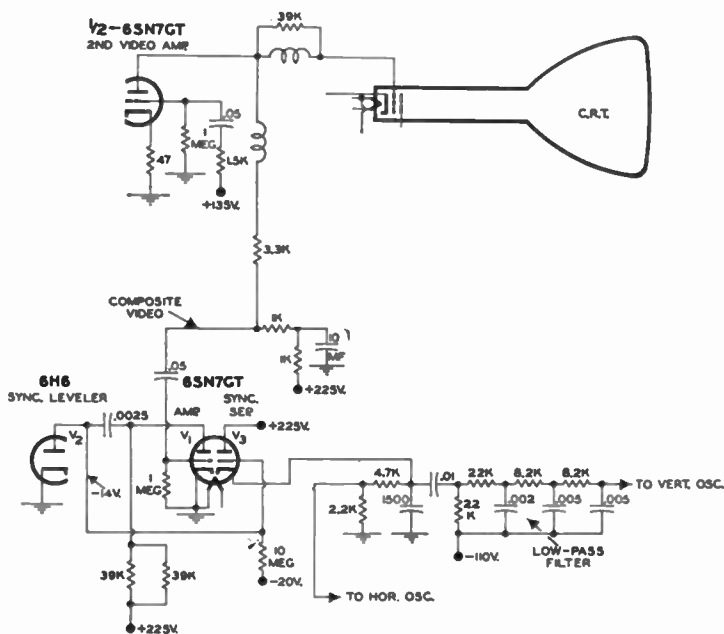


FIG. 12.10. Another triode sync separator circuit.

rise of the pulse beyond 14 volts and leveling off the tips of all sync pulses. For signals too weak to develop sufficient voltage at the output of V_1 to drive V_2 into conduction, leveling does not occur and the sync amplitude variations reach the sweep oscillators, impairing sync control.

Pentode Sync Separators. To obtain clean-cut sync separation, sharp cut-off pentodes are usually more satisfactory than either triodes or diodes. The circuit of Fig. 12-11 was shown, in part, in Chapter 8. It was noted there that the diode d-c restorer circuit (V_1) delivered 14 times more sync voltage to V_2 than video detail signal. However, an appreciable amount of video signal still remains, and consequently further clipping is necessary. Complete pulse separation is achieved by the following three-stage system containing a sync amplifier, a sync separator, and a second sync amplifier or limiter. Let us analyze each in turn.

The first sync amplifier is a 6SK7 pentode which has a remote cut-off characteristic. The signal received from the d-c restorer network has a polarity such that the sync pulses are in the negative direction, which tend

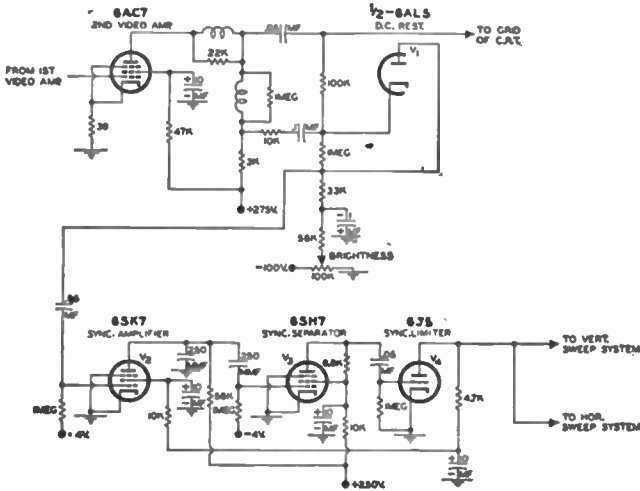


FIG. 12.11. A pentode sync separator circuit.

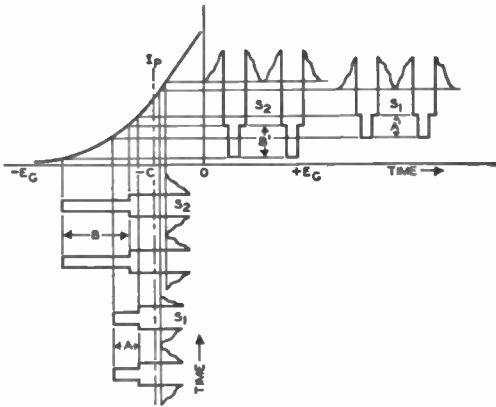


FIG. 12.12. The action of V_2 in Fig. 12.11 in stabilizing the sync pulse output of V_1 .

to drive the tube toward plate cut-off. With a weak signal, such as S_1 of Fig. 12-12, the sync pulse operates over the relatively high-gain portion of the curve. With large signals, such as S_2 , the sync pulse extends into the curved, or low-gain, portion of the tube characteristic. The result is a more constant sync pulse output. Noise pulses are also suppressed by this action, improving the sync-to-noise ratio.

The sync separator stage V_3 clips the signal just above the blanking level and removes all the video portion of the signal. This is accomplished by using the sync pulses of the signal to drive the grid of V_3 positive and by establishing a negative grid-leak bias. Although some d-c bias is applied to this stage from the power supply, it is considerably smaller than the grid-leak bias and serves mainly to prevent noise peaks from operating the sweep oscillators when no sync signal is reaching the sync separator. The 6SH7 tube in this stage has a sharp cut-off characteristic.

The sync pulses reaching the grid of the third sync amplifier are negative in polarity and must be inverted before they can be applied to the sweep oscillators. These pulses are sufficiently strong to drive V_4 into cut-off, clipping the signal. The result is a square-topped wave. Through the use of this fairly elaborate network, a constant sync pulse output is obtained with peak-to-peak video signal variations of from 6 to 60 volts on the grid of the picture tube. Systems which utilize fewer stages of sync separation are more sensitive to signal strength variation, and the stability of their synchronizing circuits diminishes rapidly with lowering of the signal strength.

Fig. 12-13A is a final illustration of a pentode sync separator. The tube, a sharp cut-off 6AU6 pentode, clips the sync from the composite signal and produces a d-c restoration voltage across its cathode bias resistor. The tube contains a large load resistor in its plate circuit and a relatively large resistor in its cathode leg. The result of this combination, particularly the large plate resistor, is to produce a characteristic such as shown in Fig. 12-13B. The cathode bias produced varies with the average amplitude of the applied signal and will be great enough to place all video detail information beyond cut-off, permitting only the sync pulses to appear in the plate circuit. (See Fig. 12-13C.) The pulses are sufficiently strong, with normal signals, to extend into the flat region of the curve. Thus, sync clipping at either end is obtained. Because of the rather large amplitude of the output pulses, a voltage divider network is connected to the plate load resistor, and the sync pulses are tapped off at appropriate points and then fed to the horizontal and vertical oscillators through filter networks.

Special Single-stage Sync Separators. In recent circuit designs several new methods of achieving sync separation have been employed. In one receiver, a 6BE6 pentagrid tube is the sole sync separator stage employed. (See Fig. 12-14.) Grid No. 1 receives the output of the video detector. The signal here, with the sync pulses negative, is on the order of 2 volts peak-to-peak. A potentiometer, labeled the fringe lock control, applies a small positive voltage to grid No. 1. The control is adjusted so that the bias applied to grid No. 1 is close to zero, thereby preventing the negative video signal from cutting the plate current off in the tube.

In the same tube, grid No. 3 receives the video signal after it has been

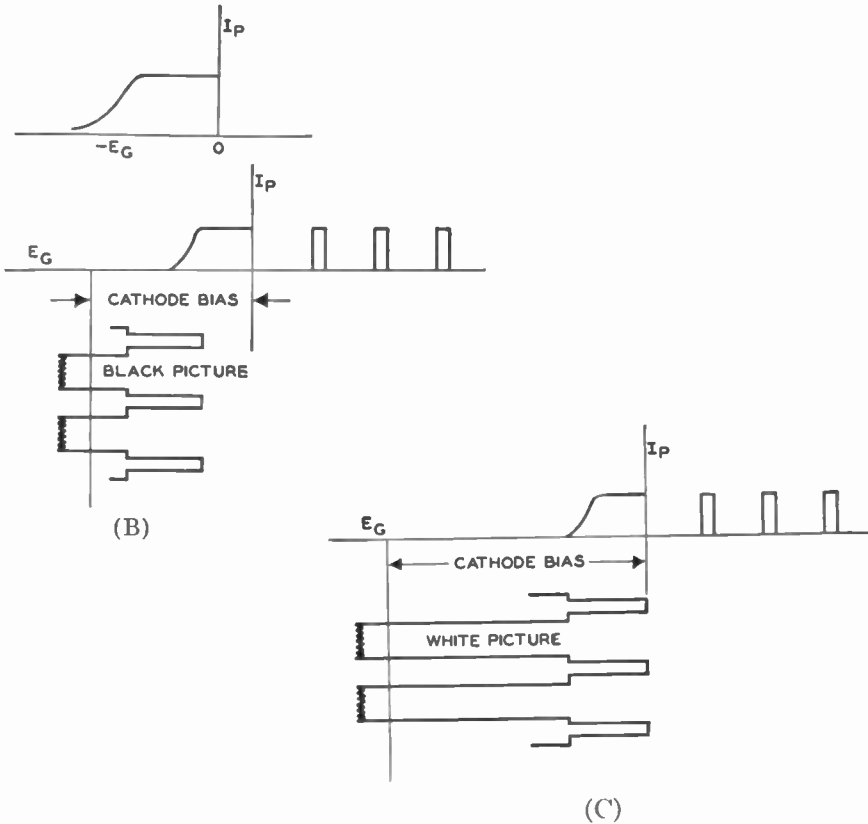
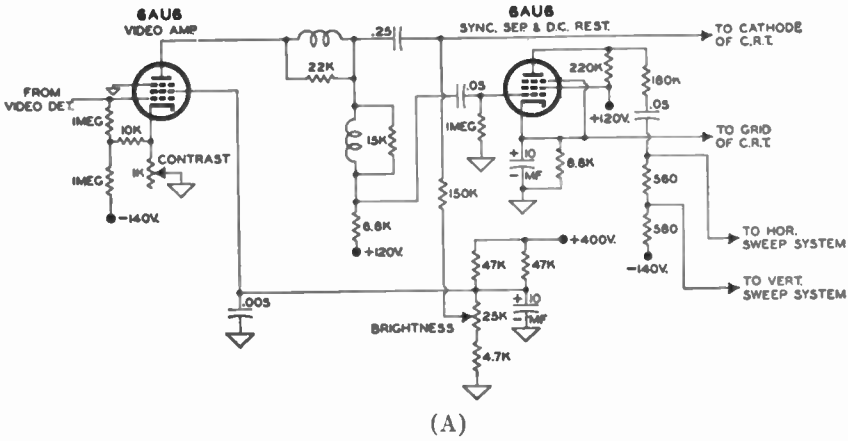
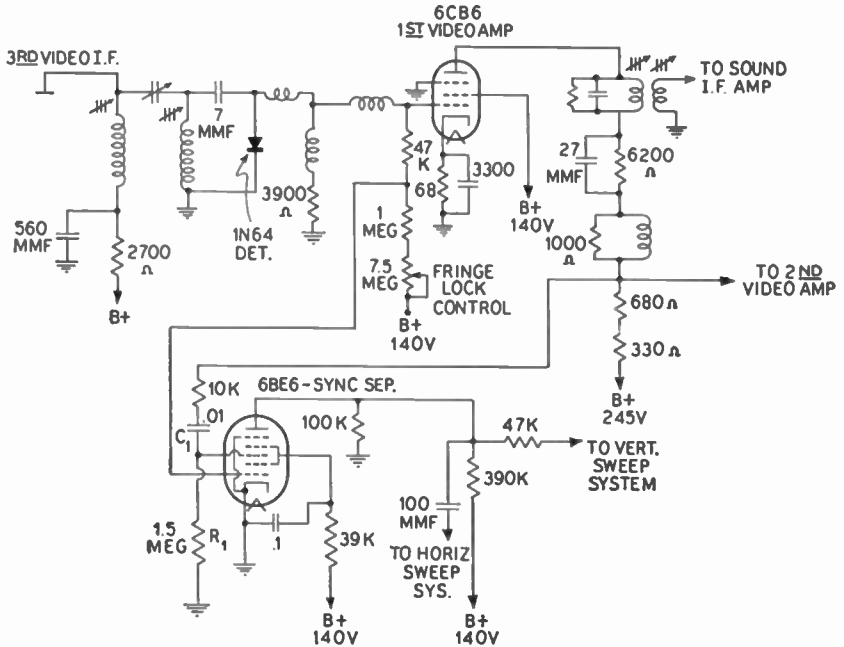


FIG. 12.13. (A) A pentode sync separation circuit using a 6AU6. (B) Characteristic of 6AU6 with the load resistance and voltages specified. (C) Operation of 6AU6 with black and white pictures.

amplified and inverted by the first video amplifier. The signal now has an amplitude of 40 volts peak-to-peak and the sync pulses are positive. Initially, grid No. 3 has zero bias. When the 40-volt peak-to-peak video signal is applied to it, grid leak bias is developed by C_1 and R_1 to an average value of -10 volts. This bias is maintained by the positive sync pulses, and only when the sync pulses are active can electrons flow past grid No. 3.



Courtesy Zenith Radio Corp.

FIG. 12.14. A 6BE6 pentagrid employed as a sync separator.

For the more negative portions of the signal containing the video information, the grid voltage is too negative to permit electrons to pass. Under these conditions, no current reaches the plate of the tube.

At grid No. 1, electrons are normally able to pass because the fringe lock control is adjusted so that electrons pass grid No. 1 even at the sync pulse tips which are most negative. However, grid No. 3 permits electrons to pass it only when the sync pulses are active. Effectively, therefore, it serves to clip the sync pulses from the rest of the video signal. Both grid No. 1 and No. 3 must be in condition to allow electrons to pass, otherwise none will reach the plate.

If a noise pulse drives grid No. 1 beyond the 2-volt level, plate current cut-off occurs, and the noise pulse cannot get through to trigger the sweep oscillators falsely. On rare occasions, a strong noise pulse may occur at

the time of the sync pulse and the tube likewise will cut off. However, the flywheel action of the sweep oscillators will maintain sync during this brief period. The entire fringe lock system is based on the fact that the loss of an occasional sync pulse is to be preferred to having a noise pulse get through to trigger the sweep oscillators falsely.

Another single-stage sync clipper uses the special 6BN6 tube. This tube possesses a characteristic such that, when the grid voltage changes from negative to positive values, the plate current rises rapidly from zero to a sharply defined maximum level. This same maximum value of plate current remains, no matter how positive the grid voltage becomes. Current cut-off is achieved when the grid voltage goes about 1.5 to 2 volts negative.

The reason for this particular behavior of the tube stems from its construction. (See Fig. 12-15A.) The focus electrode, together with the first

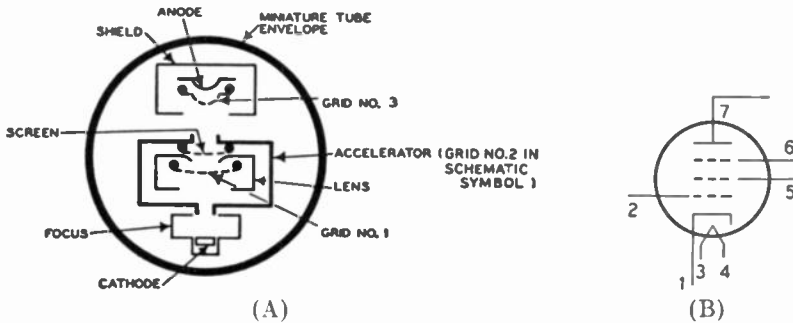


FIG. 12.15. The beam-gated tube. (A) The internal construction of the beam-gated tube. (B) Schematic symbol of beam-gated tube.

- 1 — Cathode, focus electrode, and internal shields
- 2 — Grid No. 1 (Signal or limiter grid)
- 3 } Heaters
- 4 }
- 5 — Grid No. 2 (Accelerator grid)
- 6 — Grid No. 3 (Quadrature grid)
- 7 — Plate or Anode

accelerator slot, forms an electron gun which projects a thin-sheet electron stream upon grid No. 1. The curved screen grid, together with the grounded lens slot and aided by the slight curvature of grid No. 1, refocuses the beam and projects it through the second accelerator slot upon the second control grid. This grid and the anode which follows are enclosed in a shield box. Internally, the focus, lens, and shield electrodes are connected to the cathode. The accelerator and the screen grid receive the same positive voltage because both are connected internally.

The foregoing design is such that the electrons approaching the first grid

do so head-on. Hence, when grid No. 1 is at zero potential or slightly positive, all approaching electrons pass through the grid. Making the grid more positive, therefore, cannot increase the plate current further. When, however, grid No. 1 is made negative, those electrons that are stopped and repelled back toward the cathode do so along the same path followed in their approach to the grid. Because of the narrowness of the electron beam and its path of travel, electrons repelled by the grid form a large space charge directly in the path of other approaching electrons, thus causing an immediate cessation of current flow through the tube.

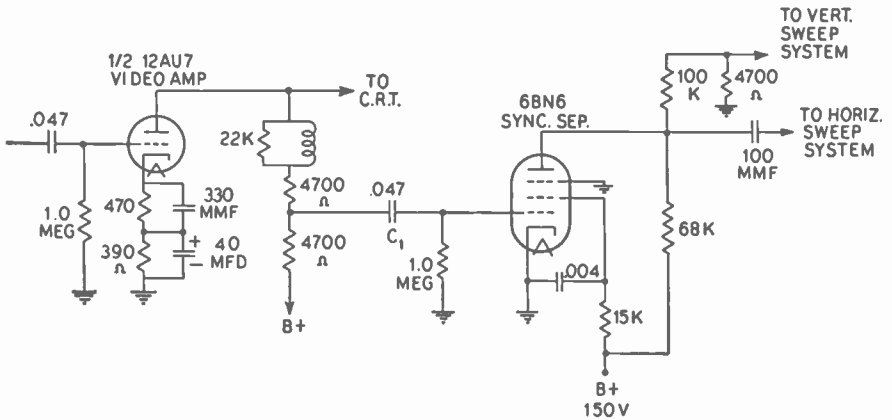


FIG. 12.16. A sync separator circuit using the 6BN6 tube.

In the sync separator application of this tube, the full video signal with the sync pulses positive is fed to the input of the tube. (See Fig. 12-16.) The positive sync pulses draw grid current, charging up C_1 and establishing grid-leak bias for the tube. Only the sync pulses are able to overcome this bias sufficiently to permit electrons to flow through the tube. The rest of the video signal, extending in the negative direction, is without effect. Sync clipping is particularly effective since plate current ceases when the grid is only 1.5 to 2 volts negative. Noise is also clipped by this tube because the plate current reaches saturation readily, and no positive voltage beyond this can increase the amount of plate current.

The portion of the sync pulse which does get through the tube appears in the plate circuit amplified. From here it is applied directly to the sweep oscillators or their control circuits.

Pulses and Their Form. So far, only general terms have been used when discussing the synchronizing pulses of video waves. Their purpose has been stated time and again, but nothing definite has been given as to the actual means of accomplishing their objective. There is nothing in

sound receivers that even closely resembles this action, and a detailed examination becomes necessary. The pulses, separated from the rest of the wave as outlined above, will be held in abeyance while we develop in greater detail the form and functions of the horizontal and vertical pulses.

It is already known that, as each horizontal line signal arrives at the grid of the picture tube, the electron beam should be in correct position, ready to sweep out the information contained in the signal. The position of the electron beam is controlled by saw-tooth oscillators. In order that the oscillator shall have the beam in the correct position, horizontal synchronizing pulses are inserted into the video signal. They could have been sent separately, but the present method is cheaper and simpler in operation. It is to be noted and continually kept in mind that the function of the horizontal synchronizing pulses is to trigger an oscillator in order to bring the electron beam from the right-hand side of the screen to the left-hand side. Once the beam is at the left-hand side, the oscillator is no longer directly under the control of a pulse and goes about its normal function of sweeping the beam across the screen. Thus each horizontal pulse that precedes the line detail sets up the beam in readiness for the scanning out of this information. The next pulse arrives when the beam is at the far right-hand side of the screen, at the end of a line.

There are 525 lines sent out every $\frac{1}{30}$ of a second. In one second, then, we have 525 times 30, or 15,750 lines. This means that the frequency of the horizontal pulses is 15,750 per second, or one arrives every $1/15,750$ sec. The time interval is quite small, being only 0.000064 sec.

In similar manner, the vertical pulses serve the purpose of bringing the electron beam back to the top of the screen for the beginning of each field. With interlaced scanning (described in Chapter 1), every other line is scanned, with each field ($\frac{1}{2}$ frame) taking $\frac{1}{60}$ of a second. The beam next sweeps out the lines that were missed, this also in $\frac{1}{60}$ of a second. The total frame, with all lines, is accomplished in $\frac{1}{60}$ plus $\frac{1}{60}$ of a second, or $\frac{1}{30}$ of a second. Thus we see that the vertical pulses must occur once every $\frac{1}{60}$ of a second, or 60 times in one second. This frequency is considerably less than that of the horizontal pulses and it is because of this fact that they can be separated with comparative ease.

With the preceding ideas in mind, let us closely examine the construction of the video signal with its synchronizing pulses. In Fig. 12-17 several lines of an image are shown, complete with the detail information, blanking voltages, and horizontal synchronizing pulses. The blanking and synchronizing voltages occupy approximately 20 to 25 per cent of the total signal amplitude. Notice that the blanking voltage retains its control over the cathode-ray tube grid for some time before and after each synchronizing pulse. This is done to make certain that no beam retrace is visible at all

on the screen. As soon as the blanking voltage relinquishes control of the grid, the line detail becomes active once again. All the lines of one field follow this form, the only difference occurring in the camera detail of the various sections of the image.

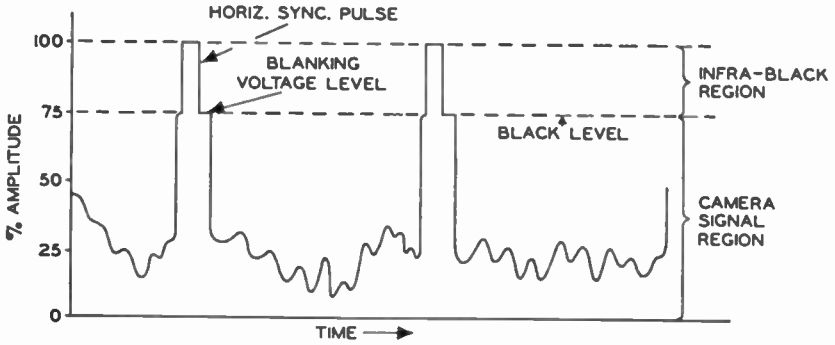


Fig. 12.17. A video signal with its synchronizing (horizontal) pulses.

At the bottom horizontal line, it is necessary to insert a vertical impulse that will bring the beam back to the top of the screen again. During the period that the vertical pulse is active, it is imperative that the horizontal oscillator should not be neglected. For, if this did occur, the horizontal generator probably would slip out of synchronization. To avoid this the vertical pulses are arranged in serrated form and accomplish vertical and horizontal synchronization simultaneously.

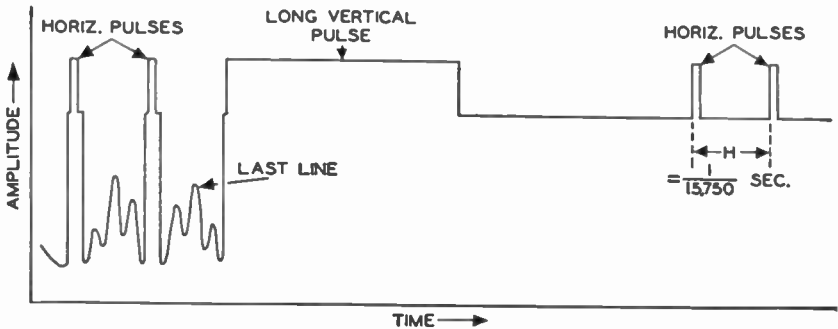


Fig. 12.18. The basic form of the vertical synchronizing pulse.

Serrated Vertical Pulses. To understand the form of the vertical pulse that has finally been evolved, start with the voltages shown in Fig. 12-18. At the bottom of the image, a long vertical pulse is inserted into the signal. This controls the vertical synchronizing oscillator and forces the beam to be brought back to the top of the screen. No provision is made in the

signal, in this preliminary form, to provide horizontal oscillator control while the vertical pulse is acting. As stated above, such a condition is undesirable as it permits the horizontal oscillator to slip out of control. To prevent this, the vertical pulse is broken up into smaller intervals and now both actions can occur simultaneously. The vertical synchronizing pulse, in the modified form, is shown in Fig. 12-19 and is known as a serrated vertical pulse.

While the vertical pulse is broken up to permit the horizontal synchronizing voltages to continue without interruption, the effect on the verti-

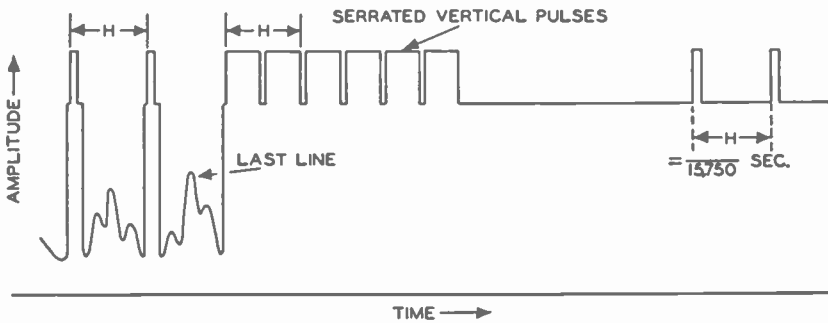


FIG. 12.19. The serrated vertical synchronizing pulse.

cal pulse is substantially unchanged. It still remains above the blanking voltage level practically all of the time it is acting. The interval is much longer than the previous horizontal pulse frequency. The two pulses are still capable of separation because their wave forms are different, as can be seen in Fig. 12-19.

Due to the fact that an odd number of lines is used for scanning, the form of the signal just prior to the application of the serrated vertical pulse must be still further modified. With an odd number of lines, 525, each field contains $262\frac{1}{2}$ lines from the beginning of its field to the start of the next. This is important and has not been overly stressed before. In Fig. 12-20, reprinted from Chapter 1, the notation is made that the end of the visible portion of each field occurs at the bottom of the image. However, the actual end of that field is not reached until the beam has been brought back to the top of the screen again. At the end of the visible portion of the first field, the beam must be interrupted at point D and the vertical synchronizing pulse inserted. Point D, we can see, occurs during the middle of a horizontal line. From D, the beam is brought up to point E, and the second field is begun. The visible portion of the latter field is completed at point F, the end of a complete horizontal line, and is returned to point A to repeat the entire sequence. These events are men-

tioned here for review. The reasons for employing this particular method of scanning were explained in Chapter 1.

When the beam is blanked out at the bottom of an image and returned to the top, it does not move straight up, but instead it moves from side to side during its upward swing. The reason is due to the rapidity with which a horizontal line is traced out as compared to the vertical retrace period. In fact, there are approximately 20 horizontal lines traced out while the vertical synchronizing pulses are bringing the beam back to the top of the picture.

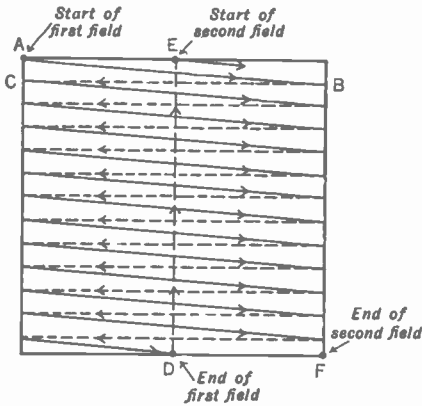


FIG. 12.20. The motion of the electron beam in interlaced scanning. For simplicity, the retrace from point D to point E has been shown as a straight line.

Thus, in each field, 20 horizontal lines are lost in the blanking interval between fields. Of the 525 lines which are specified, only $525 - 2(20)$ or 485 lines are actually effective in forming the visible image.

The method for arriving at the figure of 20 horizontal lines is quite simple. The electron beam is blanked out for approximately 1,250 microseconds between fields while the beam is being shifted from bottom to top of the image. During this interval, the horizontal sweep oscillator is also active. Thus, the beam, while it moves up under the influence of the vertical deflection voltage, is also moving back and forth because of the horizontal deflection oscillator. One horizontal line requires $1/15,750$ secs or 64 microseconds. Dividing this 64 into 1,250, we find that approximately 20 horizontal lines are traced out. In a frame, which contains two fields, 40 lines are thus lost. To see these retrace lines, turn up the brightness control on a television receiver when no station is being received and only the scanning raster is visible.

The fact, brought out above, that the vertical pulse is once inserted into the video signal when a horizontal line is half completed and once at the end of a complete line, necessitates a further modification of the video signal just prior to the arrival of the vertical pulse. A serrated video signal for each case is illustrated in Fig. 12-21. The half-line difference between the two diagrams may not affect the horizontal synchronizing generator operation, but it can cause the vertical oscillator to slip out of control.

To have the vertical pulse oscillator receive the necessary triggering voltage at the same time after every field, a series of six equalizing pulses is inserted into the signal immediately before and after the vertical synchroniz-

ing pulses. These equalizing pulses, shown in Fig. 12-22, do not disturb the operation of either oscillator (as will be shown later), yet they do permit the vertical pulse to occur at the correct time after every field.

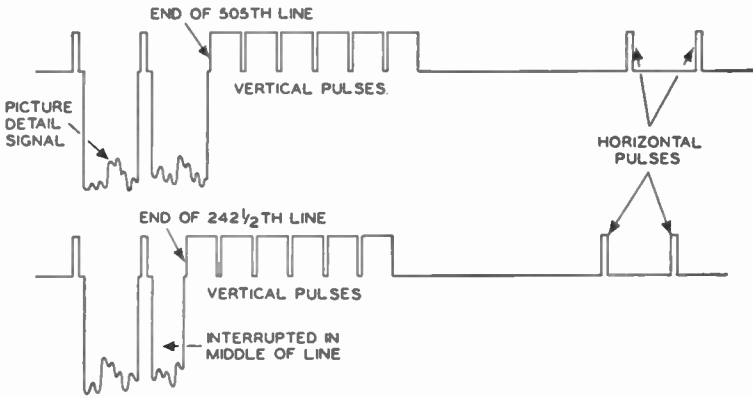


FIG. 12.21. The form of the video signal at the end of 242½ and 505 lines. The equalizing pulses are not shown here.

Once the serrated vertical pulse is ended, the six equalizing pulses are again inserted in the signal, and the line detail assumes control while the next field is swept out. One vertical pulse occurs at the end of every 262½ lines, while a horizontal pulse appears at the end of each line.

Vertical and Horizontal Separation. The separation of the vertical and horizontal pulses from each other is based on their frequency (or wave

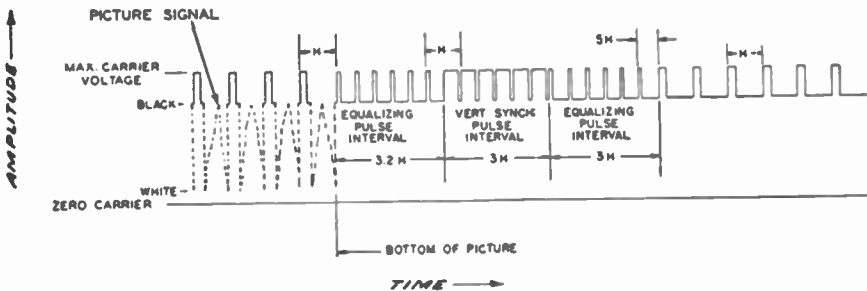


FIG. 12.22. The position of the equalizing impulses in the video signal.

form) difference and not on their amplitude since the latter is the same for both. The two pulses are compared in Fig. 12-23. Note that the horizontal pulse is much shorter in duration than the vertical pulse, rising and falling in 5 microseconds. Essentially, then, a low-pass filter will develop the vertical pulse voltage at its output, while a high-pass filter will have only the horizontal pulse voltage at its output. These two distinct pulses can then be

fed to their respective oscillators, controlling them in accordance with the dictates of the signal being received.

The operation of a filter and its effect on a wave are not difficult to understand. The filters employed separate the vertical and horizontal pulses from each other and then modify their waveform so that they are suitable for controlling the frequency of the oscillators that follow. To see how this

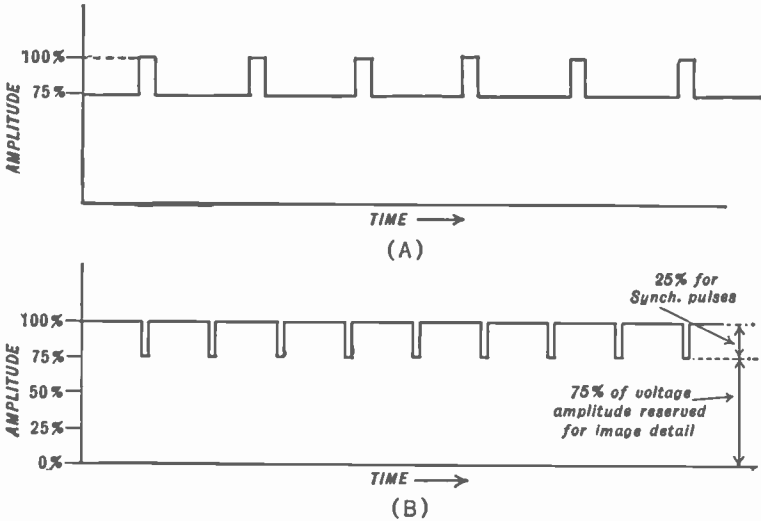


FIG. 12.23. A comparison of the waveforms of horizontal (A) and vertical (B) pulses.

occurs, let us apply a square-top pulse to the high-pass filter shown in Fig. 12-24, the output being obtained from the resistor.

At the application of the first edge of the square-wave pulse (known as the leading edge), a momentary flow of current takes place through the resistor to charge the capacitor fully to the value of the applied voltage. Once the full value is reached, nothing further occurs all along the flat portion of the pulse because a capacitor (and hence, a capacitor and resistor in series) reacts only to changing (or a-c) voltages, not to steady (or essentially d-c) voltages. The voltage along the flat top of the pulse is steady. At the next (or lagging) edge of the pulse, where the voltage drops suddenly, another momentary flow of current takes place, this time in the opposite direction, discharging the capacitor. The result of the application of the square-wave synchronizing pulse to the input of the high-pass filter is the output wave indicated in Fig. 12-24.

Each incoming synchronizing pulse gives rise to two sharp pulses at the output of the filter, with one above and one below the reference line. This,

of course, is due to the fact that one is obtained when the front edge of the incoming pulse acts on the filter and one when the lagging edge arrives.

For control of the sweep oscillator, only one of these two output pulses is required. If the first pulse at the output of the filter is negative (below the line) and a positive pulse is required, the conversion is readily made. Merely apply these pulses to an amplifier and the first pulse becomes positive. The amplifier introduces a phase shift of 180° , which is equivalent to reversing every value in a wave. The oscillators that are used, either the blocking or multivibrator types, respond to the first pulse, becoming insensitive immediately thereafter to other pulses that do not occur at the *proper point* in the oscillator frequency interval. When the next horizontal pulse

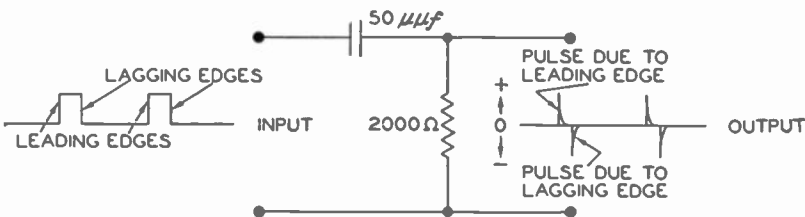


FIG. 12.24. A high-pass filter and its effect on the horizontal synchronizing pulses.

arrives, it is again in position to control the oscillator action. In this manner, any pulse occurring at an intermediate interval is without effect. One or two exceptions to this will be noted later.

The foregoing action of a high-pass filter indicates how the serrations of the vertical pulse permit control of the horizontal synchronizing oscillator during the application of the vertical pulse. In Fig. 12-25 are shown the input wave and the output pulses of a high-pass filter. Of all those present, only the positive pulses that occur at the proper time ($1/15,750$ of a sec) affect the horizontal oscillator. These active pulses are indicated by "A" in the figure. Note that all active pulses are evenly spaced and differ $1/15,750$ sec. The conditions shown in Fig. 12-25A occur only when the vertical pulses are inserted at the end of a line. Fig. 12-25B shows the situation when the field ends on a half line. Now the same equalizing and serrated pulse pips are not active in controlling the horizontal oscillator. Because of the difference in field ending, the control has shifted to those pips which were inactive in Fig. 12-25A. However, the shift has in no way interfered with the timing in the control pips. This shift from field to field illustrates why all the equalizing and vertical pulses are designed to produce pips twice in each horizontal line interval.

The long vertical pulses are without effect on this filter because of its small time constant and, further, because the output is obtained from across

the resistor. As soon as a vertical pulse is applied across the terminals of the filter, a short, sharp, current fully charges the capacitor. With the capacitor charged to the full voltage value, no further current flows through the resistor until another change occurs. The output is taken from across the resistor and, with current flowing only a very short time, a short, sharp pulse of voltage is obtained. At the lagging edge of the input wave, an-

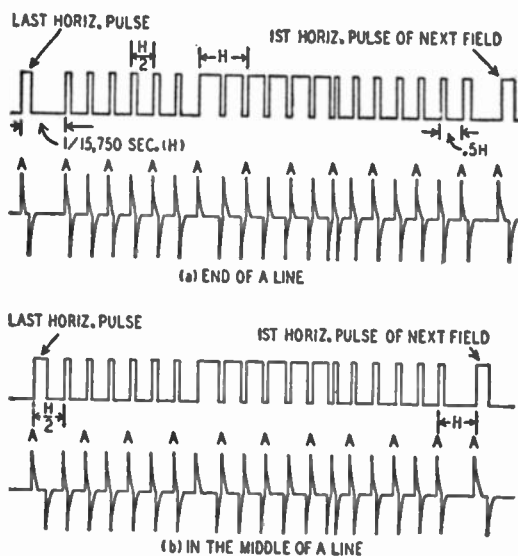


FIG. 12.25. Conditions during vertical pulses. The pips labeled "A" control the horizontal sweep oscillator.

other quick flow of current brings the capacitor voltage back to its previous value and again a voltage pulse develops across the resistor. Hence, only *changes* in the input wave appear across the output resistor, because it is only at these times that a current flows in the filter, either to charge or discharge the capacitor. The serrations inserted in the vertical pulse provide the changes that cause current to flow in the high-pass filter. Thus control can be maintained at the horizontal oscillator even when the vertical pulse is acting.

Vertical Pulse Filters. For vertical pulse separation, we use a low-pass filter of the type shown in Fig. 12-26. This appears identical with the high-pass filter, except that the positions of the condenser and resistor have been interchanged and the output is obtained from the capacitor. Besides the difference of position, the time constant of the capacitor and resistor is much greater than that of the previous filter. A long time constant means that the capacitor will charge and discharge slowly and will not respond as readily

as the previous filter to rapid changes in voltage. Hence, when a horizontal pulse arrives at the input of this filter, its leading edge starts a *slow* flow of current through the resistor, and the capacitor begins to charge. But this charging process is slow and, almost immediately afterward, the lagging edge of the wave reaches the filter and reverses the current flow, bringing the capacitor back to its previous value. Very little change has occurred during this short time interval. And the vertical synchronizing oscillator is designed so that it does not respond to these small fluctuations.

What is true of the effect of the horizontal pulses on the vertical filter is even more true with respect to the equalizing pulses, which rise and fall much more rapidly. Essentially, then, we have eliminated the possibility of the higher frequency pulses affecting the operation of the vertical synchronizing generator. Fig. 12-27 shows the output voltage of the filter on the application of these higher frequency waves. Their voltage level is below the dotted line which represents the point that the voltages must reach in order to affect the generator.

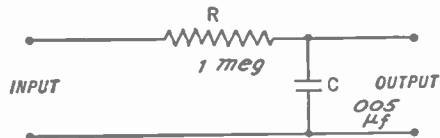


FIG. 12.26. A low-pass filter. The capacitor combines (or integrates) all the serrated vertical pulses until the output voltage rises to the necessary level where the vertical sync oscillator will react.

The building up of the voltage across the capacitor for the output begins when the serrated vertical pulses are reached. Even though the pulse is serrated, it still remains above the reference line for a relatively long time. The capacitor charges slowly in the manner indicated in Fig. 12-27. The small notches in the wave are due to the serrations. At these points, for a small fraction of a second, the voltage drops and then rises again. As previously noted, these changes affect the horizontal filter but leave the vertical filter output substantially unchanged because of their rapid disappearance.

Equalizing Pulses. We can pause for a moment here and determine more clearly the reason for the equalizing pulses. In Fig. 12-28 is shown the build-up of vertical deflecting voltage across the output of the vertical filter, once for the vertical pulse that comes at the end of a line, and once for the pulse that comes in the middle of a line. In the top illustration of Fig. 12-28, we see that each horizontal pulse causes a slight rise in voltage across the output of the vertical filter, but this is reduced to zero by the time the next pulse arrives. Hence there is no residual voltage across the vertical filter due to the horizontal pulses. Only when the long, serrated vertical pulse arrives is the desired voltage increase obtained.

However, the situation in the lower illustration of Fig. 12-28 is slightly different. Here the last horizontal pulse is separated from the first vertical

pulse by only half a line. Any horizontal voltage developed in the vertical filter will thus not have as much time to reach zero before the arrival of the first vertical pulse. This means that the vertical build-up does not start from zero, as in the top illustration, but from some small voltage value. As

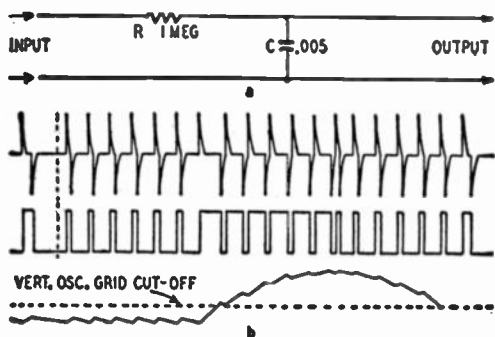


FIG. 12.27. (a) Low-pass filter for separating the vertical and horizontal pulses; (b) the waveform of the rise in voltage across the capacitor due to vertical pulses.

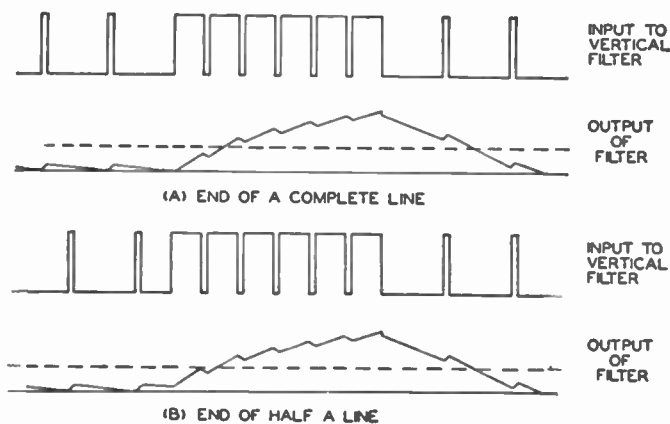


FIG. 12.28. The difference in voltage conditions before a vertical pulse when no equalizing pulses are used.

a result, the dotted line is reached sooner than if the voltage rise had started from zero. Since the dotted line represents the firing point of the vertical oscillator, we see that this oscillator is triggered a small fraction of a second too soon. The time actually involved is quite small, but it does prove sufficient to upset the precision interlacing of modern television images.

With the insertion of equalizing pulses before and after every vertical pulse, the voltage level established before the start of each vertical serrated

pulse is essentially the same, and the vertical oscillator is triggered at the proper moment in each instance.

After the complete vertical pulse has passed through the filter, the charge on the capacitor output gradually returns to the small value it had previously, the voltage due to the horizontal pulses (see Fig. 12-27). These pulses develop a very small voltage, far from sufficient to affect the vertical oscillator. Only the larger, longer vertical pulse $\frac{1}{60}$ of a second later can accumulate enough voltage to trigger the oscillator.

From a comparison of the vertical and horizontal pulse forms shown in Figs. 12-24 and 12-27, we may get the impression that the vertical pulse is not very sharp. This is because the vertical pulse is shown extended over quite a few horizontal pulses and the comparison exaggerates the extent of the vertical pulse. If the vertical pulse were drawn to a larger interval, then it too would appear sharp. So far as the vertical synchronizing oscillator is concerned, this pulse occurs rapidly and represents a sudden change in voltage.

The polarity of the pulses, as obtained at the output of their respective filters, may or may not be suitable for direct application to the controlled synchronizing oscillators. It all depends upon the type of oscillator to be controlled. For a blocking oscillator the leading pulse must be positive. If a multivibrator type of oscillator is employed, either a positive or negative pulse may be used, depending upon where it is introduced. This will be more fully developed presently.

Synchronizing Oscillators. From the block diagram of the components of the synchronizing section of a receiver (Fig. 12-1), we see that the oscillator is the next stage in the path of the synchronizing pulse. Before actually studying the operation of the stage itself, let us first review its function.

The electron beam must move across the screen at the regular rate of 15,750 times a second. Its path, as explained in Chapter 1, is not straight across the screen, but tilted slightly downward. At the end of the line, it is brought rapidly back to the left-hand side of the screen. The type of voltage at the horizontal and vertical deflecting plates that will accomplish this distinct motion is the saw-tooth wave drawn in Fig. 12-29. This wave gradually rises linearly and then, when it reaches a certain height, returns rapidly to its starting value.

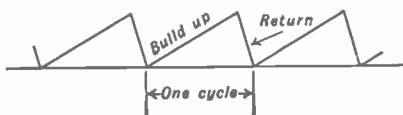


FIG. 12.29. A saw-tooth wave. Either the voltage or the current may assume this form.

A capacitor, connected to a battery (or other source of supply) through a resistor, will gradually charge in a manner approximately as shown in this figure. When the capacitor voltage reaches a predetermined value, a pulse

from the oscillator completely discharges it and the voltage build-up begins again. Without going much further into this sequence at this point, we see that the pulse from the synchronizing oscillator controls the action of the charging capacitor and, through it, the action of the electron beam across the screen. The oscillator itself is controlled by the pulse in the television signal. In this way the entire network is tied together and coordinated.

It may perhaps occur to some that, if the function of the synchronizing oscillator is merely to transmit pulses along to the charging capacitor, would it not be possible to apply the synchronizing pulses directly, without the intervening oscillator. The answer is Yes. However, the reason this is never done in practice is because there are times when no television signal is being received or when the signal is so weak that its pulses are not strong enough to actuate the capacitor. At these times, and with no oscillator, the electron beam would just remain at one point on the screen. The result — a burned out screen. With an oscillator, though, the electron beam is continually swept across the screen, signal or no.

The Blocking Oscillator. The blocking oscillator is one of two popular synchronizing oscillators used in modern television receivers. In common with all oscillators, feedback of energy from the plate to grid must occur. A transformer is employed for this purpose. Any change of current in the plate circuit will induce a voltage in the grid circuit which will act to aid this change. To examine this situation in detail, consider the operation of the oscillator when a disturbance occurs in the circuit acting to increase the plate current. To aid this increase, a positive voltage is induced in the grid through transformer T (see Fig. 12-30A). With the grid more positive than before, more plate current will flow, resulting in the grid becoming rapidly very positive. A positive grid means that electrons will flow in this circuit, charging capacitor C_2 . The electrons reaching the grid pile up on the right-hand plate of C_2 . With resistors R_2 and R_3 low in value, the charge on the capacitor would leak off rapidly and the action of the oscillator would continue. In practice, however, R_2 and R_3 are made high, combining with C_2 to give a long time constant. The electrons on C_2 discharge slowly to the cathode, placing a negative voltage on the grid, as shown in Fig. 12-30A.

Because of the slow discharge of C_2 , electrons which have accumulated on the grid remain there in sufficient numbers to give it a large negative bias, sufficient to block or stop the plate current flow. Gradually the electrons accumulated on C_2 pass through R_2 , R_3 , and R_1 back to the other plate of C_2 . Then the negative bias on the grid slowly becomes less. When the discharge is almost complete, electrons from the cathode once again reach the plate, plate current starts up, quickly reaches its high value, drives the grid positive, and the process repeats itself. Thus, during every cycle there is a

short, sharp pulse of plate current, followed by a period during which the tube blocks itself until the accumulated negative charge on the grid leaks off again. The frequency of these pulses is determined by C_2 , R_1 , R_2 , and R_3 .

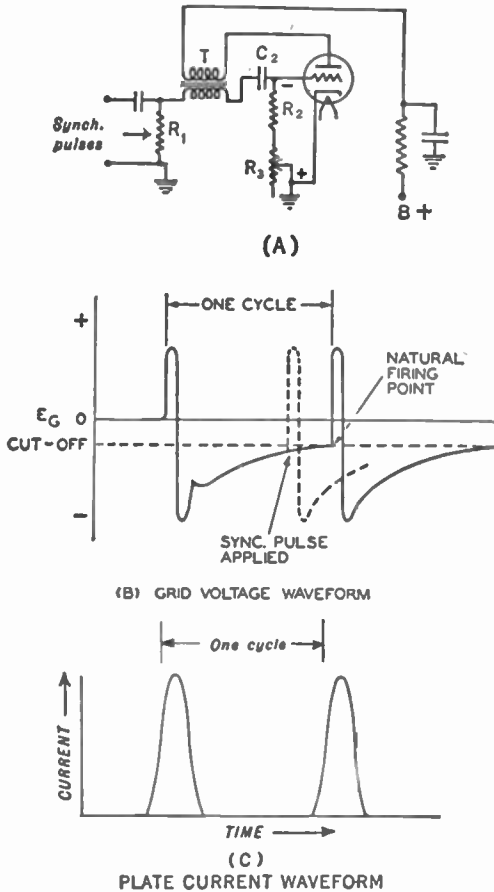


FIG. 12.30. A blocking oscillator (A). The grid voltage variations are illustrated at (B) while the form of the plate current is given at (C).

The form of the voltage drop across R_2 and R_3 is shown in Fig. 12-30B. In C, the plate current pulse occurs once in every cycle. It is possible to control the frequency of this oscillator if a positive pulse is injected into the grid circuit at the time indicated in Fig. 12-30B. To be effective, the frequency of the controlling pulse must be near and slightly higher than the free frequency of the oscillator. By free frequency, we mean the natural frequency at which it will oscillate if permitted to function alone. This is controlled by C_2 , R_1 , R_2 , and R_3 .

The point at which the synchronizing pulse should be applied to the grid of the oscillator is illustrated on the curve of Fig. 12-30B. A positive pulse, applied to the oscillator grid when it is at this point of its cycle, will bring the tube sharply out of cut-off and cause a sharp pulse of plate current to flow. Then, at the application of the negative pulse of the horizontal synchronizing voltage which follows immediately, the oscillator is no longer in any position to respond. The grid has now become so negative that it is unaffected by the second negative synchronizing pulse. It is only when the grid capacitor C_2 is almost completely discharged, that any pulse will *effectively* control the oscillator's frequency. This accounts for the firm control of the correct horizontal pulses. Equalizing pulses which occur at the half-way point in the oscillator cycle do not possess sufficient strength to bring the tube out of cut-off. It also explains why a positive synchronizing pulse is required, as stated several paragraphs before.

In short, then, it is observed that the synchronizing pulse controls the start of the oscillator's cycle. If left alone, the oscillator would function at its natural period which, more often than not, would not coincide with the incoming signal. Through the intervening action of the synchronizing pulse, both oscillator and signal are brought together, in step. Naturally, for effective control, both synchronizing pulse and oscillator frequency must be close enough together to permit locking-in.

Resistor R_3 is made variable in order to provide adjustment of the oscillator frequency. It is commonly known as the "Hold Control" since it can be varied until the frequency of the blocking oscillator is held in synchronism with the incoming pulses. It is generally placed in the rear of the television receiver where a serviceman may reach it easily for any necessary adjustment.

The output from the oscillator may be taken from either the plate or the grid circuits and used directly on the charge and discharge capacitor, or it may be applied through another tube. The simplest method of obtaining the saw-tooth deflection waves is shown in Fig. 12-32. But before this diagram is analyzed, it may be helpful if we discuss briefly the saw-tooth wave and its properties.

Saw-tooth Waves. The desired shape that the saw-tooth waves should have is shown in Fig. 12-29: a long, straight, gradual rise in voltage until a predetermined value is reached, and then a quick, sudden drop to the initial starting level. The process then repeats itself, 15,750 times a second for the horizontal oscillator and 60 times a second for the vertical oscillator.

Practically, the simplest way of obtaining the gradual rise in voltage followed by a sudden drop is by charging and discharging a capacitor. If a capacitor is placed in series with a resistor and a source of voltage, the

flow of current through the circuit will cause the voltage across the capacitor to rise in the manner shown by the curve of Fig. 12-31. This curve is not linear along its entire length, but the approximation to linearity at the beginning section of the curve is close enough for most practical purposes. Hence, if the capacitor is discharged just as it reaches point A on the curve, we will have a satisfactory saw-tooth wave suitable for application to the deflecting plates of a cathode-ray tube. The discharge of the capacitor should be as rapid as possible since during the time the capacitor is discharging the electron beam is blanked out at the tube and no picture detail is appearing on

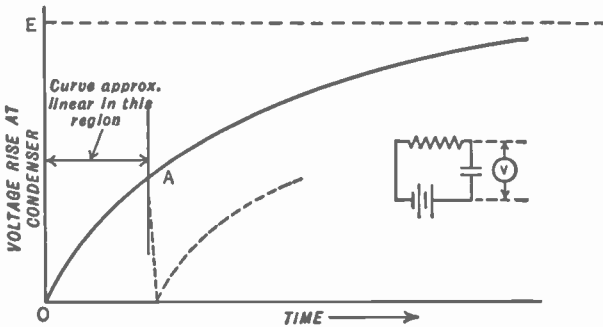


FIG. 12.31. The manner in which a voltage across a capacitor increases when a potential is applied through a series resistor.

the screen. The shorter the time spent in discharging the capacitor, the greater the interval during which the useful portion of the video signal may be acting at the screen.

A simple and inexpensive method of charging and discharging a capacitor to produce the necessary saw-tooth waves is given in Fig. 12-32. Triode V_1 is connected as a blocking oscillator, and the charge and discharge capacitor is placed in the plate circuit. From the preceding discussion of the operation of these oscillators, we know that a short, sharp pulse of plate current flows once in every cycle. During the remainder of the time, the grid is negatively biased beyond cut-off and no current flows in the plate circuit.

Throughout the time when no plate current is flowing, capacitor C_1 is charging because one side of this capacitor connects to the positive terminal of the power supply through resistors R_1 and R_2 , and the opposite plate of the capacitor is attached to ground. The charge this capacitor absorbs assumes the polarity shown in Fig. 12-32.

When plate current does start to flow, it is only for a very short period, and during this time the resistance of the tube becomes very low. Capacitor C_1 , which is actually in parallel with the tube, then quickly discharges through this low resistance path. At the end of the short pulse of plate

current, the grid has been driven very negative by the accumulation of electrons in C_2 and the tube becomes non-conducting again. C_1 no longer has this easy path for discharging and slowly starts to charge, as previously explained. The saw-tooth variation in voltage across C_1 is transmitted to the next tube, an amplifier, through coupling capacitor C_c . The process repeats itself, either at the horizontal scanning frequency or at the vertical frequency, depending upon its constants.

It will be noted from the foregoing action that, the instant the synchronizing pulse arrives at the oscillator, it triggers the oscillator, the tube becomes conducting, and the capacitor developing the saw-tooth voltage discharges.

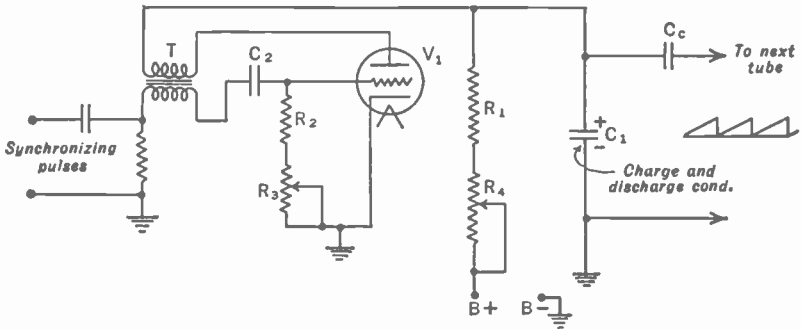


Fig. 12.32. A simple method of obtaining a saw-tooth voltage from a capacitor. This voltage, when applied across the cathode-ray deflecting plates, will control the action of the electron beam in the desired manner.

Hence, whenever a pulse arrives at the grid of the blocking oscillator, the capacitor discharges and the electron beam is brought back from the right-hand side of the screen to the left-hand side. This action is true in all such synchronizing oscillators.

Resistor R_4 is made variable to permit adjustment of the width of the picture. As more of its resistance is placed in the circuit, the amount of charging current reaching C_1 is lessened, with a subsequent decrease in the voltage developed across C_1 during its period of charging. A small voltage variation at C_1 means, in turn, a small voltage applied to the deflecting plates. The length of the left to right motion of the electron beam is consequently shortened, resulting in a narrower picture at the viewing screen. This is the reason for labeling R_4 the "Width Control." In the vertical synchronizing circuit, this same control would affect (and adjust) the height of the picture. Here it would be labeled the "Height Control."

In some television receivers, an extra tube is used to discharge the saw-tooth voltage generating capacitor. The circuit, shown in Fig. 12-33, is basically as simple as the preceding method. A 6SN7 double triode is used, with the first tube functioning as the blocking oscillator. The syn-

chronizing separator tube feeds the oscillator (through its grid circuit) the necessary pulses and synchronizes the oscillator's frequency to that of the incoming signal. The blocking oscillator then goes through its cycle, the grid becoming positive for a small fraction of a second, drawing grid current, and almost immediately thereafter developing a very negative grid charge that gradually leaks off through the variable resistor R_1 .

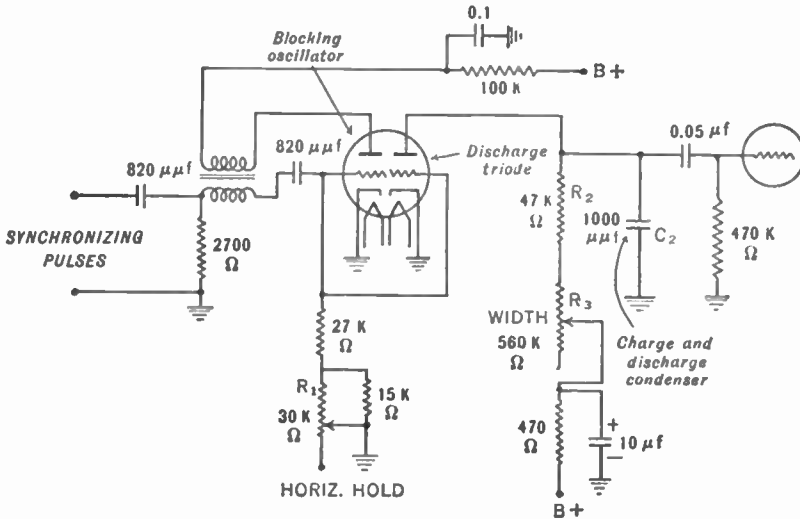


FIG. 12.33. A commercial circuit used for obtaining saw-tooth voltages in television receivers. The double triode is a 6SN7 or its equivalent.

The grid of the discharge tube is connected directly to the grid of the first triode and hence goes through the same voltage variations as the blocking oscillator grid. When the grid is negative, capacitor C_2 is charging in the plate circuit of the discharge tube, since no plate current is flowing at this time. The capacitor is charging through resistors R_2 and R_3 . The moment the grids of both tubes become positive, the discharge triode conducts, and C_2 discharges quickly. The next instant the grids are driven negative (beyond cut-off), and C_2 starts the charging process again. The remainder of the cycle is similar to that of the preceding circuit.

With the circuit of Fig. 12-33, a single tube need not be used for the dual functions of blocking oscillator and discharge tube. The use of two triodes (the two halves of 6SN7) permits each circuit to be designed and operated more efficiently.

Multivibrator Synchronizing Oscillators. In addition to the blocking oscillator, multivibrator generators have been favored by some manufacturers to serve as the synchronizing oscillator. Essentially, the multi-

vibrator is a two-stage resistance-coupled amplifier, with the output of the second tube fed back to the input of the first stage. Oscillations are possible in a circuit of this type because a voltage at the grid of the first tube will cause an amplified voltage to appear at the output of the second tube which has the same phase as the voltage at the grid of the first tube. This is always the case with an even number of resistance-coupled amplifiers, but never with an odd number. The output of an odd number of such stages is always 180° out of phase with the voltage applied at the input of the first tube. The two voltages thus oppose, rather than aid, each other.

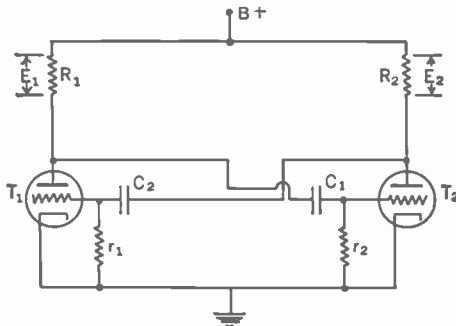


FIG. 12.34. The fundamental multivibrator circuit.

The operation of a multivibrator is best understood if we trace the voltage and current changes through the various circuit elements. To start, let us assume that the power supply has just been connected across the circuit. See Fig. 12-34. Due perhaps to some slight disturbance in the circuit, the plate current of tube T_1 increases. This produces an increase in the voltage across R_1 , with the plate end of the resistor becoming more negative. Capacitor C_1 , which is connected to R_1 at this point, likewise attempts to become more negative, and the grid of T_2 also assumes the same potential. The net result is a lowering of the current through T_2 and R_2 .

The lowered voltage across R_2 means that the plate end of this resistor becomes less negative, or relatively positive to its previous value. Capacitor C_2 transmits this positive increase to the grid of T_1 and, consequently, even more plate current flows through R_1 . The process thus continues in this manner, with the grid of T_1 becoming more and more positive and driving the grid of T_2 increasingly negative by the large negative charge built up across r_2 and C_1 . The plate current of T_2 is rapidly brought to zero by this action.

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Tube T_2 remains inactive until the negative charge on C_1 discharges and removes some of the large negative potential at the grid of T_2 . The path of discharge of C_1 is through the relatively low resistance r_p of tube T_1 and the relatively high resistance r_2 . When C_1 has discharged sufficiently, plate current starts to flow through R_2 , causing the plate end of the resistor to become increasingly negative. This now places a negative charge on the grid of T_1 , and the plate current through R_1 decreases. The lessening of the voltage drop at R_1 causes the plate end of the resistor to increase posi-

tively, and the grid of T_2 (through C_1) receives this positive voltage. The increased current through R_2 quickly raises the negative grid voltage on T_1 (through C_2) and drives this tube to cut-off. When the excess charge on C_2 leaks off, the process starts all over again. C_2 loses its accumulated negative charge by discharge through r_p of T_2 and r_1 . Contrast this path with that of C_1 .

The entire operation may be summed up by stating that first the plate current of one tube rapidly rises, driving the second tube to cut-off. This condition remains until the second tube is released from its cut-off state and

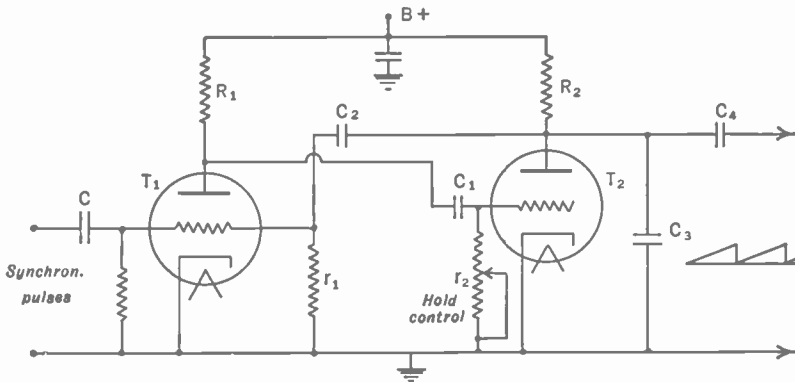


FIG. 12.35. How a multivibrator may be connected to control the charging and discharging of a capacitor to derive saw-tooth waves.

commences to conduct. It is now the first tube which is cut-off. When the first tube is again permitted to conduct, the second tube is driven into non-conduction. The switching continues in this manner, with the frequency largely determined by the grid resistors and condensers, r_1 , r_2 , C_1 , and C_2 .

If a synchronizing pulse is applied to either of the grids, and if its frequency is close to the natural frequency of the oscillator, it is possible to control the period of the multivibrator effectively.

Fig. 12-35 illustrates how the multivibrator may be employed to control the charge and discharge of a capacitor, thereby developing the required saw-tooth voltages. The same multivibrator is used, with the addition of the charge and discharge capacitor C_3 . When tube T_2 is not conducting, the power supply will slowly charge C_3 through resistor R_2 . The moment that the grid voltage of T_2 reaches the cut-off point of the tube, the tube starts to conduct and its internal resistance decreases. Capacitor C_3 then discharges rapidly through the tube. During the next cycle, T_2 is again non-conductive, and again C_3 slowly charges. C_4 transmits the voltage variations appearing across C_3 to the next amplifying tube. Resistor r_2 is made vari-

able to permit adjustment of the multivibrator so that it can be locked in with the synchronizing pulses. Hence r_2 is the hold control.

TYPICAL VALUES OF COMPONENTS SHOWN IN FIG. 12-35

<i>60 Cycles</i>	<i>15,750 Cycles</i>
T_1 — $\frac{1}{2}$ 6SN7	T_1 — $\frac{1}{2}$ 6SN7
T_2 — $\frac{1}{2}$ 6SN7	T_2 — $\frac{1}{2}$ 6SN7
r_1 —2.2 megohms	r_1 —220,000 ohms
r_2 —1.0 megohms	r_2 —100,000 ohms
R_1 —100,000 ohms	R_1 —47,000 ohms
R_2 —2.2 megohms	R_2 —500,000 ohms
C_3 —0.1 μ f	C_3 —500 μ μ f
C_1 —.01 μ f	C_1 —.001 μ f
C_2 —.05 μ f	C_2 —.005 μ f

The desired form of the saw-tooth synchronizing pulses is a slow rise in voltage, followed by a rapid decrease. Toward that end, C_1 and r_2 of Fig. 12-35 are designed to have a considerably longer time constant than C_2 and r_1 . C_1 and r_2 will discharge slowly, maintaining T_2 in cut-off while C_1 slowly charges. During this interval, T_1 is conducting. Upon the application of a negative synchronizing pulse to the grid of T_1 , this tube is forced into cut-off, while T_2 rises sharply out of cut-off and into conduction. C_3 now discharges rapidly. Because C_2 and r_1 have a small time constant, T_1 does not remain cut-off very long and as soon as C_3 has discharged, T_1 begins to conduct, again cutting off the plate current of T_2 . The ratio of the time constants of C_1 , r_2 and C_2 , r_1 is in the vicinity of 9 to 1.

In many commercial receivers, a slightly altered form of multivibrator circuit is used, although the basic operation remains the same. This oscillator is shown in Fig. 12-36. Feedback between tubes is accomplished in

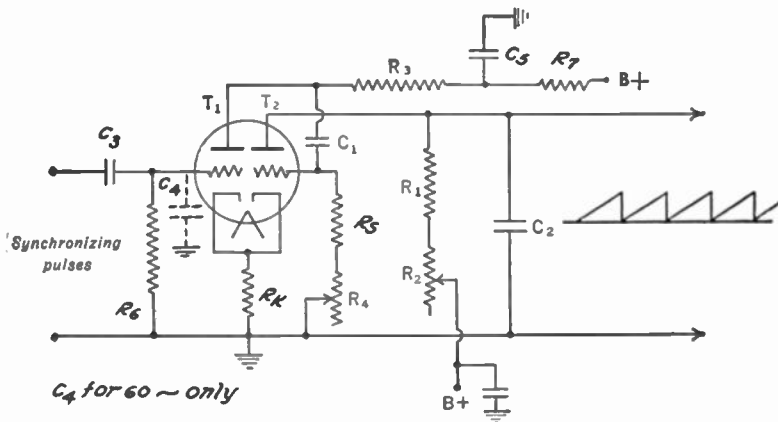


FIG. 12.36. Another widely used form of multivibrator. This is known as a cathode-coupled multivibrator.

two ways: through the coupling capacitor C_1 and the unby-passed cathode resistor, which is common to both tubes.

The charge and discharge capacitor C_2 is placed in the plate circuit of the second triode. During the portion of the multivibrator cycle when triode T_2 is not conducting, C_2 is essentially across the power supply and charges through resistors R_1 and R_2 . When a sharp negative pulse of voltage is applied to triode T_1 , the plate current of this tube decreases, causing the plate end of resistor R_3 to become increasingly positive. As the grid of T_2 is connected to this part of R_3 , it too will become more positive. The plate current through T_2 will rise sharply, developing enough voltage across the common cathode resistor to bring T_1 to cut-off. T_2 , however, continues to conduct because its grid has received sufficient positive voltage from the potential variation across R_3 to counteract partially this high negative cathode bias. T_1 , not having this positive grid voltage, is forced into cut-off. During this period, when T_2 is conducting heavily, its internal resistance is low and C_2 discharges through it.

The high positive voltage on the grid of T_2 , which resulted in a large plate current flow for an instant (and permitted C_2 to discharge), makes the grid draw current. This immediately biases the grid to cut-off (similar to the blocking oscillator), brings T_1 out of cut-off, and permits C_2 to charge again. Resistor R_4 is made variable to permit adjustment of the frequency of the multivibrator. R_2 controls the amount of the charging current flowing into C_2 , and this in turn regulates the extent of the electron beam sweep across the screen. It is the width control.

TYPICAL VALUES OF COMPONENTS SHOWN IN FIG. 12-36

<i>60 Cycles</i>	<i>15.750 Cycles</i>
T_1 } 6SN7	T_1 } 6SN7
T_2 }	T_2 }
R_1 —1.0 megohm	R_1 —470,000 ohms
R_2 —2.0 megohms	R_2 —500,000 ohms
R_3 —100,000 ohms	R_3 —47,000 ohms
R_4 —1.2 megohms	R_4 —50,000 ohms
R_5 —1.2 megohms	R_5 —33,000 ohms
R_6 —2.2 megohms	R_6 —2,000 ohms
R_7 —100,000 ohms	R_7 —100,000 ohms
C_1 —.01 μ f	C_1 —.001 μ f
C_2 —.1 μ f	C_2 —500 $\mu\mu$ f
C_3 —.01 μ f	C_3 —50 $\mu\mu$ f
C_4 —.001 μ f	C_4 —not necessary
C_5 —.1 μ f	C_5 —.006 μ f
R_k —470 ohms	R_k —470 ohms

Here, as before, the incoming synchronizing pulses serve to alter slightly the time at which a flip-over from one tube to the other takes place. With-

out these pulses, each tube would conduct for a portion of the cycle, just as in the case of the previous multivibrator.

While a negative synchronizing pulse at the grid of T_1 will cause C_2 to discharge, we may obtain the same effect if a positive synchronizing pulse were fed to the grid of T_2 . The negative pulse, however, results in a more stable arrangement and is generally used. This fact explains the statement made several paragraphs before when it was pointed out that either a positive or negative synchronizing pulse could be used to actuate a multivibrator. For the blocking oscillator, it will be remembered that a positive pulse was required.

The phrase "synchronizing an oscillator" is quite frequently used when describing the operation of television circuits. There are, however, many

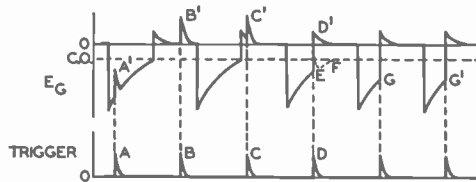


FIG. 12.37. Illustrating how the incoming pulses lock-in the synchronizing oscillators.

technicians who are not completely clear as to the exact mechanism of this synchronization. To clarify this point the following explanation is offered.

In a television receiver, the pulses of the incoming signal take control of the free-running sweep oscillators and lock them into synchronism with the pulse frequencies. We are referring, of course, to the horizontal and vertical synchronizing pulses. It is highly improbable that the first pulse, when it reaches the oscillator, arrives at such a time as to force the free-running oscillator exactly into line. Generally, this does not occur until after several pulses of the incoming signal have reached the sweep oscillator. Let us examine the means whereby the receiver oscillator is gradually forced into synchronization with the incoming pulses.

In order to synchronize an oscillator, the pulses must be applied to the oscillator input. In Fig. 12-37 we have the grid voltage waveforms of a multivibrator and, beneath them, the triggering pulses as they are received from the preceding pulse separator networks. Suppose the first pulse, at A , arrives at a time when the grid is quite negative and thus this pulse is unable to bring the tube out of cut-off. The second pulse, at B , arrives when the tube is conducting. Thus, it drives the grid more positive and has very little effect on its operation. The conditions for the third pulse are similar to those for the second pulse. The fourth pulse, at D , arrives at a time when

the grid of the tube is negative. However, this pulse is able to drive the grid positive, thereby initiating a new cycle. Thereafter, each succeeding pulse arrives at a time when it will bring the tube out of cut-off and the sweep oscillator is securely locked in as long as the pulses are active. It is important that the pulse reach the grid of the oscillator when it can raise the tube above cut-off. Unless it can do this, it will be without power to lock in the oscillator.

One final word about the foregoing oscillators. As the grid voltage approaches the cut-off value, it becomes increasingly sensitive to noise pulses which may have become part of the signal. A sufficiently strong interference pulse, arriving slightly before the synchronizing pulse, could readily trigger the oscillator prematurely. When this occurs, the electron beam is returned to the left-hand side of the screen before it should and the right-hand edge becomes uneven. Severe interference causes sections of the image to become "torn" (see Chapter 16). To prevent this form of image distortion, many television receiver manufacturers have designed synchronizing systems which respond only to long period changes in the pulse frequency. Since interference flashes seldom have regular pattern, they cannot affect these special systems. Several such systems are analyzed in Chapter 13.

We have now completed a description of the components of an ordinary television synchronizing section. In the next chapter several commercial synchronizing systems are examined in detail to illustrate how the various components are combined to form an integrated unit.

QUESTIONS

1. Do the horizontal and vertical synchronizing pulses ever reach the control grid of the cathode-ray tube? Explain.
2. Draw a diagram of a complete video signal indicating where the horizontal synchronizing pulses are located.
3. Illustrate the path taken by all synchronizing pulses in a television receiver.
4. State specifically the action of the horizontal and vertical pulses in controlling the motion of the electron beam.
5. Explain what precautions must be observed before the pulses can be separated from the rest of the video signal.
6. Draw the circuit of a diode clipper. Explain how it operates.
7. Why are pentode or triode clippers more desirable than diode clippers?
8. Draw the circuit of a pentode clipper stage indicating the values to be assigned to the B+ voltage.
9. Why do the blanking voltages last longer than either the vertical or the horizontal pulses? What would happen if the blanking voltages were too short?
10. Explain why serrated vertical pulses are employed. Draw a serrated vertical pulse, indicating its time duration in micro-seconds.
11. What is an active line? What is the approximate number of active lines per frame? Indicate how your figure was obtained.

12. Indicate the position and time duration of the equalizing pulses.
13. Why are equalizing pulses necessary? Illustrate your answer by means of diagrams.
14. Draw the diagram of a differentiating network and explain its operation.
15. Draw an integrating network and explain its operation.
16. Do the same vertical and equalizing pulses trigger the horizontal oscillator after every field? Explain.
17. Illustrate the action of a vertical pulse in triggering the vertical sweep oscillator.
18. How are saw-tooth deflection voltages developed? What are the limitations of this method?
19. Draw the diagram of a blocking oscillator.
20. Explain the operation of blocking oscillators.
21. To control effectively the sweep oscillator, should the pulse frequency be higher or lower than the oscillator frequency? Why?
22. Draw the circuit of a cathode-coupled multivibrator.
23. Explain briefly how the multivibrator functions.

Chapter 13

DEFLECTION SYSTEMS

An Electrostatic Deflection System. The schematic of an electrostatic deflection system is shown in Fig. 13-1. This circuit was chosen for analysis because it is representative of many of the electrostatic deflection systems found in television receivers.

We pick up the signal path at the grid of the final video-frequency amplifier. The video signal is amplified by the tube and applied to the cathode of the image tube and to a 6AU6 pentode functioning as a d-c restorer and sync separator. This portion of the circuit was previously described in Chapter 12, and it was then shown that a d-c restoration voltage would develop across the cathode resistor of the 6AU6 while, at the same time, only the vertical and horizontal sync pulses would pass through the tube and appear in the plate circuit. The sync pulses are then fed to the vertical and horizontal deflection systems. Thus far, note that we have only achieved sync pulse separation from the rest of the video signal. The vertical and horizontal sync pulses have not been separated from each other.

The d-c restoration voltage is applied to the grid of the cathode-ray tube, while the complete video signal is received by the cathode of the tube. This arrangement is employed here because the video signal polarity, at the output of the video amplifier, is such that the sync pulses are positive with respect to the rest of the signal.

To separate the vertical and horizontal sync pulses, integrating and differentiating filters are placed before the vertical and horizontal oscillators respectively. The differentiating network inserted in the grid circuit of the horizontal multivibrator consists of a 220- μmf capacitor and a 4,700-ohm resistor. The integrating network placed in the signal path leading to the vertical sweep oscillator (also a multivibrator) consists of two 4,700-ohm resistors and two 0.005- μf capacitors. With these filters, only the proper sync pulses reach each sweep oscillator.

Turning our attention first to the horizontal sweep system, we see that it

consists of two duo-triodes (12SN7's). The first 12SN7 is the multivibrator, operating at a frequency of 15,750 cycles. A 250,000-ohm potentiometer permits variation of the horizontal frequency and is the hold control. The saw-tooth deflection voltage is developed across C_1 in the plate circuit of the second triode section of V_1 . A width control regulates the amount of voltage developed across the capacitor. One triode section of V_2 receives the saw-tooth voltage, amplifies it, and applies it to one horizontal deflection plate of the 7JP4 image tube. The same triode also feeds a small portion of this voltage to the second triode section of V_2 whose plate is connected to the other horizontal deflection plate in the cathode-ray tube. In this manner, push-pull deflection is achieved.

The vertical system, containing V_3 and V_4 , is almost identical in design. A 1-megohm potentiometer in the multivibrator circuit controls the vertical sweep frequency, and a 10-megohm potentiometer regulates the height of the image. C_2 is the charge and discharge capacitor, developing the necessary saw-tooth deflection voltage which, after passage through V_4 , is applied equally, but in opposite phase, to the vertical deflection plates in the 7JP4. A variable resistor, R_1 , serves as the vertical linearity control. Its function is to correct the shape of the saw-tooth waves so that they rise more linearly. The name of the control is derived from this action. The need for correction arises from the tendency of the charging voltage across C_2 to increase in a manner not quite linear. This fact was pointed out in a preceding paragraph where it was stated that, if only a very small portion of the curve were used, the resulting wave would be satisfactory. This situation is still true. However, if only the most linear portion of the charging curve is used, the voltage developed across the charging capacitor is small. This requires a considerable amount of amplification, accomplished only through additional amplifiers. As a practical solution it is customary to permit the voltage across the charging capacitor to extend somewhat beyond the linear portion, and to compensate for this non-linearity by a special control. Good results are obtained in this manner and one push-pull stage of amplification following the charge-and-discharge capacitor is usually sufficient.

In this system, R_1 regulates the amount of voltage fed back from the output of the vertical sweep amplifiers to the input and in this way is able to offset tapering or rounding off of the deflection voltage developed across C_2 .

In both multivibrators (V_1 and V_3), the grid and cathode of each triode section are both returned to -140 volts instead of to ground. Since the voltage is applied equally to both elements, it does not constitute a grid bias voltage. Instead, it is used here, in conjunction with a positive 400 volts applied to the plates of the tubes, to provide an effective plate voltage of 540 volts. Thus a fairly linear saw-tooth voltage of large amplitude can be developed across C_1 and C_2 .

In addition to the saw-tooth deflection voltages, each of the deflection plates receives a d-c voltage for the purpose of centering the electron beam. Two centering controls, one for the vertical set of plates and one for the horizontal set, permit variation of the d-c voltage on one of the deflection plates of each set. The d-c voltage on the other plate is kept constant.

The high-voltage power supply shown at the bottom of the diagram is of the R.F. type described in Chapter 10. It develops 5,000 volts which are applied to the centering controls and a bleeder network. At appropriate points along the bleeder network, voltages are tapped off for the accelerating electrode in the 7JP4 and for the first anode. The latter voltage is made variable, providing a focusing adjustment of the electron beam. Low voltages (those applied to the remaining tubes in the set) are developed separately by a power supply using diode and selenium rectifiers. Voltages available are +250 volts, +400 volts, +120 volts and -140 volts. The +250 volts is used solely for powering the high-voltage R.F. supply.

Saw-tooth Current Waves. Up to this point, electrostatic deflection has been described in detail. We have seen that a saw-tooth voltage wave is necessary at the deflecting plates in order to swing the electron beam properly across the screen. The simple charging and discharging of a capacitor is sufficient to obtain the desired wave shape. If the same saw-tooth voltage is applied across the coils of an electromagnetic deflecting system, it will be found that the electrons no longer move across the screen in the desired manner. The reason is quite simple. In order to cause the electron beam to move slowly across the screen from left to right and then rapidly back to the left-hand side again, the beam must be subjected to a field of force that is varying in a saw-tooth wave manner. In the electrostatic case, a saw-tooth voltage at the plates will do this. In electromagnetic deflection, applying a saw-tooth voltage to the coils will not result in a saw-tooth current wave through the coils. And, since the magnetic flux varies directly with the current through the coil and not with the voltage across it, the flux variation will likewise differ from the necessary saw-tooth shape. If the charging and discharging of a capacitor is to be utilized at all, then a correction becomes necessary in order that the voltage applied across the deflecting coils will develop a saw-tooth current wave.

The final form of the voltage wave applied to the deflecting coils is derived by analyzing the components of the coils and their action when subjected to voltages of various shapes. Each coil contains inductance plus a certain amount of resistance. As far as the resistance is concerned, a saw-tooth voltage will result in a saw-tooth current. For the inductance, considering a pure inductance, a voltage having the form shown in Fig. 13-2B is needed for saw-tooth current flow. Combining both voltage waves, we obtain a resultant that varies in the manner shown in Fig. 13-2C. A voltage

of this type, when placed across the deflecting coils will give a saw-tooth current, and the magnetic flux, varying in like manner, will force the electron beam to sweep across the screen properly. Note carefully that the resultant wave is not obtained by combining the two voltage waves in *equal* measure.

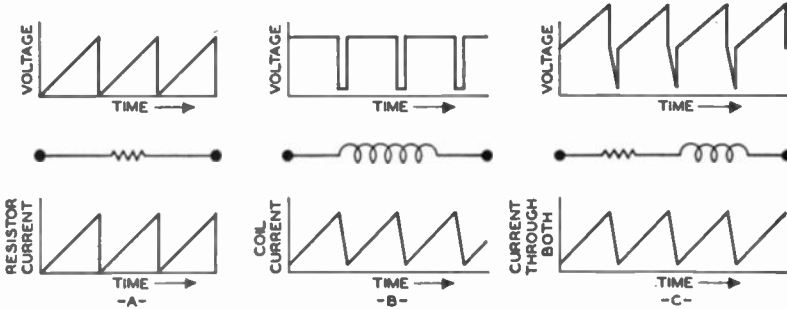


FIG. 13.2. By applying the voltage shown above electrical component, the saw-tooth waves shown below the component are obtained.

If the deflection circuit contains more inductance than resistance, the resultant wave will be closer in form to Fig. 13-2B. On the other hand, if the resistance predominates, then the resultant wave will resemble Fig. 13-2A more. Hence, one may expect to find variations of this horizontal deflection wave ranging from almost a pure saw-tooth wave to that shown in Fig. 13-2C.

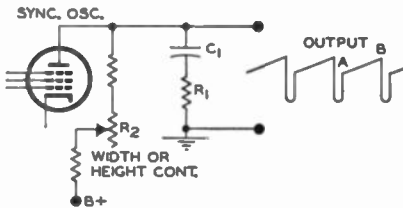


FIG. 13.3. A common method for generating deflecting voltages suitable for application to deflecting coils.

through resistor R_2 from the battery. During this period the voltage is rising from A to B. When the oscillator tube is triggered and the tube is conducting heavily, the plate voltage drops nearly to cathode potential. Capacitor C_1 discharges during this time. The conduction time, however, is short and C_1 is unable to discharge completely before the tube is again cut-off. The plate voltage does not have to rise slowly from cathode potential, but instead rises immediately to whatever voltage still remains across the capacitor. Thereafter, it rises slowly in a fairly linear manner until the arrival of the next pulse. R_2 is made variable to permit adjustment of the output voltage so that the proper size of the image may be attained.

The only difference, it is noted, between the methods for generating suitable deflecting voltages for electrostatic and electromagnetic systems is the components at the output terminals. For electrostatic deflection, the output is taken from a capacitor alone, whereas, for the electromagnetic deflection, a series resistor is included. Either combination may be used with the blocking oscillator or the multivibrator.

An Electromagnetic Deflection Unit. The system undertaken for this analysis is shown in Fig. 13-4. A portion of the entire video signal is taken from the plate circuit of the first video-frequency amplifier, 6AC7, and applied to the control grid of the 6SG7 sync separator, V_1 . At this point, the video signal polarity is such that the sync pulses possess the most positive polarity. Since the cathode of V_1 is grounded, the tube does not have any grid bias until a signal is applied. Grid current then flows for the most positive portions of the signal (here, the sync pulses), establishing grid-leak bias across C_1 and the connecting grid resistor R_1 . With the low plate voltage on the tube (9.5 volts), plate current will flow only at the most positive portions of the video signal, and only the sync pulses will appear across the plate load resistor.

The output of V_1 is now sent to two separate amplifiers. One portion of the signal goes to V_{3A} where the now negative sync pulses are amplified, inverted, and then fed to the grid of the vertical blocking oscillator V_{3B} . An integrating network consisting of R_2 (39,000 ohms) and C_2 (0.01 μ f) prevents any except the vertical sync pulses from developing enough voltage across C_2 to trigger the vertical blocking oscillator. V_{3A} , while amplifying the pulses, also functions as a clipper because of its low plate voltage (34 volts), tending further to clip off any video signal still remaining at this point in the circuit.

The vertical blocking oscillator is conventional in form, being triggered when the vertical sync pulse voltage across C_2 has attained the proper value. The peaked saw-tooth deflection wave necessary for magnetic deflection is developed across C_3 and R_3 . This wave is fed to V_5 , where two triode sections of a 6SN7, connected in parallel, amplify the deflection wave and then transfer it through an output transformer to the vertical section of the deflection coil. A variable resistor of 5,000 ohms is inserted in the cathode circuit of V_5 to provide control of vertical linearity. The vertical blocking oscillator contains a hold control to adjust its frequency, if the oscillator should drop out of synchronization, and a height control to adjust the vertical size of the image.

In the secondary circuit of T_1 , there are two resistors and a capacitor, R_5 , R_6 , and C_4 . The function of R_6 is to permit a small current to flow through the vertical deflecting coils for centering the electron beam. A fixed tap is provided for one connection to the vertical coils, while the other end

of the coil is attached to the center, movable arm. In this way it is possible to have a small current flow either in one direction or another, or to have no current at all when the movable arm is at the tap. A similar arrangement is found in the horizontal deflecting circuit (resistor R_{12}).

Capacitor C_4 and resistor R_5 are designed to eliminate any tendency on the part of the deflecting coils to set up oscillations. In winding these coils, it is impossible to eliminate distributed capacitance between turns; hence the inductance of the coils and this distributed capacitance form a parallel resonant circuit (see Fig. 13-5). Every $\frac{1}{60}$ of a second, a sharp pulse of voltage develops across the circuit, and, if one of the frequency components

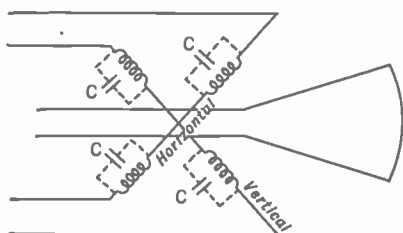


FIG. 13.5. The distributed capacitance inherent in deflecting coils.

of this pulse is near the resonant frequency of the coil inductance and distributed capacitance, oscillations will be set up. The result is the appearance of spurious lines at the top of the image. These occur here because the oscillations are set up immediately after the pulse of voltage appears across the coils and the beam, after the pulse, is at the upper side of the screen.

By inserting a simple RC circuit in parallel with the inductance of the deflecting coils (and their distributed capacitance) it becomes possible to damp quickly any oscillation that may be set up. As in any other parallel resonant circuit, the amplitude of the oscillations can be decreased if a relatively low resistance is placed across the circuit. The action resembles the loading of the input tuning circuits in the R.F. and I.F. stages. Further damping precautions consist of R_7 and R_8 , placed in parallel with each section of vertical deflecting coil (there are two connected in series), and the 6SN7 output triode. Its low plate resistance affords some damping in itself.

In the horizontal deflecting system, where the pulses have higher component frequencies, it is even more important that stringent measures be taken to eliminate these unwanted oscillations.

Turning now to the horizontal sweep system, we find that the sync pulses are received by V_{2A} from V_1 , amplified, clipped, and then fed through a differentiating network (R_9 and C_5) to the grid of the blocking oscillator V_{2B} . The peaked deflection voltage developed across C_6 and R_{13} is applied to V_4 , a 6BG6G pentode output amplifier, and then transferred through an output transformer to the horizontal deflection coils.

The output circuit of the horizontal deflection system differs in several respects from the same portion of the vertical system. For one thing, a damping tube (V_6) is employed instead of an RC damping network. An

ordinary RC parallel damping circuit would prevent very large voltages from being built up across the coils because of its loading effect, and large voltages are required when the pulse is quickly moving the electron beam from one side of the screen to the other. Changes occur more rapidly here than in the much slower moving vertical circuit. Furthermore, an RC filter at this point would lengthen the return time of the beam, an undesirable feature.

More efficient operation is obtained by using V_6 , whose purpose in this circuit is twofold. It functions as a damping circuit to suppress oscillations that are set up in the secondary during beam retrace, and, in addition, it converts some of the large amount of energy developed in this circuit into d-c potential which is combined with the d-c voltage of the power supply to provide an increased voltage to the horizontal output amplifier, V_4 .

In order properly to interpret the action which occurs in this circuit, the following relationships within the horizontal scanning interval must be known. A complete horizontal scanning cycle requires $1/15,750$ of a second, or approximately 64 microseconds. Of this, the visible portion of the horizontal trace is about 53 microseconds long. The blanking interval, then, is 11 microseconds long, and during this time the beam must be returned to the left side of the tube, the trace started and made linear. In order that all this be accomplished within the 11 microseconds, the return trace can be allotted only 7 microseconds.

During the trace period, the voltage across the yoke windings is constant, as shown in Fig. 13-6C. This will produce a linearly rising saw-tooth current in the coils. In addition, there is a small amplitude saw-tooth voltage which compensates for the resistance in the circuit. When the negative pulse of the applied wave reaches the grid of the output tube, the plate current is suddenly cut off since the tube is driven deep into cut-off.

The magnetic field, which has been steadily building up in the output transformer, begins to collapse. The rate of collapse is determined by the natural resonant frequency of the system; this is about 75 ke in order to insure a sufficiently short retrace period. The voltage generated by the collapsing field is negative on the damping tube, preventing this tube from conducting. Thus, there is actually no load across the transformer, and the system, if left in this condition, would oscillate vigorously. This is actually done for one half cycle. At the end of the first half cycle, the yoke current reaches a maximum value in the reverse direction to which it was flowing at the end of the trace period. The induced yoke voltage now reverses polarity, and the damping tube begins to conduct. By now beam retrace has been completed and the next trace must be started.

The energy which was developed in the coil by the output tube in the latter part of the last trace has not been completely dissipated. Very little

energy was lost during the first half cycle of retrace because the damping tube was non-conductive and the circuit had few losses. When, after the first half cycle, the damping tube does begin to conduct, it places such a heavy load across the deflection coil that further oscillations are prevented. The field begins to decay at a rate determined by the load of the damping tube across the coil.

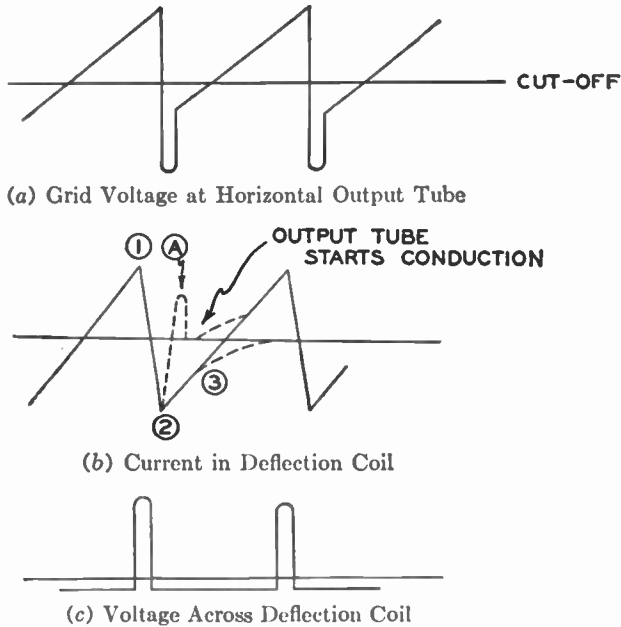


FIG. 13.6. The various current and voltage relationship in the horizontal deflection output circuit.

Before we proceed further, let us stop and note carefully the sequence of events. When the negative portion of the deflecting voltage drives the grid of the output tube to cut-off, the deflecting yoke, output transformer, and other incidental components form a resonant circuit which is shocked into oscillation. The time of a half cycle of this oscillation (75 kc) is about 7 microseconds and so the retrace occurs within its allotted time. At the end of a half cycle, the damping tube begins to conduct, causing the oscillatory voltage to decay in essentially a linear manner. Note, however, that sometime after this the output tube begins to conduct again and this additional power in conjunction with the decay current in the deflection yoke produces a linear trace motion of the electron beam.

The waveforms in Fig. 13-6 illustrate graphically the action within the circuit. Fig. 13-6A shows the voltage applied to the grid of the horizontal

output tube. In Fig. 13-6B we have the current in the deflection coil, as indicated by the heavy solid lines. At point (1), the tube is driven into cut-off, and the magnetic field collapses and reverses itself. At point (2) the retrace interval has ended, and the beam is now at the left-hand side of the screen ready for the next line scan.

The dotted curve A shows what would occur if the damping tube did not begin to conduct and prevent the oscillations from continuing. From point (2) to point (3) the energy remaining in the deflection yoke is decaying in a fairly linear manner. At point (3), this energy has begun to die off, and at this moment the output tube starts once again to conduct. The resulting interaction of these two currents is a linear rise in current. This sequence of events is repeated each cycle.

It might be of interest to the reader to note that the decay current, from point (2) to point (3), moves the beam about 40 per cent of the way across the screen starting from the left-hand edge. Then, the output tube takes over and drives the beam the remaining 60 per cent of the way to the right-hand edge.

The damping tube serves not only to prevent continued oscillations in the deflection coils after retrace, but also to convert some of this energy to a useful d-c voltage. At the time the horizontal output tube is brought into cut-off, a tremendous amount of energy is in the output transformer. Part of this energy is used to bring the electron beam from the right-hand side of the screen to the left-hand side. When this has been accomplished, a considerable amount of energy remains, which would result in continued oscillations if the damping diode did not begin to conduct at this moment, not only preventing the continuance of the oscillations, but also rectifying the pulse into d-c. This is the reason the plate of V_6 is labeled +280 volts, and the cathode, +350 volts. The plate voltage for the 6BG6G tube is supplied through the damping diode, and consequently this tube receives the benefit of the additional voltage.

A small variable coil is shunted across part of the secondary winding of the horizontal output transformer, and by varying the inductance of the coil we can vary the amount of voltage developed across this winding and hence the picture width.

The high-voltage power supply derives its power from the collapse of the field in the output horizontal transformer during the retrace interval. When the 6BG6G plate current is cut off, a positive pulse appears on the primary of the output transformer. This is stepped up by auto-transformer action and rectified by the 1B3-GT diode. Since the frequency of these pulses is 15,750 cycles per second, sufficient filtering is provided by a single 500- $\mu\mu\text{f}$ capacitor. The high voltage generated is fed to the second anode of the cathode-ray tube by way of a side seal on the side of the tube. C_8 , C_9 , and

L_1 filter out the 15,750-cycle ripple present in the 70 d-c volts developed by V_6 .

The horizontal output circuit and the attendant method of obtaining high voltage stems originally from the first commercial television receiver marketed by RCA back in 1946. This is essentially the circuit shown in Fig. 13-4. The horizontal output transformer employed a fairly large powdered-iron core. The original circuit was designed principally for 53° picture tubes requiring an accelerating potential of about 9,000 volts. The tube most frequently used in the horizontal output amplifier was the 6BG6G, a husky beam power tube capable of delivering a sizable amount of power.

As the television industry expanded and the demand for large screen tubes grew, the 10-inch screen gave way to the 12½, 14, 15, 16, 17, 19-inch and even larger screens. In order to keep the length of these large screen tubes within reason, the deflection angle was increased, first to 60°, then 66°, and finally 70°. With 27-inch tubes recently built, the deflection angle has been again increased, this time to 90°.

Now, increasing the deflection angle did reduce the overall length of the picture tube, but it required more high voltage and greater driving power. By employing two 1B3 rectifiers as voltage doublers, the high voltage could be increased; by paralleling 6BG6G's, the driving power could be made greater. But these solutions were not the most desirable since they necessitated more components and resulted in higher costs.

Through research and development, the television manufacturers developed a higher efficiency output transformer using a small ferrite core. Also, other output tubes were developed (such as the 6BQ6 and 6AU5) which are capable of greater power with less current drain on the B+ supply. Now, with these improved components, it is readily possible to develop voltages as high as 18,000 volts and to obtain sufficient power to sweep over the entire area of a 90° deflection tube.

In addition to the changes which were being made in the construction of the output transformer, changes were also appearing in the horizontal output circuit itself.

RCA, in some of their sets which followed the 630TS model, came up with the autotransformer arrangement shown in Fig. 13-7. The horizontal output transformer contains one winding to which are connected the 6BG6G tube, the 1B3-GT, the 6W4 damper, the horizontal deflection yoke and the width coil. Filament power is supplied to the 1B3-GT tube by a single-turn loop placed close to the transformer winding.

In this autotransformer, the horizontal deflection winding is connected across terminals 1 and 4, these being the terminals across which the horizontal deflection winding of the yoke best matches the autotransformer. The 6W4 damper tube connects between points 1 and 5. Note that with

the removal of the secondary windings, the damper tube is now inverted. That is, the cathode of the 6W4 connects to a higher joint on the autotransformer winding than the plate. The energy which the damper tube is able to rectify and convert into d-c is developed across the .039- μ f capacitor con-

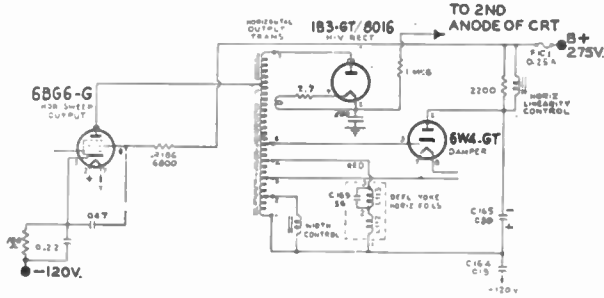


FIG. 13.7. An autotransformer type of horizontal output transformer.

nected from the plate of the 6W4 tube to terminal 1 of the output transformer. The polarity of this rectified voltage developed here is as shown in Fig. 13-7. To add this voltage to that of the low-voltage power supply, the B+ terminal of the low-voltage power supply connects to the negative end of the .039- μ f capacitor through a horizontal linearity coil. The other (or positive) terminal of the .039- μ f capacitor connects to the bottom of the

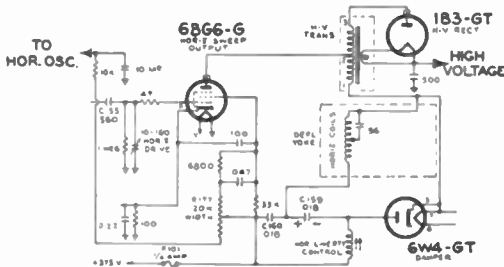


FIG. 13.8. In this horizontal output system, the deflection yoke coils are connected directly into the plate circuit of the output amplifier.

autotransformer, applying the boosted B+ to the plate of the 68G6G output amplifier.

Another development in horizontal output systems is the direct-drive system shown in Fig. 13-8. Here, the inductance of the deflection yoke is connected directly into the plate circuit of the horizontal output amplifier in place of a section of the previous autotransformer. The 6W4 damper tube is connected directly across the horizontal deflection windings and the sharp pulse which appears here during beam retrace is rectified and added in

series with the receiver B+. In Fig. 13-8, this additional d-c voltage appears across C_{159} (.018 μ f), with the polarity as shown. Note that the low-voltage B+ (375 volts here) is placed in series with the voltage across C_{159} through the horizontal linearity coil.

Picture width is controlled by potentiometer, R_{177} . This potentiometer is essentially in series with the B+ at one end and the plate circuit of the horizontal oscillator at the other end. By regulating the amount of voltage made available to the horizontal oscillator, we can control the amplitude of

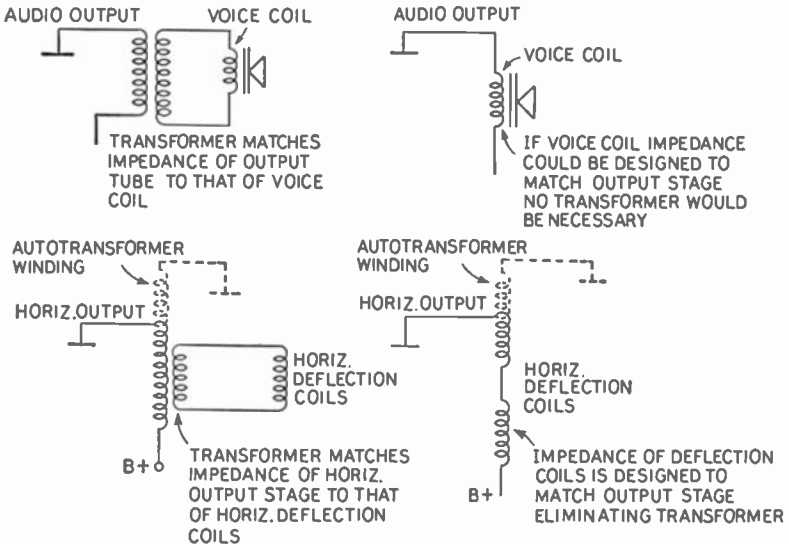


FIG. 13.9. A comparison of the coupling methods employed to connect loudspeakers to audio output stages with the coupling methods shown in Figs. 13.7 and 13.8.

the saw-tooth deflection voltage which reaches the 6BG6G output amplifier and thereby the width of the picture.

Fig. 13-9 compares the coupling methods employed to connect loudspeakers to audio output stages with those methods used to couple the deflection yoke to the horizontal output tube.

The development of larger screens tended to magnify whatever defects are present in an image. For example, the poor vertical interlacing of scanning lines may pass unnoticed on a 10-inch screen, but on a 12-inch screen it could cause a very noticeable deterioration in picture quality. By the same token, lower than normal focus at the edges of a picture may not be serious on a small screen whereas on a large screen it would stand out quite prominently.

Improved focus at the edges of a picture has been achieved by development of a cosine deflection yoke. In this yoke the horizontal deflection

windings are distributed around the neck of the tube in such a manner that the magnetic field is uniform. In the ordinary yoke, the field is non-uniform, resulting in distortion of the electron beam shape with subsequent defocusing.

Automatic Frequency Control Systems. Use of the incoming sync pulses to trigger and control the vertical and horizontal sweep oscillators represents the simplest, most economical, and most direct method of controlling the motion of the electron beam in the image tube. Unfortunately, however, this method possesses limitations and disadvantages which outweigh its economy and simplicity. Perhaps the greatest disadvantage is its susceptibility to noise disturbances arising from electrical apparatus and equipment operating in the vicinity of the receiver. The noise pulses, combining with the video signal and extending usually in the same direction as the desired sync pulses, pass through the same stages as the pulses and arrive at the sweep oscillators. They do their greatest damage when they arrive during the interval between sync pulses. If the amplitude of the noise pulses is sufficiently great, they will trigger the sweep oscillator, initiating a new cycle prior to its proper time. When the vertical oscillator is so triggered, the picture will move vertically either up or down, until the proper sync pulses in the signal can again assume control. If the horizontal oscillator is incorrectly triggered, a series of lines in a narrow band will be jumbled, giving the appearance of streaking or tearing across the image. When the interference is particularly heavy and persistent, the entire picture becomes jumbled and may even be thrown out of horizontal sync permanently, requiring manual adjustment of the horizontal hold control.

Of the two sweep systems in a television receiver, interference is particularly destructive to the horizontal system. To understand why this is so, we must examine the nature of most interference voltages and their effect upon the vertical and horizontal sweep oscillators.

Whenever a blocking oscillator is triggered, for example by a sync pulse, its grid, after a short period of conduction, becomes highly negative as a result of an accumulation of electrons on the grid capacitor. This negative voltage is sufficient to keep the tube beyond cut-off until the charge on the grid capacitor has decreased to a value at which current is permitted once again to flow through the tube. In most circuits now in use, the capacitor discharge occurs in the manner shown in Fig. 13-10. At the start, the discharge is fairly linear. However, as the amount of charge contained in the capacitor decreases, the discharging rate decreases in an exponential manner. In Fig. 13-10, the region, usually called non-linear, extends from points A to B.

Now, when the negative charge existing on the grid condenser is large, the oscillator is relatively immune to incoming positive pulses. With con-

tinued discharge, however, the immunity decreases. Experience has indicated that off-cycle triggering of the oscillator is generally concentrated in the last 15 per cent of its discharge cycle. This is true regardless of the frequency at which the oscillator is operating. Hence, one would expect to experience equal difficulty with both deflecting systems in the receiver. That this is not so is due to the nature of the noise pulses and the type of filters inserted before each sweep oscillator.

Noise pulses which are most troublesome to television receivers possess high amplitude, but are narrow in width or, what is the same thing, are of short duration. (The energy in noise pulses is distributed over a wide

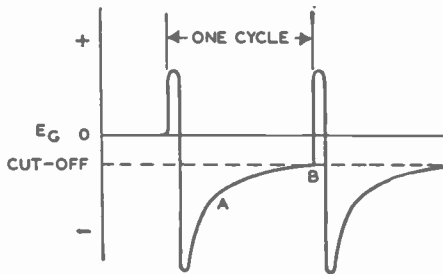


FIG. 13.10. The manner in which the grid voltage of a blocking or multivibrator oscillator varies.

range of frequencies. In order for a peak to occur, the phase relationship among the various frequencies must be such as to permit them to add, forming the high-amplitude pulse or peak. This condition, however, usually exists only for a brief interval, which explains the narrow width of these pulses.)

When the pulses reach the path leading to the horizontal sweep oscillator, they are readily passed because of the short-time constant of the filter leading to the horizontal system. A short-time-constant filter is necessary because the horizontal sync pulses themselves have a time duration of only 5 microseconds. On the other hand, the filters leading to the vertical system have a long-time constant and automatically act to suppress the effects of all horizontal sync pulses and noise pulses of short duration. The presence of this low-pass filter (called an integrating network) is largely responsible for the greater immunity to noise pulses enjoyed by the vertical system. Of course, when a wide noise pulse is received, it contains enough energy to cause offtime firing of the vertical oscillator, but the annoyance caused the viewer from this source is seldom great. To reduce the susceptibility of the horizontal sweep system to noise pulses of any type, several automatic frequency (and phase) control systems have recently been developed.

In each of these special control systems an oscillator is set to operate at 15,750 cycles, and the output of the oscillator controls the horizontal motion of the electron beam across the screen of the image tube. The next step is to synchronize the frequency of this sweep oscillator with the incoming horizontal sync pulses of the signal. This is accomplished through an intermediate stage known by one of the following names: Control tube, AFC

phase detector, or horizontal sync discriminator. Whatever the name, what this intermediate network does is to compare the frequency of the incoming horizontal sync pulses with the frequency of the receiver horizontal sweep oscillator. If a difference does exist, then a d-c voltage is developed which, when fed back to the horizontal sweep oscillator, is used to change its frequency until this is exactly equal to that of the incoming sync pulses.

Note that the incoming sync pulses are not applied directly to the sweep oscillator. They are merely compared (in frequency) with the output of the sweep oscillator and, if a frequency difference exists, then a d-c voltage is developed which, when fed back to the sweep oscillator, forces its frequency back into line with that of the sync pulses.

Now (and here is the crux of the whole matter), by having the d-c control voltage pass through a long-time-constant filter before it reaches the sweep oscillator, we can eliminate the effects of any noise impulses and permit only relatively slow changes in frequency of the sync pulses (which may occur at the transmitter) to affect the sweep oscillator. We have thus incorporated into the horizontal sweep system a long-time-constant filter somewhat similar to that present in the vertical sweep system.

Remember, it was the long-time-constant filter in the vertical sweep system that gave this network its immunity from most noise pulses (which have a short-time duration). A long-time-constant filter could not be used directly in the horizontal system because it would have prevented the desired horizontal sync pulses (together with the noise pulses) from reaching the horizontal sweep oscillator. Hence the need for the indirect method outlined above.

Now let us analyze the various automatic frequency control systems currently in use.

Horizontal A.F.C. Systems. One of the first A.F.C. systems to be used is shown in block form in Fig. 13-11. In this circuit, a stable Hartley oscillator is set at 15,750 cycles per second. Connected to the oscillator is a synchronizing discriminator circuit which receives the pulses of the incoming television signal and compares the frequency and phase of these synchronizing pulses with the generated sine wave of the Hartley oscillator. Any slight variations between the two will produce a d-c voltage which is applied to the grid of a reactance tube after passing through a long-time-constant filter. The plate circuit of this reactance tube is connected across the tuning circuit of the Hartley oscillator, and, as the d-c voltage at the grid of the reactance tube varies, it will alter the plate current of this tube and with it the oscillator frequency.

Since the reactance tube plays such a vital role in this and other A.F.C. circuits, let us pause here and determine how a tube can simulate the action of a resistance, a capacitance, or an inductance.

In Fig. 13-12A, a tube is connected with normal plate and grid voltages, so that plate current is flowing through the tube. When we increase the plate voltage, the plate current will increase and the two will be in step within fairly wide limits of plate current flow. The same is true of a plate voltage decrease; the current will decrease in step. Since the two follow

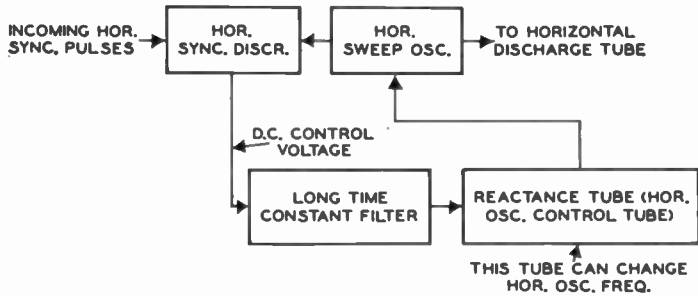


Fig. 13.11. A block diagram of the A.F.C. system shown schematically in Fig. 13.13.

each other directly, we obtain essentially the same action as the voltage and current in a resistor, and consequently the tube is functioning as a resistor, although, at some voltages, it acts as a non-linear resistor.

Now, let us alter the circuit to the form shown in Fig. 13-12B. A source of R.F. voltage is connected to the plate of the triode and the "B+" voltage

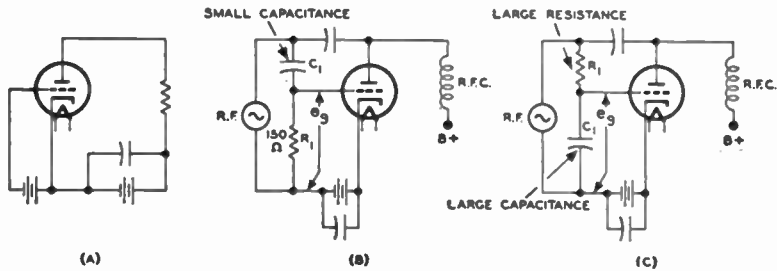


Fig. 13.12. Circuit diagrams show how a vacuum tube can be made to appear as a resistor (A), a capacitor (B), or an inductor (C).

now reaches the tube through an R.F. choke. The choke prevents the R.F. component of the plate current from passing through the d-c voltage source. In the same circuit, a small capacitor and resistor transfer a portion of the R.F. voltage to the grid of the tube. Let us investigate this section of the circuit more closely.

Since C_1 and R_1 are placed directly across the R.F. voltage source, an R.F. current will flow through both components, the amount of current being governed by the impedances of C_1 and R_1 . In the diagram it is specified

that C_1 is small, resulting in a large impedance. The resistor, on the other hand, is low in value. Thus, the circuit impedance will be largely capacitive, and the current flowing through C_1 and R_1 will lead the R.F. voltage by approximately 90° . The R.F. current, flowing through R_1 , will develop a voltage, e_p , which is in phase with the R.F. current and leading the R.F. voltage from the generator by 90° . It is further true that since e_p is also the alternating grid voltage for the tube, the plate current will lead the applied R.F. voltage by the same 90° . Whenever any electrical component exhibits the property of having the current through it lead the voltage applied across it by 90° , it is said to act like a capacitance. Thus, by properly connecting the tube, we have made it appear to the circuit as a capacitor. The tube will have this effect on any circuit in which it is placed.

To indicate how a tube can be made to function as an inductance, we can employ the circuit shown in Fig. 13-12C. Now we find that R_1 and C_1 have been interchanged and the impedance of R_1 greatly exceeds the impedance of C_1 , at the operating frequency of the R.F. voltage generator. The current now through R_1 and C_1 will be in phase with the applied R.F. voltage. However, across any capacitor, the voltage always lags 90° behind the current and, therefore, the R.F. grid voltage, in Fig. 13-12C, will lag the current through R_1 and C_1 by 90° . By the same token, it will lag the R.F. generator voltage by 90° . Within the tube, the plate current, being in phase with the grid voltage, will also lag the R.F. voltage by 90° . To the circuit, the tube will appear as an inductance in which the current lags 90° behind the applied R.F. voltage. (If desired, a high resistance can be shunted across C_1 to provide a leakage path for any electrons accumulated at the grid. With a high value of resistance, the circuit operation remains unchanged.)

In each of the foregoing instances, the amount of inductance and capacitance that the tube injects into the circuit will depend upon the amount of plate current flowing through the tube. If we increase the d-c bias for the tube in the positive direction, the plate current will increase and its reactive effect on the circuit will likewise increase. With a greater negative d-c bias, the opposite action and effect will occur.

With the foregoing in mind, let us analyze the circuit of which Fig. 13-11 is the block diagram. This circuit is shown in Fig. 13-13 and consists of the following.

1. A horizontal sync discriminator circuit using V_{1A} and V_{1B} .
2. A horizontal sweep oscillator (V_2).
3. A reactance or horizontal oscillator control tube (V_3).
4. A long-time-constant network using C_3 , C_4 , and R_4 .

The sweep oscillator (V_2) is the well-known Hartley. The oscillator coil T_2 is inductively coupled to the horizontal sync discriminator by means

of a center-tapped primary winding, T_1 . Each diode in the discriminator circuit is connected across half of T_1 , receiving sine-wave voltages which are equal in amplitude but opposite in phase. Disregarding R_1 for the moment, the voltages appearing across T_1 will affect this circuit in the following manner:

During one half cycle, the top end of T_1 is positive, forcing V_{1A} to conduct. Its current flows from cathode to plate, through the top half of the coil, through R_1 and R_2 back to the cathode again. The voltage developed across R_2 is such that the bottom end is negative and the cathode end is positive.

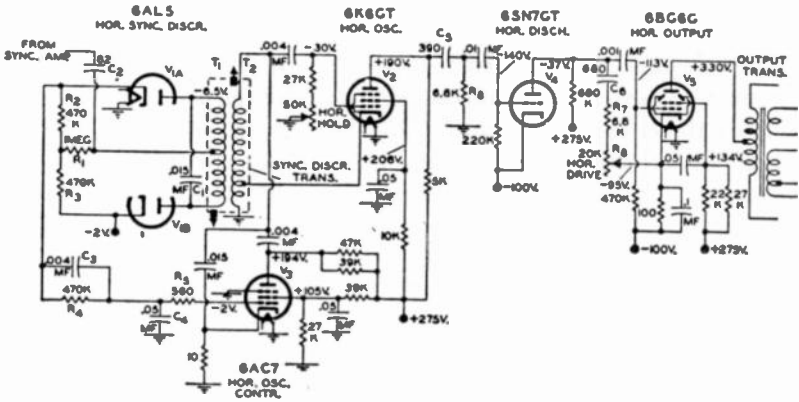


FIG. 13.13. An automatic frequency control system used in television receivers.

Throughout this interval, V_{1B} is non-conductive because the sine-wave voltage applied to its plate is negative.

During the next half cycle, V_{1B} conducts and V_{1A} is cut off. Current now flows through this tube, up through the bottom half of the primary winding, T_1 , through R_1 and R_3 back to V_{1B} again. The voltage developed across R_3 is such that its top end is negative. If the circuit is operating properly, the voltages across R_2 and R_3 will be equal. However, when these two voltages are added, the resultant will be zero because of the back-to-back placement of the two resistors. In this circuit the voltages are added, which means that no output voltage will be obtained, using this network alone. However, inspection of the circuit reveals that the horizontal sync pulses are also added to this circuit by means of C_2 and R_1 . C_2 connects to the output of the sync separator and transfers the horizontal pulses appearing here to R_1 where both V_{1A} and V_{1B} receive this pulse voltage in equal amplitude and phase. Now, let us see how this alters circuit operation.

A horizontal sync pulse can arrive at one of three times. It can appear across R_1 when the sine-wave voltage at T_1 is zero; it can arrive when

V_{1A} is conducting; or it can arrive when V_{1B} is conducting. If the pulse arrives when the sine wave applied to each tube is passing through zero, there will be no change in circuit conditions. The incoming synchronizing pulse appears across R_1 and combines with the sine-wave voltage from each half of the secondary coil T_1 to form the driving voltage for each diode. If the pulses arrive when the sine-wave voltage across T_1 is zero (Fig. 13-14A), then each diode will receive the same pulse voltage, and the same amount of rectified voltage will appear across the diode load resistors R_2 and R_3 . The total net output from both tubes will be canceled because their load resistors are connected in opposition (back-to-back).

Note again that the reason each diode produces equal voltages across R_2 and R_3 at this moment is that there is no sine-wave voltage and each tube receives the same synchronizing pulse voltage from R_1 .

Suppose, however, that the pulse arrives at some other instant. Two such situations are shown in Fig. 13-14B and C. In Fig. 13-14B, the pulse arrives when the top diode is positive (and conducting) and the bottom diode is cut off. Obviously, then, there will be, on the average, more voltage developed across R_2 than across R_3 . The average voltage, over one cycle, will be positive, and this, fed to V_3 , will make its grid more positive. On the other hand, if we consider the situation of Fig. 13-14C, we see that R_3 will receive the greater voltage and that, on the average, the voltage from the combination will be negative. The effect on V_3 will be different in each instance. It is seen, therefore, that the double-diode arrangement of V_1 is a very sensitive phase discriminator and will develop an output voltage which may be negative, zero, or positive, depending upon the phase of the pulses with respect to the voltage generated by the Hartley oscillator.

The d-c voltage developed in the phase discriminator, together with a fixed negative biasing voltage (-2 volts), is applied through the long-time-constant filter of C_3 , R_4 and C_4 to the grid of a reactance tube, V_3 . This tube is so connected that its plate current is 90 degrees out of phase with the voltage across it, and consequently it appears as a reactance. (Note that a .015- μ f capacitor is connected from the plate circuit of V_3 to the 10-ohm cathode resistor. Here, the reactance of the capacitor far exceeds the resistance of the 10-ohm resistor.) Since V_3 is connected directly across the

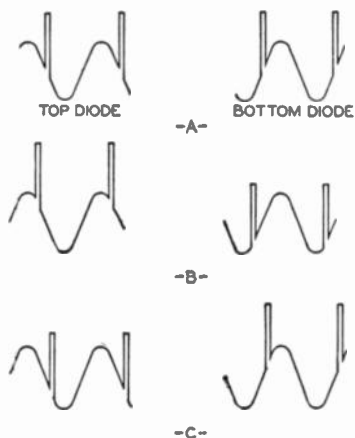
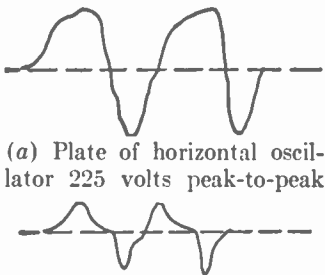


FIG. 13.14. Illustrating how the horizontal synchronizing discriminator operates.

tuning coil of the Hartley oscillator, changes in its plate current will produce changes in the frequency of the oscillator and force the oscillator to operate at the frequency of the synchronizing pulses. The d-c voltage developed at the phase discriminator is fed directly into the grid of the reactance tube and thereby controls the plate current flow. In this manner, all differences in frequency between the synchronizing pulses and the Hartley oscillator are instantly corrected.

C_3 , R_4 , and C_4 form a long-time-constant filter to block the passage of rapid changes in d-c from the phase discriminator produced by noise pulses.



(a) Plate of horizontal oscillator 225 volts peak-to-peak

(b) Input to horizontal discharge tube 100 volts peak-to-peak

FIG. 13.15. Waveforms

contains only a resistance, and consequently the waveform found here is not the symmetrical sine wave that would be produced by a resonant circuit; rather it has the form shown in Fig. 13-15A. The peak-to-peak voltage on the grid of V_2 is approximately 130 volts. This grid swing produces a square wave in the plate circuit, with a peak-to-peak voltage of 225 volts. The square wave is differentiated by C_5 and R_6 , resulting in the waveform shown in Fig. 13-15B. The positive portion of the differentiated waveform is sufficiently sharp to trigger the discharge tube. The discharge tube is normally at cut-off due to bias produced by grid rectification of these pulses from the oscillator. The positive pip of the pulse overcomes this bias and drives the tube into heavy momentary conduction. During this period, the plate voltage of V_4 falls to cathode potential, and C_6 does not discharge completely because of R_7 and R_8 , both of which are in series with C_6 . When V_4 becomes non-conductive again, its plate voltage rises quickly to a value determined by the charge remaining on C_6 . From this point the plate voltage rises slowly, and we get the desired deflection waveform. This is transferred to the horizontal output tube V_5 .

Saw-tooth A.F.C. System. Another approach to automatic frequency control of the horizontal sweep oscillator is the system shown in Fig. 13-16.

Only the relatively slow changes in frequency caused by differences between the Hartley oscillator and the incoming horizontal sync pulses are permitted to reach V_3 . A hold control is available in case the Hartley oscillator frequency should be too far from the frequency of the incoming pulses.

Horizontal Discharge. A horizontal discharge tube, V_4 , follows the horizontal oscillator and produces, in its plate circuit, the peaked deflecting voltage which will drive the horizontal deflecting coils. In V_2 , the Hartley oscillator is connected between the screen grid and the cathode. The plate circuit contains

V_{2A} , V_{2B} , and V_3 correct the frequency of the horizontal multivibrator, V_4 , whenever it drifts away from the frequency of the incoming sync pulses. V_{2A} and V_{2B} receive voltages from the horizontal sweep system and from the incoming pulses. The two voltages are compared as to frequency, and any difference produces either a positive or negative voltage at the grid of V_3 . These changes are amplified and then transferred to the multivibrator, whose frequency is changed accordingly.

In detail, the automatic horizontal frequency control network functions as follows: The incoming horizontal sync pulses are transferred by means of T_1 to the two diodes,* V_{2A} and V_{2B} , with the polarity as shown in Fig. 13-16. (Although V_{2A} and V_{2B} are triodes, they are connected to function as diodes.) The top end of the secondary of T_1 develops a positive pulse voltage, and the bottom end, a negative pulse. The positive pulse causes V_{2A} to conduct, and the negative pulse on the cathode of V_{2B} causes this tube to conduct also. The current flowing through V_{2A} charges capacitor C_3 to approximately the peak value of the applied pulse while the current flowing through V_{2B} charges C_4 . The polarity of each voltage is indicated in Fig. 13-16. During the interval between pulses, each capacitor discharges, the electrons moving from C_3 down through R_1 and R_3 to C_4 and from the other plate of C_4 through the secondary transformer winding back to C_3 . The discharge is slow and the voltages developed across R_1 and R_3 prevent V_{2A} and V_{2B} , respectively, from conducting until the arrival of the next pulse.

Feeding into this circuit is another voltage. This is a saw-tooth voltage which develops across C_5 from pulses which are applied to it from the secondary of the horizontal output transformer. This saw-tooth voltage possesses the same frequency as the horizontal sweep oscillator since it is this stage which drives the horizontal output amplifier. The saw-tooth voltage is applied in equal measure to each tube; this means that the plate of V_{2A} and the cathode of V_{2B} receive the same polarity voltage (of the saw-tooth wave) at the same time. Thus, at the discriminator, we have both the ingredients needed to check the operating frequency of the horizontal oscillator against the frequency of the incoming pulses.

The comparison of the two frequencies occurs only at the instant the sync pulses arrive, for it is only at this moment that V_{2A} and V_{2B} conduct and therefore are in a position to respond to the saw-tooth voltage. As in the previous A.F.C. circuit, three situations are possible.

First, if the sync pulses arrive at a time when the saw-tooth wave is passing through zero, then current will flow through V_{2A} and V_{2B} , recharging C_3 and C_4 for any voltage that they may have lost during the interval between pulses. This flow of current will remain within the branch of the

* Another name for this stage is *phase detector*.

circuit formed by the two tubes and the secondary of T_1 . No voltage will appear between point A and ground to affect the d-c amplifier, V_3 , and by this, the horizontal multivibrator. Actually this represents the desired condition since it indicates that the frequency of the sweep oscillator and the sync pulses are in step with each other.

The second situation occurs when the sync pulses arrive and the saw-tooth voltage is positive at this moment. Under this condition V_{2A} will receive a positive pulse and a positive saw-tooth voltage. V_{2A} will conduct more strongly than usual, with current not only charging C_3 , but also flowing through R_1 to ground and from ground up through C_1 , R_2 , and C_2 to the cathode again. This flow of current through R_2 , C_1 , and C_2 will establish a voltage here which is positive with respect to ground.

At the same moment that V_{2A} is conducting more strongly, conduction through V_{2B} is reduced because the positive saw-tooth voltage partially offsets the negative sync pulse applied to the cathode of V_{2B} . Since the current flow through V_{2B} is reduced instead of increased as it is through V_{2A} , it cannot offset the positive voltage at point A. This potential, applied to the grid of V_3 , increases the current through V_3 , driving the plate of this tube more negative. Since the grid of the horizontal multivibrator is connected to the plate of V_3 , it, too, becomes more negative, thereby altering its frequency. In this instance, the change is toward a lower frequency.

The third situation occurs when the pulses arrive and the saw-tooth voltage is negative. Now, V_{2B} conducts more strongly than V_{2A} , its current flowing not only into C_4 , but also through R_2 , C_1 , and C_2 to ground and from here to R_3 and back to V_{2B} . The voltage developed this time at point A is negative with respect to ground which reduces the current flow through V_3 and acts to speed up or raise the frequency of the horizontal multivibrator.

Filter R_2 , C_1 , and C_2 responds only to slow changes, thereby preventing fast-acting noise pulses from affecting the operation of the multivibrator. V_5 is the horizontal output amplifier. It receives the peaked deflection voltage, amplifies it, and uses this voltage to drive a high-voltage rectifier (8016) and the horizontal deflection coils. The tuned circuit in the cathode leg of the multivibrator is resonant to 15,750 cycles and serves to stabilize the frequency of this unit.

D-c Control of Oscillator Frequency. In the previous A.F.C. system of Fig. 13-13, the d-c control voltage was applied to a reactance tube and this, in turn, varied the sweep oscillator frequency. In the present A.F.C. system (Fig. 13-16), the d-c control voltage developed by the sync discriminator is used directly to alter the frequency of the horizontal sweep oscillator. This direct method of frequency control is readily adapted to multivibrators and blocking oscillators.

To understand what happens when the d-c control voltage is applied directly to an oscillator, consider the operation of a blocking oscillator.

It was noted in Chapter 12 that the length of time a blocking oscillator was cut off was determined essentially by the time constant of the grid circuit. When the grid resistor and capacitor values are high, the charge accumulated across the grid capacitor diminishes slowly and the tube is kept cut off for a longer period of time. When these values are low, the cut-off interval of the tube is shortened accordingly.

If now, we raise the negative bias on the grid (by adding a negative voltage from some external source), we can see that to reach the cut-off bias value of the tube, using the same grid resistors and capacitor, will require a longer time than if no such negative voltage had been added to the circuit. And if the tube is kept cut off for a longer length of time, then obviously the frequency generated by this oscillator will be lower.

On the other hand, injecting a small positive voltage on the grid of the oscillator will decrease the total negative voltage developed here by the circuit operation and permit the tube to return to conduction sooner. The result: Generation of a higher frequency. It is precisely in this manner that the d-c control voltage developed by the above sync discriminator circuit varies the frequency of the sweep oscillator. While only the action of a blocking oscillator has been analyzed, the reasoning for a multivibrator is similar.

The circuit in Fig. 13-17 is an automatic frequency control network which operates in a manner very similar to that of the preceding saw-tooth system. The incoming sync pulses are fed, in equal amplitude but opposite polarity, to two diodes, V_2 and V_3 . The plate of V_2 receives a positive horizontal sync pulse at the same time that the cathode of V_3 is receiving a negative sync pulse from V_1 . Because of the polarity of these pulses, both diodes will conduct at this instant, with the coupling capacitors charging up to essentially the peak pulse value. During the interval between pulses, the two diodes will be kept cut-off.

In addition to the sync pulses, V_2 and V_3 also receive a saw-tooth voltage from the horizontal sweep amplifier. As in the previous system, the saw-tooth wave will be going through zero at the time the sync pulses reach V_2 and V_3 if the frequency of the horizontal sweep oscillator is properly synchronized to the incoming pulses. If a frequency difference exists, the saw-tooth voltage will not be going through zero when the sync pulses arrive. If the saw-tooth voltage has some negative value at this instant, V_3 will conduct more strongly than V_2 , and a negative resultant voltage will appear across C_1 and R_3 . (A negative saw-tooth voltage favors V_3 because this voltage is applied to the cathode of the tube.) The negative voltage across C_1 is then passed on to the following sweep oscillator, changing its point of operation and consequently its frequency.

By the same token, arrival of the sync pulses when the saw-tooth voltage is positive will cause V_2 to conduct more strongly than V_3 , producing a resultant negative voltage across C_1 . (A positive saw-tooth voltage favors V_2 because it is being applied to the plate of this tube.) The effect of the sweep oscillator of the positive voltage across C_1 will be opposite to that produced by a negative voltage.

C_1 and R_3 form a fairly long-time-constant filter, permitting only the voltage variations due to differences between the frequency of the sync pulses and the saw-tooth voltage to develop here. Momentary voltage variations due to noise pulses are effectively suppressed.

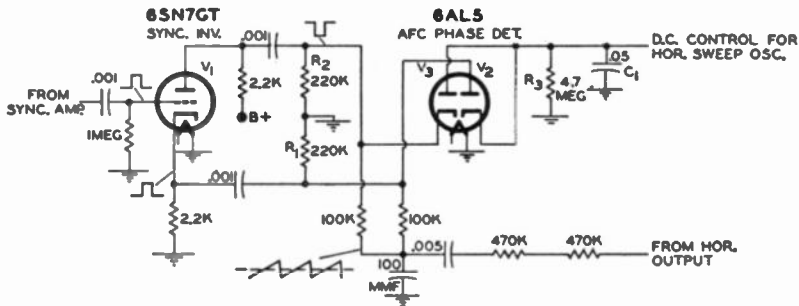


FIG. 13.17. An A.F.C. system closely similar to the one shown in Fig. 13.16.

The d-c control voltage obtained from C_1 could be applied directly to the horizontal sweep oscillator and, in some receivers, it is. On the other hand, a more sensitive arrangement is obtained when the d-c control voltage is amplified before being applied to the horizontal oscillator.

Additional Applications of Saw-tooth System. The d-c control voltage obtained from the phase detector in Fig. 13-17 can be employed to regulate the frequency of almost any type of oscillator which periodically biases itself to cut-off. One combination which has enjoyed considerable popularity is shown in Fig. 13-18. Here the d-c control voltage from the 6AL5 phase detector is applied through a long-time-constant filter (R_{428} , R_{429} , C_{415} , C_{416}) to the grid of one triode section of a multivibrator.

If the multivibrator is examined closely, it will be seen to contain a parallel resonant circuit in the plate lead of the first triode section. This combination is tuned to 15,750 cycles, and by its presence serves to stabilize the operation of the multivibrator. How it does this will be discussed presently.

Pulse-width A.F.C. System.* Another A.F.C. circuit which has gained wide acceptance is that shown in Fig. 13-19. It consists of a single control tube, a long-time-constant filter and a blocking oscillator. The

* Also known as the "synchroguide A.F.C. system."

6BG6G horizontal output tube is added to the diagram because part of its output voltage is fed back to the control tube. Basically, the horizontal oscillator is a free-running oscillator and discharge circuit. It does not receive the incoming pulses directly, but, should its frequency differ from that of the pulses, then the control tube V_1 will alter the negative bias on the grid of the blocking oscillator and thereby change its frequency. It can do this because the cathode resistor R_1 of the control tube is common to the grid of the blocking oscillator. The incoming sync pulses, positive in polarity, are applied to the grid of V_1 through a $120\text{-}\mu\mu\text{f}$ capacitor. This same grid also receives a pulse from the horizontal output transformer and a small portion of the voltage from the output of the blocking oscillator. The pulse from the output transformer and the voltage from the oscillator form a combined wave, which then reacts with the incoming sync pulses to maintain the blocking oscillator on frequency. (Fig. 13-19A illustrates the shape of the various voltages reaching V_1 . The wave obtained from the output of V_2 is originally a saw-tooth voltage, but it is converted into a parabolic wave by the $120,000\text{-ohm}$ resistor and C_4 before reaching V_1 .)

The combined wave is specifically designed to have fairly steep sides * (see Fig. 13-20) in order that any difference in frequency between the blocking oscillator and the incoming sync pulses will have a marked effect on the circuit. Here is how this occurs.

A portion of the bias from the blocking oscillator is applied to the grid of the control tube and is sufficient to keep the control tube cut-off except when the incoming sync pulse is high on the slope of the grid waveform (the combined wave mentioned previously) as shown in Fig. 13-20A. If the blocking oscillator changes phase so that the pulse arrives at a time when it is down the slope, the amount of time that V_1 conducts will decrease. This is indicated in Fig. 13-20B by the narrow width of the waveform extending above the cut-off point of tube V_1 .

On the other hand, if the blocking oscillator frequency changes so that the sync pulse arrives at a time when it is closer to the top of the combined wave (Fig. 13-20C), then the plate conduction time of V_1 will increase. When the control tube conducts, C_1 and C_2 in its cathode circuit will charge to a d-c potential proportional to the length of time that current flows through

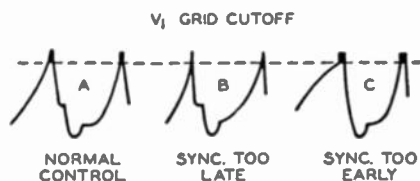


FIG. 13.20. Horizontal control waveforms. Shaded area above dotted line is portion of waveform effective in producing oscillator control voltage across V_1 .

* In some designs of this circuit, the voltage fed back from the output transformer is omitted. This changes the form of the control wave somewhat but does not alter basic circuit operation.

the tube. This d-c potential is applied as a bias to the grid of the blocking oscillator, thereby altering its frequency and tending to bring it back into line. The components of the cathode circuit of V_1 form a fairly long-time-constant filter which averages the plate current pulses.

Stabilizing Circuit. Examination of the blocking oscillator, Fig. 13-19, reveals a parallel resonant circuit (L and C) in the plate lead of V_2 . Its purpose is to increase the stability of the oscillator and to give it greater noise immunity. How it does this can be seen from the following explanation.

L and C are placed in series with the $B+$ lead going to the plate of the horizontal oscillator. The oscillator, being a blocking oscillator, is cut off for more than 90 per cent of the cycle and conducts heavily for 10 per cent of the time or less. Let us consider the oscillator when it is conducting heavily. During this interval, the heavy current drawn through L and C prevents this circuit from oscillating. At the same time a voltage is developed across these components with the polarity as indicated in Fig. 13-21A.

When the blocking oscillator drops into cut-off, the current flowing through the tube and the resonant circuit — L,C — ceases abruptly. This sudden stoppage of current excites L,C into resonance, the coil attempting to maintain the flow of current. The polarity of the voltage developed across the coil required to do this is shown in Fig. 13-21B and is seen to be exactly opposite to its previous polarity in Fig. 13-21A.

Referring to the waveform developed in this circuit at point Y , at time A in Fig. 13-21C the blocking oscillator drops into cut-off, and at time B the resonant circuit reverses its voltage polarity in an attempt to maintain the flow of current.

Thereafter the current in the resonant circuit L,C flows first into the lower plate of C , then reverses and flows into the other plate. The frequency of this circuit is close to 15,750 cycles, and so it has time to complete one cycle before the blocking oscillator again conducts heavily, at time E , damping out the oscillations in L and C .

Now, the ability of this additional resonant circuit to improve the noise immunity of the blocking oscillator is due to the fact that near the end of the discharge cycle, when the cut-off voltage on the oscillator grid is low, the voltage developed across L,C is negative. In Fig. 13-21C, this is the region CDE. This negative voltage opposes the $B+$ voltage from the power supply and, in so doing, reduces the susceptibility of the blocking oscillator to any noise pulses that may appear at this time.

Here is why this is so. When the plate voltage of a triode decreases, the negative voltage required to cut off the plate current decreases. In other words, the cut-off point approaches closer to zero volts. With the $B+$ plate voltage going down (because of the opposition voltage across L and C), we achieve the same effect as bringing the cut-off level of the tube closer to zero.

See Fig. 13-22. With the cut-off level thus moving upward, it will require a stronger noise pulse to raise the grid voltage up to this new cut-off level in order to have current flow through the tube again for the start of the next cycle.

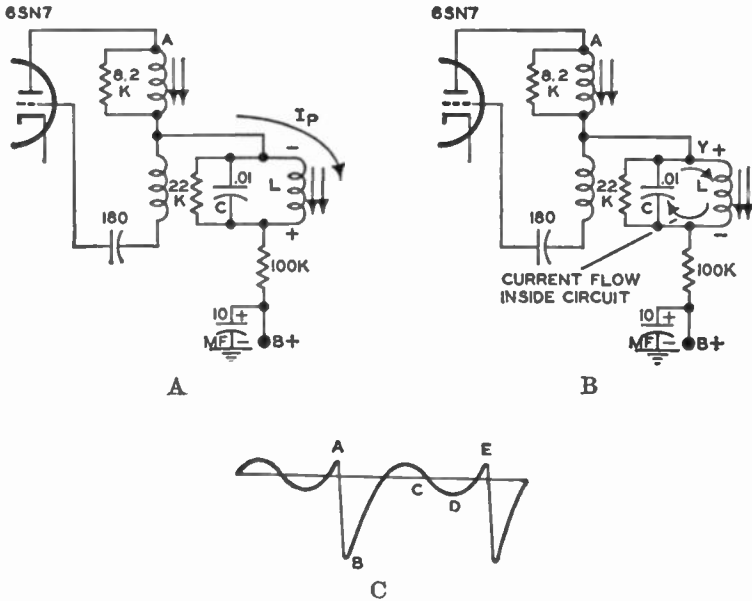


FIG. 13.21. Diagram illustrating how L and C (of Fig. 13.19) help stabilize the pulse-width A.F.C. system.

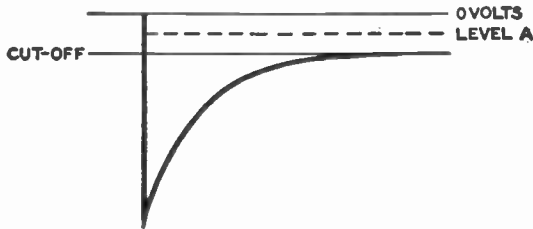


FIG. 13.22. See text for explanation.

On the other hand, in the region near E of Fig. 13-21C, the voltage across L and C is going in the positive direction, aiding the $B+$ voltage and preparing the plate circuit for the triggering pulse which is about to occur.

Resonant stabilizing circuits have been used in the multivibrator as well as in the blocking oscillator. One such circuit was shown in Fig. 13-18. The additional resonant coil and capacitor are placed in the plate circuit

of V_1 and adjusted to a frequency of 15,750 cycles. The presence of this circuit alters the manner in which the grid of the second triode comes out of cut-off. In Fig. 13-23A we have the waveform appearing at the plate of the first triode in the absence of the stabilizing circuit. Fig. 13-23B shows the grid waveform of the second triode under the same condition. Now, when we insert the stabilizing circuit, its waveform (Fig. 13-23C) will add to those existing in the circuit to produce the modified waveforms shown in Fig. 13-23,

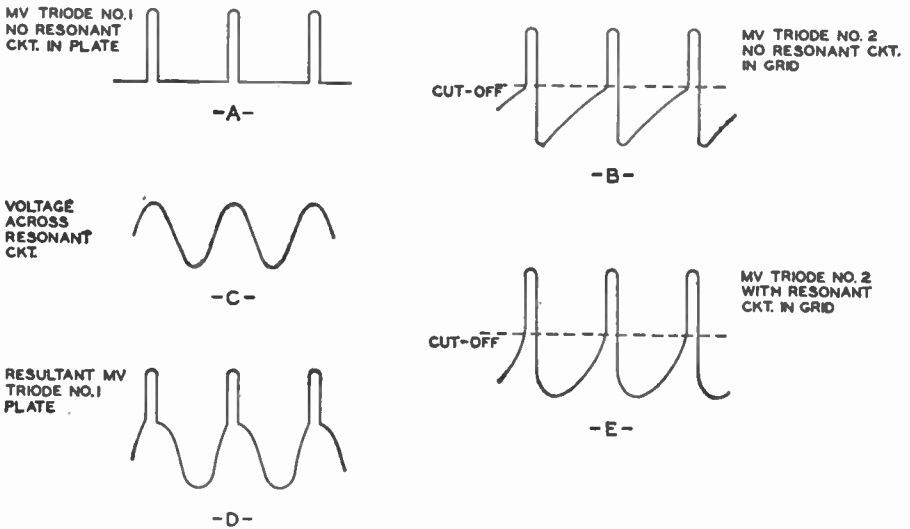


Fig. 13.23. The effect of a resonant stabilizing circuit on the operation of the multivibrator shown in Fig. 13.18. *After DuMont Service News*

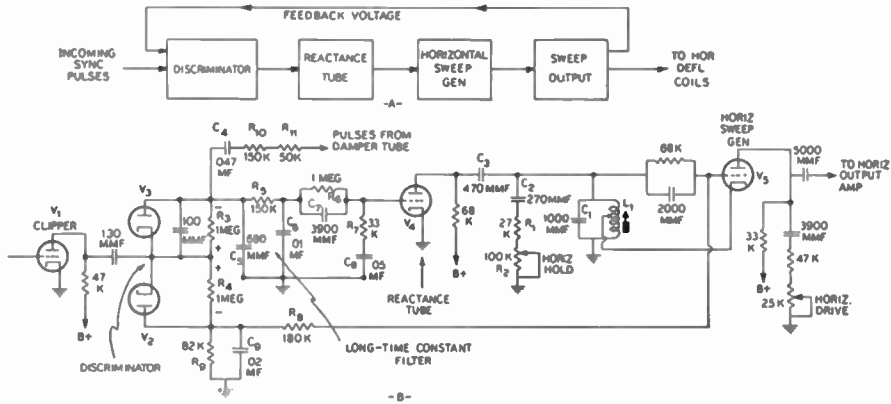
D and E. Of particular importance is the grid waveform of the second triode. Note that it now comes out of cut-off quite sharply, and it will require a considerably stronger noise pulse to trigger this tube prematurely than it would have without the stabilizing circuit.

The controls in the pulse-width system are as follows: The blocking oscillator transformer, T_1 , is slug-tuned to permit coarse adjustments in oscillator frequency. Stabilizing coil L is also slug-tuned, and it affects not only the stability of the blocking oscillator but also its frequency. The horizontal hold control will alter the plate voltage of the control tube and, in this manner, the amount of voltage developed across R_1 . This is the only front-panel control of the group. C_4 is part of a capacitive voltage divider network that controls the amplitude of the waveform on the grid of the control tube. C_4 will therefore control the point at which V_1 starts to conduct. Finally, C_3 is part of a capacitance voltage divider and regulates the amount of deflection voltage reaching the horizontal output amplifier. Because of this it is fre-

quently called the drive control. It governs the width of the picture and, to some extent, its linearity. An additional linearity control located at a subsequent point in the horizontal sweep system regulates the overall horizontal linearity.

Gruen A.F.C. Circuit. General Electric (and several other manufacturers) are using an A.F.C. system which differs in some respects from the systems already discussed.

The basic circuitry and associated block diagram are shown in Fig. 13-24. The horizontal sweep oscillator is a triode-connected Hartley. L_1 , in the



Courtesy G.E.

FIG. 13.24. (A) Block diagram and (B) schematic circuit of Gruen A.F.C. system.

tank circuit, is the coarse frequency control. Shunted across L_1 are three capacitors which, with the coil, serve to establish the frequency at which the circuit operates. The most obvious capacitance is C_1 ($1,000 \mu\text{mf}$), and this is fixed in value. The second capacitance is C_2 ($270 \mu\text{mf}$), and this unit, in conjunction with R_1 and R_2 , also affects the oscillator frequency. The variation, of course, is achieved by the horizontal hold control potentiometer. As this places more or less resistance in the circuit, it changes the effect of C_2 on the resonant circuit and, in consequence, alters the frequency.

The third capacitance across the tuning circuit is C_3 ($470 \mu\text{mf}$), and this, like C_2 , is in series with a resistor. In this case, however, the resistance is that contained within V_4 , and just what value this resistance is will be determined by the bias on the tube. Thus, V_4 is the automatic control of the oscillator frequency, and it is V_4 which receives the d-c control voltage from the phase detector.

V_2 and V_3 form the phase detector where the incoming sync pulses are compared in frequency with pulses obtained from the plate of the damping tube. If the frequency relationship is not right, a resultant d-c voltage will

be produced which will be used to regulate the resistance of V_4 . This, in turn, will vary the Hartley oscillator frequency.

The arrangement of the phase detector circuit differs somewhat from previous phase detectors. The incoming sync pulses are applied to the cathodes of V_2 and V_3 in negative polarity. Since a negative pulse at the cathode of a tube is equivalent to a positive pulse at the plate, tube conduction will occur. Both tubes will conduct equally, resulting in equal voltage drops across their respective plate load resistors, R_3 and R_4 . (R_3 is for V_3 , R_4 is for V_2 .) However, both resistors are placed back-to-back with the result that their net voltage is zero.

Also applied to V_2 and V_3 is a series of pulses fed back from the damper plate (or any other point *after* the sweep oscillator). These pulses are converted into saw-tooth waves by R_{10} , R_{11} , C_4 , and C_5 . Thus, we have here somewhat the same condition as that existing in the previous saw-tooth A.F.C. system. V_2 and V_3 compare the saw-tooth wave with the incoming pulses and develop from this a resultant d-c voltage in the same manner as described for the previous A.F.C. system. This d-c control voltage is fed through a long-time-constant filter C_5 , R_5 , and C_6 and then through an anti-hunting circuit, R_6 , R_7 , C_7 , and C_8 , before it is applied to the grid of V_4 . The purpose of the anti-hunting circuit is to prevent the voltage from overswinging or hunting back and forth. This will occur when a correcting voltage is fed to the control circuit.

The purpose of R_8 , R_9 , and C_9 is to feed into the control circuit enough d-c bias to place V_4 at the correct operating point. The voltage developed by the phase detector then shifts V_4 above or below this point, as required. A portion of the large negative bias developed at the oscillator grid is employed for this purpose.

QUESTIONS

1. List and explain the purpose of every variable resistor and capacitor found in the circuit of Fig. 13-1.
2. Draw a low-voltage power supply which would be suitable for the deflection system of Fig. 13-1.
3. If the upper triode section of V_1 (Fig. 13-1) stopped functioning, how would it affect the image?
4. The video signal is reaching the control grid of the cathode-ray tube, but the clipper tube breaks down. What is the effect on the screen?
5. If a vertical line is obtained on the screen when the set is turned on, indicate all the possible points where the trouble could exist. Use Fig. 13-1.
6. What is the purpose of the linearity control in Fig. 13-1?
7. Why are saw-tooth voltage waves not generally suitable for use with deflection coils?
8. How is the proper deflection voltage for deflection coils developed?
9. Draw the differentiating circuit (with parts values) used in Fig. 13-4. Do the same for the integrating network.

10. What is a damping tube? Why do we use a damping tube only in the horizontal system? What is used in the vertical system to accomplish the same purpose?

11. If a leakage path developed in one of the .005- μ f capacitors which couple the horizontal deflection voltage to the 7JP4 of Fig. 13-1, how would this affect the image?

12. Draw the beam-centering system used in Fig. 13-4.

13. List and explain the purpose of every control found in the circuit of Fig. 13-4.

14. Why are automatic frequency control systems usually found only in the horizontal sweep system and not in the vertical system?

15. How does the A.F.C. system of Fig. 13-13 operate?

16. What does the horizontal discharge tube of Fig. 13-13 do?

17. How does the phase detector in the saw-tooth A.F.C. system operate?

18. Where is the long time constant filter of the pulse-width system located? List its components.

19. What is the purpose of an anti-hunting network? Where is one located in the circuits described in this chapter?

20. List the various controls in the pulse-width A.F.C. system. Indicate briefly what they do.

Chapter 14

FREQUENCY MODULATION *

General Outline. Frequency modulation, although only a newcomer so far as radio broadcasting history is concerned, has been chosen as the method of transmitting the audio portion of television programs. This choice was the result of several factors. Of the two systems available today, A-M and F-M, the latter has been proved capable of better reception under adverse conditions. It is easier to minimize interference from other near-by stations operating on the same frequency with frequency modulation than with amplitude modulation. Finally, there is the matter of cost, a factor especially applicable to transmitters. Because of the arrangement of the circuits in a frequency-modulated transmitter, it is more economical to produce a given wattage signal with this equipment than to produce the same wave with amplitude modulation. Specifically, the large difference in cost between the two systems lies in the audio power required to produce a certain strength signal. With A-M, the audio power is generally 50 per cent of the carrier power, which may entail many thousands of watts for a powerful station. On the other hand, in F-M, the audio required represents only a fraction of the output power and can be more easily generated.

The power relationship that exists in an amplitude-modulated wave between the sidebands and the carrier is in the ratio of 1:2 for 100 per cent modulation. This is only the average power and, when the equipment is designed, it must be capable of handling the much higher peak (or surge) power. Naturally, this requirement materially increases the cost of the station. In F-M transmission, the power output does not increase with modulation and no additional provision for handling excess power need be made.

The matter of fidelity is not stressed because, contrary to popular opinion,

* Due to space limitation, only the most important aspects of F-M can be considered here. For a more detailed analysis, the reader is directed to M. S. Kiver, *F-M Simplified*, 2nd Edition, published by D. Van Nostrand Co., Inc., 1951.

just as much fidelity is available with A-M as with F-M. It is only on the present crowded broadcast band (500 to 1,500 kc) that space is not available to permit the full 10,000 or 15,000 audio cycles to be reproduced. Given sufficient ether space, both systems have equal fidelity.

In order to discuss radio using frequency modulation, it first might be best to point out the differences between A-M and F-M waves and then to see why these differences exist. Toward that end, the following list has been tabulated:

<i>Factor</i>	<i>F-M</i>	<i>A-M</i>
Amplitude of Signal...	Remains constant	Varies with per cent of modulation
Audio Voltage	The frequency spread of signal is determined by the strength of the audio voltage	Determines the amplitude of the wave
Audio Frequency	The frequency of the audio modulating voltage will determine how rapidly the F-M wave will change from one frequency to another	The audio frequency controls the speed with which the amplitude of the wave changes
Signal Spread	The number of sidebands depends upon the amplitude of the modulating signal. In television, this spread is restricted to 25 kc on either side of carrier	Limited to 5 kc on either side of the carrier frequency. It is determined by the frequency of the modulating audio wave

The ordinary amplitude-modulated waveform, as it might appear for 100 per cent modulation, is shown in Fig. 14-1A. The audio-modulating signal adds to or subtracts from the amplitude of the carrier. When the modulating signal becomes too strong, overmodulation occurs and the carrier

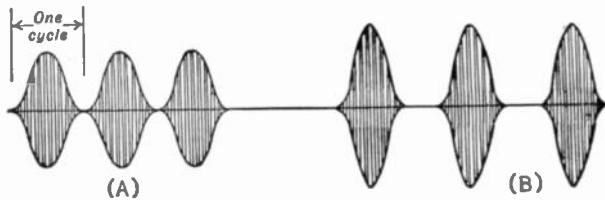


FIG. 14.1. Amplitude-modulated waves. (A) 100% modulation, (B) overmodulation.

is driven to zero for some short time. This is illustrated in Fig. 14-1B. Note that, whereas the amplitude of the wave may increase as much as possible, it can only decrease to zero. Whenever overmodulation occurs, the waveform becomes distorted and the greater number of frequencies that are generated by this process causes the bandwidth of the station to increase.

Hence all commercial broadcasting stations are careful to see that their output never reaches 100 per cent modulation.

A frequency-modulated wave is constant in amplitude, but varies in frequency. It would appear as pictured in Fig. 14-2. The property of constant amplitude makes the frequency-modulated wave so important. Most of man-made and natural interference has been found to affect the amplitude of a wave much more than its frequency. For the A-M signal, the interference distorts the waveform and, with this, the intelligence contained therein. F-M, on the other hand, contains its intelligence in its changing

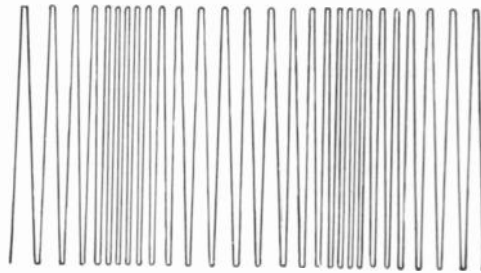


Fig. 14.2. Frequency-modulated wave. The changing spacing between cycles indicates different modulating frequency intensities.

frequencies. At the F-M receiver, one of the I.F. stages, called the limiter, smooths any irregularities in the amplitude of the incoming signal and by this process eliminates the interference.

The frequency bandwidth of an F-M wave depends upon the strength of the impressed audio voltage. At the transmitter, the carrier frequency is fixed by a self-excited oscillator. This frequency is the mean or center frequency of the broadcast station. When the sounds that are to be transmitted are fed into the microphone, the mean frequency of the transmitter is varied. The louder the audio signal, the greater the deviation. For example, a frequency deviation (or change) of 50 kc in the output might occur for a strong audio voltage, whereas only a 1-ke change would occur if the audio voltage were weak. In the A-M case, the amplitude and not the frequency of the wave changes for different audio sound levels.

The rapidity with which the F-M transmitter frequency moves from one point to another is determined by the frequency of the modulating sound. A high-pitched sound would cause the frequency of the F-M transmitter to change more rapidly than if 60 or 100 cycles were used.

One definite advantage obtained with frequency modulation is due to the observed (and calculated) fact that, if two signals are being received simultaneously, the effect of the weaker signal will be eliminated almost entirely if it possesses less than half the amplitude of the other stronger

signal. This means that for one signal to completely override another at the receiver, their amplitudes need be in the ratio of 2:1, or more. With a good antenna, it is easy most times to tune in one station in sufficient strength so that the other interfering station or stations are eliminated entirely. No

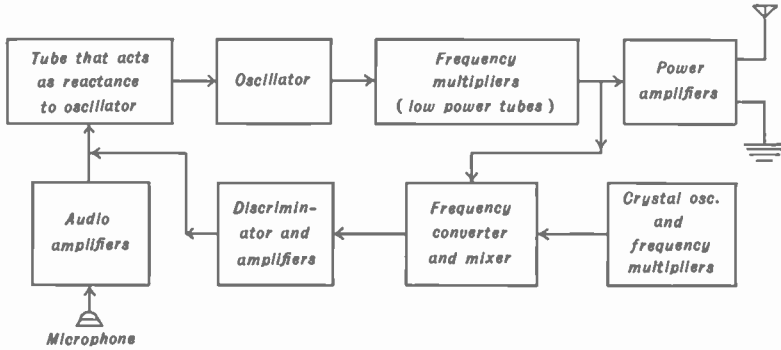


FIG. 14.3. Crosby system or reactance-tube method of producing F-M waves.

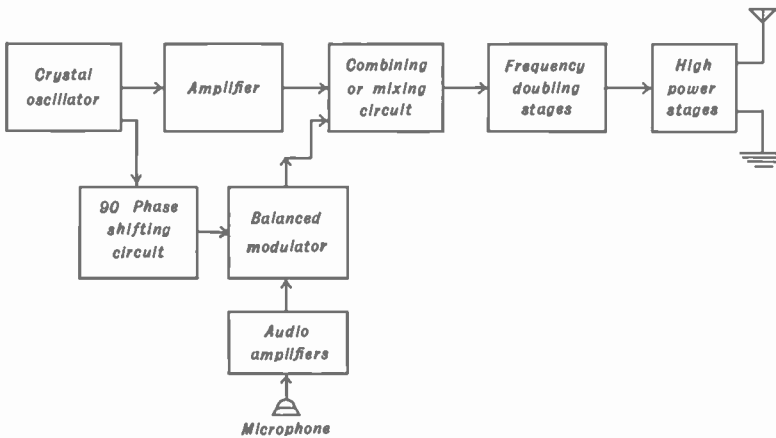


FIG. 14.4. Armstrong system of frequency modulation.

such situation exists with A-M signals, where interfering stations can be heard when even a 100:1 relationship exists between the various carrier amplitudes.

Transmitters. It would be difficult to analyze the functions of a frequency-modulated transmitter by comparing it with the more familiar amplitude-modulated transmitter because of the great dissimilarity between them. Consider the two basic methods used to develop an F-M signal shown in block form in Figs. 14-3 and 14-4. In Fig. 14-3 we have the reactance-tube, or Crosby, system, whereas in Fig. 14-4 we have the Armstrong method

of frequency-modulated transmission. Although both methods are highly involved and mathematically complex in design, a sufficiently simple explanation is possible to indicate the general functions of the various stages of each of the systems.

In the Crosby transmitter, the heart of the process is to be found in the reactance tube. This tube and its circuit are directly connected to a single frequency oscillator, the latter being perhaps some form of Hartley oscillator. As long as no audio signal is impressed on the grid of the reactance tube, no plate current variations will occur in the output of the tube, which is directly connected across the tuning coil of the oscillator. Under such conditions, the oscillator generates its mean or carrier signal frequency. Any receiver tuned to this transmitter would not, at this particular moment, receive any audio output.

Upon speaking into the microphone, however, a varying voltage is generated which is amplified by conventional audio stages and placed on the grid of the reactance tube. These audio variations cause the reactance tube plate current to assume different values. Because the plate current must pass through the oscillator coil and because of the phase relations of the associated circuit, the reactance tube may be made to appear either inductive or capacitive to the oscillator, thus causing its frequency to increase or decrease. This is the first step in the formation of a frequency-modulated signal. The audio variations have directly affected the frequency of the oscillator, with a stronger audio signal causing a greater change in frequency than a weaker audio signal.

The frequency variations or deviations from the mean or carrier frequency are at this point quite small. For example, with a mean frequency of 1 mc, the audio voltage may cause a frequency variation of perhaps plus and minus 10 kc about the 1 mc. Feeding this signal into a doubler stage will result in an output of 2 mc having a frequency deviation about this carrier value of plus and minus 20 kc. By applying more of these multipliers, the final carrier frequency is obtained having a frequency variation of plus and minus 100 kc (200 kc in all). While the values used here as illustrations may not be the exact figures found in practice, they do illustrate how the final signal spread is obtained. In any one stage, the frequency range must be a small percentage of the mean or carrier frequency at that point in order to prevent distortion.

The other sections of the transmitter shown are concerned with keeping constant the mean frequency of the self-excited oscillator. The manner in which this is accomplished is indicated in Fig. 14-3. A crystal oscillator generates a single, stable frequency. The output from the crystal oscillator is passed through a sufficient number of frequency multiplier stages so that a frequency equal to the exact carrier frequency of the transmitter (as fixed

by the F.C.C.) is reached. This signal from the crystal oscillator section is then fed to a mixer where it combines with the actual carrier of the transmitter itself. If there is no difference between these two values, no voltage is sent back to the reactance tube. However, if something has caused the transmitter carrier frequency to drift, a resultant voltage will be obtained from the discriminator, and the bias of the reactance tube will be varied accordingly. A positive change in bias of the reactance tube will affect the oscillator frequency one way, while a negative change will result in the opposite reaction. In either case, the output of the discriminator will always act so as to bring the mean frequency of the main oscillator back to its correct value. With accurate adjustment, the transmitter carrier will always remain at its assigned value, the limits being determined by the correcting crystal oscillator.

A second system commonly used for producing frequency modulation is shown in block diagram form in Fig. 14-4. In this transmitter, the fundamental frequency is determined by a crystal-controlled oscillator operating at about 200 to 300 kc. A portion of the output of this oscillator is fed to an amplifier while the other portion is sent through a 90° phase-shifting circuit and a balanced modulator. The modulator merely combines the phase-shifted carrier with the audio-modulating voltage. Due to the action of a balanced modulator, sidebands are generated, just as in any ordinary amplitude-modulated transmitter. The one difference, however, lies in the fact that the carrier is automatically suppressed and only the sidebands appear at the output of the stage. These sidebands, when recombined with the original carrier, will cause frequency modulation, although at this point the amount of frequency variation about the carrier is not very great.

From this point on, frequency multipliers increase the amount of frequency deviation obtained until it is as large as desired. Then isolating power amplifiers increase the strength of the signal and feed it to the transmitting antenna. The heart of this system is in the 90° phase-shifting network, for it is through its action that frequency modulation is obtained.

In the description of the two systems, only the basic outlines have been given. The number and type of stages that are found depend upon the output power desired and the frequency of the transmitter. However, except for the last few power amplifiers, the intervening tubes may be of the small inexpensive receiving type commonly found in home sets. Thus we can see that the audio power required in either one of these two systems is very small, much less than the amount necessary with amplitude modulation.

F-M Receivers. Although it is difficult to draw a comparison between A-M and F-M transmitters, it is possible to show the similarity between the respective receivers. The F-M receiver is a superheterodyne in practically all instances, although tuned-radio-frequency (TRF) sets are also

possible. The superheterodyne offers so many more advantages than a simple tuned-radio-frequency receiver that the latter is seldom used.

The block diagrams of Fig. 14-5 illustrate the differences between A-M and F-M superheterodynes. Besides the limiter and discriminator stages in the F-M receiver, both sets would appear to be exactly alike, and indeed might easily be mistaken for each other from an ordinary schematic. Up to the limiter stage, the essential difference between the two types of receivers resides almost wholly in the tuning circuits that connect each stage. In F-M, these circuits must be capable of receiving higher frequencies and

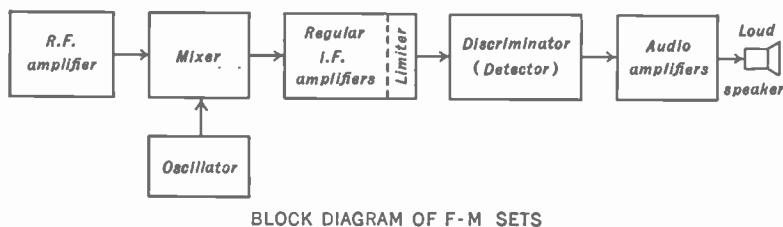
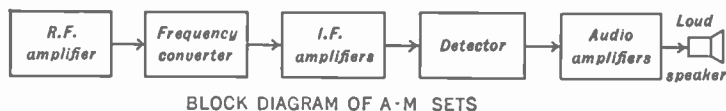


Fig. 14.5. A comparison of the block diagrams of A-M and F-M receivers.

also be capable of passing a wider band of side frequencies associated with the F-M carrier. In the ordinary F-M receiver, designed for use between 88 and 108 mc, each station is allowed sidebands ranging up to 75 kc on either side of the carrier. For television audio, merely 25 kc is used, the narrower bandwidth simplifying somewhat the problem of receiving both the video and the audio carriers simultaneously.

One special feature common to many F-M receivers is the use of a separate oscillator that feeds the mixer tube and helps generate the I.F. At the higher frequencies employed for the television audio, normal drifting by the oscillator produces more marked effects than it does in the lower frequency broadcast receivers. At 60 mc, 2 per cent frequency drift would shift the signal beyond the bandpass of the audio I.F. circuits; at 1 mc, the same percentage shift is only 20 kc and would not shift a signal beyond the bandpass of circuits designed for a ± 75 -kc signal spread. The separate tube arrangement results in greater stability, with drift reduced to a smaller fraction than would be present in designs using the same tube for mixing and generating the oscillator voltage. Often, such additional devices as compensating ceramic capacitors are placed in the oscillating tank circuit in

order to counteract tendencies on the part of the other frequency determining components to change with operating conditions.

The most common I.F. in use today for the F-M receiver (not associated with television) is 10.7 mc. Television sound I.F. values fall at or near 4.5 mc, 21.25 mc, or 41.25 mc.

Limiters. The first significant difference between the A-M and F-M superheterodynes is noted at the limiter stage or stages. Essentially, the purpose of a limiter is to eliminate the effects of amplitude variations in the F-M signal. While it may have been true that the frequency-modulated

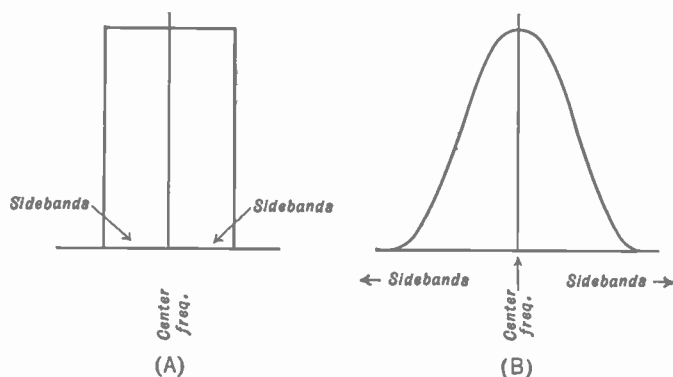


FIG. 14.6. Receiver response curves. (A) is the ideal curve, (B) a typical practical result.

signal left the transmitter with absolutely no amplitude variations, this is almost never true by the time the signal reaches the limiter.

To digress for a moment, let us see where, in the receiver itself, various portions of the F-M signal could have received more amplification than other parts of the signal, the result naturally causing some frequencies to possess greater amplitudes. An ideal response curve for a tuned circuit is shown in Fig. 14-6A. With such a characteristic, each frequency within the signal would receive uniform amplification. Such a happy situation, however, is seldom encountered in practice. The more usual state of affairs is illustrated by the curve of Fig. 14-6B. Here it is apparent that the center frequencies receive more amplification than those located farther away. Hence, even if the incoming signal were perfectly uniform, by the time it arrived at the limiter amplitude variations would be present. The result would be distortion if this wave were allowed to reach the speaker. It is for the limiter to remove the amplitude variation.

A typical limiter stage is given in Fig. 14-7. Inspection reveals that low plate and screen voltages are used, in addition to grid-leak bias in the input or grid circuit. The low electrode voltages cause the tube to reach

current saturation with moderate signals at the grid. The use of grid-leak bias aids in keeping the output plate current (and hence the output signal) constant for different input voltage levels. It is readily apparent that, with

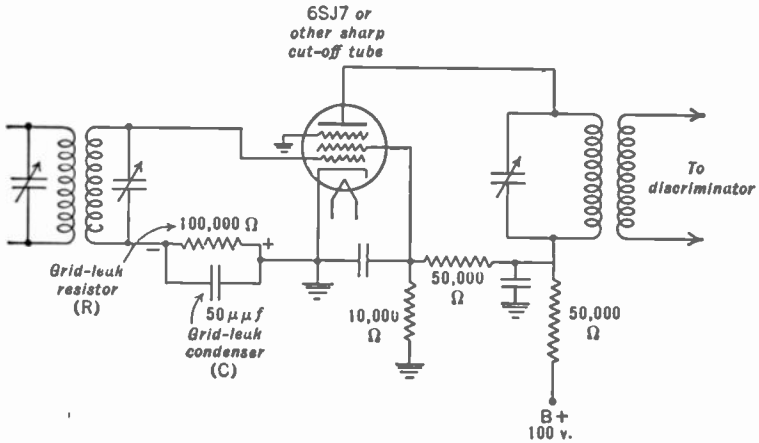


FIG. 14.7. A familiar limiter stage circuit.

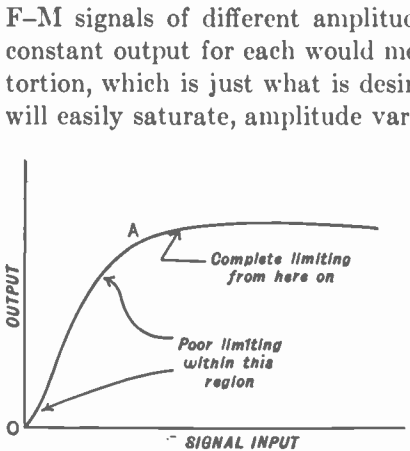


FIG. 14.8. The characteristic curve for a limiter.

the output signal of the tube increases with input signal until a certain voltage is reached. Beyond this point, known as the knee of the curve, point A, the limiter plate current remains substantially constant for all stronger input voltages. Since complete limiting begins at this point, the signals at the antenna of the receiver must receive sufficient amplification to force the limiter tube to operate beyond point A. From this point, the output of the

most disturbing noises. This is all possible because of one fact — namely, that much man-made or natural interference affects the amplitude of the radio signal more than its frequency. By the simple device of smoothing out the amplitude differences of the incoming waves (without affecting their fundamental frequency), we eliminate the noise or interference. This is one reason for the extensive use of F-M.

A limiter characteristic curve is shown in Fig. 14-8. Notice that the

limiter will remain constant. Any signal which is so weak that it is unable to operate the tube beyond OA will have its noise appear in the limiter output.

The situations for weak and strong signals may be demonstrated graphically. The curve OAB in Fig. 14-9 is the relationship between the input grid voltage or signal and the resulting plate current in the output of the limiter. With the tube biased to point C, the input signal voltage will vary

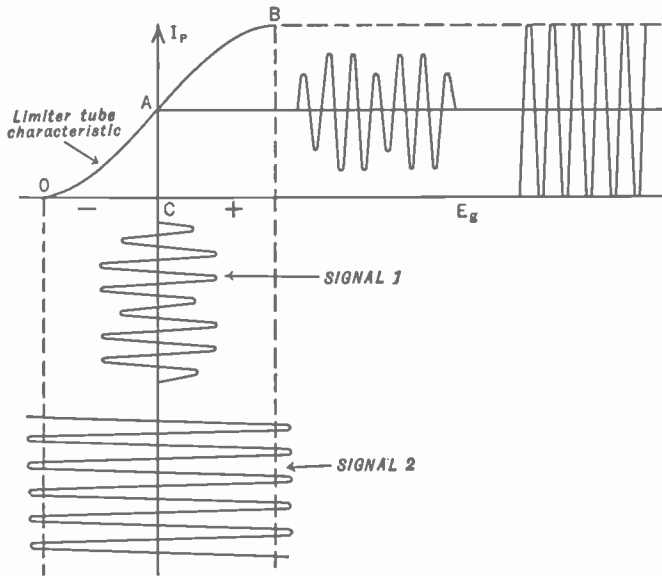


FIG. 14.9. These curves illustrate that the incoming signal must reach a certain amplitude before the limiter stage will saturate.

about this point. Consider the first small signal coming in. As it varies the grid bias, corresponding changes take place in the plate circuit and at no time will the plate current be forced to its saturation value. This means that any noise and amplitude distortion contained in this signal will be amplified and reproduced in the plate circuit and, from here, go to the discriminator. Frequency modulation in this case will not eliminate the interference.

Now consider the second signal voltage. At all peak points of the signal, plate current saturation is reached on the positive peaks, while current cut-off is responsible for smoothing out the negative peaks. In the output circuit, all trace of amplitude distortion has been clipped or eliminated. When this signal is fed to the discriminator, it should give noise-free operation. Thus, while a limiter provides F-M with its greater advantages, care must be taken to see that it is properly operated; otherwise its useful-

ness is lost. The F-M receiver must be so designed that all desired signals to the input receive sufficient amplification. When this is done, the plate current of the limiter will give constant output.

It should be mentioned that, by clipping off the top of the waves in order to have them all reach the same level, we are introducing amplitude distortion into the signal. Ordinarily this would be objectionable but, in the present case, the new frequencies generated by this clipping are harmonics of the frequencies arriving at the limiter, and the resonant circuit in the output of the stage does not react to them. Thus the harmonics are filtered out and can cause no damage. Only signals having frequencies within the band-pass limits of the resonant coil and condenser develop sufficient voltage to be passed on to the next inductively coupled circuit.

It is possible to design limiters on the basis of low plate and screen voltages alone, but better results and more amplification are obtained if grid-leak bias is added to this combination (see Fig. 14-7). With the insertion of grid-leak bias, it is possible to raise the electrode voltages, increasing the gain somewhat. The tube initially has zero bias with no signal at the grid. As soon as a signal acts, the grid is driven slightly positive, attracts electrons, and charges the capacitor C , Fig. 14-7. This capacitor attempts to discharge through R but, due to the relatively long time constant of R and C , the discharge occurs slowly. Because of current flow through R , a voltage is developed, with the end nearest the grid becoming negative. This voltage will act as a bias, varying in value as the incoming signal varies and in this way tending to keep the plate current steady within rather wide limits of input voltage. A strong signal causes the grid to become more positive, resulting in a greater current flow through R . A larger bias is developed. A weaker signal will cause less voltage, resulting in essentially the same amount of plate current. Usual values of C range from 30 to 60 μmf and, for R , between 50,000 and 200,000 ohms.

Because the voltage across R , the grid-leak resistor, will vary with the amplitude of the incoming signal, this point of the limiter is generally used for two purposes: one is for aligning the preceding I.F. amplifiers, and the other is for magic-eye tuning devices. The tuning eye operates on differences in voltage and this point in an F-M receiver is well suited to indicate the strength of the incoming signal.

For alignment, the vertical cable of the oscilloscope, when placed across the ends of the grid-leak resistor, will indicate the form of the signal and will permit the adjustment of the preceding I.F. transformers for maximum response. It will be seen that this process is used in aligning the receiver described in Chapter 15.

Although one limiter stage serves satisfactorily, better results can be obtained with two stages, one following directly behind the other. The

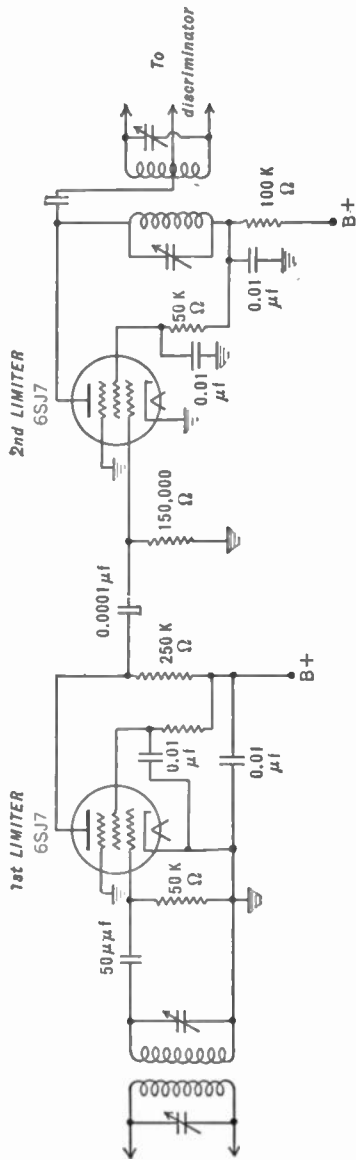


Fig. 14.10. A dual limiter circuit.

circuit of two limiters in cascade is shown in Fig. 14-10, and their characteristic curve is given in Fig. 14-11. With two limiters, the knee of the resulting curve becomes sharper and provides better limiting action. In all limiter circuits, sharp cut-off pentodes are used, with tubes having the highest values of G_m preferred.

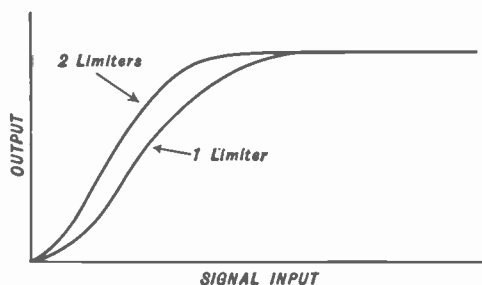


Fig. 14.11. A comparison of limiter characteristic curves for single and dual stages.

The Discriminator. The purpose of the second detector in an amplitude-modulated set is to obtain the audio variations from the incoming modulated signal. The same stage in a frequency-modulated receiver must derive the audio variations from the different incoming frequencies. Thus, although the end product in both cases is the same, the methods used are quite different. We know that with F-M a large frequency deviation from the carrier means a loud audio signal, whereas a small frequency deviation means a weak audio note. Hence, some circuit must be devised that will develop voltages proportional to the deviation of the various incoming frequencies about the carrier.

A simple circuit that discriminates against various frequencies is the elementary parallel (or series) resonant circuit. As is well known, this circuit will develop maximum voltage at the resonant frequency, with the response falling off as the frequency separation increases on either side of the central or resonant point.

One of the first discriminators used in F-M receivers contained two resonant circuits in an arrangement as shown in Fig. 14-12. The primary coil L_1 is inductively coupled to L_2 and L_3 , each of which is connected to a diode tube. Each tube has its own load resistor, but the output of the discriminator is obtained from the resultant voltage across both resistors.

In order to determine the frequencies to which L_2 and L_3 must be tuned, it should be recalled that when an audio-modulating signal alters the frequency of an F-M transmitter it varies this frequency above and below one central or carrier value. Thus, for a sine wave, the maximum positive portion would increase the frequency, say by 40 kc, while the maximum negative section would decrease the carrier frequency by the same amount.

At intermediate points, less voltage would cause correspondingly less frequency deviation.

To have the discriminator function in a similar manner over the same range, L_2 and L_3 are each peaked to one of the two end points of the I.F.

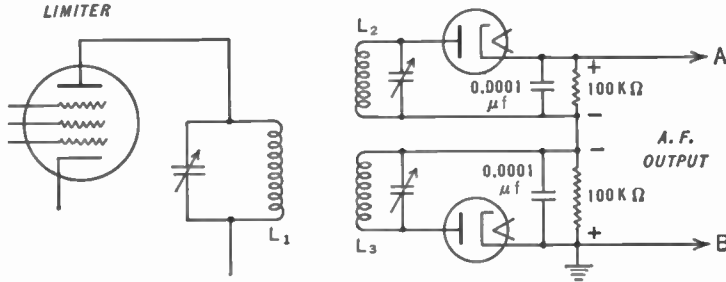


FIG. 14.12. A simple discriminator circuit.

band. For example, if the I.F. band spread extends from 21.15 mc to 21.35 mc (with 21.25 as the mean or carrier frequency), L_2 might be peaked to 21.15 mc, and L_3 would be peaked to 21.35 mc. The response curves would then appear as in Fig. 14-13.

The two curves are positioned in the manner shown because of the way their load resistors are connected in the circuit. According to the arrange-

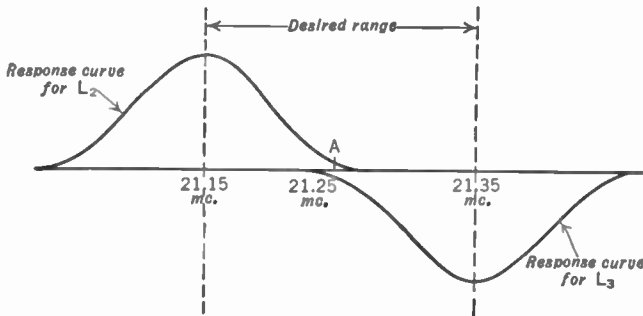


FIG. 14.13. The overall response curve for the discriminator of Fig. 14.12.

ment, the voltages developed across them tend to oppose each other, as indicated by the polarities across the resistors in the circuit of Fig. 14-12.

At the center frequency, point A of Fig. 14-13, the two voltages developed across the load resistors cancel each other and the resultant voltage is zero. By similarly adding the voltages at other points about the carrier, we obtain the overall resultant curve shown in Fig. 14-14. This is the familiar S-shaped curve of all frequency discriminators and shows how the output voltage of the second detector will vary as the incoming frequencies

change. Specifically, suppose the signal acting at the input to the discriminator at any one instant has a frequency of 21.31 mc. The amount of voltage developed at the output is given by point 1 on the vertical axis. Then, at the next instant, if the frequency should change to 21.19 mc, the output voltage is indicated by point 2. Notice that all frequencies below 21.25 mc result in positive output voltages, whereas all those above 21.25 mc give rise to negative output voltages. In this way, the audio voltages

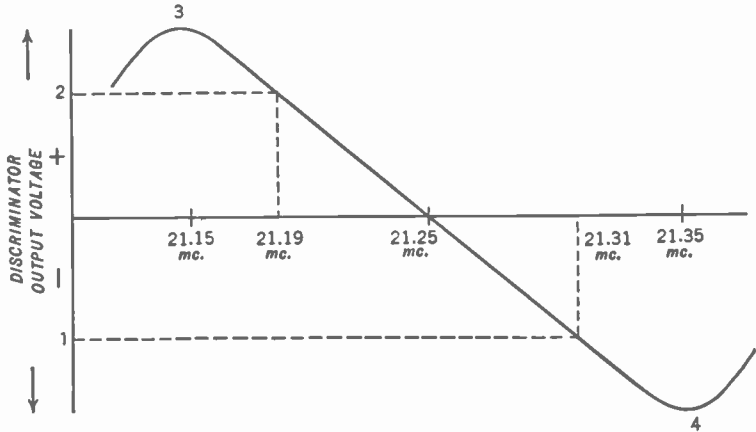


FIG. 14.14. The resultant S-shaped discriminator characteristic curve obtained by adding the two separate curves of Fig. 14.13.

that modulated the carrier frequency at the transmitter are extracted in the receiver.

The useful portion of this discriminator characteristic curve is the linear portion included between the two maximum points, 3 and 4. Any non-linearity along this section of the curve would produce amplitude distortion in the output audio signal. When discriminators are designed, the maximum points 3 and 4 are generally set much farther apart than required in the particular receiver. This insures a linear curve at those frequencies that are actually used, since the response characteristic has a tendency to curve near the maximum peaks. By utilizing a smaller range, amplitude distortion in the output signal is kept to a minimum. The sections of the curve of Fig. 14-14 beyond points 3 and 4 are completely disregarded. The frequency of the output voltages is determined by how rapidly the frequency of the incoming I.F. signal varies. A large frequency deviation in the input signal gives rise to a strong output wave, and the rapidity with which this incoming frequency changes determines whether the strong output will be high or low pitched.

A Modified Discriminator. It may be wondered why the preceding circuit was described in such detail if it is not used in modern receivers. The reason lies simply in the ease with which this circuit brings out the fundamental conversion process at the second detector of an F-M receiver and because it is basically the same as the present-day discriminator of Fig. 14-15. Instead of employing two separate condensers for the sec-

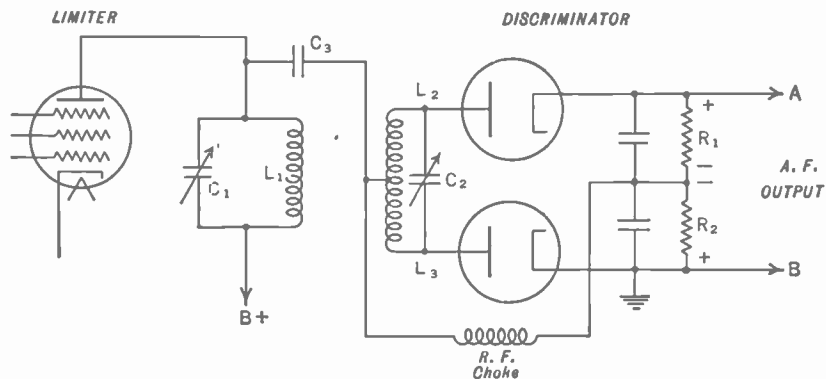


FIG. 14.15. A common discriminator in F-M receivers.

ondary circuits, only one is used. R_1 and R_2 are the load resistors, one for each diode, and the resultant output audio voltage is still obtained across points A and B. The use of one capacitor, instead of two, results in greater ease in aligning the circuits and economy in construction. The tap divides the secondary coil into essentially two identical coils, L_2 and L_3 .

Circuit operation depends upon the voltages developed across L_2 and L_3 for the various incoming frequencies. The voltages add vectorially with the voltage in the primary coil L_1 , which is brought over to the secondary circuit through condenser C_3 . A complete discussion of the operation of this circuit would involve determining the phase relationships between the various coils, for it is only in this way that the different voltages appearing at the output of the discriminator can be computed. The different phase relations are brought about by the changing frequencies that enter the discriminator. For example, at resonance the secondary tuning circuit acts purely resistive to an incoming signal. For frequencies above resonance, the inductive reactance of the secondary circuit predominates whereas, for frequencies below resonance, the capacitive reactance determines the phase of this circuit. As these phase relations fluctuate, the output voltage taken from across R_1 and R_2 likewise changes. A characteristic curve similar to the S-shaped one of Fig. 14-14 is obtained for this discriminator.

There are many variations of this discriminator schematic, two of the more common ones being given in Fig. 14-16 A and B. The output termi-

nals for each type are clearly indicated. They connect to the following audio amplifiers. The discriminator of Fig. 14-16B is used in General Electric television receivers. This design, while still being fundamentally the same as the others, permits a tube with a single cathode to be used, which is advantageous for some circuits.

The need for limiter stages arises because the discriminators noted in the preceding paragraphs are sensitive to the amplitude of the incoming signal. Stating this in other words, these discriminators are not pure

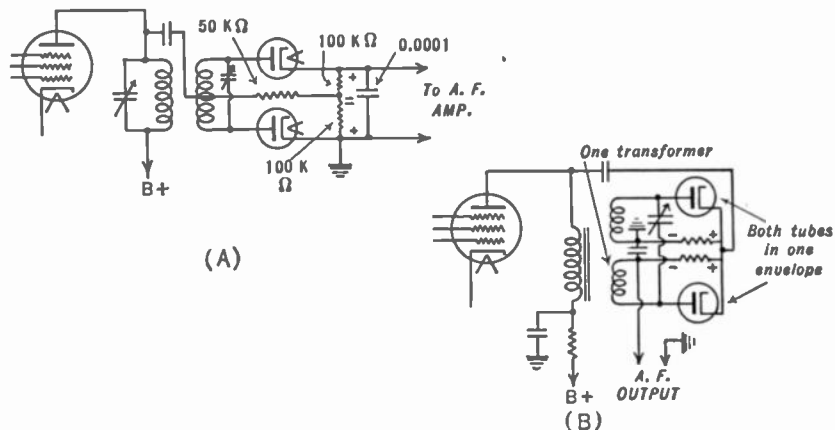


FIG. 14.16. Two additional discriminator circuits found in use. The arrangement in (B) has the advantage of placing both cathodes at the same potential. This permits use of duo-diodes with one common cathode.

F-M detectors. It is the purpose of the limiter to eliminate any amplitude variations contained in the arriving signal and present to the following discriminator a signal which is wholly F-M. It has long been recognized that the development of a detector which was more immune to A-M would simplify the circuit of the F-M receiver. Excessive amplification must be given to a signal before it reaches the limiter in order to have the signal drive the limiter into saturation. But, if we could either do away with the limiter entirely or, at least, use only a partial limiter, then the demand for amplification would be lessened. Such a detector has been developed and its operation is considered in the following paragraphs.

F-M Ratio Detector. To understand why a ratio detector enjoys greater immunity from A-M distortion in the incoming F-M signal, let us compare its operation with that of the ordinary discriminator.

In the discriminator circuit of Fig. 14-15, let the signal coming in develop equal voltages across R_1 and R_2 . This would occur, of course, when the incoming signal is at the center I.F. value. Suppose that each

voltage across R_1 and R_2 is 4 volts. When modulation is applied, the voltage across each resistor changes, resulting in a net output voltage. Say that the voltage across R_1 increases to 6 volts and the voltage across R_2 decreases to 2 volts. The output voltage would then be equal to the difference between these two values, or 4 volts.

However, let us increase the strength of our carrier until we have 8 volts, each, across R_1 and R_2 , at mid-frequency. With the same frequency shift as above, but with this stronger carrier, the voltage across R_1 would rise to 12 volts and that across R_2 decrease to 4 volts. Their difference, or 8 volts, would now be obtained at the output of the discriminator in place

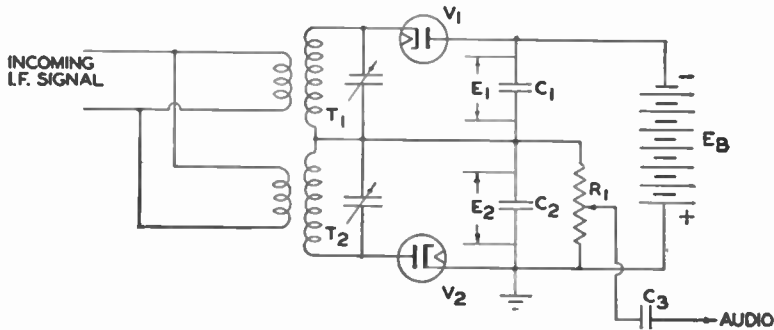


Fig. 14.17. Preliminary form of the ratio detector.

of the previous 4 volts. Thus the discriminator responds to both F-M and A-M. It is for this reason that limiters are used. The limiter clips all amplitude modulation off the incoming signal and an F-M signal of constant amplitude is applied to the discriminator.

When unmodulated, the carrier produced equal voltages across R_1 and R_2 . Let us call these voltages E_1 and E_2 respectively. With the weaker carrier, on modulation, the ratio of E_1 to E_2 was 3 to 1 since E_1 became 6 volts and E_2 dropped to 2 volts. With the stronger carrier, on modulation, E_1 became 12 volts and E_2 dropped to 4 volts. Their ratio was again 3 to 1, the same as with the previous weaker carrier. Thus, while the difference voltage varied in each case, the ratio remained fixed. This demonstrates, in a very elementary manner, why a ratio detector could be unresponsive to carrier changes.

An elementary circuit of a ratio detector is shown in Fig. 14-17. In this form, the detector is similar to the detector of Fig. 14-12, where each tube has a completely separate resonant circuit. One circuit is peaked slightly above the center I.F. value (say T_1); the other peaked to a frequency below the center (say T_2). The output voltage for V_1 will appear across C_1 , and the output voltage for V_2 will be present across C_2 . The

battery, E_b , represents a fixed voltage. Since C_1 and C_2 are in series directly across the battery, the sum of their voltages must equal E_b . Also, due to the manner in which the battery is connected to V_1 and V_2 , no current can flow around the circuit until a signal is applied. Now, while $E_1 + E_2$ can never exceed E_b , E_1 does not have to equal E_2 . In other words, the ratio of E_1 to E_2 may vary. The output voltage is obtained from a resistor connected across C_2 .

When the incoming signal is at the I.F. center value, E_1 and E_2 will be equal. This is similar to the situation in the previous discriminator. However, when the incoming signal rises in frequency, it approaches the resonant point of T_1 and the voltage across C_1 likewise rises.

For the same frequency, the response of T_2 produces a lower voltage. As a consequence, the voltage across C_2 decreases. However, $E_1 + E_2$ is still equal to E_b . In other words, a change in frequency does not alter the total voltage, but merely the ratio of E_1 to E_2 . When the signal frequency drops below the I.F. center point, E_2 exceeds E_1 . The sum, however, of $E_1 + E_2$ must equal E_b . The audio variations are obtained from the change of voltages across C_2 . Capacitor C_3 prevents the rectified d-c voltage in the detector from reaching the grid of the audio amplifier. Only the audio variations are desired.

The purpose of E_b in this elementary explanatory circuit is to maintain an output audio voltage which is purely a result of the F-M signal. E_b keeps the total voltage ($E_1 + E_2$) constant, while it permits the ratio of E_1 to E_2 to vary. As long as this condition is maintained, we have seen that all amplitude variations in the input signal will be without effect.

The problem of deciding upon a value for E_b is an important one. Consider, for example, that a weak signal is being received. If E_b is high, the weak signal would be lost because it would not possess sufficient strength to overcome the negative polarity placed by E_b on the tubes V_1 and V_2 . The tubes, with a weak input voltage, could not pass current. If the value of E_b is lowered, then powerful stations are limited in the amount of audio voltage output from the discriminator. This is due to the fact that the voltage across either capacitor — C_1 or C_2 — cannot exceed E_b . If E_b is small, only small audio output voltages are obtainable. To get around this restriction, it was decided to let the average value of each incoming carrier determine E_b . Momentary increases could be prevented from affecting E_b by a circuit with a relatively long time constant.

The practical form of the ratio detector is shown in Fig. 14-18. It uses the phase-shifting properties of the discriminator of Fig. 14-15. R and C_3 take the place of E_b and the voltages developed across R will be dependent upon the strength of the incoming carrier. Note that V_1 and V_3 form a series circuit with R (and C_3) and any current flowing through these tubes

must flow through R . However, by shunting the $8\text{-}\mu\text{f}$ electrolytic condenser across R we maintain a fairly constant voltage. Thus, momentary changes in carrier amplitude are merely absorbed by the condenser. It is only when the *average* value of the carrier is altered that the voltage across R is

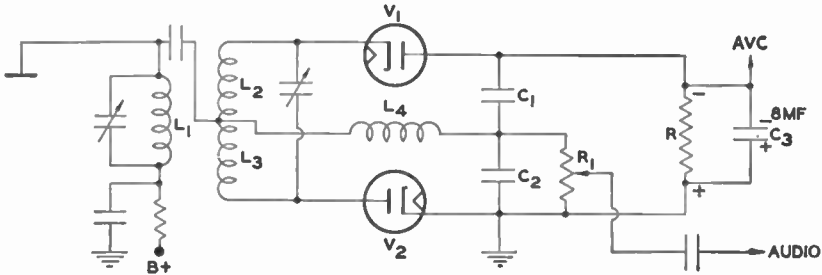


FIG. 14.18. Practical form of the ratio detector.

changed. The output audio-frequency voltage is still taken from across C_2 by means of the volume control.

Since the voltage across R is directly dependent upon the carrier strength, it may also be used for A.G.C. voltage. The polarity of the voltage is indicated in Fig. 14-18.

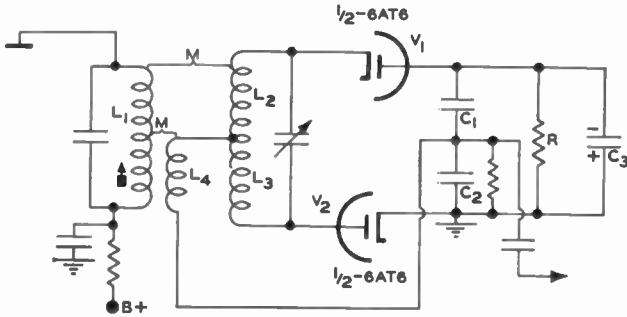


FIG. 14.19. In this ratio detector, L_4 receives its voltage from L_1 by inductive coupling.

Ratio Detector Modifications. Another form of the ratio detector is shown in Fig. 14-19. L_4 , instead of receiving its voltage from the primary L_1C_1 through a direct capacitive connection, is now coupled inductively to L_1 . Circuit operation remains unchanged because the voltage across L_4 still depends directly upon the voltage existing across L_1 .

The two foregoing ratio detectors (Figs. 14-18 and 14-19) are unbalanced circuits, so-called because V_1 and V_2 are not equally balanced against ground. We can transform these circuits into a balanced ratio detector by

moving the position of the ground connection, as shown in Fig. 14-20. In place of one stabilizing condenser (C_3) and resistor, R , we now have two of each. Their function, however, remains the same.

To see how this circuit works, let us assume that the voltage coming in is at the I.F. mid-frequency. Equal voltages will be applied to V_1 and V_2 , and equal voltages will appear across C_4 and C_5 , with the polarity noted. At the same time, there will be a voltage developed across A-B equal to the average value of the signal. Half this voltage will appear across R_1C_6 , and half of the total voltage will be across R_2C_7 . Since C_4 is in parallel with R_1 and C_6 , all will have the same voltage. The same is true of C_5 ,

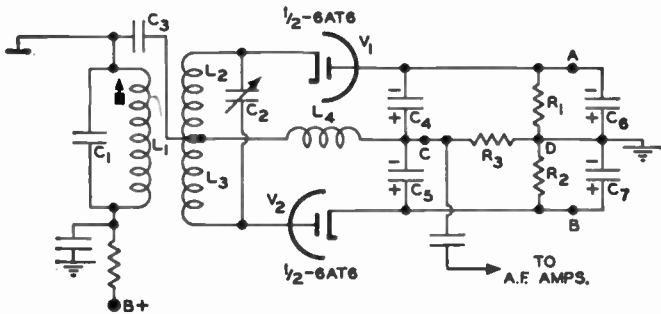


FIG. 14.20. A balanced ratio detector.

R_2 , and C_7 . As a consequence of these conditions, there will be no *difference of potential* between points C and D and no audio output. This, again, is similar to the previous discriminators.

Now let the signal frequency swing below the mid-I.F. value. V_1 will receive more voltage than V_2 and more voltage will develop across C_4 and less across C_5 . The carrier amplitude has not changed, however, because this is an F-M signal, and consequently the voltage between points A and B remains the same. To use numerical values, assume that the voltage across R_1C_6 is 6 volts. Because of the signal frequency shift, the voltage across C_4 rose from 6 to 9 volts, while C_5 dropped from 6 to 3 volts. Since E_{C_4} is now greater than E_{R_1} , and E_{C_5} is less than E_{R_2} , currents will flow in these circuits. These currents, flowing through R_3 , will develop a voltage drop of 3 volts here, with point C positive and point D negative. If now we add the voltages around both branches, we see that they check out. The rise in voltage across R_3 results in an audio output. When the signal swings in the opposite direction, C_5 receives more voltage than C_4 , and the polarity of the voltage across R_3 reverses. In this manner, for signal frequency swings above and below the center point, positive, zero, and negative voltages will develop across R_3 . This is the audio output voltage. At all times, the sum of the voltages across C_4 and C_5 must equal the average

carrier voltage present between points A and B. Changes in frequency do not alter the total voltage but merely the ratio of E_{C_4} to E_{C_5} . That is why this is known as a ratio detector. Changes in signal amplitude will not change the ratio of E_{C_4} to E_{C_5} .

To illustrate this, consider the example used in the foregoing paragraph. A signal-frequency shift caused the voltage across C_4 to rise from 6 to 9 volts, while the voltage across C_5 dropped from 6 to 3 volts. The ratio of E_{C_4} to E_{C_5} is $9/3$ or $3/1$. Now let us assume that the carrier amplitude is doubled, momentarily. This would double both sides of our ratio but leave the value of the ratio unchanged. Thus, $9/3 = 18/6 = 3/1$.

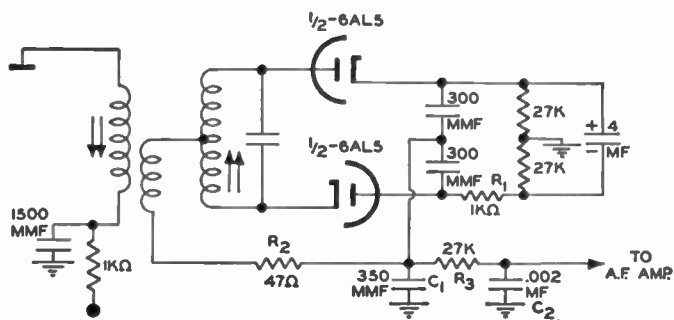
Any momentary increase in carrier will affect the numerator and denominator of this ratio in like measure and consequently leave the basic value of the ratio unaltered. The same is true of carrier decreases. Actually, because of the presence of the long-time-constant network of R_1C_6 and R_2C_7 , the momentary changes in carrier amplitude only *tend* to make the voltages across C_4 and C_5 go up or down. By the time the voltage in the circuit actually changes, the pulse or disturbance has passed.

Several forms of the balanced ratio detector commonly encountered are shown in Fig. 14-21. In the first circuit (Fig. 14-21A) R_1 is inserted to provide better balance between both halves of the circuit, and R_2 limits the peak plate current drawn by each diode section of the 6AL5. C_1 shunts I.F. voltages away from the audio output, while R_3 and C_2 comprise a de-emphasis filter to equalize the audio signal back to its original form. An A.G.C. voltage can be obtained from the negative side of the $4\text{-}\mu\text{f}$ stabilizing capacitor.

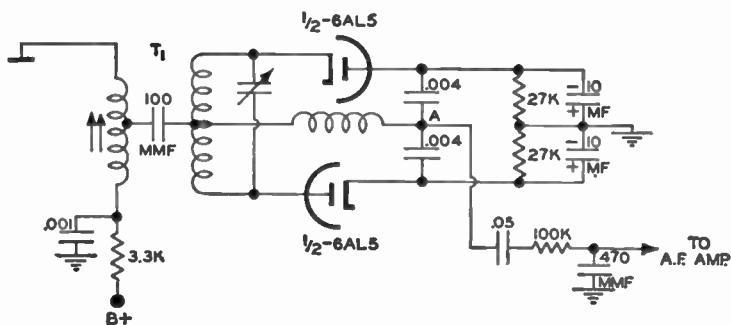
The second circuit of Fig. 14-21 is a slight variation of the first circuit. The principal difference lies in the manner in which the ratio detector transformer T_1 supplies the reference voltage against which the induced secondary voltages are compared. The $100\text{-}\mu\mu\text{f}$ capacitor and the R.F. choke (the latter connected between the center-tap of the secondary of T_1 and point A) are similar in their action to C_3 and the R.F. choke of Fig. 14-15.

There is one further circuit found in F-M receivers that is not used in A-M sets — the so-called “de-emphasizing filter.” It was required because the greatest amount of audio-frequency noise is generated in the transmitter at the higher frequencies, from 5 kc up. To reduce the effect of the noise, a pre-emphasis network is inserted in the audio system of the transmitter. The function of the circuit is to favor the frequencies above 1500 cycles. It accomplishes this by proportionately attenuating the lower frequencies more than the higher frequencies of the signals passing through the network. A typical accentuator filter is shown in Fig. 14-22A. The higher frequencies, in passing through the network, lose less voltage than the accompanying low frequencies.

Pre-emphasis is applied to the audio signals at the first audio amplifier. Beyond this network, the audio voltages combine in the usual manner with whatever noise is present in the system. At the receiver there is a de-



(A)



(B)

FIG. 14.21. Two commonly used variations of the balanced ratio detector.

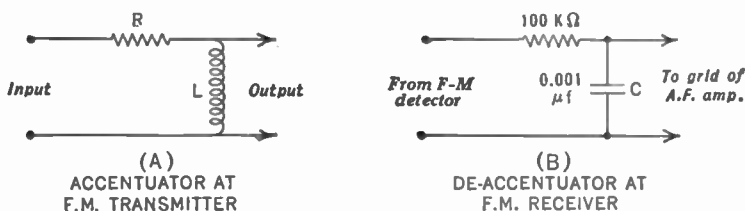


FIG. 14.22. Two circuits that help maintain the fidelity of the higher audio frequencies in F-M systems.

emphasis circuit (Fig. 14-22B) with the reverse properties of the pre-emphasis circuit. The frequencies above 1500 cycles are reduced to their original values. At the same time a similar reduction in noise occurs.

The overall effect is a return of the signal to its proper relative proportions, but with a considerable reduction in noise.

The audio amplifiers that follow the discriminator are engineered for flat response up to 15,000 cycles. This frequency response extension is easier to achieve here than in video amplifiers, and little change in basic amplifier form will be noticed. A single or push-pull power amplifier provides sufficient power to drive 12-inch speakers. For the home, this permits adequate volume.

The Audio System in Television Receivers. The sound signal, once it has been divorced from the video signal, is passed through one or more of its own I.F. amplifiers before it reaches the F-M detector. At the detector output, the resultant audio voltage is further amplified until it is sufficiently strong to drive a loudspeaker. The elaborateness of the final audio system will be governed, in large measure, by the price of the receiver. In table model receivers, 12-inch speakers are rare, and the audio system is patterned closely after the current A-M midget sets. When the unit is housed in a console cabinet, a larger speaker is used, with a subsequent rise in the quality of the audio output.

Servicemen who are not familiar with F-M circuits may be inclined to believe that F-M, I.F. amplifiers differ, radically, from A-M, I.F. stages. Actually, this is not so. From a servicing standpoint, the F-M, I.F. system of a television receiver differs in only two respects from its A-M counterpart — in the frequency of operation and in the bandpass of its tuned circuits. The center frequency is 20 mc or more, and the bandpass is between 150 kc and 350 kc. In all other respects A-M and F-M systems are similar, and knowledge of one system may be transferred to the other.

The television F-M signal, when fully modulated, occupies a bandwidth of approximately 50 kc (plus or minus 25 kc about the carrier). The audio bandpass, however, is designed for a greater frequency variation. This is done in conventional-type receivers to minimize audio distortion arising from frequency drift of the local oscillator. At 100 mc, a drift in frequency of 0.1 per cent produces a change of 100 kc, which is more than sufficient to throw the audio signal completely out of the bandpass of the audio I.F. amplifiers. However, with normal frequency drift and a fine-tuning control at the oscillator, the audio signal can be kept within the circuit frequency boundaries. In this respect, it is important for the serviceman to realize that the fine-tuning control in conventional-type sets is adjusted until the sound comes through, clear and undistorted. When this is accomplished, if detail does not appear as sharp as desired, little can be done to correct this except realignment of the video I.F. circuits. The fine-tuning control here is incorporated primarily for the audio signal.

In Intercarrier receivers, the sound I.F. frequency is 4.5 mc, and this

remains constant regardless of the setting of the fine-tuning control. Hence, in these receivers, the control is adjusted for best image and, if the set is operating properly, the sound is best here, too. You do *not* tune for loudest sound in Inter-carrier receivers because so much sound signal would reach the video second detector that the video signal would be affected. It will be recalled from Chapter 11 that the sound voltage at the video detector must be kept small.

The number and type of sound I.F. amplifiers will depend upon the F-M detector used and the circuitry of the TV receiver itself. In conventional TV receivers, the sound signal is separated from the video signal early, and in these sets you will find two to three sound I.F. stages.

In Inter-carrier sets, the sound signal requires less amplification after separation because it has penetrated more deeply into the video system. Thus, in these sets, one or at most two sound I.F. stages will be found.

In common use are two F-M detectors, the Foster-Seeley discriminator and the ratio detector. The latter further appears in two forms, balanced and unbalanced. The Foster-Seeley circuit, while essentially an F-M demodulator, will, at the same time, also respond to A-M. Consequently, one of the I.F. stages preceding this detector must be operated as a full limiter. For the ratio detector, either in balanced or unbalanced form, complete limiting is unnecessary because the detector is fairly unresponsive to A-M over wide ranges of signal amplitude. Consequently, in a system employing the latter detector, every I.F. stage will be an amplifier, although partial limiting is quite common in the last I.F. stage in order to maintain a fairly constant input to the detector. Engineering tolerances in ratio detector transformers are often such as to require this limiting.

A television sound system of a conventional-type set employing a Foster-Seeley discriminator is shown in Fig. 14-23. The first and second stages have single iron-slug tuned coils, which resonate with a fixed capacitor of 20 $\mu\mu\text{f}$ and the tube capacitances to the audio intermediate frequency of 21.9 mc. Shunting resistors are not required for these coils because at 21.9 mc a bandpass of several hundred kilocycles can be readily achieved by regulating the Q of the coil. The third stage is a limiter, using grid-leak bias and low plate and screen voltages to cause the tube to saturate readily. The limiter grid resistor is kept small, so that the grid time constant will be small. In this way, the effect of sharp pulses of interference (generally from ignition systems) are reduced or eliminated.

Figure 14-24 illustrates a television sound system using a ratio detector. This is also from a conventional-type set.

Very similar to the foregoing is the sound system, shown in Fig. 14-25, from an Inter-carrier set with the sound I.F. value at 4.5 mc. Other than the sound I.F. value, there is no basic difference between all these sound systems.

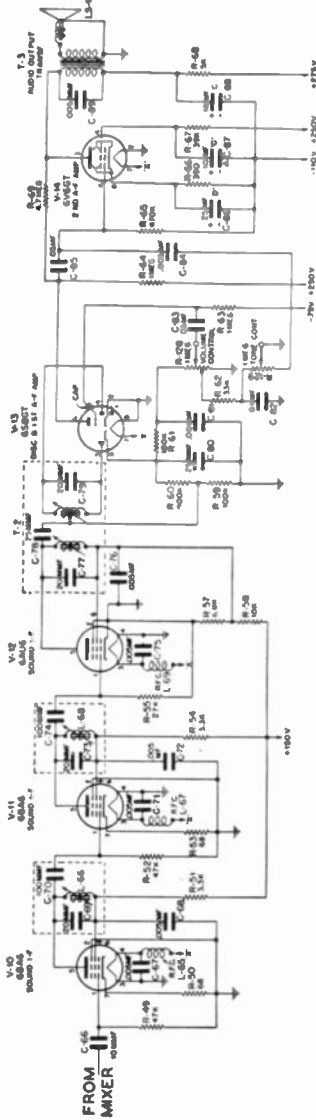


Fig. 14.23. The complete audio system in a television receiver using a Foster-Seeley F-M detector. Courtesy Motorola

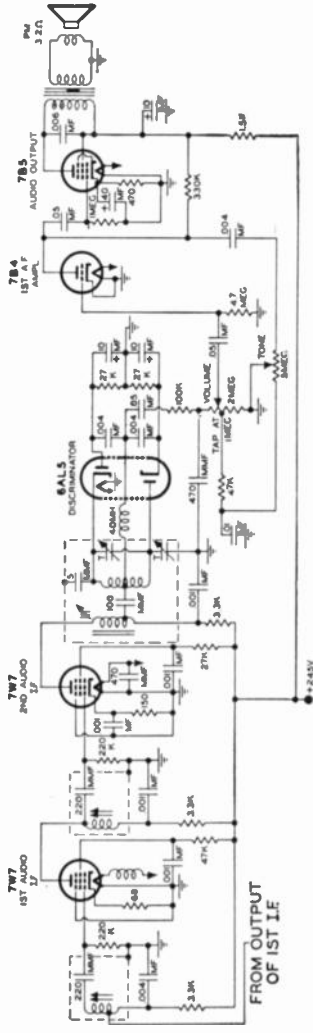


Fig. 14.24. An audio system using a balanced ratio detector. Courtesy Philco

Mention might also be made of the 6BN6 beam-gated tube which is used by some manufacturers as a combined limiter-discriminator. The reader will recall that the operation of this tube was discussed in Chapter 12 in connection with its use as a sync separator.

A typical circuit of the 6BN6 as a limiter-discriminator is shown in Fig. 14-26. Incoming sound I.F. signals are applied to control grid 1. If

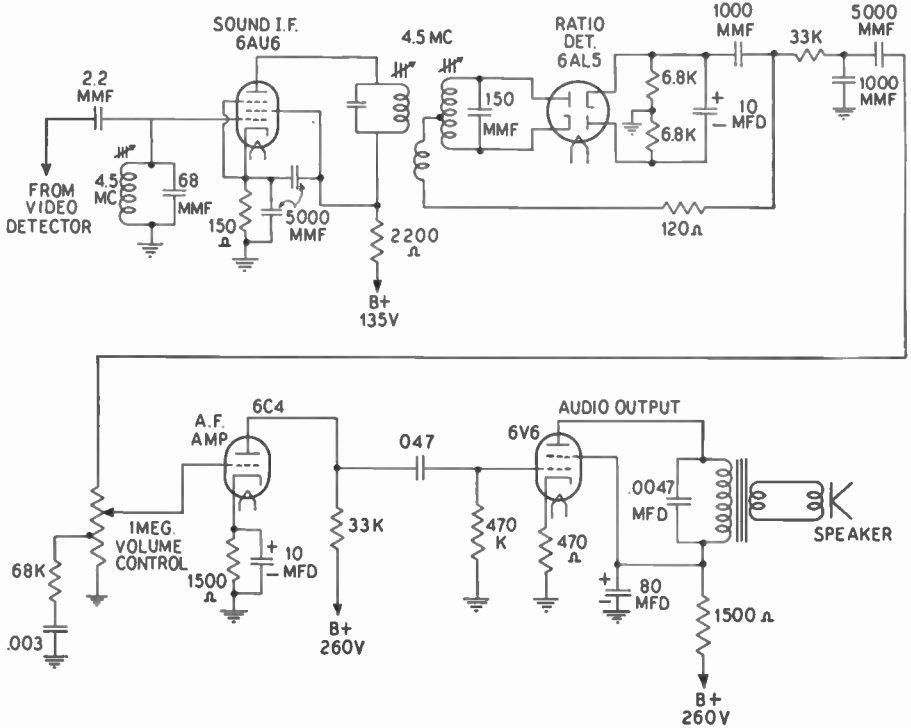


FIG. 14.25. A 4.5-mc I.F. sound system in an Intercarrier receiver.

the signal has received sufficient prior amplification, it will have a peak-to-peak value of several volts. Upon application to grid 1, current through the tube will start to flow only during the positive part of the cycle and will remain essentially constant no matter how positive the signal may become or what amplitude variations it may contain. Thus, signal limiting is achieved in this section of the tube, the electron beam being passed during the positive half-periods of the applied signal and cut-off occurring during negative half-periods. The groups of electrons that are passed then travel through the second accelerator slot and form a periodically varying space charge in front of grid 3. By electrostatic induction, currents are made to flow in the grid wires. A resonant circuit is connected between this grid and ground, and

a corresponding voltage of approximately 5 volts is developed at grid 3. The phase of this voltage is such that it will lag the input voltage on grid 1 by 90°, assuming that the resonant circuit is tuned to the intermediate frequency. (Because of this 90° difference between grid voltages, grid 3 is often referred to as the quadrature grid.)

Electrostatic induction, while it may be new to the reader, is an effect which finds wide application in all vacuum tubes.

Whenever a group of electrons approach an element in a tube, electrons

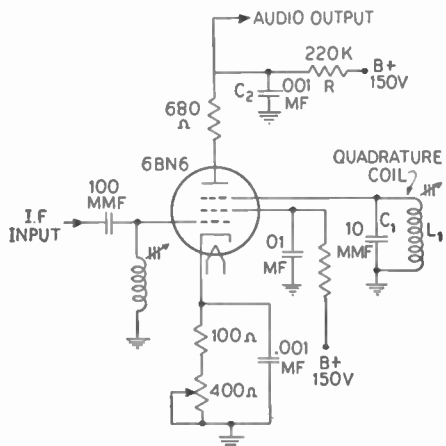


FIG. 14.26. The beam-gated tube connected as a limiter-discriminator.

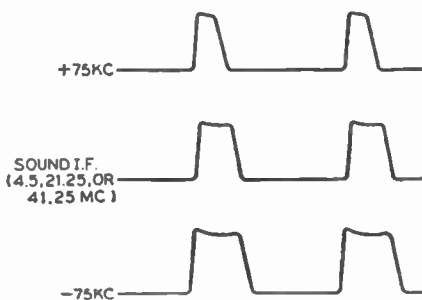


FIG. 14.27. The variation in current pulse duration with different incoming frequencies in a 6BN6.

on the surface of that element will be repelled, resulting in a minute flow of current. By the same token, electrons receding from an element will permit the displaced electrons to return to their previous positions. Again, a minute flow of current results, this time in a direction opposite to that of the first flow. If a sufficient charge periodically approaches and recedes from an element, the induced current can be made substantial. This is precisely what occurs at grid 3 in the 6BN6.

In the gated-beam tube, grids 1 and 3 represent electron gates. (Grid 2 has a fixed positive voltage for electron acceleration.) When both are open, current passes through the tube. When either one is closed, there is no current flow. In the present instance, the second gate lags behind the first. Plate-current flow starts with the delayed opening of the second gate and ends with the closing of the first gate. Now, when the incoming signal is unmodulated, and L_1C_1 of Fig. 14-26 is resonated at the I.F. frequency, the voltage on grid 3 will lag the voltage on grid 1 by 90°. However, when the incoming signal is varying in frequency, the phase lag between the two grid voltages will likewise vary. This, in turn, varies the length

of the period during which plate current can flow. (See Fig. 14-27.) Thus, plate current varies with frequency, and the circuit is designed so that the current varies in a linear manner. By placing a resistor in the plate lead, R of Fig. 14-26, we can obtain an audio voltage to feed the audio amplifiers that follow. A typical discriminator response for an F-M receiver with a 10.7-mc center frequency is shown in Fig. 14-28.

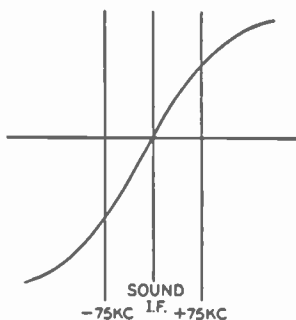


FIG. 14.28. Discriminator response of the 6BN6 when connected as shown in Fig. 14.26.

In the circuit of Fig. 14-26, a 680-ohm resistor is inserted between the load R and the plate of the tube. By-passing of the I.F. voltage is accomplished by C_2 , but since this capacitor is placed beyond the 680-ohm resistor, a small I.F. voltage appears at the anode of the tube. Through the interelectrode capacitance that exists between the anode and grid 3, the I.F. voltage developed across the 680-ohm resistor is coupled into L_1C_1 . The phase relations existing in this circuit are such that this feedback voltage aids in driving the tuned circuit.

Bias for control grids 1 and 3 is obtained by placing a resistor in the cathode leg of the tube. Since A-M rejection, especially at low input signals near the limiting level, is a function of the correct cathode bias, the cathode resistor is made variable. This permits adjustments to be made in the field in order to compensate for tube or component changes.

QUESTIONS

1. In an A-M wave, where is the intelligence contained? How does this differ from the conditions prevailing in frequency modulation?
2. What is a discriminator? What is the accepted application of this word?
3. What influence does the audio-modulating signal frequency have in the production of an F-M signal?
4. Where do the F-M sidebands obtain their power? Contrast this with the situation existing in an A-M signal.
5. Differentiate between phase modulation, amplitude modulation, and frequency modulation.
6. What is the significance of the 2 to 1 ratio in F-M reception?
7. Why is oscillator stability so important in a high frequency receiver?
8. What purpose do the I.F. amplifiers serve in a superheterodyne?
9. What advantages are obtained through the use of an I.F. amplifier?
10. Why are limiters necessary in F-M receivers?
11. Do all F-M receivers require limiters? Explain.
12. What advantage do double limiters possess over single limiters?
13. Explain the operation of a grid-leak bias limiter.
14. In what other circuits in radio do we find grid-leak bias?
15. What effect does the use of lowered voltages have on tube operation?

16. Contrast the purpose of an A-M detector with an F-M detector. State clearly why each should differ.

17. Draw the schematic diagram of an early type of discriminator which employed two secondary windings.

18. Explain the operation of the circuit drawn for Question 17.

19. Would the foregoing discriminator function if one of the diodes became inoperative? Give the reasons for your answer.

20. Draw the circuit of a modified discriminator (Foster-Seeley type) widely used today.

21. Draw the circuit of a ratio detector.

22. Explain briefly the operation of a ratio detector.

23. What is the Armstrong System of generating frequency modulation? Why is it useful?

24. Draw a block diagram of the basic Armstrong System. Explain the function of each stage.

25. What properties of a 6BN6 enable it to be used as a limiter? In what other section of a television receiver could these same properties be put to use?

Chapter 15

TYPICAL TELEVISION RECEIVERS — ANALYSIS AND ALIGNMENT

Television Receiver Cabinets. Television home units may contain not only the television receiver itself, but may also incorporate additional A-M and F-M chassis to take care of the other broadcast services available to the public. Usually both A-M and F-M signals are handled by a single combination circuit which receives the sound signals, converts them down to suitable intermediate frequencies, amplifies them, and then detects them. After this, the demodulated voltages are transferred to the audio section of the television receiver where they subsequently reach the loudspeaker.

With the present methods of building midget sound receivers, the A-M, F-M chassis fits easily in one corner of the cabinet. And last, but not least, is the record player, a welcome addition to any receiver. These combinations, which are certain to be continued, offer products having good sales appeal to the public. It is much easier (and neater) for the home to have one general-purpose receiver, than several separate units scattered about, each for a specific purpose. The popularity of the modern multi-purpose meter in the radio laboratory indicates that the technician feels the same way about his laboratory.

Receiver Panel Controls. The front panel controls that are associated with television receivers fall into two categories: those dealing with the audio section of the receiver, and those associated with the video chassis. Straddling both groups would probably be the regular tuning control and the fine-tuning control, adjusting the reception of both signals simultaneously. Of the audio controls, little need be added to what is already known. One volume and one tone control permit all the variation ordinarily desired, even with the greater audio range available with F-M.

For adjustment of the video portion of the television receiver, the number of front panel controls varies with each manufacturer. Some manufacturers provide only contrast and brilliancy controls; others include the hold con-

trols. All try to keep the number of front panel controls to a minimum; anything that will simplify the operation of the receiver for the layman is desirable and certain to make the television receiver a more popular instrument.

a. Contrast. The most important video control on the front panel is the contrast control. This adjustment on a television set is similar in its action to the volume control on a sound receiver. The contrast potentiometer varies the amplification that the video signal receives. The greater the strength of the video voltage applied to the control grid of the cathode-ray tube, the more intense will be the image on the screen. A high setting might be desirable if the surrounding light is strong, or at least bright enough to interfere with the clear perception of an ordinary lighted screen.

In some receivers, the contrast control does not vary the output voltage directly. Rather it accomplishes this by controlling the gain of several tubes throughout the set. If the control is advanced too far, the regulated tubes are placed on a portion of the characteristic curve that is not linear. The result is distortion. It is well known that the human eye is more critical of distortion than the ear. In sound receivers, distortion percentages can run as high as 15 per cent and still be tolerated. In the case of the eye, this amount would prove unsatisfactory.

b. Brilliancy Control. Another panel adjustment for the video section of the television receiver is the "brilliancy or brightness control." This works in close conjunction with the contrast control, and it should not be set until the contrast potentiometer has been set to the desired point. It will be recalled that the brilliancy potentiometer varies the bias on the grid of the cathode-ray tube and, in this manner, the operating point of the tube. With a strong video signal, it seems reasonable that the bias on the grid of the viewing tube should be different than when the incoming signal is weaker. The correct setting of the brilliancy control will just bias the cathode-ray tube so that the blanking voltage level causes cut-off of the electron beam. Refer to Chapter 8. Too low a setting causes the picture to appear washed out; too high a setting eliminates some of the darker detail of the image.

c. Hold Control. The hold controls, one for the vertical system and one for the horizontal system, enable the viewer to bring the picture back into sync should the occasion arise. Nearly all sets manufactured today possess automatic frequency control of the horizontal oscillator. When this is functioning properly there should be little need for adjustment of the horizontal hold control. If the manufacturer feels that the horizontal system is sufficiently stable to require only occasional adjustment, then he may place the horizontal hold control on the rear panel. Another favorite spot is behind a small, hinged panel on the front of the cabinet. However, if the manufacturer feels that the hold control will be needed more than just occasionally, he may place it on the front panel.

The vertical system, because of the integrating network that precedes it, is less susceptible to disturbances than a free-running horizontal system. The vertical oscillator may fall out of sync momentarily, causing the picture to roll, but it usually locks itself in again. As a general practice, the vertical hold control is placed near the horizontal hold control.

d. Fine Tuning. In addition to the controls just described for the video chassis, and the two for the audio chassis, we find another knob labeled "fine tuning." This is a vernier variable capacitor connected across the oscillator tuning circuit. With it, small variations in the oscillator frequency can be compensated. This control is especially necessary at the high frequencies where small percentage variations mean greater frequency changes. The regular tuning is accomplished by means of push buttons or a selector switch. (With continuous tuning, no fine-tuning control is needed.)

Secondary Controls. The other variable controls associated with the video chassis of the television receiver are placed within easy reach at the rear of the set. They are placed there only for the convenience of the serviceman and are generally not to be touched by the ordinary user of the set. A list of these rear secondary controls includes the following:

1. Vertical linearity.
2. Horizontal linearity.
3. Vertical centering (if used).
4. Horizontal centering.
5. Vertical size. (Height Control.)
6. Horizontal size. (Also known as Width or Drive Control.)

The exact number found, of course, will depend upon the elaborateness of the receiver design. No discussion of the action of the controls is deemed necessary here as it would merely be a repetition of what has been given in previous chapters.

A Modern Television Receiver. In view of the fact that there is in use today both conventional-type and Intercarrier television receivers, an analysis of a typical receiver in each category will be given. Since the conventional-type circuit appeared first, it will be considered first. An interesting example of such a television receiver is the circuit shown in Fig. 15-1. This is a 30-tube, direct-viewing television receiver. Twelve channels are covered, with an F-M sound system. The various oscillator frequencies for each of the twelve channels are as follows:

<i>Channel Number</i>	<i>Channel Freq. Mc</i>	<i>Picture Carrier Freq. Mc</i>	<i>Sound Carrier Freq. Mc</i>	<i>Receiver R.F. Osc. Freq. Mc</i>
2	54-60	55.25	59.75	81
3	60-66	61.25	65.75	87
4	66-72	67.25	71.75	93
5	76-82	77.25	81.75	103
6	82-88	83.25	87.75	109
7	174-180	175.25	179.75	201
8	180-186	181.25	185.75	207
9	186-192	187.25	191.75	213
10	192-198	193.25	197.75	219
11	198-204	199.25	203.75	225
12	204-210	205.25	209.75	231
13	210-216	211.25	215.75	237

Operating controls, those on the front panel, are (see Fig. 15-2):

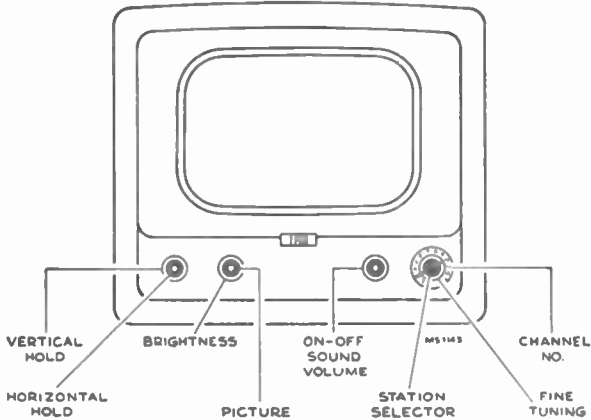


FIG. 15.2. The front panel operating controls of the television receiver in Fig. 15.1.

- Channel selector }
- Fine tuning }
- Picture (contrast) }
- Brightness }
- Horizontal hold }
- Vertical hold }
- Sound volume and on-off switch..... Single knob

The non-operating controls, those which are mounted on the rear of the chassis, include:

- Vertical centering
- Width
- Height
- Horizontal linearity
- Vertical linearity
- Horizontal drive
- Horizontal oscillator frequency
- Horizontal oscillator phase
- Focus

The purpose and use of each of the controls will become evident as the analysis of the circuit proceeds.

A block diagram of the receiver is shown in Fig. 15-3. The full schematic is shown in Fig. 15-1. In the explanations that follow, it is recommended that both diagrams be consulted.

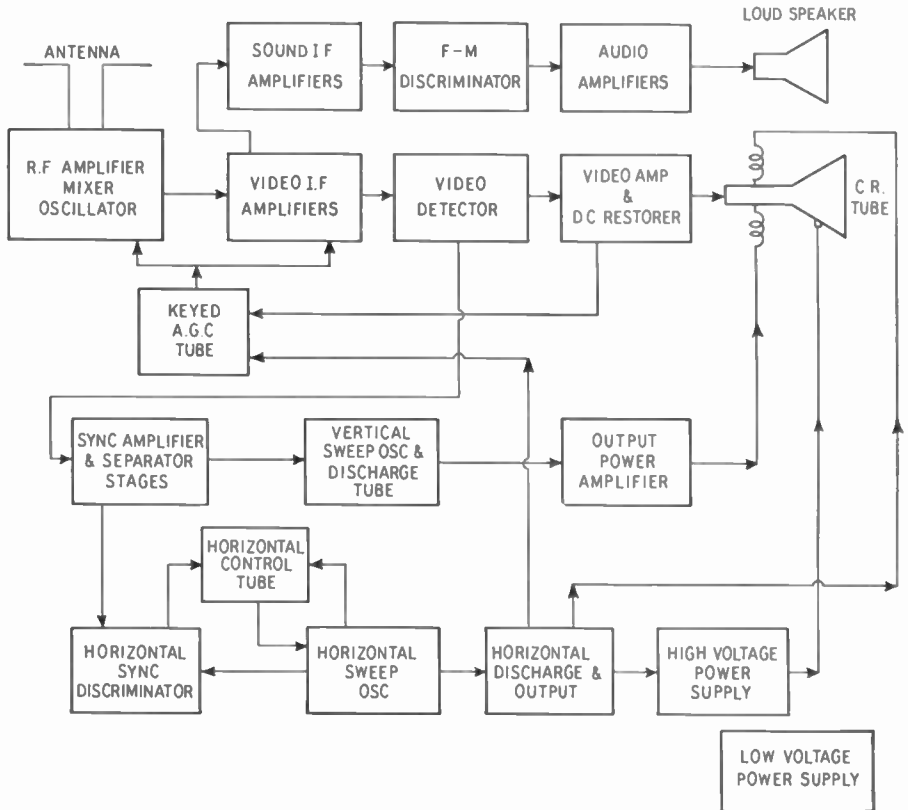


Fig. 15.3. A block diagram of the television receiver shown in Fig. 15.1.

The tuner employed in this receiver is a cascode Standard Coil tuner. The two triode sections of the 6BK7 form the cascode R.F. amplifier, while the 6J6 has its triodes functioning as the oscillator and mixer, respectively. Operation of this tuner was previously covered in Chapter 4, and the reader is referred there for further details.

Output of the mixer is fed into a video I.F. system consisting of four 6CB6 amplifiers stagger-tuned. The various I.F. tuning coils are peaked to the following frequencies:

L_{111}	21.8 mc
T_{108}	25.3 mc
T_{104}	22.3 mc
L_{183}	25.2 mc
L_{188}	23.4 mc

In addition, there are three trap or attenuation circuits provided. One is for the sound take-off, and it is coupled to the plate coil of the second video I.F. amplifier. A second trap circuit is coupled to the plate coil of the first video I.F. amplifier. This is for the sound signal of the next lower adjacent channel. The final trap is located in the cathode leg of the fourth video I.F. amplifier. This circuit is to attenuate whatever sound signal (of the same channel) still remains in the system.

The gain of the first two video I.F. amplifiers is controlled automatically by a keyed A.G.C. system. The level of the bias is established by the strength of the incoming signal and partially by the setting of the A.G.C. level control. More on this presently. The final two video I.F. amplifiers are cathode-biased in the conventional manner of a class A amplifier. Omission of the by-pass capacitor across R_{116} , R_{121} , and R_{126} and the use of a small 82- $\mu\mu\text{f}$ capacitor across R_{133} tend to stabilize amplifier operation by degeneration.

The Sound System. As noted, the separation of the sound and video signals takes place in the plate circuit of the second video I.F. amplifier. Thereafter the video signal travels toward the picture tube while the sound signal moves toward the speaker.

The sound system contains three I.F. amplifiers, a Foster-Sceley type of discriminator, and two stages of audio amplification. The sound discriminator bandwidth between peaks is 350 kc. The signal is received by a 6BA6 amplifier which is functioning as a Class A amplifier. The second I.F. amplifier, also a 6BA6 tube, is operating as a partial limiter, using a combination of grid-leak and cathode bias. The limiting action is completed in the third I.F. amplifier, utilizing grid-leak bias, which has a very short time constant. The grid-leak combination is composed of 22,000-ohm grid resistor and a 51- $\mu\mu\text{f}$ grid capacitor. There is no cathode bias.

The operation of an F-M system depends upon the correct adjustment of the I.F. and discriminator stages. The proper alignment procedure for the present system will be given later in this chapter. The operation of the limiter and discriminator stages was examined in detail in Chapter 14.

The output of the discriminator consists of the audio frequencies, as broadcast at the studio. These are transferred, via the volume control, to the audio amplifiers. Both audio stages are entirely conventional in their construction, acting merely to strengthen the audio signals until they are strong enough to operate the loudspeaker. The audio system contains only a vol-

ume control, there being no provision made for tone control. There is, however, a tone-compensating network placed across the volume control. The speaker will deliver $2\frac{1}{2}$ watts of undistorted power and a maximum of 4 watts with a tolerable distortion.

Video Second Detector. The video second detector, a diode, is connected to produce a positively phased signal. In this form, the synchronizing pulses are the most negative and the brightest portions of the image are the most positive. This type of output voltage is necessary because there are an even number of video amplifiers between the detector and the cathode-ray tube. The diode load resistor R_{137} is 3,900 ohms. The circuit is fully compensated, employing both series and shunt peaking, with L_{187} and L_{188} . The 22,000-ohm resistor across L_{187} is to prevent excessive peaking should the self-inductance and self-capacitance in the coil resonate within the video signal range. The response of this network decreases rapidly above 4 mc and prevents the video I.F. from reaching the video amplifiers.

A portion of the video signal is obtained from R_{137} and applied to V_{118} , the first sync amplifier. L_{103} and R_{283} serve to decrease the loading effect of V_{118} on R_{137} , especially the input capacitance of the sync amplifier tube. L_{103} is also beneficial in maintaining the steep sides of the sync pulses.

Video Amplifiers. Two stages of video amplification increase the strength of the signal until it is capable of fully modulating the cathode-ray beam. Each stage is designed with high- and low-frequency compensation. The high-frequency compensating components are the series and shunt coils, such as L_{189} , L_{190} , L_{191} , and L_{192} , etc. The low-frequency compensating components are the decoupling networks between the lower end of the load resistors and the power supply: for example, C_{223B} (10- μ f electrolytic capacitor) and R_{141} (6,800-ohm dropping resistor) in the plate circuit of V_{115} .

The first video amplifier is d-c coupled to the video detector. This is required because the keyed A.G.C. tube obtains its video signal from the plate of this amplifier. By using d-c coupling between the video detector and the amplifier we maintain all sync pulses at the same level. This, it will be recalled, was necessary for the proper operation of the keyed A.G.C. circuit.

Since the signal output of the video detector is negative, an average negative potential will be established here. This will bias the 6AU6 and permit the tube to operate properly with its cathode grounded.

One further component of interest in the first video amplifier is the 4.5-mc trap. This is formed by L_{308} together with a small fixed capacitor and a 10,000-ohm resistor (R_{277}). Its purpose is to absorb any 4.5-mc voltage that may have been developed in the video detector by the beating of the sound and video carriers. All of the sound signal should have been absorbed by the sound traps in the video system. However, this may not always occur completely, and L_{308} is inserted as another safeguard.

The second video amplifier, V_{116} , uses a 6V6 power output tube. Bias for this tube is established in the grid circuit by direct connection to the negative side of the power supply. In the cathode circuit, bias is developed by tube current flow through R_{146} , R_{262} , and R_{131} and by returning R_{144} to the -100 -volt tap of the power supply. Note that most of this -100 volts is dropped across R_{144} , and this portion will not affect the grid-to-cathode bias of the 6V6. In fact, the strong tube currents flowing through R_{146} , R_{262} , and R_{131} ordinarily counteract whatever negative power supply voltage appears across these three resistors and establishes a net positive voltage. However, by varying R_{131} , we can use the negative supply voltage to offset part or all of this positive cathode voltage and in this way regulate the stage gain. R_{131} is thus the contrast control.

D-c Restorer. R - C coupling between the 6AU6 and the 6V6 and between the 6V6 and the picture tube will not pass the d-c component of the detected video signal. Hence, d-c restoration is desirable. The method employed in this receiver is almost identical with that described in Chapter 8 and requires little additional explanation. R_{148} applies the video signal to V_{114B} and also serves as the d-c path by which the d-c restorer feeds its correcting voltage to the grid of the picture tube.

Synchronizing Amplifier and Separator. The input to this section of the receiver is obtained from the load resistor of the video second detector. The synchronizing pulses, at the grid of V_{118} , extend in the negative direction. As a first step toward effecting clear-cut sync separation, the entire signal is amplified by V_{118} . The next tube, a 6AU6, is biased to cut-off, and only voltages which are going positive will produce a flow of current in its plate circuit. Since the 6AC7 inverts the incoming negative pulses, these are most positive at the grid of the 6AU6. The remainder of the signal, however, is less positive or negative and hence is removed. The final stage (V_{120A}) in this system tends further to square off the sync pulses.

At the output of the second sync amplifier, the vertical and horizontal pulses must be separated from each other and fed to their respective systems. The horizontal pulse is of very short duration (5 microseconds); the vertical pulse lasts for the relatively long interval of 190 microseconds. The low-pass integrating network consisting of R_{163} , R_{164} , R_{165} , C_{151} , C_{152} , and C_{153} will by-pass the quickly rising and falling horizontal pulse. The vertical pulse, on the other hand, will pass through the network and reach the grid of the 6J5 vertical blocking oscillator and discharge tube.

During the negative portion of the cycle, the grid of the 6J5 is held beyond cut-off and C_{158} is charging through R_{169} and R_{170} . When the synchronizing pulse arrives, it drives V_{121} into conduction, and C_{158} discharges through the secondary winding of T_{106} and the 6J5 tube. The current in T_{106} induces a positive voltage on the grid of the 6J5, which further reduces the plate re-

sistance of the tube and permits C_{158} to discharge even faster. The sequence follows the pattern of operation of all such blocking oscillators, and the waveform developed across C_{158} and R_{174} is shown in Fig. 15-4. This type of waveform is obtained from peaking circuits (C_{158} in series with R_{174}) and, when applied to a coil, will produce a saw-tooth current. Adjustment of R_{169} will vary the amplitude of the deflection voltage and is, of course, the height control.

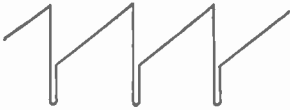


FIG. 15.4. The shape of the deflection voltage wave at the input to the vertical output amplifier.

The circuit, in its present form, appears to bear no resemblance to charge and discharge circuits previously discussed in Chapter 12. However, by a rearrangement of the components, the similarity is readily apparent (see Fig. 15-5).

To develop sufficient driving power, the output of the 6J5 is applied between the grid and cathode of the 6S4. In the cathode leg of this amplifier, a variable resistor functions as the vertical linearity control. Variation of resistance in the cathode leg has the effect of producing slight variations in the shape of the saw-tooth wave by shifting the operating point of the tube. We

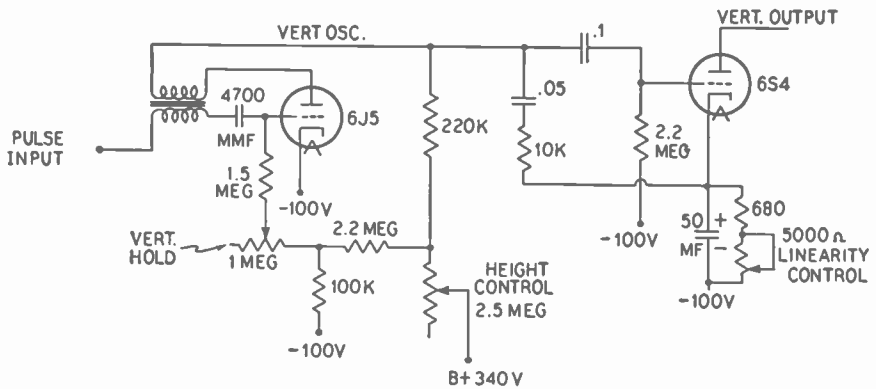


FIG. 15.5. The vertical oscillator and discharge circuit, rearranged to more conventional form.

can alter or "distort" the shape of the saw-tooth wave to a form which will produce a more linear motion of the cathode-ray beam.

In shifting the operating point of the 6S4 by the linearity control, we also vary the gain of the tube. This will affect the height of the image. Accordingly, whenever the linearity control is adjusted, we must also adjust the height control. Conversely, adjustments of the height control affect the shape of the saw-tooth voltage on the V_{121} plate and require adjustment of the linearity control. The 6S4 is matched to the vertical deflecting coils by

means of T_{107} . The vertical coils are sufficiently damped (by means of the two 560-ohm shunting resistors) to prevent any shock-excited oscillations from existing in the coils for any length of time. A centering control provides a means for sending a d-c current through the coils, in either direction for centering the beam on the screen.

Leading off the secondary of T_{107} is a resistor-capacitor network which connects, at its other end, to the cathode of the picture tube. The purpose of this circuit is to transmit positive vertical pulses to the cathode of the picture tube during the vertical retrace interval. A positive pulse at the cathode is equivalent to a negative pulse at the grid and, if these pulses are strong enough, they will prevent the electron beam from reaching the picture screen during this interval. This will remove the vertical retrace lines which, under weak signal conditions, can prove annoying to the viewer.

During vertical retrace, a large positive pulse appears across the secondary of T_{107} . (C_{242} , R_{278} , R_{279} , R_{280} , C_{243} , and C_{244} apply this pulse to the picture tube cathode.

Horizontal-Sweep Oscillator System. The horizontal-sweep oscillator system contains an automatic frequency and phase control which frames the picture and also possesses the desirable noise immunity, lack of which has proven so destructive in designs employing the usual trigger-type of scanning oscillator. In this circuit, a stable Hartley oscillator is set at 15,750 cycles per second. Coupled to the oscillator is a synchronizing discriminator circuit which receives the pulses of the incoming television signal and compares the frequency and phase of these synchronizing pulses with the generated sine wave. Any slight variations between the two will produce a d-c voltage which is applied to the grid of a reactance tube. The plate circuit of this reactance tube is connected directly across the tuning circuit of the Hartley oscillator, and, as the d-c voltage at the grid of the reactance tube varies, it will alter the plate current and with it the oscillator frequency.

The Hartley oscillator is V_{125} (6K6) and it is conventional in form. The oscillator coil (secondary of T_{108}) is closely coupled to the primary winding. The primary winding is center-tapped and tuned by means of C_{108} . Since each diode is connected across half of the primary coil T_{108} , each receives voltages which are equal in amplitude but opposite in phase. The operation of this system was described in Chapter 13 and the reader is referred there for a step-by-step analysis.

Horizontal Discharge. The horizontal discharge tube follows the horizontal oscillator and produces, in its plate circuit, the peaked deflecting voltage which will drive the horizontal deflecting coils. In V_{125} , the Hartley oscillator is connected between the screen grid and the cathode. The plate circuit contains only a resistance, and consequently the waveform found here is not the symmetrical sine wave that would be produced by a resonant cir-

cuit; rather it has the form shown in Fig. 15-6. The peak-to-peak voltage on the grid of V_{125} is approximately 130 volts. This grid swing produces a square wave in the plate circuit, with a peak-to-peak voltage of 225 volts. The square wave is differentiated by C_{179} and R_{202} , producing the wave shown in Fig. 15-7. The positive portion of the differentiated wave is sufficiently sharp to trigger the discharge tube. The discharge tube is normally cut-off due to bias produced by grid rectification of these pulses from the oscillator. The positive pip of the pulse overcomes this bias and drives the tube into heavy momentary conduction. During this period, the plate voltage of V_{120B} falls to cathode potential, and C_{179} discharges rapidly. However, the conduction period is quite short, and C_{179} does not discharge completely because

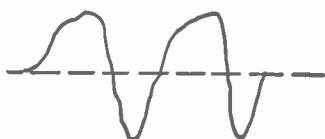


FIG. 15.6 Plate of horizontal oscillator 225 volts peak-to-peak.



FIG. 15.7 Input to horizontal discharge tube 100 volts peak-to-peak.

of R_{241} and R_{242} , both of which are in series with C_{179} . When V_{120B} becomes non-conducting again, its plate voltage rises quickly to a value determined by the charge remaining on C_{179} . From this point the plate voltage rises slowly, and we get the desired waveform. This is transferred to the horizontal output tube (V_{126}) via C_{178} .

Horizontal-Deflection and High-Voltage Circuits. The horizontal output tube is a 6BQ6-G power amplifier. The deflection voltage it receives from V_{120B} is amplified and then applied to the horizontal output autotransformer, T_{109} . Since this type of circuit was previously discussed in Chapter 13, an extended discussion is not required here. The horizontal deflection yoke connects effectively between terminals 5 and 3. B+ from the power supply is applied to this circuit via a 200-ohm and a 400-ohm resistor to the plate of the 6W4 damper tube. Added to this voltage is the "boost B+" which is developed across C_{242} by the damper tube. It is the sum of these two voltages which is applied to the plate of the 6BQ6-G.

A special horizontal linearity coil (L_{0R1}) is inserted between the top and bottom windings of the horizontal output autotransformer. The purpose of this coil is to keep the damper tube from conducting except just after beam retrace has been completed. It does this by keeping the cathode of the 6W4 positive with respect to its plate. However, after beam retrace has been completed, the oscillating voltage drives the cathode negative and the damper tube begins to conduct, absorbing the energy still remaining in the deflec-

tion yoke circuit. As noted previously on p. 331, the damper tube continues to absorb energy until the beam is about 40 per cent of the way across the screen. By this time the energy is nearly all gone, and the horizontal output tube swings into action.

Varying the inductance of the linearity coil, by means of a slug adjustment, will alter the length of time the damper tube conducts. This, in turn, will be found to affect picture linearity.

There are two adjustments for the picture width, R_{242} and L_{310} . Both are necessary because, although their adjustment alters the width of the image, they perform other functions. R_{187} determines the ratio of high peaking and saw-tooth voltage on the grid of the output tube. This affects the point on the trace at which the tube conducts. Clockwise rotation of the control increases picture width, crowds the right side of the picture and stretches the left side. L_{310} varies the output and hence the picture width by shunting a portion of the secondary winding of T_{100} . Clockwise rotation of the control increases the picture width and causes the right side of the picture to stretch slightly. Both controls must be adjusted whenever the position of either one is changed.

From the H.V. rectifier (1B3-GT), up to 16,000 volts are taken, and applied to the second-anode aquadag coating of the cathode-ray tube. Very little filtering of this high voltage is necessary since the frequency of its ripple is high (15,750 cycles). The filter capacitance is a small 500- μmf capacitor. The stored energy is small, and the high voltage is prevented from proving fatal in most instances. The remaining potentials required by the cathode-ray tube are low and are supplied either by the low-voltage power supply directly or are obtained from the "boost B+" circuit (on the side of C_{242} labeled "boost B+").

Low-Voltage Power Supply. The low-voltage power supply furnishes the power for every stage of the receiver, including the low-voltage electrodes of the cathode-ray tube. The design of the low-voltage power supply is conventional, using two 5V4G rectifiers in parallel in order to obtain the 400 volts at 290 ma required by the set. The full 400 volts are not all positive, the ground connection being so placed that 300 volts are positive and 100 volts are negative. The supply is well filtered to prevent interaction between the many stages (and their many diverse operations) in the set.

Cathode-Ray Tube. The cathode-ray tube (which may have any screen size between 16 and 21 inches) is a magnetically deflected and focused tube using a deflection yoke, a focus coil, and an ion trap magnet.* The deflecting coils receive their driving power from the output transformers of the horizontal and vertical deflecting systems. The focus coil is connected into the negative portion of the low-voltage power supply, and means are pro-

* The new electrostatic focus tubes can be used in this receiver, if desired.

vided whereby the current through the focus coil may be varied. The ion trap uses a permanent magnet, and its adjustment is accomplished by physically changing its position on the neck of the cathode-ray tube until the brightest image is obtained.

Variation of the control grid to cathode potential is obtained by means of R_{152} , the brightness control. This control is connected into the d-c restorer circuit, but has no effect on this circuit because the d-c potential introduced by this control reaches the two elements of the d-c restorer tube in equal strength.

Keyed A.G.C. System. The keyed A.G.C. tube, V_{132} , receives its video signal from across R_{257} located in the plate circuit of the first video amplifier. The plate of this tube is fed a pulse obtained from terminal No. 4 of the horizontal output transformer. Current flows through the tube whenever the incoming sync pulses are active at grid No. 1 and the positive retrace pulse is active at the plate. The current, flowing through R_{259} , R_{260} , and R_{255} , develops a negative potential across R_{255} which is applied to the grids of the first two video I.F. amplifiers and the grid of the input triode of the cascode R.F. amplifier.

A 25,000-ohm potentiometer is inserted in the cathode leg of V_{132} to enable the set user to adjust the A.G.C. voltage to his particular conditions. The adjustment is made with the strongest signal tuned in. Observe the picture and slowly rotate the control until the picture begins to distort (or exhibits excessive contrast). Now back down the control to the point where the picture distortion is no longer evident.

Alignment of the Receiver — Equipment Required. For the proper alignment of television receivers, the following basic pieces of electrical apparatus are required: a cathode-ray oscilloscope, a wide-band sweep oscillator, a signal generator, a vacuum-tube voltmeter, and a marker signal generator that is capable of indicating specific frequency points on the test pattern swept out on the oscilloscope screen.

a. Oscilloscope. The cathode-ray oscilloscope is today a necessary piece of test equipment among the serviceman's electrical testing apparatus. Its greatest use, that of observing waveforms of different voltages and frequencies in the receiver, provides the repairman with a positive means of rapidly determining exactly what is occurring at all points in the circuit under test. It eliminates guesswork and permits accurate adjustments to be made until the correct operating conditions are attained. For a television receiver, satisfactory images are observed only if the various intervening circuits are functioning properly. The requirements become more stringent as the size of the screen increases and small defects become more readily apparent.

The cathode-ray oscilloscopes that are at present on the market differ little from each other for the same size screen. As the size of the screen is made larger, the number of controls available on the front panel increase, but the basic operation remains the same. The advantage of the larger-sized oscilloscopes is their greater possible viewing screen area and the improved frequency response of the vertical and horizontal amplifiers. For television alignment work, however, even a small 3-inch oscilloscope will prove satisfactory. A popular 5-inch model is shown in Fig. 15-8.

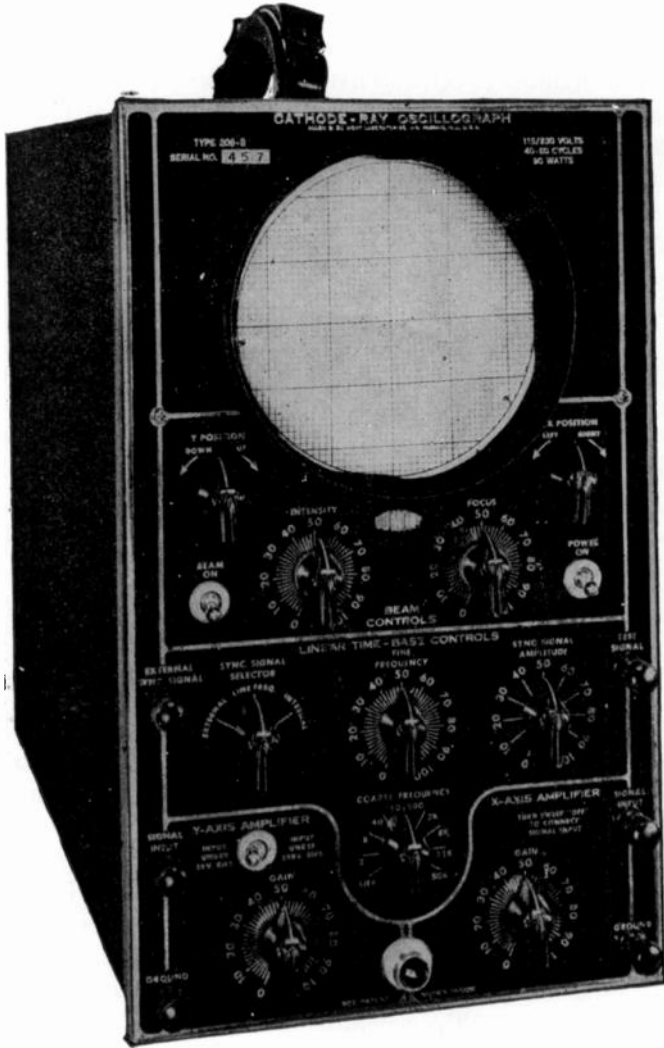
b. Alignment Oscillators. Because of the wide bandwidths that are peculiar to television receivers, the familiar signal generator, where only one frequency is available at any one time, is not especially suitable, by itself, for receiver alignment. With a single frequency entering the circuit, it is possible to determine only one point on the frequency response curve. To do this for a 6-mc or even 4-mc band would require too much time for ordinary service, where time is an important factor in determining the cost of the job. To meet the special requirements imposed by television receivers, special sweep oscillators, like the unit illustrated in Fig. 15-9, have been developed.

These generators are designed to provide an output signal that sweeps across a range of frequencies and continuously repeats this sweeping at a rate of 60 cycles per second. If this type of signal is applied to the video I.F. system of a television receiver, and an oscilloscope is connected across the output of the system (at the video detector, for example), then the pattern produced on the scope screen will represent the response curve of that system. This is in sharp contrast to the old method of using a signal generator to check circuit response by using one frequency at a time.

The range of frequencies obtainable from a sweep generator will vary with the instrument. Some are designed to cover only the video I.F. frequencies, say from 20 to 50 mc. Others cover this and the R.F. frequencies as well. In all of these instruments, there is a sweep width control which permits adjusting the range of frequencies swept out at any setting of the tuning dial. Thus, for example, suppose the generator is set for 25 mc. Then the amount by which the output signal sweeps above and below 25 mc can be varied from zero (when there is no sweep and the output frequency is 25 mc) to, say, ± 5 mc. This means the output signal varies periodically from 20 mc (25 mc - 5 mc) to 30 mc (25 mc + 5 mc).

For circuit alignment work, sweep generators are indispensable. This is especially true of the video I.F. system.

c. Single Signal Generators. Although the television receiver employs wide-band tuning circuits, the conventional amplitude-modulated signal generator is not entirely without application. In the receiver just analyzed,



Courtesy DuMont

FIG. 15.8. A modern service oscilloscope.

the I.F. single-tuned circuits are peaked with such a signal generator. Signal generators which generate one frequency at a time are also useful for marker points, as will be seen presently, and for testing the local high-frequency oscillator of the receiver. Finally, in emergencies when no wide-band signal generator is on hand, a single signal generator can be used to provide a fairly



FIG. 15.9. A signal generator covering the F-M and television bands.

satisfactory receiver alignment. All in all, the advent of television receivers has, in no way, reduced the usefulness of the standard signal generator.

d. Vacuum-Tube Voltmeters. The vacuum-tube voltmeter has always been a very handy instrument to have around, and with television receivers it becomes even more important. The vacuum-tube voltmeter, when properly constructed, has negligible loading effect on the circuit across which it is placed. In this respect it gives a true indication of the conditions in the circuit under test. Furthermore, by the use of special auxiliary probes, it will measure d-c voltages as high as 50,000 volts and R.F. voltages having

frequencies of several hundred megacycles. With the conventional 1,000-ohms-per-volt meter it is impossible to do this. Finally, the vacuum-tube voltmeter is practically immune to overloading, a feature for which even experienced servicemen frequently are grateful.

A modern vacuum-tube voltmeter is shown in Fig. 15-10.

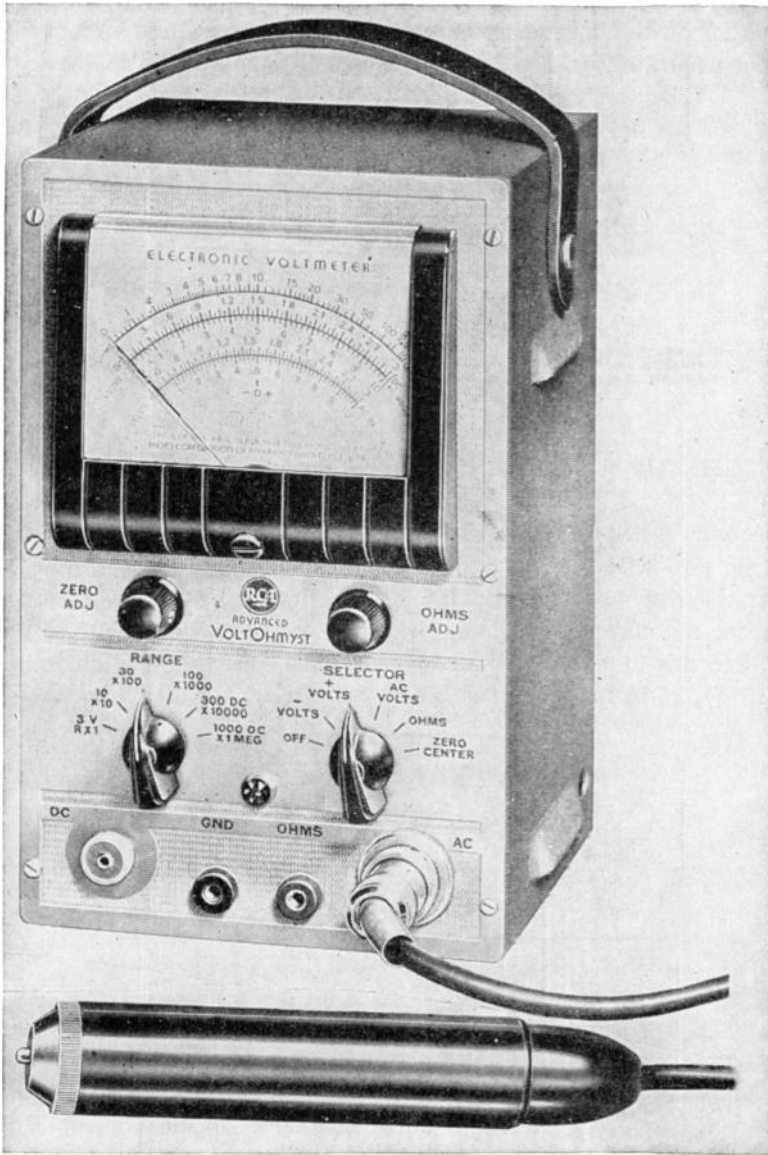
e. Marker Signals. The final piece of apparatus may be incorporated either in the sweep generator or supplied by an external signal generator that is capable of providing a single accurately calibrated signal. The purpose of a marker signal is to indicate the frequency at various points in the response curve observed on the oscilloscope screen. This will aid in adjusting the tuning slugs in the resonant circuits to the desired bandpass characteristics.

As an example, consider the response curve of Fig. 15-11, which is the response curve for the video I.F. system of the receiver shown in Fig. 15-1. This curve would be observed if we connected a sweep generator to the mixer grid and if the generator were set to sweep through the I.F. range. The oscilloscope would be connected across the video detector load resistor. What we desire to do, once we obtain the response curve, is to determine the frequencies at various points to insure that the curve rises where it should and falls where it should. It is here that the marker signal comes in handy.

If a manufacturer states in his service data that the video carrier I.F. is 25.75 mc and the other end of the response occurs at 21.75 mc, then the curve you obtain should be checked for the position of these two frequencies. To obtain marker points on the oscilloscope screen, two methods are generally employed. In the simplest method, the sweep signal generator contains an internal oscillator that superimposes its signals on the I.F. being swept out (20 to 30 mc). The indication of the marker point in the visible pattern is either a slight wiggle or else a dip in the curve at this point. (See Fig. 15-12.) Note that while two marker points are indicated in Fig. 15-12, only one would be seen at a time. First the marker oscillator would be set to 25.75 mc and its position noted on the curve and then set to 21.75 mc and its position checked again.

The above-mentioned video I.F. band limits, 21.75 mc and 25.75 mc, represent the entire 4 mc that can be employed to transmit the details of the televised scene. Many receivers are designed to pass only 3 mc in the I.F. amplifiers. In these instances, reference to the manufacturer's instructions will quickly indicate the band limits, and the marker frequencies can be changed accordingly.

If the sweep oscillator does not contain an internal device for supplying the marker points, then these may be obtained by the following method. Take another signal generator and place its output leads in parallel with



Courtesy RCA

FIG. 15.10. A vacuum-tube voltmeter capable of measuring voltages with frequencies up to 200 mc.

those of the sweep generator, using a $50\text{-}\mu\mu\text{f}$ isolating capacitor* in the signal lead of the marker generator. Set the frequency of this second oscillator accurately to one of the frequencies it is desired to check on the

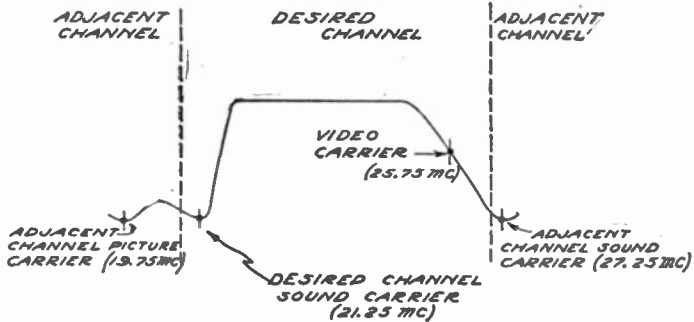


FIG. 15.11. The overall video I.F. response curve of the receiver in Fig. 15.1.

response curve. Say this is 25.75 mc. With the equipment turned on, a wiggle (or pip) will appear on the overall response curve at 25.75 mc. Note whether the response at this point is as indicated by the manufacturer in his service manual. Now change the marker frequency to 21.75 mc and note where this appears on the response curve. In this particular receiver (Fig. 15-1), 25.75 mc represent the video carrier, and 21.75 mc the end of

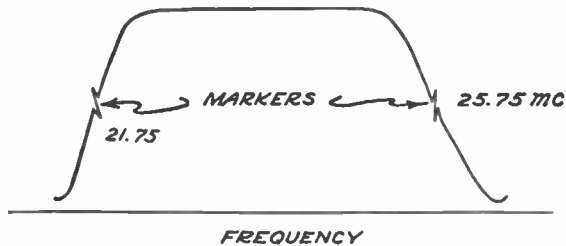


FIG. 15.12. The use of marker points for definitely indicating a frequency on a response curve.

the video bandpass. Other frequencies that should be checked are the trap frequencies. (See Fig. 15-11.) These include 21.25 mc, the sound carrier of the same channel and 27.25 mc, the lower adjacent channel sound carrier. At each of these points the response should be very low. In other sets that

* It may be found that attaching the marker generator directly to the sweep generator lead will cause the response curve to alter its shape. If this occurs, try inserting an isolating resistor of 10,000 ohms in the signal line of the marker together with the $50\text{-}\mu\mu\text{f}$ capacitor. The capacitor's value is not critical and values between $20\ \mu\mu\text{f}$ and $200\ \mu\mu\text{f}$ have been suggested by various manufacturers.

the serviceman will encounter, other frequency values will be used. The procedure, in any event, remains the same and requires only changing the frequency of the marker signal to these values.

It is advisable not to turn the amplitude of the indicating signal generator too high, but to keep it as low as possible (and still obtain a marker line).

This second method is highly flexible. While the sweep oscillator is moving back and forth across the band, the marker signal generator is set at one frequency. The setting of the marker generator can be altered at will, providing the serviceman with a means for identifying each point on the visible curve.

Video I.F. Stages. Alignment of the several stages of a television receiver does not differ basically from the procedure followed in the more familiar A-M sets. For television, it is true that more care must be exercised and the equipment is more extensive, but the alignment starts at the same place — the I.F. stages — and ends at the R.F. stages at the input of the receiver. In order to illustrate the general method, let us use the receiver shown in Fig. 15-1 as our example, pointing out the sequence of events and adjustments to be followed. While we will use the specific components of the present set, similar sections of other receivers may readily be substituted for similar results.

CONVENTIONAL-TYPE RECEIVERS

It is immaterial whether the alignment in conventional receivers starts in the video I.F. system or in the sound I.F. system. However, after one section is completed, the other is begun. The manufacturer of the set shown in Fig. 15-1 recommends the following order:

1. Video I.F. traps
2. Video I.F. transformers.
3. 4.5 mc trap in video amplifier.
4. Sound discriminator.
5. Sound I.F. transformers.
6. R.F. stages.

In our discussion we will employ this alignment order.

To start, the A-M (or single) signal generator is connected into the grid circuit of the mixer tube. The ground lead of the generator connects directly to the chassis of the receiver; the output lead is in series with a 50- μ f capacitor and attaches to test point 9 in the mixer grid circuit. The vacuum-tube volt-meter is placed across the video second-detector load resistor, or R_{137} .

The next step is to set the bias on the first two I.F. stages to the value it will have when signals of moderate strength are received. The value most

frequently suggested by manufacturers is -3 volts. In this set we pull out the keyed A.G.C. tube and then apply 3 volts from a battery to the A.G.C. line. The negative terminal of the battery connects to the ungrounded end of C_{230} , and the positive terminal goes to the chassis.

With this done, set the signal generator to each of the following frequencies and tune each specified adjustment for minimum indication on the VTVM.

T_{104}	21.25 mc
(Accompanying Sound Trap)	
T_{105} (top)	21.25 mc
(Same as preceding trap)	
T_{103} (top)	27.25 mc
(Adjacent Channel Sound Trap)	

The next step is the alignment of the video I.F. coils. The signal generator is set for each of the frequencies to which the coils are tuned and then the proper coil is adjusted for *maximum* response on the meter. The frequencies at which each coil is peaked are:

L_{11}	21.8 mc
(Converter Output)	
T_{105} (bottom)	25.3 mc
(1st Video I.F.)	
T_{104} (bottom)	22.3 mc
(2nd Video I.F.)	
L_{185} (top of chassis)	25.2 mc
(3rd Video I.F.)	
L_{185} (top of chassis)	23.4 mc
(4th Video I.F.)	

If any of the transformers (T_{103} and T_{104}) require any appreciable adjustment, it is advisable to recheck the traps at these coils. In this system, there is a possibility of I.F. oscillations if two or more of the I.F. transformers are tuned to the same frequency. I.F. oscillations will cause a voltage in excess of 3 volts to appear across the diode load resistor. This voltage is not affected by R.F. signal input and it will also be independent of the setting of the contrast control. To stop the oscillations, shunt the grids of the first three I.F. amplifiers to ground with 1,000- $\mu\mu\text{f}$ capacitors. Then connect the signal generator to the grid of the fourth I.F. amplifier tube and adjust the coil in the plate circuit of this tube (L_{185}) for maximum response. Shift the signal generator to the grid of the third I.F. amplifier. Remove the shunting capacitor here and, with the generator set to the proper frequency, adjust the plate-tuned circuit of this tube for maximum response. Repeat this procedure for the second and then the first I.F. amplifier. The oscillations should now be completely gone if the circuits are functioning properly.

The entire video I.F. system is now adjusted. An overall check should be run, with the sweep oscillator connected to test point 9 in the mixer grid circuit. A 50- μ f capacitor is inserted in series with the output lead of the generator. The vertical input terminal of the oscilloscope is connected to the ungrounded end of R_{137} through a 10,000-ohm isolating resistor. The scope ground connects to the receiver chassis. Also, if the sweep generator uses a 60-cycle sine-wave driving voltage (and nearly all do), it will have some of this voltage available at a front panel terminal. A wire should be run between this terminal and the horizontal input bending post on the scope. Then, the saw-tooth sweep of the scope is cut off, and this sine-wave voltage is substituted in its place. The reason for doing this is to insure that the pattern produced on the screen will be linear in frequency. That is, a 1-mc change in frequency at one end of the response curve will occupy as much space horizontally as a 1-mc change in frequency at the other end of the curve.

Also available on the sweep generator is a phase control. This is rotated until a single pattern is obtained on the scope screen.

With the oscilloscope power on and the unit operating properly, the I.F. response should be visible on the screen. Check the shape of the curve against the one recommended by the manufacturer. With a variable marker signal, check each of the end points of the response curve and the trap frequencies. If any slight adjustments are required, they are made with

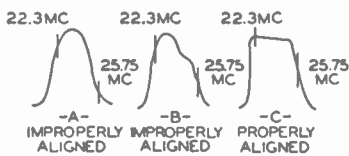


FIG. 15.13. The form of the video I.F. response curve for proper and improper alignment.

the sweep generator on so that the effect on the overall curve is seen. However, if it is found that any considerable adjustment is necessary, then the entire alignment procedure should be redone.

The form of the overall video I.F. response curve when the circuit is improperly aligned is indicated in Fig. 15-13 A and B. The proper shape is the one in Fig. 15-13C.

4.5-mc Trap Adjustment. This receiver contains a 4.5-mc trap in the plate circuit of the first video amplifier, V_{115} . To adjust this trap we would proceed as follows. A signal generator is connected to the grid of V_{115} . The generator frequency would be set to 4.5 mc. For an indicator we need two things: A VTVM and a germanium crystal detector. The latter device is needed to rectify the 4.5-mc signal so that an indication can be obtained on the VTVM. The meter will not respond directly to a 4.5-mc voltage.

If the VTVM possesses an R.F. probe, this may be used. Otherwise the crystal detector shown in Fig. 15-14 may be employed. The input terminal of the crystal detector connects to the picture tube grid. The other input

terminal attaches to the receiver chassis. The output of the detector is applied to the d-c terminals of the VTVM. Use the lowest scale here.

Now, with the generator output turned up, adjust the 4.5-mc trap for minimum meter indication.

Sound Discriminator Alignment. With the video I.F. stages functioning in good order, we next tackle the sound I.F. system and the F-M discriminator.

To align the discriminator, set the signal generator for approximately 1-volt output at 21.25 mc. Connect the generator output lead to the grid of the third sound I.F. The ground lead goes to the receiver chassis. The VTVM is set to the 10-volt scale. It is then connected to the point where R_{219} and R_{220} join. Now detune the secondary winding of T_{113} and adjust the primary winding of the same transformer for maximum indication on the meter.

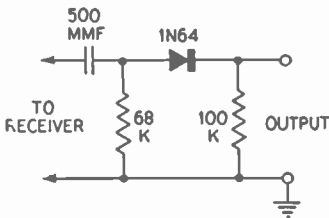


FIG. 15.14. A crystal detector network.

The reason for this procedure is that the discriminator is a balanced circuit and produces across both load resistors (R_{219} and R_{220}) no output at the center sound I.F. (21.25 mc in this case). By connecting the meter across half the load and throwing the secondary of the discriminator out of adjustment, we can tune the primary winding to

its proper peak. Next, to bring the secondary (and the discriminator) back into balance, we proceed as follows:

Connect the VTVM to the top of R_{220} . This places it across both load resistors. Adjust the secondary of T_{113} for zero indication on the meter. On one side of this zero point, the meter will read a positive d-c voltage and, on the other side, a negative d-c voltage. The best procedure, then, is to move slowly back and forth about the zero point, each time lessening the extent of the excursion until the exact zero point is reached.

The object, when aligning a discriminator, is to make its response linear over its operating range of frequencies.

The linear section is that portion of the discriminator characteristic between points *A* and *B* (Fig. 15-15). The primary winding of the discriminator transformer (T_{113}) governs the extent of the linear portion of the curve between the end points where the curve reaches its peak and begins to fall off. The secondary winding governs the position of the cross-over point. The cross-over point should be located midway between the upper and lower peaks. We can use the sweep generator to see the entire curve at one time and determine whether both of the foregoing conditions are fulfilled. This is the next step.

The sweep generator is connected to the grid of the third I.F. sound amplifier. Set the sweep bandwidth to approximately 1 mc with the center frequency at 21.25 mc and with full output. The vertical input terminals of the oscilloscope are connected between the top of R_{220} and ground. The pattern that is obtained should be as shown in Fig. 15-15. If the curve is not symmetrical or the linear portion sufficiently long, both trimmers of T_{113} will have to be readjusted. The curve should be linear for at least 200 kc. A separate generator used to provide a variable marker will indicate the extent of the linear portion.

Sound I.F. Alignment. To align the sound I.F. stages, connect the oscilloscope to the third sound I.F. grid-leak resistor, R_{217} . It is suggested that a 10,000-ohm isolating resistor be placed in series with the oscilloscope

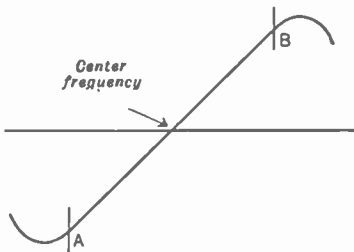


FIG. 15.15. Discriminator characteristic curve.



FIG. 15.16. The shape of the response curve and the position of the marker frequency when aligning the F-M, I.F. stages.

lead that connects to the top of R_{217} . (We are still using the vertical input terminals of the oscilloscope.)

Connect the sweep generator to the grid of the second sound I.F. amplifier and insert a 21.25-mc marker signal from the other signal generator at the same point. The pattern obtained on the oscilloscope screen should be similar to the illustration in Fig. 15-16 with the marker pip in the center of the curve. Transformer T_{112} is adjusted for maximum gain and symmetry about the 21.25-mc marker.

It is important that the sweep oscillator output be kept as low as possible — as low as it is possible to obtain a usable indication on the oscilloscope screen. Too large an input signal will produce a broadened response curve, permitting misadjustments to pass unnoticed and possibly causing distortion on weak signals.

Place the sweep oscillator and signal generator at the grid of V_{104} and adjust T_{111} for maximum gain and symmetry about the 21.25-mc marker. This completes the adjustment of the sound I.F. system.

As an alternate procedure, it is possible to align the sound I.F. stages, using only an A-M signal generator. Connect the generator to the grid of

V_{104} . Set it to 21.25 mc and turn on the audio modulation. In spite of the fact that V_{107} is an F-M detector, the audio note of the signal will be heard in the loudspeaker. Now all we need do is to peak the primary and secondary of T_{111} and T_{112} , in turn, for loudest audio sound from the speaker. If desired, the primary of T_{113} can be similarly adjusted. Then, the secondary can be set for minimum sound. This method of aligning the discriminator transformer is not as good as the sweep generator and scope method, but it may be used.

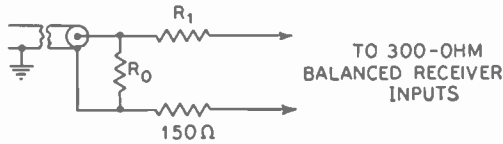
R.F. and Converter Circuit Adjustment. The front-end section of a television receiver possesses a greater overall bandpass (6 mc) than either the video or sound I.F. stages that follow it. To attain this bandpass, the tuning circuits are heavily loaded by a combination of shunting resistors and the relatively low input impedance of vacuum tubes at high frequencies. The tuner used in the receiver of Fig. 15-1 is the Standard Coil tuner containing separate removable R.F., mixer, and oscillator coils for each of the twelve VHF channels. There are only three adjustments for the R.F. and mixer circuits on all twelve channels (C_2 , C_3 , and C_4) and these are set so that the most uniform response is obtained on all local channels. The oscillator coils have a tuning slug for each channel, this being required since the oscillator generates a single frequency per channel, and it is relatively easy for this frequency to drift too far to one side or the other.

Alignment of the R.F. stages carries with it a number of precautions that must be observed if the job is to be properly carried out. First, there is the matter of matching the sweep signal generator to the receiver input terminals. Most sweep generators have an unbalanced output with impedances between 50 ohms and 100 ohms. On the other hand, television receivers possess either a 75-ohm unbalanced input or a 300-ohm balanced input. The 75-ohm unbalanced receiver input can usually be connected directly to a generator whose output is similarly unbalanced and whose impedance lies between 50 to 100 ohms without causing appreciable mismatching. However, when you attempt to connect such a sweep generator directly to the input terminals of a 300-ohm balanced receiver, the response pattern will be sufficiently affected to result in misadjustment of the circuit trimmers. (The input tuning circuits of a television receiver contribute to the shape of the overall R.F. response curve, and, when the proper matching is not employed between instrument and receiver, the tuning curve of this input circuit is affected. This, in turn, alters the overall pattern.)

To match an unbalanced sweep generator to a 300-ohm balanced television receiver, an arrangement such as shown in Fig. 15-17 would be employed. One series resistor has a fixed value of 150 ohms. The other series resistor has a value which depends upon the impedance of the sweep generator output; this is true also of the shunt resistor, R_0 . The resistors should

be of the non-inductive variety, preferably carbon or of composition construction.

As a first step, the sweep generator is connected to the receiver input terminals, as described. The scope vertical input terminal connects to test point 9 through a 10,000-ohm resistor, and the scope ground attaches to the receiver chassis.* The scope beam should be driven by a 60-cycle sine-wave voltage obtained either from the sweep generator or from the oscilloscope if the latter contains a phase control. The marker signal, when it is needed after a response curve has been obtained, would be loosely coupled to the sweep generator, either by connection through a small (5–10 $\mu\mu\text{f}$)



GENERATOR IMPEDANCE	R_0	R_1
50 Ω	56 Ω	120 Ω
72 Ω	82 Ω	110 Ω
92 Ω	110 Ω	100 Ω

FIG. 15.17. How to match signal generator output impedance to balanced receiver input (300 ohms).

capacitor, or by laying the marker generator output cable across the resistive matching network.

The negative terminal of a 3-volt battery is clipped onto the A.G.C. wire, and the positive terminal goes to the chassis. If it is found difficult to obtain a curve of sufficient amplitude, the battery can be removed, and the tuner A.G.C. wire can be simply grounded to the receiver chassis.

After an appropriate warm-up interval of 10 to 15 minutes, the receiver is tuned to channel 12. The response curve sought for this channel is shown in Fig. 15-18. If the curve observed does not possess this form, adjust C_2 , C_3 , and C_4 . Adjusting C_3 will generally shift the center of the response curve in relation to the video and sound carrier markers. C_2 and C_4 should be alternately adjusted for best gain with flat-top appearance. The response curve should have maximum amplitude and flat-top appearance consistent with proper bandwidth and correct marker location. Do not overly broaden the curves since this will result in a loss of sensitivity.

*The mixer, being essentially a detector, provides the oscilloscope with a demodulated voltage. Should the response pattern of the R.F. amplifier stage itself be desired, then an R.F. probe would be required.

After the curve for channel 12 has been obtained, each of the other channels should be checked in turn. In each instance, the marker generator frequency would be changed to the corresponding video carrier frequency and sound carrier frequency for the particular channel being tested. None of the R.F. coils possesses individual adjustments, and ordinarily the C_2 , C_3 , and C_4 settings established for channel 12 will provide satisfactory response curves on all channels. However, if reasonable alignment is not obtained on a particular channel, (a) check to see that coils have not been intermixed, or (b) try replacing the pair of coils for that particular channel, or (c) repeat the C_2 , C_3 , and C_4 adjustments for the weak channel as a compromise adjustment to favor this particular channel. If the compromise

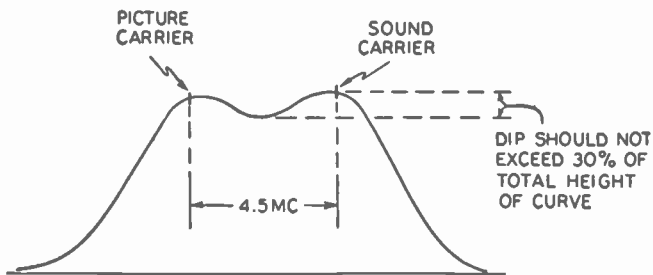


FIG. 15.18. R.F. response curve.

adjustment is made, other channels operating in the locality should be checked to make certain that they have not been appreciably affected.

R.F. Oscillator Adjustment. There are several methods that may be employed to align the oscillator, but the simplest method is as follows. Connect an A-M signal generator to the receiver input. Rotate the tuner for the desired channel. Set the fine-tuning control to mid-position. Turn the volume control fully clockwise. Finally, connect a wire jumper from test point 9 of the mixer grid to the control grid of the first audio amplifier. As a precaution against interference, remove the 6AL5 sound discriminator.

Now, set the generator to the exact oscillator frequency for the channel to which the tuner is adjusted. See p. 389. Then, carefully adjust the individual oscillator slug until a beat whistle is heard in the speaker. At this point the oscillator is correctly set. The same procedure is followed for all other channels on which the oscillator is to be adjusted.

AN INTERCARRIER RECEIVER

The block diagram of a General Electric television receiver operating on the Intercarrier principle is shown in Fig. 15-19. The associated schematic diagram is given in Fig. 15-20.

The R.F. tuner contains two R.F. amplifiers, an oscillator, and a mixer. Mechanically the tuner employs a wafer switch, and the various VHF stations are brought in by rotating the selector switch to the desired position. The coils which tune the first and second R.F. stages, the converter, and the oscillator are mounted on individual wafers of the selector switch. The local oscillator V_{103A} operates on the high-frequency side of the incoming R.F. signals. In the mixer the video carrier is brought down to 45.75 mc; the sound carrier is converted down to 41.25 mc.

The first R.F. amplifier possesses a grounded grid with the signal being applied to the cathode circuit. Coils L_{101} to L_{105} are incorporated into the cathode circuit to help maintain a 300-ohm input. L_{106} and its capacitors

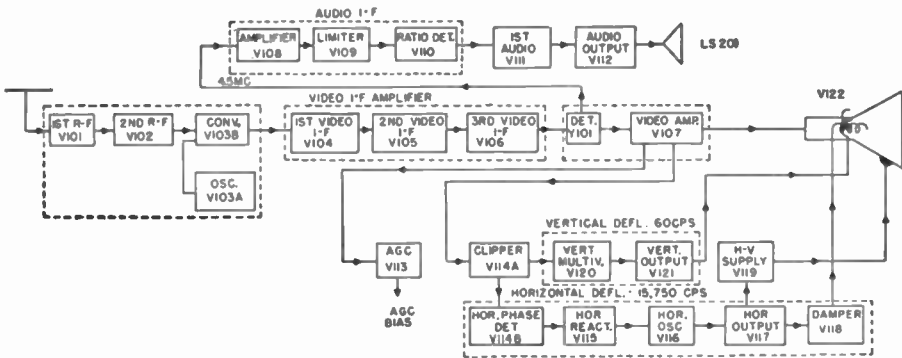


FIG. 15.19. Block diagram of the Intercarrier receiver of Fig. 15.20.

(C_{102} and C_{103}) form an I.F. trap to assist in the reduction or removal of any A-M or F-M interfering signals at the I.F. frequencies (41–47 mc) that may be picked up by the antenna system.

The second R.F. amplifier is a straightforward pentode amplifier. Grid bias is partially obtained from the cathode resistor (R_{105}) and partially from the A.G.C. network. The sound and video signals are transferred from the plate of this second R.F. amplifier to the grid of the converter via a 3.9- μmf capacitor (C_{113}). Oscillator voltage reaches the mixer at the same point via a small 1.5- μmf capacitor.

Video System. The video I.F. system is stagger-tuned to pass the video I.F. (45.75 mc) and the sound I.F. (41.25 mc) carriers with the proper amplitude relationship. The first two video I.F. stages are A.G.C.-controlled, while the third stage develops its own bias. Three trap circuits in the system serve to prevent signals from adjacent channels from interfering with the normal operation of this receiver. The two 47.25-mc traps are for the removal of any adjacent lower-channel audio voltage that may be present. The 39.75-mc trap is designed to remove the video carrier signal of a

higher adjacent channel. A fourth trap, at 41.25 mc, serves to reduce (but not eliminate) the sound carrier of the same channel to the necessary level for Intercarrier operation.

The reader may wonder why a single 39.75-mc trap will suffice to eliminate the video carrier of an adjacent channel whereas a single 41.25-mc trap will not eliminate the sound carrier of the tuned-in channel. The answer is that the R.F. tuning circuits also impose some attenuation on adjacent channel signals, and this aid, together with that obtainable from a video I.F. system trap, is generally enough to reduce these signals below a usable level. The sound signal of the same channel, however, is not attenuated; hence, it would require more than a single trap in the video I.F. system to remove it completely.

The video signal is detected by the germanium crystal detector (1N64) as well as the 4.5-mc F-M signal which is the beat frequency between the video I.F. (45.75 mc) and the sound I.F. (41.25 mc). The sound is taken off at the take-off coil L_{163} and applied to the sound system. The video signal is coupled to a 6CB6 video amplifier and from here passed on to the cathode of the picture tube. A picture contrast control is placed in the screen-grid circuit of the 6CB6 and, by varying the voltage to this element, it varies the stage amplification.

It will be seen that $R-C$ coupling is employed between the video amplifier and the picture tube, thereby eliminating the d-c component of the video signal. No d-c reinsertion network is employed, and, to avoid the appearance of the vertical retrace lines at some settings of the contrast control, a vertical blanking pulse is fed to the control grid of the picture tube. The negative pulse is obtained from the vertical output transformer and reaches the picture tube control grid via R_{178} , C_{170} , C_{169} , and R_{177} . A positive d-c voltage is also applied to the control grid to offset some of the positive cathode voltage. The brightness control is located in the cathode circuit of the picture tube.

Sound System. The sound system uses one 6CB6 as a 4.5-mc I.F. amplifier, one 6AU6 as a limiter, a ratio detector, and two stages of audio amplification. There is nothing unusual in this system to warrant special attention except possibly the placing of the focus coil and control in the cathode circuit of the 6W6GT audio output stage. This is probably done here because of the considerable current passed by the 6W6GT tube. The setting of the focus control determines how much current will actually pass through the focus coil, and in this way a suitable variation can be obtained.

Sweep Systems. A portion of the video signal is taken from the plate of the V_{107} and applied to the sync clipper. In the output of this tube the horizontal and vertical sync pulses separate. The vertical pulses pass through an integrating network (R_{313} , R_{314} , R_{315} , and C_{309} , C_{310}) to trig-

ger the vertical multivibrator. C_{304} develops the vertical deflection voltage. This is then amplified by V_{121} (two triodes in parallel for power) and applied to the vertical deflection windings of the yoke through T_{301} . The same transformer also furnishes a retrace blanking signal to the picture tube.

Returning to V_{114A} , we see that a triggering pulse is also furnished to the horizontal sweep system via C_{352} . The automatic frequency control system employed here is the Gruen circuit developed by General Electric. This, it will be recalled, was previously discussed in Chapter 13. The horizontal deflection voltage is developed across R_{366} and C_{365} and then transferred to the 6AV5 horizontal output amplifier. The signal is amplified in this stage and then transformer-coupled to the deflection coils. At the same time T_{351} utilizes the horizontal retrace current to produce a high voltage for the 1B3 which rectifies it and supplies the high voltage to the accelerating anode of the picture tube. The tube V_{118} works as a damper tube and produces a "B+ boost" voltage in conjunction with the filter in its cathode circuit. The low-voltage power supply of this receiver uses a normal power transformer and two rectifiers, V_{123} and V_{124} .

Keyed A.G.C. System. The keyed A.G.C. system in this receiver operates essentially in the same manner as the keyed A.G.C. systems described in Chapter 6. Thus, the composite video signal with sync pulses extending upward (or positive) is applied to the control grid of the 6AU6 tube, V_{113} . The plate receives its periodic horizontal pulses from the secondary of T_{351} , the horizontal output transformer. Plate current flows when these pulses are active, developing the A.G.C. bias across resistors R_{256} and R_{254} .

Since the plate current of the pentode 6AU6 is independent of plate voltage, the keying pulse amplitude has negligible effect in establishing the A.G.C. bias. This is controlled entirely by the peak amplitude of the signal applied to the grid circuit of the keyer tube. Therefore, the video signal is applied with its d-c component by taking it off the anode of the video amplifier, V_{107} .

If the station signal increases suddenly, the positive sync pulse amplitude at the grid of the keyer tube also increases; the decrease in bias causes a greater current, resulting in an increased A.G.C. bias which biases the tubes of the video I.F. amplifier so that the gain is reduced. The opposite action takes place when the signal decreases. The result is good A.G.C. action.

In order to improve this system, it is important to operate the video amplifier stage as a noise limiter. Therefore, the signal applied to the grid of the video amplifier should be of such an amplitude that plate current cut-off is almost reached on the peaks of the sync pulses. A large noise pulse is made harmless by clipping it so that it extends only slightly beyond the

tips of the sync pulses. This operation can be maintained effectively only over a narrow range of signal amplitudes: for a strong signal reception, a high A.G.C. bias voltage has to be developed; for weak signal reception, a lower A.G.C. bias is required.

To extend the noise clipping to a weaker signal level, the screen of the keyer tube is varied with the signal strength so that, for a weak signal, a low screen voltage is applied and, for a strong signal, a large screen voltage is used. This is accomplished by obtaining the screen voltage for the keyer tube through a dropping resistor R_{402} in the B+ supply. Thus when the signal is weak, the plate current of V_{104} and V_{105} is high and the voltage drop across R_{402} is large, providing a low screen voltage. In a pentode a lowering screen voltage reduces the plate current, and therefore the A.G.C. bias is lowered for the weak signal. The lower A.G.C. bias increases the gain of the video amplifier to compensate for the loss in signal strength, and better noise clipping results.

To provide an initial bias to the A.G.C.-controlled tubes, a fixed bias of approximately 0.6 volt, derived from the grid bias of the horizontal oscillator V_{116} , is applied across resistor, R_{254} . Thus when no signal is received, the R.F. and I.F. amplifier tubes supplied with A.G.C. will have this minimum bias applied to them so that their plate current will not be excessive.

QUESTIONS

1. Explain the function and operation of the focus, contrast, brilliancy, and fine-tuning controls.
2. What controls are generally classified as secondary controls? Where is their position on the television receiver chassis?
3. What type of R.F. tuning system is used in the receiver of Fig. 15-1? Explain how it operates.
4. What is the function of each of the following components in the R.F. section of Fig. 15-1: R_5 , L_{11} , L_8 , R_7 , R_1 , C_9 , and C_5 ?
5. What would happen if R_1 opened up?
6. Explain why there is an oscillator adjustment for each channel, yet C_2 , C_3 , and C_4 are the only adjustments for the R.F. circuits for all channels.
7. List all the trap circuits in the video I.F. system, together with their resonant frequencies. Explain how each trap circuit achieves its purpose.
8. Describe the operation of the contrast control of Fig. 15-1.
9. List the high frequency compensating components in the stages following the video second detector. Which are the low frequency compensating components?
10. Explain how the horizontal synchronizing discriminator controls the frequency of the horizontal sweep oscillator.
11. What type of oscillator is used in the vertical synchronizing system? Redraw the circuit so that it assumes a more conventional form.
12. Explain the operation of the horizontal output system, using the proper waveforms.
13. How does the vertical linearity control function?

14. What equipment would be needed to align thoroughly this receiver?
15. What is a marker signal? Why is it useful? How are marker signals generated?
16. Explain in detail the alignment procedure for video I.F. trap circuits.
17. Outline the alignment procedure for the video I.F. system.
18. How is the sound I.F. system aligned?
19. By what method is the oscillator adjusted?
20. What adjustments are made in aligning the R.F. amplifier and converter?
21. A receiver is brought in for repair with the complaint that dark bars appear across the screen. It is later determined that this is true only when a signal is being received. What components are at fault and how can they be corrected?
22. What type of tuner is employed in the receiver of Fig. 15-20? Explain briefly how it operates.
23. Which channel is the receiver tuned to in Fig. 15-20?
24. Make a listing of the tubes in the receiver of Fig. 15-20 together with their function.
25. What components form the vertical retrace suppression circuit?
26. If you obtained sound on this receiver but no picture, what tube would you check first? Explain the reason for your choice. (Assume a raster is present.)
27. In the absence of both sound and picture, what stages would you check?
28. What is the purpose of the connection between pin 1 of V_{116} and R_{264} ?

Chapter 16

SERVICING TELEVISION RECEIVERS *

Introduction. Television receivers, in their present state of development, are critical mechanisms that require accurately adjusted circuits if the maximum enjoyment of a viewer is to be derived. Indiscriminate replacement of component parts, a practice quite popular for many sound receivers of today, will generally cause more grief than good and should be discouraged. Careful adherence to manufacturer's values is especially important in the timing circuits of the receiver deflection systems. Hold controls permit some variation of the oscillator's frequency, but the limits are fairly narrow. Wide discrepancy between the values of the replacement parts and those specified by the manufacturer would render synchronization impossible.

Experienced servicemen utilize the indications (or lack of them) obtained from the loudspeaker of a sound receiver to their greatest extent. Probably the best example of the usefulness of the indications from the speaker in servicing work is the hoarse, rasping sound obtained when a filter capacitor in the power supply becomes defective. The set continues to operate, but with reduced volume and the characteristic distortion of sound. A receiver with this trouble, in the hands of an expert serviceman, would be repaired in less time than it takes to describe the defect. The greatest difficulty, however, in trouble-shooting is experienced when the set operates intermittently, the defect appearing and disappearing without any apparent reason.

A television receiver, in addition to having a loudspeaker for its sound, has another and better source of indication, the viewing screen. The eye is a more critical judge than the ear, and defects in television sets reveal

* The reader's attention is also directed to the author's "Television and F-M Receiver Servicing," 3rd Edition, 1953, published by D. Van Nostrand Co., Inc. Here will be found more detailed servicing procedures than can be conveniently handled in this text.

their causes far more readily than comparable defects in sound receivers. The two (sound and sight) combine to facilitate television service work to a remarkable degree. True, a certain amount of experience is required, but it can be obtained in a relatively short time. With the basic principles (as presented in previous chapters) clearly in mind, the radio technician should be able to associate the various distortions of the image with specific circuits throughout the receiver.

Servicing Guide Points in Television Receivers. All television receivers can be divided into two groups, conventional and Intercarrier sets. In each of these systems, in turn, there are sound and video systems, and it is important to know where the separation of the signals occur. The servicing significance of the separation point can be seen by the fact that, up to the separation point, both signals travel side by side. What affects one signal will also affect the other. However, beyond the separation point, each signal travels alone.

Thus, suppose a receiver exhibits a distorted video output accompanied by a distorted audio indication. This condition immediately tells the serviceman that the defect must be located in a circuit through which both signals pass together, and it directs attention to the stages preceding the separation point. On the other hand, when only one signal appears distorted at the output, then obviously this distortion must have occurred in some circuit dealing solely with this signal. If the audio is defective, then all attention is centered on the audio stages *following* the point of separation. If the image is distorted, only the video stages *beyond* the separation point need be examined.

With these simple facts in mind, tracking down troubles in television receivers can be made simply, logically, and quickly.

When servicing a receiver, your first step is to locate the signal separation point. In Intercarrier receivers you know that this occurs at the video second detector or beyond; in conventional-type sets, signal separation must take place between the mixer and some point before the video second detector. A rapid procedure for locating the stage where the signals separate is to start at the video second detector. Pull this tube from its socket and note what has been affected. If only the picture is gone, it indicates that the set is of the conventional type and that the separation point exists prior to the video second detector. On the other hand, if both sound and video are affected, then the set uses the Intercarrier system.

To locate the exact separation point, you would then proceed to pull tubes in the video I.F. system — if the set is of the conventional type — until you reach the stage where tube removal affects both sound and video. In an Intercarrier set, tube pulling would be directed toward the video amplifier stages following the video second detector. This method will work

well on all television receivers except those in which the filaments are series-wired. In these sets, removal of one tube in a string will affect all others in the string and probably prevent the signal from reaching the picture tube or loudspeaker. Under these conditions, actual circuit tracing or reference to the set schematic is necessary to locate the separation point.

Fig. 16-1 illustrates, in block diagram form, the method just described for localizing defects using the signal separation position in conventional-type sets. Fig. 16-2 does the same for Intercarrier receivers.

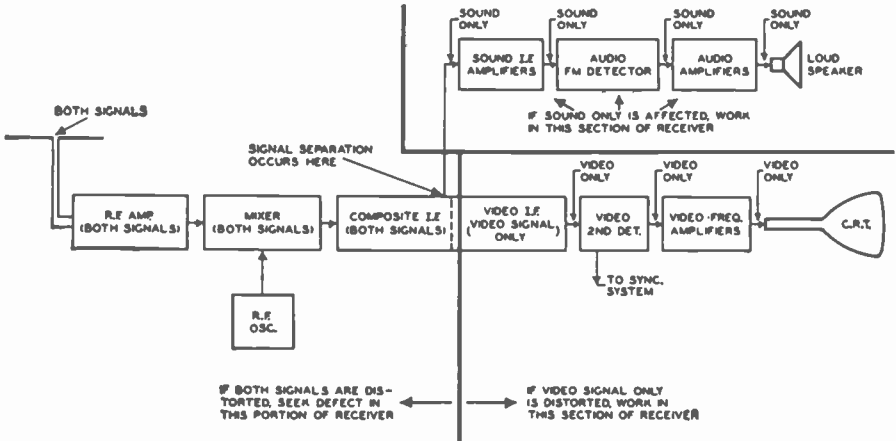


Fig. 16.1. A block diagram of a conventional-type television receiver illustrating the significance of the signal separation point.

Once the defect has been narrowed down to a specific section of the receiver, the next step is to analyze that particular section with a view toward further localization. This brings us to the block diagrams shown in Figs. 16-3 and 16-4. In the audio system, the general breakdown consists of the I.F. amplifiers, the F-M detector, the audio amplifiers, and the loudspeaker. In the video system we have, in addition to an I.F. section, a second detector and video amplifiers, the vertical and horizontal sweep systems. Furthermore, most television receivers have the high-voltage power supply located at the output of the horizontal sweep system.

Once the difficulty has been traced to a particular system in the receiver, we are in a position to conduct a further analysis of our defective receiver. In the video system, for example, breakdown of the signal path in sections A, B, or C will prevent the video signal from reaching the cathode-ray tube. If the sound separation point is prior to the breakdown, sound will be unaffected. If it is after the breakdown, it, too, will be lost. In either instance the picture tube screen will contain a raster, but no sound. The scanning raster will be due to the sweep oscillators which continue to func-

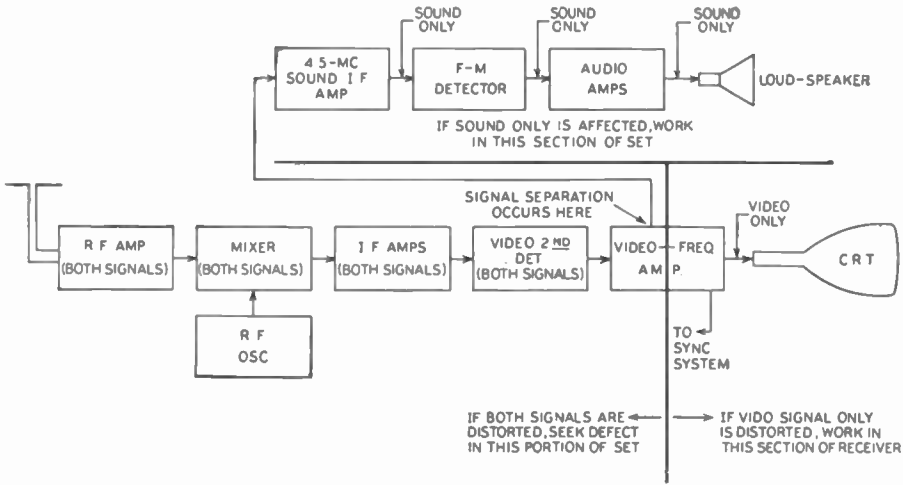


Fig. 16.2. A block diagram of an Intercarrier television receiver illustrating the significance of the signal separation point.

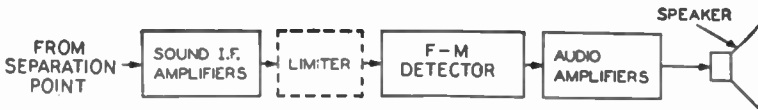


Fig. 16.3. A block diagram of the audio section of a television receiver.

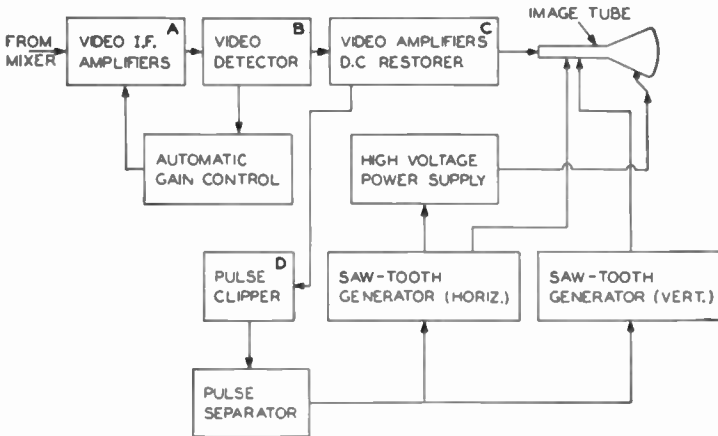


Fig. 16.4. A block diagram of the video section of a television receiver.

tion because they are self-oscillating. These oscillators are not being triggered, but they do oscillate.

As another illustration, suppose the circuit opens up in section D. The sync pulses will be prevented from reaching the sweep oscillators; consequently the oscillators will not be controlled by the incoming signal. On the other hand, image signals are reaching the cathode-ray tube. The visual result will be a picture which is scrambled because the various sections are not being placed on the screen in proper sequence. The audio section will be unaffected.



FIG. 16.5. A scanning raster.

These are but two of the many clues that the serviceman receives each time the set becomes defective. In this chapter we will undertake the analysis of many common clues that are encountered in defective television receivers. Recognition and interpretation of these signposts will help the serviceman in 90 per cent of his work.

R.F. System. A defect in the R.F. section of a television receiver will affect both the sound and video outputs. If the signal is prevented from passing through completely, no sound will be heard from the loudspeaker and no image will be seen on the screen. However, what will be visible is a scanning raster. (See Fig. 16-5.)

Since tubes are, by far, the greatest source of trouble in any section of a television receiver, they should be checked first. If the tubes are good, the trouble lies elsewhere in this section — either in the tuning circuits or in the resistors and capacitors.

The serviceman should know the method of tuning that is employed in the set because this fact will help him determine whether or not the tuned circuits are at fault. For example, in the Standard Coil tuner, a completely

separate tuning circuit is switched in for each channel. When all channels are affected, some common component, such as a tube or resistor, must be at fault. If the tubes have already been checked, then make a voltage and resistance test to find any faulty resistors.

In the RCA tuner (Fig. 3-13), all of the tuning coils are in series. An open coil in any tuning line will disrupt operation for its channel and all lower ones. It will not, however, interfere with the operation of higher frequency channels because the B+ of the unaffected section of the tuning line will be able to travel through the shorting bar to reach the triode plate in the open line. However, an open circuit in both sections of any tuning line, on any channel, will disrupt all operation. A resistance check will reveal the open coil or coils. The approach to each different tuner will be governed by its design, and no general rules can be set up to cover all tuners. However, find out first whether the tuning circuits for the various channels are independent of one another or whether what happens to one coil will affect others. This information is basic and will help immeasurably in any subsequent analysis of tuner trouble.

There are several signal tests which are useful for indicating whether or not a signal can get through the front-end stages. A rough test, but one which will reveal whether or not a signal can get through, is to detach the transmission line from the receiver input terminals and brush the line leads across the input terminal posts. Bursts of noise will be heard in the speaker (even with F-M) and flashes of light will appear in the scanning raster. If the signal cannot get through the R.F. stages, these indications will not be obtained. Admittedly this is a rough test, but it is surprising how well it works. On some sets, even disconnecting the transmission line is unnecessary. Merely turn the contrast control to its maximum clockwise position. Noise will be heard from the speaker, and snow with light flashes will be seen on the screen.

One problem which might arise is how to distinguish between a defect in the front-end stages and the complete absence of any signal reaching the television receiver. The latter could occur if there was a break in the transmission line leading from the antenna to the receiver. A simple method for checking the front-end stages is by the use of an A-M signal generator. Connect the instrument across the input terminals and set the generator frequency to a value about 1 mc above the video carrier frequency for that channel. Turn on the A-M modulation in the generator. If dark bands appear across the screen of the image tube, then a signal is able to pass through the R.F. stages. Note that the signal generator will not produce any indication on the screen if the receiver oscillator is inoperative.

The foregoing tests will indicate whether or not the tubes are at fault and whether the tuned circuits are in operating condition. The tests can

be applied in a few minutes. If the tests indicate that the tubes are O.K. and the set is completely dead on all channels, then only a voltage and resistance check will reveal the source of trouble.

Voltage and Resistance Checks. In the R.F. system trouble is most likely to originate in the oscillator. Hence, the first step in checking this section is to measure the d-c voltage at the oscillator grid with a vacuum-tube voltmeter. Typical values range from 2–10 volts, depending upon the design of the circuit. This voltage should be checked on all channels, since it is quite possible for the oscillator to function at some frequencies and not at others. If the voltage measured differs appreciably from the value specified by the manufacturer, further checking is indicated. The tube may require replacement or the B+ voltage may be low. Both defects are capable of reducing the d-c grid voltage.

If an oscillator grid voltage is present, we know the circuit is oscillating. Of course, this does not mean that the oscillator is on frequency, and it may very well happen that it is sufficiently far off frequency to give the same effect as though the circuit is inoperative. Usually this does not happen, but it can. A check of the oscillator frequency, using the method described in Chapter 15, will reveal whether or not this is so.

After the oscillator, the plate and grid voltages of the R.F. amplifier and mixer should be checked. Ninety-nine per cent of the difficulties arising in these stages will be found by one of the tests described. Also useful but not as widely employed is signal tracing. An A–M signal generator is connected to the input terminals of the receiver and is set to a frequency about 1 mc above the video carrier frequency for that channel. (The mid-frequency for the channel would be another good value to use.) A high-frequency probe in conjunction with a sensitive VTVM or an oscilloscope may then be used to trace the signal through the R.F. stages. If the VTVM is used, an unmodulated signal is obtained from the generator. If the oscilloscope is used, the signal must be modulated. The set oscillator should be disabled in order to insure that any signal picked up by the probe be that of the A–M generator only. Presence of the signal at the R.F. amplifier plate, at the mixer grid, and at the mixer plate can be determined by placing the probe at these points and noting what indications, if any, are obtained on the oscilloscope or VTVM.

To obtain definite indications, the signal generator should be capable of putting out at least 0.1 volt of signal. At the higher frequencies, this may not always be so. Also, it may not readily be possible to reach the desired points in the R.F. circuits since the tuner construction is usually quite compact.

The Video Circuits. The next group of circuits through which the signal passes are the video circuits. These include all of the video I.F.

stages, the video detector, and the video frequency amplifiers. Let us consider each, in turn.

Failure in the video I.F. system will cause the image to become distorted or to disappear entirely from the screen, but it may or may not affect the sound. If the sound passes through part of the video I.F. system, thus passing through the affected stage, then it will suffer together with the video signal. But if the sound is separated *before* the defective stage, then

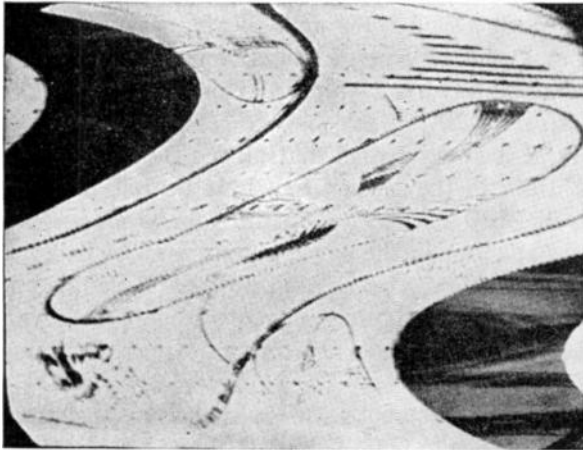


FIG. 16.6. The visual effect of 60-cycle voltage reaching the video signal path.

it will be heard without distortion. The importance of knowing where signal separation occurs in a receiver thus becomes evident.

As with other sections of the television receiver, tubes are the greatest offenders. Hence, these should be checked first, preferably by substitution. Gassy tubes and microphonic tubes are especially vulnerable to this method since they will not always be detected by a tube tester. Microphonic tubes in the video system produce a "bouncy" picture. The picture "bounces" whenever the set is jarred. The trouble may be so acute that even traffic rumbling by the house will disrupt the picture.

Gassy tubes produce a variety of effects, of which overloading appears to be the most common. When the stage becomes overloaded, the sync pulses are either depressed or clipped off entirely. This causes the set to lose synchronization. Tonal values in the picture frequently reverse so that a negative picture is produced. A gassy tube can cause a positive voltage to develop in the grid circuit. And if the A.G.C. network connects to this grid, the entire A.G.C. voltage will be disrupted. Finally, gassy tubes have been known to introduce 60-cycle hum from the filament to the cathode circuit. Visually, this will distort the picture as shown in Fig. 16-6.

Gassy tubes may sometimes reveal themselves by a bluish glow, although the absence of a glow does not mean that the tube is not gassy.

Another common tube trouble is a cathode-to-heater short. This will definitely introduce a 60-cycle hum in the picture, as shown in Fig. 16-6. Tube substitution will bring this to light, too.

In addition to tubes, any of the other components can be the cause of failure. This includes resistors, capacitors, and coils. To locate the defective unit, several methods of approach are possible. First, there is voltage checking. With a voltmeter or VTVM, measure the voltages at the screen, plate, and control grid of the various I.F. tubes. Then compare these with the manufacturer's values. This method is effective and frequently successful.

Incidentally, if it is discovered that all the B+ voltages in a system are low, excessive drain in the B+ system is indicated, and the trouble need not necessarily lie in the stages where the voltage measurements are being made. A filter capacitor in the power supply may be leaky, or the rectifier tube emission may be low, or a shorted by-pass capacitor may be at fault. Your job now may be the tedious one of disconnecting various branches leading off the B+ in an effort to determine the one containing the defective component.

Voltage tests are most successful in detecting open resistors, shorted capacitors, and incorrect A.G.C. voltages. This latter voltage is most important because an improper A.G.C. voltage can completely disrupt set operation. When this voltage is too low, the controlled stages run "wide open," resulting in signal overloading. When this voltage is too high (negatively), the tubes cut off and no signal gets through. Hence, pay particular attention to A.G.C. voltage. In fact, if you suspect that the trouble lies in the A.G.C. system, pull out the A.G.C. tube or, if this is impractical, remove the A.G.C. voltage possibly by breaking this circuit at the point where it leaves the A.G.C. tube. Then in its place insert a small 3-volt battery with the negative terminal connecting to the A.G.C. line and the positive terminal going to the chassis. If the set returns to normal operation, you know that the video system is all right and that the trouble lies in the A.G.C. circuits.

A defect which will not be found by voltage checks (such as open capacitors or shorted coils) may be brought to light by injecting a signal into the circuit. Take an A-M generator and connect it to the control grid of the last video I.F. amplifier. (This is the stage just before the video second detector.) Set the frequency to the mid I.F. value and turn on the modulation. If this stage is functioning (and all the others that follow it), then black and white horizontal bars will appear across the face of the picture.

The generator can be moved back, stage by stage, until the point of

failure is located. This method is simple, requires only an A-M signal generator, and is readily carried out.

If you possess a sensitive oscilloscope and an R.F. probe, you may be able to trace the video signal as it travels through the I.F. system. Connect the probe to the vertical input terminals of the oscilloscope and turn the vertical gain up as far as it will go. This places the oscilloscope in its most sensitive condition. Now touch the probe tip to the grid of the second video I.F. amplifier and see if you obtain the video signal waveform on the scope screen. A normal indication is shown in Fig. 16-7. (The same waveform, only reversed 180°, is also acceptable. The phase of the pattern depends upon the probe circuit and the number of vertical amplifiers in the oscilloscope.) If you obtain this, move on to the grid of the next I.F. amplifier and so on. Continue until you lose the signal or it becomes distorted.

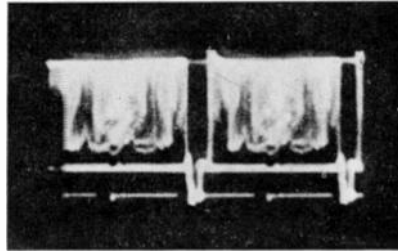


FIG. 16.7. The appearance of a normal video signal as seen on an oscilloscope screen.

When you reach the video second detector, the R.F. probe is no longer required. Detector action will be furnished now by the video detector.

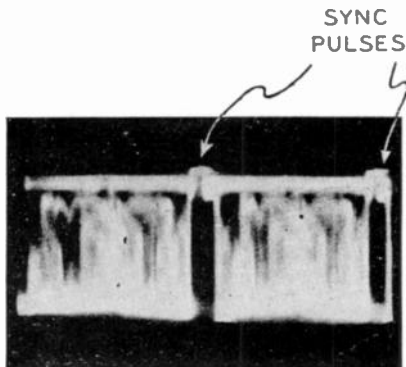


FIG. 16.8. A video signal in which the sync pulses have been partially suppressed. Note that the video signal extends up to the level of the pulses.

type sets this will affect only the picture. In Intercarrier sets, both video and audio will be missing. Partial defects seem to arise more frequently in germanium crystals than in vacuum tubes. The result may be a weak pic-

Signal tracing can be very effective, provided you have a sensitive scope and can distinguish between a normal-appearing video signal and a distorted one. Overloading in any of the video I.F. amplifiers as well as a-c hum in the signal is very clearly brought to light by this method. See Figs. 16-8 and 16-9. Signal tracing with an oscilloscope is used extensively in the video amplifiers (those beyond the video detector) and in the vertical and horizontal sweep systems.

In the video detector, a completely defective vacuum tube or germanium diode will prevent any signal from passing through it. In conventional-

ture lacking in contrast, or unstable vertical or horizontal lock in, or intermittent operation.

Following the video second detector are the video amplifiers.

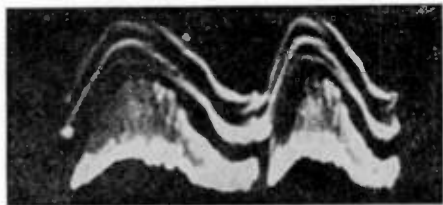
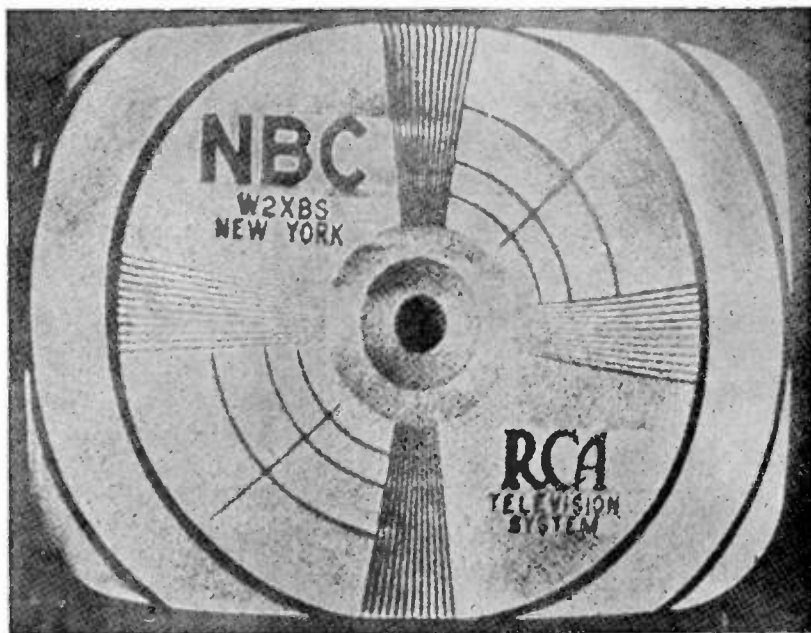


FIG. 16.9. 60-cycle hum in video signal.

Unless a video amplifier becomes completely inoperative, in which case no image at all is obtained on the screen, indications of other defects will be evident only by their effect on the image. With regard to a video-frequency amplifier tube and its circuit, the following defects may occur:

1. A defective low-frequency compensating network.
2. A defective high-frequency compensating network.
3. Improper voltages at the tube electrodes.
4. An inoperative tube or a defective component.



Courtesy RCA

FIG. 16.10. A visual indication of poor low-frequency response in video system.

When the low-frequency compensation network is defective, the larger objects in the image "smear." (See Fig. 16-10.) Check defective coupling and by-pass capacitors, or grid load resistors in the video amplifiers, such

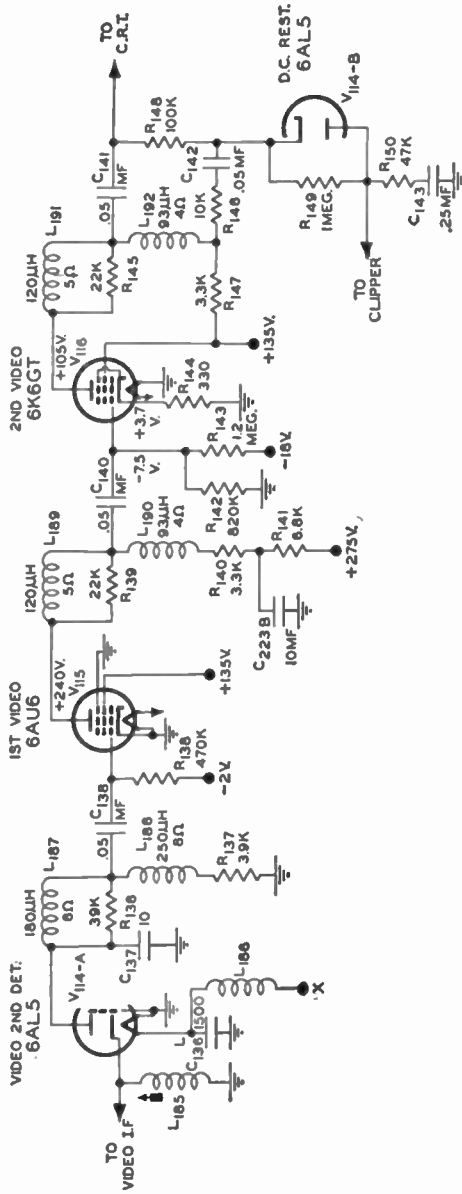


Fig. 16.11. A video-frequency amplifier system.

as V_{115} (Fig. 16-11). Check C_{138} , C_{140} , C_{141} , C_{223B} , R_{138} , R_{142} , R_{143} , R_{148} , etc. Smearing may be due to an overloaded video amplifier. Overloading of the video amplifier may also occur because of insufficient bias on the control grid, resulting in grid current. Check the bias and possible grid current of V_{115} and V_{116} .

Smearing can also occur when the bias voltage is too low. Hence measure the grid voltage at each video amplifier tube. Gassy tubes and leaky coupling capacitors are two causes for low grid bias voltage. Low plate and screen voltages produce smearing because the tube, under these conditions, is readily overloaded.

A more difficult defect to detect is the loss of fine detail due to poor high-frequency response of the video stages. High-frequency compensation is produced by the series and shunt peaking coils in the video amplifiers and by using small load resistors. The peaking coils are frequently shunted by a resistor to prevent them from sharply increasing the amplifier gain at their resonant frequency. Should this resistor increase considerably in value or perhaps open up, its shunting effect would be removed and transient oscillations would be developed in the coil and its distributed capacitance whenever the signal frequency fell within this range. The oscillations would appear on the screen as ghost lines (or multiple lines) after any small object or sharply defined line or edge in the picture. Their effect is most noticeable on test patterns.

When a peaking coil is not shunted by a resistor, and the coil opens up, the B+ path to the tube is broken and the stage becomes inoperative. On the other hand, when the coil is shunted by a resistor and the same condition occurs, one of two things can happen. In the case of a shunt peaking coil, the gain of the stage rises sharply because the load impedance on the tube has increased greatly. This causes the picture to become much brighter. However, the response of the stage is narrowed and the high-frequency response suffers. Picture quality is well below normal.

An open series peaking coil with a shunt resistor also results in a marked increase in stage gain. However, the signal transferred to the next stage is taken from a point below this resistor and so actually *less* signal is available for the next stage. As a result, the picture becomes dimmer, requiring an increased setting of the contrast control to bring it back to its former intensity. At the same time, the frequency response is reduced by the increased plate load and picture quality deteriorates.

The fine detail in a television image is produced by the higher video frequencies. Improper alignment or too narrow a response of the video I.F. amplifiers may be one cause. Other reasons are: defective video amplifier tubes, V_{115} and V_{116} ; defective peaking coils, L_{187} , L_{188} , L_{189} , L_{190} , L_{191} , and L_{192} , in Fig. 16-11; and defective interstage coupling capacitors, such as

C_{138} , C_{140} , and C_{141} . It is well to remember that lower-priced receivers are designed generally for less than the full 4.0-mc video response. Consequently, the image obtained in these sets will not contain as much detail as the images produced in more carefully designed receivers.

Another approach to video amplifier servicing when no signal is reaching the picture tube is to tune a station in and then follow the video signal from the video second detector to the point where the break occurs. The oscilloscope is an ideal instrument for this purpose. Place the vertical input leads of the scope across the plate load resistor of the video second detector. Set the scope sweeping frequency either to 30 cycles (to observe two fields)

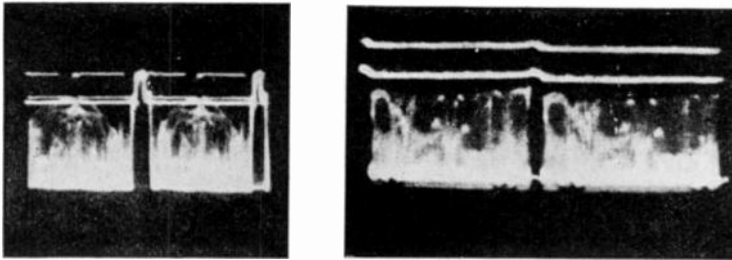


FIG. 16.12. Video signal showing (left) two lines and (right) two fields. To observe the lines, scope sweeping rate was set to 7875 cps; to observe two fields, scope rate was 30 cps.

or to 7875 cycles (to observe two lines). Typical field and line patterns of video signals are shown in Fig. 16-12. Whether the polarity of the observed signals is as shown or reversed depends upon how the detector is wired. In any event it is the presence of the signal which is important and not its polarity.

Once the signal is observed at the output of the second detector, it can be traced to the grid and plate of each succeeding video amplifier until it disappears or becomes distorted. Where this occurs represents the location of the defect, and voltage and resistance measurements should then bring it to light.

A variation of this signal tracing method is to *listen* for the signal instead of viewing it on an oscilloscope screen. This is made possible because of the presence of the low-frequency 60-cycle vertical sync pulses in the video signal. To hear them, simply take a 0.1- μ f capacitor and connect one end to the volume control in the audio section of the receiver. To the other end of the capacitor attach a long wire with an alligator clip. The clip can then serve as a probe, hooking into that portion of the video amplifier system where it is desired to check for the presence of the video signal. If the signal is present, a loud buzzing sound will be heard in the loudspeaker.

If the signal is absent, no sound will be obtained. By this simple expedient, breaks in the video signal path can be readily discovered.

Microphonic Tubes. One defect which arises from time to time and which has not been covered by the preceding material is due to microphonic tubes. A microphonic tube is one in which the elements are not rigidly fixed. Any vibration sets the elements in motion, generating transient currents. As long as nothing disturbs the elements, they operate normally. But, when the vibrations set up by the sound reach these tubes, transient voltages are generated. In the speaker, the familiar hollow sound is heard, while on the screen the image tends to flicker as if suddenly jerked out of synchronization. Some engineers describe this effect by saying that the picture seems to "bounce." Other visual effects of microphonic tubes are loss of synchronization and partial distortion of image details. Sometimes tapping helps the serviceman to locate the microphonic tube and sometimes individual replacement of each tube is necessary. Whatever the method, it follows closely the practice employed for present-day sound receivers. Noise due to resistors and partial shorts fall into the same category. ¹¹

SERVICING PICTURE TUBE CIRCUITS

The proper presentation of an image on the face of a picture tube depends first upon a signal being present at the grid (or cathode) of the tube, second that the correct d-c voltages are being applied to the various elements, third, that the neck-mounted coils and magnets are properly positioned, and, finally, that sufficient deflection power is available. Failure of any one of these conditions will either distort the picture or else result in its absence. Let us consider each category separately to see what its effect will be on the picture.

A. No Video Signal. In the absence of a signal, only a raster will be seen on the screen. The raster will tell you that the high voltage plus all the other d-c voltages are operating normally. It will tell you that the horizontal and vertical deflection systems are delivering the necessary deflection currents to the yoke. And from the sharpness of the raster lines and the absence of any shadows over the screen, you will know that the focus coil and ion trap magnet are correctly in place.

B. Loss of High Voltage. The surest indication of high-voltage failure is the appearance of a perfectly blank screen. If a blank screen is accompanied by normal sound output, then we know that the low-voltage power supply is operating and we can concentrate on the high-voltage system. However, a blank screen and no sound is generally due to a defective low-voltage supply and this should be tackled first.

High-voltage power supply failure will cause a dark screen but there

may be other reasons for the same condition. What these are will depend upon the manner in which the high voltage is generated. Thus, if the receiver possesses an R.F. high-voltage power supply, then a dark screen generally will be due to failure of the high voltage or a defective cathode-ray tube. If the filament of the picture tube is lit, check the high-voltage system. If a kilovoltmeter is available, measure the output of the high-voltage power supply. If the voltage is low or missing, replace the oscillator and/or high-voltage rectifier tubes. Measure the d-c voltages in the oscillator circuit. As a final step, make resistance (continuity) checks in the high-voltage system.

When the high voltage is developed by the inductive "flyback" method, the number of factors which can produce a blank screen increase. There is not only the high-voltage unit itself and the cathode-ray tube, but, in addition, the horizontal deflection system. To determine where the trouble exists in this system, first measure the voltage at the output of the high-voltage supply with a kilovoltmeter. If the voltage is zero, determine whether the horizontal deflection system is operating by checking the waveform at the grid of the horizontal output amplifier.

If the waveform and its peak-to-peak value are normal, check the waveform at the cathode of the tube. A normal indication here will indicate that the tube is functioning as it should. Concentrate now on the high-voltage rectifier circuit. Tube is checked first, then filter circuit components.

The absence of any deflection waveform indicates that the trouble exists in the horizontal deflection system. This system is best checked with an oscilloscope, noting the waveforms at various points, and comparing these with those given by the manufacturer. Note that a distorted waveform may still cause some high voltage to be produced, although this voltage will be low. A distorted deflection waveform will visibly affect the horizontal linearity of any image appearing on the screen.

It is well to distinguish between a dark screen, which is due to no high voltage, and a screen containing a scanning raster but no image. The latter difficulty, when accompanied by a normal functioning audio system, indicates a defective video system. In this case, the horizontal deflection system and the high-voltage power supply are both operating satisfactorily as revealed by the appearance of the scanning raster.

C. Defective Picture Tube. The picture tube is undoubtedly the single most expensive item in the television receiver. Because of this, the set owner is understandably very much concerned with its condition. Fortunately, these tubes are ruggedly built and most of them can be expected to last from 18 to 24 months. However, defects do arise and it behooves the serviceman to be familiar with them.

The brilliance of the raster (and the picture) depends upon the number of electrons striking the fluorescent screen. This, in turn, depends upon the bias between cathode and control grid and upon the number of electrons emitted from the cathode. If a tube is improperly constructed or after it has had long usage, it is possible that less than the normal number of electrons will be emitted from the cathode. The result will be a dim picture, even for advanced settings of the brightness control.

A tube with low emission should normally be replaced. However, it has been found that raising the filament voltage above its normal value will frequently cause enough additional electrons to be emitted to restore the picture brightness. In time, of course, it can be expected that the emission will again decrease below a usable value, but in the meantime useful service is being obtained from a normally unusable tube. Devices which enable the serviceman to increase picture-tube filament voltage are commercially available.

An incorrectly positioned ion trap can also be responsible for reduced brightness. Set owners who tamper with the set may be the reason for this. Or, in moving a set about the house, the ion trap magnet may have been jarred out of position. Or, an inexperienced technician may have had difficulty arranging the focus coil to obtain a shadowless picture and he finally "solved" the problem by shifting the ion trap magnet from its optimum position.

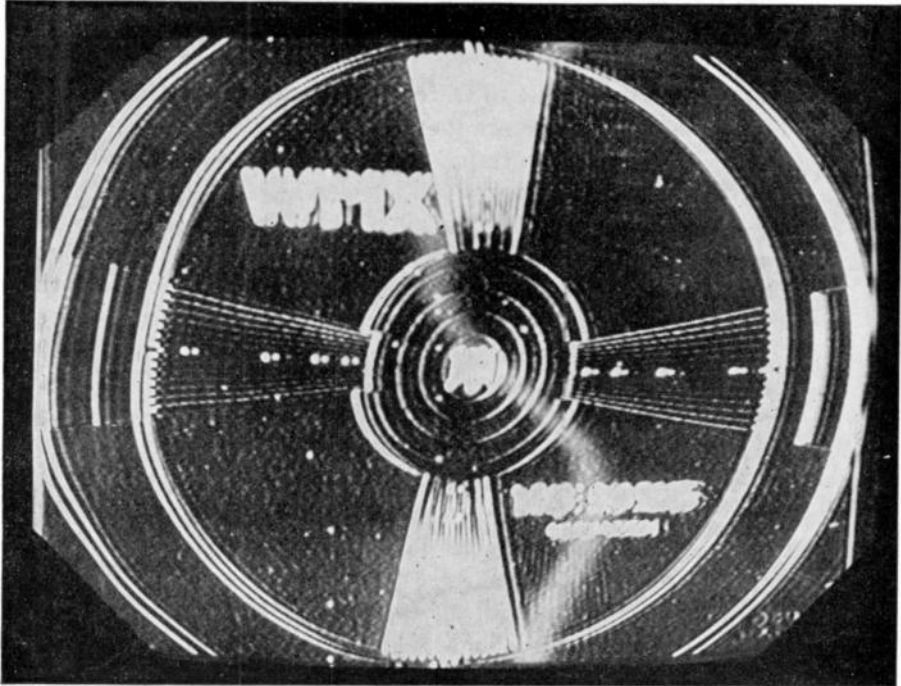
Incidentally, when the ion trap is not correctly positioned, it can lead to the eventual development of ion spots. An ion spot is an area of the screen where the picture brilliance is reduced and the picture is discolored. The color of these areas is brown and they appear most often in the center of the screen. However, they may develop at other points, too.

When a picture tube becomes gassy, several things can happen to it. It may develop negative pictures. See Fig. 16-13. Now, negative pictures may be due to a bad picture tube or to an excessively strong signal. If the signal is the cause, then negative pictures should be obtained only on the channel where this signal is being received. On all other channels, normal pictures should be seen. If the tube is at fault, then, of course, all stations will be affected.

One instance when negative pictures could be received on all stations and yet the tube not be defective arises when the A.G.C. network is defective and zero bias is placed on all controlled tubes. If all received signals are of moderate strength it is possible that, with the set running "wide open," all will cause overloading with subsequent negative pictures. Measuring the grid bias of the A.G.C.-controlled tubes will tell you whether the bias voltage is normal or not.

A tube which is somewhat gassy may sometimes cause picture blooming.

This is a condition in which the picture expands in all directions as the brightness control is turned up. Blooming is accompanied by loss of focus and sometimes by complete loss of picture and raster. The blooming or spreading out of the picture is due to a decrease in high voltage applied to the tube. If the tube is gassy, it may be drawing an excessive amount of current, and this drain on the high-voltage supply can cause a considerable reduction in high voltage.



Courtesy Sylvania News

FIG. 16.13. A negative picture in which all the tonal values are reversed.

Another cause of blooming is a rise in value of the filter resistor which is employed between the 1B3 high-voltage rectifier and the picture tube. A common value for this resistor is between 470,000 ohms and 1 meg. If this resistor rises sharply in value, it will decrease the high voltage made available to the tube and blooming will result. A defective high-voltage rectifier tube also can be responsible for blooming.

Another source of picture-tube trouble is a cathode-to-heater short. In sets where the cathode is externally connected to the filament, there will be no noticeable effect. However, in many television receiver circuits, the brightness control is placed in the cathode circuit of the picture tube. See Fig. 16-14. One side of the heater is usually grounded. Hence, when a

cathode-to-heater short develops, the cathode is placed at ground potential, and no variation of the brightness control will be able to alter this.

There are two solutions to this problem. One is to replace the picture tube. The other is to retain the same picture tube, but to use a special filament transformer for the picture tube heater. In this way the ground connection is removed from the heater and the d-c potential on the cathode is permitted to vary with the brightness control.

Note, however, that this particular tube could be used again without change in those television receivers where cathode and filament are connected together externally.

D. Picture Distortions Due to Defective Picture Tube Components.

Associated with the picture tube are the deflection yoke, the focus coil, and the ion trap. Improper placement of any of these components will have an adverse effect on the picture. By the same token, any defect within these components will also tend to distort the picture or even eliminate it entirely.

When a picture is out of focus, the cause may be either an improperly adjusted focus control or a defective focus coil or some resistor in the focus coil circuit. To focus the electron beam, the focus coil must have a certain amount of current flowing through it. Too much or too little current will produce defocusing.

If the set uses a permanent-magnet focus unit, poor focus indicates improper placement or possibly a weakened magnet.

In Fig. 16-15, the ion trap is out of position. In Fig. 16-16, the deflection yoke is not as close to the cone or bulb of the picture tube as it should be.

The illustration in Fig. 16-17 reveals distorted images which stem from the same source — a short in the deflection yoke.

Besides the two sets of windings in the deflection yoke housing, there will also be found several damping resistors (across the vertical windings) and a small capacitor (across one of the horizontal windings). Should any of these components become defective, picture distortion will be produced. Typical of the appearance of the image is Fig. 16-18, where there are several ripples at the left side of the screen, causing the pattern to appear wrinkled.

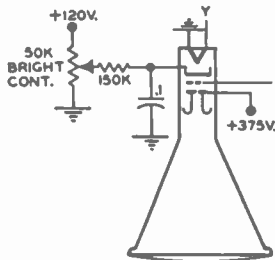


FIG. 16.14. A brightness control in the cathode circuit of a picture tube.

SERVICING LOW-VOLTAGE POWER SUPPLIES

Since the low-voltage power supply of a television receiver is, in many respects, similar to the power supply in a radio receiver, the same types of

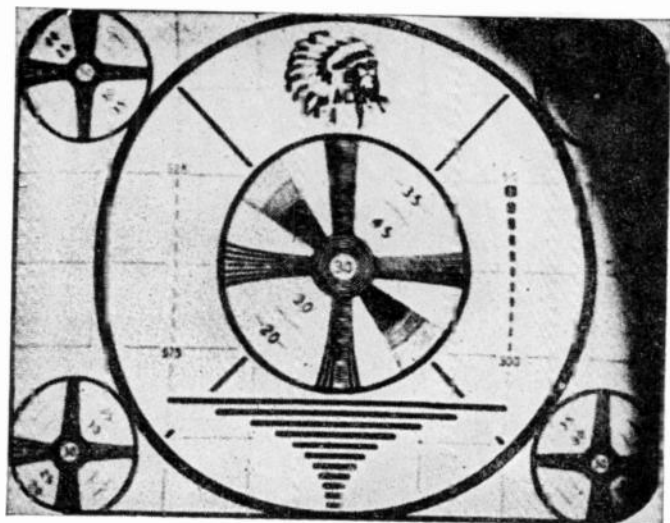


FIG. 16.15. Improper placement of ion trap.



FIG. 16.16. The shadow around the outer edge of the pattern is caused by a deflection yoke which is not as close to the cone of the tube as it will go.

Courtesy Sylvan's News

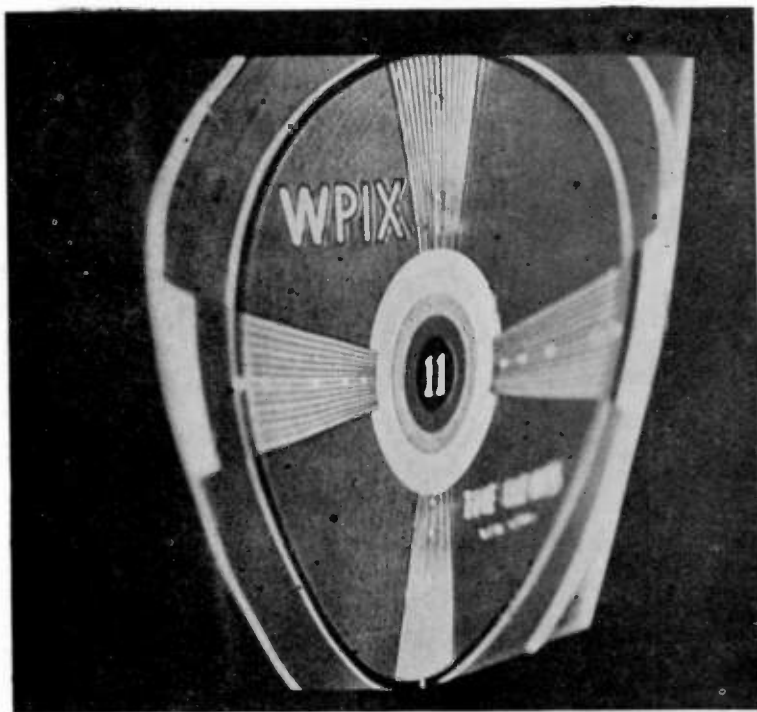
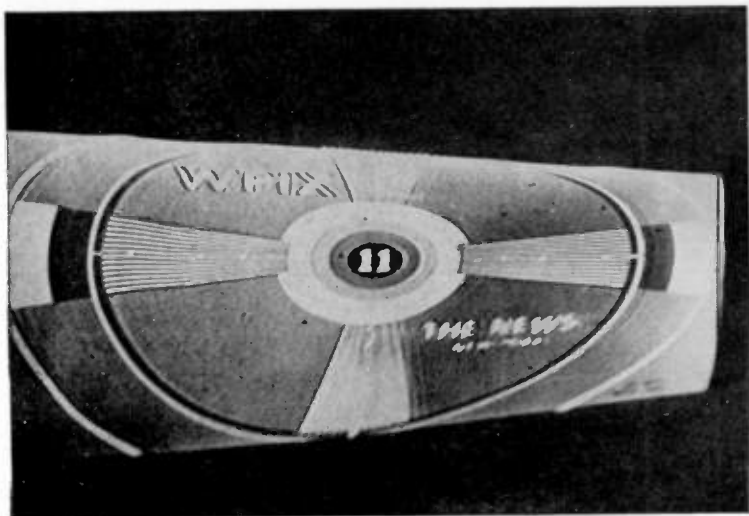


FIG. 16.17. Distortion produced by a short across part of the deflection coil windings. (A) Short circuit is across part of horizontal deflection yoke winding.



(B) Short is in vertical section of yoke winding.

troubles will be encountered in both. The following discussion will outline what effects can be expected from defective components.

A. Tubes. A rectifier tube is designed to pass a considerable amount of current, and in many television receivers these tubes are being pushed to the limit of their capacity. If, after being in use for some time, the emission level of the rectifier tube decreases somewhat, it will have an immediate



Courtesy Sylvania News

Fig. 16.18. Ripples caused by an open or wrong value of capacitor across the top half of the horizontal winding of the yoke.

effect on the value of voltage developed. Lowered voltages can lead to dim pictures, unstable picture hold-in, decreased volume and a decrease in set sensitivity, to name some of the more obvious results. It is frequently difficult from a number of symptoms to point directly to the power supply and say that here lies the cause of the trouble. However, where the trouble cannot be corrected by tube replacement, then a check should be made of the B+ voltage to determine if this possesses its normal value.

Besides lowered filament emission, rectifier tubes can burn out or short. If the set is not fused, a shorted rectifier tube can result in a burned-out transformer or even generate enough heat to set the receiver on fire. Most manufacturers fuse their sets, but it is astonishing how many do not take this simple precaution.

It will also be noted that some manufacturers, when they fuse their set, do not provide sufficient leeway for momentary current surges or for line voltages which are slightly above normal. Some servicemen deal with this situation by replacing the existing fuses in the set with others possessing higher current ratings. This is frequently dangerous unless it is definitely known that the circuit is under-fused. Generally the only source of this information is the manufacturer's service manager or his design engineers.

A better procedure and one which will generally be satisfactory is the use of a SLO-BLO or time lag fuse having the same rating as the original fuse. These fuses are designed to withstand momentary surges of current, but they will go if the overload current persists. Littlefuse makes the SLO-BLO fuse and Buss makes the time-lag Fusetrons.

B. Selenium Rectifiers. Selenium rectifiers are widely used in television receivers because they will perform the same function as vacuum tube rectifiers but with greater efficiency. They are small in size, rugged, comparatively cool in operation, and possess a long, useful life. However, from time to time, these units will become defective and require replacement.

Selenium rectifiers are sometimes tested by measuring their forward and reverse resistances. This is done by placing the ohmmeter leads across them and noting the resistance and then reversing the ohmmeter leads and again measuring the resistance. The readings will normally read greatly different values in each direction. However, what these resistance values are will depend on the ohmmeter voltage and scale. Because of this, ohmmeters should be used *only* to determine whether current can flow through the rectifier. They should never be used to gauge how well the unit will perform.

Whenever B+ voltage is below normal and no other obvious defects exist, many servicemen simply substitute other rectifiers for the ones in the receiver. In performing the changeover, the following precautions should be observed:

(1) The soldering iron or solder should not be brought into contact with the rectifier plates. Also, do not keep the iron applied to the soldering terminals any longer than necessary.

(2) Mount the selenium rectifier in the same manner as its predecessor. See that it receives adequate ventilation.

C. Electrolytic Capacitors. Electrolytic capacitors in the low-voltage power supply provide most of the filtering. They are not a permanent part of electronic equipment and, like the vacuum tube and the battery, must be replaced periodically. This inconvenience is more than compensated for by the small-volume, large-capacitance characteristics of the electrolytic capacitor.

When making a visual inspection of a receiver during any servicing job, the serviceman should check the filter capacitors for signs of corrosion, dripping, or sealing around the base of the capacitor. The capacitor may still work perfectly, but these signs will indicate that the capacitor is about to fail. It should be replaced. The failure of a filter capacitor usually produces excessive hum (in the speaker), low-volume output, and heavy black bars on the face of the screen.

Effects of Shorted or Leaky Filter Capacitors. A shorted filter capacitor can cause the rectifier tube to burn out. It will cause the fuses to blow (if the circuit is fused).

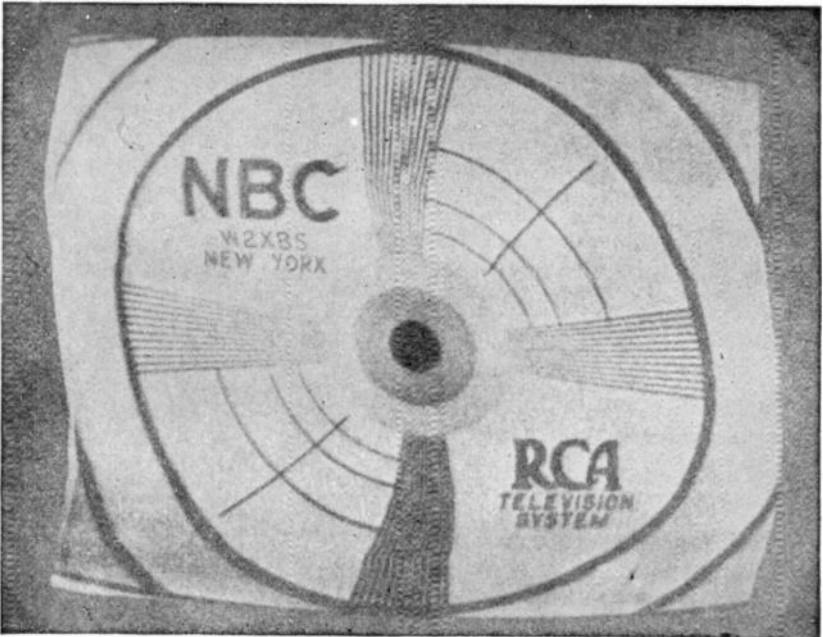
With a leaky filter capacitor, the rectifier filament glows brighter than normal. The voltage output of the power supply is low, there is hum in the speaker, dark horizontal bars appear on the pictures, and the transformers and chokes overheat.

Effects of Open Filter Capacitors. An open filter capacitor produces most of those symptoms caused by leaky capacitors except excessive rectifier filament brightness and the heating of parts. It will lower the B+ voltage. This is especially true if the capacitor in question is the input capacitor of the filter network. If the capacitors in the filters are in parallel, one open capacitor may not noticeably affect the operation of the receiver. (This will depend upon the design tolerances of the receiver. In television receivers it is very important that the power supply be well filtered. Therefore, the power supplies frequently contain enough capacitance so that failure of one unit may not be noticed.)

Open Filter Chokes. Open filter chokes can be isolated quite simply. For example, if the filter choke of Fig. 10-3 were to open, there would be no output voltage and a higher than normal voltage across C_{220} and C_{221} . When operating normally, the filter choke should be cool to warm. It should not become so hot that the hand cannot be held on it for several minutes.

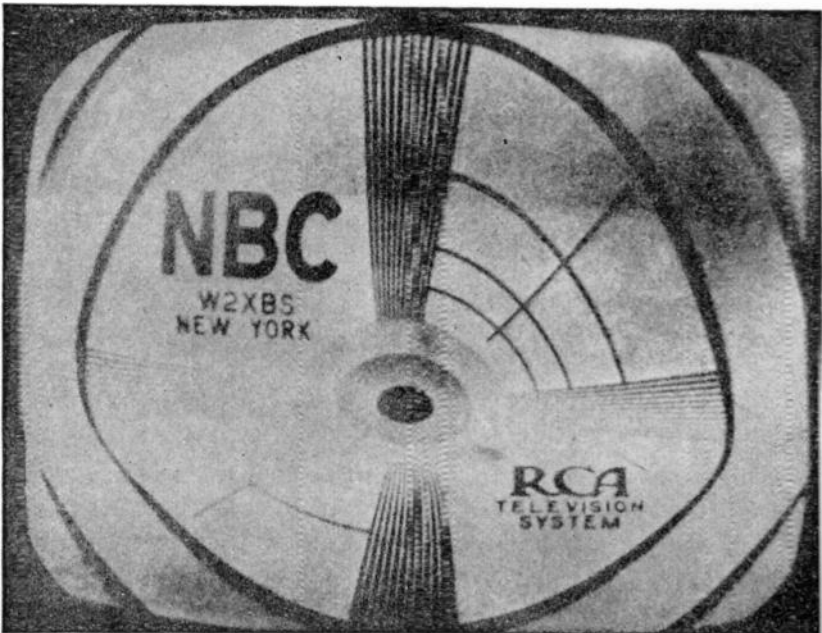
One visual effect of an improperly filtered power supply can be seen from Figs. 16-19, 16-20, and 16-21. The distortion in Fig. 16-19 is due to ripple in the voltage reaching the horizontal deflection system. The image weaves slowly from side to side. When the ripple frequency is 60 cycles, the pattern shown is observed. The curvature at the edge of the image represents one cycle of a sine wave laid on its side. If the ripple voltage is raised to 120 cycles, then the number of "bends" is similarly doubled. Fig. 16-20 indicates the visual effect of ripple in the vertical deflection system. This is an alternate spreading and crowding of the image in the vertical direction. Generally this distortion is not stationary, but moves slowly up and down the image.

The final illustration (Fig. 16-21) concerns ripple in the video system.



Courtesy RCA

FIG. 16.19. Distortion due to ripple in horizontal deflection system.



Courtesy RCA

FIG. 16.20. Unfiltered A.C. ripple in vertical deflection amplifiers.

One or two heavy bars obscure the image. The bars may be stationary or they may move slowly up or down.

Sound vs. Filter Ripple. Sound signals reaching the cathode-ray tube produce an effect which is similar in certain respects to a-c ripple. Each defect causes black bars to appear across the screen, but those due to the sound voltages are more numerous and their intensity changes in step with

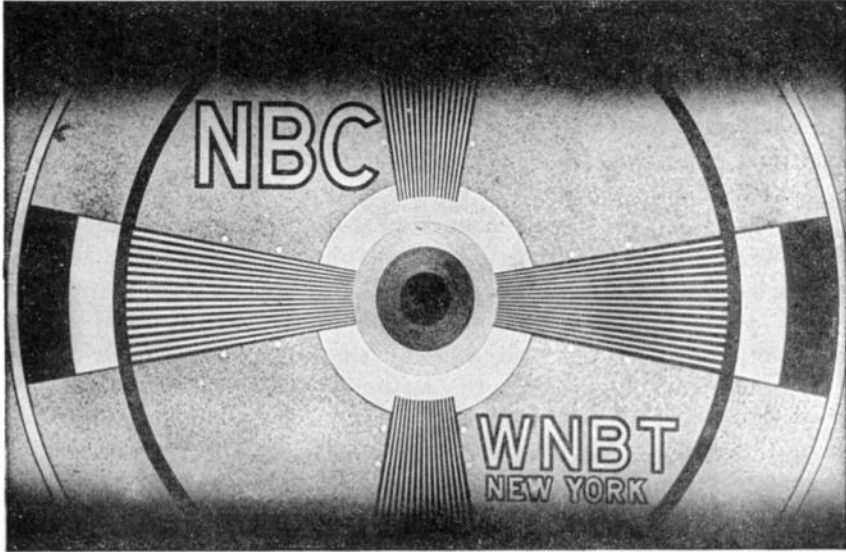


FIG. 16.21. A large percentage of ripple voltage in video amplifiers.

the amplitude of the applied audio. The black bars produced by hum in the power supply seldom exceed one or two in number, are much wider than the audio bars, do not vary in intensity, and will be visible on all channels.

Sound voltages reaching the cathode-ray tube may arise from two sources: improper alignment of the trap circuits in the video I.F. system or incorrect setting of the fine-tuning control. As a first step toward removing the sound bars, adjust the fine-tuning control. In a normally functioning receiver, this adjustment should be enough to remove the sound bars from the image without distorting the sound output. If, however, it is found that the sound bars cannot be removed or, when they are removed, the sound becomes fuzzy, then realignment of the circuits is necessary.

Transformerless Receivers. Television receivers which use selenium rectifiers do not, in many cases, use a power transformer. This means that the serviceman must be careful not to ground the chassis of such receivers unless he is certain that the side of the power line to which the chassis is

connected is also at ground potential. Unless this precaution is observed, it is quite possible that a fuse will be blown and any equipment connected to the receiver will be damaged. For this reason, isolation transformers should always be used when servicing transformerless receivers.

SERVICING HIGH-VOLTAGE POWER SUPPLIES

A good indication of high-voltage failure is the appearance of a dark screen. If this is accompanied by normal sound, then we know that the low-voltage power supply is operating normally and we can concentrate on the high-voltage system. However, a blank screen and no audio is generally due to a defective low-voltage supply.

As a first step in checking the high-voltage supply, measure the voltage at the end of the second anode connector (after the connector has been removed from the tube). If there is no high voltage, a likely source of trouble is the high-voltage rectifier. Substitute another tube and see if the voltage returns. If the tube is not at fault, then the next step depends upon the manner in which the high voltage is developed. In the fly-back type of supply, a check should be made to determine whether the proper deflection voltage is reaching the grid of the horizontal output amplifier. This is done with an oscilloscope. If this wave does not have the proper shape and peak-to-peak amplitude, work back towards the sweep oscillator to find out why the proper wave is not being developed.

When the drive voltage is normal, a check should be made of the horizontal output tube. Also, the damper tube may be faulty. The important thing is to determine whether the trouble is arising from the output circuit or from insufficient drive on the output amplifier. This will then direct you to the proper circuit.

Within the high-voltage system the usual causes of failure are the tube, the filter capacitor, and the series filter resistor. If the capacitor shorts, there will be no high voltage. If it opens, there will be some high voltage, but not very much. A common trouble developed by the series filter resistor is an increase in value. This does not remove the high voltage, but it does decrease it and produces an effect known as blooming. As the brightness control is turned up, the picture begins to lose brilliance and starts expanding. It is possible for the picture to disappear altogether at some setting of the brightness control.

A rectifier tube with low emission can also be responsible for blooming.

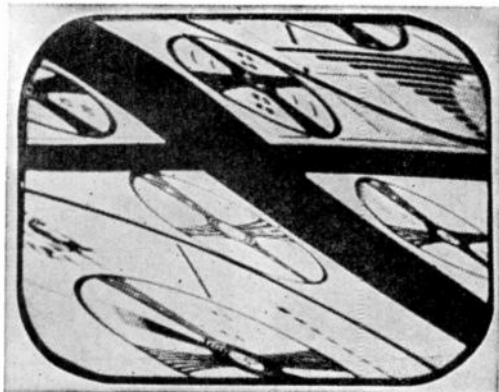
R.F. Power Supply Servicing. A defective rectifier tube, filter capacitor, or filter resistor will produce the same effects in R.F.-type power supplies as they do in the flyback type. Where the R.F. supply does differ is the manner in which it develops its high voltage. The R.F. power supply is a

self-contained unit, requiring only low voltage from the television receiver. Therefore, what happens in the horizontal deflection system is usually of no immediate importance to the R.F. power supply. When high voltage is missing and the low-voltage B + is normal, then the trouble lies within the high-voltage circuit. Tubes should be tested first, preferably by substitution. Next, determine if the R.F. oscillator is operating. Measure the grid bias on the oscillator tube. If it is low, oscillations are weak. Try adjusting the trimmer capacitor across the oscillator coil to see if the amplitude of oscillations can be increased. Do this carefully, for frequently it is easy to throw the circuit out of oscillation. If this does not help, check circuit resistances and capacitors. Sometimes a number of turns in the oscillator coil become shorted. This may or may not be detectable by a resistance check because of the low coil resistance.

SYNC SEPARATOR STAGES

The sync separator stages stand between the vertical and horizontal sweep systems and the video system.

A defect in the sync separator stages of a television receiver will almost always cause both vertical and horizontal sections to fall out of sync. The picture on the screen will be found to roll aimlessly vertically and horizontally. See Fig. 16-22. While it may be possible to lock the picture



Courtesy G.E.

FIG. 16.22. Loss of vertical and horizontal lock-in.

in momentarily by adjusting the vertical and horizontal hold controls, the picture will soon fall out of sync again.

The best way to check through a sync circuit in search of a defect is by tuning in a signal and then using an oscilloscope to check pulse waveforms at the grid and plate of each of the sync stages.

The point at which to start is the input to the sync separator section. The waveform here will most generally be the video signal complete with video voltage and sync pulses. See Fig. 16-23A. At the output of the first sync separator (V_{118}) (for a system such as that shown in Fig. 15-1)

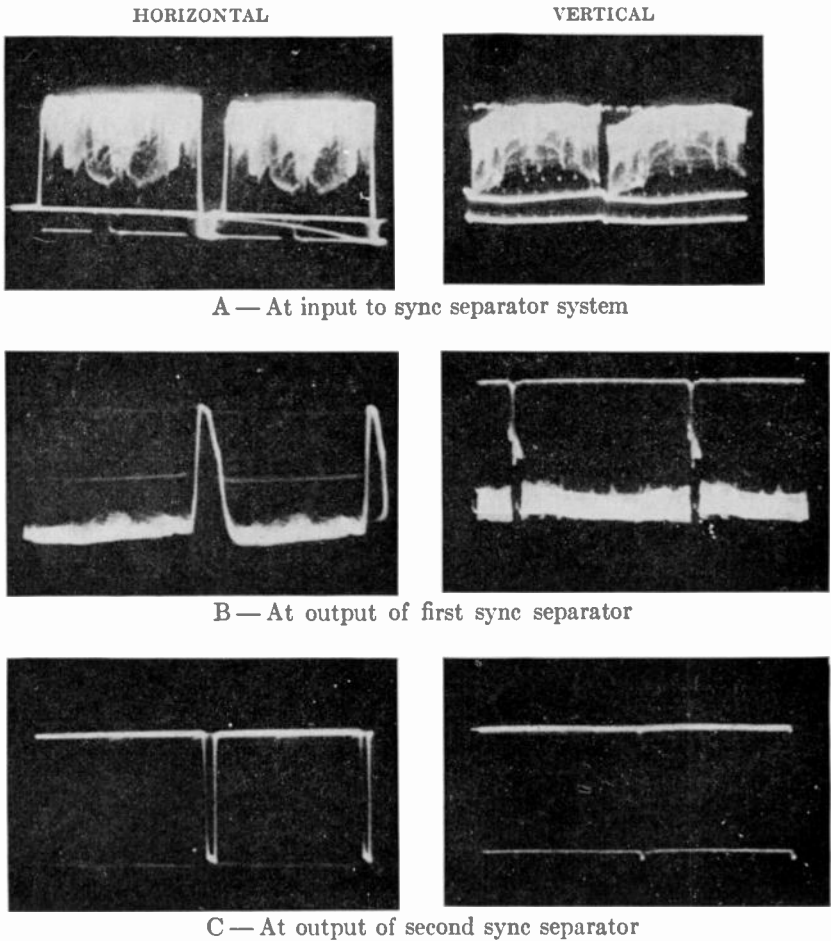


FIG. 16.23. An illustration of how the sync separator stages separate the sync pulses from the video signal.

the vertical and horizontal waveforms will appear as shown in Fig. 16-23B. (The oscilloscope sweeping rate to observe the vertical sync pulses is 30 cycles; for the horizontal sync pulses it is 7875 cycles.) Note how much the video signal has been suppressed or even eliminated while the sync pulses have been amplified. At the output of the second sync separator stage, V_{119} , the pulses are more clearly defined and the percentage of video signal

has been further reduced. See Fig. 16-23C. The separation of sync pulse from video signal becomes substantially complete at the plate of the third sync separator stage, V_{120A} . The sync pulses are now ready to be transferred to their respective sweep systems.

In viewing sync pulses on an oscilloscope screen, the serviceman will find that the horizontal pulses stand out clearer and more distinctly than the vertical pulses. One reason for this is that the horizontal pulse is simpler in structure than the serrated vertical pulse. Also, the horizontal sync pulse occupies a greater proportion of a line than the vertical sync pulse does of a field. Hence there is more of the pulse to be observed when the scope scanning rate is set to the proper value. These facts are borne out by the oscillograms of vertical and horizontal sync pulses shown in Fig. 16-23.

It is important to observe carefully the composite video signal which is applied to the input of the sync separator section. If, for example, it is found that the sync pulses have been compressed (see Fig. 16-24) and it is difficult to keep the picture in sync, then it can be surmised that it is not the sync separator stages which are at fault, but the preceding video system. It is quite possible that the defect exists in the

A.G.C. network, wherein the controlled R.F. and I.F. amplifiers are being permitted to operate at higher than normal gain due to insufficient bias. This can readily lead to overloading with subsequent sync pulse suppression. When this signal reaches the sync separators, there is a considerable amount of video signal operating at the same level as the sync pulses, and consequently it becomes impossible to effect a clear-cut separation. In the vertical system, this will show up as unstable or critical hold-in. The picture will have a tendency to roll. In the horizontal system, the automatic frequency control network may hold the picture more rigidly in place. However, the hold-in range will undoubtedly be smaller than normal. Also, a bend may appear at the top of the picture. Vertical lines or objects in the picture will be found to curve to the right or left. In extreme cases, the top of the picture will "flag-wave," or move rapidly from side to side.

If it is found that sync pulses are being compressed or clipped entirely before they reach the sync separator stages, then the path of the video signal should be traced back to the video second detector with the oscilloscope.

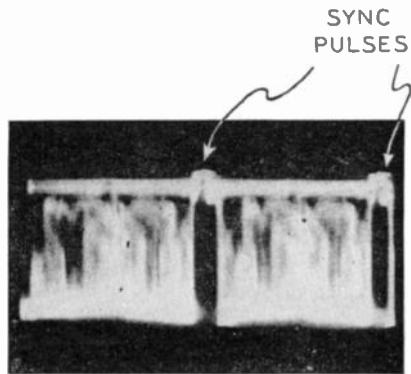


FIG. 16.24. A video signal in which the sync pulses have been partially suppressed. Note that the video signal extends up to the level of the pulses.

Sync limiting or compression may occur at almost any point from the R.F. amplifier to the sync take-off point in the video amplifiers. A gassy tube, a defective A.G.C. system, a signal which is too strong, a change in operating voltages or in component values, all can be responsible. Examine the signal in the video second detector. If it looks normal, then the preceding R.F. and I.F. stages are operating normally. If the sync pulses are absent or have been compressed, the trouble exists prior to the video second detector. Check the plate, screen, and grid bias voltages on the video I.F. tubes first. Pay particular attention to low A.G.C. voltages since these are a frequent source of trouble. Gassy tubes — sometimes identifiable by a blue glow — are also common causes. So are cathode-to-heater shorts. The servicing procedure now follows along the lines previously outlined for video I.F. amplifiers.

When poor sync action has been traced to the sync separator stages, the oscilloscope is employed first to localize the defective stage. Thereafter, voltage and resistance checks will be required to isolate the component. Of course, in *any* section of the receiver, tubes should be checked first. If this does not clear up the trouble, we then turn to voltage and resistance measurements.

VERTICAL DEFLECTION SYSTEM

Difficulties in the vertical deflection system are perhaps the easiest to analyze because the voltages developed here deal only with the vertical sections of the image. There are no high-voltage power supplies associated with this system, such as we find in the horizontal deflection system. When only the vertical deflection system is affected, analysis of the source of the defect is simple and straightforward.

1. The most positive indication of complete failure of the vertical system is the appearance of a narrow, horizontal line on the cathode-ray tube screen. See Fig. 16-25. The first components to check are the vertical oscillator and amplifier tubes. If these are working, check the vertical sweep waveforms with an oscilloscope, starting at the vertical oscillator tube and progressing forward to the vertical output amplifier. Once the defective stage is located, voltage and resistance checks should reveal the defective component. Keep in mind, when making the foregoing tests, that an open in the windings of the vertical deflection coils is also a possibility. This does not usually occur, but it can.

2. Failure of picture to lock-in vertically. See Fig. 16-26. Ordinarily, if the vertical hold control is rotated, a point will be found where the frequency of the oscillator is brought close enough to the incoming vertical synchronizing pulses to permit lock-in. The picture then becomes stable. However, if the oscillator is not functioning properly, or the vertical syn-

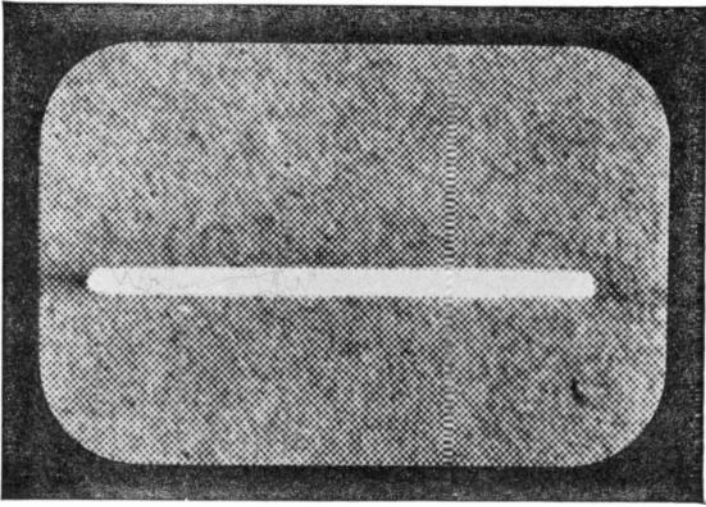


FIG. 16.25. Complete loss of vertical deflection.

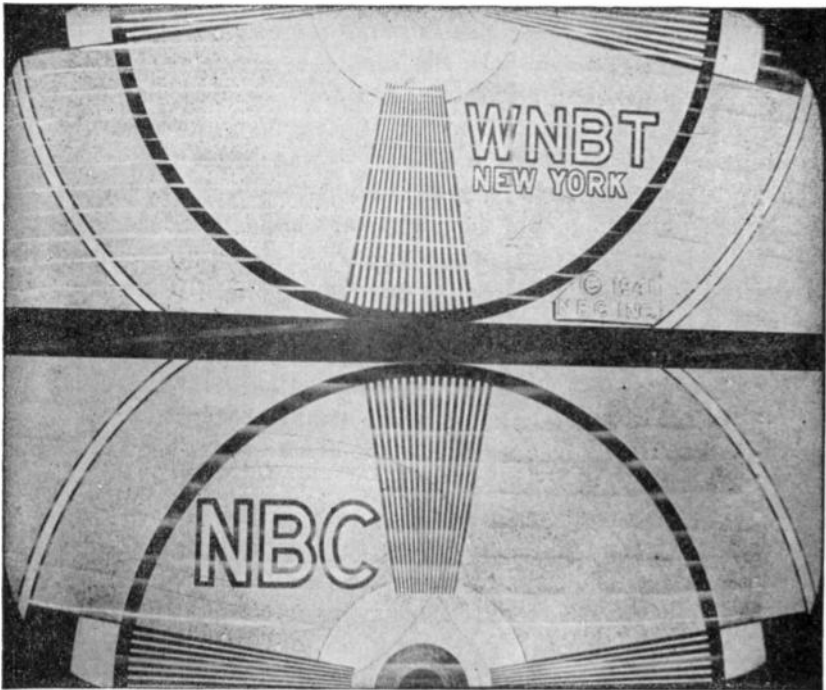


FIG. 16.26. Incorrect setting of vertical hold control. This picture was retouched slightly to bring out the loss in interlacing that generally accompanies this condition.

chronizing pulses are not reaching the oscillator, then perfect lock-in will not occur. Hence, when the vertical hold control proves ineffective, make the following tests:

a. Check the waveform at the input to the vertical sweep oscillator. Determine whether the pulses reaching the vertical sweep oscillator are sufficiently strong to maintain control. It is quite possible that the path from the sync pulse separator to the vertical oscillator does not transmit the full vertical pulse. Reasons for this may be defective coupling capacitors, open resistors, or components which have changed value appreciably. Resistors can be checked by measurement and capacitors by substitution. Check, too, the amplitude of the complete video signal at the point where the sync pulse separation occurs. The image on the screen will also help determine whether sufficient signal strength is present.

b. If the foregoing test produces normal results, check the waveform at the output of the vertical sweep oscillator. Note whether variation of the vertical hold control has any effect on the frequency of the observed waveform. A lack of such frequency variation indicates an open resistor in the grid circuit containing the hold control. Check the resistors and capacitors for the defective component.

c. Sometimes poor vertical lock-in comes from a completely unsuspected source. In one set brought into the shop the picture could not be held vertically. All tubes and suspected parts were checked and found to be O.K. However, when the serviceman started repositioning some of the wires near the vertical sweep oscillator, it was found that a lead from the vertical system had moved near the filament leads of another tube. The 60-cycle field radiated by the filament wires induced sufficient voltage in the vertical sweep oscillator circuit to trigger this circuit prematurely, resulting in lock-in instability. When the offending filament leads were dressed away from the vertical sweep oscillator, the set returned to normal operation.

3. Picture "bounces" when set is jarred. Check all the tubes in the vertical synchronizing system for microphonics. Mere tapping of each tube, while in the set, is inconclusive. The best test is to replace each tube, in turn, with one known to be good.

Incidentally, too strong a signal or too high a setting of the contrast control can produce a jumpy picture. When this condition is present, the tubes are driven to saturation, compressing or otherwise distorting the synchronizing pulse waveforms. The result is poor sync control of the sweep oscillators. In sets where the horizontal system employs some form of automatic control, noise disturbances or weak signal will affect the vertical sweep system to a greater degree than the horizontal system. However, when such control is absent, both will be affected almost equally.

4. Picture height insufficient. In a normally functioning receiver, rotation of the height control will produce the proper picture height. Inability to obtain this result may be due to one or more of the following reasons:

a. Weak vertical output amplifier tube.

b. Lowered plate and screen voltages on vertical oscillator and vertical amplifier tubes.

c. Low line voltage.

d. Improper placement of the deflection yoke.

Items *c* and *d* will affect the width of the image, too. Hence, if the width is normal, these items can be disregarded.

5. Picture compressed at top. This is an indication of poor linearity of the vertical deflection voltage. The vertical saw-tooth deflection wave is developed in the plate circuit of the vertical sweep oscillator, amplified by the vertical output amplifier, and then applied to the vertical deflection coils. If this wave does not have the proper form, if portions of it curve or bend more than they should, then the electron beam in the cathode-ray tube will not travel down at an even rate. The visual result will be a bunching of lines in some sections of the image and the spreading apart in others. This is known as poor linearity. When the image displays this type of distortion, the vertical sweep and amplifier circuits contain the trouble.

To correct the defect, try adjusting the vertical linearity control. Most receivers contain such a control. If these adjustments fail, change the vertical output tube. If the nonlinearity still persists, check (preferably replace) the capacitor and/or resistor which develop the deflection wave. In Fig. 16-27, these would be C_{101} , and R_{82} . Check, too, R_{87} , R_{131} , and C_{105} , because these also influence the shape of the deflection wave. Finally, grid, plate, and screen voltages of the vertical output amplifier should be checked against the manufacturer's recommended values. Line overlapping and image compression can arise from lowered tube potentials.

A precise vertical linearity check can be made with an audio oscillator. Place the receiver in operation but have it tuned so that no signal is being received. This will leave the screen with a blank raster. Now connect the audio oscillator across the load resistor of the video second detector and set the frequency to 660 cycles. As the output of the generator is turned up, a series of alternate black and white stripes will appear. See Fig. 16-28. If the bars are evenly spaced, we know that the scanning of the screen is linear. On the other hand, a non-linear scanning rate will cause the bars to bunch together at some points.

The figure 660 was suggested because it is an integral multiple of the 60-cycle vertical scanning rate. Almost any multiple of 60 may be used, although it is desirable to have at least 10 or more bars on the screen.

The audio signal was injected at the video second detector because this same signal will also lock-in the vertical sweep system and produce a steady pattern. The sync signals are almost always taken from some point beyond

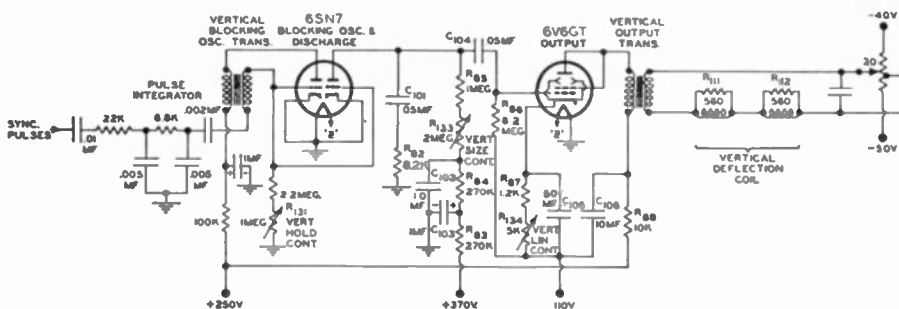


Fig. 16.27. A vertical deflection system.

the second detector and in this way part of the injected signal will reach the vertical sweep oscillator and lock it in.

6. Inability to center image vertically. (A similar analysis may also be applied to the horizontal centering control.)

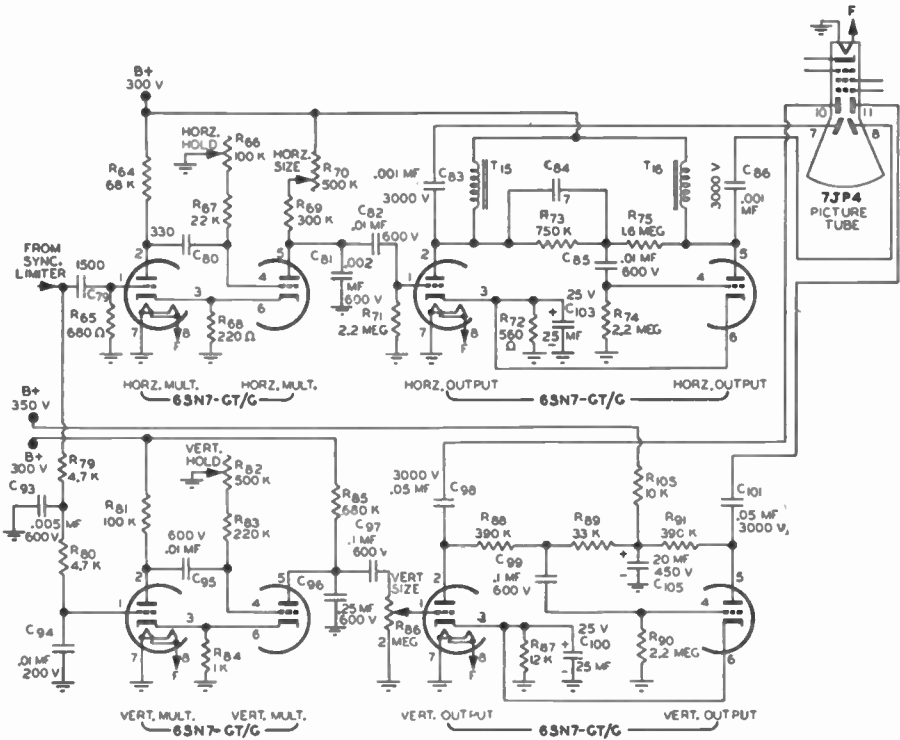


Fig. 16.28. Black and white horizontal bars for checking vertical linearity.

a. Electrostatic Deflection C.R. Tubes. The circuit shown in Fig. 16-29 will be used to illustrate the common troubles encountered in such circuits.

With all power off, check coupling capacitors C_{98} and C_{101} for leakage, using a capacitor checker. On one side of each of these capacitors we have high voltage; on the other side, low voltage. Any leakage will result in a

flow of current between capacitor plates, effectively reducing the high voltage that reaches the cathode-ray tube deflection plate connected to that capacitor. If the other coupling capacitor is good, the other deflection plate will receive the proper amount of positive high voltage and swing the elec-



Courtesy Belmont Radio

FIG. 16.29. The vertical and horizontal deflection circuits for an electrostatic deflection tube.

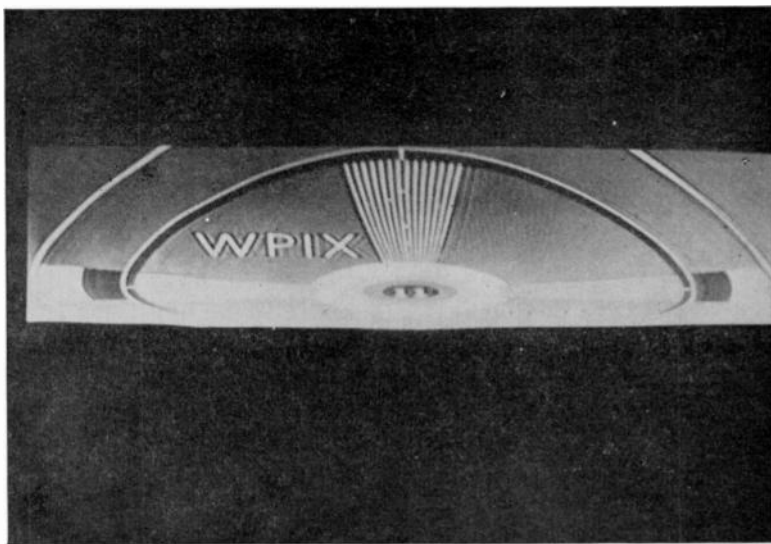
tron beam sharply to its side. The centering control will then be ineffective.

If the coupling capacitors check O.K., measure the resistances and voltages in the vertical centering circuit (not shown here).

Electrostatic deflection tubes sometimes are enclosed in a magnetic shield. If this shield should become magnetized, then its magnetic field may have sufficient strength to deflect the beam and prevent the centering controls from bringing it back to the center of the screen. Should the shield become magnetized, a de-magnetizer may be made by removing the laminations from one side of a filter choke, connecting the choke to the 110-volt line, and passing the open end of the choke coil over the surface of the shield.

The rapidly reversing field of the choke should remove all the magnetization of the shield. (Shields which are magnetized will attract magnetic materials similar to any other magnet.)

b. *Electromagnetic Deflection C.R. Tubes.* Test for open resistors in centering network. If the position of the focus coil is the sole centering method in the receiver, then adjust the focus coil position. Where a centering potentiometer is employed, check the voltage across this control. If



Courtesy Sylvania Electric

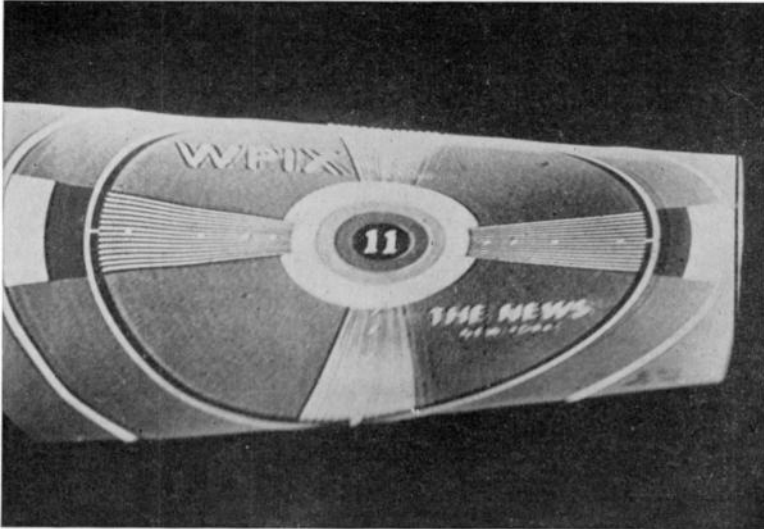
FIG. 16.30. An illustration of vertical foldover.

this is low, measure the voltage of the low-voltage power supply. A low voltage at either point will produce a restricted centering range. Finally, if all these tests do not remove the trouble, check the cathode-ray tube itself. The electron gun may have become tilted because of some sudden jarring of the tube.

7. Vertical foldover in a picture might be considered as an aggravated case of non-linearity. In form, the picture will appear somewhat as shown in Fig. 16-30. The bright horizontal stripe across the bottom of the image represents the point where the scanning beam stopped moving downward. Causes of this condition generally arise from some defect between the output circuit of the vertical oscillator and the grid circuit of the output tube. Thus, a leaky coupling capacitor between oscillator and output amplifier or a substantial decrease in the value of the grid resistor will often be responsible. It might also be wise to check the filter capacitor in the cathode leg of the output amplifier. Finally, a gassy output tube is also a distinct possibility and a new tube should be tried.

Sometimes a condition will be obtained in which there is a bright horizontal line across some intermediate point in the picture. This is usually due to heater-to-cathode leakage in the vertical output tube.

8. Shorted turns in the vertical output transformer will result in a loss in picture height. It may sometimes give the same indication as insufficient driving voltage to the vertical output amplifier. This type of trouble is usually difficult to detect because it reveals itself neither to normal voltage nor resistance checks. If the shape and peak-to-peak amplitude of the



Courtesy Sylvania Electric

FIG. 16.31. Vertical keystoneing.

deflection wave are correct at the input to the vertical output amplifier and if the amplifier appears to be operating normally, then the output transformer could be a logical suspect.

Measuring the peak-to-peak value of a wave can be a very powerful servicing tool, especially in the sweep sections of a television receiver. Make it a habit to measure this value at the input to the output amplifier. Then compare the value you get with that recommended by the manufacturer.

Shorted turns in the vertical deflection coils will produce the trapezoidal or keystone-shaped pattern shown in Fig. 16-31. (Which side is wider depends upon which section of the coils contains the short.)

HORIZONTAL DEFLECTION SYSTEM

It is evident from what has been discussed in Chapter 13 that distinctive waveforms are produced at each point in the vertical and horizontal de-

lection systems. Obvious, too, is the fact that the reception of a signal does not appreciably alter the shape of these waves.

What better method is then available than comparing the wave shapes of the voltages in the defective receiver with the corresponding waveforms given by the manufacturer in his service manual. For example, refer to the waveforms shown in Fig. 16-32. These are the proper waveforms present in the vertical and horizontal circuit of this receiver when everything is operating normally. The peak-to-peak voltage values are also important in waveform checking and these should be carefully noted. When the wave amplitudes are found to be appreciably smaller than recommended by the manufacturer, tubes and the B+ power supply voltages should be checked. On the other hand, distorted or improperly shaped waves usually indicate defective capacitors and/or resistors in the circuits.

It has been stated before and it will be stated again that tubes are, by far, the most important source of trouble in present-day television receivers. Hence, when trouble is traced to the deflection systems, check the tubes first. If the tubes test O.K., then the next job is waveform checking.

Take an oscilloscope and connect the grounded vertical input terminal to the receiver chassis. Connect a test prod to the other vertical input terminal. Then, starting at the plate of the synchronizing oscillator, check plate and grid waveforms of each tube, working toward the deflection plates or coils of the cathode-ray tube. At the point where the waves disappear or are not in their proper form, voltage and component checks should be made to determine the reason for the wave change or disappearance. It is desirable to use a ruled plastic mask for the oscilloscope screen and note approximately the peak-to-peak voltage values of each of the waveforms checked. Variations of 10 to 15 per cent from the recommended values can be accepted since the adjustment of various controls can readily affect the wave amplitude by this amount.

Loss of Synchronization. Loss of horizontal synchronization is indicated when the picture assumes the distorted appearance of Fig. 16-33. In the latter, the image appears to slip in a horizontal direction. If the slippage is great enough, it results in several images overlapping, and no detail can be clearly distinguished. In the absence of too large an input signal, slow rotation of the horizontal hold control will bring the oscillator back to the correct operating frequency at which one stationary image is again visible. Failure of the hold control to correct the oscillator frequency generally indicates a faulty component in this generator or its automatic control network.

Other Horizontal Defects. A fairly common type of picture defect, and one which gives many technicians a considerable amount of trouble, is horizontal non-linearity. The picture may be compressed or elongated at

the left- or right-hand sides (Fig. 16-34). There may be a series of ripples across a portion of the screen (Fig. 16-18), or the picture may possess two or three closely spaced dark lines (Fig. 16-35).

To track down the portion of the horizontal circuit most likely to contain the seat of the trouble, the technician must recall something that was stated on p. 331. That is, that the first 40 per cent of horizontal beam travel is controlled by the damper tube (and its circuit) while the remaining 60 per

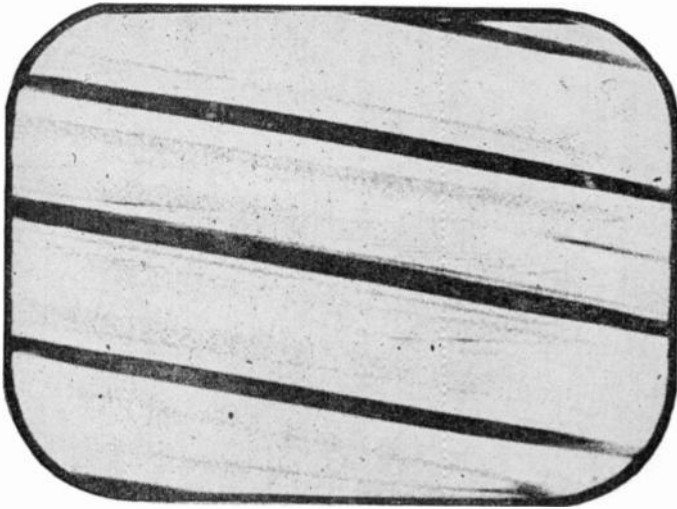


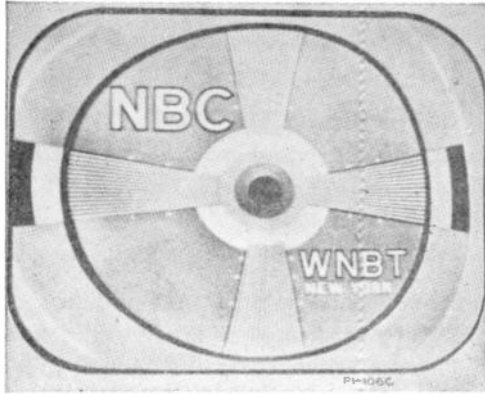
Fig. 16.33. Loss of horizontal synchronization.

cent of beam travel across the screen is controlled by the horizontal output amplifier.

From this we can see that, if the picture is impaired on the left-hand side, the defect is most likely to exist in the damper tube circuit (i.e., from the plate of the horizontal output amplifier to the deflection yoke). On the other hand, if it is the right-hand side of the picture which is distorted, the most likely place to look for the defect is in that portion of the horizontal sweep system extending from the horizontal oscillator up to and including the output tube.

Thus, from the foregoing reasoning, ripples in the picture caused by a defective capacitor in the deflection yoke, or foldover caused by a defective damper tube, or light and dark stripes caused by misadjustment of the linearity coil, all quite naturally fall at the left-hand side of the picture because they arise from the damper-tube circuit. In place here, too, are the dark stripes of Barkhausen oscillations* since these occur after the output tube

* These are spurious oscillations set up within the horizontal output amplifier tube just after the tube has been cut off.



Courtesy RCA

FIG. 16.34. An image possessing horizontal non-linearity.



Courtesy Radio News

FIG. 16.35. The visual effect of a Barkhausen oscillation.

has been cut off and the beam has just been returned to the left-hand side of the screen.

On the other hand, foldover or compression at the right-hand side of the picture is produced when the value of the grid resistor of the horizontal output tube is decreased sharply.

Most frequent defects in the damper stage include the damper tube, any

damping resistors that may be used, the two filter capacitors, C_{79} and C_{78} in Fig. 16-36, and finally, the linearity coil.

In the horizontal output stage, check the following:

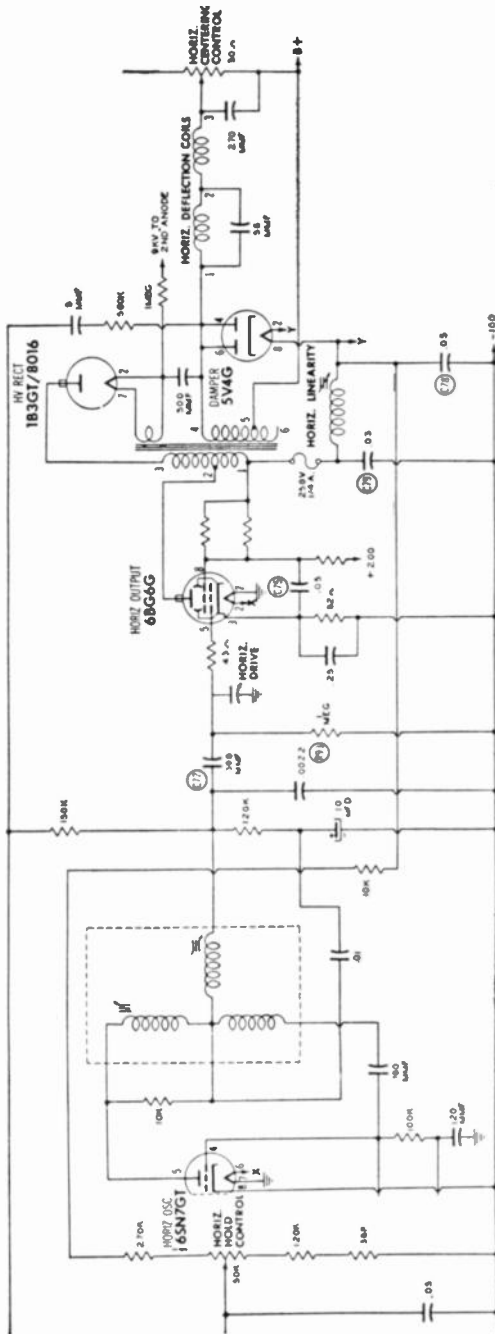
1. Lowered grid resistor (R_{93} in Fig. 16-36).
2. Defective screen-grid by-pass capacitor (C_{75} in Fig. 16-36).
3. Leaky coupling capacitor (C_{72} in Fig. 16-36).
4. Bad horizontal output tube.
5. Defective cathode bias resistor or capacitor.

In most instances the general location of the defect, as revealed by the section of the picture which is affected, is fairly well defined. However, because a "B+ boost" voltage developed by the damper tube is fed back to the horizontal output tube (and frequently to other tubes in the horizontal system as well), a certain amount of interaction between the various sections is bound to occur. For example, changes in the "B+ boost" filter network components will have an effect on both sides of the picture, although the left-hand side will be the section principally affected. Or, again, insufficient driving voltage applied to the grid of the output tube will have its greatest effect on the right-hand side of the picture. But since the "B+ boost" voltage developed depends upon the drive voltage, the left-hand side of the picture will suffer, too. Thus, while the circuits directly associated with each portion of the picture will have their greatest effect on that section, the close relationship between all circuits will produce disturbances in other portions of the picture as well.

There is still another feature of horizontal-sweep-system operation that is worth noting and this is the peak-to-peak amplitude of the driving voltage fed to the grid of the output amplifier tube. The width of the picture is governed by this voltage as well as the high-voltage for the picture tube and the "B+ boost" in the damper circuit. It is an important check point in the horizontal system and measuring its value early in the servicing analysis is strongly recommended.

SERVICING A.F.C. CIRCUITS

The purpose of the A.F.C. circuit in the horizontal sweep system is to keep the horizontal oscillator locked-in with the incoming sync pulses. When this system is operating properly, random noises and momentary disturbances will have no visible effect on the horizontal sync stability; the picture will remain securely in place. However, when the system is not operating normally, it may slip out of synchronization occasionally, or the setting of the horizontal hold control will be critical, or the picture may refuse to be locked-in at all.



Courtesy H. W. Sams & Co.

Fig. 16.36. A typical horizontal output system.

Failure of the A.F.C. system to maintain the picture in synchronization may be due to a defect in the control circuit or it may be due to a defect in some prior circuit. If the latter is true, then the sync pulses reaching the control tube are distorted in some manner or they may even be missing altogether.

If it is determined that the proper pulses are reaching the A.F.C. circuit, then any reason for sync instability must be due to faulty operation of the control circuit itself. To localize the source of the trouble, several methods of approach are open to the serviceman. As a start, he can measure the voltages within the control circuit. This lends itself quite readily to the location of fairly obvious defects caused by resistances which are either open or have changed radically in value and leaky capacitors. Tubes are not mentioned because it is assumed that these will receive first attention whenever work is done on *any* circuit.

Another approach to the service problem is by waveform checking within the A.F.C. circuit. To do this effectively, the serviceman should be familiar with the correct waveforms. The various waveforms in each of the more popular A.F.C. systems were shown in Chapter 13. Perhaps a better place to find what the waveform should be for a particular receiver is in the manufacturer's service manual. Slight variations in the circuit will frequently cause some modifications in the wave shapes developed therein.

Another item to check in your examination of an A.F.C. system is whether the d-c control voltage is being fed to the controlled tube. Connect a VTVM between the d-c control path and chassis. Then, with a signal coming in, slowly rotate the horizontal hold control from end to end. In a properly operating system, the needle on the VTVM should swing back and forth in step with the hold control variation. This indicates that the control network is developing corrective voltages to offset the changes in horizontal sweep frequencies produced by rotating the hold control. Failure to observe these voltage swings on the VTVM indicates that either no d-c control voltage is being developed or that what is developed is not reaching the controlled stage.

Changes in component values are frequently the cause of instability in A.F.C. systems. Be especially mindful of this when making voltage and resistance checks.

QUESTIONS

1. What is the significance of the video-sound signal separation point in a television receiver?
2. What indications would be obtained if each of the following defects occurred in a television receiver? (Consider each one separately.)
 - a. No vertical deflection voltage.
 - b. The R.F. amplifier tube became defective.

- c. The clipper tube became inoperative.
- d. No low voltage.
- e. The horizontal deflection yoke opened up.
- f. A coupling capacitor in the video amplifier system opened.
- g. The A.G.C. line was accidentally grounded.
- h. The horizontal oscillator stopped functioning.

3. What is the visual effect of low-voltage a-c ripple in the video amplifiers? How do we distinguish between this effect and that obtained when sound voltages reach the cathode-ray tube?

4. Describe a method for measuring the amount of ripple in the low-voltage power supply.

5. What controls are associated with the horizontal deflection system? Consider that a pulse-width A.F.C. system is being employed.

6. How would you go about checking an A.F.C. system which you believed was defective? Use the system in Fig. 15-1 as your example.

7. In checking through a horizontal or vertical deflection system, what do you look for?

8. Which controls are associated with the vertical deflection system? State the visual effect of an improper adjustment of each control (considered separately).

9. How can we distinguish between vertical and horizontal non-linearity?

10. What possible image distortions are due to defects in the video I.F. system?

11. List the equipment which should be on hand for thoroughly testing a television receiver.

12. What defects could exist in sets employing electromagnetic deflection that could not exist in sets using electrostatic deflection?

13. Outline several tests which will indicate whether or not a signal can get through the front-end stages.

14. What visual effects do gassy tubes produce? (This does not include picture tubes.)

15. Why is it frequently important to measure the A.G.C. voltage early in an analysis?

16. What defects can occur in a video amplifier?

17. Indicate some of the troubles you might encounter with picture tubes.

18. Outline your servicing approach to a receiver containing no high voltage. Assume a flyback system is employed.

19. What will cause picture smearing?

20. List some of the causes of insufficient picture height.

Chapter 17

UHF TELEVISION

The Federal Communications Commission, in its original allocation plan issued in 1946, set aside 12 * channels for commercial television broadcasting. It did not take long to demonstrate, however, that this was far too few channels for extensive nationwide coverage and, in 1952, an additional 70 channels in the UHF band (470-890 mc) were added to the original 12.

The problems facing the television serviceman in the UHF band do not differ in principle from those presently facing him in the VHF bands. They do, however, differ in degree. Thus, he must still erect an antenna system to capture as much signal as possible except that UHF signals reaching the antenna are weaker and the losses presented by transmission lines are greater. Within the receiver, UHF signals are first fed to a mixer crystal where they are converted down to a much lower frequency and then they are amplified by suitable amplifiers. Note that there is no stage of pre-amplification before the mixer (as in VHF receivers) simply because tubes capable of operating in the UHF range are, at the present time, quite expensive. Thus, for the moment, we again find UHF reception at a disadvantage, and this portion of a television receiver will be less sensitive than the VHF section. Where the sensitivity of a VHF receiver might be on the order of 25 microvolts, its sensitivity when tuned to UHF would be perhaps 150 microvolts.

In order to present as much UHF signal to the receiver as possible, careful erection of the antenna is necessary. This means that you not only have to find the best spot horizontally, but you also have to determine what the best height is. In the choice of antennas, the serviceman has a considerable number from which to pick, and, fortunately, high-gain arrays are more feasible at UHF because antenna dimensions are smaller. Thus, a half-wave dipole at 550 mc will have roughly $\frac{1}{10}$ the overall dimensions of a half-wave dipole at 55 mc. This means that more elements can be added

* At first 13 channels were allocated, but channel 1 (44-50 mc) was subsequently dropped. See page 36.

to the UHF array without causing it to become unwieldy. Moreover, since the gain of an antenna generally rises with the number of elements, higher gain can be anticipated from UHF arrays.

In the sections that follow a number of popular UHF arrays will be examined.

ANTENNAS

Fan Dipole.* The half-wave dipole is the simplest type of VHF antenna and likewise it is the simplest UHF antenna. See Fig. 17-1. By using triangular sheets of metal instead of rods, the unit becomes a broadband affair, capable of receiving all signals within the television UHF band. The response pattern of a fan dipole is a figure-8 unless a screen reflector is placed behind the array. See Fig. 17-2. In this case it becomes uni-directional.

The impedance of a fan dipole possessing a corner angle of 70° is about 300 ohms. Overall length for greatest gain is of the order of 16 inches. Fan dipole gain is slightly higher than that of a rod dipole, and for this reason these units provide satisfactory reception only in strong signal areas where there are relatively few ghost signals.

Fan dipoles can be stacked two high and four high to provide increased gain and better discrimination against ground reflected signals. The latter property is useful in reducing the number of interfering signals capable of affecting the receiver.

Note that the di-fan reflector in Fig. 17-2 is a mesh screen instead of a rod as is customary at VHF. Screens are far more efficient reflectors than rods and the only reason they are not used extensively at VHF is because they would be too bulky. Mesh screens are as effective as solid metallic sheets, provided the mesh openings are on the order of 0.2 of a wavelength or less at the highest operating frequency. Reflector dimensions are not critical, but the edges should extend for a short distance beyond the dipole elements.

Rhombic Antennas. The rhombic antenna, Fig. 17-3, has been used to some extent for VHF reception in weak signal areas. It has broad band-pass properties, is uni-directional, possesses a sharply defined directivity pattern, and is capable of good gain. The length of each of the four sides (also known as legs) should be at least several wavelengths long. In fact, the longer each side, the greater the gain and the sharper the horizontal directivity pattern of the array. In order to obtain the uni-directional effect, a resistor having a value of 470 to 500 ohms is inserted at one end of the array (Fig. 17-4). The input impedance at the other end is equal in value to

* Also known as a "Bow-Tie" antenna and as a "di-fan" antenna.

this resistor, but connecting a 300-ohm line here will, for all practical purposes, serve very nicely with only a slight loss in gain.

In the UHF band, it is customary to make each leg length equal to two or more wavelengths. Suitable mechanical support can be obtained by a cross-

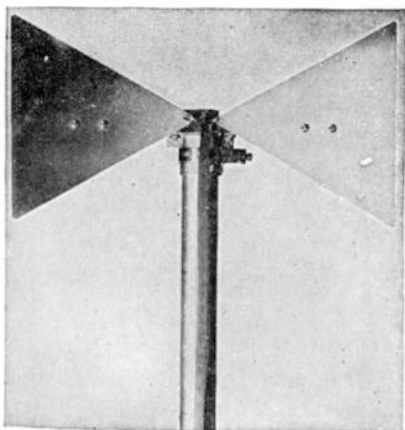


FIG. 17.1. A UHF fan dipole.

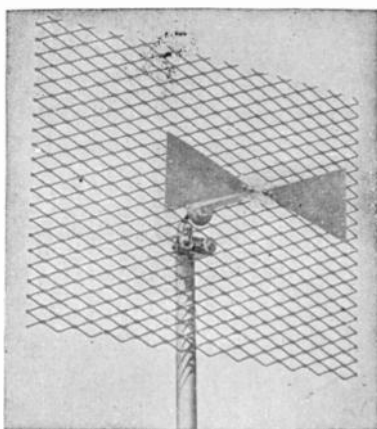


FIG. 17.2. A fan dipole with a screen deflector.
Courtesy Walter L. Schott Co.

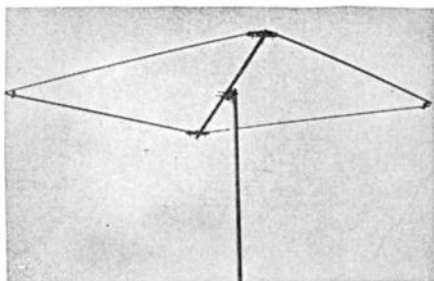


FIG. 17.3. A rhombic UHF antenna.

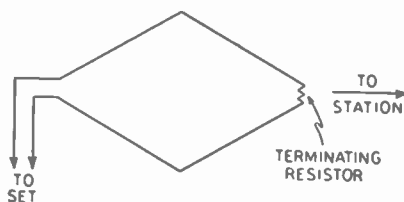
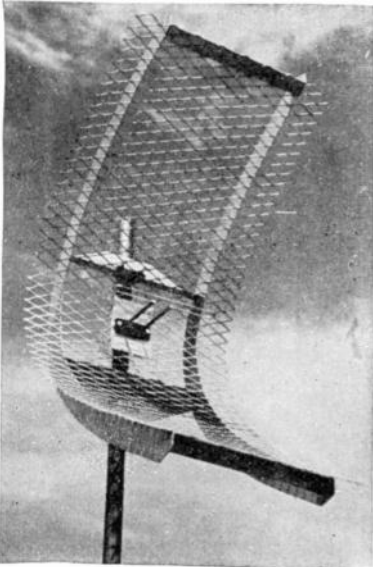


FIG. 17.4. By inserting a resistor in one end of a rhombic antenna, a unidirectional pattern is obtained.

arm mast which is mounted at the top end of a pipe. The array is then oriented with the resistor end pointing in the direction of the station. Stacking of rhombic antennas will provide increased gain and is employed in weak signal areas.

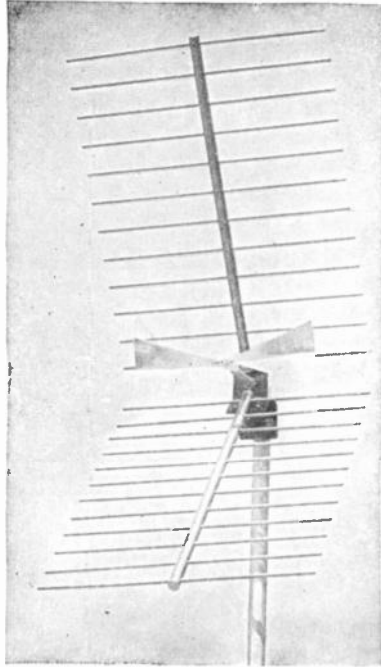
Cylindrical Parabolic Reflector. Probably everyone is familiar with the fact that the headlights on a car use parabolic reflectors in order to obtain a high concentration of light. In much the same fashion, parabolic reflectors can be (and have been) used to receive and transmit radio waves with high gain and sharp directivity.

Instead of using an entire parabolic "dish," it is possible to use only a section. The cylindrical parabolic reflector in Fig. 17-5 alone can provide a gain of 8 db over a resonant half-wave dipole. The vertical directivity of this antenna structure is sharp, but the horizontal directivity is somewhat broad. Where high gain is desired and the ghost problem is not serious, this array will provide excellent results.



Courtesy TACO

FIG. 17.5. A cylindrical parabolic reflector.



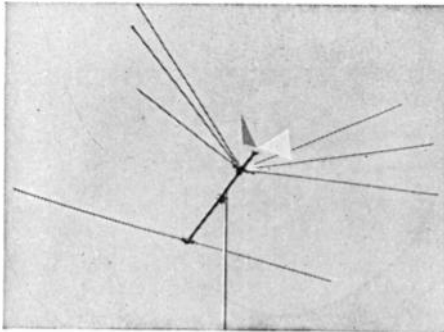
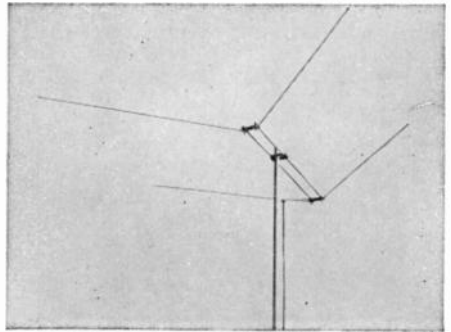
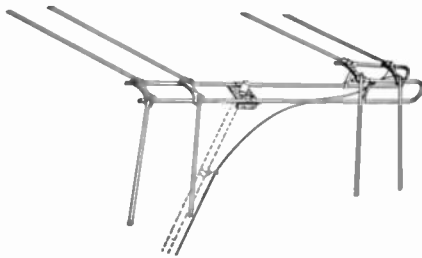
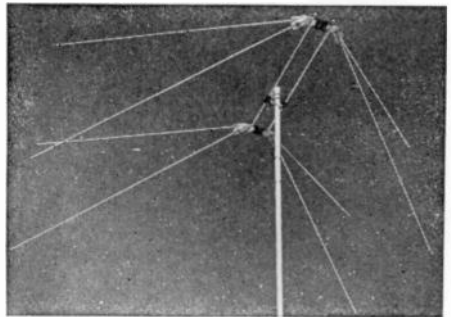
Courtesy Walter L. Schott Co.

FIG. 17.6. A corner reflector for UHF antenna.

Corner Reflectors. Instead of using curved surfaces as reflectors, it is possible to use two flat surfaces which are so positioned as to intersect at some angle, forming a corner. This type of reflector, shown in Fig. 17-6, is known as a "corner reflector" antenna. The receiving element, usually a dipole antenna, is placed at the center of this corner angle and at some distance from the vertex of the angle.

The response pattern of this array depends not only on the corner angle but also on the distance between the antenna and the vertex of the reflector corner. If the antenna is positioned too far from the vertex, the response pattern will have several lobes. If it is brought in too close, the vertical response will be broadened and the susceptibility of the array to ground reflected signals will increase.

The corner angle in the commercial array of Fig. 17-6 is 90° , and a similar bend is placed in the accompanying fan dipole. Gain over the entire UHF-TV band is high, ranging from about 7 db at 500 mc to 13 db at 900 mc.

(A) *Courtesy J.F.D.*(B) *Courtesy J.F.D.*(C) *Courtesy Ward*

(D)

FIG. 17.7A. A combination VHF-UHF antenna consisting of a low-band conical and a broad-band fan dipole.

B. A double "V" type of antenna which will receive VHF and UHF signals.

C. Another combination antenna derived from the basic "V."

D. The RMS CVA-500 VHF-UHF antenna.

Combination Antennas. In many parts of the country, both VHF and UHF stations will be in operation and, for those areas, combination antennas, if suitable, would serve to simplify the installation problem. A number of combination VHF-UHF arrays are available, and the ones shown in Fig. 17-7 are indicative of what the technician will find.

The J.F.D. array, shown in Fig. 17-7A, consists of a low-band conical antenna for VHF signals and a broad-band fan dipole for UHF signals. A single lead-in line delivers signals to the receiver through the use of a special coupling device which is mounted directly at the antenna itself.

"V" type arrays have appeared in a number of forms, of which the one shown in Fig. 17-7B is typical. The antenna contains four rods: two at the front and two at the rear. The two front rods may be considered as the directors; the two rear rods act as the active elements, supplying the signal to the receiver via a 300-ohm transmission line. However, in distinction to other arrays containing directors, the two front rods, Fig. 17-7B, are electrically connected to the two rear rods. The purpose of the connecting rods is twofold.

First, the rods serve as a transmission line to conduct to the two rear rods whatever signal is picked up by the two front rods. In addition, the two rear rods also pick up that portion of the signal which passed over the front rods and combine this with the energy received from the front rods via the transmission line.

Second, the connecting rods support both front and rear rods and produce a mechanically sturdy array.

It is characteristic of these antennas that the longer each side of the "V" becomes as compared to the operating frequency of the signal, the narrower the angle between the sides must be made for best gain and a single-lobe pattern. This means that, for low-band operation, the sides are spread out farthest, generally until the included angle is 90° . For the high-band operation, the sides are brought in closer, generally to 60° or even 45° . The best angle for a particular installation is determined at the time the antenna is erected. 60° is a good compromise angle for both VHF and UHF signals, although if signals above 750 mc are to be received, a smaller angle may be necessary.

Two variations of the "V" type array are shown in Fig. 17-7C and Fig. 17-7D. Stacking of the sections increases the gain (about 2 db per additional section) and sharpens the vertical directivity.

Other Antennas. While not specifically stated in this discussion, we could use any of the antennas previously described in Chapter 2. Of course, the size of these antennas would have to be scaled down to the proper dimensions, but other than that no other changes would be required. Design formulas remain unaltered, too.

Because of the smaller dimensions of UHF antennas, stacking of arrays is more common. A 4-bay array designed for reception at 60 mc would be in the neighborhood of 8 feet wide and 16 feet high. At 500 mc, it would only be 1 foot wide and 2 feet high, certainly a considerable difference! And since the more elements you use on an array, the more gain you obtain from it, the trend toward more elaborate structures is understandable.

TRANSMISSION LINES

Transmission lines in common use at VHF frequencies include a 450-ohm open-wire line, 300-ohm flat and tubular lines, and two coaxial cables, the RG-59U and RG-11U. All these lines, with the exception of the 300-ohm tubular, were illustrated in Chapter 2. The tubular line is shown here in Fig. 17-8. At 100 mc, the signal loss per 100 feet of the open-wire line is 0.35 db, in the 300-ohm line it is about 1.2 db, in the RG-11U cable it is 1.8 db, and in RG-59U it is 3.8 db. The open-wire line has, by far, the lowest attenuation and, as such, is eminently suitable for fringe area installations. Its characteristic impedance is 450 ohms, and this high value may



Courtesy Amphenol

FIG. 17.8. 300-ohm tubular twin-lead transmission line.

frequently require a matching network between the line and the usual 300-ohm receiver input impedance. That it does not always do so stems from the fact that receiver impedances may vary considerably from their stated value of 300 ohms.*

The 300-ohm line (flat) is the one which is most extensively employed in VHF because it is economical, it matches receiver input impedances directly, and its attenuation is still low. In areas where the surrounding noise is high, use of the unshielded open-wire or 300-ohm lines becomes impractical and one of the coaxial cables must be employed.

With increase in frequency, line db attenuation rises, too. Thus, at 500 mc, the open-wire loss mounts to 0.78 db, the 300-ohm line loss becomes 3.2 db, RG-11U attenuation increases to 5.0 db, and RG-59U loss reaches the rather high figure of 9.4 db. Comparative figures at 100 mc, 500 mc and 1000 mc are given in Table 17-1, and it can well be appreciated how carefully the amount of line needed must be figured in order that no more than necessary be employed.

An interesting sidelight on line attenuation is the rapid rise in this value in unshielded lines when they become wet. The 300-ohm flat line appears to be most vulnerable in this respect, jumping from a value of 3.2 db (at 500 mc), when the line is dry, to 20.0 db when it becomes wet. What this rise would do even to a strong signal is quite obvious. The 300-ohm tubular is considerably better in this respect and hence would be more desirable for installation purposes. No data are available on the attenuation increase in

* Recently a 375-ohm open-wire line was announced. This brings the line impedance closer in value to the 300-ohm receiver input, and it is doubtful whether a matching network will ever be required.

TABLE 17-1. TRANSMISSION LINE LOSS
(Db Loss Per 100 Feet)

Type	100 Mc		500 Mc		1000 Mc	
	Wet	Dry	Wet	Dry	Wet	Dry
450-ohm open-wire*	...	0.35	...	0.78	...	1.1
300-ohm tubular	2.5	1.1	6.8	3.0	10.0	4.6
300-ohm flat	7.3	1.2	20.0	3.2	30.0	5.0
RG-59U	...	3.8	...	9.4	...	14.2
RG-11U	...	1.8	...	5.0	...	7.6

* Estimated values — unknown for wet conditions.

wet open-wire lines, although it is not considered to be appreciable. Shielded cables, such as the RG-11U and RG-59U, are not affected at all by inclement weather.

An important factor in keeping overall attenuation down is the careful routing of the line from antenna to receiver. This is especially true in the case of unshielded lines. Keep the line away from gutters, pipes, or other metal objects as much as possible. Avoid sharp bends in the line. If a bend must be made, have it occur gradually. Finally, secure the line tightly by means of stand-off insulators so that it does not sway in the wind or otherwise alter its position.

Impedance matching at the antenna and, more importantly, at the receiver will require more careful attention in UHF installations. Mismatching at the receiver results in energy being reflected back along the line with resultant standing waves. It has been found that the attenuation of a line may be increased by as much as 2 db over its normal rating when standing waves are present. In strong signal locations the additional loss may not be serious, but in moderate and weak signal areas it can mean the difference between usable and unusable pictures.

UHF CONVERTERS

The allocation of television channels in the UHF region was not announced until the spring of 1952 and the actual granting of station permits did not occur until July of the same year. By that time, VHF television, on channels 2 to 13, had been in operation for six years or more and nearly 18 million receivers had been manufactured and sold. None of these sets

were equipped to receive UHF signals, and unless some method was evolved to adapt these sets for UHF reception, they would, in time, become obsolete.

The industry attacked the problem in a twofold manner. The most flexible solution was the construction of converters. These receive the UHF signal, convert it down in frequency to one of the VHF channels, and then transfer this signal to the input terminals of a VHF receiver where it is treated as any other VHF signal. Most converters are housed in separate cabinets and contain their own power supply. For convenience in operation, they can be placed on top of the VHF receiver cabinet. The incoming UHF signals are fed to a suitable terminal at the rear of the converter and then the transformed signal is routed to the VHF receiver by another line.

Some converters are especially designed by some receiver manufacturers for their own VHF sets, and in these cases power for the converter can be obtained from the VHF receiver power supply. In still other modifications, the UHF converters are designed to fit inside the VHF receiver cabinet. Tuning of the converter can then be accomplished either by mechanical linkage to the VHF tuner (if such provision has been made) or by using a separate front panel dial. A number of manufacturers follow this latter course with their own VHF receivers.

A second solution for adapting VHF receivers to UHF reception was by the addition of auxiliary tuning elements to the VHF tuner. Examples are the UHF channel strips that can be substituted for unused VHF strips in turret tuners. Zenith and Standard Coil are advocates of this approach because their tuners fall within this category.

With either of these two methods, all existing VHF receivers can be adapted to receive UHF signals. At the same time, the receivers now being produced contain either combination VHF-UHF tuners or else make provision for the installation of a UHF tuner when this becomes necessary. Undoubtedly, the next few years will see a variety of tuning methods until the field becomes stabilized and the number of different tuners is reduced.

In the paragraphs to follow, a number of UHF converters will be examined to determine how they function and how they can be combined with VHF receivers. UHF tuning strips will also be considered.

General Electric Converter. The General Electric converter,* Model UHF-101, shown in 17-9, consists of a preselector tuned input circuit, a 1N72 crystal mixer stage, a 6AF4 oscillator, and a 12AT7 I.F. amplifier. The block diagram of this unit is given in Fig. 17-10 and the schematic diagram in Fig. 17-11.

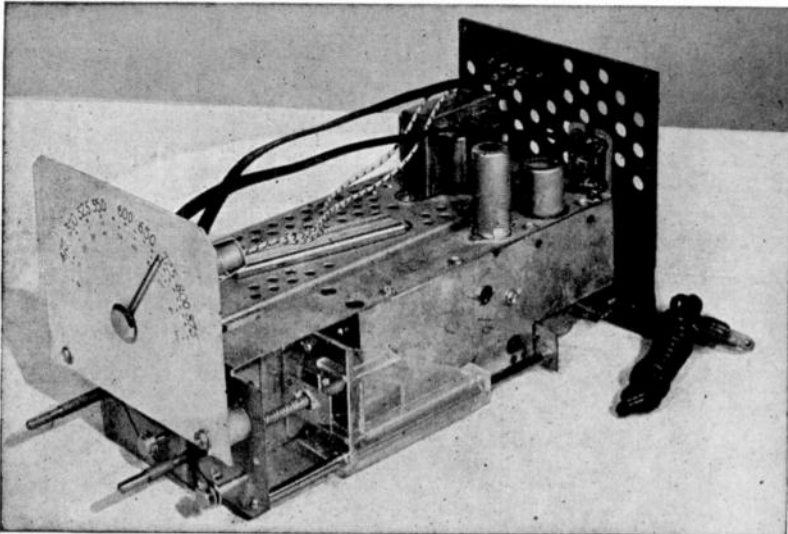
In detail, incoming UHF signals are applied to a balanced input circuit

* G.E. calls its converter a "UHF Translator."



Courtesy G.E.

FIG. 17.9A. General Electric model UHF-101 UHF converter.



Courtesy G.E.

FIG. 17.9B. General Electric without cabinet. The screw drive for the transmission-line tuners can be seen in the lower left-hand section of the chassis.

having an impedance of 300 ohms. The input tuning circuit, called a pre-selector, is a quarter-wave shorted section of parallel-wire transmission line. In Fig. 17-11 this line is labeled L_1 . The lowest tuning frequency is obtained

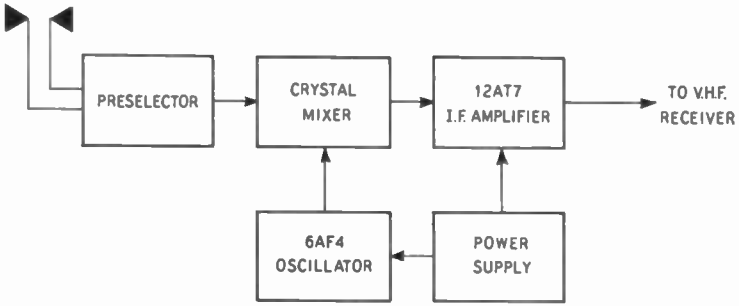


FIG. 17.10. Block diagram of General Electric UHF converter.

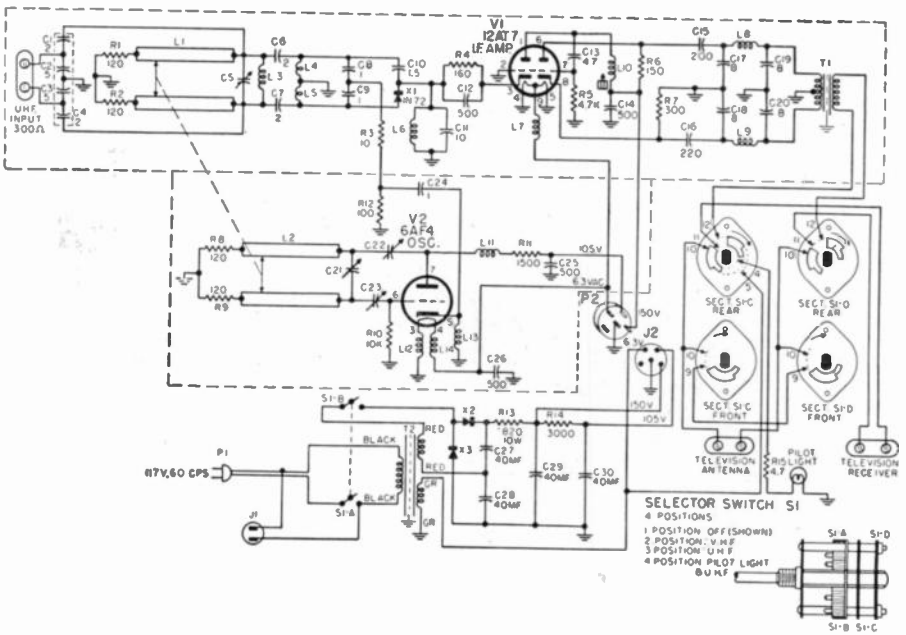


FIG. 17.11. Schematic diagram of the General Electric Model UHF-101 converter.

when the shorting bar is at the resistor (R_1 and R_2) end of the line; the highest frequency is obtained when the shorting bar is near C_5 . This preselector circuit determines which of the UHF signals are to be permitted to enter the converter and which signals are to be attenuated. Beyond L_1 is a high-pass

Courtesy G.E.

filter which acts in conjunction with L_1 to establish the R.F. pass band of the converter and to prevent VHF signals from penetrating very far into the converter circuit and interfering with its normal operation.

Following the preselector is the crystal mixer, a 1N72. Here the incoming R.F. signal is combined with a locally generated voltage to produce the difference or I.F. signal. A 6AF4 triode is the oscillator, and its output voltage is capacitive-coupled to the 1N72. The oscillator uses the same type of quarter-wave tuning circuit that the preselector does and the two are ganged together. To permit tracking over the UHF band, both lines have high- and low-end tracking adjustments. C_5 (for L_1) and C_{21} (for L_2) are the low-end tracking adjustments. For the high end, the positions of the individual shorting bars must be separately adjusted and special nuts are provided for this purpose.

The frequency generated by the oscillator *must* be below the incoming signal frequency by an amount equal to the desired I.F. This is necessary in order that the video carrier frequency, after mixing, remain below the sound carrier frequency. Remember that the output of this converter goes to a VHF receiver and the relative positions of the two carriers must be the same as they are in directly received VHF signals.

The signal at the mixer output is now applied to an I.F. amplifier. The notation of I.F. is borrowed from the ordinary superheterodyne circuit where the amplifiers following the mixer are known as I.F. amplifiers. Actually, here, the I.F. amplifier will have a frequency equal to that of VHF channel 5 (or 6) because that is what this converter is going to feed to a subsequent VHF receiver. Therefore, a converter I.F. is not the same value usually as the I.F. in a VHF receiver and this distinction should be kept in mind.

In the G.E. converter there are two I.F. stages formed by the two triode sections of the 12AT7. The first triode functions as a grounded-grid amplifier, while the second amplifier has its cathode and plate so set up that a balanced output is obtained. (In other words, the plate supplies half the output signal while the cathode supplies the other half.)

The output signal of the converter may fall either in channel 5 or channel 6, depending upon the adjustment of coil L_{10} . A choice of two channels is available so that if one channel is occupied (locally) the other channel may be used.

The power supply uses selenium rectifier in a voltage doubler arrangement. Output B+ voltages of 150 and 105 volts are available.

On the back of the converter, three sets of antenna terminals are available. To one set is connected the UHF antenna. To another, we apply the VHF antenna that would normally go to the VHF receiver. The third set of terminals is for connection of the converter output to the VHF receiver. When the converter front panel switch is set to "VHF," then the VHF an-

tenna is connected to the VHF receiver directly through the switch. In this position the converter is by-passed.

When the converter front panel switch is set to "UHF," then the converter output is passed on to the VHF receiver. In the final switch position labeled "LIGHT," the dial light of the converter comes on, enabling the set owner to see the converter dial more easily. UHF signals still reach the VHF receiver in this switch position.

Fig. 17-12 shows the connection diagram of the G.E. converter for the antennas, the converter, and the VHF receiver. Note that the power line plug of the VHF receiver can be plugged into the back of the converter. The power plug of the converter then goes into the wall outlet and the converter determines whether or not both units receive power.

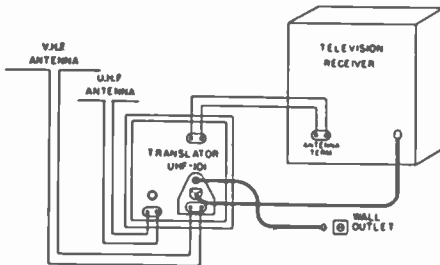


FIG. 17.12. Connection diagram for the General Electric converter.

Before we leave this converter, it might be of interest to note that it uses a crystal (rather than a vacuum tube) mixer.

The practice of using crystals for the mixer stage in place of vacuum tubes is widespread in spite of the fact that crystals introduce an 8- or

9-db loss in signal strength. A vacuum tube used as a mixer would provide some gain, but it would introduce more noise, which is undesirable. If we add to this the fact that crystals are cheaper and their circuits simpler, their popularity becomes understandable.

Another Converter. A second converter, Fig. 17-13, is built somewhat along the lines of the G.E. converter except that its tuning circuits are different and it makes use of a cascode I.F. amplifier.

The converter input is designed for a 75-ohm coaxial cable. (To use a 300-ohm transmission line, a special matching device would be necessary. These are available.) All incoming signals are first passed through a high-pass filter before reaching the tuned preselector circuits. The low frequency cut-off of the filter is in the neighborhood of 400 mc and is designed, as in the G.E. unit, to prevent VHF signals from entering the converter circuits and causing interference. The double-tuned preselector which follows next provides whatever UHF selectivity this unit possesses. Two of the three concentric-line sections tune the preselector elements to the desired frequency. Each section of the preselector is shielded thoroughly, and whatever signal passes from the first to the second section must do so by the small capacitive coupling that exists between them. The diagram indicates that an actual physical capacitor is not used, but that the coupling is achieved by placing

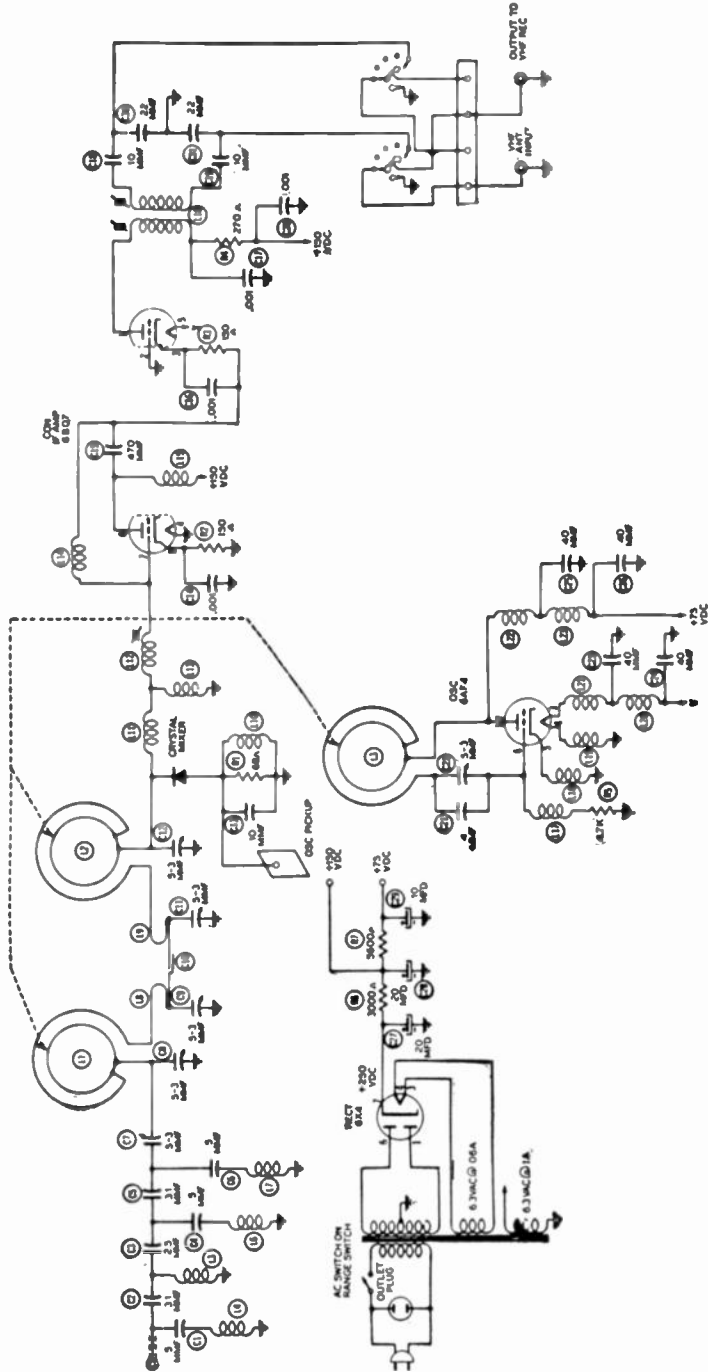
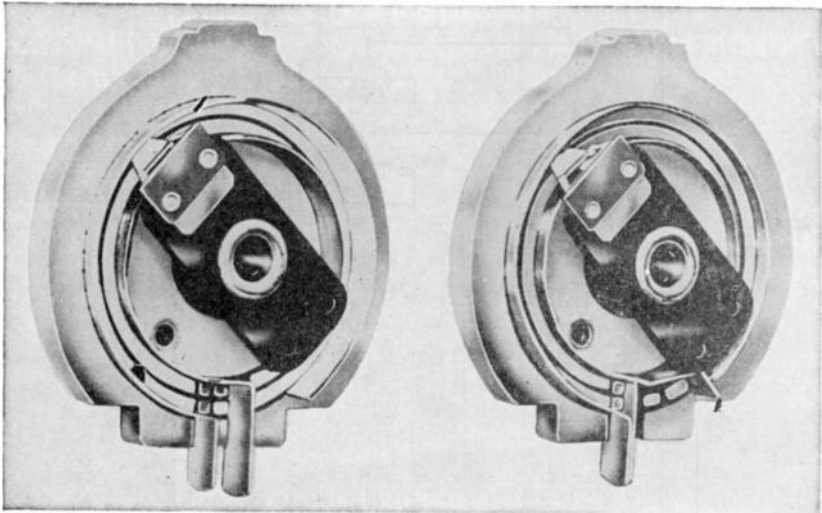


Fig. 17.13. Schematic diagram of a second UHF converter. *Courtesy H. W. Sams & Co.*



Courtesy P. R. Mallory Co.

FIG. 17.14. The tuning element employed in the converter of Fig. 17.13.

leads from both sections close together. The end result, of course, is the same.

A closer look at the tuning elements themselves indicates they consist of two metal strips which are placed in parallel circular grooves extending for 270° of the total 360° . See Fig. 17-14. The far ends of the two concentric lines are permanently shorted together; the other ends are the points to which the circuit components are connected. The length of line which is then active is that section between the open ends and the movable shorting bar. The reason for shorting the far terminals together is to short-circuit any low-frequency interference which may be present. This short also tends to keep the lines balanced and thereby reduce oscillator radiation. This is especially important in the R.F. preselector section.

Three and four concentric line sections can be ganged together on one shaft. To permit tracking, the various tuning elements can be shaped differently from one another.

From the output of the preselector, the signal goes to a 1N72 germanium diode crystal. Voltage from the oscillator is also injected into this circuit by using a small metal plate placed near the oscillator tuning circuit. This type of coupling is essentially capacitive and it possesses the advantage of injecting an increasing amount of voltage as the frequency rises. This will offset the tendency of the oscillator output to drop as the frequency goes up and maintain a fairly constant injection voltage.

The oscillator uses the third of the tuner sections. The circuit is of the ultraudion type which is so widely used in VHF receivers. C_{22} is for the

purpose of adjusting the oscillator so that it tracks with the other two sections of the tuner. C_{21} is used for temperature compensation. Note the extensive use of small R.F. chokes in all sections of this oscillator. In the plate and heater circuits, their primary purpose is to prevent the R.F. signal from reaching other sections of the converter from which it might radiate to near-by receivers. In the cathode circuit, the choke helps to maintain this element at R.F. ground potential. And in the grid circuit the choke again prevents the R.F. energy from straying out of its regular channels.

Finally, as a further safeguard against unwarranted radiation, all sections of the tuner are carefully shielded, especially the oscillator.

At the output of the mixer, the difference frequency (or converter I.F.) is transferred to a 6BQ7 cascode amplifier. The coupling network of L_{11} , L_{12} , and L_{13} permits the I.F. signal to pass but prevents the higher R.F. signal from reaching the 6BQ7. The cascode amplifier sections strengthen the signal and then couple it to the output via transformer L_{16} . From here the signal is transferred to the VHF receiver. Both 72-ohm unbalanced and 300-ohm balanced outputs are available.

Either channel 5 or channel 6 on the VHF receiver may receive the converted UHF signal. The one chosen will again depend upon which one is least disturbed by incoming VHF signals.

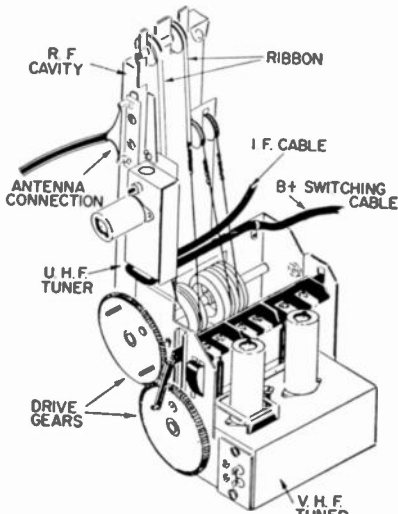
The power supply in this converter employs a 6X4 full-wave rectifier to provide +75 and +150 voltages for the 6AF4 oscillator tube and the 6BQ7 cascode amplifier. A resistance-capacitance filter removes the a-c component of the rectified voltages.

Raytheon Converter. Another approach to UHF conversion is demonstrated by the Raytheon UHF tuner. This unit does not possess a separate cabinet, as the other converters do, but rather is installed directly on the VHF receiver chassis. To make the installation, the VHF chassis is removed from its cabinet, a drive gear is fitted onto the VHF tuner, and then the UHF unit is fastened to the VHF chassis with four screws. See Fig. 17-15. With the drive gears, the receiver owner can tune both UHF and VHF units with the same front panel knob. (This arrangement was devised especially by Raytheon for the conversion of their own VHF television receivers.)

Electrically, there are two cables running from the UHF tuner to the VHF chassis. One cable couples the converted signal to the VHF I.F. section while the other cable is for switching B plus voltage to either UHF or VHF tuners. Note that here the converted UHF signal does not pass through the VHF tuner section (as it did in the previous converters), but instead feeds directly to the video I.F. amplifiers of the VHF receiver. This represents a single conversion process in distinction to the double conversion procedure employed in the preceding converters. Note that in order to do this, however, the UHF section must step the incoming UHF signal down to the video I.F. value of the VHF receiver. In Raytheon sets, this is

45.75 mc. The reason this can be done in this instance, of course, is that the UHF tuner is being used with a specific type of set. If it were designed to be used with *all* makes of television receivers, it would have to follow the conversion method of the previous converters.

The schematic diagram of the Raytheon tuner is shown in Fig. 17-16.* It contains a preselector input circuit, a crystal mixer, a 6AF4 ultraudion oscillator, and a 6BQ7 cascode amplifier.



Courtesy Raytheon

FIG. 17.15. The Raytheon UHF converter is installed on the VHF tuner assembly.

Frequency tuning is accomplished by a double-tuned coaxial line or R.F. cavity. These tuning circuits are basically nothing more than a quarter-wave, shorted, tuned stub. The electrical length of the lines is varied by a ribbon which is attached to a dial cord and pulley arrangement. In this manner tuning is accomplished, a process similar to varying the length of a tuned stub, which changes the resonant length for different frequencies. Signal transfer between pre-selector tuning sections is by inductive coupling furnished by short lengths of wire.

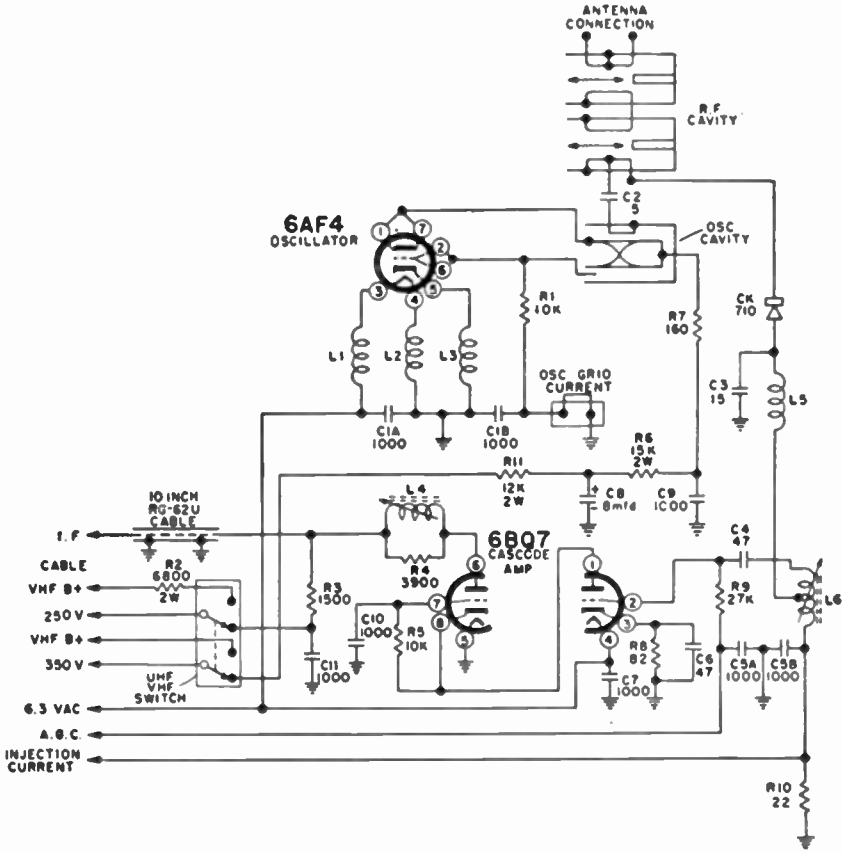
The oscillator resonant circuit is similar to that of the preselector. Here a quarter-wave, shorted, parallel-wire transmission line is used and a shorting bar varies the electrical length of the

line. There is provision in the grid circuit of the oscillator for inserting a microammeter to measure grid current. This will serve to indicate not only whether the oscillator is functioning but also how well.

The mixer combines the oscillator voltage and the incoming signal and produces a difference frequency equal to the video I.F. This signal is fed to the 6BQ7 cascode amplifier where it is amplified and then coupled to the I.F. section of the VHF receiver through 10 inches of RG-62U coaxial cable.

The input impedance of this UHF tuner is 300 ohms and the overall bandwidth is 6 to 8 mc. Since there is no R.F. amplifier ahead of the mixer, the sensitivity of the receiver on UHF will be less than it is on VHF. The manufacturer states that the UHF sensitivity is approximately 150 microvolts.

* From information furnished by the Belmont Radio Corporation which is a subsidiary of the Raytheon Manufacturing Company.



Courtesy Raytheon

Fig. 17.16. Schematic diagram of the Raytheon UHF converter.

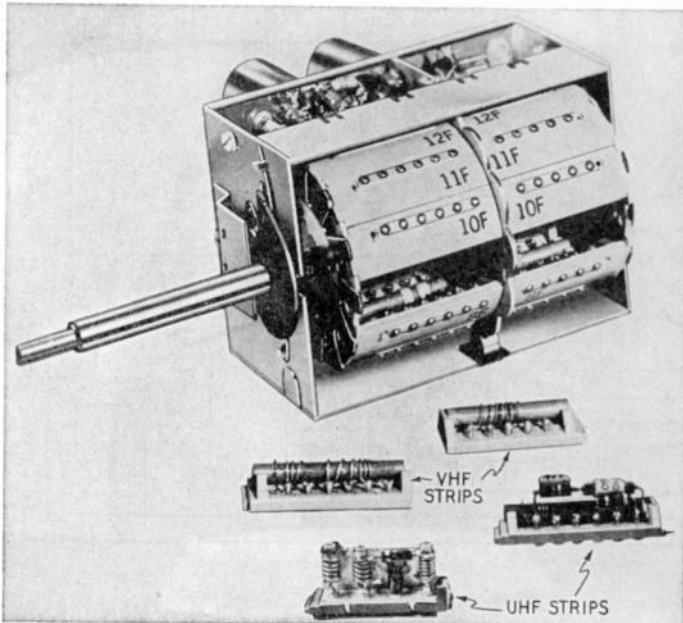
TUNER STRIPS

The second approach to UHF conversion is by means of tuner strips. There are, on the market, turret-type tuners in which the R.F. tuning elements are mounted on strips which can be removed from the tuner drum. In the Standard Coil tuner, Fig. 17-17, there is a set of two strips, mounted edge to edge, for each VHF channel. In any one locality, however, the maximum number of VHF stations is 7. This will then leave 5 unused strips and these may be taken out and a UHF set of strips inserted in their place. One UHF strip will be required for each UHF station to be received.

The block and schematic diagrams of the circuits contained on the Standard Coil UHF strips are shown in Fig. 17-18 A and B. Referring to the block diagram first, the incoming signal from the antenna is applied initially to a preselector circuit which is designed to pass only the desired signal and to

attenuate all others. The next stage to receive the signal is the mixer and in all current UHF equipment this is a crystal.

Also feeding into the mixer circuit is a locally generated signal which beats with the received signal to produce the stepped-down or difference frequency. This oscillator signal could be developed by a separate UHF oscillator, but for the sake of simplicity, the VHF oscillator already in the set is



Courtesy Standard Coil

FIG. 17.17. The Standard Coil turret tuner. The UHF strips, shown at the bottom of the illustration, would replace the two VHF strips shown just above.

utilized. What happens is simply this. The VHF oscillator is set to oscillate at some appropriate frequency. Let us say this is 139.25 mc. This frequency is then passed through a crystal (not the mixer) where harmonics of 139.25 mc are produced. A multiplier tuned circuit is peaked to the third harmonic, insuring that this frequency will predominate to the exclusion of other harmonics. This third harmonic, which is 417.75 mc, then beats against the incoming signal to produce a difference frequency which will be accepted by the R.F. amplifier of the VHF receiver.

To take a specific example, let us say that a UHF signal in the 530-536 mc channel is to be received. In this channel, the video carrier is 531.25 mc and the sound carrier is at 535.75 mc. The signal which is used to beat against this is 417.75 mc, using the figure of our example. This gives us the following difference frequencies:

For the video carrier

$$531.25 - 417.75 = 113.50 \text{ mc}$$

For the sound carrier

$$535.75 - 417.75 = 118.00 \text{ mc}$$

These two signals are then accepted by the R.F. amplifier, amplified, and then fed to the regular VHF mixer where they beat with the fundamen-

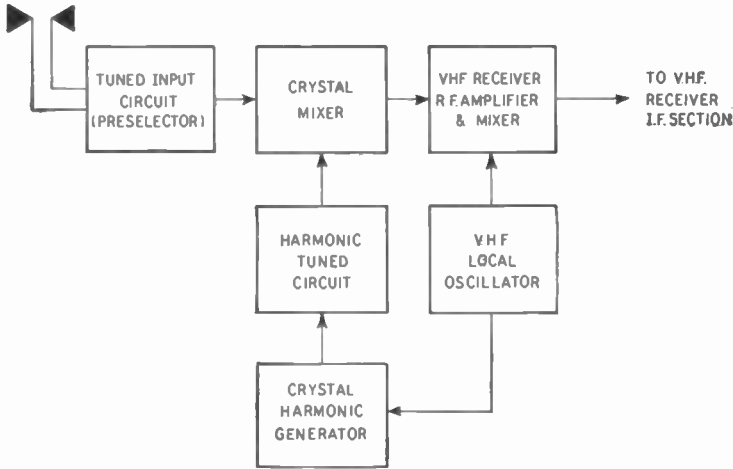
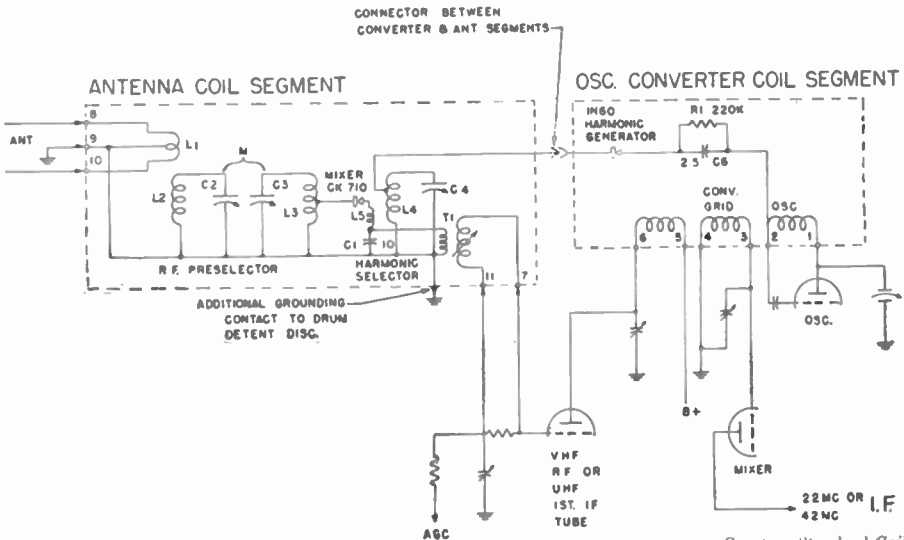


FIG. 17.18A. Block diagram of the UHF circuits contained on the Standard Coil plug-in strips.



Courtesy Standard Coil

FIG. 17.18B. Schematic diagram of the components on Standard Coil UHF strips.

tal frequency generated by the local VHF oscillator. This frequency is 139.25 mc and the result of this second beating is as follows:

For the video carrier

$$139.25 - 113.50 = 25.75 \text{ mc}$$

For the sound carrier

$$139.25 - 118.00 = 21.25 \text{ mc}$$

Thus, at the output of the VHF mixer, the video and carrier I.F. values are what they normally are for a directly received VHF signal.

From the figures used in this illustration, several interesting facts come to light. First, by using a harmonic of the VHF oscillator frequency, we do away with the need for a separate UHF oscillator. A simple germanium crystal produces the desired harmonics from the VHF oscillator signal.

Second, since this strip determines what circuits the VHF R.F. amplifier, mixer, and oscillator will have, the frequencies of these circuits can be of *any* desired value. In the foregoing example, the VHF oscillator operated at 139.25 mc; the R.F. amplifier was tuned to accept a frequency band extending from 113 mc to 118 mc.

Third, in this type of double conversion, the UHF oscillator mixing frequency must be *below* the received UHF signal. This is required in order that the position of the video and the sound carrier frequencies are not reversed. Coming into the VHF R.F. amplifier, the video carrier frequency must be below the sound carrier frequency. This is their relative position in VHF signals and must be their position in this type of arrangement.

With this preliminary discussion we can examine in detail the actual circuit on the Standard Coil UHF strip. See Fig. 17-18B. The UHF antenna is connected to terminals 8 and 10 of the strip where it actuates coil L_1 with whatever signal is received. Through mutual inductance, the signal is transferred to L_2 and then, from here, passes on to the mixer coil, L_3 . The mixer crystal is tapped down on L_3 (for best impedance match) so that it receives whatever signal is present in this circuit.

Also feeding into the mixer circuit is a harmonic of the oscillator signal. L_4 and C_4 are resonated to the harmonic which it is desired to pass on to the mixer. Transfer of the oscillator harmonic to the crystal mixer occurs by inductive coupling between L_4 and L_5 .

Both signals are now mixed and the difference frequency signal is passed on to the grid of the normal R.F. amplifier of the VHF receiver. The signal is amplified here, then transferred to the mixer where it combines with the fundamental frequency signal of the VHF oscillator. The result of this second mixing should be the normal intermediate frequency. In some receivers this will be in the 20-30 mc range; in others it will fall within the 40-50 mc range.

Note from Fig. 18B that the tuning coils and capacitances for the VHF R.F. amplifier, mixer, and oscillator are also contained on this same strip. This, of course, is in keeping with the VHF practice on turret tuning assemblies.

The multiplier crystal receives its oscillator voltage via R_1 and C_6 . The crystal distorts the oscillator currents to develop a series of harmonics. These currents flow through L_4 and C_4 where the desired harmonic is accentuated, then transferred on to L_5 for mixing with the incoming signal.

Another UHF strip is that developed by Zenith for use in its VHF turret tuners. The method of approach is somewhat similar to that described previously, except that incoming UHF signals are immediately reduced to their video I.F. values (either in the 20 or 40 mc ranges). The signal is then amplified by the VHF R.F. amplifier (which is now operating at the video I.F. frequencies) and the mixer. This latter stage has also been converted to an I.F. amplifier. Following this, the signal is transferred into the normal VHF I.F. system. On UHF operation, the VHF oscillator functions only for the UHF crystal mixer; its signal is not required for the VHF mixer tube since this is now a straight amplifier.

The foregoing discussion does not, by any means, exhaust the subject of UHF converters, but it does indicate the present trend in such units and provides the technician with the basic knowledge he needs to deal with them. In time, UHF converters, as separate devices, will disappear entirely, and each television receiver will be capable of receiving both VHF and UHF signals. The converter is merely a transitional device, designed to keep existing sets from becoming obsolete with the introduction of UHF-TV broadcasting in their locality.

QUESTIONS

1. Why are multiple element antennas more feasible at UHF than at VHF?
2. List four UHF antennas. Outline the properties of each briefly.
3. Why is a 300-ohm flat twin lead transmission line less suitable for UHF application than the 300-ohm tubular line?
4. What two general approaches did the television industry adopt to enable VHF sets to receive UHF signals?
5. Describe in detail how a Standard Coil VHF turret tuner can be converted to receive UHF signals.
6. Describe the operation of the General Electric Model UHF-101 Translator.
7. What type of tuning elements does the converter of Fig. 17-13 employ? Indicate briefly how they operate.
8. Which of the UHF units described in this chapter use a single conversion process in order to bring the UHF signal down to the video I.F. values?
9. Why are UHF installations more difficult to perform than VHF installations?
10. Describe a VHF-UHF combination antenna.

Chapter 18

COLOR TELEVISION

Introduction. Emphasis throughout the preceding chapters has been directed entirely to the underlying principles of transmission and reception of so-called black-and-white images. In such a system, only black, white, or intermediate shades appear on the receiver viewing screen. The result is similar in all respects to the ordinary motion picture. Although the reproduced image is certainly far from being an exact duplicate of the full-colored scene originally televised at the studio, it does impart sufficient information to prove highly fascinating. The public has long been accustomed to black-and-white pictures in the motion-picture theater and accepts with little or no objection the same type of image in a television receiver. Undoubtedly black-and-white images will be standard in television receivers for several years at least. But nearly all television engineers are willing to concede that colored images are more desirable and will eventually replace the present system.

The appeal of color television lies in its greater naturalness. We live in an environment that contains many varieties of color, and to desire the same life-like qualities in television is quite understandable. Color in an image heightens the contrast between elements, brightens the highlights, deepens the shadows, and appears to add a third dimension to an otherwise flat reproduction. It is interesting to note that more detail appears to be present in colored images containing fewer lines than corresponding black-and-white pictures. Perhaps many readers have noticed the remarkable differences between color motion films and ordinary motion pictures. Similar differences are observed with television.

The FCC has been investigating color television for a number of years, dating back to 1946. At first it rejected all systems shown to it as being unsuitable for commercialization. This was in 1947. For a time the question of color television lay dormant, but then it was revived and, after long months of hearings and viewing demonstrations, the FCC in October,

1950, adopted a field-sequential system proposed by the Columbia Broadcasting System (CBS). This system was known as a non-compatible system because its line and field frequencies differed from those of the black-and-white television receivers. Non-compatibility meant that, if a station was broadcasting a CBS color signal, a black-and-white picture could not be properly presented on a black-and-white television receiver unless certain changes were made in the receiver circuitry.

The adoption of the CBS color system met with opposition from the public and the television manufacturers, chiefly because of its incompatibility, and the system did not progress very far. Almost concurrently with the FCC decision, the Korean War broke out, and in the interests of national defense (and the conservation of critical materials) all commercial manufacturing of color television receivers was temporarily discontinued. When this ban was subsequently lifted several years later, CBS admitted that industry opposition to its system was too great to be overcome by CBS singlehandedly; hence CBS indicated it would be willing to go along with whatever system received the backing of the industry. This turned out to be the NTSC system, which, in December, 1953, superseded the CBS system.

Before we investigate the NTSC color system, let us refresh our memory on some elementary aspects of color.

Elements of Color. Color, physicists tell us, is a property of light. If we take sunlight and pass it through a glass prism, a variety of colors are produced. White sunlight contains all colors, but, owing to the limitations of the human eye and to the fact that the colors produced by the prism blend into one another, we can count only six fairly distinct colors (red, orange, yellow, green, blue, and violet). Upon closer inspection of this color distribution, numerous fine gradations may be distinguished, both between different colors and within any one color itself. For example, red, when it first becomes definitely distinguishable from its neighbor, orange, possesses a different shade than one finds at the other end of the red band where infrared wavelengths are approached.

It is a common experience with all persons who are not color blind to find that objects which possess one color under an electric light may assume a considerably different color when examined in the sunlight. The difference is due to the fact that the color of an object is a function of the wavelengths of the light which the object does *not* absorb. Thus, if we shine white light on a body and none of it is absorbed, we see a white body. However, if under the same white light the object appeared blue, then the object would be absorbing all the other components of white light and be reflecting blue.

To obtain the true color of an object, we must examine it under a light which contains all the wavelengths of the visible spectrum. Thus, a blue

object appears much darker under an ordinary incandescent lamp than it does in sunlight. This is because the lamp has an excess of red light and a deficiency of blue. Since a blue object will reflect only blue rays, it will reflect less light under an incandescent lamp and give a darker appearance. In sunlight, blue and red are present to the same extent and the object assumes its proper color.

With objects that are transparent, the color is determined by the light which is transmitted through the object. Thus, in a green piece of glass, green is permitted to pass through, whereas the other colors are absorbed.

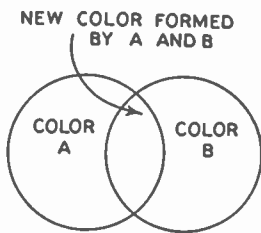


FIG. 18.1. Two circles of light, A and B; where they overlap, they form a third color different from either A or B.

Color Primaries. Anyone who has ever experimented with projector lamps has discovered that when differently colored lights from several projectors are combined, the resultant color seen by an observer will differ in hue from any of the projected lights. Thus, for example, yellow can be formed by combining red and green light; white light can be produced by combining red, green, and blue. The color of the mixed light will appear to the eye as a complete color, and the eye will be unable to distinguish the various components of the mixture that produced that color.

This method of color formation is illustrated in Fig. 18-1. Two circles of colored light are projected onto a screen and positioned so that they overlap to some extent. Within the overlapping region, a new color will be produced by the addition of color A and color B. Where the circles of light do not overlap, each light will retain its original color. If a third circle of light is added, as shown in Fig. 18-2, then additional colors can be obtained. These would be

Color A	Color D (formed from A and B)
Color B	Color E (formed from A and C)
Color C	Color F (formed from B and C)
	Color G (formed from A, B and C)

and each would differ from the other. In the areas where the circles of light overlapped, the eye would not be able to distinguish each of the colors forming the mixture, but instead would see the final color produced. Furthermore, as we varied the relative intensities of colors A, B and C, we would see that colors D, E, F, and G would likewise vary in shade.

The number of different colors that can be formed by the use of three colored lights will depend upon the colors chosen. Experience has indicated that the colors red, blue, and green, when combined with each other in various proportions, will produce a wider range (or gamut) of colors than

any other combination of three colors. Note, however, that, if we used four different colors in our mixing process, we could produce an even wider range of different colors. With the addition of more and more colors to our mixing scheme, the reproducible range would increase somewhat. Obviously, however, a line must be drawn, and the use of three colors has been standardized. The three colors chosen, red, green, and blue are thus referred to as the "primary" colors.

Chromaticity Chart. A diagram which is convenient to use for color mixing is the tongue-shaped (or horseshoe-shaped) curve shown in Fig. 18-3. This is known as a chromaticity chart. The positions of the various spectrum colors from blue at one end to red at the other are indicated around the curve. Any point not actually on the solid-line curve but within the area enclosed by the curve represents not a pure spectrum color but some *mixture* of spectrum colors. Since white is such a mixture, it, too, lies in this area,

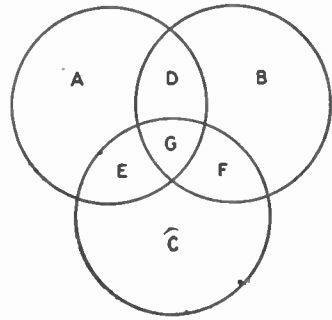


FIG. 18.2. The mixing of three colors, A, B, and C results in four new ones, D,E, F,G.

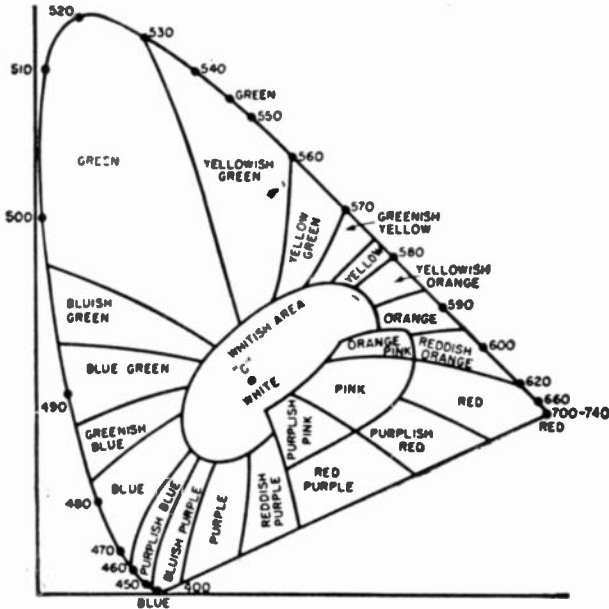


FIG. 18.3. A chromaticity diagram. The numbers listed around the perimeter of the chart represent the wavelengths of the various colors in millimicrons.

specifically, at point C. This particular point was chosen at an international convention in England and is generally referred to as "illuminant C." Actually, of course, there is no specific white light, since sunlight, skylight, and daylight are all forms of white light and yet the components of each differ considerably. The color quality of a conventional black-and-white television receiver tube is represented by some point in the central region of the diagram about point C.

(If Fig. 18-3 were reproduced in full color, it would be seen that the color changes gradually from point to point. The deepest and most intense

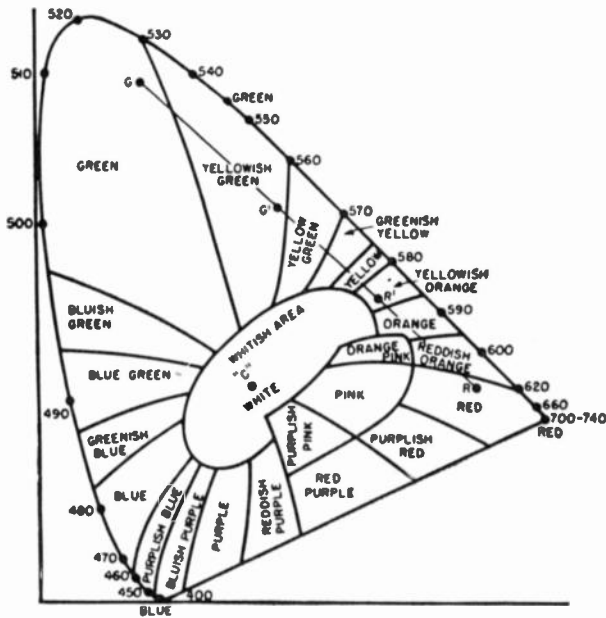


FIG. 18.4. The line drawn between points "R" and "G" passes through all the colors that can be obtained by mixing these two shades of the red and green hues.

colors are obtained at the outer edge of the diagram. Here we find the real deep red, deep blue, and deep green shades which we actually see very seldom in everyday life. More familiar are the lighter colors, appearing as we move in toward the center. These are the pastels such as pink, light green, pale blue, etc. Finally at the center come the whites with point C as the reference white, or, for our purpose here, the "whitest" white. Actually this is a rather nebulous shade, entirely arbitrary in value and simply chosen for certain conveniences.)

The chromaticity chart lends itself readily to color mixing because a straight line joining any two points on the curve will indicate all the

different color variations that can be obtained by combining these two colors additively. Thus, in Fig. 18-4 consider the line drawn connecting points R and G representing certain shades of red and green respectively. If there is more red light than green light, the exact point representing the new color will lie on the line, but be closer to R than G. Point R_1 might be such a color. On the other hand, if a greater percentage of green light is employed, the new color will still lie on the line connecting R and G, but now be closer to G than R. Point G_1 might be such a color. This method can be carried out for any two colors on the chart.

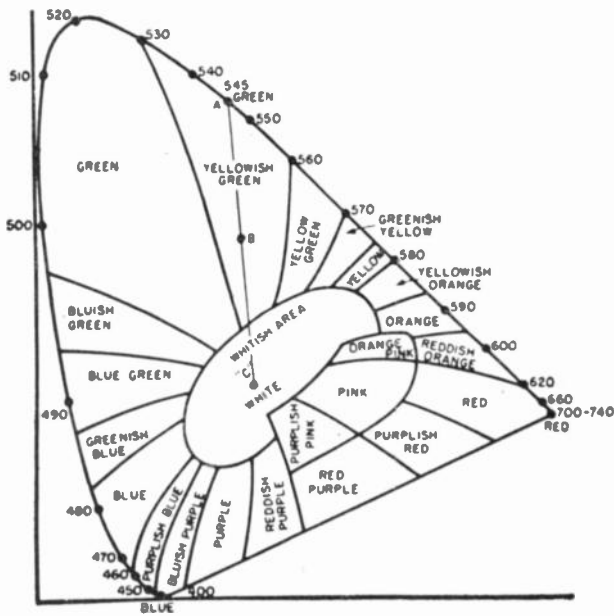


Fig. 18.5. When moving from "A" to "C," the green becomes less and less saturated, or lighter in intensity.

It is possible to specify the purity of a color by its distance from point C. Thus, consider point B in Fig. 18-5. This is halfway along the line between point C (white) and point A (green). Hence, point B represents a mixture of green diluted 50 per cent with white light, and we can say that the purity of this color is 50 per cent. Had the distance between point C and point B been 75 per cent of the total distance between point C and point A, we would have stated that the purity of the color at point B was 75 per cent. As point B moves closer and closer to the spectrum curve, the purity of the color it represents increases until it becomes 100 per cent at

the curve — point A. As point B moves closer to point C, its purity decreases. At point C, the purity is said to be zero.

In place of purity, the word *saturation* is frequently employed. Any point located on the tongue-shaped curve is said to be completely saturated. As we leave the curve and approach closer to point C, more and more white light is added to the color and it becomes less saturated or, what is the same thing, more desaturated. And, at point C, the saturation is zero.

In connection with saturation, the word *hue* is frequently heard. Hue represents color, such as red, green, orange, etc. The term is associated with color wavelength, and when we call a certain color green, or orange, or red, we are specifying its hue. Thus, hue refers to the basic color, as it appears to us, while saturation tells us how deep the color is. If the color is highly saturated, we say that it is a deep color, such as deep red or deep green. If it contains a considerable amount of white light, we say it appears faded, as a faded red or a faded green.

THE NTSC COLOR TELEVISION SYSTEM

We are now ready to investigate the NTSC color television system, the system officially adopted by the FCC. This system has been so designed that its signal occupies no more than 6 mc (video and sound) and it carries not only the full black-and-white (or monochrome) signal but, in addition, the color information.

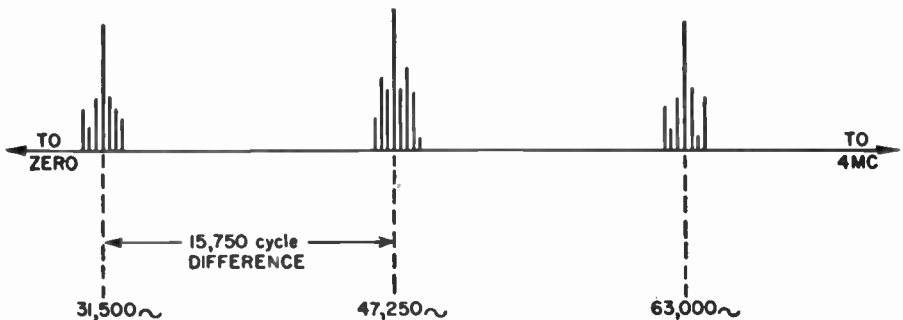


Fig. 18.6. Spectrum distribution of a monochrome signal.

The question is: "How is all this information compressed into a 6-mc bandwidth?" The answer is found in the nature of a television signal. It was discovered as far back as 1929 that a video signal does not occupy every "cycle" of the 4 mc assigned to it. Rather it appears in the form of clusters of energy located near harmonics of the 15,750-cycle line scanning frequency. The energy is grouped around these points, with relatively wide

gaps between them. (See Fig.18-6.) Since these empty spaces are not being used, they can be employed for the transmission of additional information, and here is specifically where the color information of the NTSC color television signal is placed. (See Fig. 18-7.) The practice of placing the energy of one signal between the clusters of energy of another signal is known as interleaving.

Thus, a total color signal consists of two components — a monochrome signal and the signal which carries information concerning the color. Let us examine each component, in turn.

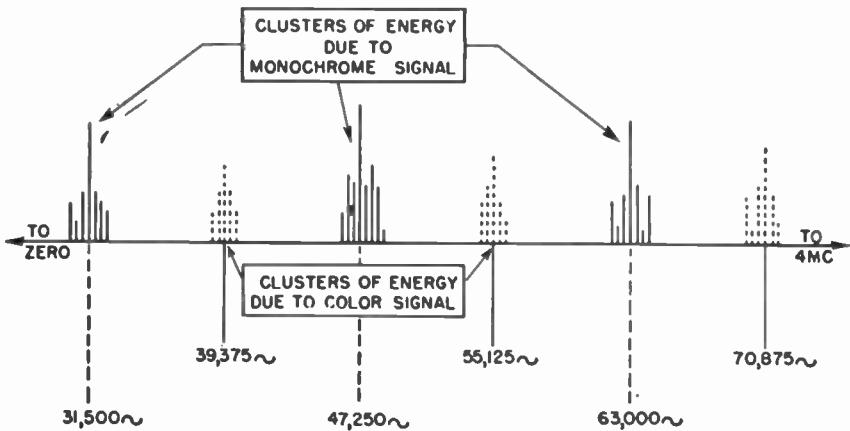


FIG. 18.7. The energy of the color signal is inserted in the gaps between the energy of the monochrome signal.

The Monochrome Signal. The black-and-white, or monochrome, portion of the total color signal is equivalent in all respects to present black-and-white signals. It is formed by combining the red, green, and blue signals from their respective color cameras in the proportions of:

$$Y = 0.59G + 0.30R + 0.11B$$

where

Y = a mathematical symbol representing the monochrome signal,

G = green signal,

R = red signal,

B = blue signal.

This particular proportion was chosen because it closely follows the color sensitivity of the human eye. That is, if you take an equal amount of green light, an equal amount of red light, and an equal amount of blue light and superimpose the rays from these lights on a screen, you will see

white. However, if you then look at each light separately, the green would appear to be twice as bright as the red, and six to ten times as bright as the blue. This is because the eye is more sensitive to green than to red and more sensitive to red than to blue. It is in recognition of this fact that the proportions given above were chosen.

Thus, the monochrome signal is composed of 59 per cent green signal (that is, 59 per cent of the output of the green camera), 30 per cent red, and 11 per cent blue and contains frequencies from 0 to 4 mc. (The use of the letter *Y* to denote the monochrome portion of the color signal is a common practice and should become familiar to the reader.)

Other names for this monochrome signal are luminance or brightness signal. These terms were chosen because they perhaps indicate more clearly the action of this signal. Every monochrome video signal contains nothing but the variations in amplitude of the picture signal, and these amplitude variations, at the picture tube, produce changes in light intensity at the screen.

The Color Signal. The second component of the television signal is the color signal itself. This, as we have just seen, is interleaved with the black-and-white signal. To determine what information this portion of the total signal must carry, let us first see how the eye reacts to color, since it is the eye, after all, for which the color image is formed.

A number of men have investigated the color characteristics of the human eye, and, briefly, here is what they found. The typical human eye sees a full color range only when the area or object is relatively large. When the size of the area or object decreases, it becomes more difficult for the eye to distinguish between various colors. Thus, where the eye formerly required three primary colors, now it finds that it can get by very well with only two. That is, these two colors will, in different combinations with each other, provide the limited range of colors that the eye needs or can see in these medium-sized areas.

Finally, when the detail becomes very small, all that the eye can discern are changes in brightness. Colors cannot be distinguished from gray, and, in effect, the eye is color blind.

These properties of the eye are put to use in the NTSC color system. First, only the large and medium-sized areas are colored; the fine detail is rendered in black-and-white. Second, as we shall see later, even the color information sent is regulated according to bandwidth. That is, the larger objects receive more of the green, red, and blue than the medium-sized objects.

The color signal takes the form of a subcarrier and an associated set of sidebands. The subcarrier frequency is approximately 3.58 mc. This represents a figure which is the product (approximately) of 7875 cycles

multiplied by 455. The value 7875 is one half of 15,750, and if we use an odd multiple (1, 3, 5, etc.) of 7875 as a carrier, then it will fall midway between the harmonics of 15,750 cycles. If we used even multiples of 7875, we would end up with 15,750 or one of its harmonics, and this would place the color signal at the same points (throughout the band) as those occupied by the black-and-white signal. Refer back to Fig. 18-7. By taking an *odd* multiple of 7875, we cause the second signal to fall in between the bundles of energy produced by the first signal and the two do not interfere.

Now that we have a color carrier (or subcarrier, as it is known), the next step is to provide it with enough modulation to enable the receiver to develop a color picture. Ordinarily, the information required would con-

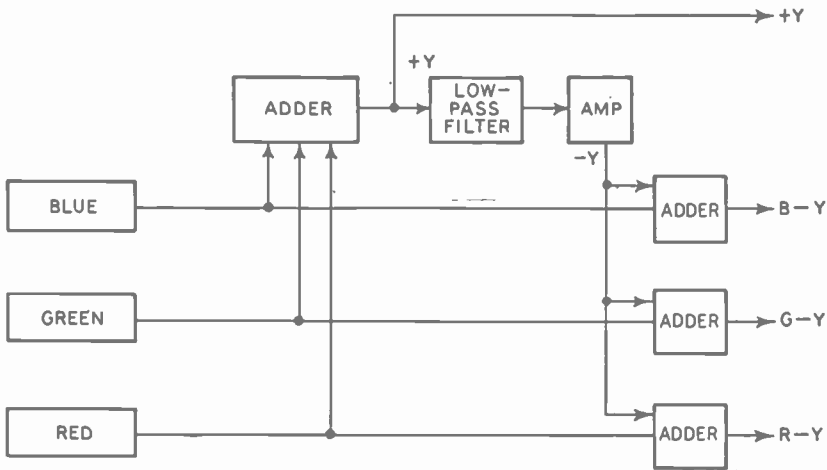


FIG. 18.8. Block diagram illustrating how color-minus-brightness signals are formed.

sist of R , G , and B since these are the three primary colors from which all of the other colors are derived. This means modulating the color subcarrier with three different voltages. Actually, however, we can do the same job by using only two quantities if we resort to the following modification. Take the R , G , and B voltages and combine each with a portion of the monochrome signal after the latter has been inverted 180 degrees. This produces $R-Y$, $G-Y$, and $B-Y$ signals. We can do this by taking a portion of the brightness signal (Y signal) and passing it first through a low-pass filter. (See Fig. 18-8.) This permits only the low-frequency components to get through, which is satisfactory since the color signals are also concerned only with the low frequencies. Then the brightness signal is passed through an amplifier and inverted. This gives us the desired $-Y$. This is then added to each of the three color signals or voltages to produce a $G-Y$, an $R-Y$, and a $B-Y$ signal.

At the receiver, the original R , G , and B can be reobtained by adding Y to $G-Y$ to obtain G , by adding Y to $R-Y$ to get R , and by combining Y with $B-Y$ to get B .

Thus far, it would seem that we have only exchanged R , G , and B for $R-Y$, $G-Y$, and $B-Y$. However, once this is done, it turns out that, instead of requiring all three color-difference signals, all we really need are two, say $R-Y$ and $B-Y$. This is so because G information is already present in the Y or brightness signal since the latter contains voltages from all three colors ($Y = 0.59G + 0.30R + 0.11B$). Hence, if we send along only $R-Y$ and $B-Y$ in the color signal to the receiver, we can use these to obtain the $G-Y$ information we need.

Thus, we now have only two pieces of color information to send, and somehow the 3.58-mc color subcarrier must be modulated by $R-Y$ and $B-Y$ voltages without conflict to each other.

The best solution to this problem, designers found, was to take the $B-Y$ and $R-Y$ signals and apply each to a separate modulator. At the same time, 3.58-mc carriers were also applied to each modulator, but with this difference. Their frequencies were the same, but one carrier was 90 degrees

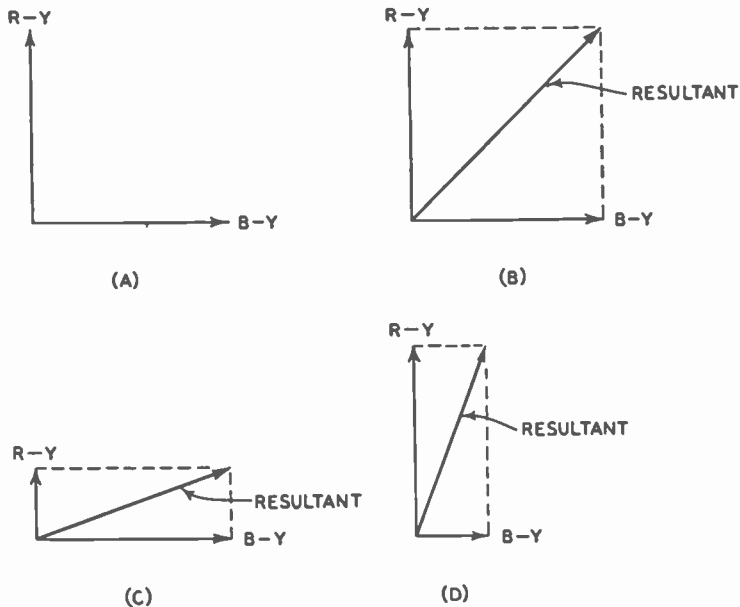


FIG. 18.9. The angular position and amplitude of the resultant carrier for various amplitudes of $B-Y$ and $R-Y$. (A) the $R-Y$ and $B-Y$ vectors; (B) the resultant when $R-Y$ and $B-Y$ are equal; (C) when $B-Y$ is stronger than $R-Y$; and (D) when the $R-Y$ signal is stronger than the $B-Y$ signal.

out of phase with the other. After the carriers were amplitude-modulated, they were then combined to form a resultant carrier.

This is best illustrated by means of vectors. In Fig. 18-9A, the $B-Y$ vector represents the $B-Y$ modulated carrier; the $R-Y$ vector represents the carrier modulated by the $R-Y$ voltage. When these voltages or signals are combined, a resultant is formed. If the $R-Y$ and $B-Y$ signals are equally strong, the resultant will occupy the position shown in Fig. 18-9B. If the $B-Y$ signal is predominant, the resultant will be drawn closer to it. (See Fig. 18-9C.) On the other hand, if the $R-Y$ signal is the stronger, the position of the resultant vector will shift toward it. (See Fig. 18-9D.) Thus, we can see that the phase angle of the resultant will be governed by the coloring or hue of the picture, whereas the amplitude (or length) of the vector will determine how saturated the colors are.

This particular fact is of great importance in the receiver because, if we somehow change the phase of the resultant with respect to $B-Y$ or $R-Y$, then the colors reproduced on the screen will be incorrect. Hence, present circuit designs incorporate a special phasing control which enables us to compensate for any phase shift that may occur. The position of this control in the circuit will be discussed presently.

Note that the $B-Y$ and $R-Y$ signals amplitude-modulate their separate carriers prior to the addition, and so each modulated signal possesses a 3.58-mc carrier and a series of sidebands (like every A-M signal). When the resultant is formed, the sidebands are brought along with it.

If we were to pause now and reconstruct our total color signal, here is what we would find. First, there would be the Y , or monochrome, signal and it would extend over the entire video frequency range from 0 to 4.0 mc. Second, there would be a color subcarrier, with a frequency of 3.58 mc. This carrier is modulated by the $R-Y$ and $B-Y$ signals, and the modulation intelligence is contained in a series of sidebands that stretch away above and below 3.58 mc. Just how far above and below is dependent on the band of frequencies contained in the $R-Y$ and $B-Y$ modulating voltages. It was discovered that the eye is satisfied by the color image produced if we include color information only up to 1.5 mc, while the portion of the image from 1.5 mc to 4.0 mc is rendered in black and white. Hence the sideband frequencies of the color-modulating voltages (so far, $R-Y$ and $B-Y$) need extend only from 0 to 1.5 mc. Furthermore, we can even modify this set of conditions somewhat because the three primary colors are required only for large objects or areas, say, those produced by video frequencies up to 0.5 mc. For medium-sized objects, say, those produced by video frequencies from 0.5 to 1.5 mc, only two primary colors need be employed.

In other words, to take advantage of this situation, we need two color signals: one of which has a bandpass only up to 0.5 mc, while the other has

a bandpass from 0 to 1.5 mc. The next problem, then, is to determine the composition of these two color signals.

To appreciate the answer to this problem, let us return to the vector diagram (Fig. 18-9) showing the $R-Y$ and $B-Y$ signals. This diagram is redrawn in Fig. 18-10A, and we have added the equivalent equation for Y .

$$Y = 0.59G + 0.30R + 0.11B$$

For $R-Y$, then, we have

$$R-Y = R - 0.59G - 0.30R - 0.11B$$

or

$$R-Y = 0.70R - 0.59G - 0.11B$$

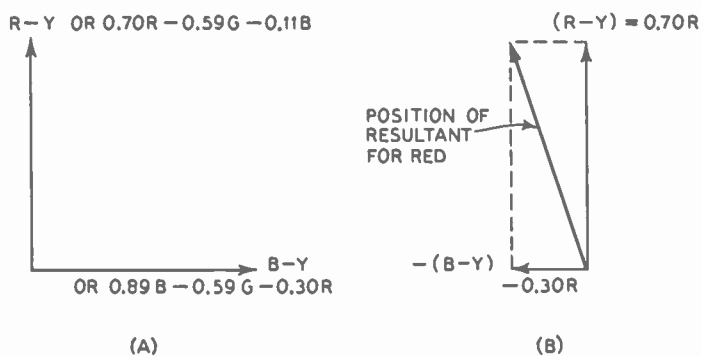


FIG. 18.10. How color determines the position of a resultant vector. (A) equations showing compositions of $B-Y$ and $R-Y$ in terms of R, G , and B ; (B) position of signal vector when red field is being scanned. (See Text)

and, for $B-Y$, we obtain

$$B-Y = B - 0.59G - 0.30R - 0.11B$$

or

$$B-Y = 0.89B - 0.59G - 0.30R$$

This means that the $R-Y$ and $B-Y$ vectors contain R, G , and B voltages in the proportions shown.

Now, let us suppose that the color camera is scanning a scene containing only red. Then, no green or blue voltages would be present and the $R-Y$ signal becomes simply $0.70R$, while the $B-Y$ signal reduces to $-0.30R$.

This set of conditions is shown in Fig. 18-10B, with the position, too, of the resultant vector. In other words, this is the position the vector would occupy when red only is being sent.

By following the same process, we can obtain the position that the resultant vector occupies when only green is being sent, or blue, or any

other color formed by combining these three colors in any combination. A number of colors are shown in Fig. 18-11, and we see, perhaps more clearly than before, how the phase of the color subcarrier changes as the color to be transmitted varies. To reiterate, then: The phase angle of the resultant is governed by the coloring of the picture, whereas the amplitude (or length) of the vector determines how saturated the colors are.

The designers of the NTSC system found that, while they could use $R-Y$ and $B-Y$ for the color signals, better system operation would result if they chose two other signals situated not far from the $R-Y$ and $B-Y$ signals* These two other signals were labeled I and Q signals, and their position with respect to $R-Y$ and $B-Y$ is shown in Fig. 18-12.

Thus, where before we had $R-Y$ and $B-Y$ voltages modulating the 3.58-mc color subcarrier, we now substitute I and Q signals. Furthermore, the Q signal possesses frequencies up to 0.5 mc, whereas the I signal is permitted to have signals up to 1.5 mc.

Now, what do we gain from this? For all color signal frequencies up to 0.5 mc, both I and Q are active and since they are 90 degrees apart as were $R-Y$ and $B-Y$, they will act just the way $R-Y$ and $B-Y$ acted. That is, they will produce, in combination with each other, all of the colors shown in Fig. 18-12. Hence, whether we use I and Q or $R-Y$ and $B-Y$ as our modulating voltages for color frequencies up to 0.5 mc, we obtain precisely the same results.

Consider, however, the situation for color frequencies from 0.5 mc to 1.5 mc. The Q signal drops out, and only the I signal remains to produce color on the picture tube screen. From

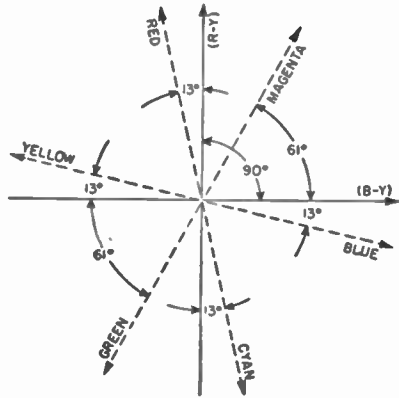


FIG. 18.11. The phase of the the color subcarrier depends upon the color to be sent.

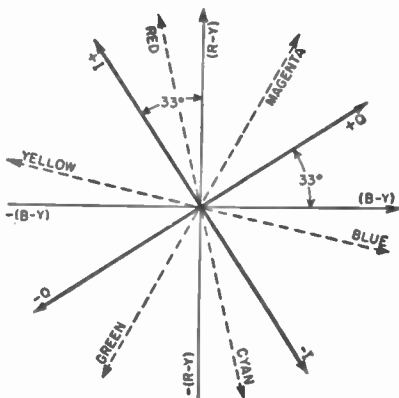


FIG. 18.12. The positions of the I and Q signals with respect to $R-Y$ and $B-Y$.

* Some receivers operate as $R-Y, B-Y$ systems. In this way they gain certain advantages of economy. $I-Q$ sets, however, produce truer color pictures.

Fig. 18-12 we see that positive values of the I signal will produce colors between yellow and red or actually a reddish orange. On the other hand, negative values of I will produce colors between blue and cyan or, in general, in the bluish-green range. Hence, when only the I signal is active, the colors produced on screen will run the gamut from reddish-orange to bluish-green.

But why do we want this? If you go back to an earlier paragraph, you will recall that, for medium-sized objects (say, those produced by video signals from 0.5 mc to 1.5 mc), the sensitivity of the eye is reduced. Actually, for medium-sized objects, it was found that the eye is sensitive principally to the bluish-greens or the reddish-oranges. The NTSC signal (via its I component) is fashioned to take advantage of this fact by producing only blue-greens or reddish-oranges for medium-sized objects.

We are now in a position to consider the color signal in all its aspects.

1. There is a monochrome signal with components that extend from 0 to 4 mc. This is the Y signal.

2. The color subcarrier frequency is set at 3.58 mc (actually it is 3.579545 mc).

3. This color subcarrier is modulated by two color signals called the I and Q signals.

4. The Q signal has color frequencies that extend from 0 to 500 kc, or 0.5 mc. This means that the upper Q sideband extends from 3.58 mc up to $3.58 + 0.5$, or 4.08 mc. The lower Q sideband goes from 3.58 mc down to $3.58 - 0.5$, or 3.08 mc.

5. The I signal has color frequencies that extend from 0 to 1.5 mc. When this modulates the color subcarrier, upper and lower sidebands are formed. The lower sideband extends from 3.58 mc down to $3.58 - 1.5$, or 2.08 mc. If the full upper sideband were permitted to exist, it would extend all the way up to $3.58 + 1.5$, or 5.08 mc. Obviously this would prevent the use of a 6.0-mc overall band for the television signal (video and sound). To avoid this spilling over beyond the limits of the

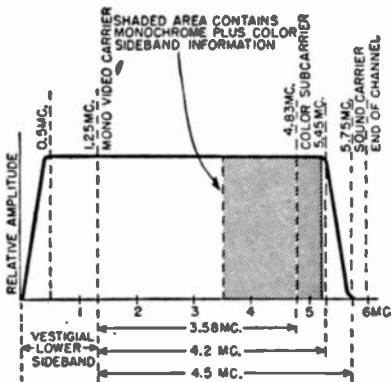


Fig. 18.13. The distribution of the full color signal within its allotted band.

already established channels, the upper sideband of the I signal is limited to about 0.6 mc. This brings the upper sideband of the I signal to 4.2 mc. The video pass band then ends rather sharply at 4.5 mc. (See Fig. 18-13.)

There is one further fact of importance in the make-up of a color tele-

vision signal and this concerns the color subcarrier. We know that the 3.58-mc carrier is modulated by the *I* and *Q* color signals. Now, in conventional modulation methods, both the carrier and the sidebands are present when the signal is finally sent out over the air. The intelligence (or modulation) is contained in the sidebands and actually is all that interests us. However, the carrier is sent along because it is required in the receiver to reverse the modulation process and re-create the original modulating voltages.

In the NTSC color system, the color subcarrier is not sent along with its sidebands (after the latter have been formed). Instead, it is suppressed by using a balanced modulator. This particular practice is followed for two reasons. First, by suppressing the color subcarrier, we reduce the formation of a 920-kc beat note between it and the 4.5-mc sound carrier, which is also part of every television broadcast. This 920-kc note would appear as a series of interference lines on the face of the picture tube. Now, it is true that the color sidebands are present and that they can (and do) beat with the 4.5-mc sound carrier to produce similar low-frequency beat notes. However, in any signal, the carrier usually contains far more energy than any of its sidebands; hence, when we suppress the carrier, in effect we are suppressing the chief source of this interference. Whatever other interference may be produced by some of the stronger sidebands near 3.58-mc can be more easily dealt with by using traps in the I.F. system. This will be seen when we examine the circuitry of a receiver.

The second reason for using the suppressed carrier method is that it leads to an automatic removal of the entire color signal when the televised scene is to be sent wholly as a black-and-white signal. For when this occurs, *I* and *Q* decrease to zero, and since the balanced modulators suppress the carrier, no color signal at all is developed.

When these advantages of carrier suppression comes one disadvantage.

When the color sidebands reach the color section of the receiver, a carrier must be reinserted in order to permit detection to take place. Offhand, one might suppose that an oscillator operating at 3.58 mc would be needed. This is one requirement. A second and vitally important consideration is the phase of this reinserted carrier. Remember that, at the transmitter,

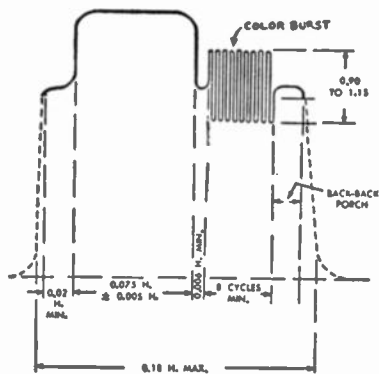


FIG. 18.14. The position of the color burst for subcarrier oscillator sync on the back porch of a horizontal sync pulse.

attention was given to the phase of I and Q as they were introduced into the modulator. If the same relative phase were not maintained in the reinserted carrier, the colors obtained at the output of the color circuits would not possess the proper hue.

To provide information concerning the frequency and phase of the missing color subcarrier, a color burst is sent along with the signal. This burst follows each horizontal pulse and is located on the back porch of each blanking pedestal. (See Fig. 18-14.) It contains the minimum of 8 cycles of the subcarrier and it is phased in step with the color subcarrier used at the station. In the receiver, this burst is used to lock in the frequency and phase of a 3.58-mc oscillator, and thus we are assured at all times that the reinserted carrier will correctly do its job when it recombines with the color sidebands.

COLOR TELEVISION RECEIVERS

A block diagram illustrating the major divisions of a color television receiver is shown in Fig. 18-15. Roughly there are nine such sections:

1. R.F. tuner.
2. Video I.F.
3. Sound I.F., ratio detector, and audio amplifiers.
4. Video detector and video amplifiers.
5. Chrominance section.
6. Color sync circuits.
7. Sync separators and A.G.C.
8. Vertical and horizontal deflection and high voltage.
9. Tri-color picture tube and associated convergence circuits.

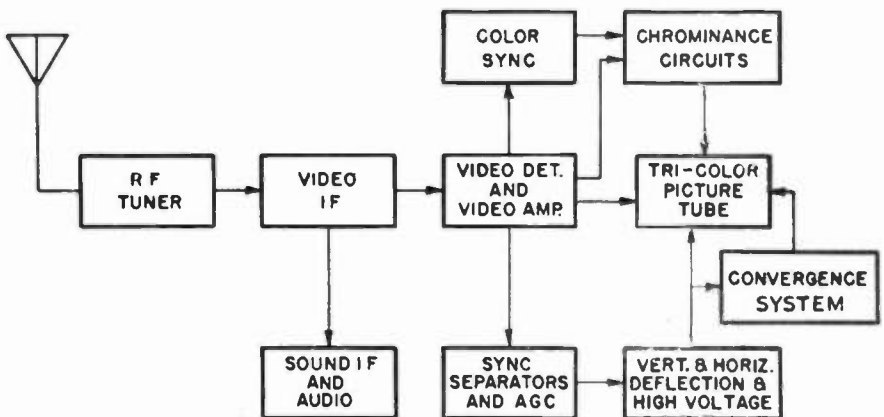


FIG. 18.15. Color television receiver master block diagram.

If we compare this briefly with a conventional black-and-white receiver, we see that all items with the exception of numbers 5, 6, and 9 are common to both types of receivers. Therefore, the circuitry in sections 1, 2, 3, 4, 7, and 8 should contain many familiar aspects to anyone who has worked with black-and-white receivers. That this is so will become evident as we proceed through a detailed description of a color television receiver.

(The actual circuits that the blocks represent are given in Fig. 18-36 at the end of this chapter. After the material has been covered once, reread the chapter, this time linking each block with its associated circuit.)

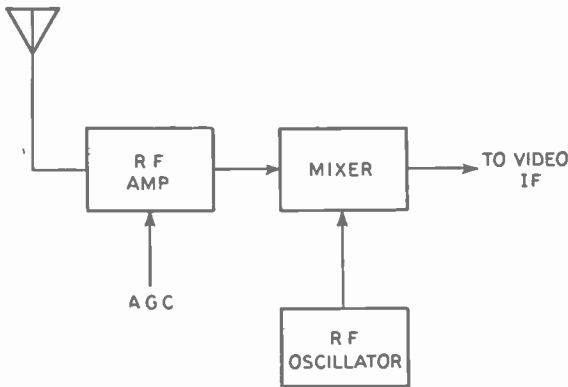


Fig. 18.16. Block diagram of R. F. tuner for color receiver.

1. R.F. Tuner. The R.F. tuner consists of an R.F. amplifier, an oscillator, and a mixer stage. (See Fig. 18-16.) This section of the receiver is similar to that employed in the black-and-white receivers since the R.F. requirements of both types of sets are alike. Any station which is now sending monochrome signals is permitted by the F.C.C. to send color broadcasts, provided the necessary equipment is installed.

The circuits ordinarily employed in the tuner consist of a cascode amplifier for the R.F. stage and a modified Colpitts circuit for the oscillator. The mixer combines both incoming and oscillator signals and from them produces the difference or I.F. signals. These are then fed to the video I.F. system.

No UHF tuning circuits are shown in Fig. 18-16, but they would be treated in the same way as they are in current black-and-white receivers.

2. Video I.F. System. The video I.F. system in color television receivers will, in general, contain four and sometimes five separate amplifiers (see Fig. 18-17). In form these will closely resemble the I.F. section of of a monochrome receiver because the change to color reception in no way

has altered the basic function of this section of the receiver, namely, that of establishing the overall bandpass and sensitivity of the receiver.

The stages in the video I.F. system are stagger-tuned, generally in the 41-to-46-mc range, with suitable traps for the accompanying sound (41.25 mc), for the sound carrier of the adjacent lower channel (47.25 mc), and for the video carrier of the adjacent higher channel (39.75 mc). The sound carrier of the same channel (41.25 mc) generally has more than one trap to insure that this signal is kept down at the proper level. Failure to observe this precaution will tend to produce a noticeable 920-kc beat on the picture tube screen, especially in those sections where the colors are highly saturated.

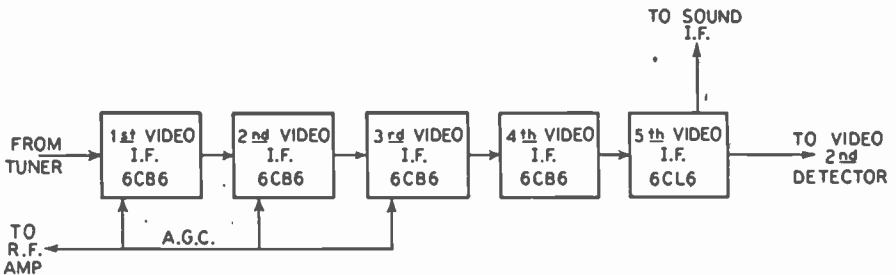


FIG. 18.17. The video I. F. system.

The overall I.F. response curve extends to approximately 4.2 mc (at the 50 per cent points) in order to include the color subcarrier and all its sidebands. It will be remembered from the previous discussion that the upper sidebands of the color subcarrier extend about 0.6 mc above 3.58 mc, or up to 4.2 mc. Hence, to reproduce the picture in full and true color, it is necessary that the upper color sidebands be permitted to pass. Beyond 4.2 mc, the I.F. response drops sharply to the level of the sound carrier.

A.G.C. voltage is applied to two or three of the I.F. amplifiers in order to control their gain in accordance with the level of the received signal. A portion of the same voltage is also fed to the control grid of the R.F. amplifier for similar control.

From the video I.F. system, the signal is generally applied to two points: The video second detector and the sound system. Let us consider the latter first.

3. Sound I.F., Ratio Detector, and Audio System. It is the practice in color television receivers to separate the sound and video carriers just as soon as it becomes feasible to do so. In Intercarrier color receivers, this generally occurs around the fourth or fifth I.F. stage. The reason for this early separation, of course, is to keep the amount of 920-kc beat inter-

ference voltage as small as possible so that its effect on the screen will be negligible.

The practice generally followed is to employ a separate detector (usually a germanium crystal diode) in which the sound and video carriers are mixed to produce the 4.5-mc sound I.F. signal. (See Fig. 18-18.) This signal is then amplified by one or two stages and applied to a ratio detector from which we obtain the original audio intelligence. This is then followed by a voltage amplifier, a power amplifier, and, finally, the loudspeaker. This system is identical in all respects to the sound section of present monochrome receivers.

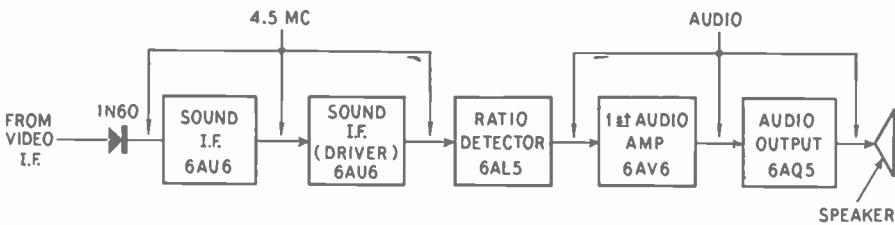


FIG. 18.18. Sound I. F., detector and audio amplifier stages.

4. Video Detector and Video Amplifiers. Returning to the video system, we find that the signal enters the video second detector after leaving the I.F. stages. Here the signal is demodulated, giving us back the 0-to-4-mc monochrome or luminance signal, and the *I* and *Q* color sidebands. Present, too, are the various synchronizing pulses plus the color burst. The latter, it will be recalled, is needed to reestablish the proper frequency and phase of the missing color subcarrier.

At the video detector output, a number of things occur. First, the brightness portion of the total signal must be fed to a separate amplifier. Second, the *I* and *Q* color sidebands must be separated from the full signal and transferred to a separate chrominance section. Third, the color burst must be made available to the color sync circuits. And, finally, there are the sync separator and A.G.C. systems that must also be tied into the signal path.

There are a number of different ways in which all of the foregoing functions can be carried out. The block diagram in Fig. 18-19 illustrates one suitable method. The signal from the video second detector is fed to a video amplifier. Here both chroma and monochrome signals are amplified. The monochrome signal is then transferred to a video second amplifier and from this stage to the matrix network. The chroma signal is taken from the cathode circuit of the first video amplifier and transferred to the band-pass amplifier which stands at the head of the chrominance section.

A 3.58-mc resonant network in the plate circuit of the first video amplifier transfers the 3.58-mc burst to a burst amplifier. The same arrangement also attenuates the amount of 3.58-mc voltage reaching the second video amplifier.

Signal voltages for the sync and A.G.C. circuits are also obtained at the plate of the first video amplifier. Also, a 1.0-microsecond delay line is inserted in the path of the luminance signal between the first and second video amplifiers. The delay line is terminated in a 3000-ohm potentiometer which serves as a contrast control for the luminance signal.

A contrast control for the chrominance portion of the signal is mechanically ganged to the luminance contrast, thereby insuring that both signals will be varied in equal amount. This is needed to maintain the proper voltage relationship between the two signals.

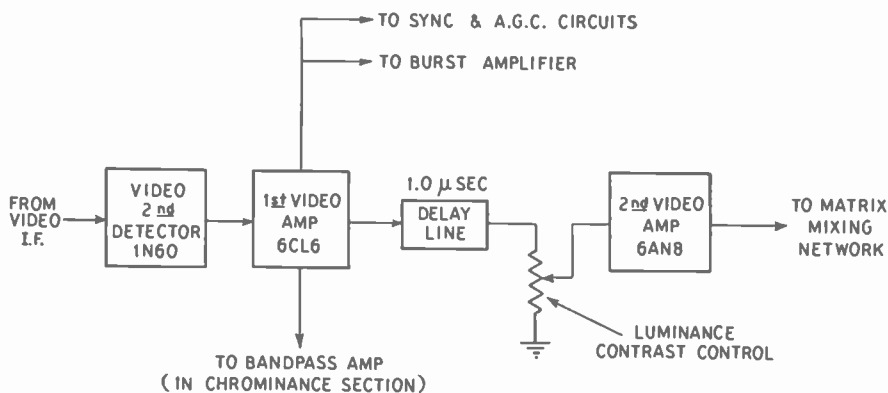


FIG. 18.19. Block diagrams of video second detector and video amplifiers (brightness channel).

A 4.5-mc trap in the cathode of the first video amplifier attenuates any 4.5-mc voltage that may develop in the video detector through the beating of the video and sound carriers.

Following the luminance signal first, we find that after the 1-microsecond delay line the signal is further amplified and then applied to a matrix network. (We will discuss this network in more detail presently.) The need for the delay line arises from the fact that the *Q* signal passes through a rather narrow bandpass filter in its system which acts to slow down its passage. To avoid having the *I* and *Y* (or luminance) signals arrive at the matrix before the associated *Q* signal, each of these other signals must be similarly delayed. In the luminance channel this is done by inserting an artificial delay line having a 1-microsecond delay. After the luminance signal has passed through the delay line, it is transferred to the matrix

where it combines with the I and Q signals to reform the red, green, and blue voltages required by the picture tube.

It is interesting to note that the bandpass of the luminance channel beyond the color take-off point is usually reduced below 3.58 mc. This is done to minimize further any visual dot pattern which the 3.58-mc color subcarrier signal may possibly develop on the face of the picture tube screen.

5. Chrominance Channel. We come now, for the first time, to a section of the receiver which has no counterpart in any monochrome receiver and which is completely devoted to color. This usually is known by the name of chrominance channel and it covers a considerable number of stages. Its principal function is first to demodulate the color signal so that the *original* I and Q voltages are reobtained. Then, these signals, in conjunction with the Y signal, are fed to a mixing network (the matrix) and combined in such a fashion that the red, green, and blue components contained in the I , Q , and Y signals are reobtained. These three voltages are then amplified and fed to separate control grids of a tri-gun color tube.

This, then, is the function of the chrominance channel. Now let us study the system in greater detail. (See Fig. 18-20.)

The full signal is obtained from the video system and fed to a bandpass amplifier. Also applied to this amplifier is a gating pulse which keys the amplifier off by applying a pulse derived from the horizontal deflection output transformer. The pulse arrives only during the horizontal retrace interval when the color burst is passing through the system. By gating out the burst from the bandpass amplifier, we prevent color background unbalance in the picture tube. This unbalance may arise if the d-c restorers in the chrominance channel clamp on to the color bursts rather than on to the tips of the sync.

The bandpass amplifier also receives a d-c biasing voltage from a stage known as a color killer. This stage is located in the color sync section of the receiver, and its purpose is to bias the bandpass amplifier to cut-off in the absence of a color signal (that is, when black-and-white signals only are being received). This precaution is taken to insure that no random color appears on the picture tube screen during a monochrome transmission. Such random color would be produced by noise or monochrome signals reaching the I and Q demodulators.

In the plate circuit of the bandpass amplifier there is a bandpass filter which permits signals from 2.1 to 4.1 mc to pass, but strongly attenuates all others. This filter thus serves to separate that portion of the signal containing the color sidebands from the section of the signal containing only monochrome information.

The bandpass filter is terminated in a color intensity control (called the

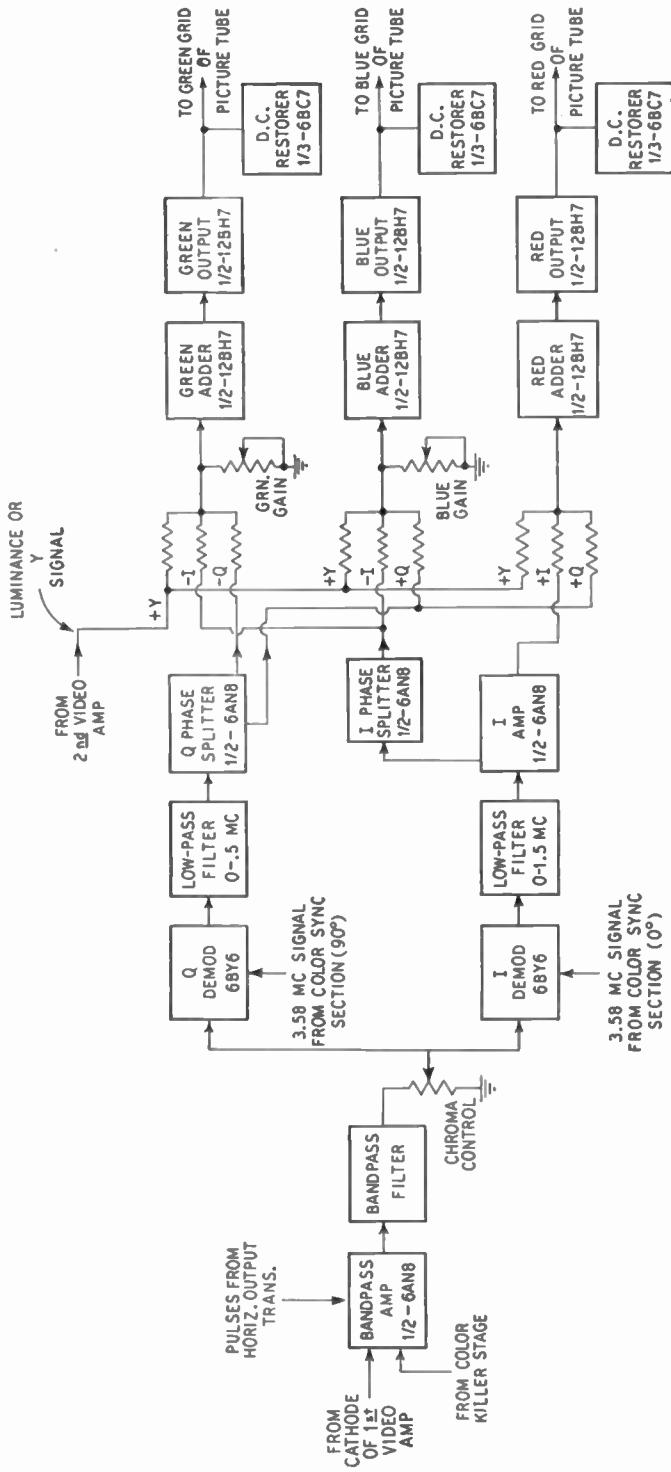


Fig. 18.20. The chrominance section of a color receiver. No special delay line is shown for the I stages because the required delay is in distributed form.

chroma control) from which point we can take as much of the color signal as we feel is required and feed it to the following chroma amplifier. This color intensity control actually determines how saturated (deep) the colors appear on the screen. It is a front panel control available to the set user to adjust as he sees fit.

The chroma stage following this control amplifies the color signal and then feeds it, in equal measure, to the *I* and *Q* color demodulators. Also arriving at these demodulator stages is another voltage: a 3.58-mc signal.

The 3.58-mc signal represents the missing color subcarrier and it must be recombined with each *I* and *Q* signal before demodulation or detection can take place. Both *I* and *Q* stages receive a 3.58-mc voltage, the only difference being that one 3.58-mc voltage lags the other 3.58-mc voltage by 90 degrees. This particular phase relationship is required because it is the way the color signal was modulated at the transmitter.

With a color signal being received, the entire chrominance channel is operative, and the output of the two demodulators represents the original *I* and *Q* color signals that were originally developed at the transmitter. The *I* signal, then, is passed through a 0–1.5-mc bandpass filter and a special 0.5-microsecond delay line. After this it may receive additional amplification before being made available to the mixing or matrix network in positive and negative polarity. The double polarity *I* signals are required in the final mixing process from which red, green, and blue voltages are re-created. A single-phase splitter provides the positive and negative *I* signals.

The use of a 0.5-microsecond delay network in the *I* channel again stems from the narrow 0-to-0.5-mc bandpass filter through which the *Q* signal is sent. The *Y* signal, it will be remembered, had to be delayed 1 microsecond for the same reason. The difference in delay between the *Y* and *I* signals arises from the different characteristics of their respective networks. In the *Y* channel, the bandpass of the circuits extends from 0 up to 3.5 or 4.0 mc. In the *I* channel, the bandpass is only from 0 to 1.5 mc. The narrower bandpass introduces some delay, requiring less additional delay in order to slow the *I* signal down to the *Q* signal.

In the *Q* channel, the demodulated *Q* signal passes through a 0-to-0.5-mc bandpass filter and reaches a phase splitter from which positive and negative *Q* signals are made available to the matrix.

We now have at the matrix the *I*, *Q*, and *Y* signals and, by properly combining them, we can reobtain the red, green, and blue voltages that were originally combined to form the *I*, *Q*, and *Y* signals.* The addition is

* Remember that the original *R*, *G*, and *B* voltages at the transmitter were manipulated to form the *I*, *Q*, and *Y* signals. Here, in the receiver matrix, the reverse action is taking place.

carried out in rather simple fashion by using a series of resistors connected as shown in Fig. 18-21. At the output of the matrix section, each of the three color voltages is separately amplified by triode or pentode amplifiers and then transferred via separate d-c restorers to the appropriate control grid of the tri-gun picture tube.

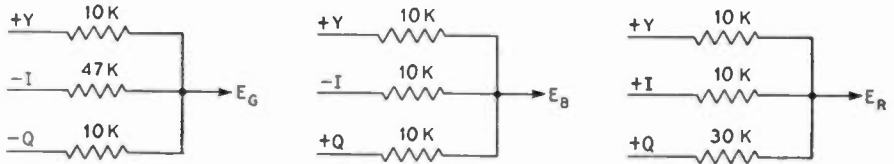


Fig. 18.21. Matrix network of color receiver (Fig. 18.36) to add I , Q , and Y signals to re-create original red, green, and blue signal voltages.

6. Color Sync Section. It was indicated, in the discussion of the chrominance channel, that a 3.58-mc color subcarrier must be reinserted into the I and Q signals before they can be demodulated. It is the function of the color sync section not only to develop such a 3.58-mc signal, but to make certain it possesses the proper frequency and, what is equally important, the proper phase. For if the phase is wrong, the reproduced color will likewise be wrong, and in a color television system this is a very noticeable form of distortion.

A block diagram of the color sync section of the receiver is shown in Fig. 18-22. The input to this section is obtained from some point in the

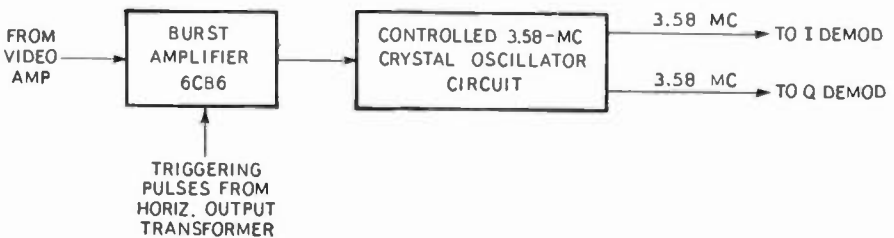


Fig. 18.22. An abbreviated block diagram of the color sync section of a color receiver. Figs. 18.23 and 18.24 illustrate the components of this section in greater detail.

luminance system, generally at the master contrast control or shortly thereafter. What the color sync section is primarily interested in obtaining is the color burst which is placed on the back porch of every horizontal sync pulse. (See Fig. 18-14.) It is this burst which contains information concerning the proper frequency and phase of the color subcarrier. The 3.58-mc burst take-off circuit generally consists of a resonant coil or trans-

former placed in the plate circuit of one of the video amplifiers in the luminance channel. This circuit is resonant to 3.58 mc and therefore it responds best to the color burst signal and progressively less to the other components of the incoming signal.

The voltage developed by the take-off circuit is applied to a burst amplifier which is gated by a pulse obtained from the horizontal output transformer. That is, the stage is cut-off except when gated by the pulse, and during this interval the color burst should be active in the system and

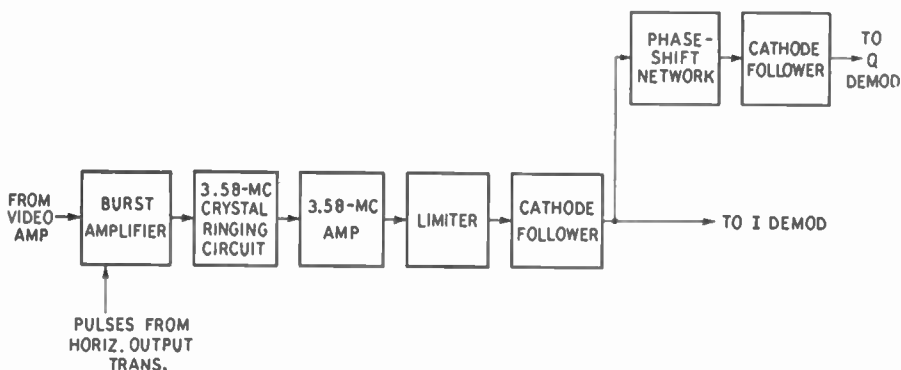


FIG. 18.23A. Block diagram of a crystal ringing system for generating 3.58-mc oscillations.

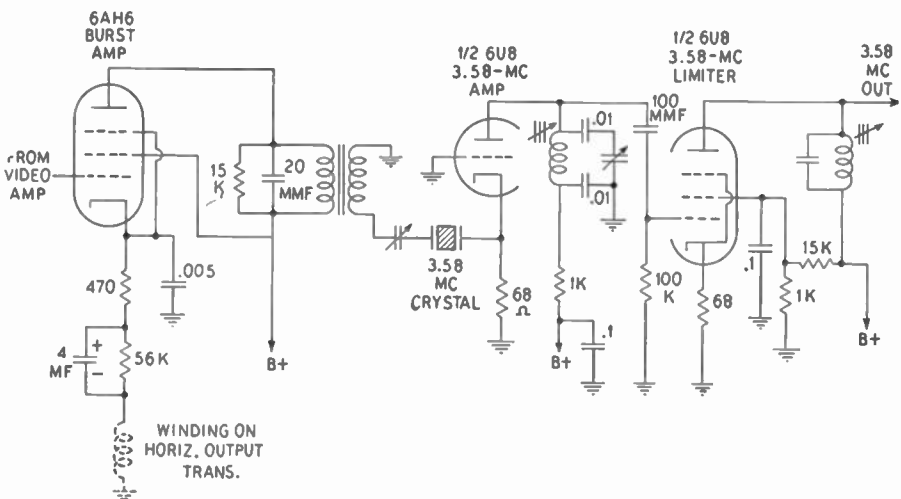


FIG. 18.23B. Partial schematic circuit of crystal ringing system.

hence pass through the burst amplifier. At the end of the gating pulse, the burst amplifier once again lapses into cut-off.

At the output of the burst amplifier, the color burst may be applied either to a crystal ringing circuit or to an automatic phase control (A.P.C.) circuit. Both systems are in use and both will be described here.

The Crystal Ringing Circuit. This system uses a quartz crystal which, when excited by the color burst at the start of each horizontal line, will continue to "ring" or oscillate at its natural frequency (here 3.58 mc) for the duration of the line. A typical circuit is shown in Fig. 18-23. The burst from the burst amplifier activates the quartz crystal and, because of its extremely high Q , it continues to oscillate with very little decrease in amplitude until the next burst arrives. The trimmer in series with the crystal can change its resonant frequency by several hundred cycles and thus take care of crystal tolerances.

The stage following the crystal is an amplifier stage, and the stage beyond that is generally a limiter to smooth out variations in output of the ringing circuit. Output from the limiter may be used as one of the 3.58-mc driving voltages for the I or Q demodulators, while the same output, passed through a 90-degree phase-shift network will provide the reference voltage for the other demodulator.

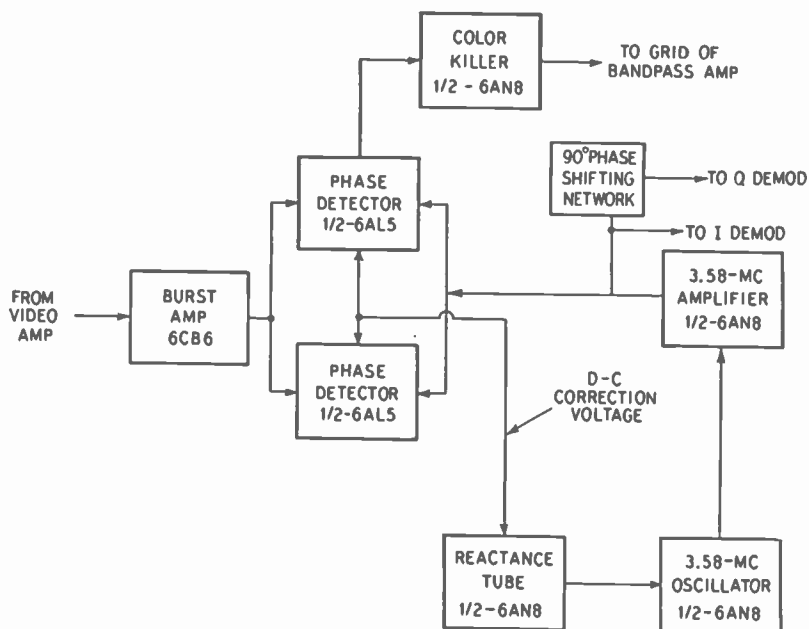


Fig. 18.24. A color sync section using an automatic phase control system. This is the method employed in the receiver of Fig. 18.36.

Automatic Phase Control System. A second approach to the development of a 3.58-mc subcarrier whose frequency and phase are locked-in to that of the color burst is by means of an automatic phase control system. This system, shown in Fig. 18-24, closely resembles the horizontal A.F.C. systems currently employed in black-and-white receivers. That is, the frequency and phase of the color burst are compared with the frequency and phase of the signal developed by a free-running, 3.58-mc oscillator. If any difference exists between the two, a corrective d-c voltage is developed and applied to a reactance tube which is connected across the oscillator's resonant circuit. The reactance tube, in response to the correction voltage, varies its plate current in such a way as to bring the oscillator frequency in line with the color burst.

The oscillator output is fed to a 3.58-mc amplifier which serves to strengthen the signal and, at the same time, acts as a buffer between the oscillator and the rest of the circuit. From the amplifier, one portion of the signal goes to the phase detector, one portion to the *I* demodulator, and one portion to the *Q* demodulator.

A color killer circuit is also part of the color sync section, and its

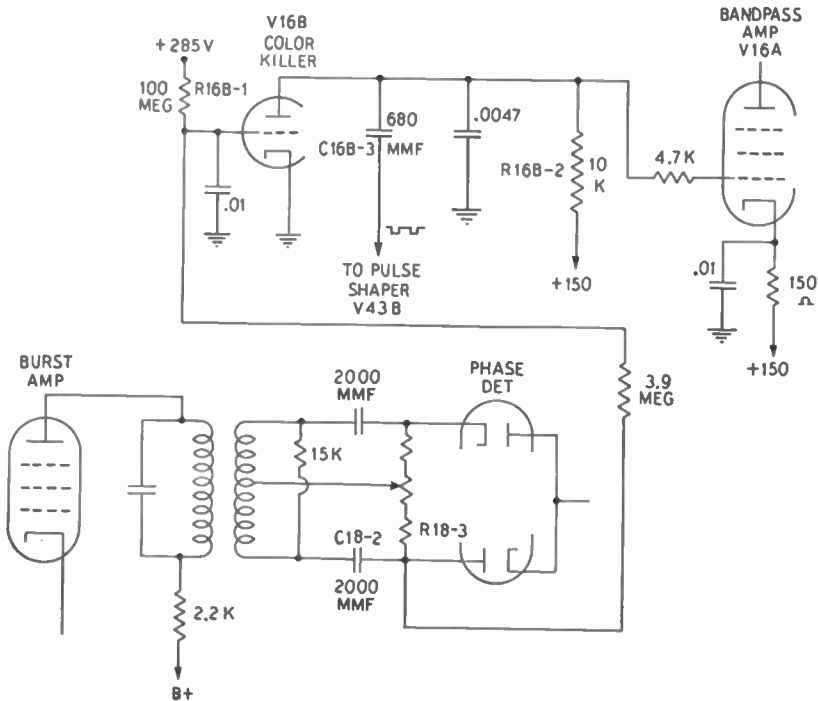


FIG. 18.25. The color killer circuit of Fig. 18.36. (See text for explanation.)

operation in the receiver of Fig. 18-36 is as follows. (A simplified diagram of the color killer stage is given in Fig. 18-25.)

The grid of the color killer tube, V16B, is tied into one side of the phase detector. With a color signal being received, the voltage at the tie-in point (junction of C18-2 and R18-3) is sufficiently negative to bias the color killer tube to cut-off. This removes its effect from the circuit and permits the 6AN8 bandpass amplifier, V16A, to function normally. The cathode of this tube has a positive 150 volts applied to it. This is offset by the +150 volts applied to the grid of the tube through resistor R16B-2, 10,000 ohms.

When the color burst is active, a negative pulse is fed into the grid circuit of V16A via capacitor C16B-3. The pulse drives the bandpass amplifier to cut-off and prevents the color burst from passing through.

In the absence of a color signal and, consequently, color bursts, the highly negative potential at the junction of C18-2 and R18-3 disappears, and now the color killer conducts heavily because its grid is made positive by the presence of the positive voltage applied to one end of grid resistor R16B-1. This flow of plate current causes the plate voltage to drop considerably below 150 volts. Since the grid of V16A is tied to the plate of the color killer, it assumes the same potential. The cathode of V16A is 150 volts positive, and the difference between this value and the less positive grid is sufficient to prevent V16A from conducting.

7. Sync Separators and A.G.C. Stages. The sync separators and A.G.C. section of a color television receiver do not differ in any important aspects from the same stages in monochrome receivers. (See Fig. 18-26.)

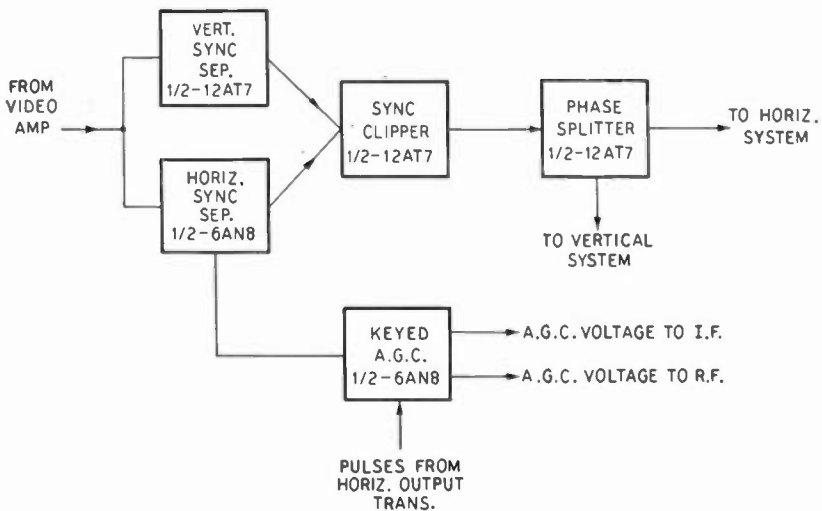


Fig. 18.26. The sync separators and A.G.C. stage of the color receiver of Fig. 18.36.

Thus, the sync separators have, as their function, the separation of the horizontal and vertical sync pulses from the rest of the video signal. Once this is accomplished, then the pulses are applied to the horizontal and vertical sweep systems through appropriate integrating and differentiating networks.

For the A.G.C. section, any method which has been employed in monochrome sets may be utilized in color receivers. At the present time, keyed A.G.C. is favored, but this preference stems not from the fact that color television is being used, but from the inherent characteristics of keyed A.G.C. itself.

The tubes controlled by the A.G.C. voltage include the R.F. amplifier and one or more video I.F. stages. Clamping of the A.G.C. voltage fed to the R.F. amplifier may also be employed.

8. Vertical and Horizontal Deflection Systems and the High Voltage Circuits. In the horizontal and vertical deflection circuits we again en-

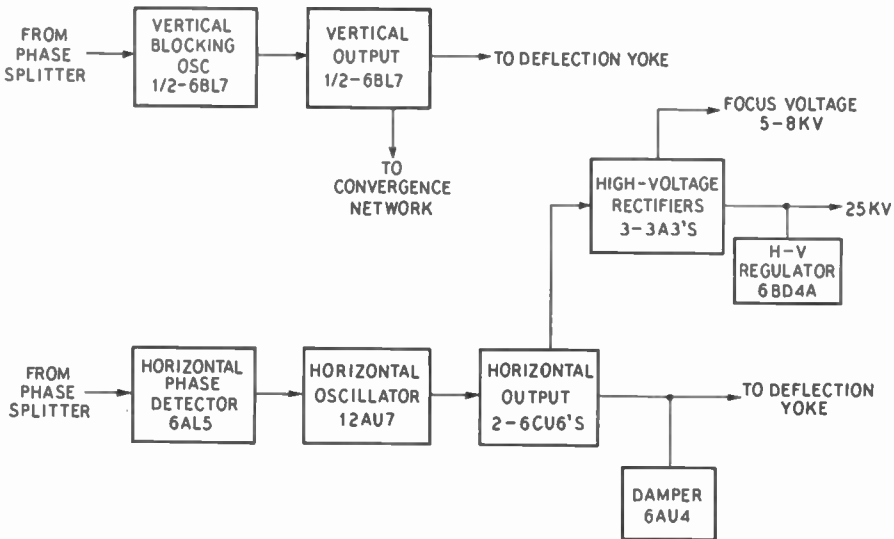


FIG. 18.27. Block diagram of the vertical and horizontal deflection systems of the color receiver of Fig. 18.36.

counter circuits familiar from monochrome receiver practice. (See Fig. 18-27.) The vertical system consists of a blocking oscillator and an output amplifier. In the horizontal section, the oscillator is preceded by an automatic frequency control system. Beyond the oscillator there may or may not be a discharge tube, after which we have the output amplifier and the horizontal output transformer. A damper tube is connected across the horizontal windings of the deflection yoke to eliminate any oscillations

which occur during beam retrace. The energy absorbed by the damper tube in this process is converted into an additional voltage which, when added to the normal B+ voltage, results in a boost B+.

The high voltage requirements of the three-gun, shadow-mask picture tube are considerably more critical than they are for a conventional black-and-white picture tube. The color tube requires 25,000 volts at a maximum current drain of 3000 microamperes. There also must be available to the tube a focus voltage variable between 5 to 8 kilovolts.

The heavy requirements of the picture tube in regard to beam current result in some serious problems when designing a combination deflection and high voltage system. Since the power used by the high voltage circuit is now an appreciable portion of the total, changes in beam current due to picture brightness changes can cause variations in scanning linearity and in the various operating potentials of the tube itself.

To avoid such variation it is necessary to maintain the high voltage load constant regardless of whether the picture is bright or dim. A simple method of achieving this stabilization is by means of a special vacuum-tube triode (6BD4-A). The plate of the tube connects to the 25,000-volt output of the high voltage system, while the cathode is returned to a positive low-voltage point — in this instance, 400 volts (Fig. 18-28). The grid is then tapped into a bleeder network hung between the high voltage system and ground.

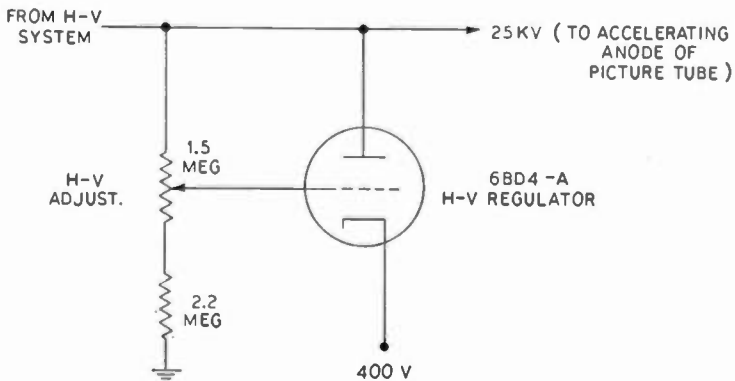


FIG. 18.28. Simplified circuit of high-voltage regulator.

Operation of the circuit is as follows: If the high voltage rises, because, perhaps, of less current drain by the picture tube, then this increase, in part, will be transmitted to the grid of the regulator triode because of the grid tap on the high voltage bleeder string. A more positive grid means

increased tube current flow and, if the circuit has been properly designed, this increased current will just take up the slack shed by the picture tube and bring the high voltage down to its correct level.

On the other hand, when the picture tube draws more current, the high voltage has a tendency to drop. This drop lowers the voltage across the bleeder, providing less positive voltage for the shunt regulator and thereby driving its grid more negative. The more negative grid reduces the current drawn by the regulator and tends to counteract the increased picture tube current. Again, the high voltage system sees a fairly constant load, and its voltage value remains stable.

Still remaining in the color receiver diagram of Fig. 18-15 is a block labeled convergence circuits. Why these are needed and what they do are best discussed after certain features of the tri-gun color picture tube have been examined. This follows in the next section.

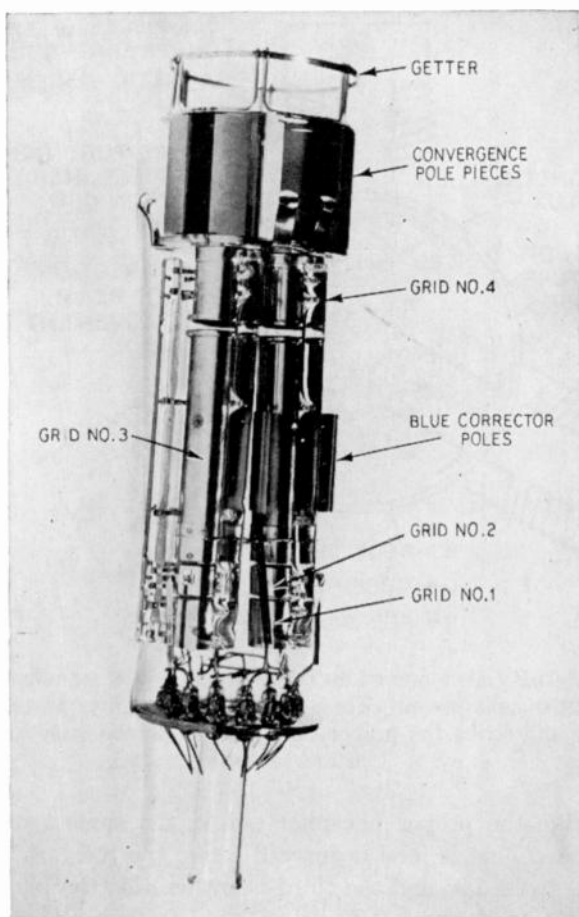
9. Tri-Gun Color Picture Tubes. All of the color receivers which have been commercially marketed thus far have employed a tri-gun, tri-color picture tube. The tri-gun portion of the name indicates that the tube possesses three electron guns. By comparison, the conventional black-and-white tube employs only one electron gun.

The second half of the name, tri-color, is linked to the fact that the screen of the tube possesses three different color-emitting phosphors. This, of course, is basic to the entire color television system since we employ the three primary colors — red, green, and blue — to synthesize the wide range of hues and tints required for the satisfactory presentation of a color picture.

The tie-in between the three electron guns and the three different types of screen phosphors now becomes evident. Each gun is concerned with only one type of phosphor. Thus, one of the electron guns develops an electron beam which strikes, say, the phosphor which emits red light. This gun may be labeled as the red gun. A second electron gun directs its beam only at the green phosphor dots and it is the green gun. The final gun is concerned in similar manner only with the blue dots. Note that in each case, it is not the color of the phosphor to which we refer, but to the light which this phosphor gives off when actuated by an electron beam. The actual color of the substance and the color of the phosphorescent light it emits do not necessarily bear any relationship to each other.

The overall color that is seen on the screen is determined by two general factors: (1) The phosphors which are being bombarded by the three guns, and (2) the number of electrons which are contained in each beam. Thus, suppose you turn one beam off completely — say, the beam from the red gun. Then, only the blue and green dots will be emitting light, and what you will see is a mixture of blue and green light which can range all the

control grid (grid No. 1), an accelerating (or screen) grid (grid No. 2), a focusing electrode (grid No. 3), and a converging electrode (grid No. 4). The heaters of all three guns are in parallel and require only two external connections to the tube base. Each control grid has its own base pin, and the same is true of each screen electrode. The focusing electrodes (grid No. 3) of all the guns are electrically connected together because one overall voltage variation will bring all three beams to a focus at the phosphor dot screen.*

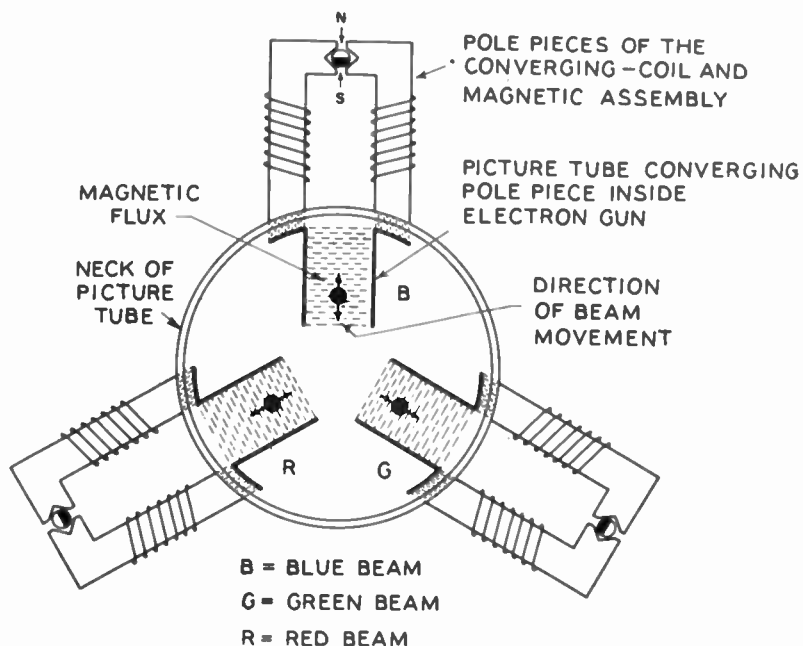


Courtesy CBS

FIG. 18.30. Electron gun assembly of 19-inch color picture tube. Three electron gun structures are employed although only two are clearly visible.

*Grid No. 3 of the blue gun has two special built-in pole pieces. These function in conjunction with an external magnet to assist in converging the beams at the center of the screen. More on this presently.

The final electrode in the gun structure is grid No. 4, the converging grid. This is a small-diameter cylinder which is internally connected to (and operated at the same high potential as) the aquadag coating (25,000 volts). Also associated with each No. 4 grid is a pair of pole pieces. These are mounted above each grid. External coils on the neck of the tube induce magnetic fields in each set of poles pieces, as shown in Fig. 18-31. These fields force the three beams to converge or come together so that each



Courtesy CBS

FIG. 18.31. External coils mounted on the neck of the picture tube induce magnetic fields in each set of pole pieces. These fields force the three beams to converge so that each beam will strike the proper phosphor dot in the same group at any one instant of time.

beam will strike the proper phosphor dot in the same group at any one instant of time. That is, one beam will strike the red dot, a second beam will strike the green dot, and the third beam will hit the blue dot, all three dots being in the same group. The three dots are bunched so close together that the light they produce combines and appears to the eye as a single color. In the absence of this converging action, it would be possible for the beams to hit phosphor dots at sufficiently separated points so that an observer would see three individual points of light. Under these conditions, mixing of colors to obtain different hues would not be possible.

The Shadow Mask. Proper beam convergence is an important aspect of tri-gun picture tube operation. Further to insure that each beam strikes only one type of phosphor dot, a mask (known as a shadow mask) is inserted between the electron guns and the phosphor dot screen. (See Fig. 18-32.) The mask is positioned in front of and parallel to the screen. The mask contains circular holes, equal in number to the dot trios. Each hole is so aligned with respect to its group that any one of the approaching beams can only "see" and therefore strike one phosphor dot. The remain-

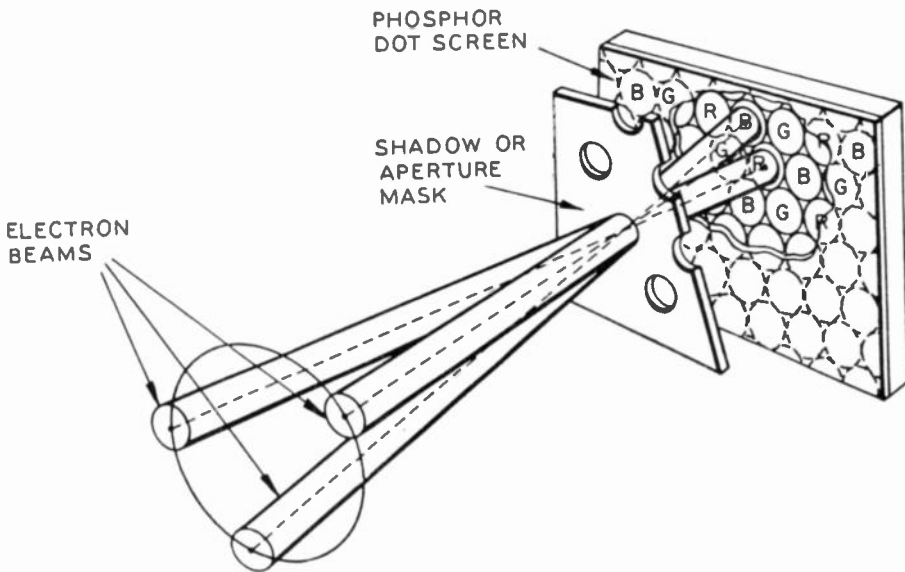


FIG. 18.32. Diagrammatic illustration of the mask and screen showing the convergence of the three beams at a single hole in the aperture mask. Note that the converged beams pass through the hole and strike their respective phosphor dots.

ing two dots of the trio are hidden by the mask. That is, the two other dots are in the "shadow" of the mask opening—hence the name of shadow mask.

What is true for one beam is true for the other two beams. Each of these can also "see" one phosphor dot. In this way, we are able to minimize color contamination which would occur when a beam either hit the wrong dot or overlapped several dots at the same time.

Static and Dynamic Convergence. In the foregoing introductory discussion, beam convergence was covered in a general manner. Actually, as we shall see now, there are two types of convergence; static convergence and dynamic convergence. In static convergence, we adjust the positions

of the beams, using either fixed d-c voltages or fixed magnetic fields, depending on the type of gun. As a further aid in this action, the electron guns in the 19- and 21-inch tubes are tilted inward slightly. And if the adjustments are made carefully, the beams will converge properly over the central area of the screen.

However, to maintain this converged condition of the beams as they swing away from the center, it is also necessary to vary their relative angles slightly. This process of changing the beam angle in step with the scanning is referred to as dynamic convergence. It is required because the

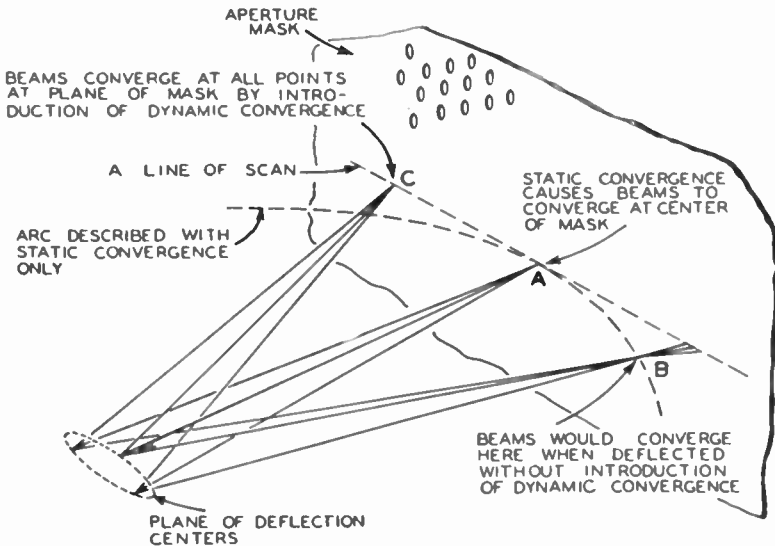


Fig. 18.33. Dynamic convergence voltages are required to cause the three beams to converge at the picture edges because the screen and aperture (or shadow) mask are not perfectly spherical.

distance traveled by the beams increases as they swing away from the center of the screen. The swing away from the center, in turn, occurs because the curvature of the screen is not perfectly spherical, and beams which are converged at the screen center will tend to converge in front of the shadow mask at points away from the center. (See Fig. 18-33.)

A moment's reflection will reveal that the extent of convergence changes the farther the beams are from the center of the screen. Furthermore, there is a direct relationship between the convergence needed at any one point and the instantaneous horizontal and vertical deflection voltage values. Thus it is possible to obtain whatever correction voltages are

needed from the vertical and horizontal deflection systems. These additional voltages are known as dynamic convergence voltages in distinction to the d-c or static convergence adjustment which is made over the central area. Where magnetic convergence means are employed, the static adjustment is made with a permanent magnet. The dynamic convergence is then achieved by introducing varying magnetic fields via converging coils mounted on the neck of the picture tube.

The basic form of the correcting current is parabolic, as shown in Fig. 18-34. When the three beams are in the center of the screen, the correction current is zero. On either side of center, however, the current varies, and the combined effect of the correction (that is, dynamic) and static fields is to keep the beams properly converged at every point of the screen.

We can now appreciate what purpose the block labeled "convergence circuits" in Fig. 18-15 serves. In

Fig. 18-36, tubes V41A, V41B, V42A, V42B, and V43A represent the stages which develop the convergence currents required by the color picture tube.

External Picture Tube Components. We come now to the components which are mounted on the neck of the tri-gun, tri-color picture tube. (See Fig. 18-35.) The first item that we recognize is the deflection yoke. This is, to a considerable extent, similar to the deflection yoke used with a black-and-white tube. However, its design is more complex because three beams must be deflected instead of one, and it is of the utmost importance that a symmetrical and uniform magnetic field be maintained throughout the deflection area.

A second component found on the neck of the color picture tube is the purity coil or magnet. This device adjusts the axis of each electron beam so that it approaches each hole in the shadow mask at the right angle to strike the appropriate color phosphor dot. In other words, the purity magnet provides for the proper alignment of the three beams with respect to the phosphor-dot plate and the shadow mask. When this component is properly set, a uniform color field will be obtained for each gun. For example, with only the red gun in operation a uniform red raster should be observed. Any departure from pure red at any point on the screen indicates that the beam is striking phosphor dots other than red. Similarly, when only the green gun is in operation, a uniform green raster should be obtained, and when only the blue gun is active, a blue field should be visible.

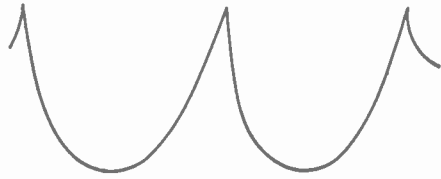
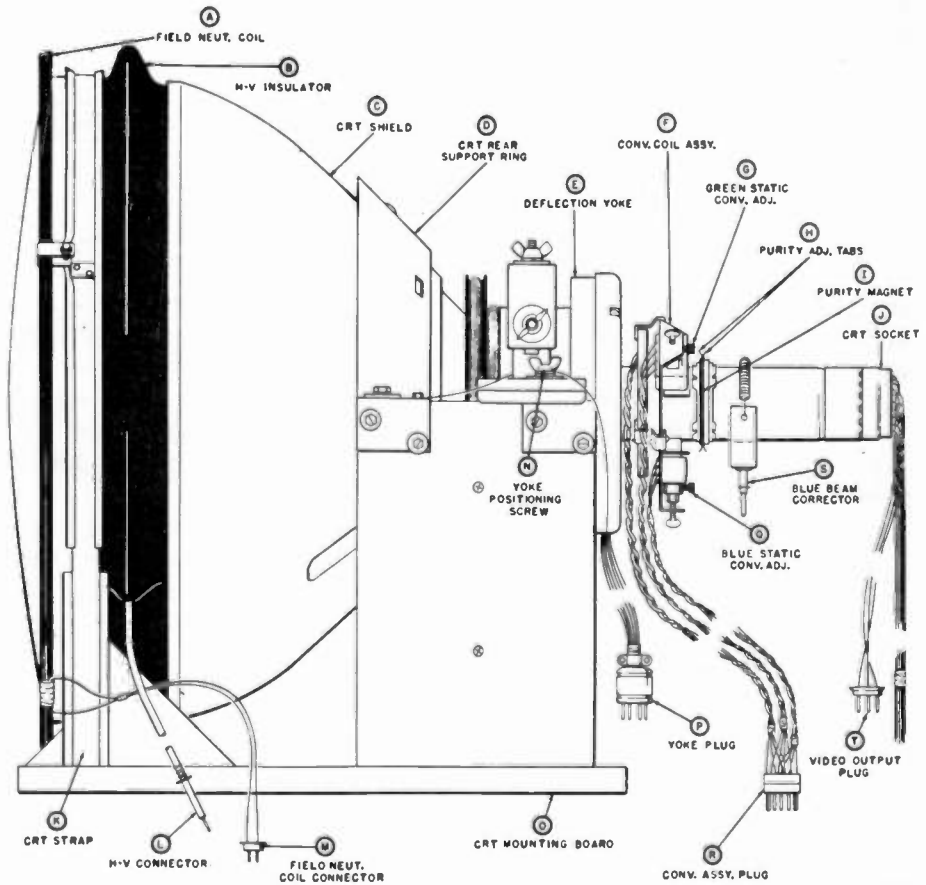


FIG. 18.34. The form of the dynamic convergence voltage (or current) is parabolic.

The larger screen color tubes which we are most concerned with utilize magnetic convergence and, toward that end, employ three sets of convergence coils, each positioned directly over the pole pieces which are internally associated with each grid No. 4. The magnetic fields set up by the coils are coupled through the glass neck of the tube to the internal pole pieces. The latter serve to shape and confine the fields so as to affect only the particular electron beams to which the individual pole pieces correspond. For example, the change in convergence angle of the red beam is a function only of the current through the external coil which couples to the internal set of pole pieces adjacent to the red beams. Likewise, the current through the green and blue external magnets affects respectively only the green and blue beams.



Courtesy CBS

FIG. 18.35. 19-inch color picture tube with its external components.

Each external coil possesses two separate windings to provide for horizontal and vertical dynamic convergence correction. For the static convergence adjustment, each coil has associated with it a small permanent magnet whose position can be varied.

A diagram of the individual dynamic convergence controls is shown in Fig. 18-31. The heavy dots represent the individual electron beams as they pass through the gun on their way to the screen. The arrows at these beams indicate their direction of movement. Note that the red and green beams are confined to paths which make an angle of 30 degrees on either side of a perpendicular axis. The blue beam, on the other hand, can move only vertically, up or down.

Now it can readily happen that, although the color dots of the green and red beams fall within the trio of phosphors, that of the blue beam does not. This means that, while we can always cause the red and green beams (or color dots) to converge, it may not be possible to have the blue beam meet the other two. Still required is another adjustment, that of being able to move the blue beam from side to side (or laterally). To effect this, a special blue beam positioning magnet is also found on the neck of the tube. Now perfect convergence of the three beams at the center of the screen is always achievable.

Note that no ion traps are used in this tube, principally because the color screen is aluminized. The layer of aluminum presents a barrier to any oncoming ions and prevents them from reaching and damaging the screen. Electrons, having only $\frac{1}{1800}$ th of the mass of an ion, encounter little difficulty in passing through this aluminum layer.

QUESTIONS

1. What advantages are offered by the use of color in presenting a television picture?
2. Name the primary colors. What is their significance?
3. What is a chromaticity chart and what purpose does it serve?
4. What colors are found near the edges of the chromaticity chart? What colors near the center?
5. Explain the meanings of hue, saturation, and purity.
6. Does the normal eye see color equally well in large as well as small areas? Explain.
7. When we say that a television signal extends from 0 to 4 mc, do we mean that it occupies every cycle of the 4-mc band? Explain.
8. How is it possible to transmit a full color signal in the same space occupied by a monochrome signal?
9. What information does the monochrome signal carry? the color subcarrier and its sidebands?
10. Why are *I* and *Q* signals better suited to color rendition than *R-Y* and *B-Y*?

11. How is information concerning the frequency and phase of the missing subcarrier supplied to the receiver?
12. Draw a block diagram of a color television receiver using a tri-gun picture tube.
13. Which sections of the color block diagram would not be found in a monochrome receiver? Explain why in each instance.
14. What precautions must be observed with respect to sound take-off from the video system?
15. List all of the sections in the receiver to which a portion of the detected video signal must be fed.
16. With which portion of the incoming signal is the color sync section specifically concerned? How does it use this information?
17. Trace the path of the monochrome portion of a color video signal from the second detector to the picture tube.
18. What is the function of the chrominance section of the receiver?
19. Describe where each of the following stages are found and indicate what they do.
 1. Bandpass amplifier
 2. burst gate amplifier
 3. color killer
 4. *I-Q* demodulators
20. Describe briefly two methods of developing suitable 3.58-mc subcarrier signals.
21. Why are special convergence circuits required in the color receiver?

GLOSSARY OF TELEVISION TERMS

A

- Active Lines**—In scanning an image, those lines that are responsible for imparting the information of the image. The beam is inactive when moving rapidly from right to left, or from the bottom of the picture to the top.
- Amplitude Modulation**—The process of varying the amplitude of a carrier in accordance with the amplitude of the intelligence (sound or video) to be broadcast.
- Aquadag**—A substance composed of graphite in solution. A coating of this substance is usually placed on the inside of a cathode-ray tube to prevent the accumulation of free electrons on the walls of the tube.
- Array**—A group or combination of two or more antenna elements, functioning as a unit. Antenna arrays are generally designed to provide greater directivity and gain than are available from a single-element antenna.
- Aspect Ratio**—A term used to denote the ratio of the image width to the image height.
- Automatic Frequency Control**—A method of automatically maintaining the frequency of an oscillator at a desired value. In television receivers, automatic frequency control is extensively applied to the horizontal sweep oscillator.
- Automatic Gain Control (A.G.C.)**—A method whereby the overall gain of a television receiver is automatically adjusted to produce a constant output signal for varying input signals.

B

- Back Porch**—The section of the synchronizing signal extending from the back end of the horizontal sync pulse to the start of the next line.
- Band**—A group of frequencies having definite upper and lower frequencies.
- Bandwidth**—The numerical difference between the highest and lowest frequencies passed by an electrical circuit.
- Beam**—When referring to television, a stream of electrons traveling in a vacuum tube. The tube to which this expression is most generally applied is the image tube.
- Bi-directional**—A term that applies to the response characteristic of an antenna and signifies that signals may be received from (or transmitted to) two different directions.

- Blacker-than-Black Region**—The region where the blanking and synchronizing voltages are found in the video signal. The voltages in this region prevent any electrons in the cathode-ray tube from reaching the viewing screen. The result is an absence of light on the screen.
- Black Level**—The point in the video signal at which the electron beam in the image tube is cut off or prevented from reaching the screen. This point normally occurs at 75% of the maximum amplitude of the signal.
- Blanking Pulses**—Sharp rises in voltage that bias the viewing tube control grid beyond cut-off. This action, when properly correlated with the brightness control, prevents the electron beam retraces from appearing on the viewing screen.
- Blooming**—The mushrooming of an electron beam (with consequent defocusing) produced by too high a setting of the brightness control.
- Booster Amplifier**—An R.F. amplifier which is not an integral part of a television receiver but which is connected to the front end of the set in order to “boost” or raise the strength of the received signal. Booster units usually contain more than one R.F. amplifier.
- Brightness Control**—A potentiometer control that varies the average or background illumination of the received image. When properly set, this control prevents any beam retraces from appearing on the screen.
- Brilliance Control**—Another name for the brightness control.

C

- Camera Tube**—A tube which has the ability to transform incoming light rays into equivalent electrical charges. A scanning beam then utilizes these charges to produce a video signal.
- Cathode-ray Tube**—A vacuum tube that contains a fluorescent screen at one end. By directing an electron beam at this screen, visible traces are formed and a combination of these gives rise to the reproduced image. Also known as Kinescope (RCA trade name).
- Centering Control**—A potentiometer which enables the viewer to move the image back and forth or up and down on a screen. Horizontal and vertical centering each requires a separate control.
- Channel**—The group of frequencies over which the video and audio signals may be transmitted. Each television channel is 6 mc wide.
- Clamping Circuit**—Another name for a d-c restorer. Clamping circuits in American television receivers maintain the tips of the sync pulses at a common level.
- Clipper**—A circuit which does not permit the positive (or negative) level of signal to exceed a certain value. The most extensive use of clippers in television receivers is the separation of the sync pulses from the rest of the video signal.
- Coaxial Cable**—A transmission line in which one conductor (the outer) completely surrounds the other conductor (the inner). The two conductors are separated by air or some other insulator.
- Contrast**—The ratio between the brightest and darkest portions of an image.
- Contrast Control**—A potentiometer that permits variations of the intensity of the various elements of an image. May be used to accentuate the highlights and shadows in an image.
- Cross-over Area**—In an electron gun, a region in the first lens system where the cathode-emitted electrons are brought together under the influence of electric (and sometimes magnetic) fields.

D

- Damping Tube**—A tube used with magnetic deflecting coils to prevent any transient oscillations from being set up in the coils or the associated circuits.
- D-c Reinsertion**—The readjusting of the video signal at the image tube to the form it possessed when transmitted by the broadcast station. Passage of the video signal through the a-c coupling networks of the video amplifiers results in the removal of the d-c component of the signal. A d-c restorer circuit just prior to the cathode-ray tube reinserts this lost d-c component.
- Deflection**—The shifting of the electron beam in a cathode-ray tube away from the center of the screen.
- Demodulation**—Another name for detection, or obtaining the intelligence contained in a signal from the carrier.
- Detail**—The smallest or minutest portions of an image that can be distinguished from each other.
- Differentiating Circuit**—A grouping of components that possess the ability to produce an output voltage proportional to the rate of change of the input signal.
- Dipole**—An antenna wire, one-half wavelength long at the frequency it is to be used. It is usually divided into two equal quarter-wave sections.
- Directive Antenna**—Any antenna system that tends to receive signals best from one or more but not all directions.
- Director**—An antenna wire placed in front of a dipole. Its purpose is to sharpen the directivity of the dipole and increase its gain.
- Discharge Tube**—A tube which is biased to cut-off and therefore non-conducting except when triggered by a positive pulse. A condenser, connected in the plate circuit of the tube, charges when the tube is non-conducting and discharges when the tube is triggered and forced into conduction.
- Discriminator**—The second detector in an F-M superheterodyne receiver. The frequency variations in the F-M signal are here converted to amplitude variations, suitable to be heard on a loudspeaker.
- Driven Element**—The wire or wires of an antenna which are connected to the receiver (or transmitter).

E

- Electric Field**—The region surrounding charged particles. An electric field is set up also whenever a magnetic field varies. Radio waves traveling through space are composed of electric and magnetic fields.
- Electromagnetic Deflection**—Deflecting an electron beam by applying varying currents to coils mounted on the neck of a cathode-ray tube.
- Electromagnetic Deflection Coil**—A circular coil placed around the neck of some cathode-ray tubes to cause deflection of the electron beam. Generally enclosed in an iron core known as a yoke.
- Electron Gun**—The cathode, grid, and cylindrical plates in a cathode-ray tube that produce the electron beam.
- Electron Multiplier**—A series of anodes used in the Image Dissector camera tube to increase the intensity of the video signal.
- Electron Scanning**—The periodic deflection of an electron beam across the screen of a cathode-ray tube, following a definite pattern.
- Electrostatic Deflection**—Deflecting an electron beam by applying varying voltages to plates mounted inside the glass envelope of a cathode-ray tube.

Equalizing Signals—A series of six pulses before and after a serrated vertical pulse. The action of these pulses causes the vertical deflection to start at the same time in each interval.

F

Field Frequency—This term is used in interlaced scanning and refers to the portion of a complete frame when either the even or odd lines are scanned. Requires $1/60$ of a second.

Flicker—The variation in brightness that can be seen by the human eye when the images presented on a screen do not follow each other with sufficient rapidity.

Fluorescent Screen—The coating located at one end of the cathode-ray tube on which the image is produced.

Flyback—The rapid retrace motion of an electron beam after every horizontal line from the right- to the left-hand side of the image and after every field, from the bottom to the top of the image.

Focusing Control—The control whereby the electron beam is made to meet the fluorescent screen in a small, well-defined spot.

Folded Dipole—A half-wave antenna consisting of two dipoles in parallel with each other. Connection to the array is made at the center of one of the dipoles.

Frame Frequency—The rate at which a complete image is scanned. This includes both even- and odd-line fields. The rate is 30 frames per second.

Frequency Modulation—A means of transmitting radio intelligence by varying the frequency of the wave.

Front Porch—The portion of the synchronizing signal extending from the end of a line to the start of the horizontal sync pulse.

G

Gamma—The ratio of brightness variation in the reproduced image to the brightness variation in the original scene.

Ghost Image—A second image appearing on the receiver screen, superimposed on the desired signal. These images are caused by reflected rays arriving at the receiving antenna some small time interval after the desired wave.

Ground Grid Amplifier—An amplifier in which the control grid is at R.F. ground potential and the signal is fed into the cathode circuit.

Ground Wave—A radio wave that travels close to the earth.

H

Halation—A ring or halo of light surrounding a point where the electron beam is impinging and producing a dot of light. The ring is due to multiple reflections between the front and back surfaces of the glass screen.

Height Control—The control whose setting determines the vertical amplitude of the image.

Hold Control—The variable resistor that permits adjustment of the synchronizing oscillator until the latter frequency nearly equals that of the incoming synchronizing pulses.

Horizontal Blanking—The interruption of the electron beam during horizontal retrace.

- Horizontal Centering Control**—A control that enables the operator to move the television image in a right or left direction across the screen.
- Horizontal Line Frequency**—The number of horizontal lines per second: 15,750.
- Horizontal Resolution**—The number of light variations or picture elements along a line which can be distinguished from each other.
- Horizontal Retrace**—The rapid right-to-left motion of the electron beam at the end of each line.

I

- Iconoscope**—An image camera tube that receives the light rays of the scene being televised and converts this energy into electrical charge.
- Image Dissector**—Another television camera tube that serves the same function as the Iconoscope and Orthicon tubes, but operates differently.
- Image Orthicon**—A highly sensitive camera tube used extensively for indoor and outdoor telecasting.
- Implode**—The action of the highly evacuated cathode-ray tube when the glass envelope is suddenly broken. Due to the atmospheric pressure against all sides of the tube, the glass moves inward with tremendous force.
- Impulse**—A sudden rise and fall of current (or voltage) in an electrical circuit.
- Integrating Circuit**—A circuit, consisting generally of condensers and resistors, whose output voltage is equal to the sum of a number of input pulses.
- Intercarrier System**—A system in which the 4.5-mc difference between video and sound carriers is used as the I.F. frequency of the sound system.
- Interlaced Scanning**—A method whereby an image is scanned first along the odd-numbered lines and then along the even-numbered lines. The result of interlaced scanning is an apparent increase in the rate at which the picture is sent. Flicker, by this means, is reduced to a minimum.
- Ion Spot**—An area on the fluorescent screen of an electromagnetic deflection image tube which has become desensitized because of a bombardment by ions in the electron beam.
- Ion Trap**—An arrangement whereby the ions within an electron beam are prevented from bombarding the screen and producing an ion spot.

L

- Limiters**—The last I.F. stage (or two) in an F-M receiver. The purpose of this stage is to eliminate all amplitude distortion or variation in the F-M signal.
- Linearity Control**—An adjustment that tends to correct any distortion in the sawtooth current or voltage waves used for deflection.
- Line Scanning Frequency**—The rate at which the lines or sections of an image are scanned. Present standards set the rate at 525 horizontal lines for each 1/30 of a second, or 15,750 per second.

M

- Marker Pip**—An identification pulse which is superimposed over a response curve to indicate the position of a definite frequency.
- Monitor**—An auxiliary circuit containing a cathode-ray tube upon which appears the image being broadcast by the transmitter. The use of this auxiliary cir-

cuit permits the engineers in the station to maintain a continual check of the signal being transmitted.

Monoscope—A television camera tube that contains a fixed pattern within the tube. The tube is used to produce pre-broadcast test patterns for testing and alignment purposes.

Mosaic—The photosensitive surface in an Iconoscope or Orthicon camera tube. It is here that the light rays are transformed into equivalent electrical charges.

Multivibrator—An oscillator used to generate saw-tooth voltage (and current) waves.

N

Negative Picture Modulation—A method of transmitting the television video signal so that all the picture values are reversed. The brightest portions of the image are represented by the least amount of voltage, while the dark sections of the image have large voltage (or current) values.

Noise—Undesirable voltages, either generated in the set or received from some external source, which produce a multitude of very small black and white spots over the entire image. These spots are often referred to as "snow."

P

Pairing—An effect in which the lines of one field do not fall exactly between the lines of the next field. When the effect is pronounced, the lines of the two fields fall directly over each other, effectively reducing the vertical definition in half.

Peaking Coil—A coil inserted in the video amplifier circuits for the purpose of improving its high-frequency response.

Pedestal—The level of the video signal at which blanking of the beam occurs.

Period—The time of one complete cycle of an alternating wave.

Persistence of Vision—A phenomenon of the human eye whereby variations in light intensity, occurring sufficiently close, tend to blend into each other and appear continuous.

Phosphor—The fluorescent material coated over the inside surface of the viewing end of a cathode-ray tube.

Pickup Tube—Another name for camera tube.

Picture Elements—The smallest portions of an image that are distinguishable from each other.

Picture Frequency—This term is synonymous with frame frequency.

Positive Picture Modulation—A method of transmitting video signals in which the brightest portions of an image are represented by the most positive voltages. Opposite to negative picture modulation.

Pulse—Another term for impulse and denotes a sudden rise and fall of voltage.

Pulse Repetition Rate—The number of times a pulse reappears each second.

Pulse Width—The time duration of a pulse.

R

Raster—The pattern of a television screen that is traced out by the electron beam. When no image is being received and the brightness control is turned up, the various lines comprising the raster can be plainly seen.

Ratio Detector—A circuit which is capable of converting frequency-modulated signals into equivalent audio signals.

- Reactance Tube**—A tube which is so connected that it appears as an inductance or capacitance to its circuit.
- Reflector**—An antenna wire placed behind a dipole and designed to improve the directional characteristics and gain of the dipole.
- Relaxation Oscillator**—An oscillator through which current flows strongly for part of a period and then ceases to flow entirely for the remainder of the period. Blocking and multivibrator oscillators perform in this manner and hence fall within this category.
- Resolution**—A term which refers to the smallest detail which can be distinguished or seen separately in a television image. Resolution is generally expressed as the maximum number of alternate black and white lines which can be distinguished in a test pattern.
- Response Characteristic**—The manner in which a circuit reacts to a set of signals. In an antenna, the response characteristic refers to the manner in which signals arriving from different directions are received. In a receiver, this same phrase refers to the way certain frequency signals are received and amplified while others are rejected.
- Retrace**—The return of the electron beam either from the right-hand side of the image to the left-hand side, or from the bottom to the top of the picture.

S

- Scanning**—The process of breaking down an image into a series of elements or groups of elements and transmitting this information in a logical manner.
- Schmidt Optical System**—An optical system employing reflectors and a corrective lens to magnify an image produced on a small cathode-ray tube screen.
- Secondary Electrons**—Electrons that are obtained from a substance because of the intensity with which other electrons (known as primary electrons) are impinging on the surface of this substance. The primary electrons arrive with sufficient force to knock loose one or more secondary electrons.
- Separation Circuit**—A circuit which will separate signals having different properties, such as amplitude, frequency, etc.
- Sequential Color Transmission**—The sending of the signals arising from differently colored sections of an image one after the other, in a certain order.
- Serrated Vertical Pulse**—The manner by which the relatively long vertical pulse is broken up into a series of pulses to permit simultaneous control of vertical and horizontal synchronizing oscillators.
- Single Side-band Transmission**—See vestigial side-band transmission.
- Smear Ghost**—Ghost images which follow the desired image so closely that the overall effect is one of smearing.
- Spurious Signal Voltages**—These are the voltages caused by the secondary emission effects of a mosaic plate.
- Staggered Tuning**—The peaking or resonating of each of the tuned circuits in an I.F. system to different frequencies. This is done to achieve a broad response.
- Synchronizing Pulses**—Voltage (and current) waveforms that keep the electron beam at the receiver in step with the camera tube electron beam.

T

- Tearing**—A synchronizing circuit disruption which causes the displacement of lines from their normal position. The visual effect is as though portions of the image had been physically torn away.
- Televisé**—To train a television camera on a scene and convert the light reaching the camera into equivalent video signals.
- Test Pattern**—A special chart on which lines and other detail are so arranged as to indicate certain characteristics of the system through which the television signal obtained from this chart passes.
- Time Constant**—The time required for the voltage or current in a circuit to rise to 63 per cent of its final value or fall to 37 per cent of its initial value.
- Time Delay**—The time required by a specific voltage or current to travel through a circuit.
- Transmission Line**—A set of two conductors possessing evenly distributed resistance, capacitance, and inductance and employed for transporting electrical energy from one point in a system to another.
- Trigger**—To set off or initiate a certain action in an electrical circuit by the application of a pulse of voltage.

U

- UHF**—An abbreviation for ultra-high frequencies. By definition, these extend from 300 to 3,000 mc.

V

- Vertical Blanking**—The interval which the electron beam is being shifted from the bottom of the image back to the top and the electron beam is prevented from reaching the screen.
- Vertical Centering Control**—An adjustment control for moving the image up or down on the viewing screen.
- Vertical Resolution**—The number of active lines in an image.
- Vertical Retrace**—The return path of the electron beam during the vertical blanking interval.
- Vestigial Side-band Transmission**—A method of transmission whereby one set of side bands is eliminated from the modulated transmitted signal.
- VHF**—Abbreviation for very high frequencies. By definition, these extend from 30 to 300 mc.
- Video Amplifiers**—Any amplifier having a uniform frequency response over a wide range. Generally this range starts at 20 cycles and extends for several megacycles.
- Video Frequency**—Any frequency obtained from the scanning by a camera tube. At the present time, the highest value is restricted to 4 megacycles, but it could be higher.

W

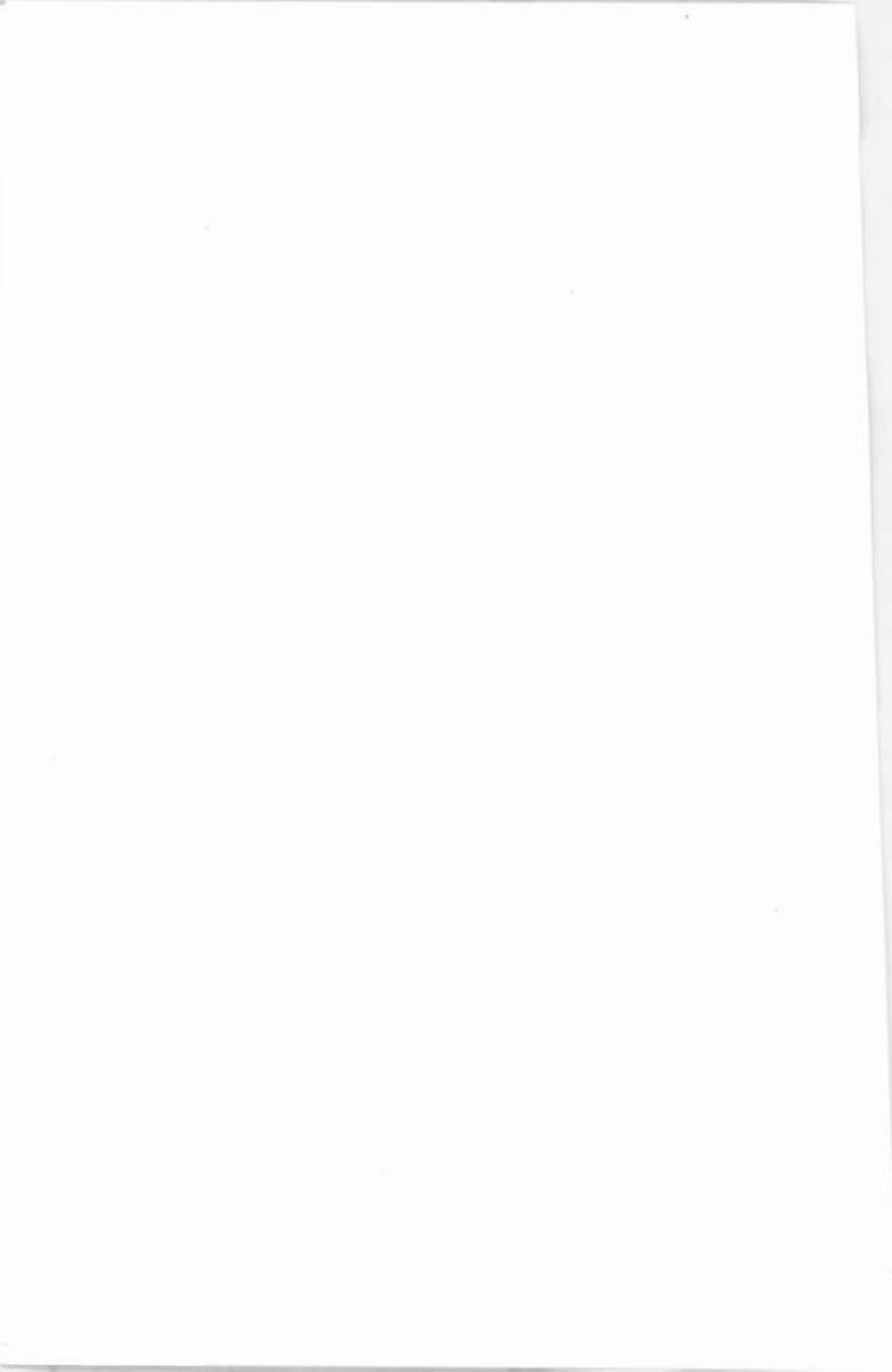
- Wedge**—Those portions of a test pattern containing a series of lines which converge at one end, this end generally positioned at the center of the pattern.

Width—The horizontal dimension of an image. It is generally expressed in inches or feet.

Width Control—The electrical component or circuit that controls the width of the image.

Y

Yoke—The frame upon which is wound the horizontal and vertical deflecting coils.



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